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# SIGNAL PROCESSING

## Through Practice

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# Contents

Introduction	iii
<b>1 Two Dice</b>	<b>1</b>
1.1 Problem . . . . .	1
1.2 Uniform Distribution: Rectangular Function . . . . .	1
1.3 Sum of Random Variables: Convolution . . . . .	2
1.4 The Triangular function . . . . .	3
1.5 The $Z$ -transform . . . . .	4
<b>2 Pingala Series</b>	<b>7</b>
2.1 JEE 2019 . . . . .	7
2.2 Pingala Series . . . . .	8
2.3 Linear Time Invariant System . . . . .	10
2.4 Power of the $Z$ transform . . . . .	12
2.5 Convolution . . . . .	14
<b>3 Circuits and Transforms</b>	<b>17</b>
3.1 Definitions . . . . .	17
3.2 Laplace Transform . . . . .	17
3.3 Initial Conditions . . . . .	22

3.4	Bilinear Transform . . . . .	25
4	Frequency Modulation	33
4.1	Message . . . . .	33
4.2	Transmitter . . . . .	38
A	Discrete Fourier Transform	41
A.1	FFT . . . . .	41
A.2	FFT . . . . .	42
A.3	Power Spectral Density(PSD) . . . . .	46
B	Filter Design	47
B.1	Introduction . . . . .	47
B.2	Filter Specifications . . . . .	47
	B.2.1 The Digital Filter . . . . .	47
	B.2.2 The Analog filter . . . . .	48
B.3	The IIR Filter Design . . . . .	48
	B.3.1 The Analog Filter . . . . .	49
	B.3.2 The Digital Filter . . . . .	51
B.4	The FIR Filter . . . . .	55
	B.4.1 The Equivalent Lowpass Filter . . . . .	55
	B.4.2 The Kaiser Window . . . . .	57
	B.4.3 The FIR Bandpass Filter . . . . .	58



# Introduction

This book introduces digital communication through probability.



# Chapter 1

## Two Dice

### 1.1. Problem

Two dice, one blue and one grey, are thrown at the same time. The event defined by the sum of the two numbers appearing on the top of the dice can have 11 possible outcomes 2, 3, 4, 5, 6, 6, 8, 9, 10, 11 and 12. A student argues that each of these outcomes has a probability  $\frac{1}{11}$ . Do you agree with this argument? Justify your answer.

### 1.2. Uniform Distribution: Rectangular Function

1.2.1. Let  $X_i \in \{1, 2, 3, 4, 5, 6\}$ ,  $i = 1, 2$ , be the random variables representing the outcome for each die. Assuming the dice to be fair, the probability mass function (pmf) is expressed as

$$p_{X_i}(n) = \Pr(X_i = n) = \begin{cases} \frac{1}{6} & 1 \leq n \leq 6 \\ 0 & \text{otherwise} \end{cases} \quad (1.2.1.1)$$



## 1.3. Sum of Random Variables: Convolution

1.3.1. The desired outcome is

$$X = X_1 + X_2, \quad (1.3.1.1)$$

$$\implies X \in \{1, 2, \dots, 12\} \quad (1.3.1.2)$$

The objective is to show that

$$p_X(n) \neq \frac{1}{11} \quad (1.3.1.3)$$

1.3.2. Convolution: From (1.3.1.1),

$$p_X(n) = \Pr(X_1 + X_2 = n) = \Pr(X_1 = n - X_2) \quad (1.3.2.1)$$

$$= \sum_k \Pr(X_1 = n - k | X_2 = k) p_{X_2}(k) \quad (1.3.2.2)$$

after unconditioning,  $\because X_1$  and  $X_2$  are independent,

$$\Pr(X_1 = n - k | X_2 = k) = \Pr(X_1 = n - k) = p_{X_1}(n - k) \quad (1.3.2.3)$$

From (1.3.2.2) and (1.3.2.3),

$$p_X(n) = \sum_k p_{X_1}(n - k) p_{X_2}(k) \triangleq p_{X_1}(n) * p_{X_2}(n) \quad (1.3.2.4)$$

where  $*$  denotes the convolution operation.

1.3.3. Substituting from (1.2.1.1) in (1.3.2.4),

$$p_X(n) = \frac{1}{6} \sum_{k=1}^6 p_{X_1}(n-k) = \frac{1}{6} \sum_{k=n-6}^{n-1} p_{X_1}(k) \quad (1.3.3.1)$$

$$\because p_{X_1}(k) = 0, \quad k \leq 1, k \geq 6. \quad (1.3.3.2)$$

From (1.3.3.1),

$$p_X(n) = \begin{cases} 0 & n < 1 \\ \frac{1}{6} \sum_{k=1}^{n-1} p_{X_1}(k) & 1 \leq n-1 \leq 6 \\ \frac{1}{6} \sum_{k=n-6}^6 p_{X_1}(k) & 1 < n-6 \leq 6 \\ 0 & n > 12 \end{cases} \quad (1.3.3.3)$$

## 1.4. The Triangular function

1.4.1. Substituting from (1.2.1.1) in (1.3.3.3),

$$p_X(n) = \begin{cases} 0 & n < 1 \\ \frac{n-1}{36} & 2 \leq n \leq 7 \\ \frac{13-n}{36} & 7 < n \leq 12 \\ 0 & n > 12 \end{cases} \quad (1.4.1.1)$$

satisfying (1.3.1.3).

## 1.5. The $Z$ -transform

1.5.1. The  $Z$ -transform of  $p_X(n)$  is defined as

$$P_X(z) = \sum_{n=-\infty}^{\infty} p_X(n)z^{-n}, \quad z \in \mathbb{C} \quad (1.5.1.1)$$

From (1.2.1.1) and (1.5.1.1),

$$P_{X_1}(z) = P_{X_2}(z) = \frac{1}{6} \sum_{n=1}^6 z^{-n} \quad (1.5.1.2)$$

$$= \frac{z^{-1}(1 - z^{-6})}{6(1 - z^{-1})}, \quad |z| > 1 \quad (1.5.1.3)$$

upon summing up the geometric progression.

1.5.2. Show that

$$p_X(n) = p_{X_1}(n) * p_{X_2}(n) \implies P_X(z) = P_{X_1}(z)P_{X_2}(z) \quad (1.5.2.1)$$

The above property follows from Fourier analysis and is fundamental to signal processing.

1.5.3. The unit step function is defined as

$$u(n) = \begin{cases} 1 & n \geq 0 \\ 0 & n < 0 \end{cases} \quad (1.5.3.1)$$

Show that the  $Z$  transform of  $u(n)$  is

$$U(z) = \frac{1}{1 - z^{-1}}, \quad |z| > 1 \quad (1.5.3.2)$$

1.5.4. Show that

$$nu(n) \xleftrightarrow{\mathcal{Z}} \frac{z^{-1}}{(1 - z^{-1})^2}, |z| > 1 \quad (1.5.4.1)$$

1.5.5. Show that

$$p_X(n - k) \xleftrightarrow{\mathcal{Z}} P_X(z) z^{-k} \quad (1.5.5.1)$$

1.5.6. From (1.5.1.3) and (1.5.2.1),

$$P_X(z) = \left\{ \frac{z^{-1} (1 - z^{-6})}{6 (1 - z^{-1})} \right\}^2 = \frac{1}{36} \frac{z^{-2} (1 - 2z^{-6} + z^{-12})}{(1 - z^{-1})^2} \quad (1.5.6.1)$$

From (1.5.5.1) and (1.5.4.1), it can be shown that

$$\begin{aligned} \frac{1}{36} [(n - 1) u(n - 1) - 2 (n - 7) u(n - 7) + (n - 13) u(n - 13)] \\ \xleftrightarrow{\mathcal{Z}} \frac{1}{36} \frac{z^{-2} (1 - 2z^{-6} + z^{-12})}{(1 - z^{-1})^2} \end{aligned} \quad (1.5.6.2)$$

1.5.7. From (1.5.1.1), (1.5.6.1) and (1.5.6.2)

$$p_X(n) = \frac{1}{36} [(n - 1) u(n - 1) - 2 (n - 7) u(n - 7) + (n - 13) u(n - 13)] \quad (1.5.7.1)$$

which is the same as (1.4.1.1). Note that (1.4.1.1) can be obtained from (1.5.6.2) using contour integration as well.

1.5.8. The experiment of rolling the dice was simulated using Python for 10000 samples.

These were generated using Python libraries for uniform distribution. The frequencies for each outcome were then used to compute the resulting pmf, which is plotted in

Figure 1.5.8.1. The theoretical pmf obtained in (1.4.1.1) is plotted for comparison.



Figure 1.5.8.1: Plot of  $p_X(n)$ . Simulations are close to the analysis.

1.5.9. The python code is available in

/codes/sum/dice.py

## Chapter 2

# Pingala Series

## 2.1. JEE 2019

$$a_n = \frac{\alpha^n - \beta^n}{\alpha - \beta}, \quad n \geq 1 \quad (9.1)$$

$$b_n = a_{n-1} + a_{n+1}, \quad n \geq 2, \quad b_1 = 1 \quad (9.2)$$

where  $\alpha$  and  $\beta$  ( $\alpha > \beta$ ) are the roots of the

$$z^2 - z - 1 = 0 \quad (9.3)$$

Verify the following using a python code.

2.1.1

$$\sum_{k=1}^n a_k = a_{n+2} - 1, \quad n \geq 1 \quad (2.1.1.1)$$

2.1.2

$$\sum_{k=1}^{\infty} \frac{a_k}{10^k} = \frac{10}{89} \quad (2.1.2.1)$$

2.1.3

$$b_n = \alpha^n + \beta^n, \quad n \geq 1 \quad (2.1.3.1)$$

2.1.4

$$\sum_{k=1}^{\infty} \frac{b_k}{10^k} = \frac{8}{89} \quad (2.1.4.1)$$

**Solution:**

```
$ python3 pingala/codes/1.py
```

## 2.2. Pingala Series

2.2.1 The Pingala series is generated using the difference equation

$$x(n+2) = x(n+1) + x(n), \quad x(0) = x(1) = 1, n \geq 0 \quad (2.2.1.1)$$

Generate a stem plot for  $x(n)$ .

**Solution:** The following code generates Fig. 2.2.1.1.

```
$ python3 pingala/codes/2.1.py
```

2.2.2 The one sided  $Z$ -transform of  $x(n)$  is defined as

$$X^+(z) = \sum_{n=0}^{\infty} x(n)z^{-n}, \quad z \in \mathbb{C} \quad (2.2.2.1)$$

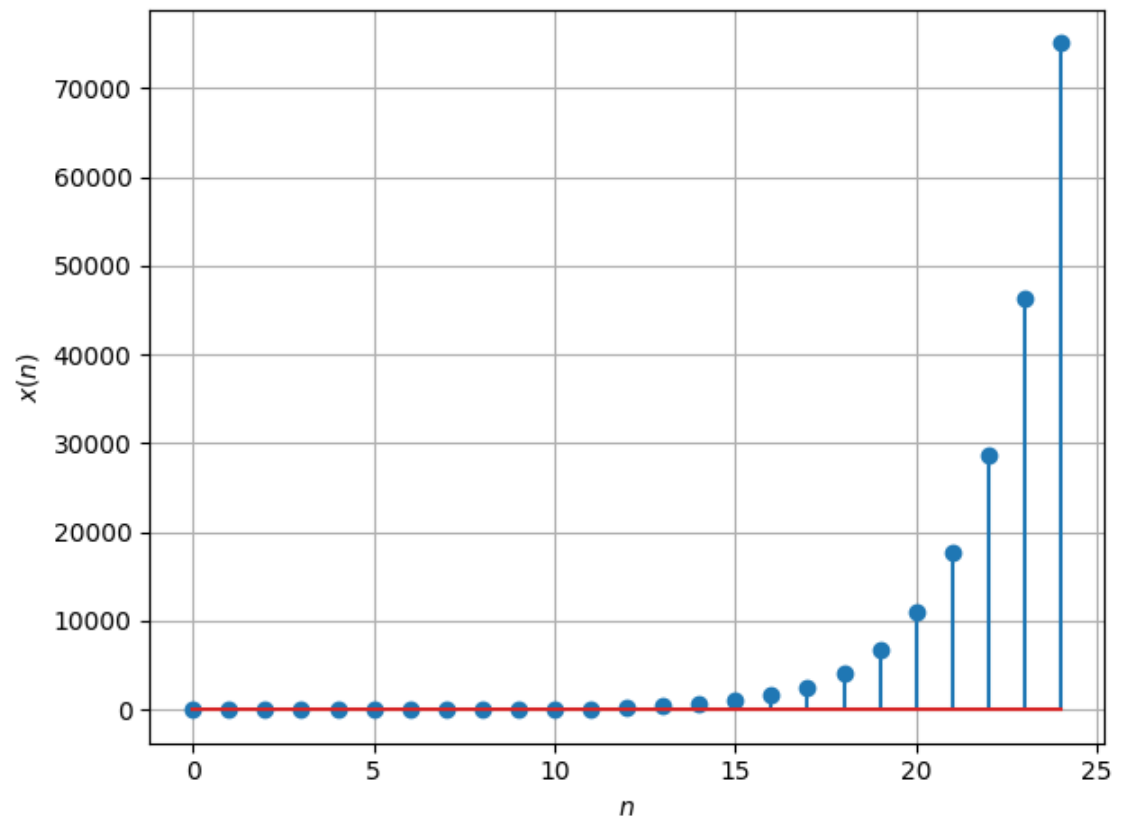


Figure 2.2.1.1: Plot of  $x(n)$



Find  $X^+(z)$ .

**Solution:** Taking the one-sided  $Z$ -transform on both sides of (2.2.1.1),

$$\mathcal{Z}^+[x(n+2)] = \mathcal{Z}^+[x(n+1)] + \mathcal{Z}^+[x(n)] \quad (2.2.2.2)$$

$$\implies z^2 X^+(z) - z^2 x(0) - zx(1) = zX^+(z) - zx(0) + zX^+(z) \quad (2.2.2.3)$$

$$\implies (z^2 - z - 1) X^+(z) = z^2 \quad (2.2.2.4)$$

$$\implies X^+(z) = \frac{1}{1 - z^{-1} - z^{-2}} = \frac{1}{(1 - \alpha z^{-1})(1 - \beta z^{-1})}, \quad |z| > \alpha \quad (2.2.2.5)$$

2.2.3 Find  $x(n)$ .

**Solution:** Expanding  $X^+(z)$  in (2.2.2.5) using partial fractions, we get

$$X^+(z) = \frac{1}{(\alpha - \beta)} \left[ \frac{z}{1 - \alpha z^{-1}} - \frac{z}{1 - \beta z^{-1}} \right] \quad (2.2.3.1)$$

$$\implies x(n) = \frac{\alpha^{n+1} - \beta^{n+1}}{\alpha - \beta} u(n) \quad (2.2.3.2)$$

$$= a_{n+1} \quad (2.2.3.3)$$

upon comparing with (9.1).

## 2.3. Linear Time Invariant System

2.3.1 Sketch

$$y(n) = x(n-1) + x(n+1), \quad n \geq 0 \quad (2.3.1.1)$$

**Solution:** Execute

```
$ python3 pingala/codes/2.2.py
```

to obtain Fig. 2.3.1.1



Figure 2.3.1.1: Plot of  $y(n)$

2.3.2 Show that

$$x(n+1) \xleftrightarrow{\mathcal{Z}} zX^+(z) - zx(0) \quad (2.3.2.1)$$

$$x(n-1) \xleftrightarrow{\mathcal{Z}} z^{-1}X^+(z) + zx(-1) \quad (2.3.2.2)$$

2.3.3 Find  $Y^+(z)$ .

**Solution:** Taking the one-sided  $Z$ -transform on both sides of (2.3.1.1),

$$\mathcal{Z}^+[y(n)] = \mathcal{Z}^+[x(n+1)] + \mathcal{Z}^+[x(n-1)] \quad (2.3.3.1)$$

$$Y^+(z) = zX^+(z) - zx(0) + z^{-1}X^+(z) + zx(-1) \quad (2.3.3.2)$$

$$= \frac{z + z^{-1}}{1 - z^{-1} - z^{-2}} - z = \frac{1 + 2z^{-1}}{1 - z^{-1} - z^{-2}}, \quad |z| > \alpha \quad (2.3.3.3)$$

2.3.4 Show that

$$y(n) = b_{n+1}. \quad (2.3.4.1)$$

2.3.5 Find the impulse response of (2.3.1.1)

## 2.4. Power of the $Z$ transform

2.4.1 Show that

$$\sum_{k=1}^{\infty} \frac{a_k}{10^k} = \frac{1}{10} \sum_{k=0}^{\infty} \frac{x(k)}{10^k} = \frac{1}{10} X^+(10) \quad (2.4.1.1)$$

**Solution:**

$$\sum_{k=1}^{\infty} \frac{a_k}{10^k} = \frac{1}{10} \sum_{k=0}^{\infty} \frac{a_{k+1}}{10^k} = \frac{1}{10} \sum_{k=0}^{\infty} \frac{x(k)}{10^k} \quad (2.4.1.2)$$

$$= \frac{1}{10} X^+(10) = \frac{1}{10} \times \frac{100}{89} = \frac{10}{89} \quad (2.4.1.3)$$

Thus, (2.1.2.1) is correct.

2.4.2 Show that

$$\sum_{k=1}^{\infty} \frac{b_k}{10^k} = \frac{1}{10} \sum_{k=0}^{\infty} \frac{y(k)}{10^k} = \frac{1}{10} Y^+(10) \quad (2.4.2.1)$$

**Solution:**

$$\sum_{k=1}^{\infty} \frac{b_k}{10^k} = \frac{1}{10} \sum_{k=0}^{\infty} \frac{b_{k+1}}{10^k} = \frac{1}{10} \sum_{k=0}^{\infty} \frac{y(k)}{10^k} \quad (2.4.2.2)$$

$$= \frac{1}{10} Y^+(z) = \frac{1}{10} \times \frac{120}{89} = \frac{12}{89} \quad (2.4.2.3)$$

Thus, (2.1.4.1) is incorrect.

2.4.3 Show that

$$\alpha^n + \beta^n, \quad n \geq 1 \quad (2.4.3.1)$$

can be expressed as

$$w(n) = (\alpha^{n+1} + \beta^{n+1}) u(n) \quad (2.4.3.2)$$

and find  $W(z)$ .

**Solution:** Putting  $n = k + 1$  in (2.4.3.1) and using the definition of  $u(n)$ ,

$$\alpha^n + \beta^n = (\alpha^{k+1} + \beta^{k+1}) u(k) \quad (2.4.3.3)$$

Hence, (2.4.3.1) can be expressed as

$$w(n) = (\alpha^{n+1} + \beta^{n+1}) u(n) \quad (2.4.3.4)$$

Therefore,

$$W(z) = Y(z) = \frac{1 + 2z^{-1}}{1 - z^{-1} - z^{-2}} \quad (2.4.3.5)$$

Thus, by invoking (2.3.4.1), we find that (2.1.3.1) is correct

## 2.5. Convolution

2.5.1 Show that

$$\sum_{k=1}^n a_k = \sum_{k=0}^{n-1} x(k) = x(n) * u(n-1) \quad (2.5.1.1)$$

**Solution:** From (2.2.3.3), and noting that  $x(n) = 0 \ \forall \ n < 0$ ,

$$\sum_{k=1}^n a_k = \sum_{k=0}^{n-1} x(k) = \sum_{k=-\infty}^{n-1} x(k) \quad (2.5.1.2)$$

$$= \sum_{k=-\infty}^{\infty} x(k)u(n-1-k) = x(n) * u(n-1) \quad (2.5.1.3)$$

2.5.2 Show that

$$a_{n+2} - 1, \quad n \geq 1 \quad (2.5.2.1)$$

can be expressed as

$$[x(n+1) - 1]u(n-1) \quad (2.5.2.2)$$

**Solution:** From (2.2.3.3),

$$a_{n+2} - 1 = [x(n+1) - 1], \quad n \geq 1 \quad (2.5.2.3)$$

and so, using the definition of  $u(n)$ ,

$$a_{n+2} - 1 = [x(n+1) - 1] u(n-1) \quad (2.5.2.4)$$

2.5.3 Show that

$$[x(n+1) - 1] u(n-1) \xleftrightarrow{\mathcal{Z}} \frac{z^{-1}}{(1 - z^{-1} - z^{-2})(1 - z^{-1})} \quad (2.5.3.1)$$

**Solution:** The Z transform of the above signal can be expressed as

$$\sum_{n=1}^{\infty} x(n+1)z^{-n} - \frac{z^{-1}}{1 - z^{-1}} = \sum_{n=2}^{\infty} x(n)z^{-n+1} - \frac{z^{-1}}{1 - z^{-1}} \quad (2.5.3.2)$$

$$= z [X^+(z) - x(0) - x(1)z^{-1}] - \frac{z^{-1}}{1 - z^{-1}} \quad (2.5.3.3)$$

$$= \frac{z}{1 - z^{-1} - z^{-2}} - z - 1 - \frac{z^{-1}}{1 - z^{-1}} \quad (2.5.3.4)$$

$$= \frac{z}{1 - z^{-1} - z^{-2}} - \frac{z}{1 - z^{-1}} \quad (2.5.3.5)$$

$$= \frac{z^{-1}}{(1 - z^{-1} - z^{-2})(1 - z^{-1})} \quad (2.5.3.6)$$

From (2.2.3.3), we get

$$\sum_{k=1}^n a_k = a_{n+2} - 1 \quad (2.5.3.7)$$



## Chapter 3

# Circuits and Transforms

### 3.1. Definitions

1. The unit step function is defined as

$$u(t) = \begin{cases} 1 & t > 0 \\ \frac{1}{2} & t = 0 \\ 0 & t < 0 \end{cases} \quad (3.1.1.1)$$

2. The Laplace transform of  $g(t)$  is defined as

$$G(s) = \int_{-\infty}^{\infty} g(t)e^{-st} dt \quad (3.1.2.1)$$

### 3.2. Laplace Transform

In the circuit in Fig. 2.1, the switch  $S$  is connected to position  $P$  for a long time so that the charge on the capacitor becomes  $q_1 \mu C$ . Then  $S$  is switched to position  $Q$ . After a long time, the charge on the capacitor is  $q_2 \mu C$ . Use variables  $R_1, R_2, C_0$  for the passive elements.



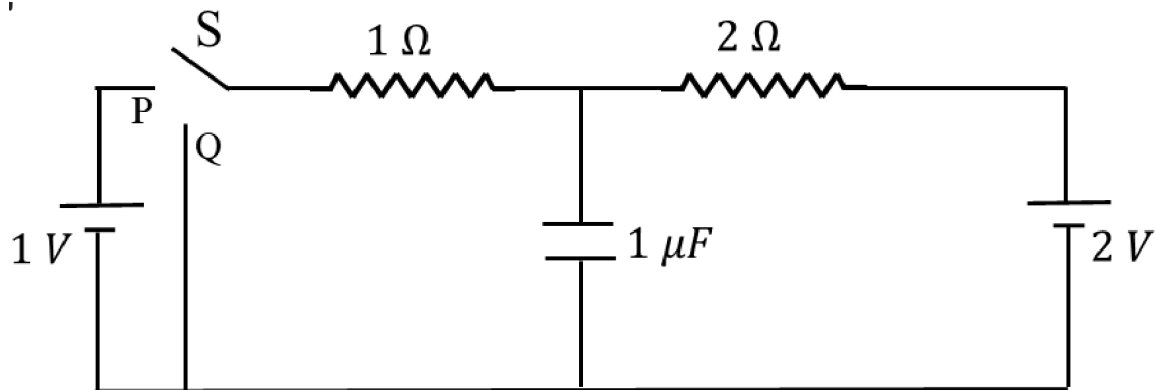


Figure 2.1:

1. Draw the circuit in position  $P$ .

**Solution:** See Fig. 3.2.1.1 drawn using the code in

cktsig/figs/ckt-q1.tex

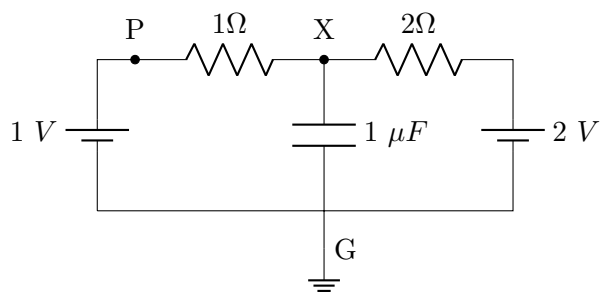


Figure 3.2.1.1:

2. Simulate the circuit in Fig. 3.2.1.1 using ngspice.

**Solution:** The following ngspice script simulates the given circuit.

ngspice codes/2\_7.cir

3. Find  $q_1$  in Fig. 3.2.3.1.

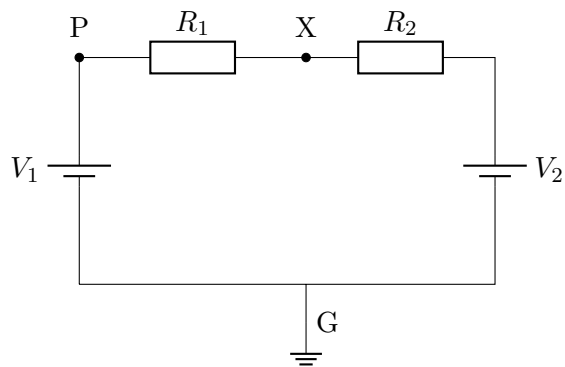


Figure 3.2.3.1:

**Solution:** Assuming the circuit to be grounded at G and the relative potential at point X to be  $V$ , we use KCL at X and get

$$\frac{V - V_1}{R_1} + \frac{V - V_2}{R_2} = 0 \quad (3.2.3.1)$$

$$(3.2.3.2)$$

$$\Rightarrow V = \frac{V_1 R_2 + V_2 R_1}{R_1 + R_2} = \frac{4}{3} \text{ V} \quad (3.2.3.3)$$

upon substituting numerical values. Hence,

$$q_1 = CV = C \frac{(V_1 R_2 + V_2 R_1)}{R_1 + R_2} = \frac{4}{3} \mu\text{C} \quad (3.2.3.4)$$

4. Show that the Laplace transform of  $u(t)$  is  $\frac{1}{s}$  and find the ROC.

**Solution:** We have,

$$u(t) \xleftrightarrow{\mathcal{L}} \int_0^\infty u(t)e^{-st}dt \quad (3.2.4.1)$$

$$= \int_0^0 \frac{1}{2}e^{-st}dt + \int_0^\infty e^{-st}dt \quad (3.2.4.2)$$

$$= \frac{1}{s}, \quad \Re(s) > 0 \quad (3.2.4.3)$$

5. Now consider the following resistive circuit in Fig. 3.2.5.1 transformed from Fig. 2.1 where

$$v_1(t) = u(t) \xleftrightarrow{\mathcal{L}} V_1(s) = \frac{1}{s} \quad (3.2.5.1)$$

$$v_2(t) = 2u(t) \xleftrightarrow{\mathcal{L}} V_2(s) = \frac{2}{s} \quad (3.2.5.2)$$

Find the voltage across the capacitor  $V_{C_0}(s)$ .



Figure 3.2.5.1:

**Solution:** Applying KCL at  $X$ ,

$$\frac{V(s) - \frac{1}{s}}{R_1} + \frac{V(s) - \frac{2}{s}}{R_2} + sC_0V(s) = 0 \quad (3.2.5.3)$$

$$\implies V(s) \left( \frac{1}{R_1} + \frac{1}{R_2} + sC_0 \right) = \frac{1}{s} \left( \frac{1}{R_1} + \frac{2}{R_2} \right) \quad (3.2.5.4)$$

$$\implies V(s) = \frac{\frac{1}{R_1} + \frac{2}{R_2}}{s \left( \frac{1}{R_1} + \frac{1}{R_2} + sC_0 \right)} \quad (3.2.5.5)$$

$$= \frac{\frac{1}{R_1} + \frac{2}{R_2}}{\frac{1}{R_1} + \frac{1}{R_2}} \left( \frac{1}{s} - \frac{1}{\frac{1}{C_0} \left( \frac{1}{R_1} + \frac{1}{R_2} \right) + s} \right) \quad (3.2.5.6)$$

6. Show that

$$e^{-at}u(t) \xleftrightarrow{\mathcal{L}} \frac{1}{s+a}, \quad a > 0 \quad (3.2.6.1)$$

and find the ROC.

**Solution:** Note that by substituting  $s := s + a$  in (3.2.4.3), and considering  $a \in \mathbb{R}$ ,

$$e^{-at}u(t) \xleftrightarrow{\mathcal{L}} \int_0^\infty u(t)e^{-(s+a)t}dt \quad (3.2.6.2)$$

$$= \frac{1}{s+a}, \quad \Re(s) > -a \quad (3.2.6.3)$$

7. Find  $v_{C_0}(t)$ .

**Solution:** Taking the inverse Laplace transform in (3.2.5.6),

$$V(s) \xleftrightarrow{\mathcal{L}^{-1}} \frac{2R_1 + R_2}{R_1 + R_2} u(t) \left( 1 - e^{-\left(\frac{1}{R_1} + \frac{1}{R_2}\right) \frac{t}{C_0}} \right) \quad (3.2.7.1)$$

from (3.2.6.1).

8. Verify your result numerically.

**Solution:** The following python code plots the graph in Fig. 3.2.8.1

```
python3 codes/2.6-sim.py
```

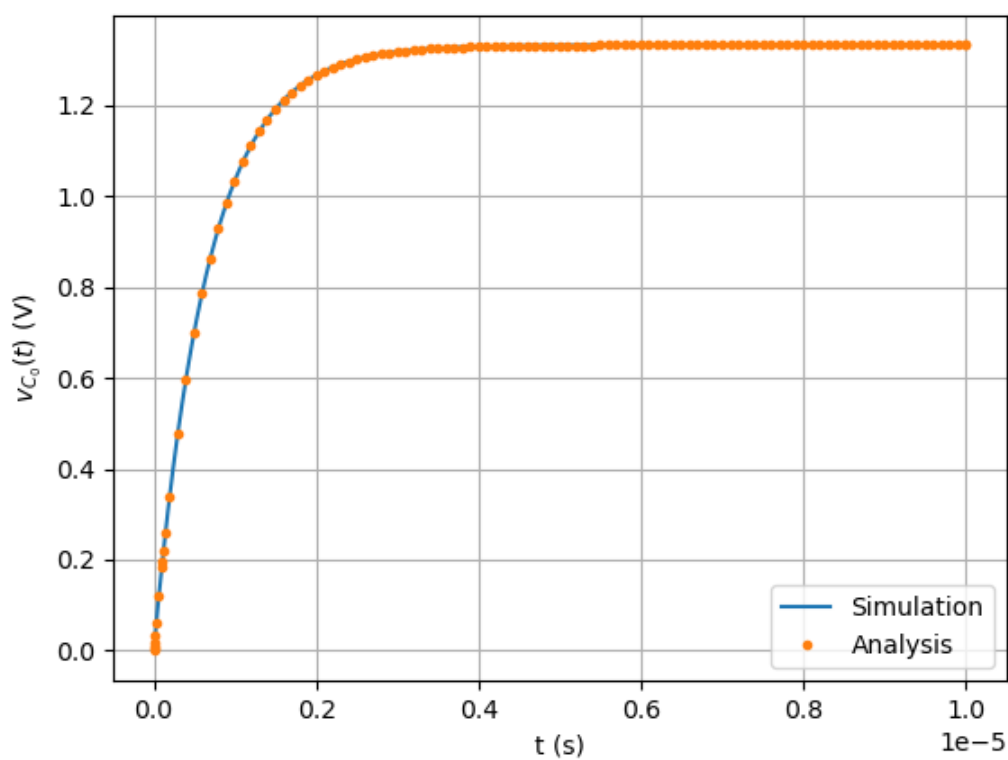


Figure 3.2.8.1:  $v_{C_0}(t)$  simulation

### 3.3. Initial Conditions

1. Draw the equivalent circuit when the switch is at Q.

**Solution:** See Fig. 3.3.1.1. below.



Figure 3.3.1.1:

2. Find  $q_2$  in Fig. 3.3.2.1.

**Solution:** The equivalent circuit at steady state when the switch is at Q is shown in Fig. 3.3.2.1. Since the capacitor behaves as an open circuit, we use voltage division at



Figure 3.3.2.1:

$X$  to obtain

$$V = V_1 \frac{R_1}{R_1 + R_2} = \frac{2}{3} \text{ V} \quad (3.3.2.1)$$

upon substituting numerical values. Hence,

$$q_2 = CV = \frac{2}{3} \mu\text{C}. \quad (3.3.2.2)$$

3. Draw the equivalent  $s$ -domain resistive circuit when  $S$  is switched to position Q.

See Fig. 3.3.3.1.



Figure 3.3.3.1:

4.  $V_{C_0}(s) = ?$

**Solution:** Using KCL at node  $X$  in Fig. 3.3.3.1,

$$\frac{V(s) - 0}{R_1} + \frac{V(s) - \frac{2}{s}}{R_2} + sC_0 \left( V(s) - \frac{4}{3s} \right) = 0 \quad (3.3.4.1)$$

$$\Rightarrow V_{C_0}(s) = \frac{\frac{2}{sR_2} + \frac{4C_0}{3}}{\frac{1}{R_1} + \frac{2}{R_2} + sC_0} \quad (3.3.4.2)$$

5.  $v_{C_0}(t) = ?$

**Solution:** From (3.3.4.2), using partial fractions,

$$V_{C_0}(s) = \frac{4}{3} \left( \frac{1}{\frac{1}{C_0} \left( \frac{1}{R_1} + \frac{1}{R_2} \right) + s} \right) + \frac{2}{R_2 \left( \frac{1}{R_1} + \frac{1}{R_2} \right)} \left( \frac{1}{s} - \frac{1}{\frac{1}{C_0} \left( \frac{1}{R_1} + \frac{1}{R_2} \right) + s} \right) \quad (3.3.5.1)$$

Taking the inverse Laplace Transform,

$$v_{C_0}(t) = \frac{4}{3} e^{-\left(\frac{1}{R_1} + \frac{1}{R_2}\right) \frac{t}{C_0}} u(t) + \frac{2}{R_2 \left(\frac{1}{R_1} + \frac{1}{R_2}\right)} \left(1 - e^{-\left(\frac{1}{R_1} + \frac{1}{R_2}\right) \frac{t}{C_0}}\right) u(t) \quad (3.3.5.2)$$

from (3.2.6.1).

6. Verify your result numerically.

**Solution:** Substituting numerical values,

$$v_{C_0}(t) = \frac{2}{3} \left(1 + e^{-(1.5 \times 10^6)t}\right) u(t) \quad (3.3.6.1)$$

This is verified by the following code that plots Fig. 3.3.6.1 using ngspice and python.

```
python3 codes/3_4-sim.py
```

7. Find  $v_{C_0}(0-)$ ,  $v_{C_0}(0+)$  and  $v_{C_0}(\infty)$ .

**Solution:** From the initial conditions,

$$v_{C_0}(0-) = \frac{q_1}{C_0} = \frac{4}{3} \text{ V} \quad (3.3.7.1)$$

Using (3.3.6.1),

$$v_{C_0}(0+) = \lim_{t \rightarrow 0+} v_{C_0}(t) = \frac{4}{3} \text{ V} \quad (3.3.7.2)$$

$$v_{C_0}(\infty) = \lim_{t \rightarrow \infty} v_{C_0}(t) = \frac{2}{3} \text{ V} \quad (3.3.7.3)$$

## 3.4. Bilinear Transform

1. Formulate the differential equation for Fig. 3.4.1.1.



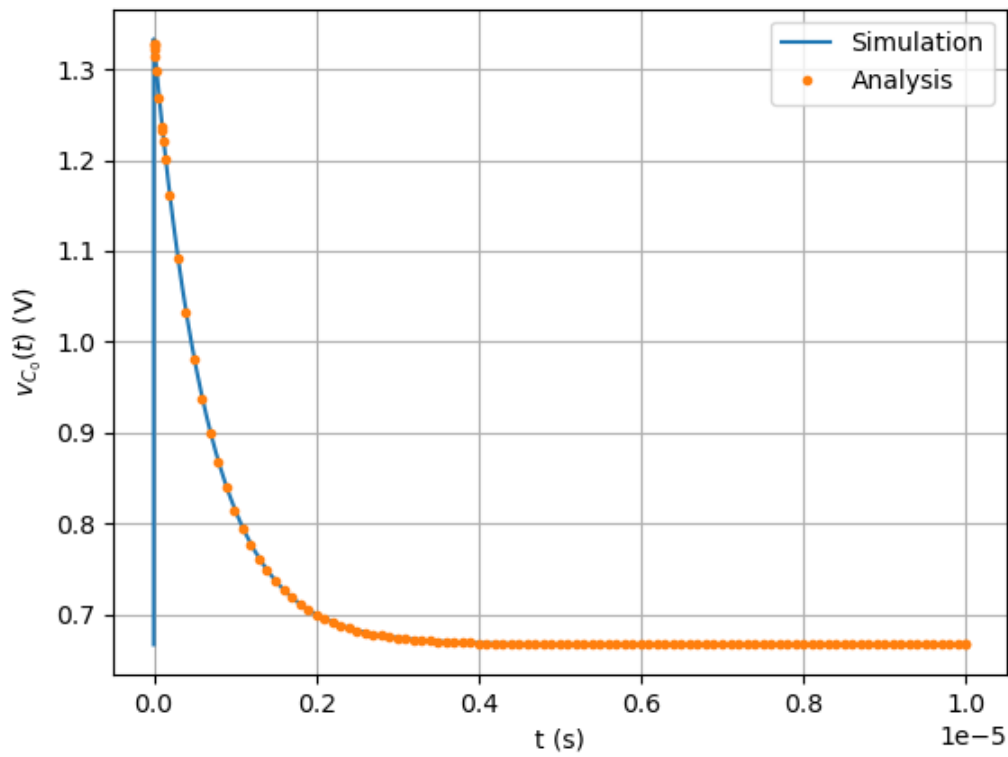


Figure 3.3.6.1:  $v_{C_0}(t)$  after the switch is flipped



Figure 3.4.1.1:

**Solution:** Applying KVL on the loop,

$$V - iR_2 - \frac{1}{C_0} \int_0^t i dt = 0 \quad (3.4.1.1)$$

where  $i(0) = 0$ ,  $V_C(0) = 0$ . Denote by  $V_C$  the voltage at the capacitor. Then,

$$i = C \frac{dV_C}{dt} \quad (3.4.1.2)$$

and therefore using (3.4.1.2) in (3.4.1.1), we get the differential equation

$$V - \tau \frac{dV_C}{dt} - V_C = 0 \quad (3.4.1.3)$$

where  $\tau \triangleq R_2 C_0$  is the time constant of the circuit.

2. Fig. 3.4.1.1 is transformed to Fig. 3.4.2.1 in the  $s$  domain. Find the system transfer function

$$H(s) = \frac{V(s)}{V_c(s)} \quad (3.4.2.1)$$

**Solution:** Using the voltage division formula, the voltage across the capacitor is

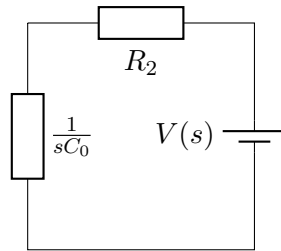


Figure 3.4.2.1:

given by

$$V_{C_0}(s) = V(s) \frac{\frac{1}{sC_0}}{\frac{1}{sC_0} + R_2} = V(s) \frac{1}{1 + sC_0 R_2} \quad (3.4.2.2)$$

$$\Rightarrow H(s) = \frac{V_{C_0}(s)}{V(s)} = \frac{1}{1 + sC_0 R_2} \quad (3.4.2.3)$$

3. Plot  $|H(j\omega)|$ . What kind of filter is it?

**Solution:**

4. Using trapezoidal rule for integration, formulate (3.4.1.3) as a difference equation by considering

$$V_C(n) = V_C(t)|_{t=n} \quad (3.4.4.1)$$

**Solution:** Integrating (3.4.1.3) between limits  $n-1$  to  $n$  and applying the trapezoidal formula,

$$\frac{V_C(n) + V_C(n-1)}{2} + \tau (V_C(n) - V_C(n-1)) = \frac{V(n) + V(n-1)}{2} \quad (3.4.4.2)$$

$$\Rightarrow V_C(n) \left( \frac{1}{2} + \tau \right) + V_C(n-1) \left( \frac{1}{2} - \tau \right) = \frac{V(n) + V(n-1)}{2} \quad (3.4.4.3)$$

5. Find

$$H(z) = \frac{V_C(z)}{V(z)}. \quad (3.4.5.1)$$

**Solution:** Applying the Z-transform on both sides of (3.4.4.3),

$$V_C(z) [(2\tau + 1) - z^{-1}(2\tau - 1)] = V(z) (1 + z^{-1}) \quad (3.4.5.2)$$

Hence,

$$H(z) = \frac{1 + z^{-1}}{(2\tau + 1) - (2\tau - 1)z^{-1}} \quad (3.4.5.3)$$

6. Is the system defined by (3.4.4.3) stable?

**Solution:** From (3.4.5.3),  $H(z)$  has a pole at

$$z = \frac{2\tau - 1}{2\tau + 1} \quad (3.4.6.1)$$

Since

$$\left| \frac{2\tau - 1}{2\tau + 1} \right| < 1, \quad (3.4.6.2)$$

the ROC of  $H(z)$  is  $|z| > 1$ , which implies that (3.4.4.3) represents a stable system.

7. How can you obtain  $H(z)$  from  $H(s)$ ?

**Solution:** Substituting

$$s = 2 \left( \frac{1 - z^{-1}}{1 + z^{-1}} \right) \quad (3.4.7.1)$$

in (3.4.2.3), we obtain (3.4.5.3). This is a special case of the the bilinear transformation for  $T = 1$  where

$$s = \frac{2}{T} \frac{1 - z^{-1}}{1 + z^{-1}} \quad (3.4.7.2)$$

8. Find  $V_C(n)$ . Verify using ngspice.

**Solution:** Since  $V(n) = Vu(n)$ ,

$$V(z) = \frac{V}{1 - z^{-1}} \quad (3.4.8.1)$$

Therefore,

$$V_C(z) = H(z)V(z) = \frac{V(1+z^{-1})}{(1-z^{-1})((2\tau+1)-(2\tau-1)z^{-1})} \quad (3.4.8.2)$$

$$(3.4.8.3)$$

Taking the inverse,

$$V_C(n) = \begin{cases} \frac{V}{2} [u(n)(1-p^n) + u(n-1)(1-p^{n-1})] & n > 0 \\ 0 & n \leq 0 \end{cases} \quad (3.4.8.4)$$

where

$$p = \frac{2\tau-1}{2\tau+1} \quad (3.4.8.5)$$

The following python code

```
cktsig/codes/1.7.py
```

plots  $V_C$  for the sampling interval  $T = 10^{-7}$  in Fig. 3.4.8.1.



Figure 3.4.8.1: Representation of output across  $C$ .



## Chapter 4

# Frequency Modulation

### 4.1. Message

Play the message signal using

```
sudo apt install ffmpeg  
ffplay fm/input-audio/Sound.wav
```

1. Find the sampling rate of the message.

**Solution:** Executing

```
python3 fm/msg/codes/sample_rate.py
```

gives the sampling rate of the input signal as 44100Hz.

2. Plot the spectrum of the message signal using the builtin FFT algorithm.

**Solution:** The following code plots the spectrum in Fig. 4.1.2.1 using the builtin FFT algorithm of the python library 'Numpy.'

Executing

```
python3 fm/msg/codes/msg_spec.py
```



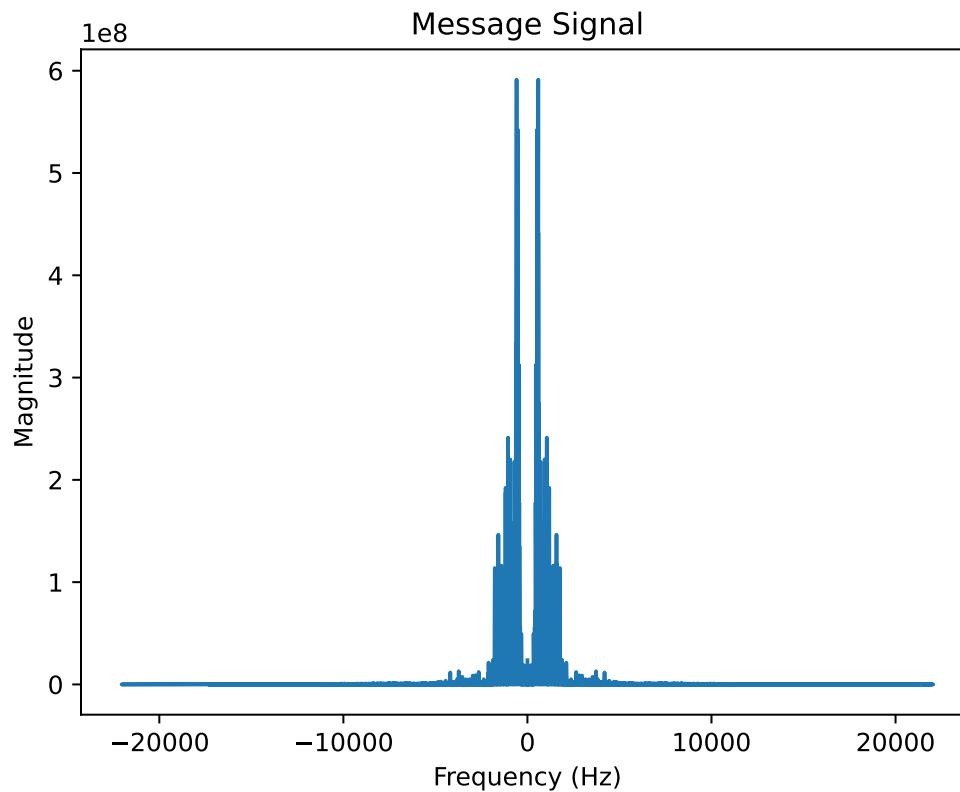


Figure 4.1.2.1: Plot of spectrum of message signal using builtin FFT algorithm.

3. Find the number of samples used to compute the FFT.

**Solution:** The following code finds the number of samples used to compute the FFT

```
python3 fm/msg/codes/no.of.samples.py
```

and gives number of samples as 1226536

4. What does the following command do?

```
f_i = np.fft.fftfreq(len(audio_data), d=1/sample_rate)
```

Parameter	Description	Value
f	Frequency	Varies with k
k	index of the DFT output component	$0 \leq k \leq n - 1$
n	Length of the input signal	1226536
d	Sample space	1/44100

Table 4.2: Parameters of Message signal

**Solution:** The `np.fft.fftfreq` function calculates the frequencies corresponding to the discrete Fourier transform (DFT) output by using the formula:

$$f = \frac{k}{nd} \quad (4.1.4.1)$$

for  $k=100$ ,

$$f = \frac{100}{1226536(1/44100)}$$

$$f = 3.595Hz$$

The `np.fft.fftfreq` function generates an array of length  $n$ , where each element represents the frequency corresponding to the DFT output component at the respective index. The index  $k$  ranges from 0 to  $n-1$ , and each index corresponds to a specific frequency component.

5. Plot the spectrum of the message signal by writing your own FFT algorithm.

**Solution:**

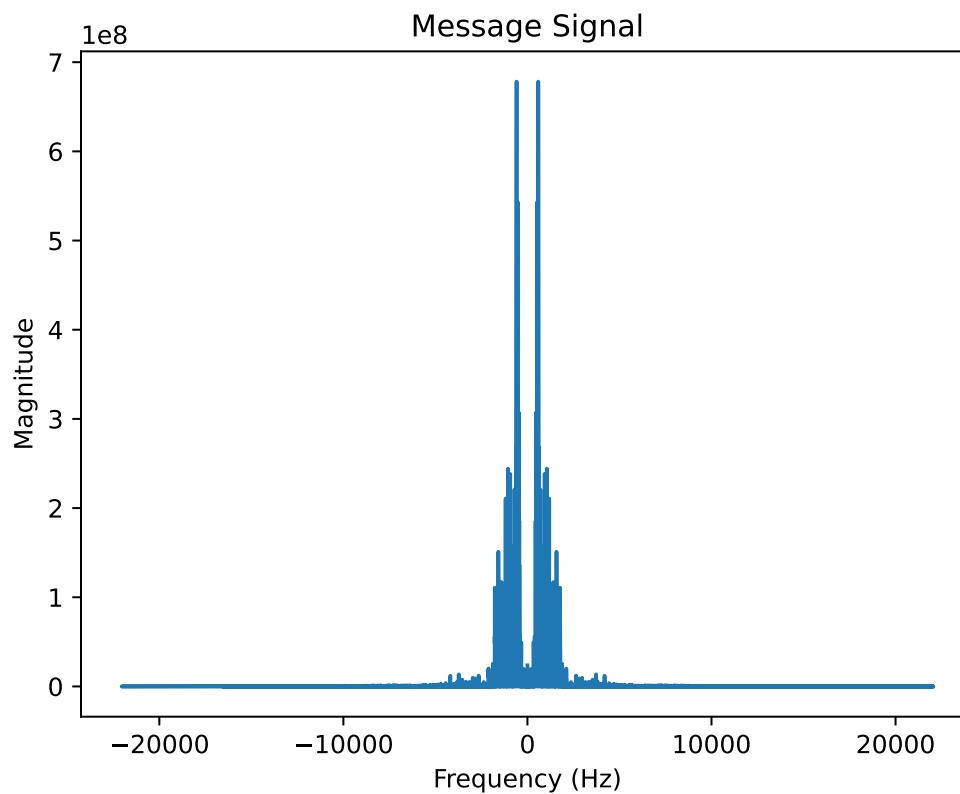


Figure 4.1.5.1: Plot of spectrum of message signal using own FFT algorithm.

The following code plots the spectrum in Fig. 4.1.5.1 using the DFT defined in (A.1.1.1).

```
python3 fm/msg/codes/FFTalgorithm.py
```

6. Compute and plot the PSD of the message signal using (A.3.1.1).

**Solution:** Executing

```
python3 fm/msg/codes/msg_psd.py
```

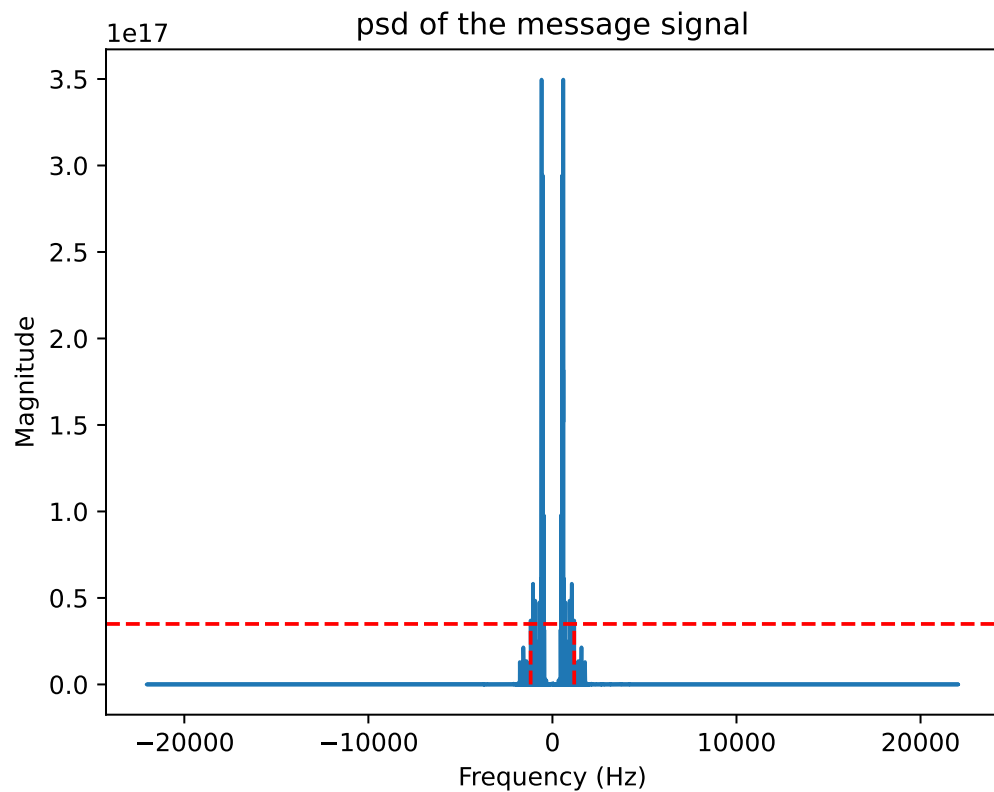


Figure 4.1.6.1: Plot of PSD of the message signal.

7. Find the bandwidth of the message signal. Through a plot, explain the principle used for computing the bandwidth.

**Solution:** Executing

```
python3 fm/msg/codes/msg_bw.py
```

gives the bandwidth of the message signal as 2366.62454 Hz

The threshold of the message signal from the plot is  $3.4965 \times 10^{16}$

The maximum and minimum frequencies are 1183.31227 Hz and -1183.31227 respectively.

Therefore bandwidth is

$$f_{max} - f_{min} = 1183.312 - (-1183.312) = 2366.624 Hz \quad (4.1.7.1)$$

## 4.2. Transmitter

1. The modulated signal is given by

$$s(t) = \cos(2\pi f_c t + \phi(t)) \quad (4.2.1.1)$$

where

$$\phi(t) = 2\pi k_f \int_0^t m(\tau) d\tau \quad (4.2.1.2)$$

List the various parameters in a table.

**Solution:** Parameters are in the table

Parameter	Description	Value
$F_s$	Sampling rate	44100 Hz
$K_f$	Frequency sensitivity	20 Hz/volt
$A_c$	Amplitude of the carrier signal	1
t	Sampling time	22 $\mu$
T	Threshold	0.1
$F_c$	Frequency of the carrier signal	102 MHz

Table 4.4: Parameters of signal

2. Obtain a difference equation for computing  $\phi(t)$ . Suggest a sampling rate.

**Solution:** To obtain the difference equation for computing  $\phi(t)$ , we need to discretize the integral in the given equation. We can use the rectangular rule for numerical integration.

Let us divide the interval  $[t_0, t]$  into  $N$  equal subintervals of width  $\Delta t = (t - t_0)/N$ . Then, we can approximate  $m(\tau)$  by its value at the midpoint of each subinterval,  $\tau_n = t_0 + (n + 1/2)\Delta t$ , where  $n = 0, 1, 2, \dots, N - 1$ . This gives:

$$m(\tau_n) \approx m(n\Delta t)$$

we can approximate the integral in the given equation ??

$$\phi(t) \approx 2\pi k f_c \Delta t \sum_{n=0}^{N-1} m(n\Delta t)$$

$\phi$  at the  $n$ th time step  $t_n = t_0 + n\Delta t$ . Then, we can write:

$$\phi_n = 2\pi k f_c \Delta t \sum_{k=0}^{n-1} m(k\Delta t) \quad (4.2.2.1)$$

$$\phi_{n-1} = 2\pi k f_c \Delta t \sum_{k=0}^{n-2} m(k\Delta t) \quad (4.2.2.2)$$

Subtracting the ?? from ??, we get:

$$\phi_n - \phi_{n-1} = 2\pi k f_c \Delta t, m((n - 1)\Delta t) \quad (4.2.2.3)$$

3. Plot the spectrum of the transmitted signal.

**Solution:** The following code plots the spectrum of transmitted signal in Fig. 4.2.3.1

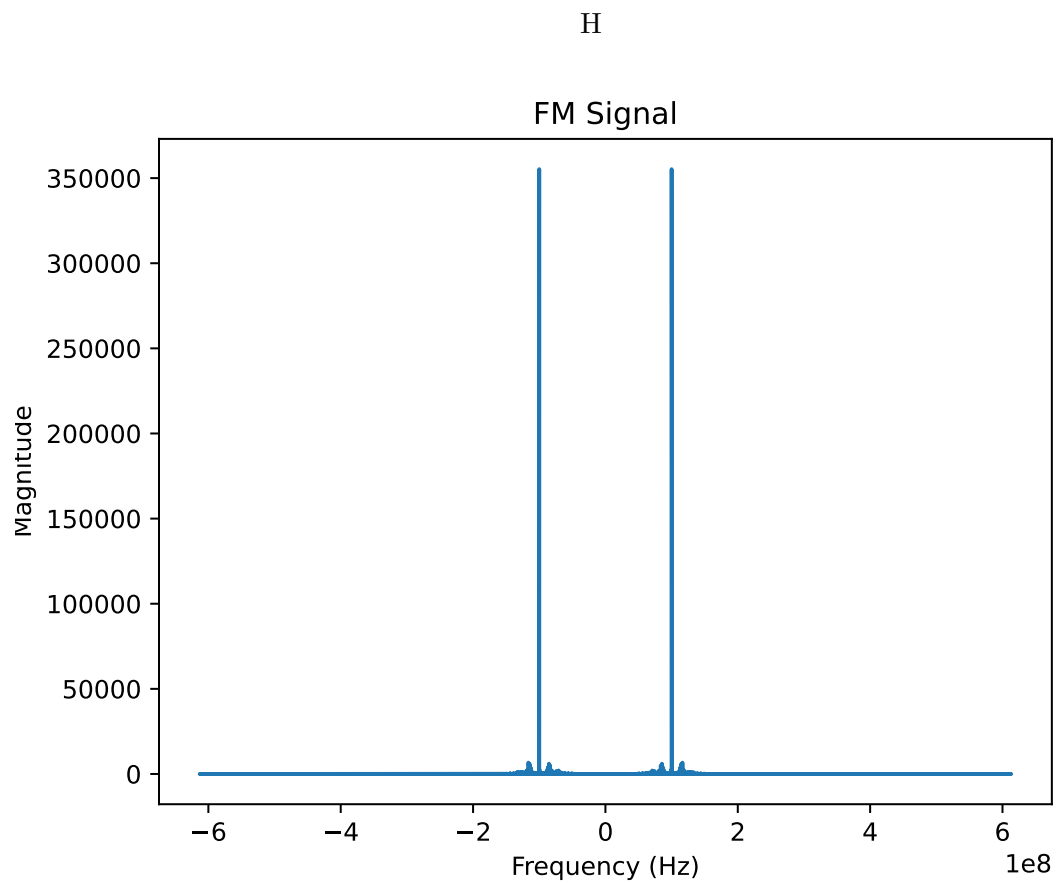


Figure 4.2.3.1: Plot of spectrum of transmitted signal.

Executing

```
python3 fm/tx/codes/tx_spec.py
```

4. Compute and plot the PSD of the message signal using (A.3.1.1).

**Solution:** Executing

```
python3 fm/tx/codes/tx_psd.py
```

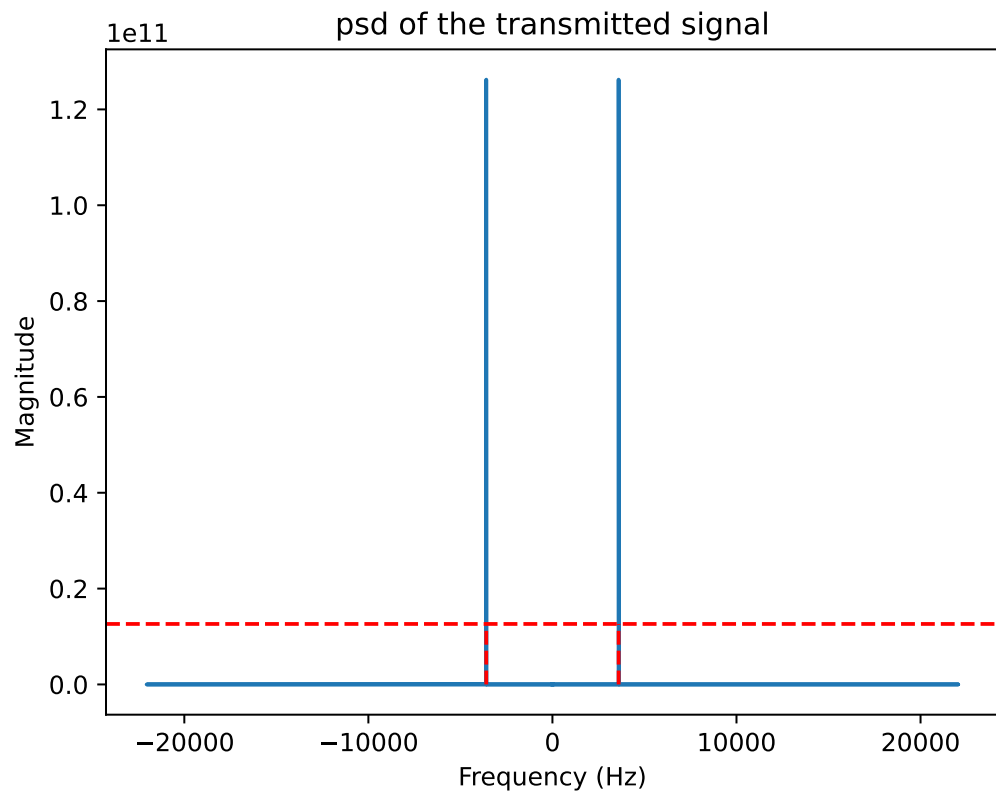


Figure 4.2.4.1: Plot of PSD of the message signal.

5. Compute the bandwidth of the transmitted signal.

**Solution:** Executing

```
python3 fm/tx/codes/tx_bw.py
```

gives the bandwidth of the transmitted signal as 7190.69574 Hz

The threshold of the transmitted signal from the plot is  $1.2621 \times 10^{10}$

The maximum and minimum frequencies are 3595.34787 Hz and -3595.34787 respectively.

Therefore bandwidth is



$$f_{max} - f_{min} = 3595.34787 - (-3595.34787) = 7190.69574Hz \quad (4.2.5.1)$$

## 4.3. Receiver

1. List the various components used to implement receiver.

**Solution:**

Components are listed in the table Table ??

Component	Type
RTL-SDR chip	Hardware
GNU radio	Software
Antenna	Hardware

Table 4.6: Components Required

The picture of RTL-SDR and Antenna is given in Fig. ?. This set is used to receive the Fm signals.



Figure 4.3.1.1: RTL-SDR

2. Install and open GNU Radio using the following commands

```
sudo apt update  
sudo apt install gnuradio  
gnuradio-companion
```

3. How to construct the block diagram in GNU radio?

**Solution:**

**Step-1:**

Search for Low pass filter block and add drag it to the work space.

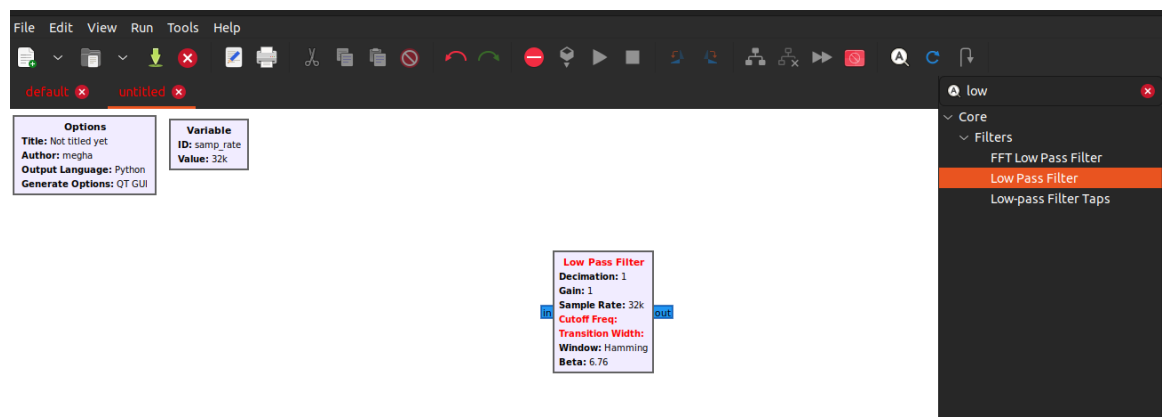


Figure 4.3.3.1: Adding blocks

Similarly do for RTL-Source block, WBFM and Audio sink blocks. **Step-2:** Connect them according to the flowgraph shown in Fig. ??.

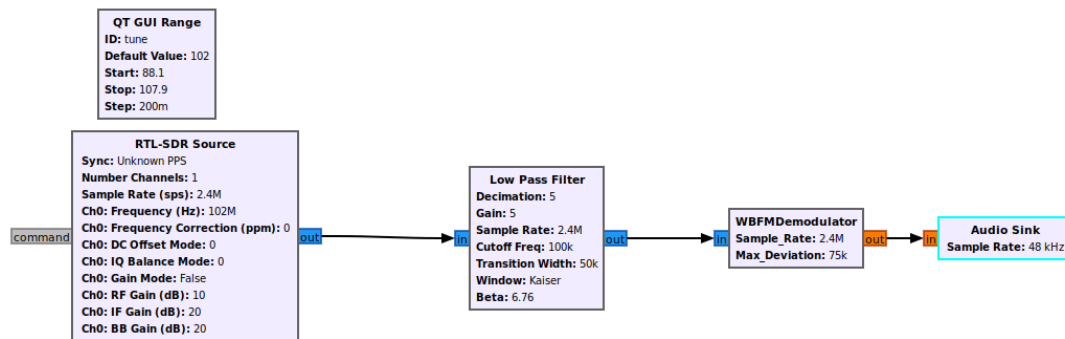


Figure 4.3.3.2: Block diagram of GNU Radio

**Note:** Refer the following website for any queries.

[https://wiki.gnuradio.org/index.php?title=Creating\\_Your\\_First\\_Block](https://wiki.gnuradio.org/index.php?title=Creating_Your_First_Block)

4. Explain each block in block diagram

**Solution:**

#### 1. RTL-SDR Source:

The RTL-SDR Source block is used to stream samples from RTL-SDR device.

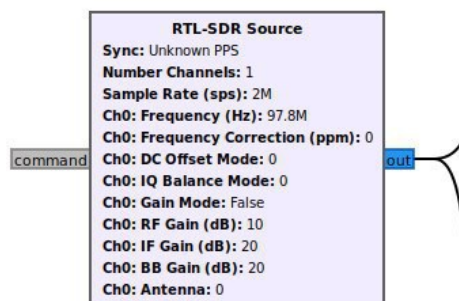


Figure 4.3.4.1: RTL-SDR Source Block

## 2. Low Pass Filter:

Low Pass Filter removes frequencies that are higher than the cutoff.



Figure 4.3.4.2: Low Pass Filter Block

## 3. WBFM:

WBFM Receive is a block for demodulating a broadcast FM signal. The output is the demodulated audio.



Figure 4.3.4.3: WBFM Block

## 4. Audio sink:

Audio Sink block provides audio output signal at the computer speakers.

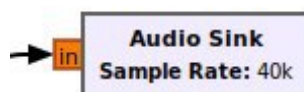


Figure 4.3.4.4: Audio Sink Block

Connect the RTL-SDR to the system and execute the flowgraph in Fig. ??.

We can listen to the sound which we have transmitted.

A default python code is generated and stored as

```
fm/rx/codes/fm_receiver.py
```

5. How to replace LPF default code with custom code?

**Solution:**

**Step-1:** Disable Low pass filter block and search for the python block and add it to the workspace.



Figure 4.3.5.1: search for blocks

**Step-2:** Double-click the block to edit the properties as in Fig. ??

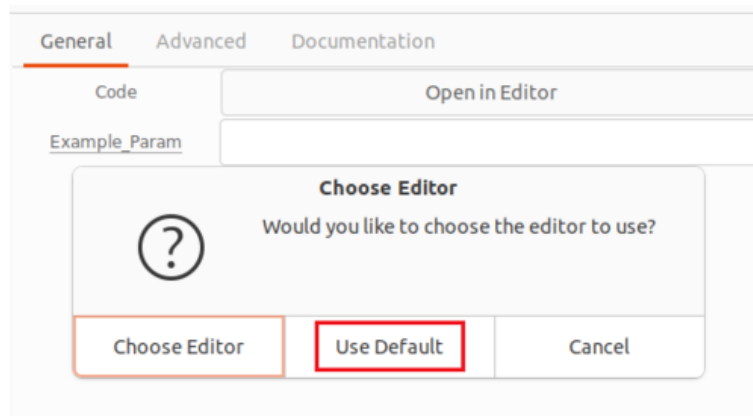


Figure 4.3.5.2: Edit properties

**Step-3:** Change the block name as required by editing in the code as shown in Fig. ??

```

1  """
2  Embedded Python Blocks:
3
4  Each time this file is saved, GRC will instantiate the first class it finds
5  to get ports and parameters of your block. The arguments to __init__ will
6  be the parameters. All of them are required to have default values!
7  """
8
9  import numpy as np
10 from gnuradio import gr
11
12 class blk(gr.sync_block): # other base classes are basic_block, decim_block, interp_block
13     """Embedded Python Block example - a simple multiply const"""
14
15     def __init__(self, additionFlag=True): # only default arguments here
16         """arguments to this function show up as parameters in GRC"""
17         gr.sync_block.__init__(
18             self,
19             name='Add or Multiply Block', # will show up in GRC
20             in_sig=[np.complex64,np.complex64],
21             out_sig=[np.complex64]
22         )
23         # if an attribute with the same name as a parameter is found,
24         # a callback is registered (properties work, too).
25         self.additionFlag = additionFlag
26
27     def work(self, input_items, output_items):
28         """example: multiply with constant"""
29         output_items[0][:] = input_items[0] * self.additionFlag
30         return len(output_items[0])
31
32

```

Save the file. GRC displays the block with a second input and the block name is updated:

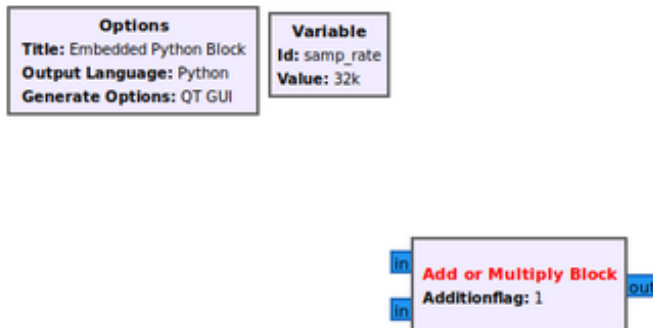


Figure 4.3.5.3: Changing block name

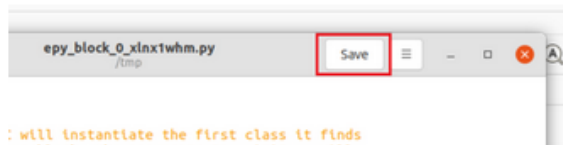
**Step-4:** Change the parameter name as required by editing in the code as shown in Fig. ??

```

4 Each time this file is saved, GRC will instantiate the first class it finds
5 to get ports and parameters of your block. The arguments to __init__ will
6 be the parameters. All of them are required to have default values!
7 """
8
9 import numpy as np
10 from gnuradio import gr
11
12
13 class blk(gr.sync_block): # other base classes are basic_block, decim_block, interp_block
14     """Embedded Python Block example - a simple multiply const"""
15
16     def __init__(self, additionFlag=True): # only default arguments here
17         """arguments to this function show up as parameters in GRC"""
18         gr.sync_block.__init__(
19             self,
20             name='Embedded Python Block', # will show up in GRC
21             in_sig=[np.complex64],
22             out_sig=[np.complex64]
23         )
24         # if an attribute with the same name as a parameter is found,
25         # a callback is registered (properties work, too).
26         self.additionFlag = additionFlag
27
28     def work(self, input_items, output_items):
29         """example: multiply with constant"""
30         output_items[0][:] = input_items[0] * self.additionFlag
31         return len(output_items[0])
32

```

Save the file:



Return back to the GRC window.

The *Embedded Python Block* displays the *Additionflag* parameter instead of *example\_param*:

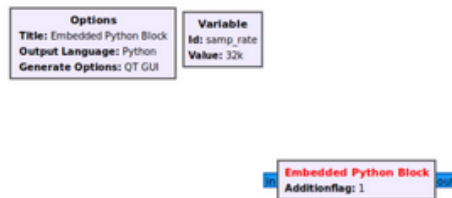


Figure 4.3.5.4: Changing parameter name

Replace the below code with default code

```
fm/rx/codes/lpf_block.py
```

6. Read the data from RTL SDR using python code and without using GNU radio.

**Solution:**

The following code reads the data from RTL-SDR without GNU radio.

```
fm/rx/codes/source_own.py
```

7. Design Low pass filter block without using GNU Radio

**Solution:**

The following code works as low pass filter.

```
fm/rx/codes/lpf_own.py
```





## Appendix A

# Discrete Fourier Transform

### A.1. FFT

1. The DFT of  $x(n)$  is given by

$$X(k) \triangleq \sum_{n=0}^{N-1} x(n)e^{-j2\pi kn/N}, \quad k = 0, 1, \dots, N-1 \quad (\text{A.1.1.1})$$

2. Let

$$W_N = e^{-j2\pi/N} \quad (\text{A.1.2.1})$$

Then the  $N$ -point DFT matrix is defined as

$$\mathbf{F}_N = [W_N^{mn}], \quad 0 \leq m, n \leq N-1 \quad (\text{A.1.2.2})$$

where  $W_N^{mn}$  are the elements of  $\mathbf{F}_N$ .

## A.2. FFT

1. Let

$$\mathbf{I}_4 = \begin{pmatrix} \mathbf{e}_4^1 & \mathbf{e}_4^2 & \mathbf{e}_4^3 & \mathbf{e}_4^4 \end{pmatrix} \quad (\text{A.2.1.1})$$

be the  $4 \times 4$  identity matrix. Then the 4 point DFT permutation matrix is defined as

$$\mathbf{P}_4 = \begin{pmatrix} \mathbf{e}_4^1 & \mathbf{e}_4^3 & \mathbf{e}_4^2 & \mathbf{e}_4^4 \end{pmatrix} \quad (\text{A.2.1.2})$$

2. The 4 point DFT diagonal matrix is defined as

$$\mathbf{D}_4 = \text{diag} \begin{pmatrix} W_8^0 & W_8^1 & W_8^2 & W_8^3 \end{pmatrix} \quad (\text{A.2.2.1})$$

3. Show that

$$W_N^2 = W_{N/2} \quad (\text{A.2.3.1})$$

4. Show that

$$\mathbf{F}_4 = \begin{bmatrix} \mathbf{I}_2 & \mathbf{D}_2 \\ \mathbf{I}_2 & -\mathbf{D}_2 \end{bmatrix} \begin{bmatrix} \mathbf{F}_2 & 0 \\ 0 & \mathbf{F}_2 \end{bmatrix} \mathbf{P}_4 \quad (\text{A.2.4.1})$$

5. Show that

$$\mathbf{F}_N = \begin{bmatrix} \mathbf{I}_{N/2} & \mathbf{D}_{N/2} \\ \mathbf{I}_{N/2} & -\mathbf{D}_{N/2} \end{bmatrix} \begin{bmatrix} \mathbf{F}_{N/2} & 0 \\ 0 & \mathbf{F}_{N/2} \end{bmatrix} \mathbf{P}_N \quad (\text{A.2.5.1})$$

6. Find

$$\mathbf{P}_4 \mathbf{x} \quad (\text{A.2.6.1})$$

7. Show that

$$\mathbf{X} = \mathbf{F}_N \mathbf{x} \quad (\text{A.2.7.1})$$

where  $\mathbf{x}, \mathbf{X}$  are the vector representations of  $x(n), X(k)$  respectively.

8. Derive the following Step-by-step visualisation of 8-point FFTs into 4-point FFTs and so on

$$\begin{bmatrix} X(0) \\ X(1) \\ X(2) \\ X(3) \end{bmatrix} = \begin{bmatrix} X_1(0) \\ X_1(1) \\ X_1(2) \\ X_1(3) \end{bmatrix} + \begin{bmatrix} W_8^0 & 0 & 0 & 0 \\ 0 & W_8^1 & 0 & 0 \\ 0 & 0 & W_8^2 & 0 \\ 0 & 0 & 0 & W_8^3 \end{bmatrix} \begin{bmatrix} X_2(0) \\ X_2(1) \\ X_2(2) \\ X_2(3) \end{bmatrix} \quad (\text{A.2.8.1})$$

$$\begin{bmatrix} X(4) \\ X(5) \\ X(6) \\ X(7) \end{bmatrix} = \begin{bmatrix} X_1(0) \\ X_1(1) \\ X_1(2) \\ X_1(3) \end{bmatrix} - \begin{bmatrix} W_8^0 & 0 & 0 & 0 \\ 0 & W_8^1 & 0 & 0 \\ 0 & 0 & W_8^2 & 0 \\ 0 & 0 & 0 & W_8^3 \end{bmatrix} \begin{bmatrix} X_2(0) \\ X_2(1) \\ X_2(2) \\ X_2(3) \end{bmatrix} \quad (\text{A.2.8.2})$$

4-point FFTs into 2-point FFTs

$$\begin{bmatrix} X_1(0) \\ X_1(1) \end{bmatrix} = \begin{bmatrix} X_3(0) \\ X_3(1) \end{bmatrix} + \begin{bmatrix} W_4^0 & 0 \\ 0 & W_4^1 \end{bmatrix} \begin{bmatrix} X_4(0) \\ X_4(1) \end{bmatrix} \quad (\text{A.2.8.3})$$

$$\begin{bmatrix} X_1(2) \\ X_1(3) \end{bmatrix} = \begin{bmatrix} X_3(0) \\ X_3(1) \end{bmatrix} - \begin{bmatrix} W_4^0 & 0 \\ 0 & W_4^1 \end{bmatrix} \begin{bmatrix} X_4(0) \\ X_4(1) \end{bmatrix} \quad (\text{A.2.8.4})$$

$$\begin{bmatrix} X_2(0) \\ X_2(1) \end{bmatrix} = \begin{bmatrix} X_5(0) \\ X_5(1) \end{bmatrix} + \begin{bmatrix} W_4^0 & 0 \\ 0 & W_4^1 \end{bmatrix} \begin{bmatrix} X_6(0) \\ X_6(1) \end{bmatrix} \quad (\text{A.2.8.5})$$

$$\begin{bmatrix} X_2(2) \\ X_2(3) \end{bmatrix} = \begin{bmatrix} X_5(0) \\ X_5(1) \end{bmatrix} - \begin{bmatrix} W_4^0 & 0 \\ 0 & W_4^1 \end{bmatrix} \begin{bmatrix} X_6(0) \\ X_6(1) \end{bmatrix} \quad (\text{A.2.8.6})$$

$$P_8 \begin{bmatrix} x(0) \\ x(1) \\ x(2) \\ x(3) \\ x(4) \\ x(5) \\ x(6) \\ x(7) \end{bmatrix} = \begin{bmatrix} x(0) \\ x(2) \\ x(4) \\ x(6) \\ x(1) \\ x(3) \\ x(5) \\ x(7) \end{bmatrix} \quad (\text{A.2.8.7})$$

$$P_4 \begin{bmatrix} x(0) \\ x(2) \\ x(4) \\ x(6) \end{bmatrix} = \begin{bmatrix} x(0) \\ x(4) \\ x(2) \\ x(6) \end{bmatrix} \quad (\text{A.2.8.8})$$

$$P_4 \begin{bmatrix} x(1) \\ x(3) \\ x(5) \\ x(7) \end{bmatrix} = \begin{bmatrix} x(1) \\ x(5) \\ x(3) \\ x(7) \end{bmatrix} \quad (\text{A.2.8.9})$$

Therefore,

$$\begin{bmatrix} X_3(0) \\ X_3(1) \end{bmatrix} = F_2 \begin{bmatrix} x(0) \\ x(4) \end{bmatrix} \quad (\text{A.2.8.10})$$

$$\begin{bmatrix} X_4(0) \\ X_4(1) \end{bmatrix} = F_2 \begin{bmatrix} x(2) \\ x(6) \end{bmatrix} \quad (\text{A.2.8.11})$$

$$\begin{bmatrix} X_5(0) \\ X_5(1) \end{bmatrix} = F_2 \begin{bmatrix} x(1) \\ x(5) \end{bmatrix} \quad (\text{A.2.8.12})$$

$$\begin{bmatrix} X_6(0) \\ X_6(1) \end{bmatrix} = F_2 \begin{bmatrix} x(3) \\ x(7) \end{bmatrix} \quad (\text{A.2.8.13})$$

9. For

$$\mathbf{x} = \begin{pmatrix} 1 \\ 2 \\ 3 \\ 4 \\ 2 \\ 1 \end{pmatrix} \quad (\text{A.2.9.1})$$

compute the DFT using (A.2.7.1)

10. Repeat the above exercise using the FFT after zero padding  $\mathbf{x}$ .

11. Write a C program to compute the 8-point FFT.

## A.3. Power Spectral Density(PSD)

1. Power spectral density (PSD) is a measure of the power distribution over frequency components of a signal. The PSD of (A.1.1.1) is given by

$$X(k) = |X(k)|^2 \quad (\text{A.3.1.1})$$

# Appendix B

## Filter Design

### B.1. Introduction

We are supposed to design the equivalent FIR and IIR filter realizations for filter number 114. This is a bandpass filter whose specifications are available below.

### B.2. Filter Specifications

The sampling rate for the filter has been specified as  $F_s = 48$  kHz. If the un-normalized discrete-time (natural) frequency is  $F$ , the corresponding normalized digital filter (angular) frequency is given by  $\omega = 2\pi \left( \frac{F}{F_s} \right)$ .

#### B.2.1. The Digital Filter

1. Tolerances: The passband ( $\delta_1$ ) and stopband ( $\delta_2$ ) tolerances are given to be equal, so we let  $\delta_1 = \delta_2 = \delta = 0.15$ .
2. Passband: The passband of filter number  $j$ ,  $j$  going from 109 to 135 is from  $\{3 + 0.6(j-109)\}$ kHz to  $\{3 + 0.6(j-107)\}$ kHz. Since our filter number is 114, substituting



$j = 114$  gives the passband range for our bandpass filter as 6 kHz - 7.2 kHz. Hence, the un-normalized discrete time filter passband frequencies are  $F_{p1} = 7.2$  kHz and  $F_{p2} = 6.0$  kHz. The corresponding normalized digital filter passband frequencies are  $\omega_{p1} = 2\pi \frac{F_{p1}}{F_s} = 0.3\pi$  and  $\omega_{p2} = 2\pi \frac{F_{p2}}{F_s} = 0.25\pi$  kHz. The centre frequency is then given by  $\omega_c = \frac{\omega_{p1} + \omega_{p2}}{2} = 0.275\pi$ .

3. Stopband: The transition band for bandpass filters is  $\Delta F = 0.3$  kHz on either side of the passband. Hence, the un-normalized stopband frequencies are  $F_{s1} = 7.2 + 0.3 = 7.5$  kHz and  $F_{s2} = 6.0 - 0.3 = 5.7$  kHz. The corresponding normalized frequencies are  $\omega_{s1} = 0.3125\pi$  and  $\omega_{s2} = 0.2375\pi$ .

## B.2.2. The Analog filter

In the bilinear transform, the analog filter frequency ( $\Omega$ ) is related to the corresponding digital filter frequency ( $\omega$ ) as  $\Omega = \tan \frac{\omega}{2}$ . Using this relation, we obtain the analog passband and stopband frequencies as  $\Omega_{p1} = 0.5095$ ,  $\Omega_{p2} = 0.4142$  and  $\Omega_{s1} = 0.5345$ ,  $\Omega_{s2} = 0.3914$  respectively.

## B.3. The IIR Filter Design

Filter Type: We are supposed to design filters whose stopband is monotonic and passband equiripple. Hence, we use the Chebyshev approximation to design our bandpass IIR filter.

### B.3.1. The Analog Filter

1. Low Pass Filter Specifications: If  $H_{a,BP}(j\Omega)$  be the desired analog band pass filter, with the specifications provided in Section 2.2, and  $H_{a,LP}(j\Omega_L)$  be the equivalent low pass filter, then

$$\Omega_L = \frac{\Omega^2 - \Omega_0^2}{B\Omega} \quad (1.1)$$

where  $\Omega_0 = \sqrt{\Omega_{p1}\Omega_{p2}} = 0.4594$  and  $B = \Omega_{p1} - \Omega_{p2} = 0.0953$ . The low pass filter has the passband edge at  $\Omega_{Lp} = 1$  and stopband edges at  $\Omega_{Ls1} = 1.4653$  and  $\Omega_{Ls2} = -1.5511$ . We choose the stopband edge of the analog low pass filter as  $\Omega_{Ls} = \min(|\Omega_{Ls1}|, |\Omega_{Ls2}|) = 1.4653$ .

2. The Low Pass Chebyshev Filter Paramters: The magnitude squared of the Chebyshev low pass filter is given by

$$|H_{a,LP}(j\Omega_L)|^2 = \frac{1}{1 + \epsilon^2 c_N^2(\Omega_L/\Omega_{Lp})} \quad (2.1)$$

where  $c_N(x) = \cosh(N \cosh^{-1} x)$  and the integer  $N$ , which is the order of the filter, and  $\epsilon$  are design paramters. Since  $\Omega_{Lp} = 1$ , (2.1) may be rewritten as

$$|H_{a,LP}(j\Omega_L)|^2 = \frac{1}{1 + \epsilon^2 c_N^2(\Omega_L)} \quad (2.2)$$

Also, the design paramters have the following constraints

$$N \geq \left\lceil \frac{\frac{\sqrt{D_2}}{c_N(\Omega_{Ls})} \leq \epsilon \leq \sqrt{D_1},}{\frac{\cosh^{-1} \sqrt{D_2/D_1}}{\cosh^{-1} \Omega_{Ls}}} \right\rceil, \quad (2.3)$$

where  $D_1 = \frac{1}{(1-\delta)^2} - 1$  and  $D_2 = \frac{1}{\delta^2} - 1$ . After appropriate substitutions, we obtain

$N \geq 4$  and  $0.3184 \leq \epsilon \leq 0.6197$ . In Figure 1, we plot  $|H(j\Omega)|$  for a range of values of  $\epsilon$ , for  $N = 4$ . We find that for larger values of  $\epsilon$ ,  $|H(j\Omega)|$  decreases in the transition band. We choose  $\epsilon = 0.4$  for our IIR filter design.

3. The Low Pass Chebyshev Filter: Thus, we obtain

$$|H_{a,LP}(j\Omega_L)|^2 = \frac{1}{1 + 0.16c_4^2(\Omega_L)} \quad (3.1)$$

where

$$c_4(x) = 8x^4 + 8x^2 + 1. \quad (3.2)$$

The poles of the frequency response in (2.1) lying in the left half plane are in general obtained as  $r_1 \cos \phi_k + jr_2 \sin \phi_k$ , where

$$\begin{aligned} \phi_k &= \frac{\pi}{2} + \frac{(2k+1)\pi}{2N}, k = 0, 1, \dots, N-1 \\ r_1 &= \frac{\beta^2 - 1}{2\beta}, r_2 = \frac{\beta^2 + 1}{2\beta}, \beta = \left[ \frac{\sqrt{1 + \epsilon^2} + 1}{\epsilon} \right]^{\frac{1}{N}} \end{aligned} \quad (3.3)$$

Thus, for  $N$  even, the low-pass stable Chebyshev filter, with a gain  $G$  has the form

$$H_{a,LP}(s_L) = \frac{G_{LP}}{\prod_{k=1}^{\frac{N}{2}-1} (s_L^2 - 2r_1 \cos \phi_k s_L + r_1^2 \cos^2 \phi_k + r_2^2 \sin^2 \phi_k)} \quad (3.4)$$

Substituting  $N = 4$ ,  $\epsilon = 0.5$  and  $H_{a,LP}(j) = \frac{1}{\sqrt{1+\epsilon^2}}$ , from (3.3) and (3.4), we obtain

$$H_{a,LP}(s_L) = \frac{0.3125}{s_L^4 + 1.1068s_L^3 + 1.6125s_L^2 + 0.9140s_L + 0.3366} \quad (3.5)$$

In Fig. 4.2 we plot  $|H(j\Omega)|$  using (3.1) and (3.5), thereby verifying that our low-pass Chebyshev filter design meets the specifications.

4. The Band Pass Chebyshev Filter: The analog bandpass filter is obtained from (3.5) by substituting  $s_L = \frac{s^2 + \Omega_0^2}{Bs}$ . Hence

$$H_{a,BP}(s) = G_{BP}H_{a,LP}(s_L) \Big|_{s_L = \frac{s^2 + \Omega_0^2}{Bs}}, \quad (4.1)$$

where  $G_{BP}$  is the gain of the bandpass filter. After appropriate substitutions, and evaluating the gain such that  $H_{a,BP}(j\Omega_{p1}) = 1$ , we obtain

$$H_{a,BP}(s) = \frac{2.7776 \times 10^{-5} s^4}{s^8 + 0.1055s^7 + 0.8589s^6 + 0.0676s^5 + 0.2735s^4 + 0.0143s^3 + 0.0383s^2 + 0.001s + 0.002} \quad (4.2)$$

In Fig 4.3, we plot  $|H_{a,BP}(j\Omega)|$  as a function of  $\Omega$  for both positive as

well as negative frequencies. We find that the passband and stopband frequencies in the figure match well with those obtained analytically through the bilinear transformation.

### B.3.2. The Digital Filter

From the bilinear transformation, we obtain the digital bandpass filter from the corresponding analog filter as

$$H_{d,BP}(z) = GH_{a,BP}(s) \Big|_{s = \frac{1-z^{-1}}{1+z^{-1}}} \quad (4.3)$$

where  $G$  is the gain of the digital filter. From (4.2) and (4.3), we obtain

$$H_{d,BP}(z) = G \frac{N(z)}{D(z)} \quad (4.4)$$

where  $G = 2.7776 \times 10^{-5}$ ,

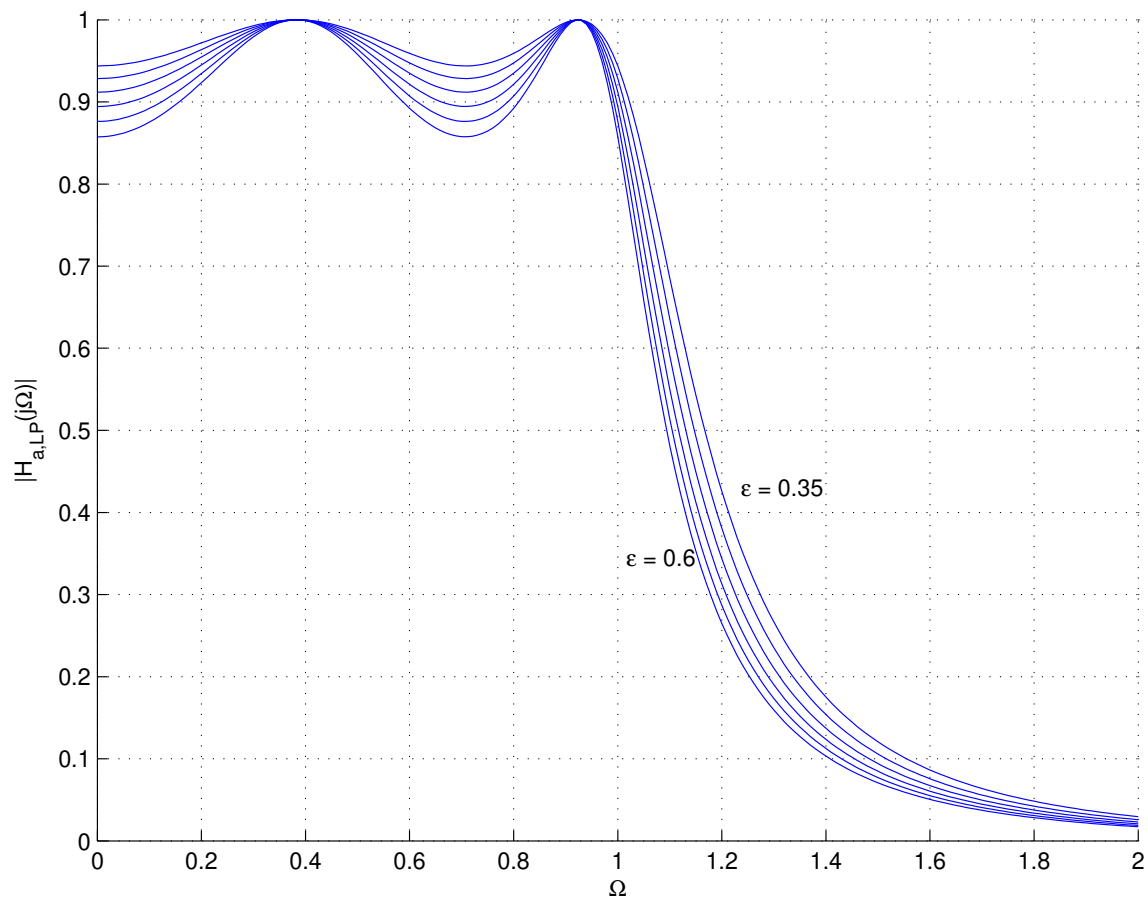


Figure 4.1: The Analog Low-Pass Frequency Response for  $0.35 \leq \epsilon \leq 0.6$

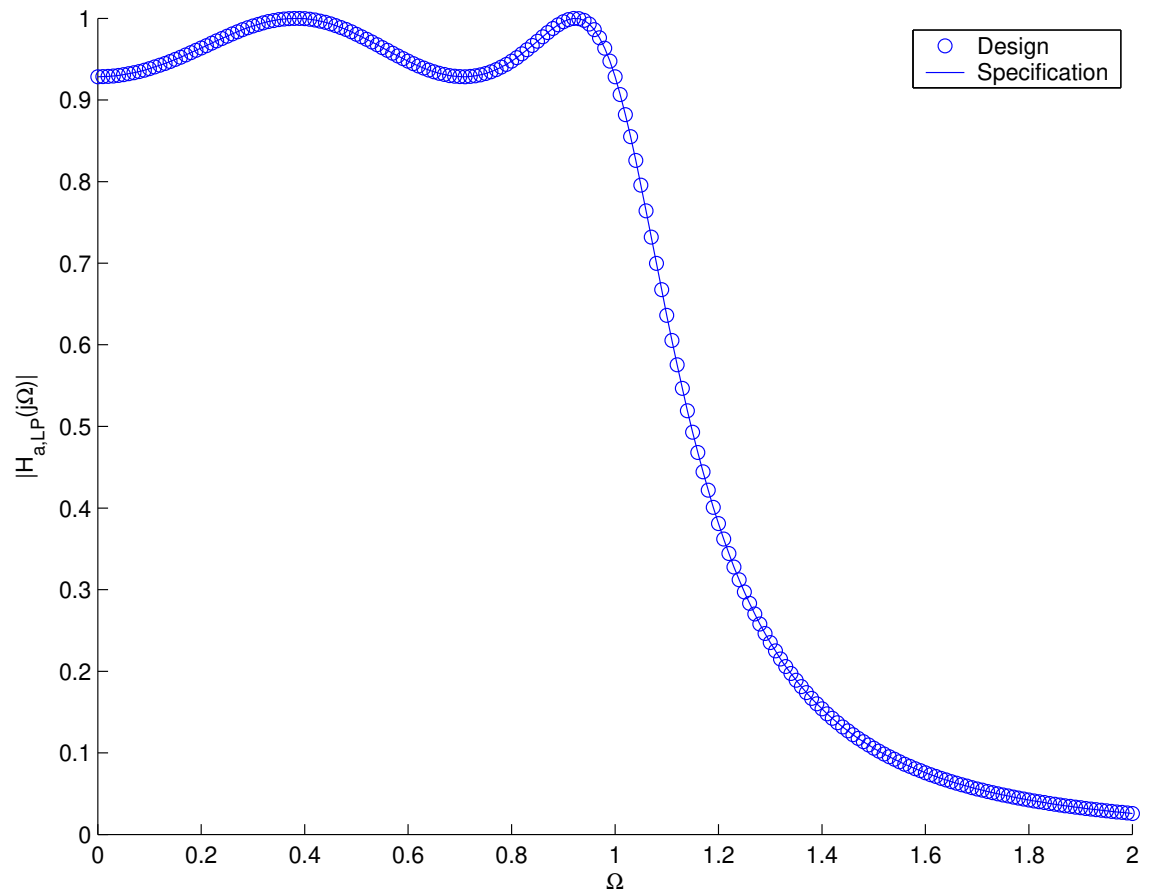


Figure 4.2: The magnitude response plots from the specifications in Equation 3.1 and the design in Equation 3.5

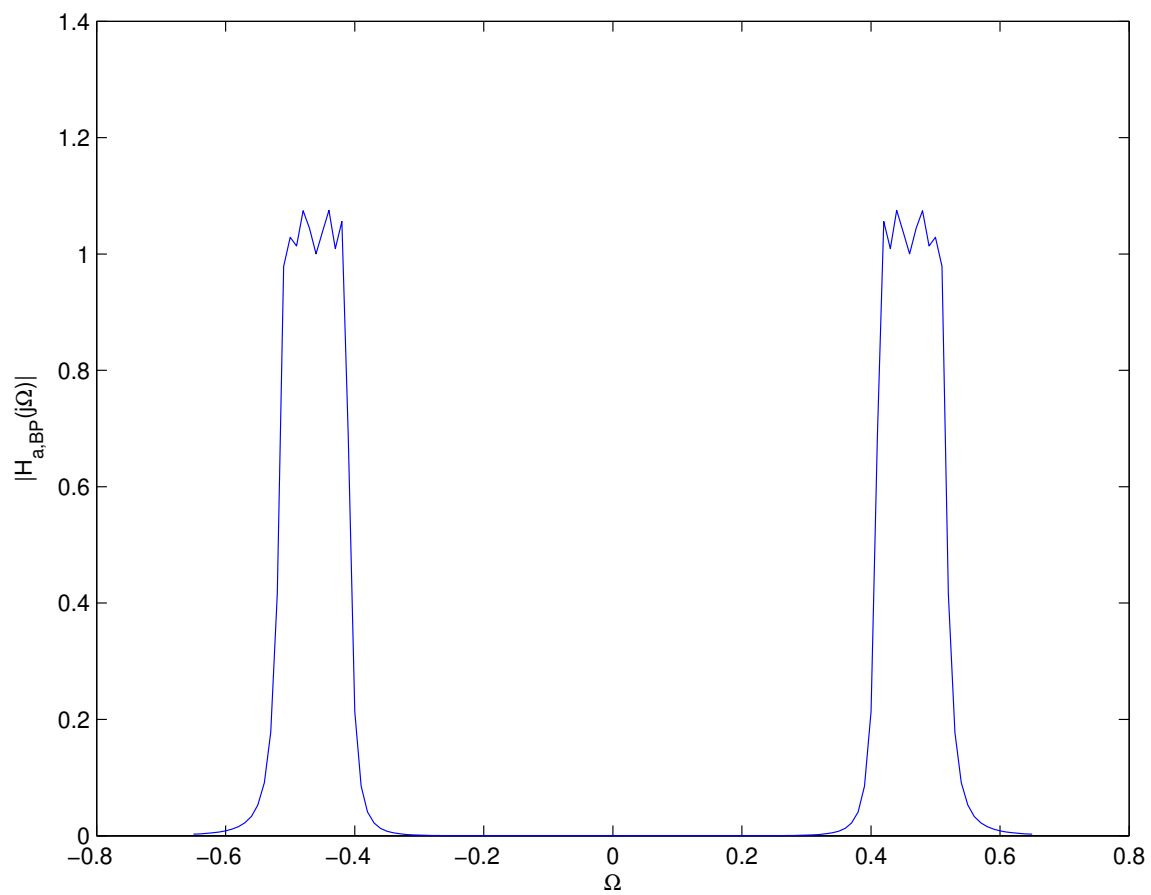


Figure 4.3: The analog bandpass magnitude response plot from Equation 4.2

$$N(z) = 1 - 4z^{-2} + 6z^{-4} - 4z^{-6} + z^{-8} \quad (4.5)$$

and

$$\begin{aligned} D(z) = & 2.3609 - 12.0002z^{-1} + 31.8772z^{-2} - 53.7495z^{-3} + 62.8086z^{-4} \\ & - 51.4634z^{-5} + 29.2231z^{-6} - 10.5329z^{-7} + 1.9842z^{-8} \end{aligned} \quad (4.6)$$

The plot of  $|H_{d,BP}(z)|$  with respect to the normalized angular frequency (normalizing factor  $\pi$ ) is available in Fig. 4.4. Again we

find that the passband and stopband frequencies meet the specifications well enough.

## B.4. The FIR Filter

We design the FIR filter by first obtaining the (non-causal) lowpass equivalent using the Kaiser window and then converting it to a causal bandpass filter.

### B.4.1. The Equivalent Lowpass Filter

The lowpass filter has a passband frequency  $\omega_l$  and transition band  $\Delta\omega = 2\pi\frac{\Delta F}{F_s} = 0.0125\pi$ . The stopband tolerance is  $\delta$ .

1. The passband frequency  $\omega_l$  is defined as  $\omega_l = \frac{\omega_{p1} - \omega_{p2}}{2}$ . Substituting the values of  $\omega_{p1}$  and  $\omega_{p2}$  from section 2.1, we obtain  $\omega_l = 0.025\pi$ .
2. The impulse response  $h_{lp}(n)$  of the desired lowpass filter with cutoff frequency  $\omega_l$  is



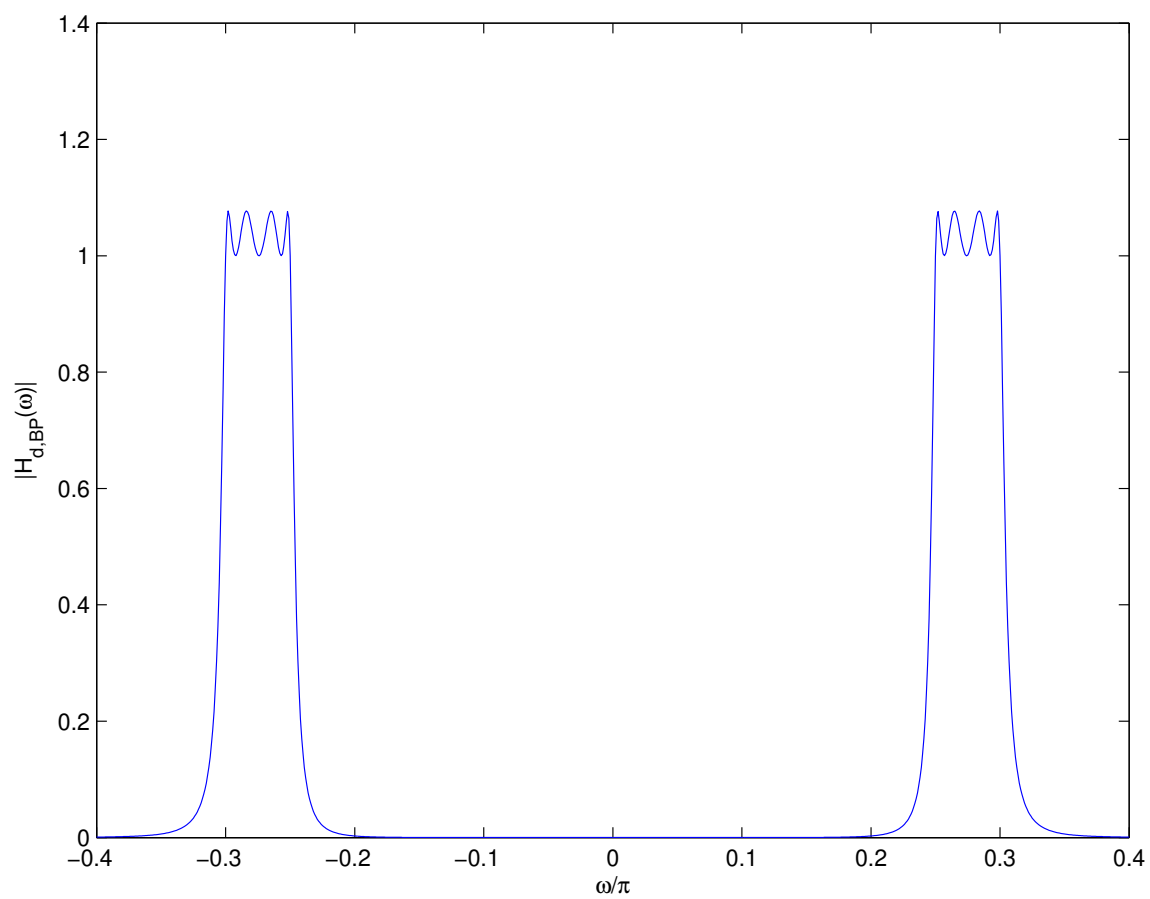


Figure 4.4: The magnitude response of the bandpass digital filter designed to meet the given specifications

given by

$$h_l(n) = \frac{\sin(n\omega_l)}{n\pi} w(n), \quad (2.1)$$

where  $w(n)$  is the Kaiser window obtained from the design specifications.

## B.4.2. The Kaiser Window

The Kaiser window is defined as

$$\begin{aligned} w(n) &= \frac{I_0 \left[ \beta N \sqrt{1 - \left( \frac{n}{N} \right)^2} \right]}{I_0(\beta N)}, & -N \leq n \leq N, & \quad \beta > 0 \\ &= 0 & \text{otherwise,} & \end{aligned} \quad (2.2)$$

where  $I_0(x)$  is the modified Bessel function of the first kind of order zero in  $x$  and  $\beta$  and  $N$  are the window shaping factors. In the following, we find  $\beta$  and  $N$  using the design parameters in section 2.1.

1.  $N$  is chosen according to

$$N \geq \frac{A - 8}{4.57\Delta\omega}, \quad (1.1)$$

where  $A = -20 \log_{10} \delta$ . Substituting the appropriate values from the design specifications, we obtain  $A = 16.4782$  and  $N \geq 48$ .

2.  $\beta$  is chosen according to

$$\beta N = \begin{cases} 0.1102(A - 8.7) & A > 50 \\ 0.5849(A - 21)^{0.4} + 0.07886(A - 21) & 21 \leq A \leq 50 \\ 0 & A < 21 \end{cases} \quad (2.1)$$

In our design, we have  $A = 16.4782 < 21$ . Hence, from (2.1) we obtain  $\beta = 0$ .

3. We choose  $N = 100$ , to ensure the desired low pass filter response. Substituting in (2.2) gives us the rectangular window

$$\begin{aligned} w(n) &= 1, -100 \leq n \leq 100 \\ &= 0 \quad \text{otherwise} \end{aligned} \tag{3.1}$$

From (2.1) and (3.1), we obtain the desired lowpass filter impulse response

$$\begin{aligned} h_{lp}(n) &= \frac{\sin(\frac{n\pi}{40})}{n\pi} \quad -100 \leq n \leq 100 \\ &= 0, \quad \text{otherwise} \end{aligned} \tag{3.2}$$

The magnitude response of the filter in (3.2) is shown in Figure 5.

### B.4.3. The FIR Bandpass Filter

The centre of the passband of the desired bandpass filter was found to be  $\omega_c = 0.275\pi$  in Section 2.1. The impulse response of the desired bandpass filter is obtained from the impulse response of the corresponding lowpass filter as

$$h_{bp}(n) = 2h_{lp}(n)\cos(n\omega_c) \tag{3.3}$$

Thus, from (3.2), we obtain

$$\begin{aligned} h_{bp}(n) &= \frac{2\sin(\frac{n\pi}{40})\cos(\frac{11n\pi}{40})}{n\pi} \quad -100 \leq n \leq 100 \\ &= 0, \quad \text{otherwise} \end{aligned} \tag{3.4}$$

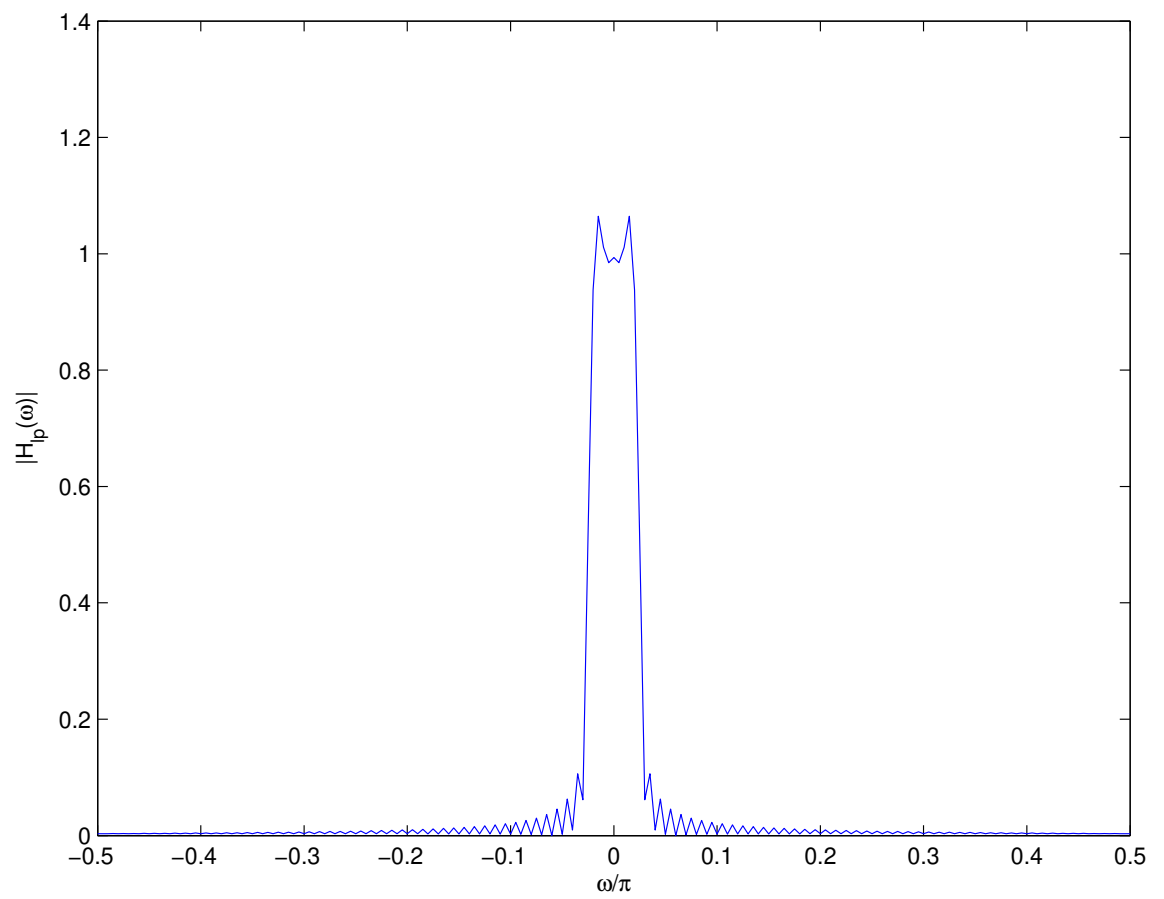


Figure 3.1: The magnitude response of the FIR lowpass digital filter designed to meet the given specifications

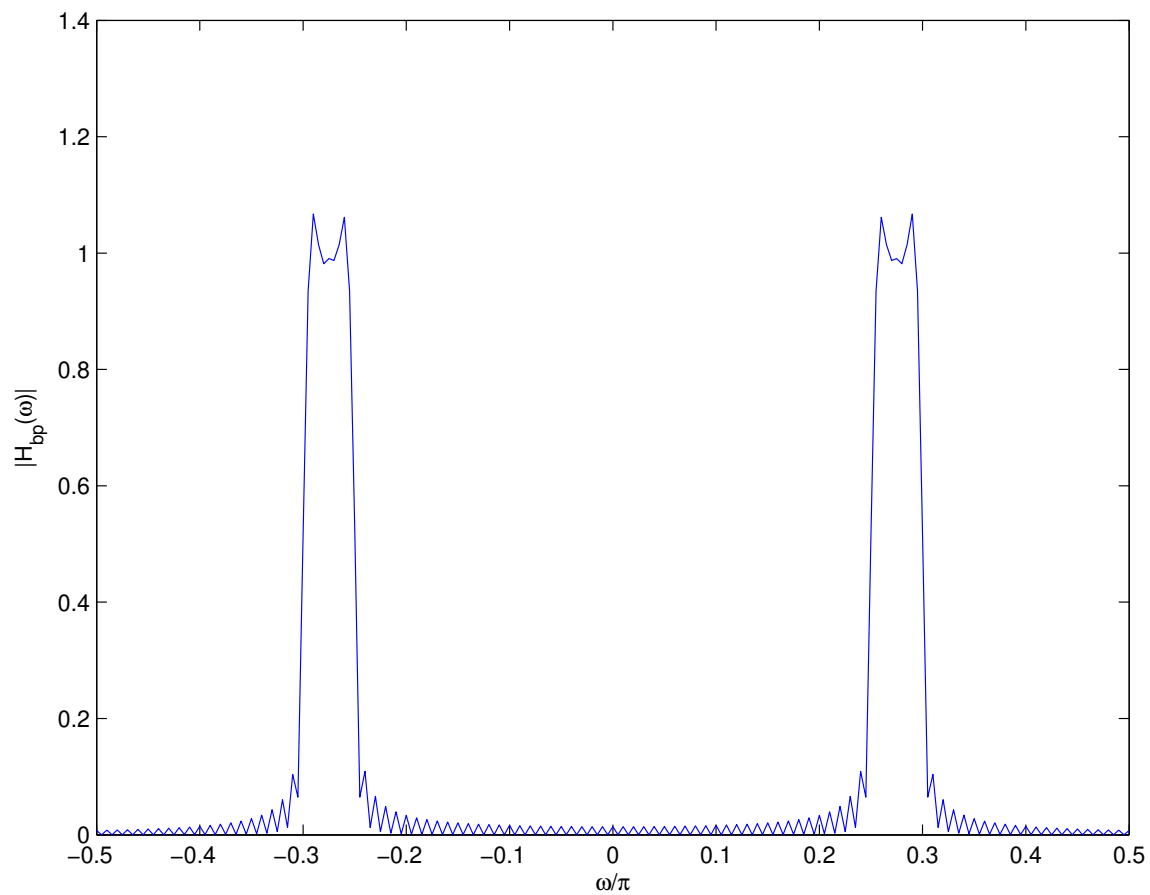


Figure 3.2: The magnitude response of the FIR bandpass digital filter designed to meet the given specifications

The magnitude response of the FIR bandpass filter designed to meet the given specifications is plotted in Fig. 3.2.