Norwegian University of Science and Technology
Department of Electronics and Telecommunications

TTT4120 Digital Signal Processing
Suggested Solutions for Problem Set 2

Problem 1

(a) The spectrum $X(\omega)$ can be found as follows.

$$\begin{aligned} X(\omega) &= \sum_{n=-\infty}^{\infty} x(n)e^{-j\omega n} \\ &= e^{j\omega} + 2 + e^{-j\omega} \\ &= 2 + 2\cos\omega \end{aligned}$$

It is shown in Figure 1.

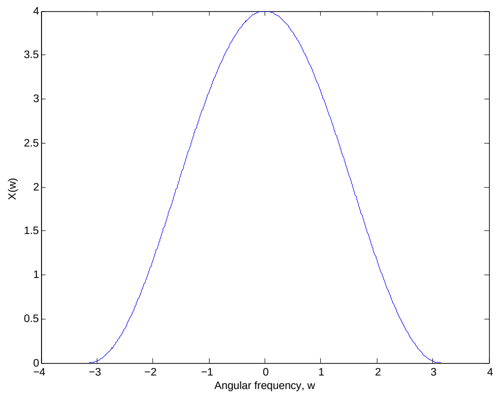

Figure 1: The spectrum $X(\omega)$

(b) The spectrum $Y(\omega)$ can be found as follows.

$$\begin{aligned}
Y(\omega) &= \sum_{n=-\infty}^{\infty} y(n)e^{-j\omega n} \\
&= \sum_{n=-M}^{M} e^{-j\omega n} \qquad l = n + M \\
&= \sum_{l=0}^{2M} e^{-j\omega(l-M)} \\
&= e^{j\omega M} \sum_{l=0}^{2M} e^{-j\omega l} \\
&= e^{j\omega M} \frac{1 - e^{-j\omega(2M+1)}}{1 - e^{-j\omega}} \\
&= \frac{e^{j\omega M} - e^{-j\omega(M+1)}}{1 - e^{-j\omega}} \\
&= \frac{e^{-\frac{j\omega}{2}}}{e^{-\frac{j\omega}{2}}} \frac{\left(e^{j\omega(M+\frac{1}{2})} - e^{-j\omega(M+\frac{1}{2})}\right)}{\left(e^{\frac{j\omega}{2}} - e^{-\frac{j\omega}{2}}\right)} \\
&= \frac{\sin\left(\omega(M+\frac{1}{2})\right)}{\sin(\frac{\omega}{2})}
\end{aligned}$$

The sketch is shown in Figure 2.

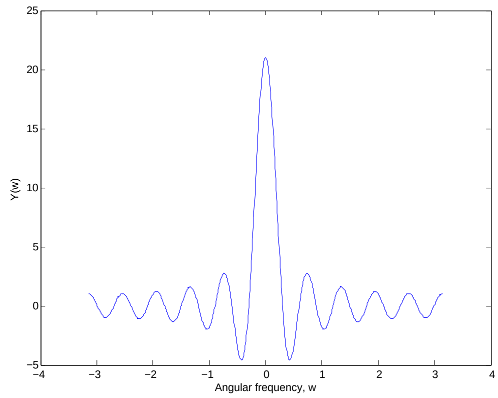

Figure 2: The spectrum $Y(\omega)$ for M=10

(c) Because they are even signals.

(d) A sketch of $z(n)$ for N=5 is shown in Figure 3. The Fourier coefficients are given by:

$$c_k = \frac{1}{N} \sum_{n=0}^{N-1} z(n)e^{-j2\pi kn/N}, \quad k = 0, \cdots, N-1.$$

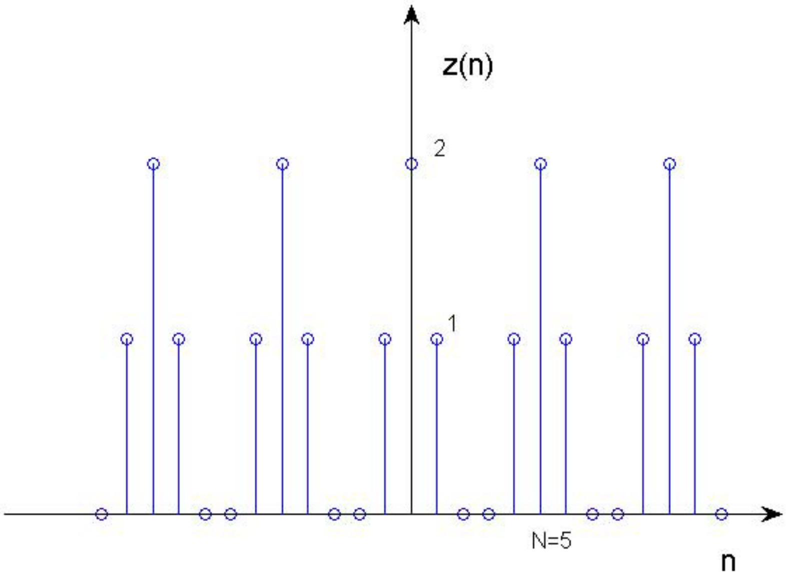

Figure 3: The signal $z(n)$, periodic extension of $x(n)$

Note that we sum from 0 up to $N-1$. Thus, the first two samples are 2 and 1 respectively, and the last sample is 1. All other samples are 0. The coefficients could be calculated over any other period.

$$\begin{aligned}
c_k &= \frac{1}{N}\sum_{n=0}^{N-1} z(n)e^{-j2\pi kn/N} \\
&= \frac{1}{N}(2 + e^{-j2\pi k/N} + e^{-j2\pi k(N-1)/N}) \\
&= \frac{1}{N}(2 + e^{-j2\pi k/N} + e^{-j2\pi k}e^{j2\pi k/N}) \\
&= \frac{1}{N}(2 + e^{-j2\pi k/N} + e^{j2\pi k/N}) \\
&= \frac{1}{N}(2 + 2\cos(2\pi k/N))
\end{aligned}$$

The Fourier coefficients are displayed in Figure 4.

(e) We have the following.

$$\begin{aligned}
X(f) &= 2 + 2\cos(2\pi f) \\
c_k &= \frac{1}{N}(2 + 2\cos(2\pi k/N))
\end{aligned}$$

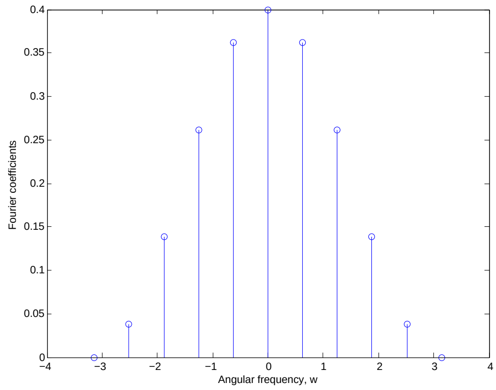

Figure 4: The Fourier coefficients c_k of $z(n)$ for $k = -5, ..., 5$

Thus, we see that

$$c_k = \frac{1}{N} X\left(\frac{k}{N}\right).$$

This means that the Fourier coefficients are (scaled) samples of the continuous spectrum $X(f)$. This always holds true: a periodic extension in the time domain equals sampling in the frequency domain.

Problem 2

(a) For the first case, we use the time-shift property of the DTFT, and get

$$X_1(\omega) = e^{j3\omega} X(\omega)$$

(b) For the second case, we use the time-reversal property of the DTFT, and it follows that

$$X_2(\omega) = X(-\omega)$$

(c) For the third case notice that:

$$x_3(n) = x(3-n) = x(-(n-3)) = x_2(n-3)$$

so that by the time-reversal and time-shift properties, it follows that

$$X_3(\omega) = e^{-j3\omega} X_2(\omega) = e^{-j3\omega} X(-\omega)$$

(d) For the last case, we have that

$$X_4(\omega) = \mathrm{DTFT}\{x(n) * w(n)\} = X(\omega)W(\omega).$$

Problem 3

(a) By taking the DTFT of both sides of the first difference equation, we get

$$\begin{aligned}
Y(\omega) &= X(\omega) + 2e^{-j\omega}X(\omega) + e^{-2j\omega}X(\omega) \\
H_1(\omega) &= \frac{Y(\omega)}{X(\omega)} = 1 + 2e^{-j\omega} + e^{-2j\omega} \\
&= e^{-j\omega}(e^{j\omega} + 2 + e^{-j\omega}) \\
&= e^{-j\omega}(2 + 2\cos\omega).
\end{aligned}$$

And for the second case, we get

$$\begin{aligned}
Y(\omega) &= -0.9Y(\omega)e^{-j\omega} + X(\omega) \\
H_2(\omega) &= \frac{Y(\omega)}{X(\omega)} = \frac{1}{1 + 0.9e^{-j\omega}}.
\end{aligned}$$

(b) We already have the frequency response $H_1(\omega)$ on polar form. Thus, the magnitude is simply

$$|H_1(\omega)| = 2 + 2\cos\omega.$$

Since $2 + 2\cos\omega \geq 0$ for all ω, the phase is simply

$$\Theta_1(\omega) = \measuredangle H_1(\omega) = -\omega.$$

The magnitude response of the second system can be found as follows.

$$\begin{aligned}
|H_2(\omega)| &= \left|\frac{1}{1 + 0.9e^{-j\omega}}\right| \\
&= \frac{1}{|1 + 0.9e^{-j\omega}|} \\
&= \frac{1}{\sqrt{(1 + 0.9\cos\omega)^2 + (0.9\sin\omega)^2}} \\
&= \frac{1}{\sqrt{1 + 1.8\cos\omega + 0.81}}
\end{aligned}$$

To find the phase, we can write $H_2(\omega)$ as

$$H_2(\omega) = \frac{1}{W(\omega)},$$

where $W(\omega) = 1 + 0.9e^{-j\omega}$. Then, the phase is given by

$$\Theta_2(\omega) = \measuredangle H_2(\omega) = -\measuredangle W(\omega).$$

Since $\mathrm{Re}\{W(\omega)\} > 0$ for all ω, we have

$$\begin{aligned}
\measuredangle H_2(\omega) &= -\tan^{-1}\left(\frac{-0.9\sin\omega}{1 + 0.9\cos\omega}\right) \\
&= \tan^{-1}\left(\frac{0.9\sin\omega}{1 + 0.9\cos\omega}\right).
\end{aligned}$$

We notice that all magnitude functions are even and that all phase functions are odd. This is a property of real signals.

(c) The frequency response of the first filter can be found and plotted by the following code.

```
[H_1, w] = freqz([1 2 1], [1]);
subplot(2, 1, 1);
plot(w, abs(H_1));
xlabel('Angular frequency, w');
ylabel('Magnitude');
subplot(2, 1, 2);
plot(w, angle(H_1));
xlabel('Angular frequency, w');
ylabel('Phase');
```

For the second filter, we change the `freqz` command as follows.

```
[H_2, w] = freqz([1], [1 0.9]);
```

This gives the plots shown in Figures 5 and 6.

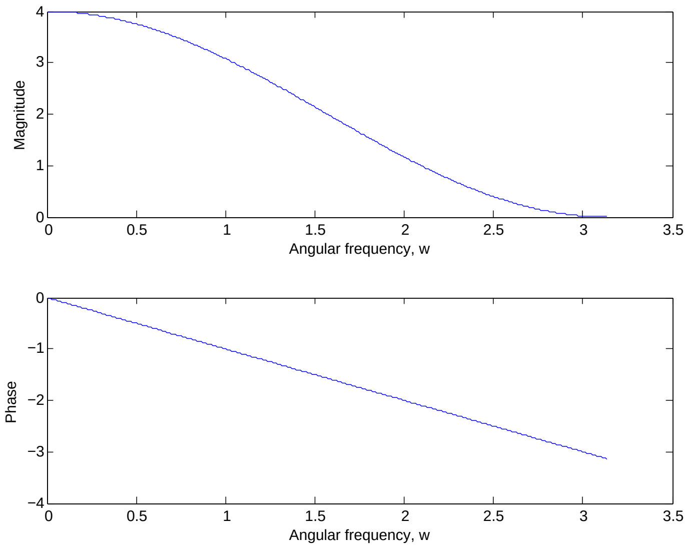

Figure 5: Magnitude and phase response of $H_1(\omega)$

(d) From the plots of the magnitude responses, we can see that the first filter attenuates high frequencies more than low frequencies. Thus, this is a lowpass filter. The second filter attenuates low frequencies more than high frequencies. Thus, this is a highpass filter.

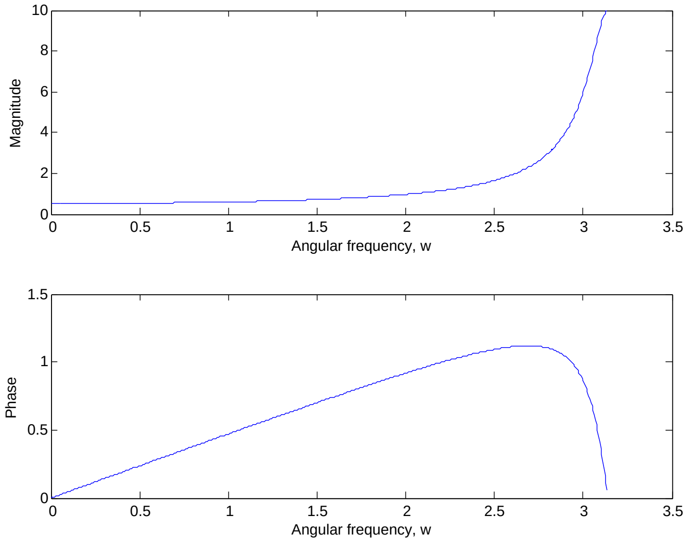

Figure 6: Magnitude and phase response of $H_2(\omega)$

(e) The response of a LTI-system $H(\omega) = |H(\omega)|e^{j\Theta(\omega)}$ to a sinusoidal input signal $x(n) = A\cos(\omega_0 n + \theta)$ equals

$$y(n) = A|H(\omega_0)|\cos(\omega_0 n + \theta + \Theta(\omega_0)).$$

Thus, the output of the first system is

$$\begin{aligned} y_1(n) &= \frac{1}{2}|H_1(\frac{\pi}{2})|\cos(\frac{\pi}{2}n + \frac{\pi}{4} + \Theta_1(\frac{\pi}{2})) \\ &= \frac{1}{2}\cdot 2\cos(\frac{\pi}{2}n + \frac{\pi}{4} - \frac{\pi}{2}) \\ &= \cos(\frac{\pi}{2}n - \frac{\pi}{4}). \end{aligned}$$

Likewise, the output of the second system is

$$\begin{aligned} y_2(n) &= \frac{1}{2}|H_2(\frac{\pi}{2})|\cos(\frac{\pi}{2}n + \frac{\pi}{4} + \Theta_2(\frac{\pi}{2})) \\ &= \frac{1}{2}\frac{1}{\sqrt{1.81 + 1.8\cos(\frac{\pi}{2})}}\cos(\frac{\pi}{2}n + \frac{\pi}{4} + \tan^{-1}(\frac{0.9\sin(\frac{\pi}{2})}{1 + 0.9\cos(\frac{\pi}{2})}) \\ &= \frac{1}{2}\frac{1}{\sqrt{1.81}}\cos(\frac{\pi}{2}n + \frac{\pi}{4} + \tan^{-1}(\frac{9}{10})) \\ &\approx \frac{1}{2}\frac{1}{\sqrt{1.81}}\cos(\frac{\pi}{2}n + 1.52)). \end{aligned}$$

Problem 4

(a) The spectra of the sampled signals are shown in Figures 7 and 8. The latter has a wider range of frequencies than the required $f \in [-\frac{1}{2}, \frac{1}{2}]$ to help making difference between alias components and signal components. The theory behind this is in ch.6.

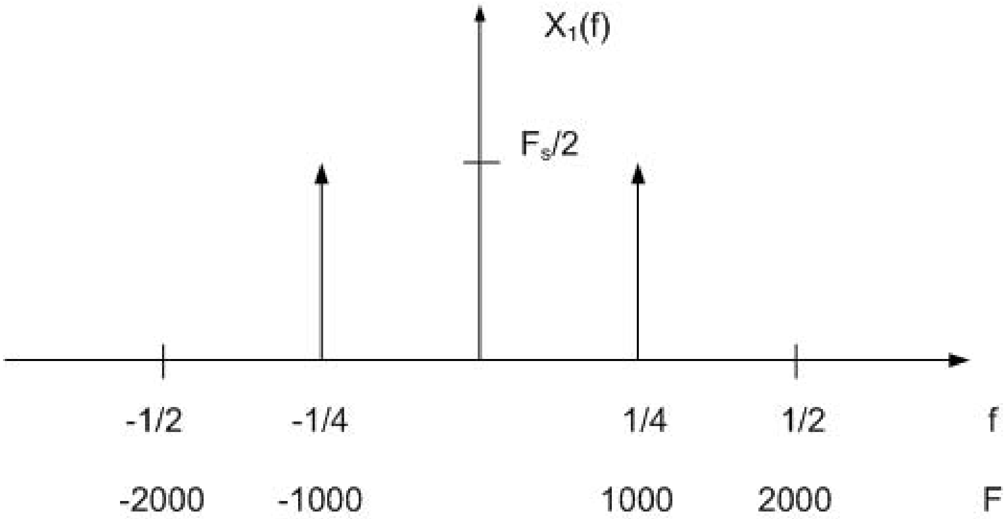

Figure 7: Spectrum of the signal $x(n)$ when $F_s = 4000$Hz

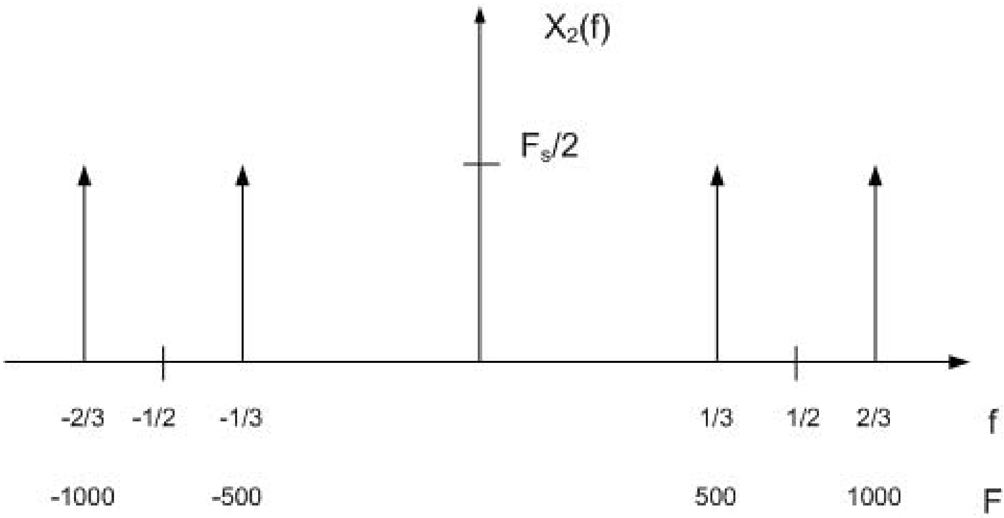

Figure 8: Spectrum of the signal $x(n)$ when $F_s = 1500$Hz

(b) Matlab-code for generating the signal corresponding to $F_s = 4000$:

```
t = [0:1/4000:1-1/4000];
cos4000 = cos(1000*2*pi*t);
```

And for the signal corresponding to $F_s = 1500$:

```
t = [0:1/1500:1-1/1500];
cos1500 = cos(1000*2*pi*t);
```

The sounds can be played with the commands:

```
sound(cos4000,4000);
pause(1);
sound(cos1500,1500);
```

They sound different because the signal incurred aliasing in the sampling. To be able to reconstruct $x_a(t)$ from a sampled signal, the sampling theorem requires that $F_s > 2F_{\max}$, where $F_{\max}$ is the highest frequency component of the signal. In this case,the signal has only one frequency component, at 1000Hz. Thus, we require:

$$F_s > 2000\text{Hz}$$