4 Following the dynamics

Now that we have at least some idea of what stochastic signals are, we can begin to analyze the behavior of systems with stochastic inputs. Let’s recall that a system is anything that transforms an input signal into an output signal. Let’s keep things simple and work with continuous-time input and output signals, and consider the case of deterministic inputs first. A system \( S \) transforms an input signal \( x : \mathbb{R} \to \mathbb{R} \) into an output signal \( y : \mathbb{R} \to \mathbb{R} \), and we denote this fact by writing \( y = S[x] \). This notation signifies the fact that the input to \( S \) is the entire signal \( x \), and the output of \( S \) is the entire signal \( y \). Thus, the value \( y(t) = S[x](t) \) of the output at time \( t \) may, in principle, depend on all \( x(s), s \in \mathbb{R} \).

Clearly, at this level of abstraction there is not a whole lot that can be done. So, it is useful to single out various types of systems:

- **causal** — when the current value of the output is not affected by the future values of the input. Formally, for any \( t \in \mathbb{R} \) and for any two inputs \( x_1 \) and \( x_2 \), such that \( x_1(s) = x_2(s) \) for all \( s \leq t \), we have
  \[
  S[x_1](t) = S[x_2](t).
  \]

- **memoryless** — when the current value of the output depends only on the current value of the input. Formally, for any \( t \in \mathbb{R} \) and for any two inputs \( x_1 \) and \( x_2 \), such that \( x_1(t) = x_2(t) \), we have
  \[
  S[x_1](t) = S[x_2](t).
  \]
  Any memoryless system is causal.

- **time-invariant** — when the output due to a time-shifted version of the input is the time-shifted version of the output. Formally, given a signal \( v : \mathbb{R} \to \mathbb{R} \) and an arbitrary \( \tau \in \mathbb{R} \), define its time shift \( v_{\tau} : \mathbb{R} \to \mathbb{R} \) by \( v_{\tau}(t) \triangleq v(t-\tau) \). Then
  \[
  S[x_{\tau}] = (S[x])_{\tau}.
  \]

- **linear** — when the output due to a superposition of inputs is the superposition of the outputs. Formally, given any two input signals \( x_1, x_2 : \mathbb{R} \to \mathbb{R} \) and any two real coefficients \( \alpha_1, \alpha_2 \),
  \[
  S[\alpha_1 x_1 + \alpha_2 x_2] = \alpha_1 S[x_1] + \alpha_2 S[x_2].
  \]

We have already seen how Markov chains can be viewed as outputs of deterministic systems driven by stochastic inputs. We will come back to this description a bit later, but first we will examine the scenario where a continuous-time stochastic signal is used as an input to a linear system, and describe the properties of the resulting output signal in terms of the properties of the input and of the system.
4.1 Linear time-invariant systems: a quick review

In ECE 210, we mostly deal with systems that are linear and time-invariant (LTI), because for such systems the relationship between the input and the output is particularly easy to write down. Each linear system is described by a function $h : \mathbb{R} \times \mathbb{R} \to \mathbb{R}$, such that, for any input signal $x$, the output is given by the superposition integral

$$S[x](t) = \int_{-\infty}^{\infty} h(t, \tau)x(\tau)d\tau \tag{4.1}$$

(as a memory refresher, prove that Eq. (4.1) specifies a linear system). The function $h$ is called the impulse response of the system because it is explicitly given by

$$h(t, \tau) = S[x](t) \quad \text{when } x(t) = \delta(t - \tau).$$

When the system is also time-invariant, $h(t, \tau) = h(t + t_0, \tau + t_0)$ for any $t_0 \in \mathbb{R}$, so, overloading the notation a bit, we can rewrite (4.1) as a convolution:

$$S[x](t) = h \ast x(t) = \int_{-\infty}^{\infty} h(t - \tau)x(\tau)d\tau. \tag{4.2}$$

While computing convolutions is an important character-building part of ECE 210, things are much easier when we pass to the frequency domain using Fourier transforms. The Fourier transform of a function $g : \mathbb{R} \to \mathbb{R}$ (whenever it exists) is defined as

$$\hat{g}(\omega) \triangleq \int_{-\infty}^{\infty} g(t)e^{-i\omega t}dt, \tag{4.3}$$

where the argument $\omega \in \mathbb{R}$ is the frequency\(^1\) and $i = \sqrt{-1}$ is the imaginary unit. To get back to the time domain, we use the Fourier inversion formula

$$g(t) = \frac{1}{2\pi} \int_{-\infty}^{\infty} \hat{g}(\omega)e^{i\omega t}d\omega. \tag{4.4}$$

The Fourier transform $\hat{h}$ of the impulse response $h$ is called the transfer function of the system, and the time-domain expression $y = h \ast x$ becomes

$$\hat{y} = H\hat{x}$$

in the frequency domain, where $\hat{x}$ is the Fourier transform of the input $x$, and $\hat{y}$ is the Fourier transform of the output $y$.

**Remark on notation.** In ECE 210, you were probably used to writing uppercase letters for Fourier transforms, like this: $X(\omega)$ instead of $\hat{x}(\omega)$. However, this will cause confusion when we start

\(^1\)Not to be confused with the generic element of the sample space $\Omega$ — hopefully, the meaning will be clear from the context.
dealing with stochastic inputs, since in that case $X$ will already denote the time-domain input. I will still write $H$ instead of $h$ because one must pay at least some respect to tradition, and because we will hardly ever use $H$ to denote stochastic signals.

So, our next order of business is to adopt this formalism to the case when the input is a stochastic signal.

### 4.2 Systems with stochastic inputs: the LTI case

Consider an LTI system with impulse response $h$ and fix a stochastic signal $X = (X_t)_{t \in \mathbb{R}}$. Then the output signal $Y = (Y_t)_{t \in \mathbb{R}}$ is related to $X$ via the convolution integral, as in (4.2):

$$Y_t = \int_{-\infty}^{\infty} h(t-\tau)X_\tau \, d\tau. \quad (4.5)$$

Of course, this is purely formal, since each $X_\tau$ is a random variable, and it takes some care to endow an integral like (4.5) with rigorous meaning. We will happily ignore all this fuss and just pretend that the above integral is well-defined. Nevertheless, it is still a random quantity, and may not admit a closed-form expression. However, as we will now see, the output mean $m_Y(t)$, the input-output crosscorrelation $R_{XY}(s,t)$, and the output autocorrelation $R_Y(s,t)$ can be expressed in terms of the input mean $m_X(t)$, the input autocorrelation $R_X(s,t)$, and the impulse response $h$.

First, let’s do the mean:

$$m_Y(t) = \mathbb{E}[Y_t] = \mathbb{E} \left[ \int_{-\infty}^{\infty} h(t-\tau)X_\tau \, d\tau \right]. \quad (4.6)$$

In general, interchanging expectations and integrals is a delicate matter, but we will be cavalier about it and just do it:

$$\mathbb{E} \left[ \int_{-\infty}^{\infty} h(t-\tau)X_\tau \, d\tau \right] = \int_{-\infty}^{\infty} \mathbb{E}[h(t-\tau)X_\tau] \, d\tau = \int_{-\infty}^{\infty} h(t-\tau)\mathbb{E}[X_\tau] \, d\tau.$$

Now, $\mathbb{E}[X_\tau] = m_X(\tau)$, so, putting everything together, we arrive at the pleasing formula

$$m_Y(t) = \int_{-\infty}^{\infty} h(t-\tau)m_X(\tau) \, d\tau = h * m_X(t). \quad (4.7)$$

That is, the output mean is given by the convolution of the input mean with the impulse response! We immediately note that if $m_X(t)$ is constant, i.e., $m_X(t) = \mu_X$ for all $t$, then $m_Y(t)$ is also constant and equal to

$$\mu_Y = \mu_X \int_{-\infty}^{\infty} h(t) \, dt.$$

Here, we assume that the integral of $h$ over the entire real line exists and is finite. We can also express this in terms of the transfer function $H$ by noting that

$$\int_{-\infty}^{\infty} h(t) \, dt = \int_{-\infty}^{\infty} h(t)e^{-i\omega t} \, dt \bigg|_{\omega=0} = H(0).$$
so we arrive at the following formula: if $m_X(t) = \mu_X$ for all $t$, then $m_Y(t) = H(0)\mu_X$.

Encouraged by our success, let’s compute the input-output crosscorrelation next:

$$R_{XY}(s, t) = \mathbb{E}[X_s Y_t] = \int_{-\infty}^{\infty} h(t - \tau) \mathbb{E}[X_s X_\tau] d\tau = \int_{-\infty}^{\infty} h(t - \tau) R_X(s, \tau) d\tau.$$ 

This almost looks like a convolution, except that $R_X$ is a function of two arguments. Note, however, that we are integrating over the second argument of $R_X$. So, we can write

$$R_{XY}(s, t) = h *_2 R_X(s, t),$$

where the subscript 2 on the asterisk indicates that we convolve only over the second argument of the input autocorrelation $R_X$:

$$h *_2 R_X(s, t) \triangleq \int_{-\infty}^{\infty} h(t - \tau) R_X(s, \tau) d\tau.$$ 

This simplifies considerably when $X$ is weakly stationary: in that case, $R_X(s, t) = R_X(t - s)$, and so

$$R_{XY}(s, t) = h *_2 R_X(s, t) = \int_{-\infty}^{\infty} h(t - \tau) R_X(s, \tau) d\tau = \int_{-\infty}^{\infty} h(t - \tau) R_X(s, \tau - s) d\tau = h * R_X(t - s)$$

is just the ordinary convolution. Now we note that $R_{XY}(s, t)$ depends only on $t - s$, so we can write $R_{XY}(s, t) = R_{XY}(t - s)$. Consequently, we arrive at the formula

$$R_{XY}(\tau) = h * R_X(\tau),$$

where $R_{XY}(\tau) = \mathbb{E}[X_t Y_{t+\tau}]$ for all $t \in \mathbb{R}$. Thus, when the input $X$ is weakly stationary, the input-output crosscorrelation is given by the convolution of the impulse response and the input autocorrelation!

Finally, let’s look at the output autocorrelation. Again, blithely interchanging the order of
expectation and integration, we have

\[ R_Y(s,t) = \mathbb{E}[Y_s Y_t] \]

\[ = \mathbb{E} \left[ \int_{-\infty}^{\infty} h(s-\tau)X_\tau d\tau \int_{-\infty}^{\infty} h(t-\tau')X_{\tau'} d\tau' \right] \]

\[ = \mathbb{E} \left[ \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} h(s-\tau)h(t-\tau')X_\tau X_{\tau'} d\tau d\tau' \right] \]

\[ = \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} h(s-\tau)h(t-\tau')X_\tau X_{\tau'} d\tau d\tau' \]

\[ = \int_{-\infty}^{\infty} h(s-\tau)h(t-\tau')X_\tau X_{\tau'} d\tau d\tau'. \]

If we perform the integration over \( \tau' \) first, we can recognize the \( *2 \) operation:

\[ \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} h(s-\tau)h(t-\tau')R_X(\tau,\tau')d\tau d\tau' = \int_{-\infty}^{\infty} h(s-\tau) \left( \int_{-\infty}^{\infty} h(t-\tau')R_X(\tau,\tau')d\tau' \right) d\tau \]

\[ = h \ast h \ast R_X(\tau,t) \]

\[ = \int_{-\infty}^{\infty} h(s-\tau)R_{XY}(\tau,t)d\tau. \]

So, if we now define the “partial convolution” \( *1 \) over the first argument, we can write

\[ R_Y(s,t) = h \ast_1 R_{XY}(s,t) = h \ast_1 (h \ast_2 R_X)(s,t). \]

This deceptively simple-looking formula hides a lot of complexity. But, once again, things simplify if \( X \) is weakly stationary. In that case, \( h \ast_2 R_X(s,t) = h \ast R_X(t-s) = R_{XY}(t-s) \), and therefore

\[ h \ast_1 (h \ast_2 R_X)(s,t) = \int_{-\infty}^{\infty} h(s-\tau)R_{XY}(\tau,t)d\tau \]

\[ = \int_{-\infty}^{\infty} h(s-\tau)R_{XY}(t-\tau)d\tau. \]  \hfill (4.9)

Again, this integral looks very much like a convolution, but here we hit a snag. In general, a convolution integral of two functions \( f \) and \( g \) will look like this:

\[ \int_{-\infty}^{\infty} f(t-\tau)g(\tau-s)d\tau, \]

i.e., if we add the arguments of \( f \) and \( g \) in the integrand, the integration variable \( \tau \) will cancel, and it is not hard to show that the integral is equal to \( f \ast g(t-s) \). But if we add up the arguments of \( h \) and \( R_{XY} \) in the integrand of (4.9), we get \( (s-\tau) + (t-\tau') = s + t - 2\tau \), and \( \tau \) most certainly does not cancel! Fortunately, there is a nice hack out of this conundrum: given \( h \), define another
function $\tilde{h}$ by setting $\tilde{h}(t) \triangleq h(-t)$. For obvious reasons, we call $\tilde{h}$ the \textit{time reversal} of $h$. Then, with this definition, we can rewrite (4.9) as
\[
\int_{-\infty}^{\infty} h(s-\tau) R_{XY}(t-\tau) d\tau = \int_{-\infty}^{\infty} \tilde{h}(\tau-s) R_{XY}(t-\tau) d\tau,
\]
and behold: $(\tau-s) + (t-\tau) = t-s$, and so we can write $h \ast_1 (h \ast_2 R_X)(s,t) = \tilde{h} \ast h \ast R_X(t-s)$. Thus, when the input $X$ is weakly stationary, the output autocorrelation is given by the double convolution:
\[
R_Y(\tau) = \tilde{h} \ast h \ast R_X(\tau), \tag{4.10}
\]
i.e., if the input $X$ is weakly stationary with autocorrelation $R_X(\tau)$, then the output $Y$ is also weakly stationary. Moreover, in that case the crosscorrelation $R_{XY}(s,t)$ also depends only on $t-s$, so we say that the input and the output are \textit{jointly weakly stationary}: each of them is WS, and $R_{XY}(t-s) = R_{XY}(t+s)$ (although be careful: $R_{XY}(t-s) = E[X_s Y_t] \neq E[X_t Y_s] = R_{XY}(t-s)$).

\subsection*{4.3 Input-output relations in the frequency domain: power spectral densities}

No one likes to compute convolution integrals (even though it builds character), so we pass to the frequency domain. Let $X$ be a WS stochastic signal with autocorrelation $R_X(\tau)$. The \textit{power spectral density} of $X$, denoted by $S_X(\omega)$, is simply the Fourier transform of $R_X$:
\[
S_X(\omega) \triangleq \widehat{R_X}(\omega) = \int_{-\infty}^{\infty} R_X(\tau) e^{-i\omega \tau} d\tau. \tag{4.11}
\]
The autocorrelation function can be recovered from the power spectral density via the Fourier inversion formula (4.4):
\[
R_X(\tau) = \frac{1}{2\pi} \int_{-\infty}^{\infty} S_X(\omega) e^{i\omega \tau} d\omega. \tag{4.12}
\]
The word “spectral” evokes frequency content, so this makes sense. But what do “power” and “density” mean? If we think about $X_t$ as a (random) current passing through a unit resistance, then the voltage across the resistor is also equal to $X_t$, by Ohm’s law. Consequently, the dissipated power at time $t$ is equal to
\[
\text{(voltage at time } t) \cdot \text{(current at time } t) = X_t^2.
\]
Thus, we may think of $E[X_t^2]$ as the \textit{average power} dissipated at time $t$. Now, if $X$ is WS, then $E[X_t^2] = E[X_t X_t] = R_X(0)$. On the other hand, if we substitute $\tau = 0$ into the Fourier inversion formula (4.12), we get
\[
E[X_t^2] = R_X(0) = \frac{1}{2\pi} \int_{-\infty}^{\infty} S_X(\omega) d\omega. \tag{4.13}
\]
Similarly, given two jointly WS stochastic signals $X$ and $Y$, we define their \textit{cross-power spectral density} $S_{XY}$ as the Fourier transform of the crosscorrelation function $R_{XY}$:
\[
S_{XY}(\omega) = \widehat{R_{XY}}(\omega) = \int_{-\infty}^{\infty} R_{XY}(\tau) e^{i\omega \tau} d\tau.
\]
Using power spectral densities, we can obtain frequency-domain forms of the results of the preceding section. For example, taking the Fourier transform of both sides of (4.8), we get

\[ S_{XY}(\omega) = H(\omega)S_X(\omega). \]  

(4.14)

Similarly, taking the Fourier transform of both sides of (4.10), we get

\[ S_Y(\omega) = \tilde{H}(\omega)H(\omega)S_X(\omega), \]  

(4.15)

where \( \tilde{H} \) is the Fourier transform of the time-reversed impulse response \( \tilde{h} \). However, we can express \( \tilde{H} \) in terms of \( H \):

\[
\tilde{H}(\omega) = \int_{-\infty}^{\infty} \tilde{h}(\tau)e^{-i\omega\tau}d\tau \\
= \int_{-\infty}^{\infty} h(-\tau)e^{-i\omega\tau}d\tau \\
= \int_{-\infty}^{\infty} h(\tau)e^{i\omega\tau}d\tau,
\]

where in the last line we have made the change of variable \( \tau \rightarrow -\tau \). Now, since \( h \) is real-valued, we see that \( \tilde{H}(\omega) \) is just the complex conjugate of \( H(\omega) \): \( \tilde{H}(\omega) = H^*(\omega) \). Therefore, (4.15) simplifies to

\[ S_Y(\omega) = |H(\omega)|^2 S_X(\omega). \]  

(4.16)

Thus, when a WS stochastic signal is used as an input to an LTI system, the effect in the frequency domain is to reshape the power spectral density in proportion to \( |H|^2 \).

**Example: white and colored noise.** Consider a zero-mean WS stochastic signal \( Z \) with the flat power spectral density \( S_Z(\omega) = q \), where \( q > 0 \) is some fixed constant. By analogy with white visible light that contains all spectral components, we call such a signal **white noise** with intensity \( q \). The autocorrelation function is then given by \( R_Z(\tau) = q\delta(\tau) \). Since \( Z \) is zero-mean,

\[ R_Z(\tau) = C_Z(\tau) = \mathbb{E}[Z_tZ_{t+\tau}] \]

for any \( t, \tau \in \mathbb{R} \). In this case, \( R_Z(\tau) = 0 \) for all \( \tau > 0 \), i.e., \( Z_t \) and \( Z_{t'} \) are uncorrelated for all \( t \neq t' \). Although white noise is a useful mathematical construction (for example, many stochastic signals of interest can be interpreted as outputs of deterministic systems driven by white-noise inputs), it is unphysical. To see why, let us substitute \( S_Z(\omega) = q \) into (4.13) to get

\[ \mathbb{E}[Z_t^2] = \frac{1}{2\pi} \int_{-\infty}^{\infty} q \, d\omega = +\infty. \]

That is, pure white noise has infinite average power! Of course, we can simply continue using it while remaining mindful of its unphysical nature (just like we do with the unit impulse), or we can develop various approximations. For example, it is often possible to consider stochastic signals
whose power spectra are flat and nonzero only in some band of frequencies. Thus, consider a WS stochastic signal $Z$ with

$$S_Z(\omega) = \begin{cases} q, & -\omega_0 \leq \omega \leq \omega_0 \\ 0, & \text{otherwise} \end{cases},$$

(4.17)

where $q > 0$ is the intensity parameter and $\omega_0 > 0$ is some cutoff frequency. Then

$$\mathbb{E}[Z_t^2] = \frac{1}{2\pi} \int_{-\omega_0}^{\omega_0} q \, d\omega = \frac{q\omega_0}{\pi},$$

which is finite. We can also compute the autocorrelation function of $Z$:

$$R_Z(\tau) = \frac{1}{2\pi} \int_{-\omega_0}^{\omega_0} S_Z(\omega) e^{i\omega \tau} \, d\omega$$

$$= \frac{q}{2\pi} \int_{-\omega_0}^{\omega_0} e^{i\omega \tau} \, d\omega$$

$$= \frac{q}{2\pi} \cdot \frac{1}{i\tau} \bigg|_{-\omega_0}^{\omega_0}$$

$$= \frac{q}{\pi\tau} \cdot \frac{e^{i\omega_0 \tau} - e^{-i\omega_0 \tau}}{2i}$$

$$= \frac{q \sin \omega_0 \tau}{\pi \tau}.$$

Recalling the definition of the sinc function\(^2\)

$$\text{sinc}(u) = \begin{cases} \frac{\sin \pi u}{\pi u}, & u \neq 0 \\ 1, & u = 0 \end{cases}$$

we can write

$$R_Z(\tau) = 2qf_0 \text{sinc}(2f_0 \tau),$$

(4.18)

where $f_0 = \omega_0/2\pi$ is the cutoff frequency in Hz. A plot of $R_Z$ with $q = 5$ and $\omega_0 = 4$ rad/s is shown in Fig. 1. Note that because $|R_Z(\tau)| > 0$ for all $\tau$, there are correlations between $Z_t$ and $Z_{t+\tau}$, but they decay to zero as $|\tau| \to \infty$.

The term “colored noise” refers to stochastic signals whose power spectra are roughly shaped like the corresponding colors of the visible spectrum. For example, the term “pink noise” refers to power spectra of the form $S(\omega) \propto 1/\omega$. Later on, we will return to this noise model when discussing flicker noise in electronic devices.

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\(^2\)This is the engineer’s sinc function. The mathematician’s sinc function is $\text{sinc}(u) = \frac{\sin u}{u}$. 

8
4.3.1 Properties of power spectral densities

The power spectral density $S_X$ of any WS stochastic signal $X = (X_t)_{t \in \mathbb{R}}$ is real (i.e., $S_X(\omega) \in \mathbb{R}$ for all $\omega$), even (i.e., $S_X(-\omega) = S_X(\omega)$ for all $\omega$), and nonnegative (i.e., $S_X(\omega) \geq 0$ for all $\omega$).

Indeed, recall that $S_X$ is the Fourier transform of the autocorrelation function $R_X$. The autocorrelation function $R_X$ takes real values, and it is also even:

$$R_X(-\tau) = \mathbb{E}[X_tX_{t-\tau}] = \mathbb{E}[X_{t+\tau}X_t] = R_X(\tau).$$

Therefore, using Euler’s formula, we have

$$S_X(\omega) = \int_{-\infty}^{\infty} R_X(\tau) e^{-i\omega \tau} d\tau = \int_{-\infty}^{\infty} R_X(\tau) \cos \omega \tau d\tau + i \int_{-\infty}^{\infty} R_X(\tau) \sin \omega \tau d\tau.$$

Since $R_X(\tau)$ is even and $\sin \omega \tau$ is odd, the second integral is identically zero, and we obtain the formula

$$S_X(\omega) = \int_{-\infty}^{\infty} R_X(\tau) \cos \omega \tau d\tau. \quad (4.19)$$

Since the integrand takes real values, $S_X(\omega) \in \mathbb{R}$ for all $\omega$. Moreover, the integrand is even as a
function of $\omega$, so $S_X$ is also an even function. This also implies that $R_X$ is given by

$$R_X(\tau) = \frac{1}{2\pi} \int_{-\infty}^{\infty} S_X(\omega) e^{i\omega \tau} d\omega$$

$$= \frac{1}{2\pi} \int_{-\infty}^{\infty} S_X(\omega) \cos \omega \tau d\omega + \frac{i}{2\pi} \int_{-\infty}^{\infty} S_X(\omega) \sin \omega \tau d\omega$$

$$= \frac{1}{2\pi} \int_{-\infty}^{\infty} S_X(\omega) \cos \omega \tau d\omega,$$

where we have used the same reasoning as before: $S_X(\omega) \sin \omega \tau$ is an odd function of $\omega$, so the second integral is zero.

It remains to show that $S_X(\omega) \geq 0$ for all $\omega$. Let $X$ be the input to a bandpass filter with the transfer function

$$H(\omega) = \begin{cases} 1, & \omega_0 \leq |\omega| \leq \omega_1 \\ 0, & \text{otherwise} \end{cases},$$

where $\omega_0 \leq \omega_1$ are arbitrary nonnegative constants. Let $Y$ denote the output stochastic signal. Then

$$S_Y(\omega) = |H(\omega)|^2 S_X(\omega) = \begin{cases} S_X(\omega), & \omega_0 \leq |\omega| \leq \omega_1 \\ 0, & \text{otherwise} \end{cases}.$$

Now, using the Fourier inversion formula and the evenness of $S_X$,

$$R_Y(0) = \frac{1}{2\pi} \int_{-\infty}^{\infty} S_Y(\omega) d\omega$$

$$= \frac{1}{2\pi} \int_{-\omega_1}^{\omega_1} S_X(\omega) d\omega + \frac{1}{2\pi} \int_{\omega_0}^{\omega_1} S_X(\omega) d\omega$$

$$= \frac{1}{\pi} \int_{\omega_0}^{\omega_1} S_X(\omega) d\omega.$$

Since $R_Y(0) = \mathbb{E}[Y_t^2] \geq 0$, we see that

$$\int_{\omega_0}^{\omega_1} S_X(\omega) d\omega \geq 0, \quad \forall \omega_0, \omega_1. \quad (4.20)$$

The only way for (4.20) to hold is if $S_X(\omega) \geq 0$ for all $\omega$.

In fact, for any function $S$ of frequency $\omega$ which is real, even, nonnegative, and satisfies the condition

$$a \triangleq \int_{-\infty}^{\infty} S(\omega) d\omega < \infty,$$

we can construct a WS stochastic signal $X$, such that $S = S_X$. To prove this, consider the following stochastic signal:

$$X_t = \cos(\Omega t + \Theta),$$

$$Y_t = \sin(\Omega t + \Theta).$$
where $\Omega$ and $\Theta$ are two mutually independent random variables. Let us assume that $a = \pi$ (otherwise, we can simply rescale $S$). We will take $\Theta \sim \text{Uniform}(0, 2\pi)$, and we will choose the pdf $f_\Omega$ of $\Omega$ later. First, let us prove that $X$ is WS. For that, we have to compute its mean and autocorrelation functions. Since $\Omega$ and $\Theta$ are independent, we can write

$$m_X(t) = \mathbb{E}[X_t] = \mathbb{E}[\cos(\Omega t + \Theta)]$$

$$= \frac{1}{2\pi} \int_{-\infty}^{\infty} \int_{0}^{2\pi} f_\Omega(\omega) \cos(\omega t + \theta) d\theta d\omega$$

$$= \frac{1}{2\pi} \int_{-\infty}^{\infty} f_\Omega(\omega) \left( \int_{0}^{2\pi} \cos(\omega t + \theta) d\theta \right) d\omega.$$

The integral in parentheses is equal to zero, so $m_X(t) = 0$ for all $t$. For the autocorrelation function, using the trigonometric identity $\cos u \cos v = \frac{1}{2} [\cos(u + v) + \cos(u - v)]$, we have

$$R_X(t, t + \tau) = \mathbb{E}[X_t X_{t+\tau}] = \mathbb{E}[\cos(\Omega t + \Theta) \cos(\Omega(t + \tau) + \Theta)]$$

$$= \frac{1}{2} \mathbb{E}[\cos(2\Omega t + \Omega \tau + \Theta) + \cos(\Omega \tau)]$$

$$= \frac{1}{2} \mathbb{E}[\cos(2\Omega t + \Omega \tau + \Theta)] + \frac{1}{2} \mathbb{E}[\cos \Omega \tau].$$

Using the same reasoning as in the derivation of $m_X$, we see that the first expectation is identically zero, and therefore

$$R_X(t, t + \tau) = \frac{1}{2} \mathbb{E}[\cos \Omega \tau]. \quad (4.21)$$

Since $m_X = 0$ and $R_X(t, t + \tau)$ depends only on $\tau$, $X$ is indeed WS. Now we will pick $f_\Omega$ to guarantee that $S_X = S$. To that end, we will use the Fourier inversion formula

$$R_X(\tau) = \frac{1}{2\pi} \int_{-\infty}^{\infty} S_X(\omega) \cos(\omega \tau) d\omega$$

(recall that $S_X$ is even). On the other hand, from (4.21) we have

$$R_X(\tau) = \frac{1}{2} \int_{-\infty}^{\infty} f_\Omega(\omega) \cos(\omega \tau) d\omega.$$

In particular,

$$R_X(0) = \frac{1}{2} = \frac{1}{2\pi} \int_{-\infty}^{\infty} S_X(\omega) d\omega$$

Let us take $f_\Omega(\omega) = \frac{S(\omega)}{\pi}$. This function is nonnegative and integrates to 1. Therefore, it is a valid pdf. By uniqueness of Fourier transforms, we have ensured that $S = \widehat{R_X}$, and therefore $S_X = S$. 

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4.4 Gaussian stochastic signals and Bussgang's theorem

An important class of stochastic signals consists of Gaussian stochastic signals (or Gaussian processes). To define a Gaussian stochastic signal, we first recall the definition of a Gaussian random variable: A random variable $X$ is called Gaussian if its characteristic function $\Phi_X(u) = \mathbb{E}[e^{iuX}]$ has the form

$$\Phi_X(u) = \exp\left(i\mu u - \frac{u^2\sigma^2}{2}\right)$$

(4.22)

for some $\mu \in \mathbb{R}$ and $\sigma^2 \geq 0$. It can be shown then that $\mu = \mathbb{E}[X]$ and $\sigma^2 = \text{Var}[X]$ (in the degenerate case $\sigma = 0$, $X = \mu$ with probability one). When $\sigma > 0$, the pdf of $X$ is obtained by taking the inverse Fourier transform of $\Phi_X$:

$$f_X(x) = \frac{1}{2\pi} \int_{-\infty}^{\infty} \Phi_X(u)e^{-iuX}du$$

$$= \frac{1}{\sqrt{2\pi\sigma^2}} \exp\left(-\frac{(x-\mu)^2}{2\sigma^2}\right).$$

When $\sigma = 0$, $X$ does not have a pdf.

Now consider a random vector, i.e., a stochastic signal of the form $X = (X_t)_{t \in T}$ with $T = \{1, \ldots, n\}$. We can represent such signals by column vectors, as in

$$X = \begin{bmatrix} X_1 \\ X_2 \\ \vdots \\ X_n \end{bmatrix}.$$

In the same way, we can represent the mean and the autocovariance functions of $X$:

$$m_X = \begin{bmatrix} m_X(1) \\ m_X(2) \\ \vdots \\ m_X(n) \end{bmatrix} \quad \text{and} \quad C_X = \begin{bmatrix} C_X(1,1) & C_X(1,2) & \cdots & C_X(1,n) \\ C_X(2,1) & C_X(2,2) & \cdots & C_X(2,n) \\ \vdots & \vdots & \ddots & \vdots \\ C_X(n,1) & C_X(n,2) & \cdots & C_X(n,n) \end{bmatrix},$$

where, as usual, $M_X(t) = \mathbb{E}[X_t]$ and $C_X(s,t) = \mathbb{E}[X_sX_t] - \mathbb{E}[X_s]\mathbb{E}[X_t]$. The $n \times n$ covariance matrix $C_X$ is symmetric, since $C_X(s,t) = C_X(t,s)$, and the diagonal entries are equal to the variances of the $X_t$'s:

$$C_X(t,t) = \mathbb{E}[X_t^2] - (\mathbb{E}[X_t])^2 = \text{Var}[X_t].$$

Now let $u = (u_1, \ldots, u_n)^T$ be a fixed deterministic vector, and consider the scalar random variable $Z = u^T X = u_1X_1 + \ldots + u_nX_n$. Geometrically, $Z$ is the projection of the random vector $X$ onto the deterministic vector $u$. We can easily compute its mean and variance:

$$\mathbb{E}[Z] = \mathbb{E}\left[\sum_{t=1}^{n} u_tX_t\right] = \sum_{t=1}^{n} u_t \mathbb{E}[X_t] = \sum_{t=1}^{n} u_t m_X(t) = u^T m_X$$
and

$$\text{Var}[Z] = \mathbb{E}[Z^2] - (\mathbb{E}Z)^2$$

$$= \sum_{s=1}^{n} \sum_{t=1}^{n} (\mathbb{E}[u_s u_t X_s X_t] - u_s u_t \mathbb{E}[X_s] \mathbb{E}[X_t])$$

$$= \sum_{s=1}^{n} \sum_{t=1}^{n} u_s u_t (\mathbb{E}[X_s X_t] - \mathbb{E}[X_s] \mathbb{E}[X_t])$$

$$= \sum_{s=1}^{n} \sum_{t=1}^{n} u_s u_t C_X(s,t)$$

$$= u^T C_X u.$$

We now introduce the following definition: we say that $X$ is a Gaussian random vector (or that $X_1, \ldots, X_n$ are jointly Gaussian random variables) if the scalar random variable $Z = u^T X$ is Gaussian for each choice of the vector of coefficients $u \in \mathbb{R}^n$. In light of our definition of a Gaussian random variable, this is equivalent to the following: $X$ is a Gaussian random vector if, for any $u \in \mathbb{R}^n$, the characteristic function $\Phi_Z = \Phi_{u^T X}$ of $Z = u^T X$ takes the form

$$\Phi_Z(\alpha) = \exp\left(ia\mathbb{E}[Z] - \frac{a^2 \text{Var}[Z]}{2}\right)$$

$$= \exp\left(iau^T m_X - \frac{a^2 u^T C_X u}{2}\right)$$

for all $a \in \mathbb{R}$. Note that

$$\Phi_{u^T X}(\alpha) = \mathbb{E}[e^{iau^T X}] = \mathbb{E}[e^{i(a \alpha)^T X}] = \Phi_{a \alpha^T X}(1),$$

so specifying the characteristic functions of all linear combinations $u^T X$ is equivalent to specifying the joint characteristic function of $X$, which is a function of a vector argument $u \in \mathbb{R}^n$ and takes the form

$$\Phi_X(u) \equiv \Phi_{X_1, \ldots, X_n}(u_1, \ldots, u_n) \equiv \mathbb{E}[e^{iu^T X}]. \quad (4.23)$$

Just like the characteristic function of a scalar random variable $X$ uniquely specifies the distribution of $X$, the joint characteristic function of a random vector $X = (X_1, \ldots, X_n)^T$ uniquely specifies the joint distribution of its coordinates $X_1, \ldots, X_n$. In particular, if the $X_i$’s have a joint pdf, then it is given by the multidimensional Fourier inversion formula

$$f_X(x) = f_{X_1, \ldots, X_n}(x_1, \ldots, x_n) = \frac{1}{(2\pi)^n} \int_{-\infty}^{\infty} \cdots \int_{-\infty}^{\infty} \Phi_X(u)e^{-iu^T X} du_1 \cdots du_n. \quad (4.24)$$

Because $\Phi_X(u) = \Phi_{u^T X}(1)$, we see that if we know the probability distributions of all of its projections $Z = u^T X$, $u \in \mathbb{R}^n$, then we know the joint probability distribution of $X_1, \ldots, X_n$ (this is referred to
as the Cramèr–Wold theorem). Consequently, a random vector \( X \) is Gaussian if and only if its joint characteristic function takes the form

\[
\Phi_X(u) = \exp \left( iu^T m_X - \frac{1}{2} u^T C_X u \right), \quad u \in \mathbb{R}^n. \tag{4.25}
\]

Along the way, we have established a nice fact: the projection of a Gaussian random vector onto a deterministic vector is always Gaussian.

**Important!** You should always keep in mind the crucial distinction between a Gaussian random vector and a vector of Gaussian random variables. It is easy to construct examples of random vectors \( X = (X_1, \ldots, X_n)^T \), where each \( X_i \) is Gaussian, yet \( X_1, \ldots, X_n \) are not jointly Gaussian (you will see such an example in the homework). If \( X_1, \ldots, X_n \) are not jointly Gaussian, then the projection \( u^T X \) may not be Gaussian.

We are now ready to define a general Gaussian stochastic signal: a stochastic signal \( X = (X_t)_{t \in T} \) is Gaussian if, for any \( n \in \mathbb{N} \) and any \( t_1, \ldots, t_n \in T \), the random variables \( X_{t_1}, \ldots, X_{t_n} \) are jointly Gaussian. That is, their joint characteristic function has the form

\[
\Phi_{X_{t_1}, \ldots, X_{t_n}}(u_1, \ldots, u_n) = \exp \left( i \sum_{k=1}^n u_k m_X(t_k) - \frac{1}{2} \sum_{j=1}^n \sum_{k=1}^n u_j u_k C_X(t_j, t_k) \right). \tag{4.26}
\]

This implies that a weakly stationary Gaussian stochastic signal is, in fact, strongly stationary. Indeed, suppose that \( X \) is WS. Then \( \mu_X(t) = \mu \) for all \( t \), and \( C_X(s, t) = C_X(t - s) \) for all \( s, t \). Using these facts in (4.26), we obtain

\[
\Phi_{X_{t_1}, \ldots, X_{t_n}}(u_1, \ldots, u_n) = \exp \left( i \mu \sum_{k=1}^n u_k - \frac{1}{2} \sum_{j=1}^n \sum_{k=1}^n u_j u_k C_X(t_k - t_j) \right). \tag{4.27}
\]

Now fix an arbitrary \( \tau \in \mathbb{R} \) and consider the random variables \( X_{t_1 + \tau}, \ldots, X_{t_n + \tau} \). Since \( \mu_X(t) \) is constant, and since \( C_X(s + \tau, t + \tau) = C_X(t - s) \) their joint characteristic function is also given by (4.27):

\[
\Phi_{X_{t_1 + \tau}, \ldots, X_{t_n + \tau}}(u_1, \ldots, u_n) = \exp \left( i \mu \sum_{k=1}^n u_k - \frac{1}{2} \sum_{j=1}^n \sum_{k=1}^n u_j u_k C_X(t_k - t_j) \right). \tag{4.28}
\]

Thus, for any choice of \( t_1, \ldots, t_n \) and \( \tau \), \( (X_{t_1}, \ldots, X_{t_n}) \) and \( (X_{t_1 + \tau}, \ldots, X_{t_n + \tau}) \) have the same joint characteristic function and hence the same distribution. Therefore \( X \) is strongly stationary. For this reason, we can just say “stationary” when dealing with Gaussian stochastic signals.

Now that we have collected some background about Gaussian stochastic signals, we can state and prove a remarkable result called Bussgang’s theorem.\(^\text{3}\) Let \( X \) be a stationary Gaussian stochastic

signal with zero mean, and let $Y$ be obtained by passing $X$ through a memoryless nonlinearity. That is, for each $t \in \mathbb{R}$, we have $Y_t = g(X_t)$, where $g : \mathbb{R} \to \mathbb{R}$ is a fixed transformation. Suppose that $g$ is differentiable\footnote{We understand differentiability in the generalized sense and allow functions like the unit step or the unit ramp.} and has moderate growth at infinity, i.e.,

$$
\lim_{|x| \to \infty} g(x)e^{-x^2} = 0. \tag{4.29}
$$

For example, any polynomial in $x$ would do. Bussgang’s theorem gives the following expression for the crosscorrelation function between $X$ and $Y$:

$$
R_{XY}(t, t + \tau) = E[X_t Y_{t+1}] = K \cdot R_X(\tau), \tag{4.30}
$$

where

$$
K = E[g'(X_0)].
$$

Since $R_{XY}(t, t + \tau)$ depends only on $\tau$, we can write $R_{XY}(\tau) = KR_X(\tau)$ instead. The remarkable thing about (4.30) is its simplicity, despite the fact that $g$ may be highly nonlinear and the output stochastic signal $Y$ may not even be Gaussian. Before giving the proof of (4.30), let us take a look at a couple of examples.

**Hard limiter.** Consider the following nonlinear transformation:

$$
g(x) = \begin{cases} 
-1, & x < 0 \\
0, & x = 0 \\
1, & x > 0
\end{cases}
$$

We can write $g$ as a difference of two unit-step functions: $g(x) = u(x) - u(-x)$. Therefore, $g'(x) = 2\delta(x)$. Let $X$ be a stationary Gaussian stochastic signal, and let $f_0$ denote the pdf of $X_0$. Then

$$
f_0(x) = \frac{1}{\sqrt{2\pi R_X(0)}} e^{-x^2/2R_X(0)},
$$

and therefore

$$
E[g'(X_0)] = 2 \int_{-\infty}^{\infty} f_0(x) \delta(x) dx = 2f_0(0) = \sqrt{\frac{2}{\pi R_X(0)}}.
$$

Hence, Bussgang’s theorem gives

$$
R_{XY}(\tau) = \sqrt{\frac{2}{\pi R_X(0)}} R_X(\tau).
$$
Half-wave rectifier. Consider 

\[ g(x) = \max\{x, 0\} = \begin{cases} \ 0, & x < 0 \\ x, & x \geq 0 \end{cases}. \]

Then \( g'(x) = u(x) \), so for any zero-mean Gaussian random variable \( X \) we have 

\[ E[g'(X)] = E[u(X)] = \int_{0}^{\infty} f_X(x)dx = \frac{1}{2}. \]

Consequently, Bussgang's theorem gives 

\[ R_{XY}(\tau) = \frac{1}{2} R_X(\tau). \]

Now we give the proof of Bussgang's theorem. First, we note that it is equivalent to the following statement: Let \( U \) and \( V \) be two jointly Gaussian random variables with zero mean. Then for any \( g \) satisfying (4.29) we have 

\[ E[U g(V)] = \text{Cov}(U, V) E[g'(V)]. \quad (4.31) \]

We obtain (4.30) by applying (4.31) to \( U = X \) and \( V = X_{t+\tau} \). Since \( X \) is zero-mean, \( \text{Cov}(X_t, X_{t+\tau}) = C_X(\tau) = R_X(\tau) \), and, since \( X \) is stationary, \( E[g'(X_{t+\tau})] = E[g'(X_0)] \). Thus, we proceed to prove (4.31). We will derive it from another important result, known as Stein's identity: Let \( g \) be a differentiable function satisfying (4.29), and let \( U \) be a zero-mean Gaussian random variable with variance \( \sigma^2 \). Then 

\[ E[U g(U)] = \sigma^2 E[g'(U)]. \quad (4.32) \]

Remark on Stein's identity. The importance of Stein's identity (4.32) lies in the fact that it actually fully characterizes the Gaussian distribution: It can be proved that any zero-mean, unit-variance random variable \( U \), such that 

\[ E[U g(U)] = E[g'(U)] \]

holds for all smooth \( g \), is Gaussian. This can be pushed further to imply that any random variable \( U \) with zero mean and unit variance, such that 

\[ E[U g(U) - g'(U)] \approx 0 \]

for all smooth \( g \), is “close to Gaussian.”

To prove (4.32), we note the following about the pdf of \( U \):

\[ f'_U(u) = \frac{1}{\sqrt{2\pi\sigma^2}} \frac{d}{du} e^{-u^2/2\sigma^2} = -\frac{u}{\sigma^2} f_U(u). \]

Therefore,

\[ E[U g(U)] = \int_{-\infty}^{\infty} u g(u) f_U(u)du \]

\[ = -\sigma^2 \int_{-\infty}^{\infty} g(u) f'_U(u)du. \]
Integrating by parts, we get
\[ E[Ug(U)] = -\sigma^2 g(u)f_U(u)\bigg|_{-\infty}^{\infty} + \sigma^2 \int_{-\infty}^{\infty} g'(u)f_U(u)du. \]

Since \( g \) satisfies (4.29), the first term on the right-hand side is identically zero, and we obtain
\[ E[Ug(U)] = 2 \int_{-\infty}^{\infty} g'(u)f_U(u)du = \sigma^2 E[g'(U)]. \]

Now we will use Stein’s identity to prove (4.31). To that end, we first use a trick which you can think of as “Gram-Schmidt orthogonalization for Gaussians:” Given the pair \((U, V)\), define the random variable
\[ \tilde{V} = V - \frac{c}{\sigma_U^2} U, \]
where \( c = \text{Cov}(U, V) = E[UV] \) and \( \sigma_U^2 = \text{Var}[U] \). Since \( U \) and \( V \) are jointly Gaussian with zero mean, \( \tilde{V} \) is also Gaussian with zero mean. Moreover, \( U \) and \( \tilde{V} \) are uncorrelated:
\[ E[U \tilde{V}] = E\left[U\left(V - \frac{c}{\sigma_U^2} U\right)\right] = E[UV] - \frac{c}{\sigma_U^2} E[U^2] = 0. \]

As you will prove in the homework, if two jointly Gaussian random variables are uncorrelated, then they are also independent. Therefore, \( U \) and \( \tilde{V} \) are independent. Using this fact, we can write
\[ E[Ug(V)] = E\left[Ug\left(\tilde{V} + \frac{c}{\sigma_U^2} U\right)\right] = \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} ug\left(\tilde{v} + \frac{c}{\sigma_U^2} u\right)f_U(u)f_{\tilde{V}}(\tilde{v})du d\tilde{v}. \]

We will first integrate over \( u \), and then over \( \tilde{v} \):
\[ \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} ug\left(\tilde{v} + \frac{c}{\sigma_U^2} u\right)f_U(u)f_{\tilde{V}}(\tilde{v})du d\tilde{v} = \int_{-\infty}^{\infty} f_{\tilde{V}}(\tilde{v}) \left[ \int_{-\infty}^{\infty} ug\left(\tilde{v} + \frac{c}{\sigma_U^2} u\right)f_U(u)du \right] d\tilde{v} = \int_{-\infty}^{\infty} f_{\tilde{V}}(\tilde{v}) E\left[Ug\left(\tilde{v} + \frac{c}{\sigma_U^2} U\right)\right] d\tilde{v}. \]

Using Stein’s identity and the chain rule for differentiation, we can rewrite the expectation in (4.33) as follows:
\[ E\left[Ug\left(\tilde{v} + \frac{c}{\sigma_U^2} U\right)\right] = \sigma_U^2 \cdot \frac{c}{\sigma_U^2} E\left[g'\left(\tilde{v} + \frac{c}{\sigma_U^2} U\right)\right] = cE\left[g'\left(\tilde{v} + \frac{c}{\sigma_U^2} U\right)\right]. \]
Substituting this expression back into (4.33), we get
\[
\mathbb{E}[Ug(V)] = c \int_{-\infty}^{\infty} f_{\tilde{V}}(\tilde{v}) \mathbb{E} \left[ g \left( \tilde{v} + \frac{c}{\sigma_U^2} U \right) \right] d\tilde{v}
\]
\[
= c \mathbb{E} \left[ g' \left( \tilde{V} + \frac{c}{\sigma_U^2} U \right) \right]
\]
\[
= c \mathbb{E}[g'(V)]
\]
\[
= \text{Cov}(U, V) \cdot \mathbb{E}[g'(V)],
\]
which proves (4.31) and thus Bussgang’s theorem.

### 4.5 Poisson point processes and Campbell’s theorem

Gaussian processes provide a basic model of stochastic phenomena involving continuously fluctuating quantities. Poisson point processes, on the other hand, serve as a good model of phenomena involving discrete events occurring at random times. Recall that a Poisson point process with rate \( \lambda \) is a discrete-time stochastic signal \((T_k)_{k \in \mathbb{N}}\), where the random arrival (or event occurrence) times \( 0 < T_1 < T_2 < T_3 < \ldots \) are such that the interarrival times \( Z_k = T_k - T_{k-1} \) (with the initial condition \( T_0 = 0 \)) are i.i.d. \( \text{Exp}(\lambda) \) random variables. Alternatively, if for all \( 0 \leq a \leq b \) we let
\[
N(a, b) \triangleq \#(\text{points } T_k \text{ in the interval } (a, b)),
\]
then \( N(a, b) \sim \text{Pois}(\lambda(b - a)) \), and \( N(a, b) \) is independent of \( N(c, d) \) if \( (a, b) \cap (c, d) = \emptyset \). Moreover, for any \( r \in \mathbb{R} \) such that \( a + r \geq 0 \) and \( b + r \geq 0 \), \( N(a + r, b + r) \) has the same distribution as \( N(a, b) \). This is just another way of saying that the Poisson counting process
\[
N_t = N(0, t) = \sum_{k=1}^{\infty} u(t - T_k), \quad (4.34)
\]
where \( u(\cdot) \) is the unit step function, has independent and stationary increments.

We can also associate each arrival with a unit impulse and form the stochastic signal
\[
X_t = \sum_{i=1}^{\infty} \delta(t - T_i). \quad (4.35)
\]
If \( X = (X_t)_{t \geq 0} \) is an input to an LTI system with impulse response \( h \), then the corresponding output is given by
\[
Y_t = \sum_{i=1}^{\infty} h(t - T_i). \quad (4.36)
\]
This type of stochastic signal is a model of \textit{shot noise} in electronic and optical devices, which is a manifestation of the discrete nature of charge carriers or photons. In that context, \( h \) describes the
response of the device to individual arrivals. We will look at shot noise in more detail later; for now, we will state and prove a useful result, known as *Campbell’s theorem*. The shot noise process $Y$ has the mean

$$m_Y = \lambda \int_{-\infty}^{\infty} h(\tau) d\tau = \lambda H(0) \quad (4.37)$$

and the variance

$$\text{Var}[Y_t] = \lambda \int_{-\infty}^{\infty} h^2(\tau) d\tau. \quad (4.38)$$

Let us first prove (4.37). Since

$$Y_t = \int_{-\infty}^{\infty} h(t-\tau) X_\tau d\tau, \quad (4.39)$$

we can apply the formula (4.6):

$$E[Y_t] = E\left[\int_{-\infty}^{\infty} h(t-\tau) X_\tau d\tau\right].$$

Interchanging the integral and the expectation, we can write

$$E[Y_t] = \int_{-\infty}^{\infty} E[X_\tau] d\tau, \quad (4.40)$$

so now we need to compute the mean $m_X(\tau)$. Since $\frac{d}{dt} u(t) = \delta(t)$, we can view the stochastic signal $X$ as the result of passing the Poisson counting process $N$ through a differentiator:

$$X_t = \frac{d}{dt} N_t = \frac{d}{dt} \sum_{i=1}^{\infty} u(t-T_i). \quad (4.41)$$

Since $E[N_t] = \lambda t$, we have

$$E[X_t] = E\left[\frac{d}{dt} N_t\right] = \frac{d}{dt} E[N_t] = \lambda$$

(once again, we have been somewhat cavalier about interchanging the order of expectation and differentiation, but all of these steps can be made rigorous using tools from the theory of point processes). Substituting this into (4.40), we get (4.37). Now we prove (4.38). Using the definition of variance and (4.37), we have

$$\text{Var}[Y_t] = E[Y_t^2] - (E[Y_t])^2$$

$$= E[Y_t^2] - \lambda^2 \left(\int_{-\infty}^{\infty} h(\tau) d\tau\right)^2, \quad (4.42)$$
so now we need to compute $E[Y_t^2]$. Using (4.39), we can write

$$E[Y_t^2] = E\left[\left(\int_{-\infty}^{\infty} h(t-\tau)X_{\tau}d\tau\right)\left(\int_{-\infty}^{\infty} h(t-\tau')X_{\tau'}d\tau'\right)\right]$$

$$= E\left[\int_{-\infty}^{\infty} \int_{-\infty}^{\infty} h(t-\tau)h(t-\tau')X_{\tau}X_{\tau'}d\tau d\tau'\right]$$

$$= \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} h(t-\tau)h(t-\tau')E[X_{\tau}X_{\tau'}]d\tau d\tau'. \quad (4.43)$$

Once again, we use (4.41):

$$E[X_{\tau}X_{\tau'}] = E\left[\frac{\partial^2}{\partial \tau\partial \tau'} N_{\tau}N_{\tau'}\right]$$

$$= \frac{\partial^2}{\partial \tau\partial \tau'} E[N_{\tau}N_{\tau'}]$$

$$= \frac{\partial^2}{\partial \tau\partial \tau'} R_N(\tau,\tau'). \quad (4.44)$$

The autocorrelation function of the Poisson process with rate $\lambda$ is given by

$$R_N(s,t) = \lambda^2 st + \lambda \min\{s,t\}$$

(exercise: verify this!), and therefore

$$\frac{\partial^2}{\partial s\partial t} R_N(s,t) = \frac{\partial^2}{\partial s\partial t} \left(\lambda^2 st + \lambda \min\{s,t\}\right)$$

$$= \frac{\partial}{\partial s} \left(\lambda^2 s + \lambda \frac{\partial}{\partial t} \min\{s,t\}\right).$$

For a fixed value of $s$,

$$\min\{s,t\} = \begin{cases} s, & t > s \\ t, & t \leq s \end{cases},$$

from which it follows that

$$\frac{\partial^2}{\partial s\partial t} \min\{s,t\} = \frac{\partial}{\partial s} u(s-t) = \delta(s-t).$$

Consequently,

$$\frac{\partial^2}{\partial s\partial t} R_N(s,t) = \lambda^2 + \lambda \delta(s-t).$$
Using this in (4.44) and (4.43), we have

\[ E[Y_t^2] = \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} h(t-\tau) h(t-\tau') \left[ \lambda^2 + \lambda \delta(\tau - \tau') \right] d\tau d\tau' \]

\[ = \lambda^2 \left( \int_{-\infty}^{\infty} h(\tau) d\tau \right)^2 + \lambda \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} h(t-\tau) h(t-\tau') \delta(\tau - \tau') d\tau d\tau' \]

\[ = \lambda^2 \left( \int_{-\infty}^{\infty} h(\tau) d\tau \right)^2 + \lambda \int_{-\infty}^{\infty} h^2(\tau) d\tau. \]

Substituting this expression into (4.42), we obtain (4.38).