Analog Signal Processing for Particle Detectors

PSI, LTP-Seminar, 12.06.2017
Citation of a slightly dispirited assistant I met during the early days of my studies:

“If you need an amplifier then try to build an oscillator, I am sure it will never oscillate, but if you need an oscillator then try to build an amplifier…”

Of course this statement is not true, but there are a couple of reasons why an oscillator may not oscillate whereas an amplifier does.
Analog Signal Processing for Particle Detectors

Topics

- analog signal processing
- charge and current preamplifier
- pulse shaping
- examples
Analog Signal Processing for Particle Detectors

- Mostly linear treatment of weak currents caused by small charges deposited or generated within a detector.
- Its goal is the extraction of information out of the weak generally noisy signals with the best possible S/N-ratio and/or the lowest possible timing jitter.
- It includes at least three steps:
  - signal capturing and (pre-)amplification
  - pulse shaping by bandwidth manipulations
  - signal conditioning for the following stage

Ch1 (yellow): signal after pulse shaping
Ch2 (blue): signal e.g. after a charge sensitive preamplifier

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Keywords in Analog Signal Processing are:

- Gain
- Bandwidth
- Dynamic input/output range
- Stability
- Linearity
- Pulse shaping/Filtering
- Noise
- Distortion
- Drive strength
- Powering and grounding scheme
- Power consumption and cooling
- Mechanical considerations (housing, size, integration)

Not covered in the above list are topics specific for non-linear analog processing (e.g. frequency mixing as it is usual done in RF-communication, ...)

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What has to be known to be able to develop a signal processing chain

**from the experiment:**
- parameters to be measured (energy, position, intensity, time of arrival, counts, ...)
- expected hit rate and their time distribution (poisson, bunch,...)
- other particles giving signals which has to be suppressed (e.g. by energy evaluation)
- environment conditions (magnetic fields, vacuum, ....)

**from the detector:**
- amount of charge deposited
- charge collection time
- detector recovery time
- detector capacitance and resistivity
- part of signal containing information
- readout scheme (e.g. charge division, time delay, ...)
- number of channels
- operating voltage
What has to be known to develop a signal processing chain

- **from the next DAQ stage:**
- type of stage (discriminator, counter, TDC, ADC, ...)
- input type (single ended, differential)
- input voltage or current range
- bias point
- input impedance
- input bandwidth or sampling frequency (for analog signals)
- minimum pulse width (for digital signals)
Applicable for Measurements of Energy deposited on detector

\[ E_{\text{particle}} \propto Q_s = \int \tau i_s(t) dt \quad \text{with} \quad \tau \gg T_{\text{charge\_collection}} \]

typical dynamic range: \[ \frac{Q_{s\text{max}}}{Q_{s\text{min}}} = 10^3 \ldots 10^5 \] (= 60 ... 100 dB or = 10...17 bits)

Energy information is a charge => Integration of current is needed

**Charge sensitive preamplifier:**
The most common ways are either to use a charge sensitive preamplifier based on a operational amplifier or especially when there is a cable in between detector and preamplifier a resistor with the cable impedance followed by a voltage amplifier:
Charge Sensitive amplifier using an operational amplifier:

Operational Amplifier:
Open loop gain $dV_{out}/dV_{in} = -A$ ($|A| \approx 1E3-1E4 >> 1$)
Input resistance $R_{in} \approx 10 - 100 \, \text{M}\Omega$ (or even higher)
Negative feedback forces $V_{in} \approx 0$ (but not $= 0$)
=> partial charge transfers from $C_d$ to $C_f$

Reason for the only partial transfer:
Amplifier transforms the feedback capacitor $C_f$ into a dynamic input capacitance

$$C_{in} = \frac{Q_{in}}{V_{in}} = \frac{C_f \cdot V_f}{V_{in}} = \frac{C_f \cdot V_{in} \cdot (A+1)}{V_{in}} = C_f \cdot (A + 1)$$

=> Free charges from the detector are divided on both capacitances ($C_d$ and $C_{in}$) with the ratio: $v_{in} = Q_d/C_d = Q_{in}/C_{in}$ => $Q_d = Q_{in} \cdot C_d/C_{in}$ and $Q_{tot} = Q_d + Q_{in} = Q_{in} (C_d/C_{in} + 1)$

$$\Rightarrow \frac{Q_{in}}{Q_{tot}} = \frac{1}{(\frac{C_d}{C_{in}} + 1)}$$

Charge sensitivity of the amplifier: $A_{VQ} = \frac{V_o}{Q_{in}} = -\frac{A}{A+1} \cdot \frac{1}{C_f} \approx -\frac{1}{C_f} \left[\frac{V}{C}\right]$ 

Attention: this circuit will not really work, see next slide
Charge Sensitive amplifier using an operational amplifier:
The previously shown circuit will not work, as there
is no significant discharge of the feedback capacitance.
The amplifier would integrate up until it saturates.

=> discharge mechanism needed:

\[ \tau = R_f \cdot C_f \] charge collection time
larger \( \tau \):
- increases linearity but
- reduces gain

has an exponential slope (\(dU \sim e^{-t/\tau}\))

reset by a transistor (normally not after each event)

voltage \(V_{res}\) may be generated by a
comparator with hysteresis switching
the transistor on after \(V_{out}\) reaches a
certain level
generates unpredictable dead-times

\[ I_{const} \ll I_{in} \]

reduces gain

has a linear slope (\(dU/dt = I_{const}/C_f\))

may be harder affected by temperature variations than resistor
Dynamic Range of Amplifiers:
Ratio between largest ($S_{max}$) and smallest ($S_{min}$) signal for which the amplifier works within its specifications.

It is typically expressed in [dB]: \[ DR = 20 \cdot \log_{10} \left( \frac{S_{max}}{S_{min}} \right) \text{ [dB]} \]

Rem: The dynamic range can be expressed for output or input signals. In a perfect linear system it would be the same.

For precision measurement systems:
The specification of a minimum S/N-ratio and enhanced linearity requirements reduce the available dynamic range of the system.

![Diagram showing dynamic range and parameters]
**Applicable for event counting at high counting rates**
Remark: Current preamplifier (also called transimpedance amplifier) are not faster but they have almost no tail what makes their signals shorter

**Characteristics:**
- Best for high counting rates
- Poorer noise performance compared to charge sensitive preamplifiers
- More affected by stability issues especially for detectors with high detector capacitance

**Transimpedance preamplifier (TIA):**

The working principle is quite similar than for the charge amplifier except that the feedback resistor can not store charges

=> The main topic here is stability
Stability is affected by detector capacitance

Amplifier OPA657

f_GBW 1,6E+9 Hz

DC Performance: Open Loop Voltage Gain:

<table>
<thead>
<tr>
<th>OL Gain [dB]</th>
<th>Frequency [Hz]</th>
</tr>
</thead>
<tbody>
<tr>
<td>70</td>
<td>1,0E+0</td>
</tr>
<tr>
<td>70</td>
<td>350,0E+3</td>
</tr>
<tr>
<td>0</td>
<td>1,6E+9</td>
</tr>
</tbody>
</table>

AC Performance: Small Signal Voltage Gain

<table>
<thead>
<tr>
<th>Gain [V/V]</th>
<th>Gain [dB]</th>
<th>Bandwidth [Hz]</th>
</tr>
</thead>
<tbody>
<tr>
<td>7</td>
<td>16,90</td>
<td>350,0E+6</td>
</tr>
<tr>
<td>10</td>
<td>20,00</td>
<td>275,0E+6</td>
</tr>
<tr>
<td>20</td>
<td>26,02</td>
<td>90,0E+6</td>
</tr>
</tbody>
</table>

Input Common Mode Capacitance

C_CM 4,5E-12 F

Detector Capacitance

<table>
<thead>
<tr>
<th>Capacitance</th>
<th>Capacitance [F]</th>
</tr>
</thead>
<tbody>
<tr>
<td>C_J</td>
<td>10,0E-12</td>
</tr>
</tbody>
</table>

Amplifier choose

<table>
<thead>
<tr>
<th>RF [Ohm]</th>
<th>RF 560,0E+3</th>
</tr>
</thead>
</table>

calculated added safety calculated:

<table>
<thead>
<tr>
<th>CF [F]</th>
<th>CF 42,7E-15</th>
</tr>
</thead>
<tbody>
<tr>
<td>CF min.</td>
<td>CF 57,3E-15</td>
</tr>
<tr>
<td>CF</td>
<td>CF 100,0E-15</td>
</tr>
</tbody>
</table>

Annahme: C_CM + C_J >> CF

3dB point of TF 2,84E+06 Hz

Transimpedance gain

A_V = R_F

f_3dB = 1/(2πR_F C_F)

Open loop gain

f_p = 1/(2πR_F C_F)

f_z = 1/(2πR_F C_D)

Noise gain

A_V = noise gain

detector capacitance adds noise gain and bandwidth not available to the signal

optimum gain-bandwidth and noise performance when f_p close to open loop gain curve

becomes unstable when f_p outside the open loop gain curve
Reasons for pulse shaping:

- **Shaping towards increased pulse width:**
  
  => **Bandwidth-reduction**
  - Improve of Signal-to-Noise Ratio
  - Pulse conditioning to match next stage requirements
    (e.g. stretching & rounding of sharp peaks for amplitude measurement with an ADC at a given $f_{\text{sampling}}$)

- **Shaping towards reduced pulse width**
  
  => **Improve of Pulse Pair Resolution (pile-up reduction):**
  - Increase of maximum count-rate
  - Minimization of energy-measuring errors
  - Dead-time or dead time variation reduction

- **Other, sometimes included objectives:**
  
  - Reduction of baseline problems
  - Minimization of timing errors

Conflicting objectives (e.g. best S/N-Ratio at max. count-rate):

=> “Optimum shaping” is usually an application dependent compromise and has to be specified!
**Most common pulse shaping methods:**

1. CR-RC Pulse Shaping  
2. Pole-Zero Cancellation  
3. Delay-Line Pulse Shaping  
4. Gaussian Filter  
5. Raised-cosine Filter  
6. Sinc Filter  
7. Constant Fraction Discriminator  
8. Baseline Restorer  

Blue = subjects in this talk
1. **CR-(RC)^n Pulse Shaping:**

![Diagram of Pulse Shaping Circuit]

- **High pass filter ("differentiator")**
  \[ \tau_D = R_D \cdot C_D \]

- **Low pass filter 1 ("integrator")**
  \[ \tau_{l1} = R_{l1} \cdot C_{l1} \]

- **Optional: Low pass filter n ("integrator")**
  \[ \tau_{ln} = R_{ln} \cdot C_{ln} \]
1. CR-(RC)^n Pulse Shaping:

Symbol: ▽ = voltage amplifier with high input impedance

- high pass filter ("differentiator")
  \[ \tau_D = R_D \cdot C_D \]

- low pass filter 1 ("integrator")
  \[ \tau_{i1} = R_{i1} \cdot C_{i1} \]

- optional: low pass filter n ("integrator")
  \[ \tau_{in} = R_{in} \cdot C_{in} \]

may have as well further "differentiators" in between

- high pass filter ("differentiator")
  \[ \tau_D = R_D \cdot C_D \]
CR - Pulse Shaping ("differentiator"):

![Diagram of pulse shaping circuit]

- **Time domain characteristics**
  - Input voltage: $V_I = V$
  - Output voltage: $V_O = V \exp(-t/RC)$
  - $RC \gg T$ and $RC \ll T$

- **Design limits for $R_D$:**
  - A lower limit for $R_D$ is given by the max. output drive current of the previous amplifier stage and their maximum peak to peak output voltage.
  - To be dominant, $R_D$ should be big compared to the output impedance of the previous amplifier.

- **Step response:**
  $$h(t) = \frac{V_{Dout}(t)}{V_{step}(t)} = e^{-\frac{t}{R_DC_D}}$$

- **Reduces pulse width if $RC \ll T$, but has inconveniences:**
  - Generates bipolar output pulses (positive on rising edges and negative on falling edges) => only applicable if falling edge has a much longer time constant.
  - Peak at maximum amplitude is very short in time (bad e.g. for ADC-sampling).
**CR - Pulse Shaping (“differentiator”):**

- **High pass filter (“differentiator”)**
  \[ \tau_D = R_D \cdot C_D \]

- **Step response:**
  \[ h(t) = \frac{V_{Dout}(t)}{V_{step}(t)} = e^{-\frac{t}{R_D C_D}} \]

- **Transfer function:**
  \[ G_D(s) = \frac{V_{Dout}(s)}{V_{Din}(s)} = \frac{s C_D R_D}{1 + s C_D R_D} \]

**Frequency domain characteristics**

- **Magnitude of CR high pass with \( f_{3db} = \frac{1}{2\pi RC} = 1E+6 \) [Hz]**

- **Phase of CR high pass with \( f_{3db} = \frac{1}{2\pi RC} = 1E+6 \) [Hz]**
Pulse Shaping Methods

**RC - Pulse Shaping ("integrator"):**

\[ h(t) = \frac{V_{\text{out}}(t)}{V_{\text{step}}(t)} = (1 - e^{-\frac{t}{R_i C_i}}) \]

- Input: \( V_i = V \)
- Output: \( V_0 = V[1 - \exp(-t/RC)] \)

\[ \tau_i = R_i \cdot C_i \]

- Low pass filter ("integrator")
- Step response:

\[ RC = T \]

\[ RC >> T \]

\[ RC << T \]

0.63 \( V_i \)

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Pulse Shaping Methods

**RC - Pulse Shaping (“integrator”):**

![RC Pulse Shaping Diagram]

- Frequency domain characteristics
- Step response:
  \[ h(t) = \frac{V_{\text{out}}(t)}{V_{\text{step}}(t)} = (1 - e^{-\frac{t}{\tau}}) \]
- Transfer function:
  \[ G_I(s) = \frac{V_{\text{out}}(s)}{V_{\text{in}}(s)} = \frac{1}{1 + sC_I R_I} \]

Magnitude of RC low pass with \( f_{\text{db}} = 1/(2\pi RC) = 1E+6 \text{ [Hz]} \)

Phase of RC low pass with \( f_{\text{db}} = 1/(2\pi RC) = 1E+6 \text{ [Hz]} \)
CR-RC Pulse Shaping (combination of a “differentiator” and an “integrator”):  

\[ G_{DI}(s) = \frac{V_{out}(s)}{V_{in}(s)} = G_D(s) \cdot G_I(s) = \frac{sC_DR_D}{(1 + sC_DR_D) \cdot (1 + sCIR_I)} \]

step response:

\[ h(t) = \frac{v_{out}(t)}{v_{step}(t)} = L^{-1}\left\{ \frac{G_{DI}(s)}{s}\right\} = \frac{\tau_D}{\tau_D - \tau_I} \cdot \left( e^{-\frac{t}{\tau_D}} - e^{-\frac{t}{\tau_I}} \right) \]

and for the most common case with \( \tau_D = \tau_I = \tau \):

\[ h(t) = \frac{t}{\tau} \cdot e^{-\frac{t}{\tau}} \]
**CR-RC Pulse Shaping—mathematical intermezzo:** Derivation of $h(t)$ for the case where $\tau_D = \tau_I$:

Starting point is the previous general formula for the CR-RC step response:

$$h(t) = L^{-1} \left\{ \frac{G_{DI}(s)}{s} \right\} = \frac{v_{out}(t)}{v_{step}(t)} = \frac{\tau_D}{\tau_D - \tau_I} \cdot (e^{-\frac{t}{\tau_D}} - e^{-\frac{t}{\tau_I}})$$

As the above formula for $\tau_D = \tau_I$ apparently results in a zero divided by zero division, a limes calculation with $\Delta \tau = \tau_D - \tau_I \to 0$ may help:

$$h(t) = \lim_{\Delta \tau \to 0} \left( \frac{\tau_D}{\tau_D - \tau_I} \cdot e^{-\frac{t}{\tau_D}} \cdot (1 - e^{-t(\frac{1}{\tau_I} - \frac{1}{\tau_D})}) \right)$$

factored out $e^{-\frac{t}{\tau_D}}$

$$= \tau_D \cdot e^{-\frac{t}{\tau_D}} \lim_{\Delta \tau \to 0} \left( \frac{1 - e^{-t(\frac{\Delta \tau}{\tau_I \tau_D})}}{\tau_D - \tau_I} \right)$$

some minor cosmetics

$$= \tau_D \cdot e^{-\frac{t}{\tau_D}} \lim_{\Delta \tau \to 0} \left( \frac{\frac{d}{d\Delta \tau} (1 - e^{-t(\frac{\Delta \tau}{\tau_I \tau_D})})}{\frac{d}{d\Delta \tau} (\Delta \tau)} \right)$$

l’Hospital’s rule

$$= \tau_D \cdot t \cdot e^{-\frac{t}{\tau_D}} \lim_{\Delta \tau \to 0} \left( \frac{\frac{t}{\tau_I \tau_D} \cdot e^{-t(\frac{\Delta \tau}{\tau_I \tau_D})}}{\tau_I} \right) = \frac{t}{\tau} \cdot e^{-\frac{t}{\tau}}$$

limes, then $\tau_I = \tau_D = \tau$
CR-RC Pulse Shaping (combination of a “differentiator” and an “integrator”):

\[ G_D(s) \cdot G_I(s) = \frac{V_{out}(s)}{V_{in}(s)} = \frac{V_{out}(s)}{V_{in}(s)} = \frac{sC_D R_D}{(1 + sCDR_D) \cdot (1 + sCIRI)} \]

Effect of a simple CR-RC Pulse shaping:
Signal reduced (bad), but much higher reduction of noise => enhancement of S/N-Ratio
1. **CR-(RC)^n Pulse Shaping:**

Increasing number of integrators make the output Pulse more symmetrical with a faster return to baseline.

To preserve the peaking time the time constant of the original integrator (for n=1) has to be divided by the number n: \( \tau_n = \tau_1 / n \)
2. **Pole-Zero Cancellation:**

- Controls the lower cutoff frequency
- No impact on the upper cutoff freq.
- Can be used for pulse shape and baseline recovery adjustments
- It is not the most efficient but an easy to integrate shaper

**Application:**

Charge sensitive amplifiers and stabilized transimpedance amplifiers (with a capacitance in their feedback) have a pole at $s=1/C_F R_F$ in their transfer function. This pole can be compensated with the above circuit (yellow part) which creates a zero. To compensate the pole one has to choose: $C_{PZ} R_{PZ} = C_F R_F$

The value of the Resistor $R_X$ has no effect on the zero except it would be zero or infinite.

=>$\text{pole zero cancellation can be applied as well in other networks for pulse shape and baseline optimization (e.g. for AC-coupled signal paths)}$
2. Pole-Zero Cancellation:

In frequency domain:

In time domain:
Some Examples

Conceptual study TOFTAS (ESS-WP1)

Conceptual study POLDI-Upgrade

Pictures below: Hildebrand, Stoykov, Mosset
Example: Charge Division Readout
**Schematics of the Input Amplifier:**

**OPA657:** FET-input
- GBP = 1.6 GHz
- $T_{\text{rise} \& \text{fall}} = 1$ ns (0.2 V Step)
- $T_{\text{setting}} = 20$ ns (2 V Step)
- $Z_{\text{inDiff}} = 10 \, \Omega \parallel 0.7 \, \text{pF}$
- $Z_{\text{inCM}} = 10 \, \Omega \parallel 4.5 \, \text{pF}$
- $A_{\text{OL}} = 70 \, \text{dB}$
- $V_{\text{in-noise}} = 4.8 \, \text{nV/\sqrt{Hz}}$
- $I_{\text{in-noise}} = 1.3 \, \text{fA/\sqrt{Hz}}$

**OPA658:** Current Feedback
- BW = 900 MHz (unity gain stable)
- SR = 1700V/us
- $G/\Phi_{\text{error}} = 0.025\%/0.02^\circ$
- $G_{\text{flatness}} = 0.1 \, \text{dB to 135MHz}$
- $I_{\text{out}} = 80 \, \text{mA}$
- $V_{\text{in-noise}} = 45.3 \mu V_{\text{rms}}$ (to 200 MHz)
- $I_{\text{in-noise}} = 32 \, \text{pA/\sqrt{Hz}}$
Schematics of the Track and Hold Stage:

- Track and hold stage
- Inverting amplifier: adding an offset to original signal
- Peak-detecting
  - Peak signal to ADC
  - Peak detected to FPGA
- Low level discriminator
- Event detected to FPGA
- Analog storage: Happens here
MAR-Amplifier for Photon-Detectors

Functional description:
The “MAR”-Amplifier was designed for fast photo-diode (e.g. APD) applications where due to limited space at the detector-side the photo-diodes has to be off-board and where coax cables are needed to connect them to the amplifiers. The board layout allows for the assembly of two RF gain blocks which can be chosen from a wide palette of monolithic amplifiers in .085 micro-x or SOT-89 case styles. Within 50 Ω systems the amplifier can be used either as transimpedence or voltage amplifier. It is cascadable.

The two gain stage amplifier design includes:
- π-attenuator for gain setting
- adjustable pole-zero cancellation for pulse shaping
- integrating capacitor for upper bandwidth limitation
- lower bandwidth adjustment by of AC-coupling capacitor
- filtered high voltage path for reverse bias voltage

Typical application scheme:

Specifications (with MAR-6SM on both stages):
- Input impedance 50 Ω
- primary transimpedance gain 50 V/A
- voltage gain (with no π- attenuation):

<table>
<thead>
<tr>
<th>Frequency</th>
<th>100 kHz</th>
<th>750 MHz</th>
<th>1 GHz</th>
<th>2 GHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>Gain</td>
<td>40 dB</td>
<td>37 dB</td>
<td>32 dB</td>
<td>22 dB</td>
</tr>
</tbody>
</table>

- linear output range (1dB compression) 342 mVrms @ 50Ω
- max. reverse bias voltage 200 V
- power-consumption amplifier +12 V, 32 mA
Example: MAR-Amplifier

The MAR-Amplifier
**Brief description:**
Amplifier for fast avalanche photo diodes (e.g. Silicon Sensor, AD230-8, 180 ps rise time) with integrated low and high voltage filters, a current monitor output for the APD bias current and pulse-shaping options like a Π-attenuator, an integrating capacitor and a pole-zero cancellation stage.

The board layout allows to read out the APD-signal either on the cathode or anode side. The two RF amplifier stages can be assembled with a wide palette of monolithic amplifiers in .085 micro-x or SOT-89 case styles.

**Specifications (MAR-6SM on both stages):**

**RF-Amplifier Output:**
- current to voltage conversion 50 V/A
- voltage gain at frequency

<table>
<thead>
<tr>
<th>Frequency</th>
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<th>750 MHz</th>
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<td>22 dB</td>
</tr>
</tbody>
</table>

- max. output voltage (1dB Comp): 282 mV @ 50Ω

**Current Monitor Output:**
- input current range 0 … 500 nA
- accuracy < 0.1% full scale
- current to voltage conversion 1 V/µA
- bandwidth 1 kHz

**Power-Supply:**
- Amplifier +/-12V, 300mA
- max. APD - voltage 200 V
MAR-Amplifier for STXM @ PolLux-Beamline

- High Voltage Filter (2-pol RC-Filter)
- Low Voltage Filter (Feedthrough- and Block-Capacitors)
- +V_{MAR}
- +V_{CM}
- -V_{CM}
- +V_{EXT} (+12 V)
- -V_{EXT} (-12 V)
- GND

- +V_{MAR}
- APD Signal Output
- Current Monitor Output
- +V_{CM}
- -V_{CM}
- 10-Attenuator 0.8 dB
- Integrator (optional)
- Pole-Zero Cancellation
- MAR-6 G ≈ 20 dB
- OP129 Transimp.-Amplifier G_{trans} ≈ 1 MV/A
- OP07 Inverting-Amplifier G = -1
- BUF634 Buffer-Amplifier G = 1

- Bias-Tee RF+DC
- DC RF
Static imaging and 1st movie of magnetization dynamics taken with the new APD-Detector at the PolLux-STXM

Sample: Co (50 nm) / Rh (0.75) / NiFe (50 nm)  
FM coupling due to Ne ion irradiation

Sample preparation: Sebastian Wintz, Thomas Strache

Principle of detection and data acquisition

Slide received on 24.06.2010 from Jörg Raabe