INTRODUCTION TO FRONT-END ELECTRONICS FOR PARTICLE DETECTORS

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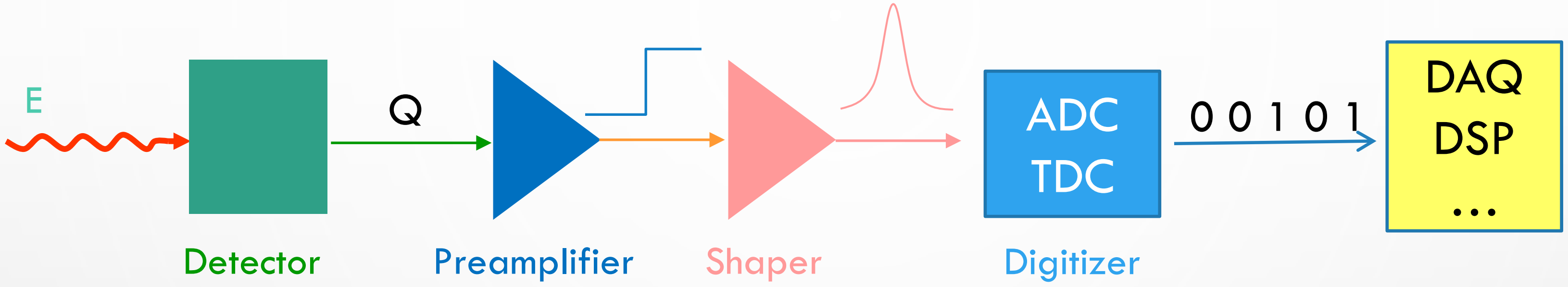
OUTLINE

- INTRODUCTION
- NOISE BASIC PRINCIPLES
- RADIATION DAMAGE
- PREAMPLIFIER SCHEMES
  - CHARGE-SENSITIVE AMPLIFIER
- SHAPERS
- HIT DISCRIMINATION AND TIME MEASUREMENT
- BIBLIOGRAPHY

INTRODUCTION

General readout architecture



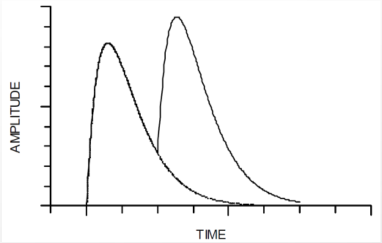
- The particle deposits energy in a detecting medium { **Gas**, **Solid**, **Liquid** }
- Energy is converted into an electrical signal: $Q = KE$
- The charge Q is typically small and must be amplified to be measured and processed
- The preamplifier converts Q into a voltage
- The shaper provides gain and shape, according to the application and trying to optimize S/N
- The Digitizer converts the "analog" information into sequence of bits, for storage and processing

FRONT-END ELECTRONICS

1. Goal: acquire an electrical signal from the detector

2. Gain and shaping time to be optimized for:

   - minimum detectable signal over the noise (maximize S/N)
   - energy measurements (linearity …);
   - event rate (pile-up, ballistic deficit, …);
   - time of arrival (time-walk, jitter …);
   - radiation tolerance;
   - power consumption;
   - cost



Example of pile-up

The requirements are often in conflict each other → the design comes out as a compromise, according to the specific application:

- Triggering (focus on timing)
- Tracking (focus on minimum detectable signal)
- Energy measurement (focus on linearity, dynamic range …)

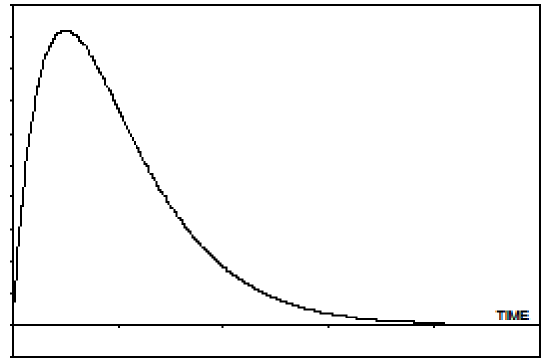
NOISE BASIC PRINCIPLES

NOISE

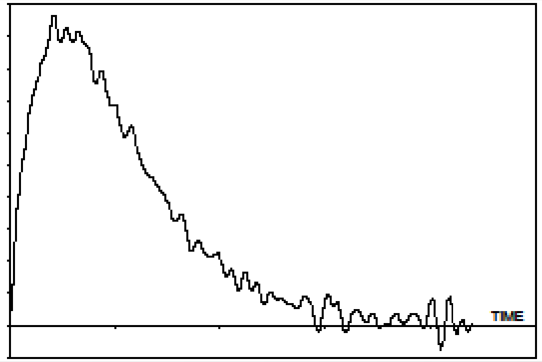
The precision of amplitude and timing measurements is limited by the NOISE

Definition

Noise is every undesirable signal superimposed to our signal of interest → fluctuations on amplitude and time measurement



Signal of ideal system



Signal + Noise

1. **External noise (interference)**

   It is generated by external sources (RF, ripple of power lines, ground loops …)

   Can be minimized by proper shielding, cabling …

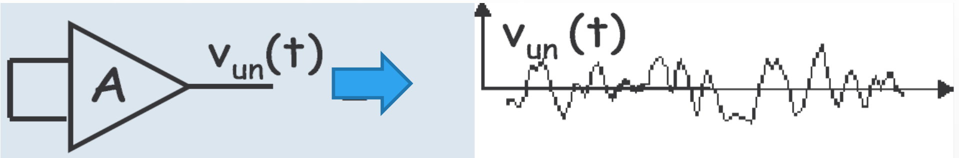
2. **Intrinsic noise**

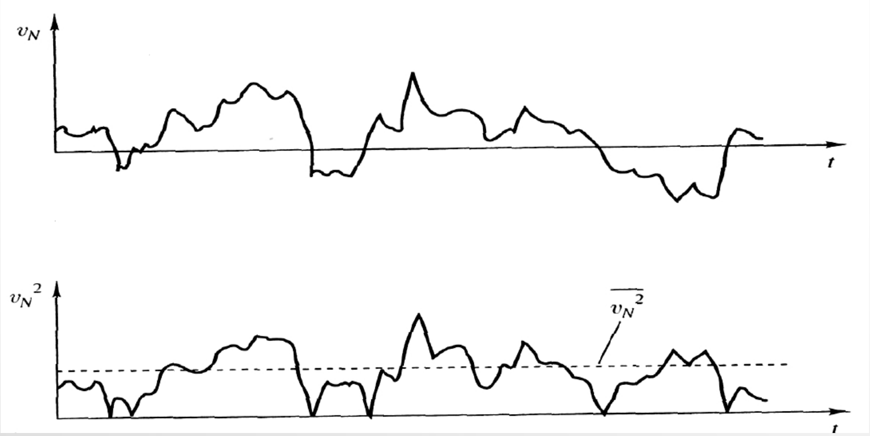
   It is a property of detector and/or electronics

   Can be reduced by proper design of front-end electronics

INTRINSIC NOISE

The output voltage of a real amplifier is never constant, even if $V_{in} = 0$
The fluctuations of $V_{un}(t)$ when $V_{in} = 0$ correspond to the noise of amplifier





- V_n has mean value = 0, but power ≠ 0
- A noise source is defined by its **POWER SPECTRAL DENSITY**: noise power per unit of bandwidth
- **If Power Spectral Density is constant with f → White Noise**

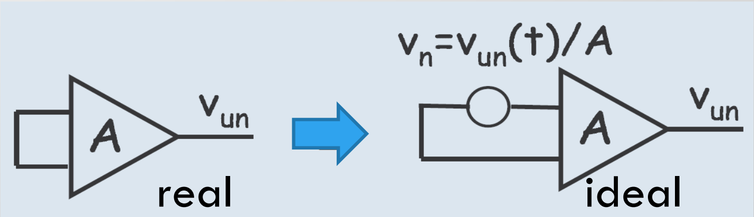
Source of **voltage noise**

$$\frac{dv_n^2(f)}{df}$$

Source of **current noise**

$$\frac{di_n^2(f)}{df}$$

The noise of a real amplifier can be attributed to a noise source in input to an ideal amplifier (noiseless)



- In the case of particle detection systems, where the input is a charge Q, we use ENC: Equivalent Noise Charge
- ENC is the signal charge which produces an output amplitude equal to rms noise
- Representing the noise with **ENC**, we can directly compare the input charge with the noise introduced by the FE amplifier

BASIC NOISE MECHANISMS

e → Material (Conductor, Semicond. … →

length = l

$$i = \frac{nev}{l}$$

i = current at the end of sample
n = number of carriers
e = unitary charge
v = velocity

The fluctuation of the current is given by: $<di>^2 = (\frac{ne}{l}<dv>)^2 + (\frac{ev}{l}<dn>)^2$

There are two basic mechanism contributing to noise:

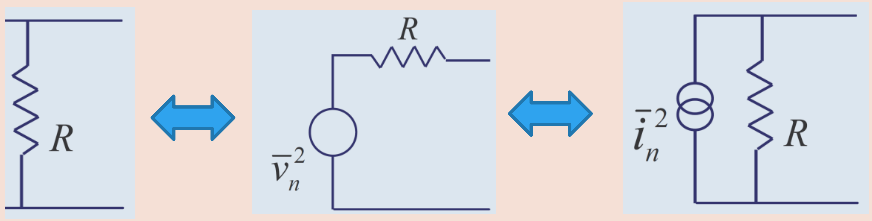
Velocity fluctuations → Thermal noise

Number fluctuations → $\begin{cases} \text{Shot noise} \\ \text{Excess (or flicker, or “1/f”) noise} \end{cases}$

THERMAL NOISE

Typical of resistors (also channel of a MOS)

- Caused by the random thermal motion of charge carriers (electrons)
- Does not depends on a DC current



A real resistor is equivalent to an ideal resistor + noise source (voltage or current)

Power spectral density

$$S_v(f) = \frac{dv_n^2}{df} = 4kTR \qquad S_i(f) = \frac{di_n^2}{df} = \frac{4kT}{R}$$

k = Boltzmann constant = $1.3806503 \times 10^{-23}$ J/K

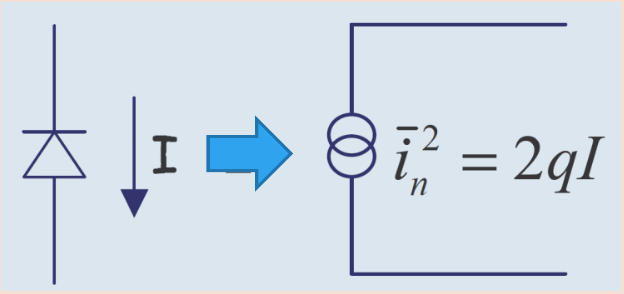
T = absolute temperature

R = resistance

Does not depend on f
→ white noise

SHOT NOISE

Caused by fluctuations in the number of charge carriers, for example in the current flowing in a semiconductor diode, where e/h cross a potential barrier



Power spectral density

$$S_i(f) = \frac{d\overline{i}_n^2}{df} = 2qI$$

Does not depend on f
→ white noise

FLICKER NOISE (1/f)

Associated to random trapping and recombination of charge carriers in the semiconductors, typically caused by imperfections in the interface regions

Power spectral density

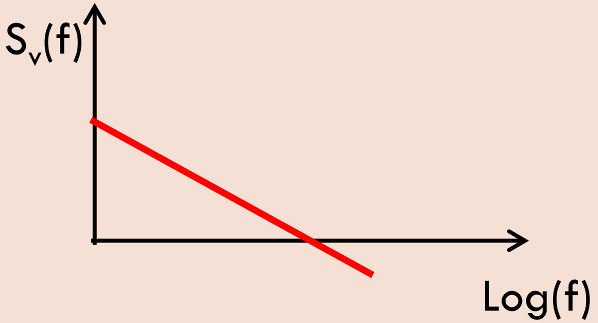
$$S_v(f) = \frac{d\overline{v}_n^2}{df} = K_f \frac{I^a}{f^b}$$

I is dc current

K_f is a constant (depends on the device)

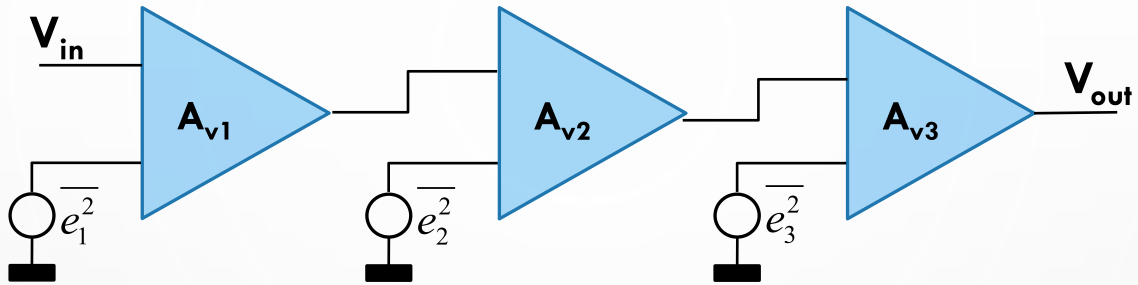
a ~ 0.5 ÷ 2

b ~ 1



Depends on f
→ important at low frequencies

INTRINSIC NOISE: IMPORTANCE OF FIRST STAGE



$$\begin{cases} V_{out} = A_{v1} * A_{v2} * A_{v3} * V_{in} \\ \overline{e^2_{out}} = A^2_{v1} * A^2_{v2} * A^2_{v3} * \overline{e^2_1} + A^2_{v2} * A^2_{v3} * \overline{e^2_2} + A^2_{v3} * \overline{e^2_3} \end{cases}$$

$$\left(\frac{Noise}{Signal}\right)^2 = \left(\frac{\overline{e^2_{out}}}{V^2_{out}}\right) = \frac{\overline{e^2_1} + \frac{\overline{e^2_2}}{A^2_{V1}} + \frac{\overline{e^2_3}}{A^2_{V1} * A^2_{V2}}}{V^2_{in}}$$

1. **Decrease as much as possible the noise contribution e_1^2 of the first stage (input transistor)**
2. Increase the gain A_{v1} of the first stage because the noise contribution of next stages are divided by the gain of previous stages

THE PROBLEM OF RADIATION DAMAGE

When an electronic device is exposed to radiation, like in HEP experiments, there is a permanent or transient modification of the electrical properties of the active devices:

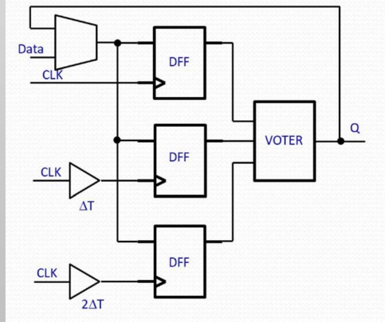
- Fake signals
- Corruption of memory content (Single Event Upset)
- Degradation of performance
- Catastrophic failure (Single Event Latch-up)

- Displacement damage: radiation (neutrons, protons, heavy ions…) change the arrangement of Si atoms in the crystal lattice → the electronic characteristic are altered

- Ionization damage: charged particles produces transient currents and entrapment of charge in SiO_2

✓ **Total dose (TID)** → Threshold shift, parasitic leakage current, mobility degradation
  - Constraints in the size of MOS (smaller device → larger damages)
  - For critical circuit, use of special layout techniques (Enclosed Layout Transistor) to minimize the leakage current. Not always applicable (i.e. not in 28 nm CMOS)

✓ **Single Event Effects (SEE)** → temporary or permanent errors
  - Adopt SEU mitigation in the digital circuitry (Triple Modular Redundancy)
  - Enough substrate contacts to minimize latch-up risk



TMR with correction

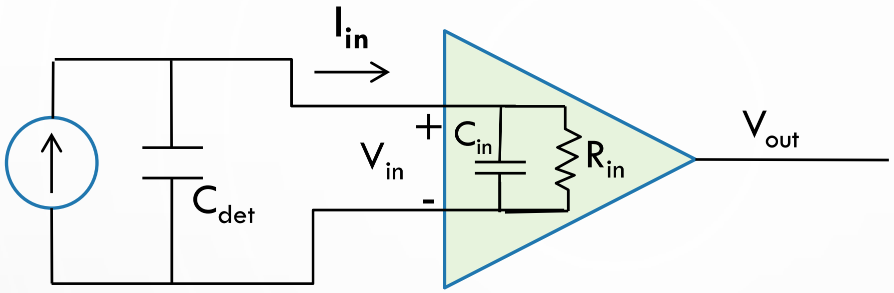
TECHNOLOGIES FOR FRONT-END ELECTRONICS

- **Most used technology for FEE is CMOS**
- Relatively "cheap" if relatively "old" technologies (nodes) are used
- Using the "multiproject foundry runs", prototyping and small productions are very affordable
- Suitable to combine on the same chip analog section, digital part and μprocessors
- Very low power consumption
- Highest transistor density
- The deep submicron CMOS tech. (≤ 130 nm) are rad-tolerant and suitable for HL-LHC, FCC
  - 130 nm: VFAT3 (CMS GEM), HGCROC (CMS HGCAL) → tens of Mrad
  - 65 nm: LpGBT, Phase-2 ATLAS/CMS pixel chips (RD53) at HL-LHC→ hundreds of Mrad
  - 28 nm: candidate technology for future upgrades → Grad
- In some applications, where higher speed is required, SiGe BiCMOS is also used. Not very large density → strip front-end chips

PREAMPLIFIER SCHEMES

SIGNAL INTEGRATION

$I_s(t)=Q_s\delta(t)$



R_{in} = amplifier input resistance
C_{in} = amplifier input capacitance

- The sensor signal is usually a short current pulse $\mathbf{I_s(t) = Q \cdot \delta(t)}$ with duration ranging from few hundreds of ps, as in Si sensors, SiPM and Resistive Plate Chambers to tens of µs, as in inorganic scintillators

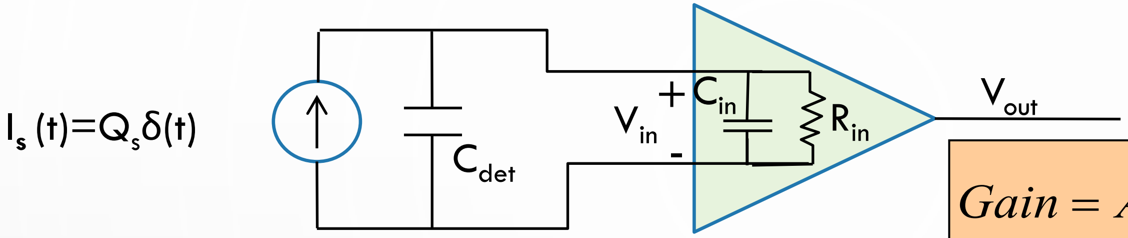
- The physic quantity of interest is the deposited energy E, that is proportional to Q

- We must integrate I to have a measurement of E: $E \propto Q_S = \int I_S(t)dt$

WHERE to integrate? Depending on charge collection time t_c and input time constant RC:

1. **Detector capacitance** → $V_{in} \propto Q_s$ → followed by voltage amplifier
2. **Current sensitive amplifier** → $V_{out} \propto I_s$ → followed by integration stage
3. **Charge sensitive amplifier** → $V_{out} \propto Q_s$

INTEGRATION ON C_{DET} (+ VOLTAGE AMPLIFIER)



$$Gain = A_v = \frac{V_{out}}{V_{in}}$$

If R_{in} is very big→ $\tau_{in} = R_{in}(C_{det}+C_{in})$ for discharging the sensor >> pulse duration (collection time)

the detector capacitance discharge slowly

$I_s(t)$ is integrated on the total capacitance $C_t = C_{det} + C_{in}$

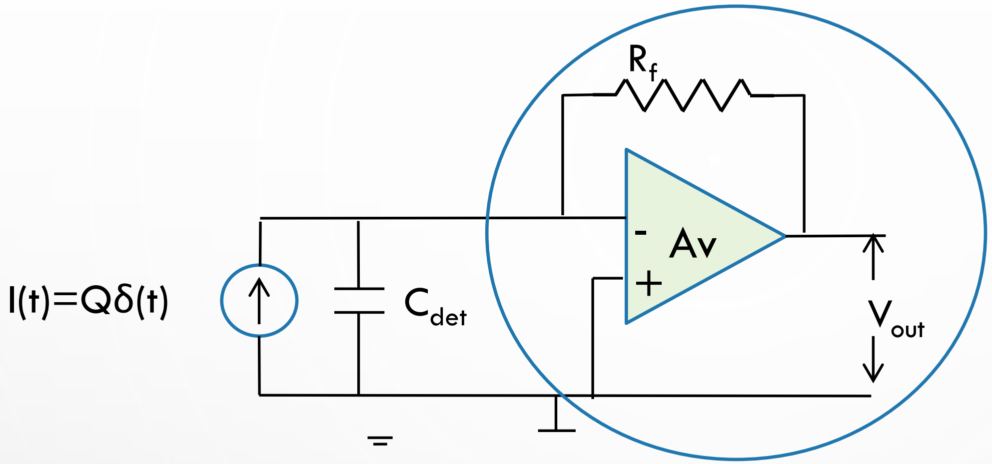
$$V_{in} = \frac{1}{C_t}\int I_s dt = \frac{Q_s}{C_{det} + C_{in}}$$

$$V_{out} = A_v \cdot V_{in} = A_v \cdot \frac{Q_s}{C_{det} + C_{in}}$$

In this method, V_{out} is proportional to Q_s, but it also depends on C_{det}

This is not desirable in the systems where C_{det} can vary:
- different strip length/width
- bias voltage
- ....

CURRENT-SENSITIVE AMPLIFIER



If R_{in} is small→ $\tau_{in} = R_{in}(C_{det}+C_{in})$ << pulse duration (collection time)

The detector capacitance discharges rapidly → the amplifier senses the current

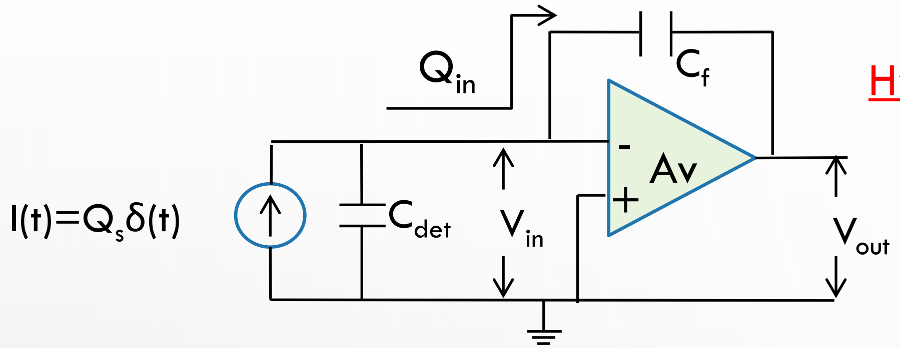
Using a transresistance amplifier (high gain operational amplifier with resistive feedback):

$$\boxed{V_{out} \propto I}$$

In this method, V_{out} is proportional to I and does not depend on C_{det}

- Input shape can be preserved (following shaping stages not strictly necessary)
- Feedback resistor introduce additional noise
- An integrating stage can follow the amplifier to provide a signal proportional to Q

CHARGE-SENSITIVE AMPLIFIER (CSA)



Hypothesis:

1. input impedance of op-amp is ∞ (i.e. MOS gate) → all current flows in the feedback network
2. A_v is very high

Voltage output: $V_{out} = -A_v V_{in}$

Voltage difference across C_f: $V_f = V_{in} - V_{out} = (A_v+1)V_{in}$

Charge deposited on C_f: $Q_f = C_f V_f = C_f(A_v+1)V_{in} = Q_{in}$ (for Hypothesis 1)

Effective input capacitance (seen by the sensor): $C_{in} = Q_{in}/V_{in} = C_f(A_v+1)$

GAIN (Charge Sensitivity):

$$CS = \frac{V_{out}}{Q_{in}} = -\frac{A_v V_{in}}{C_f(A_v+1)V_{in}} = -\frac{A_v}{C_f(A_v+1)} \approx -\frac{1}{C_f}$$

($A_v >> 1$)

BUT … not all the charge goes in the amplifier and is measured: a small fraction Q_{det} remains on C_{det} !!!

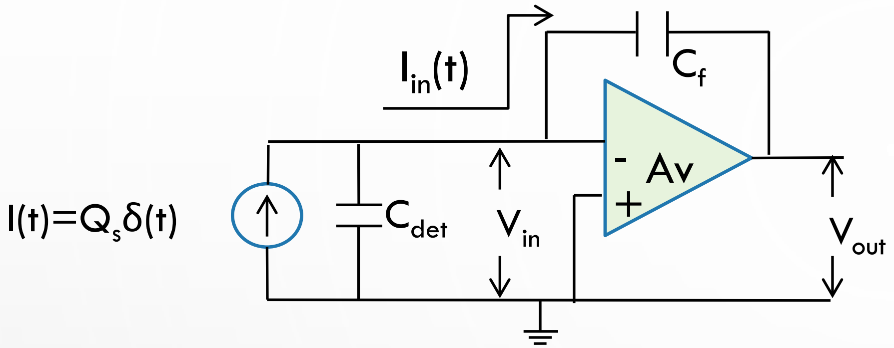
Charge transfer efficiency:

$$\frac{Q_{in}}{Q_S} = -\frac{Q_{in}}{Q_{det}+Q_{in}} = \frac{1}{1+\frac{Q_{det}}{Q_{in}}} = \frac{1}{1+\frac{C_{det}}{C_{in}}} \approx 1$$

(if $C_{in} = C_f(A_v+1) >> C_{det}$)

Example: $C_{det} = 10$ pF $A_v=10^3$ $C_f=1$ pF → $C_{in} = 1$ nF ⟹ $Q_{in}/Q_s = 0.99$

CHARGE-SENSITIVE AMPLIFIER: THE TIME RESPONSE



In the frequency domain:

$$V_{out}(\omega) = -A_v V_{in}(\omega)$$ (assuming Av constant and → ∞)

(infinite bandwidth)

$$V_{out}(\omega) - V_{in}(\omega) = -Z_f(\omega)\cdot I_{in}(\omega) = -\frac{I_{in}(\omega)}{j\omega C_f}$$

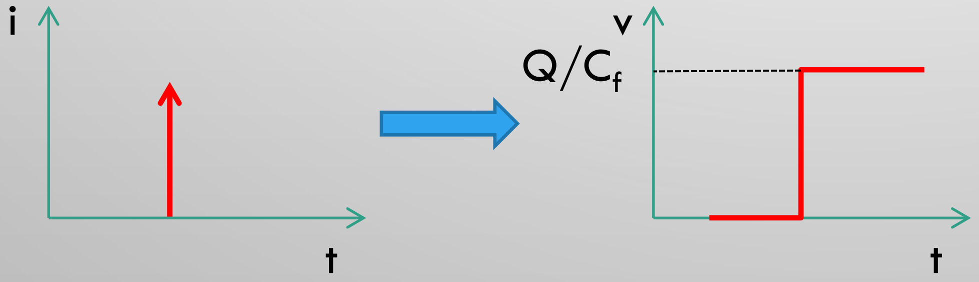
→

$$\boxed{V_{out}(\omega) = -\frac{I_{in}(\omega)}{j\omega C_f}\left(\frac{1}{1+\frac{1}{A_v}}\right) \approx -\frac{I_{in}(\omega)}{j\omega C_f}}$$

$I_{in}(t) = Q_{in}\delta(t)$ → $I_{in}(\omega) = Q_{in}$ → $\boxed{V_{out}(\omega) \approx -\frac{I_{in}}{j\omega C_f}}$ $\xrightarrow{\text{in time domain}}$ $\boxed{V_{out}(t) \approx -\frac{1}{C_f}\int I_{in}\delta(t)dt = -\frac{Q_{in}}{C_f}u(t)}$

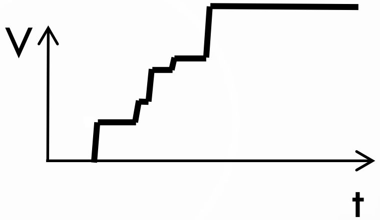
Step function

$I_{in}(t) = Q_{in}\delta(t)$

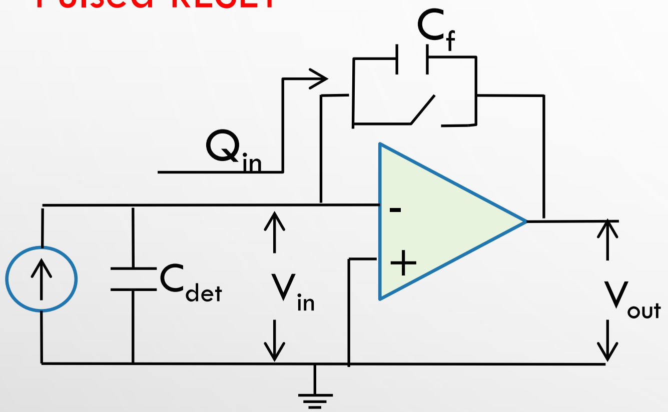


CHARGE-SENSITIVE AMPLIFIER: THE RESET

CSA feedback capacitor must be discharged, to restore the baseline of the preamplifier output and avoid the saturation



Pulsed RESET

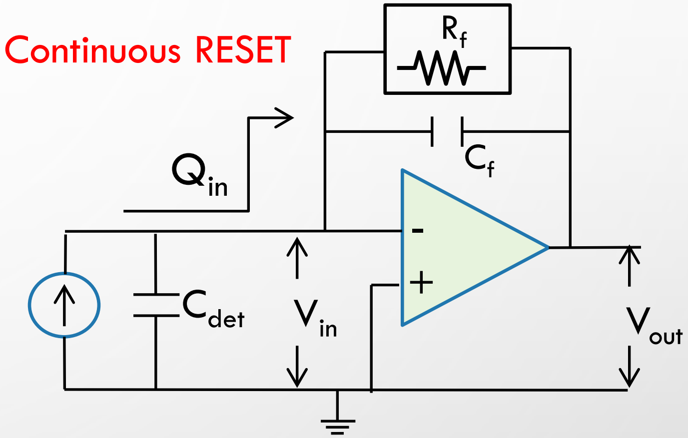


- The reset switch allows the removal of charge stored in C_f
- The switch can be closed periodically, driven by a control signal

Drawbacks:

- Dead time
- Switch noise
- Leakage current from the switch

Continuous RESET



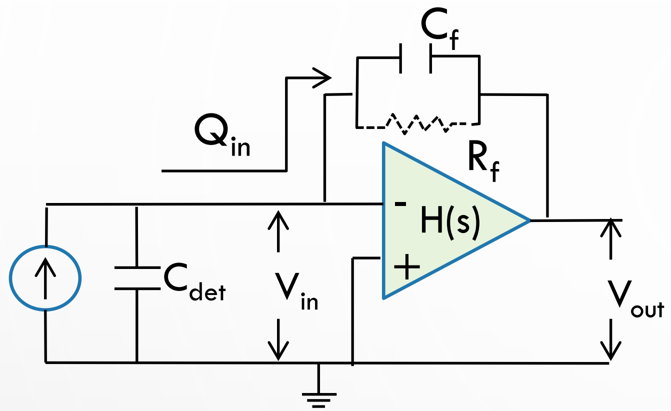
- Can be made with a resistor R_f that continuously discharges C_f
  - ➢ Discharge time constant $\tau = R_f C_f$
- CMOS circuit, also featuring detector leakage current compensation (i.e. Krummenacher circuit)

Drawbacks:

- Additional parallel noise (R_f must be very large)
- Long tail → Risk of pile-up

CHARGE-SENSITIVE AMPLIFIER: THE REAL CASE

$I(t)=Q_s\delta(t)$



<u>Two elements to be considered:</u>

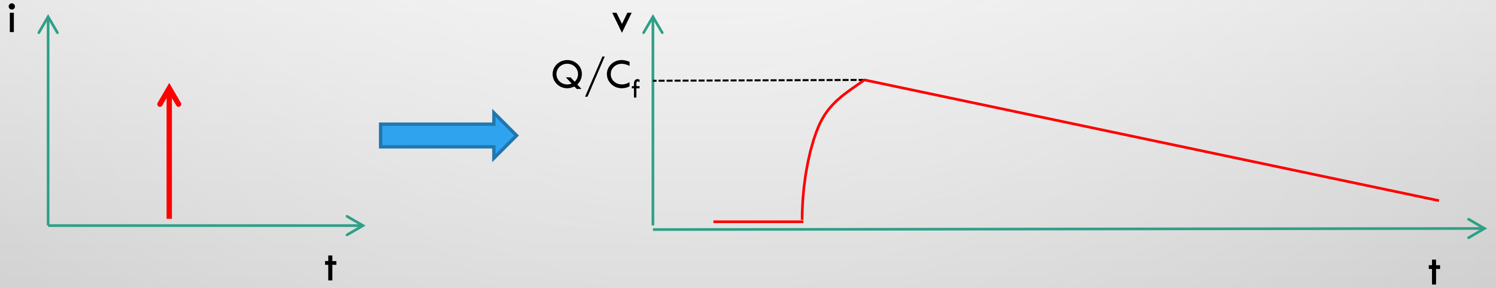
1. Resistor R_f used to discharge C_f. Since it is a source of parallel noise (inject noise current into input noise), it must made very large to decrease its contribution to noise. Typical values are several tens or hundreds of MΩ
2. Real amplifier (finite bandwidth and gain)

$$\frac{V_{out}(s)}{I(s)} = \frac{\dfrac{-g_m}{C_L C_T}}{\left(s+\dfrac{1}{R_f C_f}\right)\left(s+\dfrac{1}{R_i C_T}\right)}$$

2 poles →

$$\begin{cases} p_1 = \dfrac{1}{R_f C_f} & \text{(Low freq)} \\ p_2 = \dfrac{1}{R_i C_T} = \dfrac{\omega_0 C_f}{C_T} & \text{(High freq)} \end{cases} \quad \Rightarrow \quad \begin{cases} \tau_1 = R_f C_f & \text{Fall time constant} \\ \tau_2 = R_i C_T = \dfrac{C_T}{\omega_0 C_f} & \text{Rise time constant} \end{cases}$$

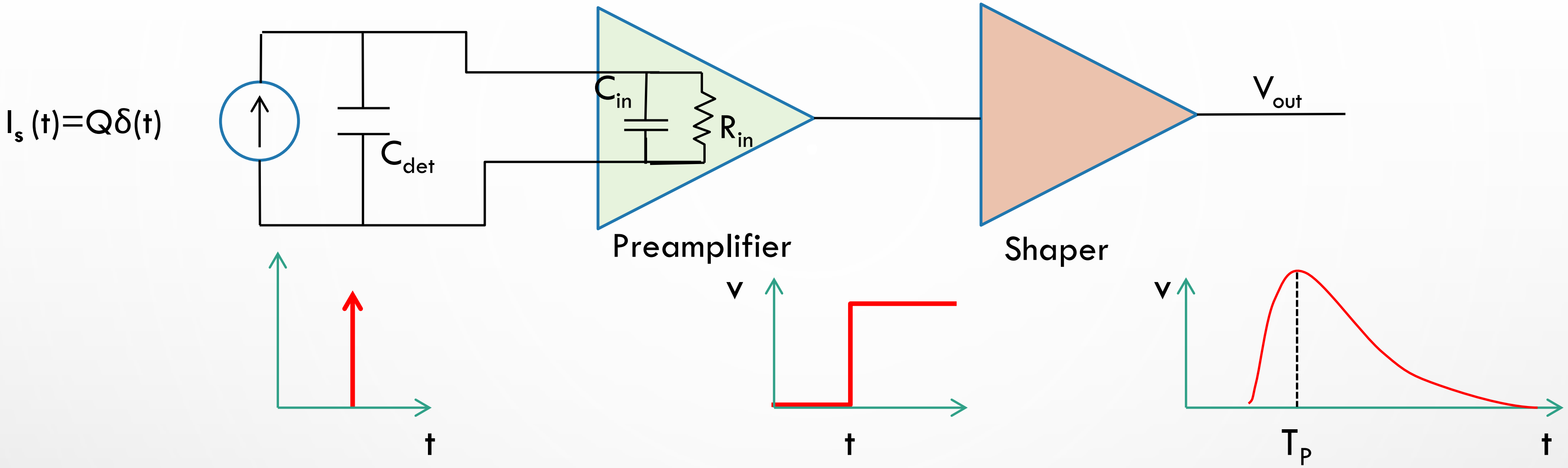
$I_{in}(t)= Q_{in}\delta(t)$



- The fall time depends on the feedback: can be very large, since R_f must be very high for low noise (>> 1 MΩ)
- The rise time depends on the input time constant, thus
  - R_i must be small to have short rise time
  - ω_0 : the amplifier GBW must be very large
  - $C_T \rightarrow C_d$: the rise time increase with detector capacitance

NOISE FILTERING: SHAPERS

PULSE SHAPING

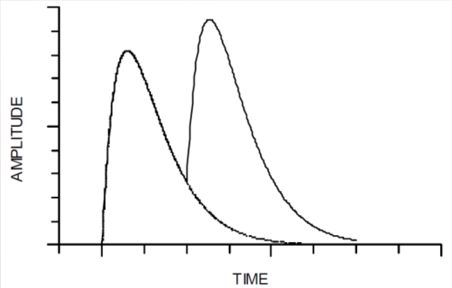


<u>The shaper</u> usually follows the preamplifier. It is a **filter** with two main purposes:

1. improve the signal-to-noise ratio S/N, restricting the bandwidth (defining the peaking time T_P)
2. tail the shape to improve the double-pulse resolution and avoid pile-up effect

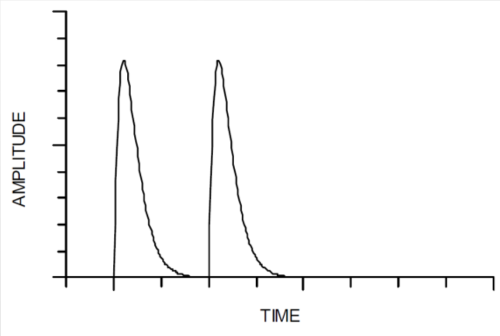
<u>Slower pulse:</u>

- Less noise
- Pile-up (distortion of amplitude measurement)



<u>Faster pulse:</u>

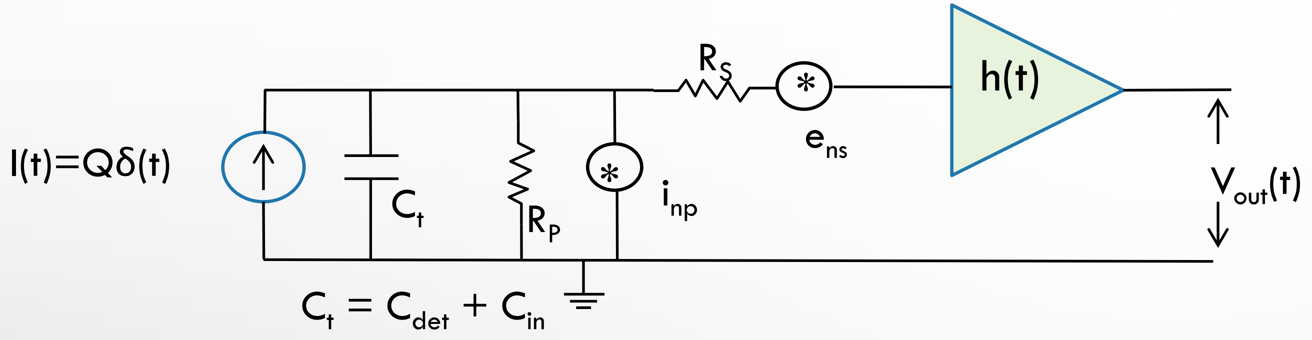
- More noise
- Double-pulse resolution



<u>The choice of the shaper (T_P, shape) derives from a compromise between the two targets</u>

EQUIVALENT NOISE CHARGE

To study the ENC, it is convenient to represent the chain with a noiseless amplifier, with transfer function h(t) and all noise sources at its input, represented by R_s and R_P



$$\overline{e_n^2} = 4KTR_S$$

$$\overline{i_n^2} = \frac{4KT}{R_P}$$

	BJT	MOSFET
R_s	$1/(2g_m)$	$2/(3g_m)$
R_P	$2h_{FE}/g_m$	$2KT/(qI_G) \sim 0$

$$g_m = \text{conductance} = \frac{\partial I}{\partial V}$$

It is possible to demonstrate that:

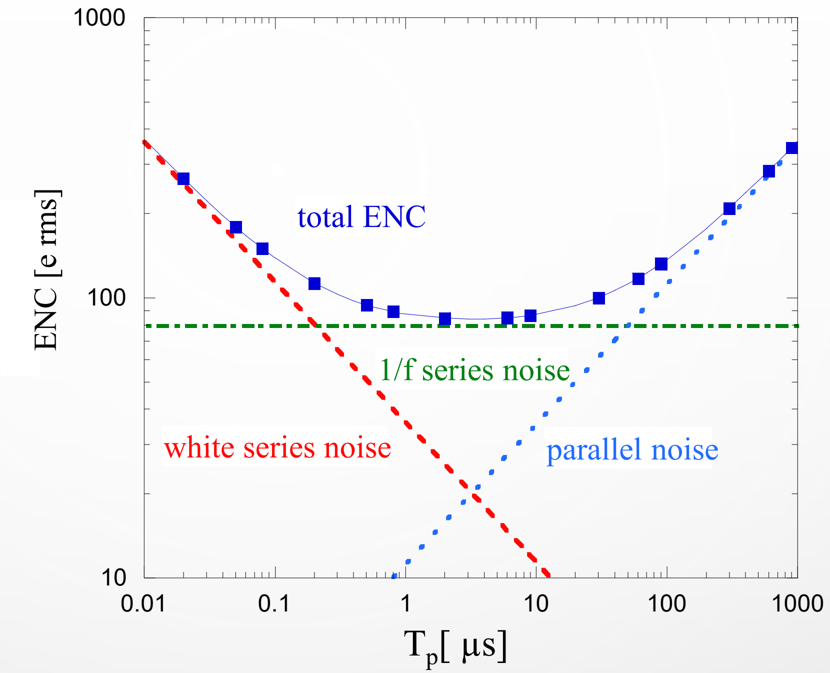
$$ENC^2 = \underbrace{\frac{2KTR_S \cdot C_t^2}{T_p}}_{\text{series noise}} + \underbrace{\frac{2KT \cdot T_P}{R_P}}_{\text{parallel noise}} + \underbrace{A_f C_t^2 A_2}_{\text{flicker noise (only MOSFET)}}$$

- **Series noise** depends on the amplifier characteristics (g_m). Increases with C_{det} and decreases with T_p
- **Parallel noise** depends on "external" factors (feedback R, shot noise of detector leakage current). Increases with T_p
- **Flicker noise** (only MOSFET): increases with C_{det} but independent of T_p

It is possible to minimize ENC by a correct choice of the dimensions of the preamplifier input device (**Capacitive matching**)

OPTIMUM FILTER

What is the best h(t) that minimizes ENC?

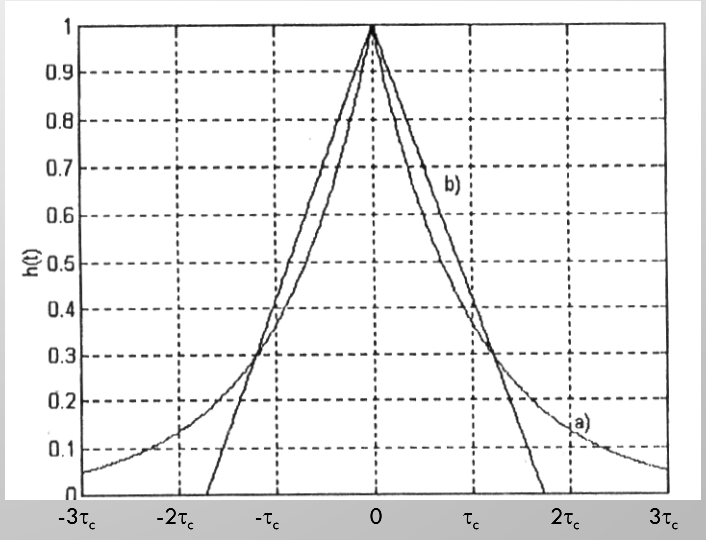


It is possible to demonstrate that

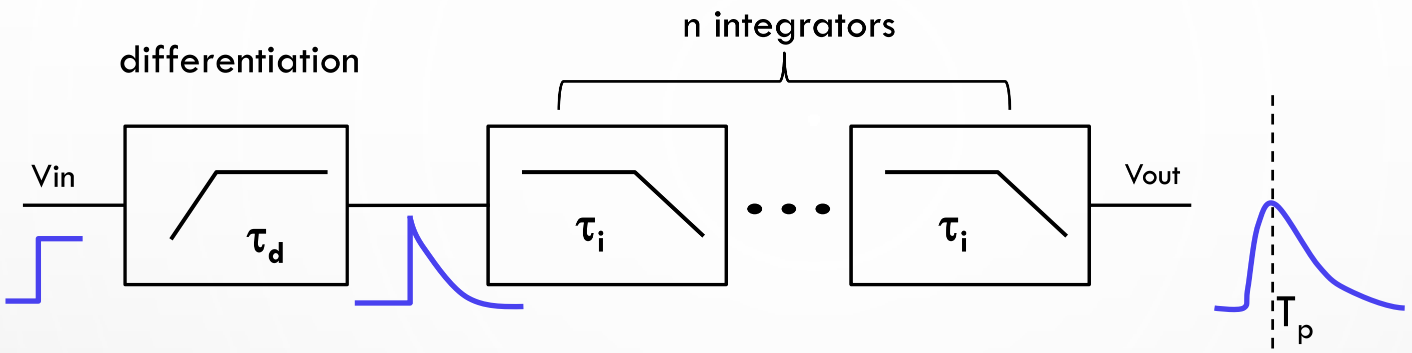
$$h_{opt}(t) = \exp\left(-\frac{|t|}{\tau_c}\right) \longrightarrow ENC^2_{opt} = 2KTR_S\frac{C_t^2}{\tau_c} = 2KTC_t\sqrt{\frac{R_s}{R_p}}$$

This function is known as cusp or matched filter (curve a in the figure)

The cusp filter is not practically feasible, but can be approximated by triangular shapers (curve b) or Pseudo-Gaussian shaper

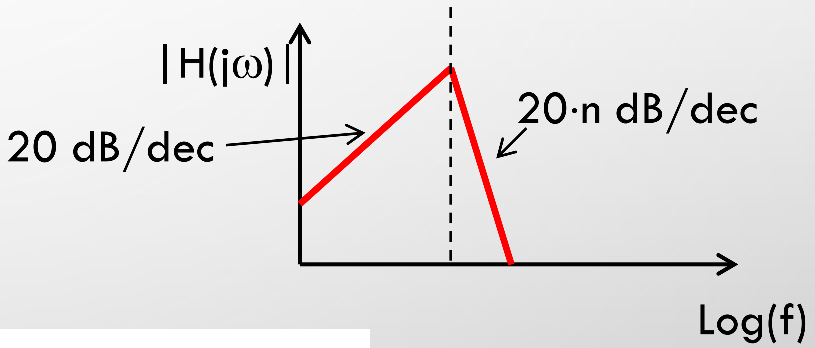


PSEUDO-GAUSSIAN SHAPER ($CR\text{-}RC^n$)

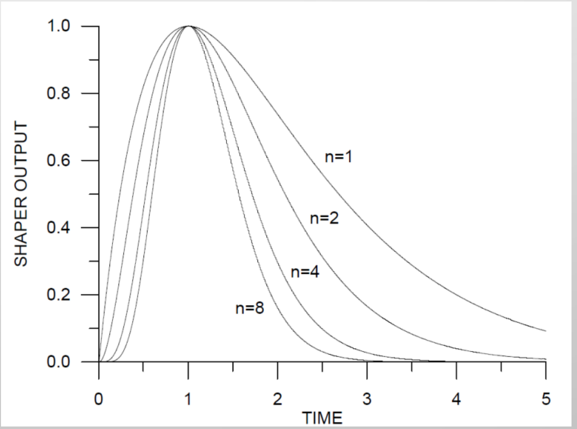


1. A high-pass filter, that makes the derivative of the input pulse and introduces the decay time τ_d
2. n low-pass filters, that limits the bandwidth (and the noise) making the integral of the signal and limiting the rise time τ_i (n is the order of the filter)

$$H(s) = \frac{u_{out}(s)}{u_{in}(s)} = \frac{s\tau_d}{(1+s\tau_d)} \frac{1}{(1+s\tau_i)^n}$$

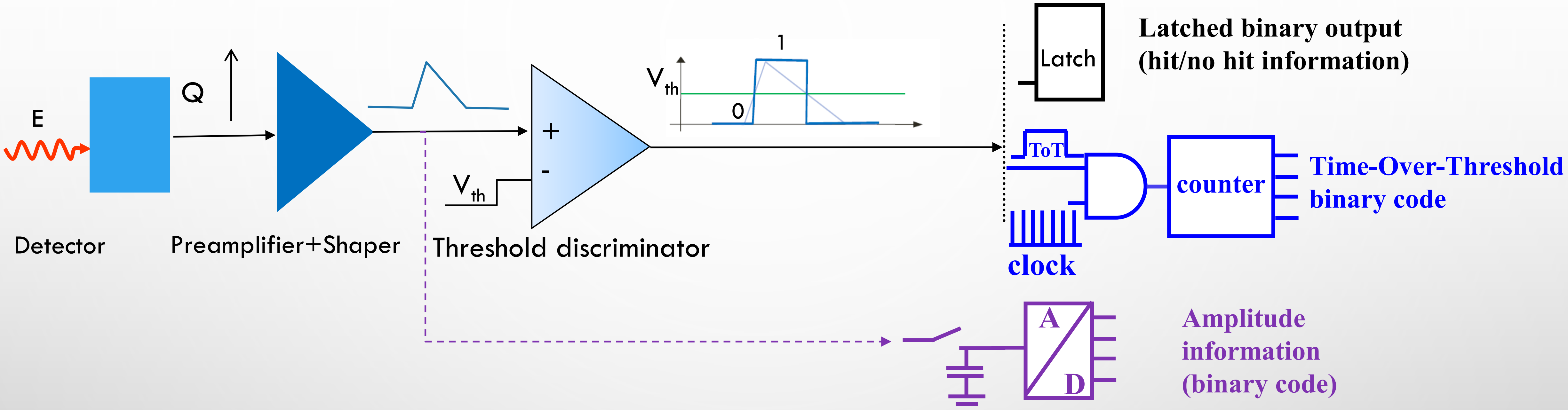


- Increasing the order keeping the same peaking time:
  1. More symmetrical
  2. Faster return to baseline
  3. Improved rate capability



HIT DISCRIMINATION AND TIME MEASUREMENT

HIT DISCRIMINATION AND ANALOG-TO DIGITAL CONVERSION



In a multichannel readout chip, channel-to-channel threshold variations due to device mismatches may degrade detection efficiency and spurious hit rate. Usually, the threshold dispersion is decreased by using:

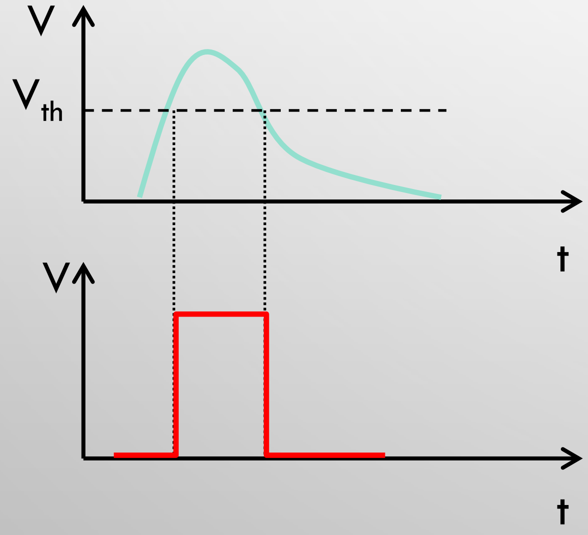
- Global threshold: common to all channels
- Local trimming threshold: channel equalization

TIME MEASUREMENT

- The hit discrimination technique contributes to the **timing resolution** and **timing accuracy**

- In HEP experiments, it is crucial to assign the hit to the correct bunch crossing (25ns at LHC)
- Sometimes, it is also requested to precisely measure the time of occurrence (tens or hundreds of ps) using Time to Digital Converters

The simplest scheme is based on: Leading edge or Threshold discriminator (comparator): when the signal crosses a threshold, the output goes from "low" to "high" level



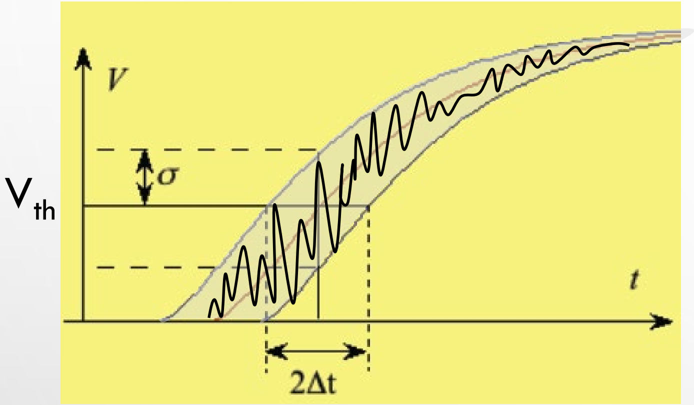
Timing measurement is limited by:

1. Jitter → Timing resolution
2. Time walk → Timing accuracy

THE JITTER

Noise has an impact in time measurements:

uncertainty in the time of crossing threshold → Jitter



$$\Delta t = \frac{\sigma_{noise}}{dV/dt}$$

slope

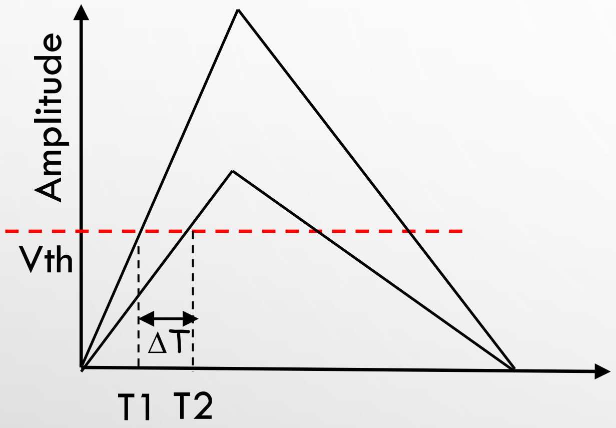
How to decrease jitter? → Conflicting conditions:

- decrease σ_{noise} → decrease bandwidth
- increase slope → increase bandwidth

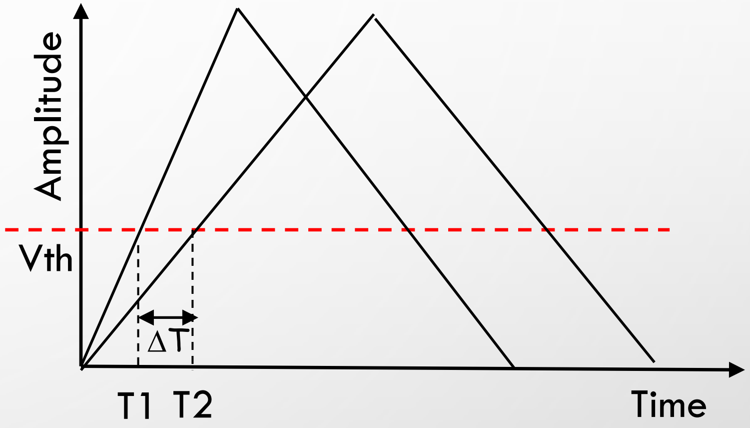
To minimize jitter, the best threshold level is where the slope is maximum

THE TIME WALK

In the leading edge discriminators, two pulses with identical shape and time of occurrence, but different amplitude cross the same threshold in different times: amplitude time walk



Even if the input amplitude is constant, time walk can still occur if the shape (rise time) of the pulse changes (for example, for changes in the charge collection time): rise time walk



The sensitivity of leading edge discriminator to time walk is minimized by setting the threshold as low as possible but it must be compatible with noise level

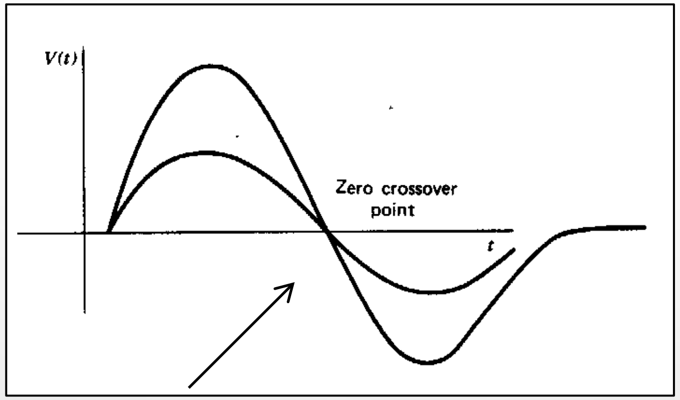
Time walk correction:

- Software: measure the pulse amplitude and apply correction to timing
- Hardware: instead of leading edge discriminator, use
  1. Crossover timing
  2. Constant Fraction timing

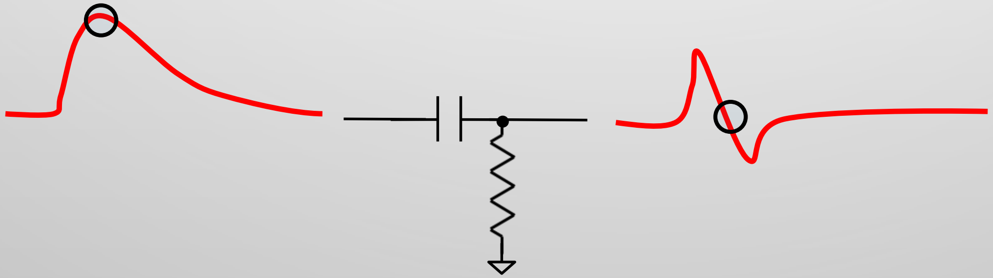
CROSSOVER TIMING

The crossover timing can greatly reduce the magnitude of the amplitude time walk

Hypothesis:
• the output of the shaper is a bipolar pulse and the time of zero-crossing is independent of the pulse amplitude

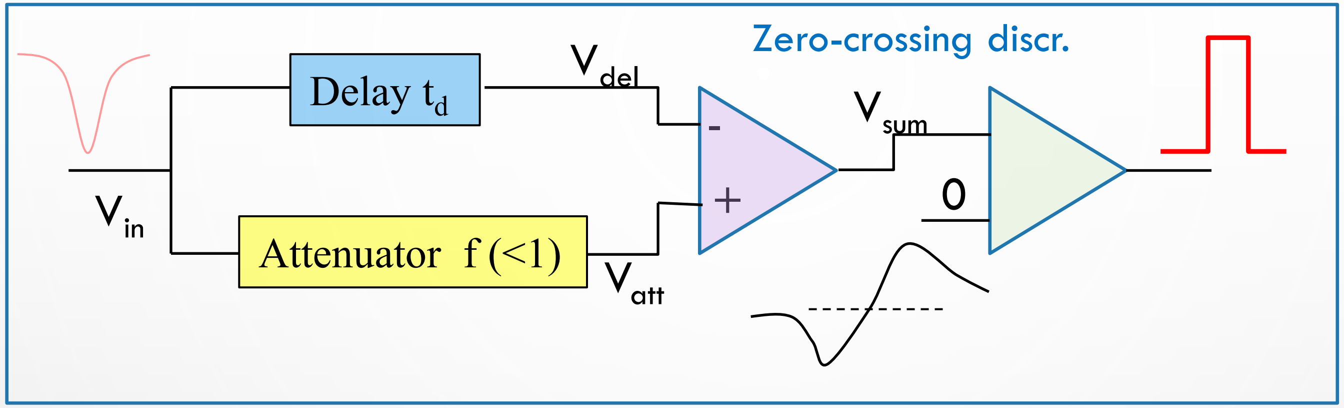


If the output of shaper is unipolar, but the peaking time is constant, adding a differentiator (C-R network) we get a bipolar pulse crossing the zero in correspondence of the signal peak



This method reduce amplitude time walk, but usually jitter is larger than leading edge triggering

THE CONSTANT FRACTION TIMING



Summing:
- inverted and delayed signal, with $t_d > t_{rise}$
- attenuated signal

It can be demonstrated that the zero-crossing time of resulting bipolar signal is independent of pulse amplitude for all pulses with constant shape
Resulting jitter for optimal parameters (t_d, f) is lower than cross-over discriminator technique

Summary

- The choice of Front-End electronics architecture is crucial to obtain the desired energy and/or time resolution

- The technology choice strongly depends on:
  - Radiation hardness
  - Speed
  - Density
  - Cost

- The most commonly used amplification chain is made of a *Charge Sensitive Amplifier*, followed by a Shaper
- Also Current and Voltage preamplifiers are often used, depending on the sensor characteristics and specific application

- The choice of pulse shape (and peaking time) comes out as a compromise between S/N optimization and double pulse resolution

- When the main goal is the time resolution, the *Constant Fraction Timing* provides the best results in terms of time walk, but requires higher circuital complexity respect to the simpler Leading Edge Timing and to the Zero-crossing Timing

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