

# Front-End electronics for particle detectors

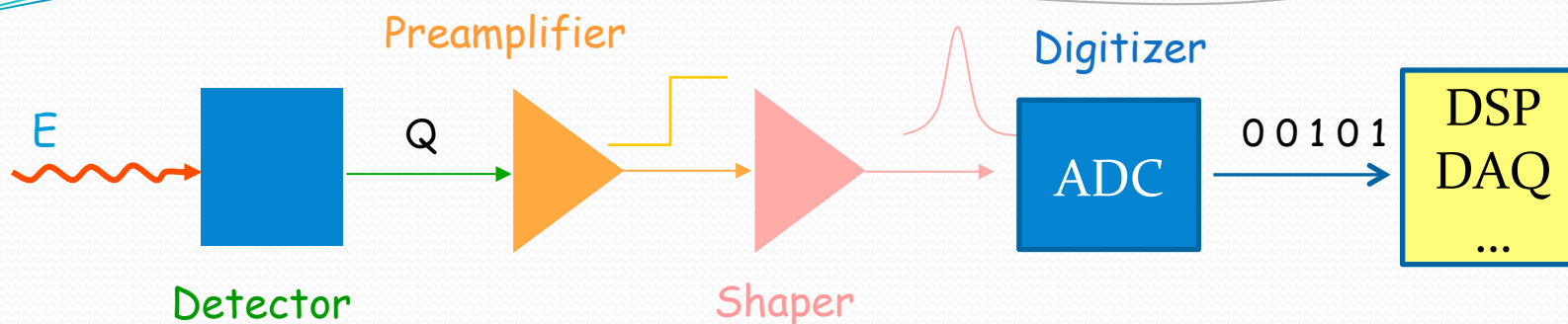
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# Outline

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- Detectors and signals
- Noise basic principles
- Radiation damage
- Front-end schemes
- Charge-sensitive amplifier
- Shapers
- Time measurement
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# Introduction

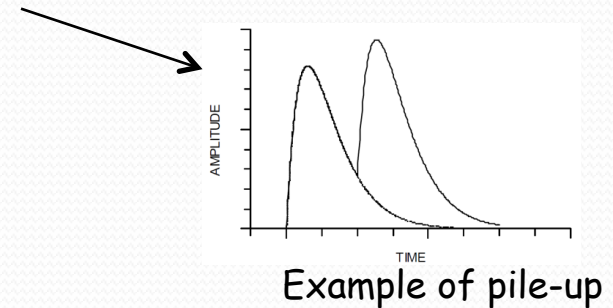


General readout architecture

- The particle deposits energy in a detecting medium  $\left\{ \begin{array}{l} \text{Gas} \\ \text{Solid} \\ \text{Liquid} \end{array} \right.$
- Energy is converted into an electrical signal:  $Q = KE$
- The charge  $Q$  is typically small and must be amplified, in order 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

# Purpose of Front-End Electronics

1. Acquire an electrical signal from the detector
2. Choose the gain and shaping time in order to optimize:
  - 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 hardness/tolerance;
  - power consumption;
  - cost



Often the requirements are in conflict each other



The final design comes out as a compromise, according to the specific application:

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





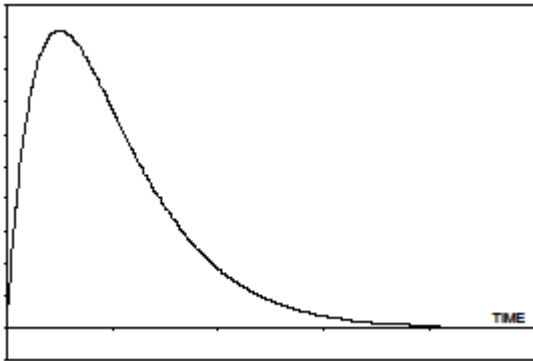
# **NOISE BASIC PRINCIPLES**

# Noise

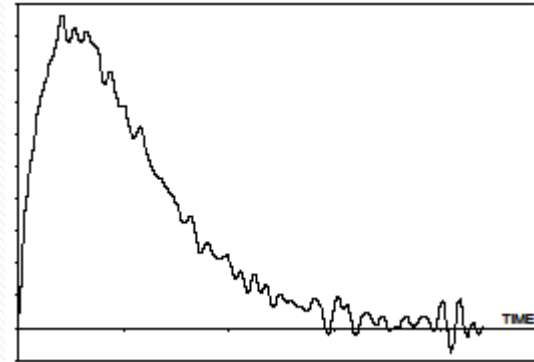
The precision of amplitude and timing measurements is limited by the **NOISE**

## 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 eliminated by proper shielding, cabling ...

## 2. Intrinsic noise

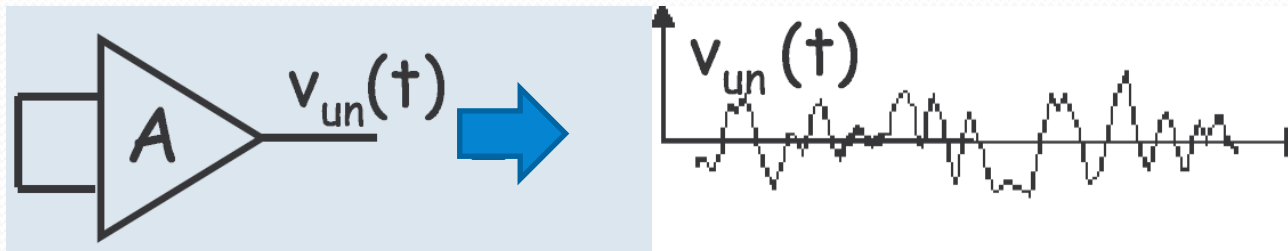
It is a property of detector and/or electronics

Can be reduced by proper design of front-end electronics

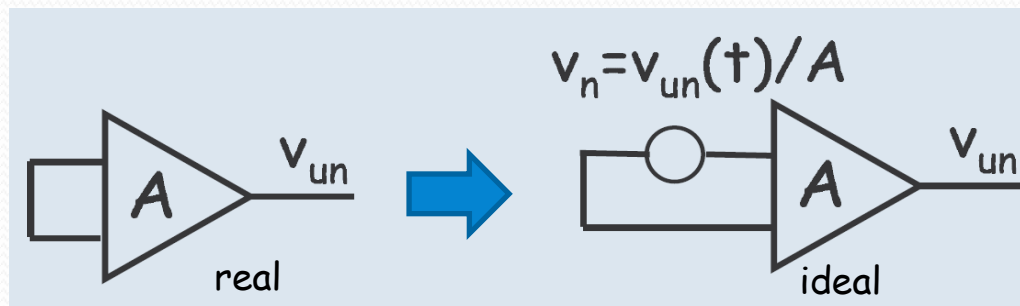
# Intrinsic noise

The output voltage of a real amplifier is never constant, even if  $V_{in} = 0$

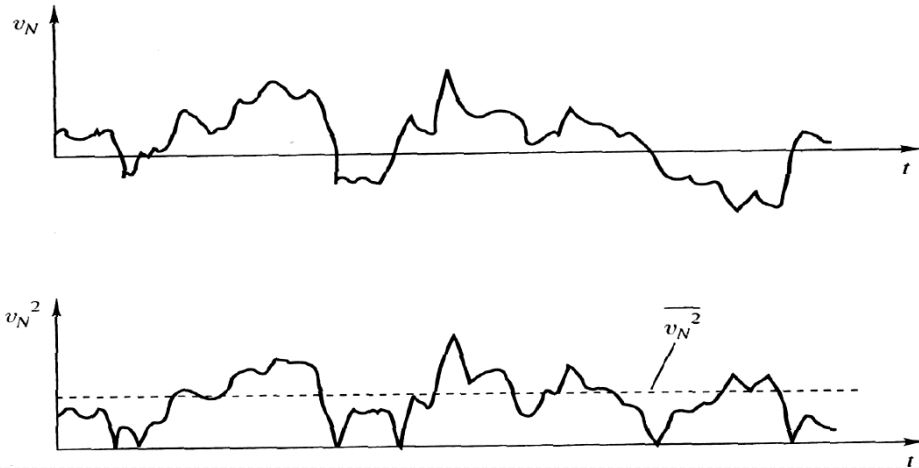
The fluctuations of  $V_{un}(t)$  when  $V_{in} = 0$  correspond to the noise of amplifier



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



# Intrinsic noise



$V_n$  has mean value = 0, but power  $\neq 0$



$$V_n^2$$

We can define:

- Source of **voltage noise**:

$$v_n = \sqrt{v_n^2(f)}$$

- Source of **current noise**:

$$i_n = \sqrt{i_n^2(f)}$$

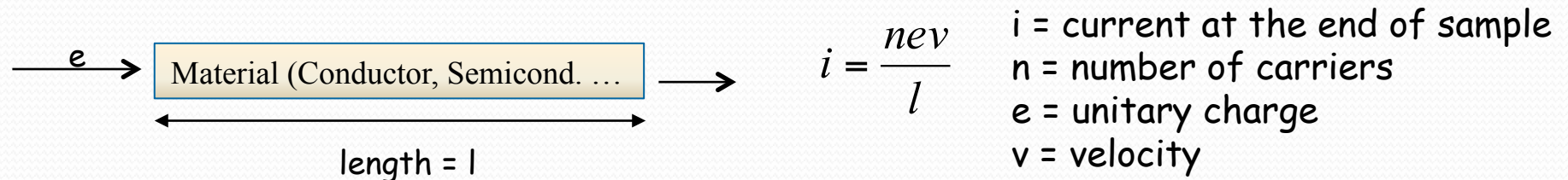
A noise source is usually defined by its **POWER SPECTRAL DENSITY** : noise power per unit of bandwidth

$$\frac{dv_n^2}{df}$$

$$\frac{di_n^2}{df}$$

If Power Spectral Density is constant  $\rightarrow$  **White Noise**

# Basic noise mechanisms



The fluctuation of the current is given by:

$$\langle di \rangle^2 = \left( \frac{ne}{l} \langle dv \rangle \right)^2 + \left( \frac{ev}{l} \langle dn \rangle \right)^2$$

There are two basic mechanism contributing to noise:

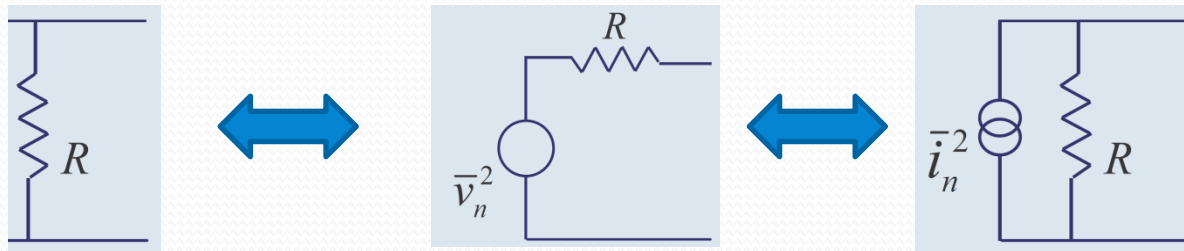
Velocity fluctuations  $\rightarrow$  Thermal noise

Number fluctuations  $\left\{ \begin{array}{l} \text{Shot noise} \\ \text{Excess (or flicker, or "1/f") noise} \end{array} \right.$

# 1. Thermal noise (Johnson noise)

It is typical of resistors

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

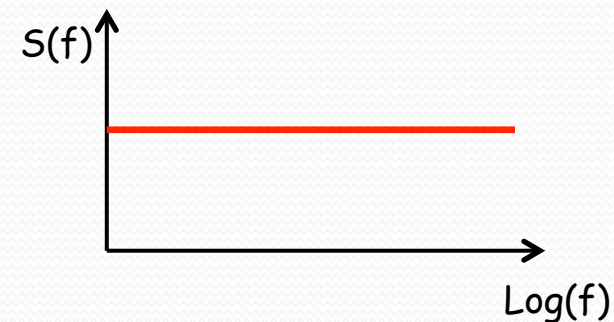


A real (noisy) resistor is equivalent to an ideal (noiseless) resistor + noise source (voltage or current)

$$S_v(f) = \frac{dv_n^2}{df} = 4kTR$$

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

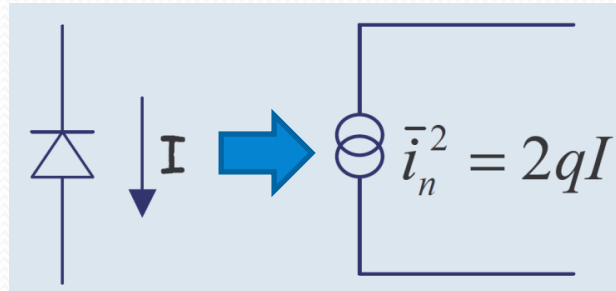
$k$  = Boltzmann constant =  $1.3806503 \times 10^{-23}$  J/K  
 $T$  = absolute temperature  
 $R$  = resistance



**Does not depend on  $f \rightarrow$  Thermal noise is a white noise**

## 2. Shot noise

It is caused by fluctuations in the number of charge carriers, for example in the current flowing in a semiconductor diode or transistor, where  $e/h$  cross a potential barrier



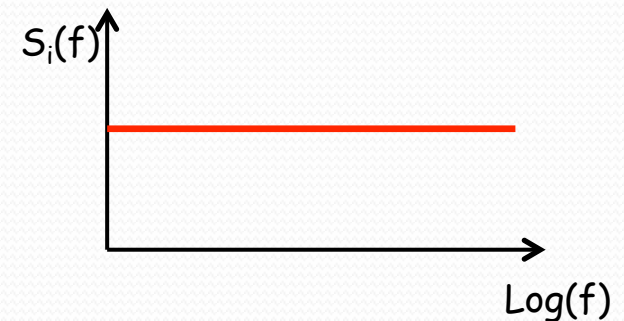
Power spectral density:

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

does not depend on  $f \rightarrow$  also shot noise is white  
(but a current  $I$  must be present)

Example: consider a reversed-biased diode, with leakage  $I = 1 \text{ nA}$

$$S_i(f) = \frac{d\bar{i}_n^2}{df} = 2 * 1.6 * 10^{-19} * 10^{-9} = 3.2 * 10^{-28} \text{ A}^2 / \text{Hz}$$



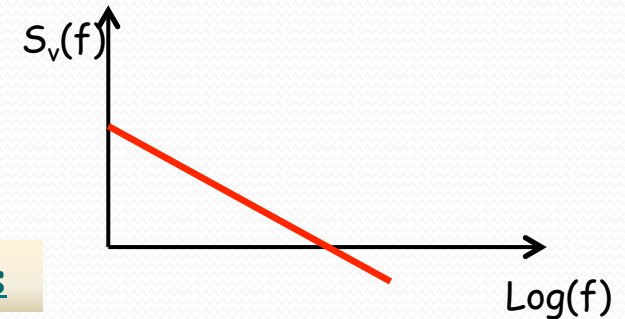
### 3. Flicker noise (1/f noise)

It is associated to random trapping and recombination of charge carriers in the semiconductors, typically caused by imperfections in the interface regions. It is also present in carbon resistors

Power spectral density:

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

$$\left\{ \begin{array}{l} I \text{ is dc current} \\ K_f \text{ is a constant (vary from device to device)} \\ a \sim 0.5 \div 2 \\ b \sim 1 \end{array} \right.$$



It depends on  $f$  and clearly it is important at low frequencies

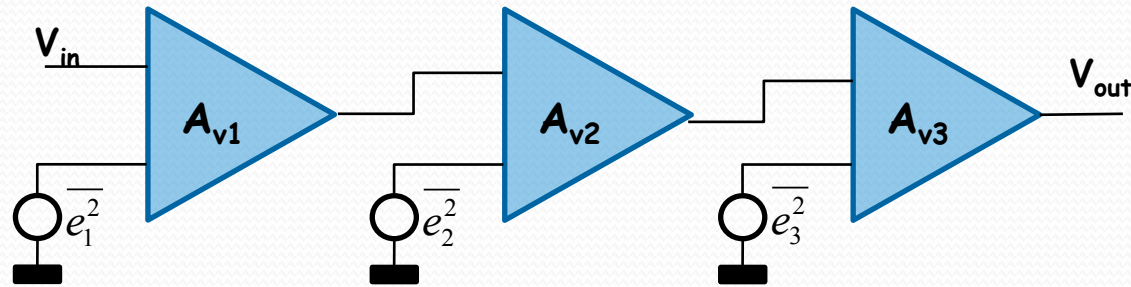
### 4. Burst noise (POPCORN noise)

Another low-frequency noise. It can be found in some integrated circuits and discrete transistor and is associated to contamination by ions of heavy metals (i.e. Au).

$$\frac{d\bar{i}_b^2}{df} = K_b \frac{I_b^c}{1 + (f / f_c)^2}$$



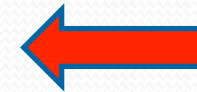
# Intrinsic noise: importance of first stage



$$\begin{cases} V_{out} = A_{v1} * A_{v2} * A_{v3} * V_{in} \\ \overline{e_{out}^2} = A_{v1}^2 * A_{v2}^2 * A_{v3}^2 * \overline{e_1^2} + A_{v2}^2 * A_{v3}^2 * \overline{e_2^2} + A_{v3}^2 * \overline{e_3^2} \end{cases}$$



$$\left( \frac{Noise}{Signal} \right)^2 = \left( \frac{\overline{e_{out}^2}}{V_{in}^2} \right) = \frac{\overline{e_1^2} + \frac{\overline{e_2^2}}{A_{v1}^2} + \frac{\overline{e_3^2}}{A_{v1}^2 * A_{v2}^2}}{V_{in}^2}$$



1. It is necessary to decrease as much as possible the noise contribution  $\overline{e_1^2}$  of the first stage
2. It is necessary to increase the gain  $A_{v1}$  of the first stage because the noise contribution of next stages are divided by the gain of previous stages

# Intrinsic noise: some practical rules

1. Uncorrelated noise sources must be added in quadrature

$$\overline{e_{tot}^2} = \overline{e_1^2} + \overline{e_2^2} + \overline{e_3^2} + \dots$$

2. In an amplifying chain, the noise generated in the first stage dominates

In first (and good) approximation, it is enough to evaluate (and decrease) the noise of the first stage

3. It is useful to represent a real (noisy) amplifier as an ideal (noiseless) amplifier with an **equivalent noise source** at its input: in this way the noise can be directly compared with input signal
4. In the case of particle detection systems, where the input is a charge  $Q$ , we use **ENC: Equivalent Noise Charge** : it is the signal magnitude 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 our amplifier



# The problem of radiation damage

# 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 signal
- Modification of memory content
- Degradation of performance
- Catastrophic failure

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

- **Ionization damage**: charged particles produce transient currents (drift, diffusion) and entrapment of charge in  $\text{SiO}_2$

Total dose (TID) → Threshold shift, parasitic leakage currents, mobility degradation

Single Event Effects (SEE) → temporary or permanent errors

# Radiation-hard electronics: which technology?

- **JFET** (no gate oxide → no entrapment of charged carriers)
- **Rad-hard CMOS** (mainly used in the '90, very expensive)
- **Deep submicron CMOS** (scaling the oxide thickness the rad-immunity is increased)
  - LHC (CMOS 0.25  $\mu\text{m}$ )
  - LHC- Upgrade Phase1 (CMOS 0.13  $\mu\text{m}$ )
  - HL-LHC (CMOS 0.065  $\mu\text{m}$ )
- **SiGe** (RF bipolar transistors: high doping levels guarantees adequate current gain and reduce the sensitivity of the surface carrier concentration to radiation-induced charge in the oxide)

# Monolithic technologies for custom applications

- **Bipolar**

- used mainly in the past
- today few foundries
- more speed and less power in analog applications
- low integration

- **BiCMOS**

- used mainly in the past
- combines advantages of bipolar and CMOS
- complex fabrication process

- **Silicon on Insulator (SOI)**

- used (mainly in the past) for rad-hard applications
- expensive

- **GaAs**

- not suitable for analog applications with high bandwidth

- **SiGe**

- RF bipolar transistors: high speed
- expensive
- rad-hard

**Most used technology is CMOS**

# CMOS technologies for Front-End electronics

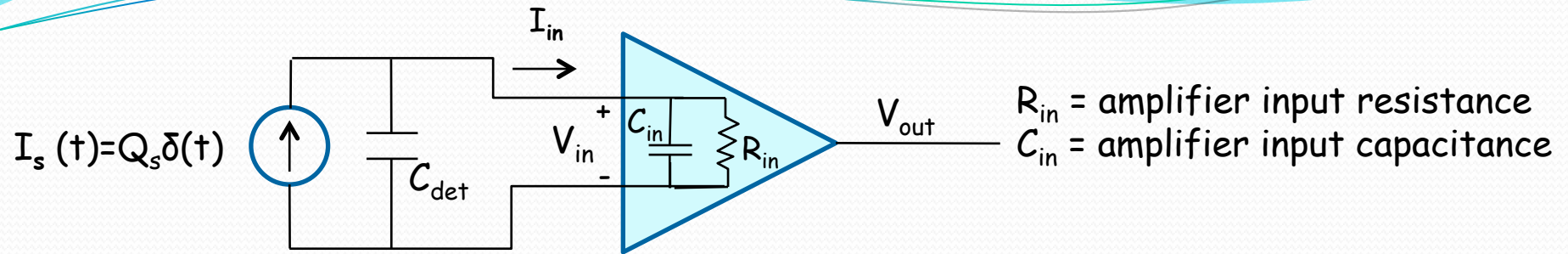
- Large part of Front-End electronics are developed in **CMOS** technology
- Relatively “cheap” if recent/“old” techn. are used
- Using the “multiproject foundry runs”, prototyping and small productions are very affordable
- Suitable to combine on the same chip analog section, digital part and  $\mu$ processors
- Very low power consumption
- The new deep submicron CMOS tech. ( $< 130$  nm) are **rad-hard** and suitable for HL-LHC, ILC, Space applications



# FRONT-END SCHEMES



# Signal integration



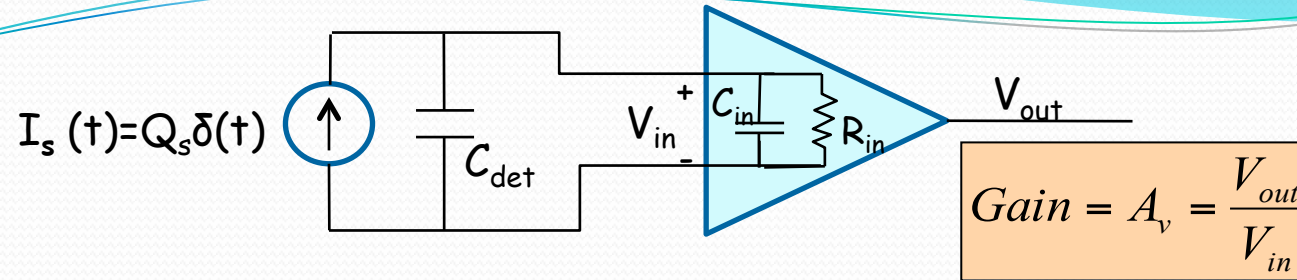
- The sensor signal is usually a short current pulse  $I_s(t) = Q \cdot \delta(t)$  with duration ranging from few hundreds of ps, as in Si sensors and Resistive Plate Chambers to tens of  $\mu s$ , as in inorganic scintillators
- The physic quantity of interest is the deposited energy  $E$ , that is 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?

OPTIONS (depending on charge collection time  $t_c$  and input time constant  $R_i C_i$ ):

1. Detector capacitance  $\rightarrow V_{in} \propto Q_s \rightarrow$  followed by voltage amplifier
2. Current sensitive amplifier  $\rightarrow V_{out} \propto I_s \rightarrow$  followed by integrating Analog-to-Digital Converter
3. Charge sensitive amplifier  $\rightarrow V_{out} \propto Q_s$

# 1. Integration on $C_{det}$ (+ voltage amplifier)



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

the detector capacitance discharge slowly

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

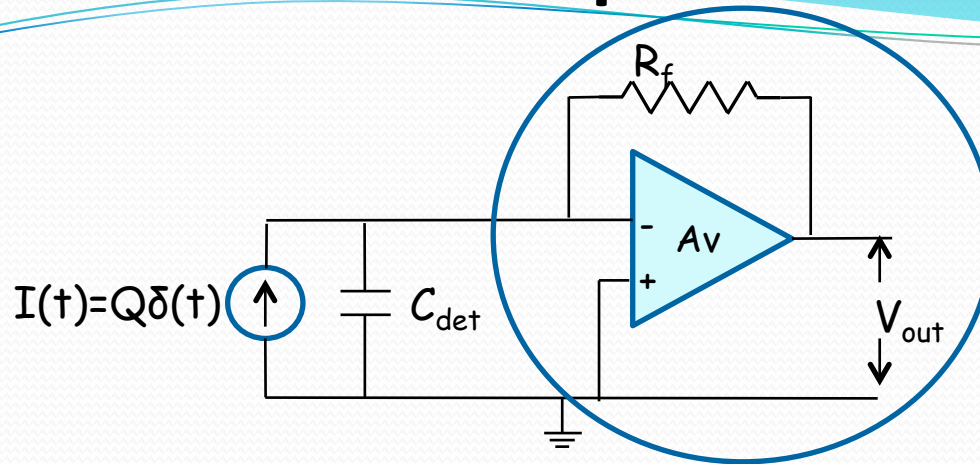
$$V_{in} = \frac{1}{C_t} \int I_s dt = \frac{Q_s}{C_{det} + C_{in}}$$

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

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:  $\left\{ \begin{array}{l} \text{different strip length/width} \\ \text{bias voltage} \\ \dots \end{array} \right.$

## 2. Current-sensitive amplifier



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



The detector capacitance discharges rapidly  $\rightarrow$  the amplifier senses the current



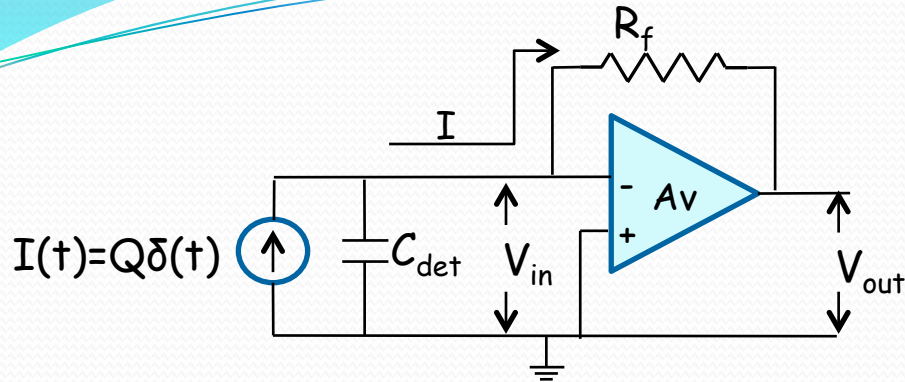
Using a transresistance amplifier (high gain operational amplifier with resistive feedback) we have:

$$V_{out} \propto I$$

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

An integrating ADC can follow the amplifier and provide a digital word proportional to  $Q$

# Current-sensitive amplifier: the ideal case



Hypothesis:

1. input impedance of op-amp is  $\infty$  (i.e. MOS gate)  
 $\rightarrow$  all current flows in the feedback
2.  $A_v$  is very big

## GAIN

$$\begin{cases} V_{out} = -A_v V_{in} \\ V_{out} - V_{in} = -R_f I \end{cases} \Rightarrow V_{out} + \frac{V_{out}}{A_v} = -R_f I \Rightarrow V_{out} = -R_f I \frac{A_v}{1 + A_v} \xrightarrow{A_v \gg 1} V_{out} \approx -R_f I$$

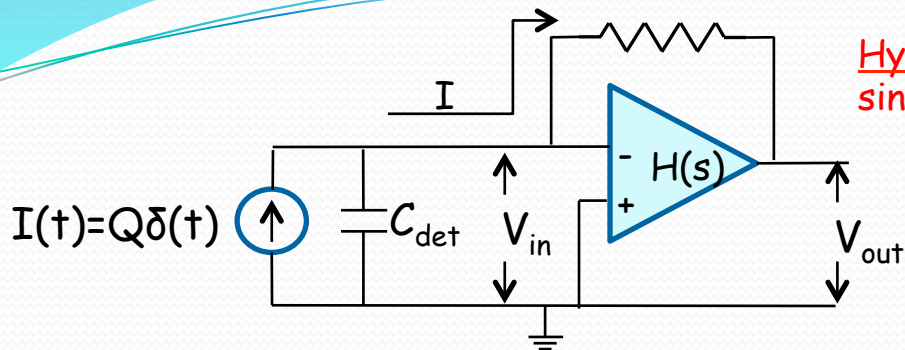
Main advantage: the shape of input signal is preserved  $\rightarrow$  shaping stage not strictly necessary

## INPUT IMPEDANCE

$$Z_{in} = \frac{V_{in}}{I} = \frac{R_f}{1 + A_v}$$

Since  $A_v$  is very big,  $Z_{in} \rightarrow 0 \Rightarrow$  Our hypothesis that  $\tau_{in} \ll$  pulse duration is confirmed

# Current-sensitive amplifier: the realistic case



Hypothesis:  
single pole transfer function

$$H(s) = \frac{g_m R}{1 + sRC}$$

$g_m$  = transconductance of the ampl.

$RC$  = internal load

$s$  = complex frequency =  $j\omega = j2\pi f$

$$V_{out}(s) = -R_f I \cdot \frac{H(s)}{1 + H(s)} = -R_f I \cdot \frac{g_m R}{(1 + g_m R) + sRC}$$

At low frequency

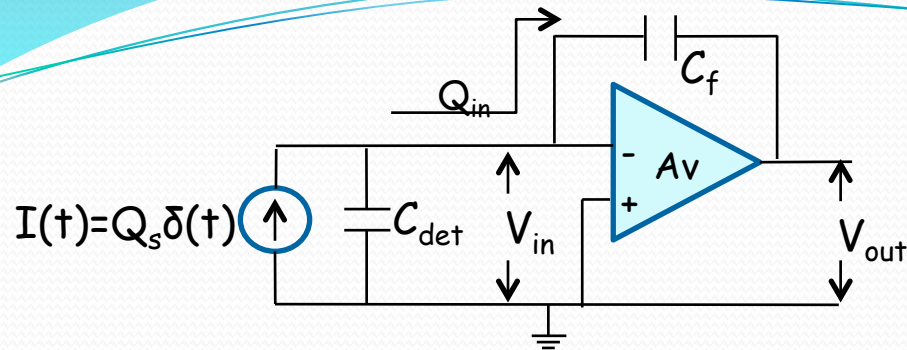
$$V_{out}(s) = -R_f I \cdot \frac{g_m R}{1 + g_m R} \approx -R_f I$$

$$Z_{in}(s) = \frac{V_{in}}{I} = \frac{R_f}{1 + H(s)} = R_f \cdot \frac{1 + sRC}{(1 + g_m R) + sRC} \approx R_f \cdot \frac{s + \frac{1}{RC}}{s + \frac{g_m}{C}}$$

$$\longrightarrow \begin{cases} \text{Zero} = 1/RC \\ \text{Pole} = g_m/C \end{cases}$$

The presence of a zero in  $Z_{in} \rightarrow$  inductive behavior  $\rightarrow$  possible instabilities (oscillations) for some values of  $C_{det}$

### 3. Charge-sensitive amplifier (CSA): the ideal case



Hypothesis:

1. input impedance of op-amp is  $\infty$  (i.e. MOS gate)  
 $\rightarrow$  all current flows in the feedback
2.  $A_v$  is very big

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  $C_f$ :

$$Q_f = C_f V_f = C_f (A_v + 1)V_{in} = Q_{in} \quad (\text{for Hypothesis 1})$$

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

**GAIN (Charge Sensitivity):**

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

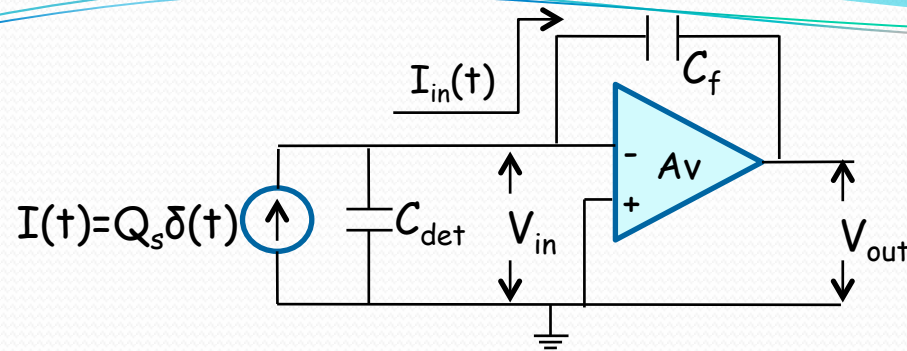
**BUT ... not all the charge goes in the amplifier and is measured: a small fraction remains on  $C_{det}$  !!!**

**Charge transfer efficiency:**

$$\frac{Q_{in}}{Q_S} = -\frac{Q_{in}}{Q_{det} + Q_{in}} = \frac{1}{1 + \frac{Q_{det}}{Q_{in}}} = \frac{1}{1 + \frac{C_{det}}{C_{in}}} \approx 1 \quad (\text{if } C_{in} = C_f (A_v + 1) \gg C_{det})$$

**Example:**  $C_{det} = 10 \text{ pF}$   $A_v = 10^3$   $C_f = 1 \text{ pF} \rightarrow C_{in} = 1 \text{ nF} \rightarrow Q_{in}/Q_S = 0.99$

# Charge-sensitive amplifier: the time response



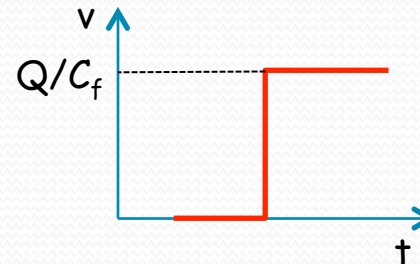
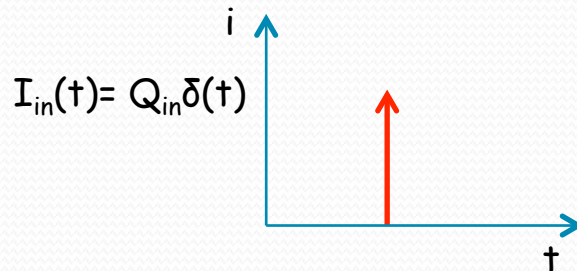
In the frequency domain:

$$V_{out}(\omega) = -A_v V_{in}(\omega) \quad (\text{assuming } A_v \text{ constant and } \rightarrow \infty \text{ (infinite bandwidth)})$$

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

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

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

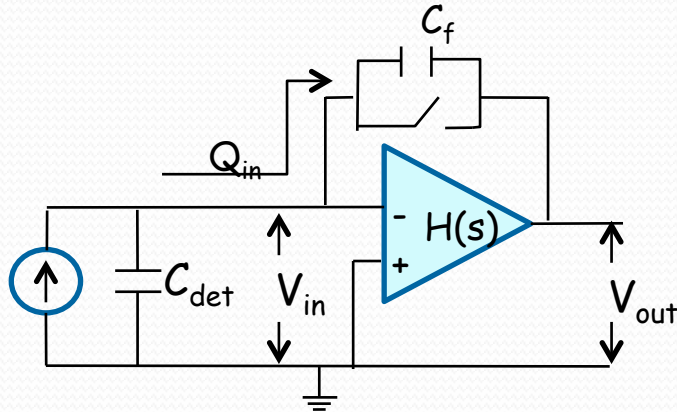


Step function



# Charge-sensitive amplifier: the reset

## Pulsed RESET

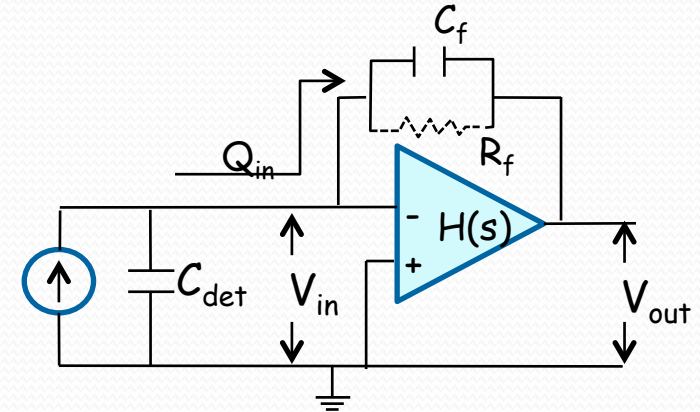


- The reset switch allows the removal of charge stored in  $C_f$
- The switch can be closed periodically or driven by some control signal

### Drawbacks:

- Dead time
- Switch noise
- Leakage current

## Continuous RESET



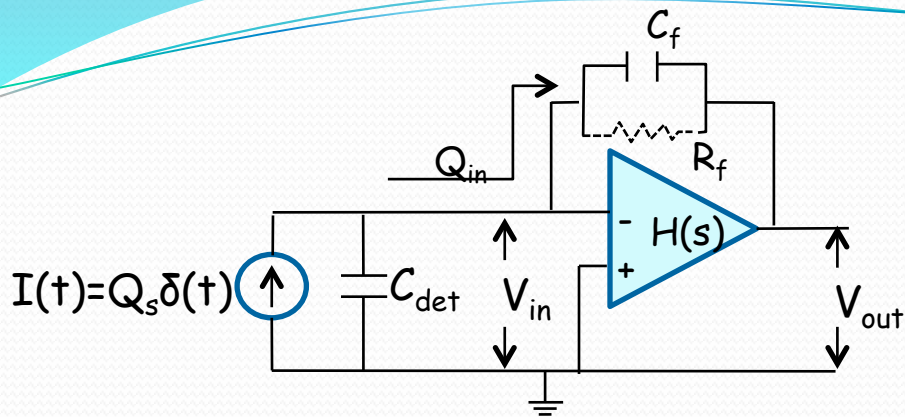
- The resistor  $R_f$  continuously discharges  $C_f$  after the pulse
- Discharge time constant  $R_f C_f$

### Drawbacks:

- Additional parallel noise
- Long tail  $\rightarrow$  Risk of pile-up



# Charge-sensitive amplifier: the realistic case



## Two elements to be considered:

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 be made very large to decrease its contribution to noise. Typical values are several tens or hundreds of  $M\Omega$
2. **Real amplifier** (finite bandwidth and gain,  $C_L$ )

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

2 poles

$$\begin{cases} p_1 = \frac{1}{R_f C_f} \\ p_2 = \frac{1}{R_i C_T} = \frac{\omega_0 C_f}{C_T} \end{cases}$$

(Low freq)

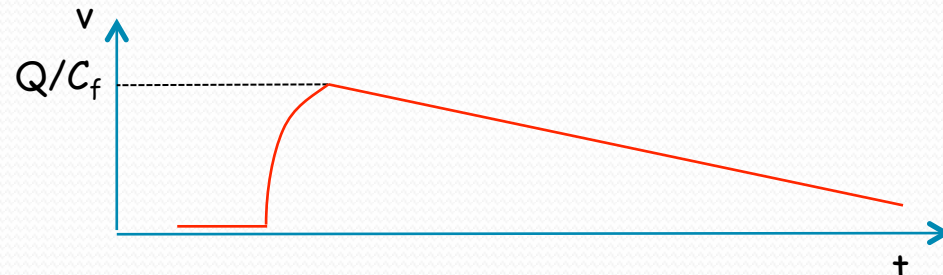
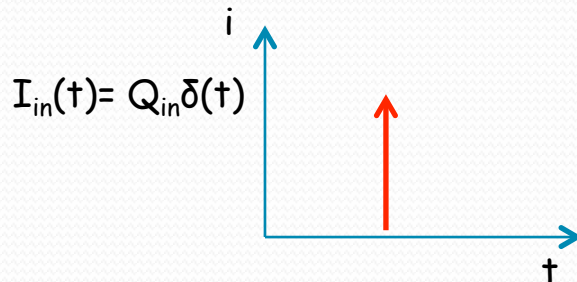
(High freq)

$$\tau_1 = R_f C_f$$

Fall time constant

$$\tau_2 = R_i C_T = \frac{C_T}{\omega_0 C_f}$$

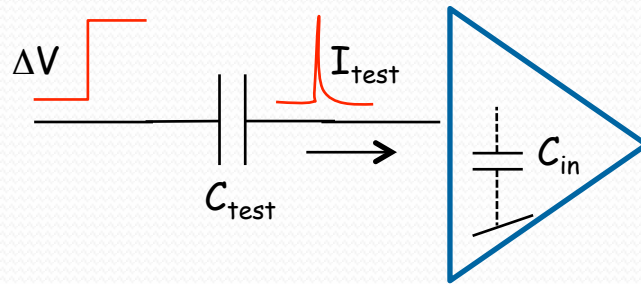
Rise time constant



- The fall time depends on the feedback: can be very large, since  $R_f$  must be very high for low noise ( $\gg 1 M\Omega$ )
- The rise time depends on the input time constant, thus
  - $R_i$  must be small to have short rise time
  - $\omega_0$ : the amplifier GBW must be very large
  - $C_T \rightarrow C_d$ : the rise time increase with detector capacitance

# Calibration

A common technique used for calibration and test is to inject a **voltage step** through a capacitor



$$\text{Ideally: } Q_{test} = C_{test} * \Delta V$$

But we must consider the dynamic input capacitance  $C_{in}$  of preamp  $\rightarrow C_{test}$  and  $C_{in}$  are in series

$$Q_{test} = C_{eq} * \Delta V = \frac{1}{\frac{1}{C_{test}} + \frac{1}{C_{in}}} * \Delta V = \frac{C_{test}}{1 + \frac{C_{test}}{C_{in}}} * \Delta V$$

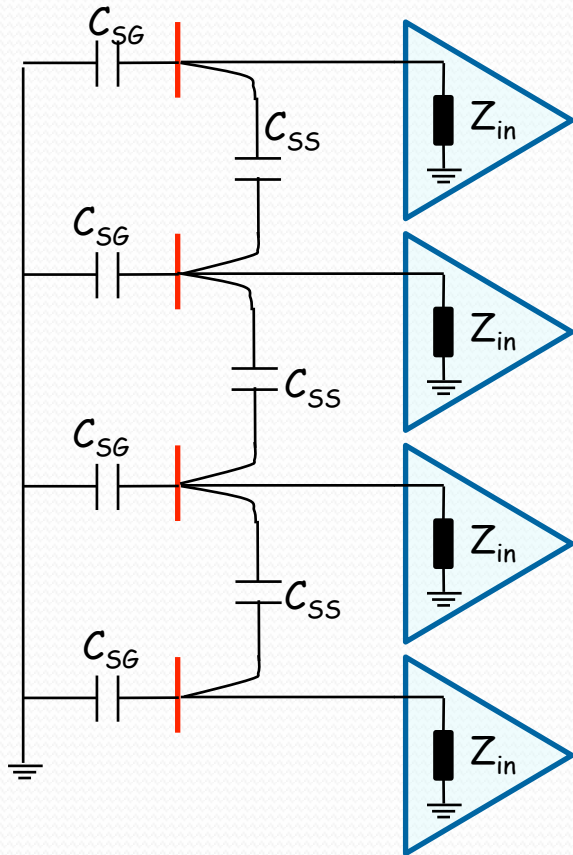
But usually  $C_{in} \gg C_{test} \rightarrow C_{test}/C_{in} \sim 10^{-3}$

$$Q_{test} \sim C_{test} * \Delta V$$

Example:  $C_{test} = 1 \text{ pF}$ ,  $\Delta V = 1 \text{ V} \rightarrow Q_{test} = 1 \text{ pC}$

\*In fact, in case of CSA,  $C_{in} = Q_{in}/V_{in} = C_f (A_v + 1)$

# Input impedance vs crosstalk



← Strip or pad

In strip or pixel detectors, where there are many adjacent channels, we must consider the following capacitive coupling:

- Strip or pad vs ground  $C_{SG}$
- Inter-strip capacitance  $C_{SS}$

If  $Z_{in} \gg Z_{ss} = \frac{1}{\omega C_{SS}}$  the charge induced on one strip is coupled into the adjacent channels through  $C_{SS}$   
 The number of affected signal depends on  $\frac{C_{SS}}{C_{SG}}$

If  $Z_{in} \ll Z_{ss} = \frac{1}{\omega C_{SS}}$  most part of the charge flows into the amplifier and only small part is coupled into the adjacent channels through  $C_{SS}$

## Summary:

low input impedance →

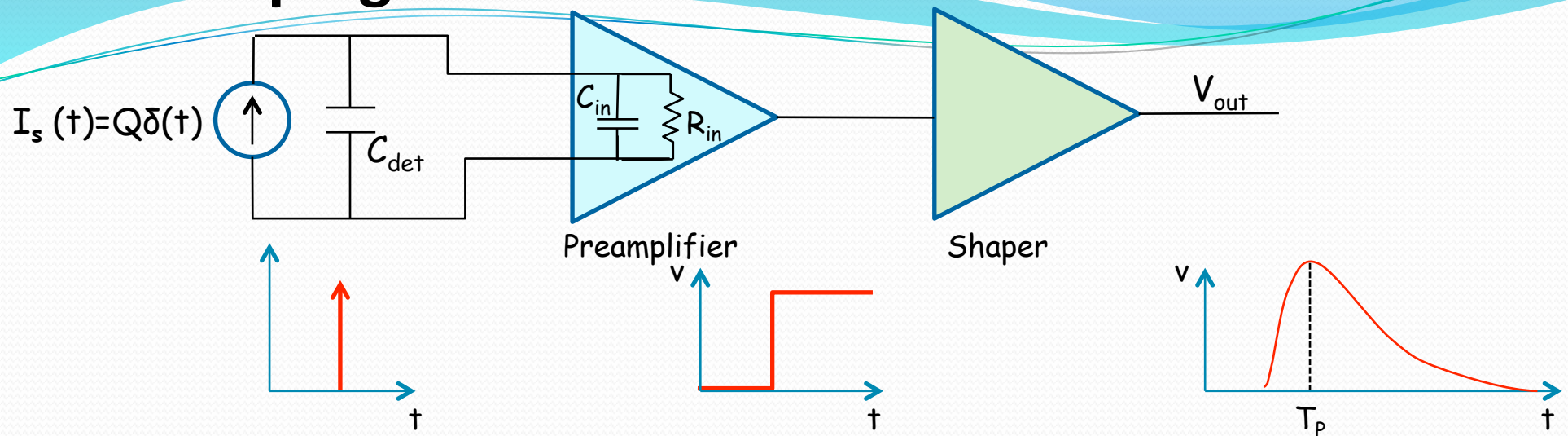
- Short rise time
- Small cross-talk



# NOISE FILTERING: SHAPERS

†

# Pulse shaping



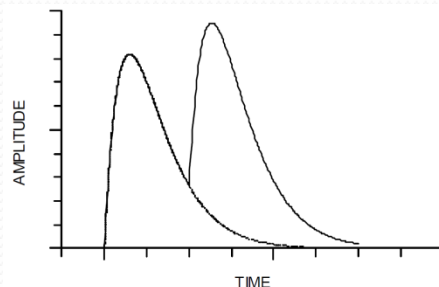
Preamplifier = input amplifier It is usually located close to detector and must have enough gain to make negligible the effects of induced noise. Typical example: Charge Sensitive Amplifier

Shaper Two conflicting objectives:

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

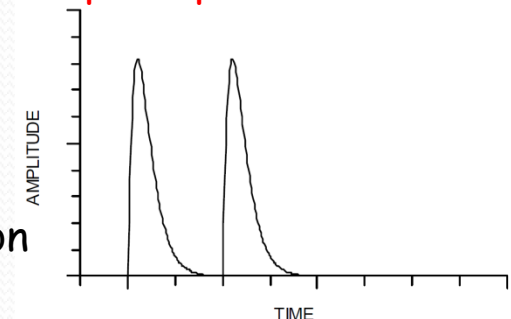
Slower pulse:

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



Faster pulse:

- More noise
- Double-pulse resolution



The choice of the shaper ( $T_p$ , shape) derives from a compromise between the two targets

# Noise through filters

$$\overline{v_n^2} \xrightarrow{H(j\omega)} \overline{v_u^2} = \overline{v_n^2} * |H(j\omega)|^2$$

$\omega = 2\pi f$

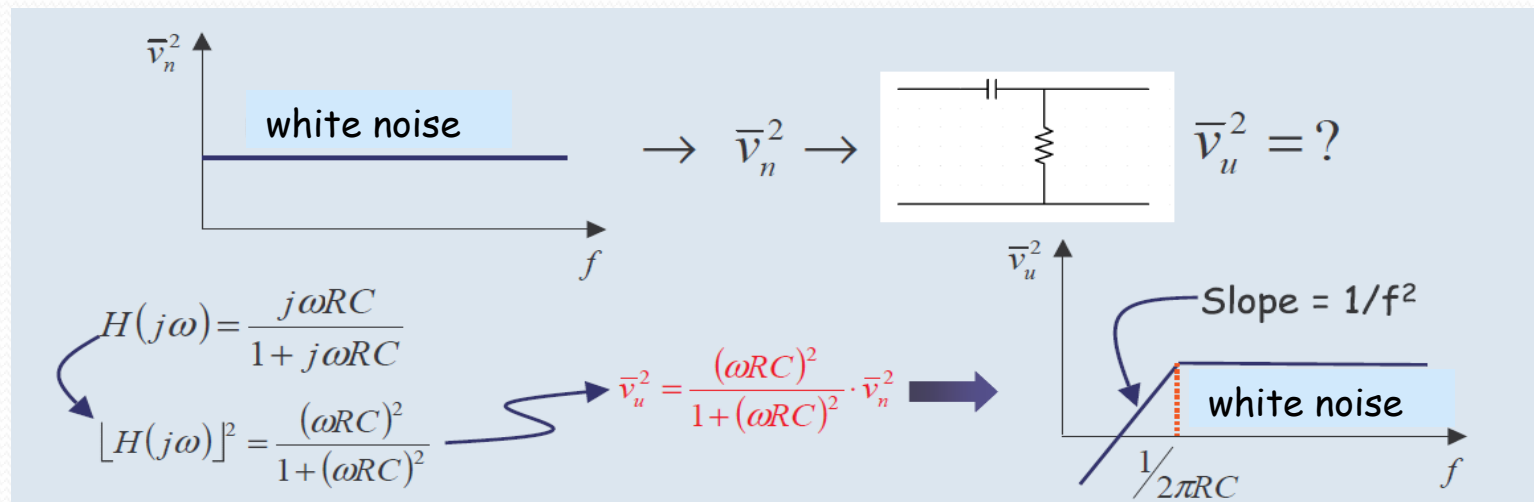
Noise power spectrum at output of a filter with transfer function  $H(j\omega)$  is equal to input power spectrum multiplied by squared transfer function

The total noise depends on the bandwidth of the system. Since spectral noise components are non-correlated, we must integrate the noise power over the frequency range of the system

$$v_{on}^2 = \int_0^{\infty} \overline{v_{un}^2} d\omega = \int_0^{\infty} \overline{v_n^2} * |H(\omega)|^2 d\omega$$

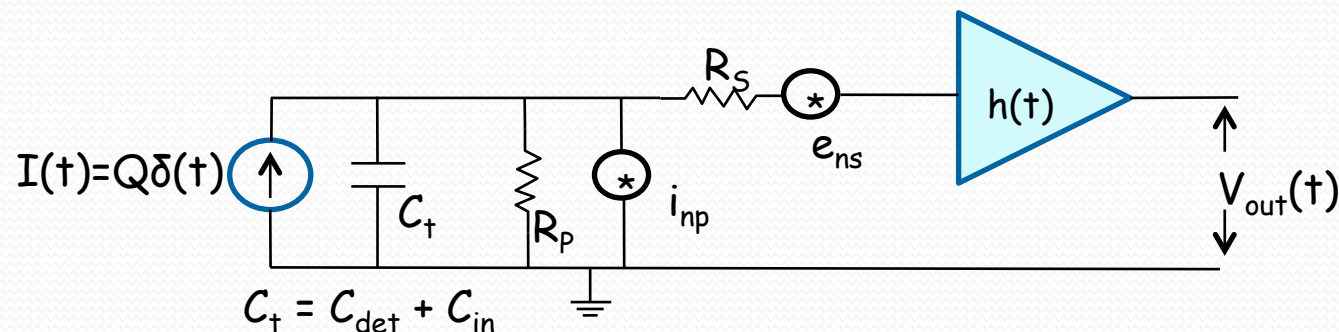
- The total noise increases with bandwidth
- Small bandwidth  $\rightarrow$  large rise-times  $\rightarrow$  less noise
- High bandwidth  $\rightarrow$  fast pulse  $\rightarrow$  more noise

## Example: white noise source connected to high-pass filter



# Optimum filter

In order to study the ENC and find the optimum filter (transfer function) of our amplifying system, it is convenient to represent our chain with a noiseless amplifier, with transfer function  $h(t)$  and all noise sources at its input, represented by  $R_s$  and  $R_p$  (we are considering only white noise source, not  $1/f$  for the moment)



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

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

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

in general  $\rightarrow R_s = \frac{a_n}{g_m}$       $a_n = \begin{cases} 0.5 \text{ in BJT} \\ 0.7 \text{ in Mosfet} \end{cases}$

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

It is possible to demonstrate that:

$$ENC^2 = 2KTR_s C_t^2 \int \left[ \frac{d}{dt} h(t) \right]^2 dt + \frac{2KT}{R_p} \int [h(t)]^2 dt$$

series noise

parallel noise

$$ENC^2 = 2KTR_s C_t^2 \left[ \int [h'(t)]^2 dt + \frac{1}{\tau_C^2} \int [h(t)]^2 dt \right]$$

$$\text{where } \tau_C = C_t \sqrt{R_p R_s}$$

noise corner time constant



# Capacitive matching

- Parallel noise depends mainly by “external” factors (Feedback resistor, detector bias and leakage)
- **Series noise depends on amplifier characteristics ( $R_s \rightarrow g_m, C_{in}$ )**

with proper design and dimensioning of preamp we can optimize  $ENC_s$

$$ENC_s^2 = 4KTR_s C_t^2 \frac{1}{t_m} = 4KT \frac{a_n}{g_m} (C_{det} + C_{in})^2 \frac{1}{t_m} = 4KT a_n C_{det} \frac{\tau_A}{t_m} \left[ \sqrt{\frac{C_{det}}{C_{in}}} + \sqrt{\frac{C_{in}}{C_{det}}} \right]^2$$

The minimum value is when  $C_{det} = C_{in}$

Input transistor capacitance must be matched to detector capacitance

$$\begin{cases} a_n = 0.7 \text{ in MOS} \\ \tau_A = \frac{C_{in}}{g_m} \end{cases}$$

$$ENC_{s\_opt}^2 = 16KT a_n C_{det} \frac{\tau_A}{t_m}$$



# ENC for Pseudo-Gaussian Shapers

Introducing also 1/f noise ...

$$ENC = \sqrt{\frac{F_v C_{in}^2 v_e^2}{T_{peak}} + F_i i_e^2 T_{peak} + F_f K_f C_{in}^2}$$

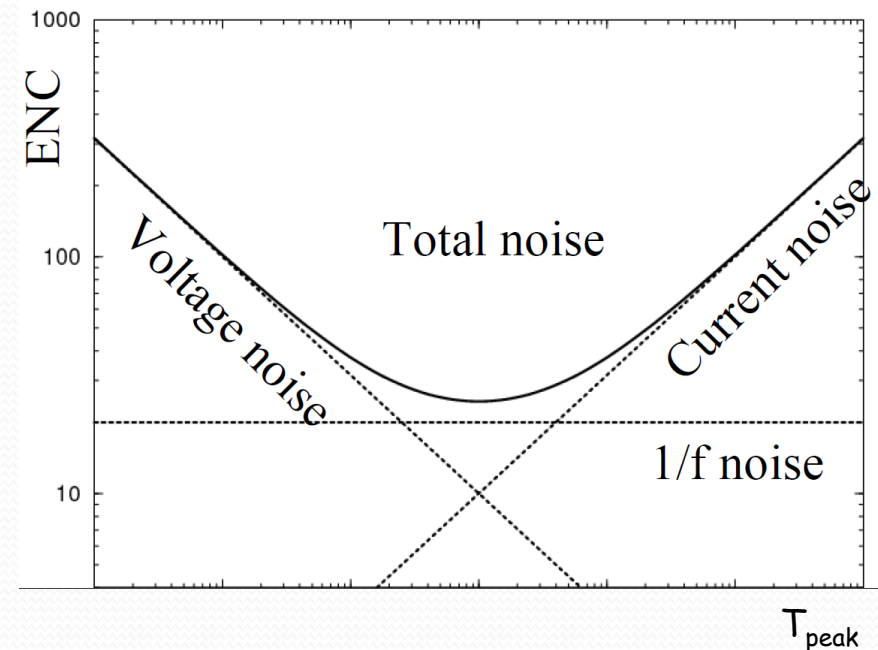
voltage noise

current noise

1/f noise

- $F_v, F_i, F_f$  - specific shape factors
- $v_e^2, i_e^2$  - voltage and current white noise densities
- $K_f$  - "1/f" noise constant
- $C_{in}$  - total input capacitance
- $T_{peak}$  - peaking time

- Voltage (serial) noise ↓ when  $T_{peak}$  ↑
- Current (paralle) noise ↓ when  $T_{peak}$  ↓
- 1/f noise does not depend on  $T_{peak}$



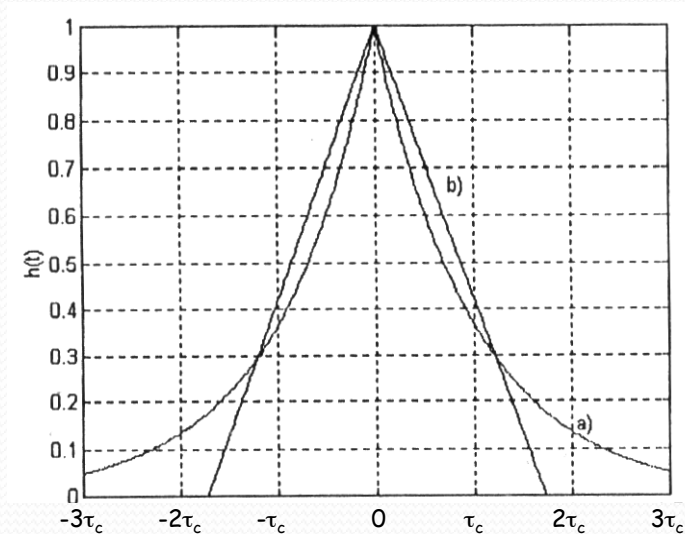
The best  $T_{peak}$  is that for  $ENC_S = ENC_P$

# Optimum filter

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

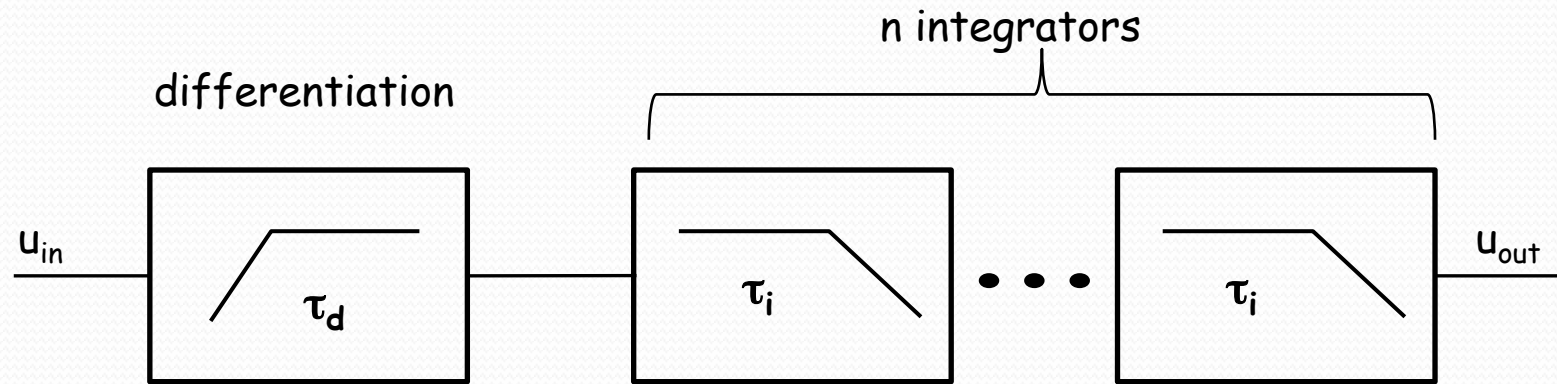
It is possible to demonstrate that  $h_{opt}(t) = \exp\left(-\frac{|t|}{\tau_c}\right) \longrightarrow ENC_{opt}^2 = 2KTR_S \frac{C_t^2}{\tau_c} = 2KTC_t \sqrt{\frac{R_s}{R_p}}$

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



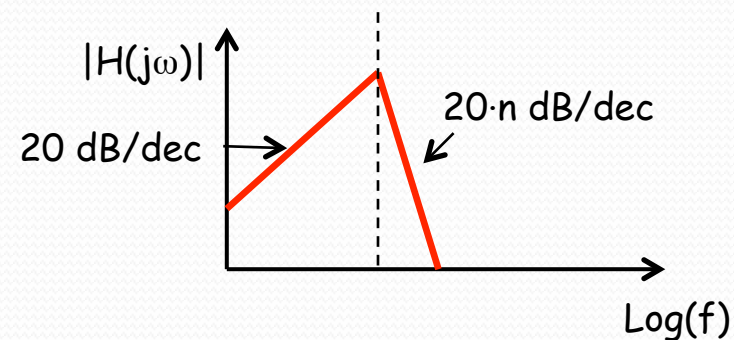
The cusp filter is not practically feasible, but can be approximated by triangular shapers (curve b) or Pseudo-Gaussian shaper

# Pseudo-Gaussian (or Semi-Gaussian) shaper



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

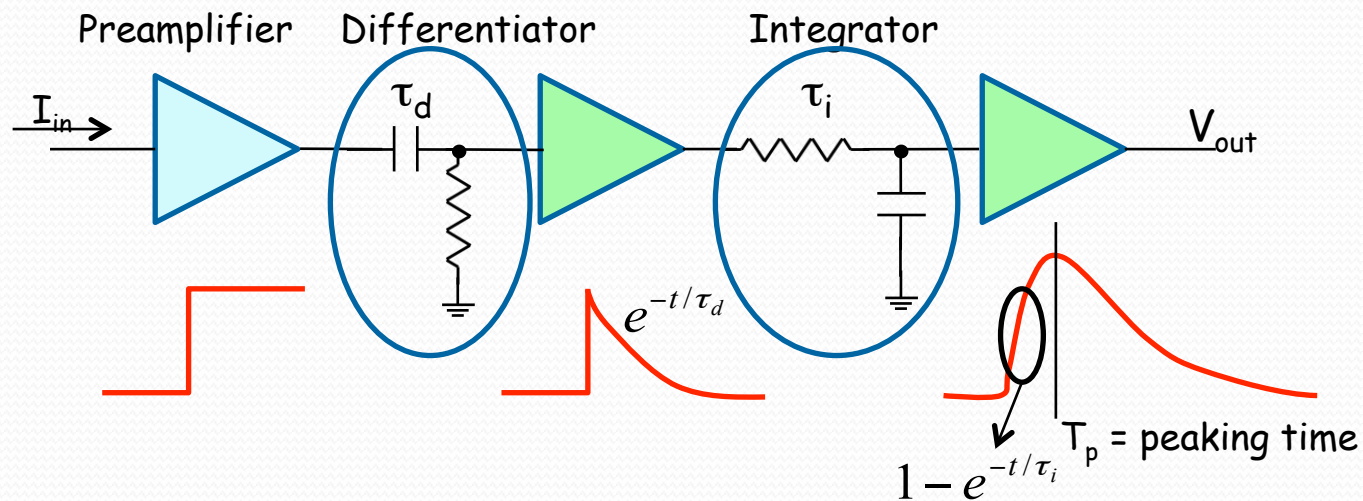
$$H(s) = \frac{u_{out}(s)}{u_{in}(s)} = \frac{s\tau_d}{(1 + s\tau_d)} \frac{1}{(1 + s\tau_i)^n}$$



# Simple shaper: CR-RC

The simplest Pseudo-Gaussian filter is the **CR-RC** shaper because :

1. The **high-pass filter** is made with CR network
2. The **low-pass filter** is made with RC network

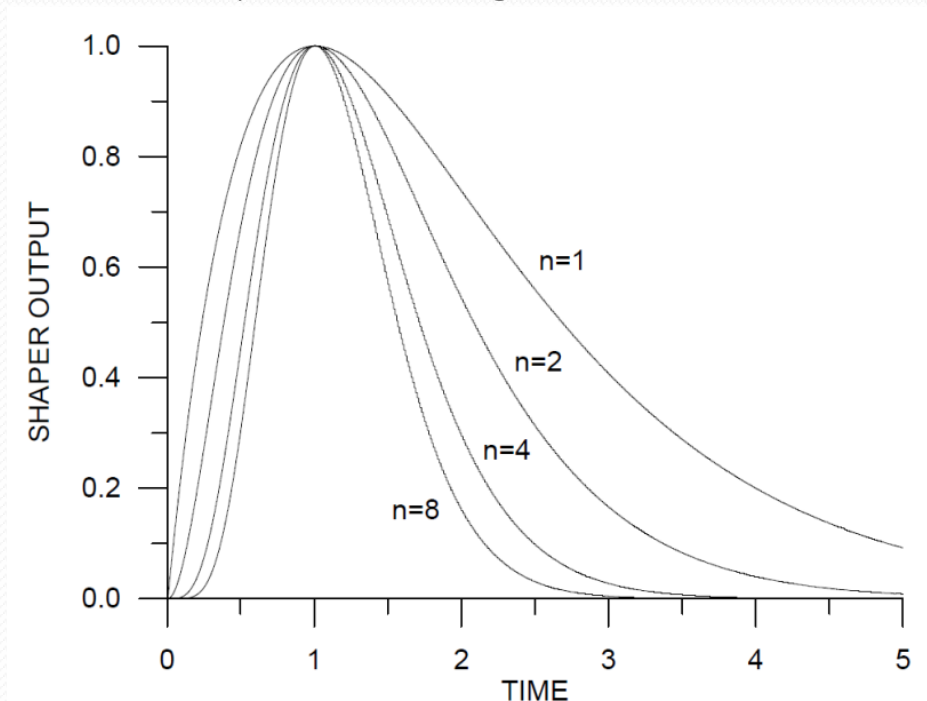


- This shaper is called **CR-RC** because the high-pass filter is made with CR network, while the low-pass filter with a RC network
- The noise is 36% worse than "optimum filter" with the same time constants

# Shaper: CR-RC<sup>n</sup>

The shapers are often more complicated, with multiple (n) integrators → CR-RC<sup>n</sup>

- Same peaking time if  $\tau_n = \tau_{(n=1)}/n$
- With same peaking time
  1. More symmetrical
  2. Faster return to baseline
  3. Improved rate capability

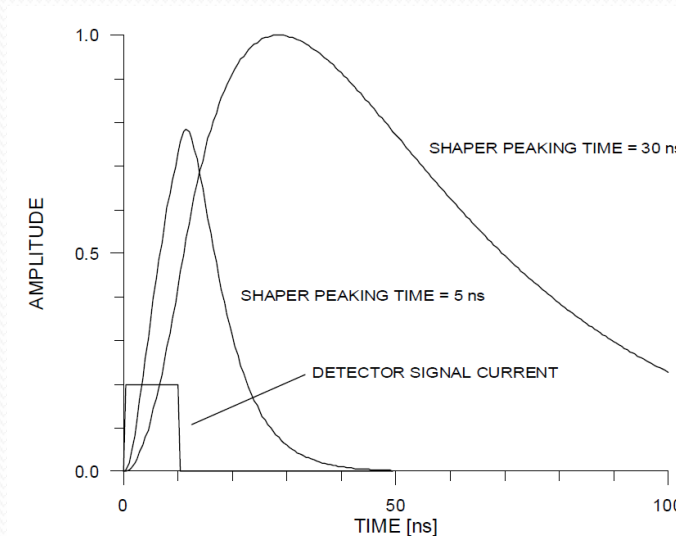


Shaper type	$F_v$	$F_i$
CR-RC	0.92	0.92
CR-RC <sup>2</sup>	0.84	0.63
CR-RC <sup>3</sup>	0.95	0.51
CR-RC <sup>4</sup>	0.99	0.45
CR-RC <sup>5</sup>	1.11	0.4
CR-RC <sup>6</sup>	1.16	0.36
CR-RC <sup>7</sup>	1.27	0.34

- Current noise decreases with shaping order

# Ballistic deficit

**Ballistic Deficit** is a Loss in Pulse Height if the peaking time  $T_p$  of the shaper is **shorter** than the detector collection time or, more in general, the rise time of its input pulse



In fact, not all the charge is collected by the amplifier because it starts to discharge before the detector signal reaches its peak

## Consequences:

- Loss of useful signal
- Increase of ENC (or decrease of S/N)

The shaping time must be carefully chosen, as a compromise among different factors:



- Short  $T_p$ : higher ENC, ballistic deficit but high sustainable event rate
- Long  $T_p$ : lower ENC but risk of pile-up

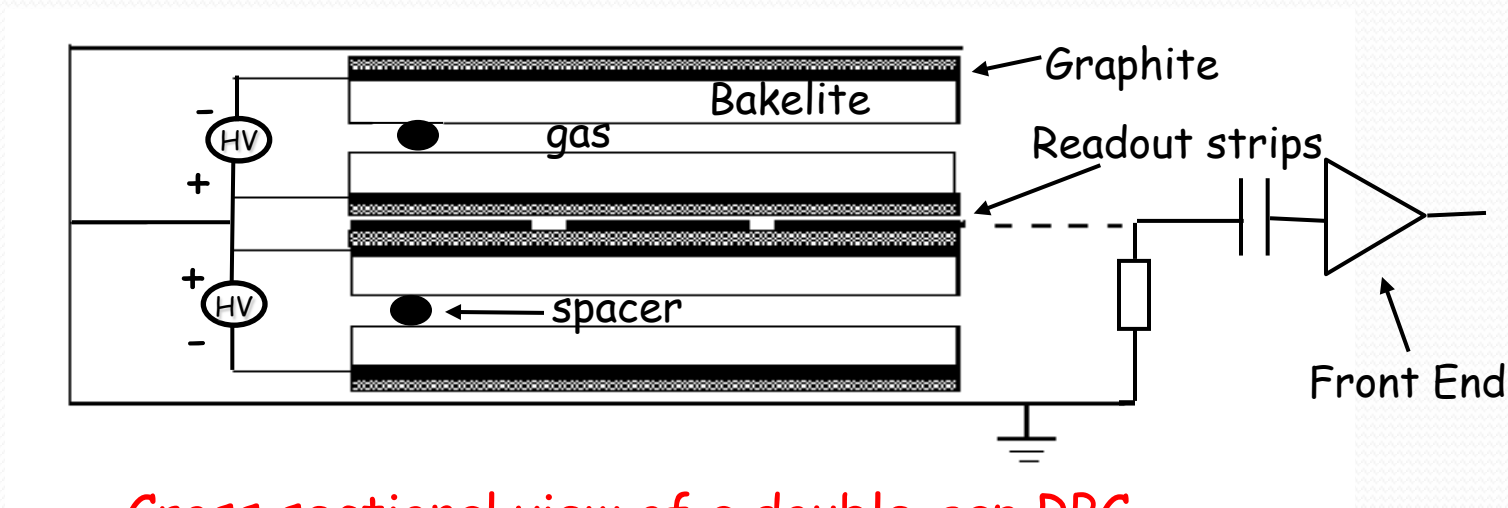
# Some examples of Front-End amplifier

1. Current sensitive amplifier: Front-end chip of Resistive Plate Chambers (RPC) for the CMS Experiment, at CERN
2. Charge sensitive amplifier: Front-end chip of Cylindrical GEM (CGEM) for the KLOE Experiment, at Frascati INFN LAB
3. VFAT3/GdSP

# CMS RPC Front-End Preamp

## Detector Characteristics

- Resistive Plate Chambers are gaseous parallel-plate detectors combining good spatial resolution with a time resolution comparable to that of scintillators ( $\sim \text{ns}$ )
- They are suitable for fast space-time particle tracking as required for the muon trigger at the LHC experiments

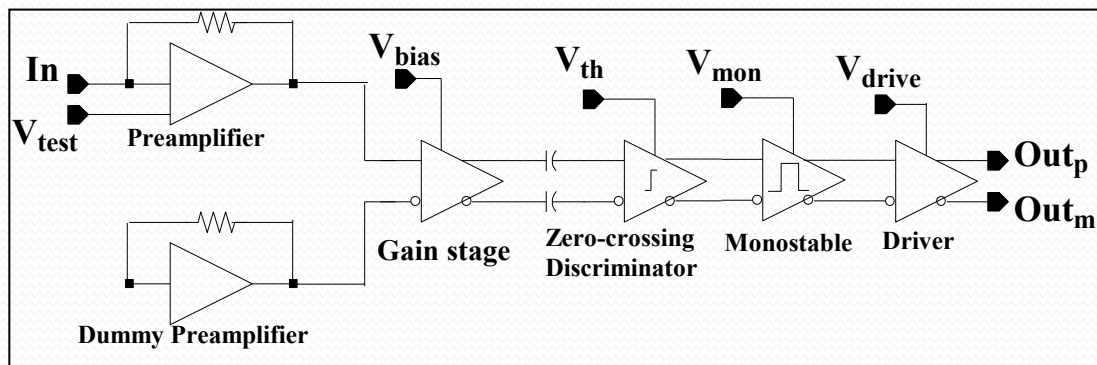


Cross sectional view of a double-gap RPC



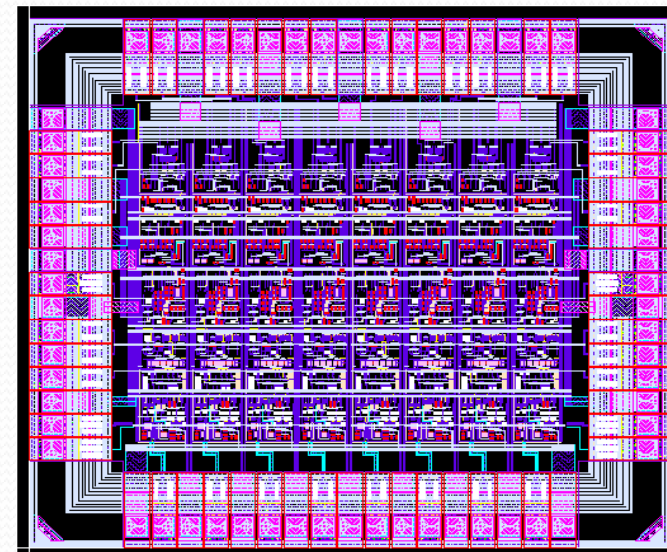
# CMS RPC Front-End

- **RPC timing information** is crucial for unambiguous assignment of the event to the related bunch crossing
- We have developed an 8-channel front-end chip in BiCMOS technology



Single channel block diagram

**Technology:** 0.8  $\mu\text{m}$  BiCMOS of AMS  
**8 channels**  
**Power supplies:** +5 V; GND  
**Power consumption:**  $\sim 45$  mW/channel



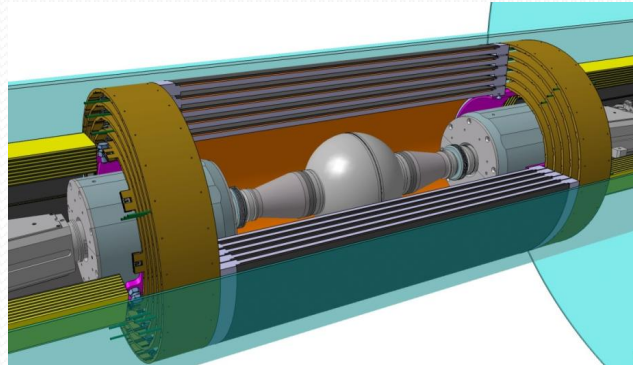
**Dimensions:** 2.9 mm X 2.6 mm  
**64 I/O pads**  
**Package:** PQFP 80

Ref.: F. Loddo et al., *New developments on front-end electronics for the CMS Resistive Plate Chambers*, Nucl. Instr. & Meth. A 456 (2000) 143-149

# KLOE Inner Tracker Front-End

**KLOE** is an experiment at DAFNE accelerator, in Frascati INFN National Laboratories (Italy)

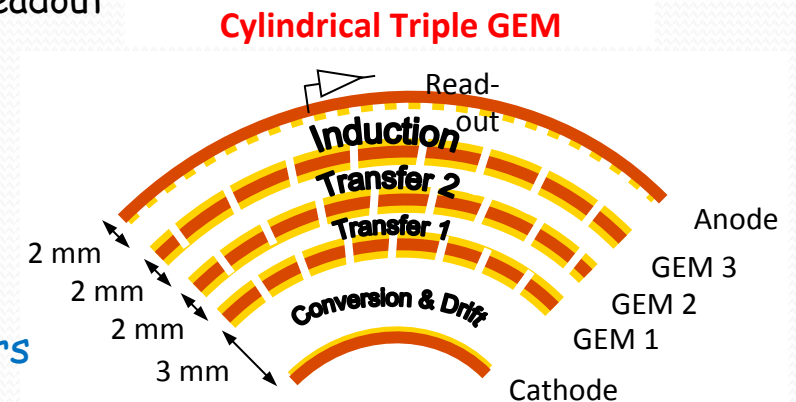
An Inner Tracker GEM-based will be inserted around the interaction point to improve the vertex resolution by a factor 3 (with respect to present detector, without IT)



Gas Electron Multiplier (GEM) technology has been adopted, for its low material budget

- 4 independent tracking layers for a fine vertex reconstruction of  $K_S$  and  $\eta$
- $200 \mu\text{m}$   $\sigma_{r\phi}$  and  $500 \mu\text{m}$   $\sigma_Z$  spatial resolutions with XV readout
- 700 mm active length
- from 130 to 220 mm radii
- 1.8%  $X_0$  total radiation length in the active region
- 5 kHz/cm<sup>2</sup> rate capability

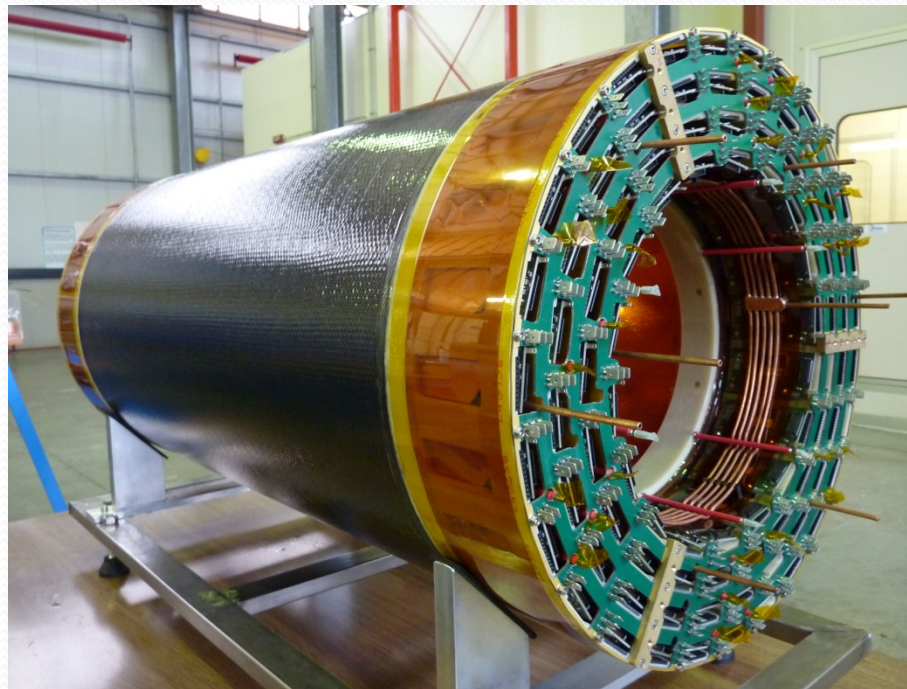
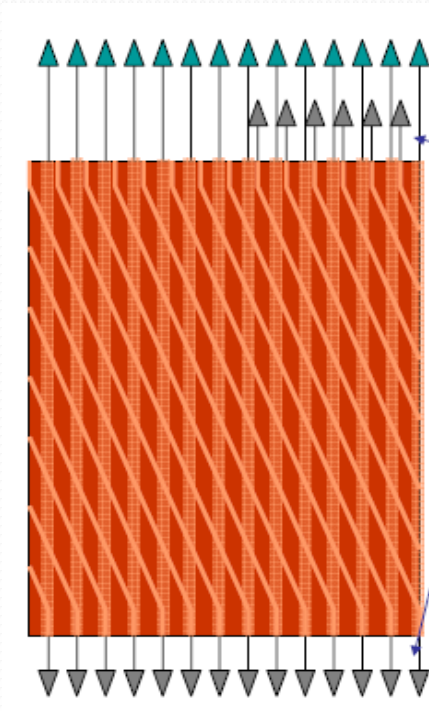
Realized with Cylindrical TRIPLE GEM detectors



# KLOE IT Readout

The Anode is shared out in a 2 dimensions layout having a XV geometry, X and V strips read out is shared out at both ends.

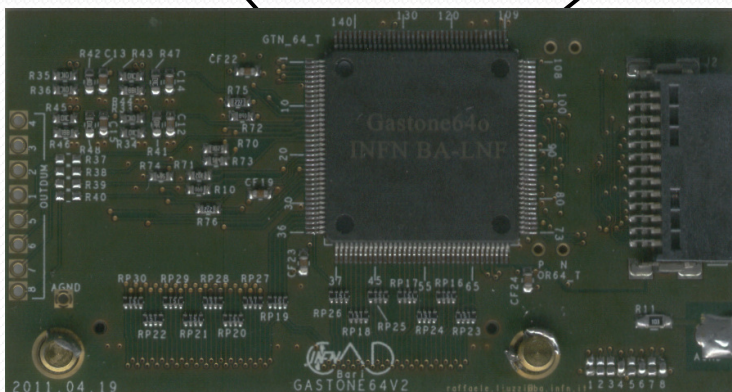
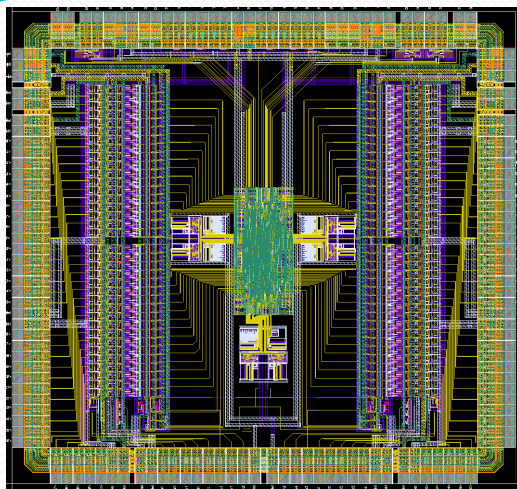
Rotated by roughly  $40^\circ$ , such strip's net provides 2D positioning for the particle passing through the layer.



KLOE-2 Inner Tracker



# KLOE IT Front-End: GASTONE64

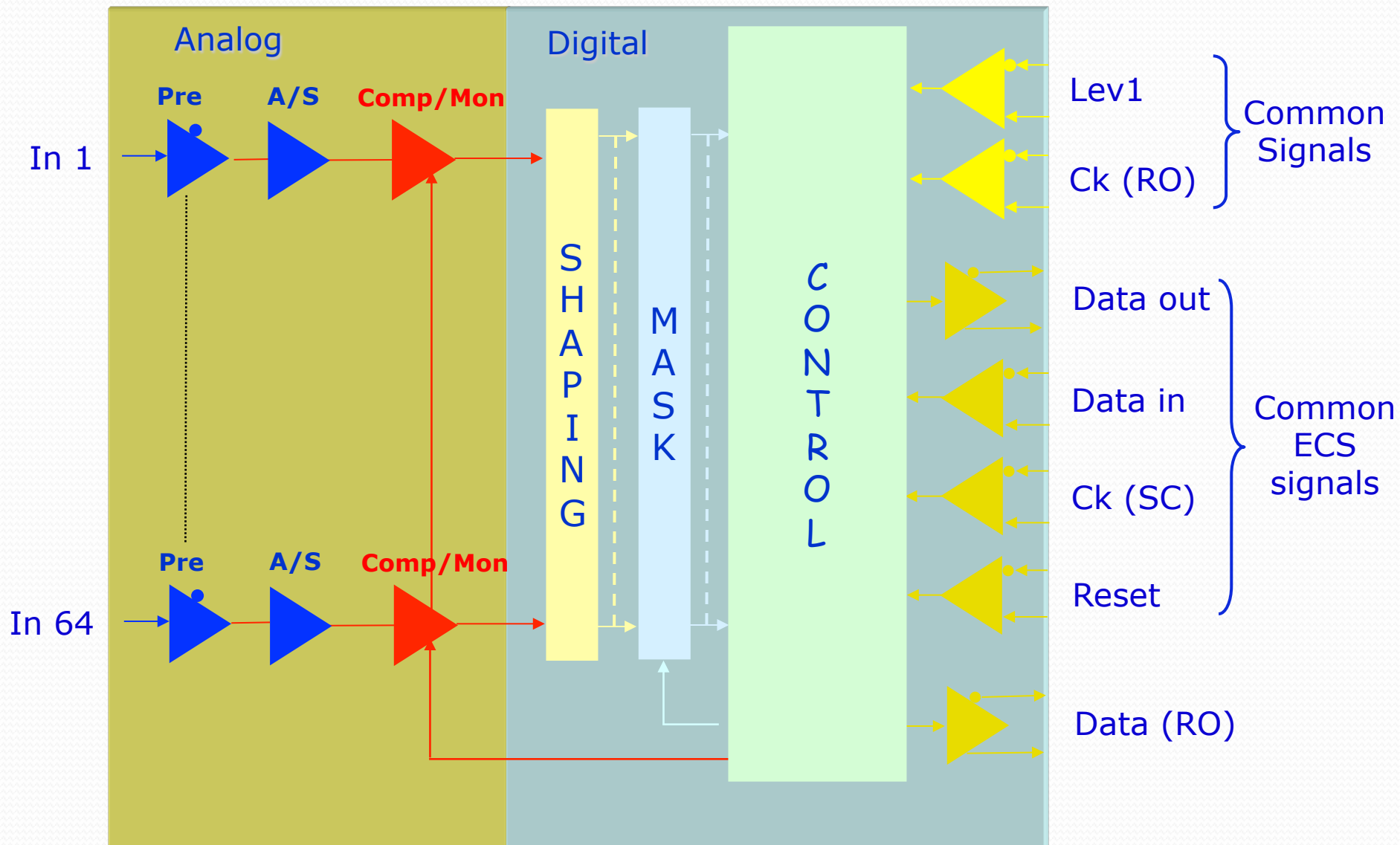


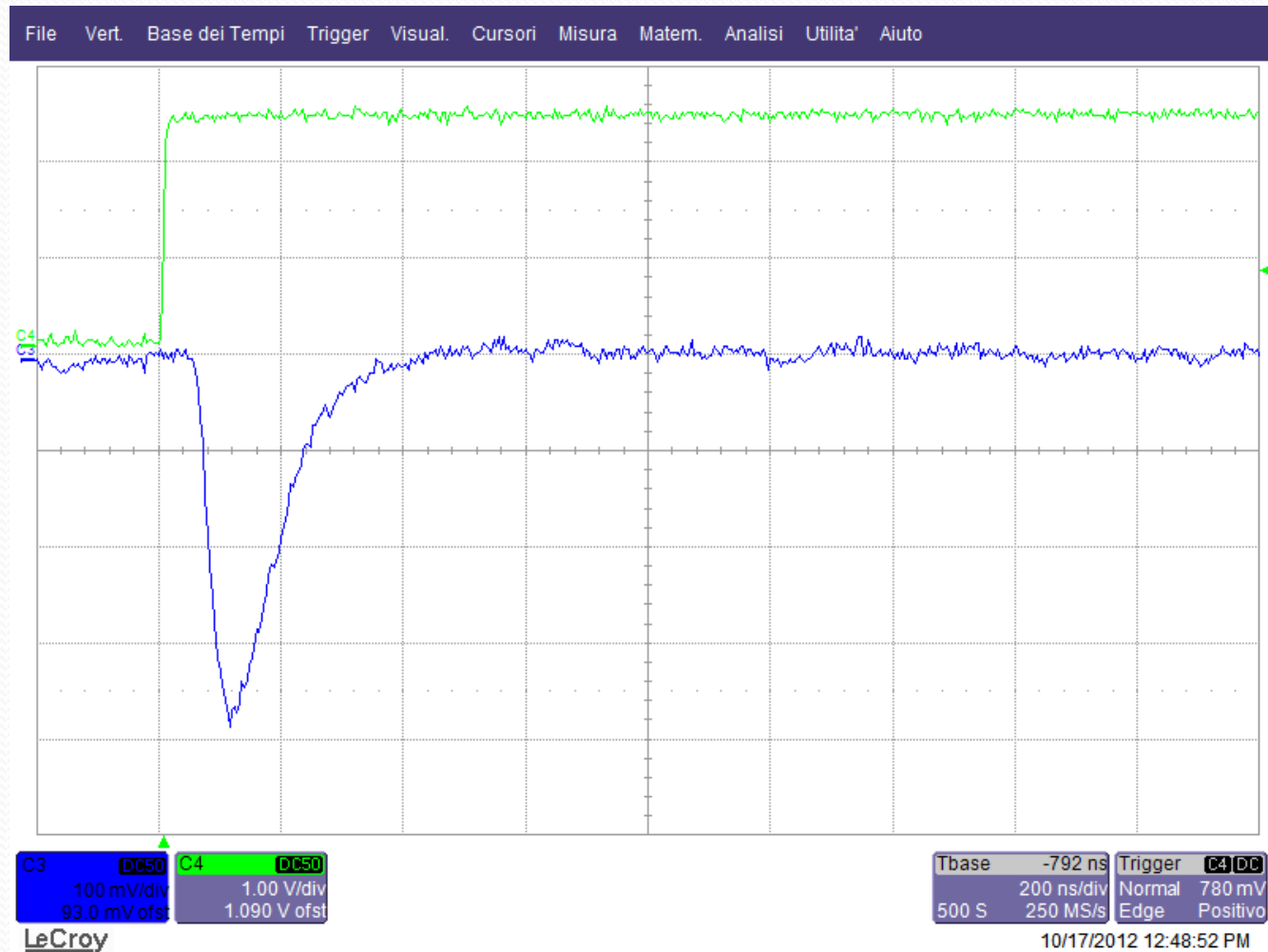
N. channels	64
Technology	CMOS 0.35 $\mu\text{m}$
Chip dimensions	4.5 X 4.5 mm <sup>2</sup>
Input impedance	120 $\Omega$
Charge sensitivity	16 mV/fC (Cdet = 100 pF)
Peaking time	~90 ns (Cdet=100 pF)
Crosstalk	< 3%
ENC	800 e <sup>-</sup> + 40 e <sup>-</sup> /pF
Power consumption	~ 6 mW/ch
Readout	Serial LVDS (100 MBps)

A. Balla et al., *A new cylindrical GEM inner tracker for the upgrade of the KLOE experiment*, Nucl. Phys. Proc. Suppl. 215:76-78,2011

A. Balla et al., *GASTONE: A new ASIC for the cylindrical GEM inner tracker of KLOE experiment at DAFNE*, Nucl. Instr. & Meth. A 604 (2009) 23-25

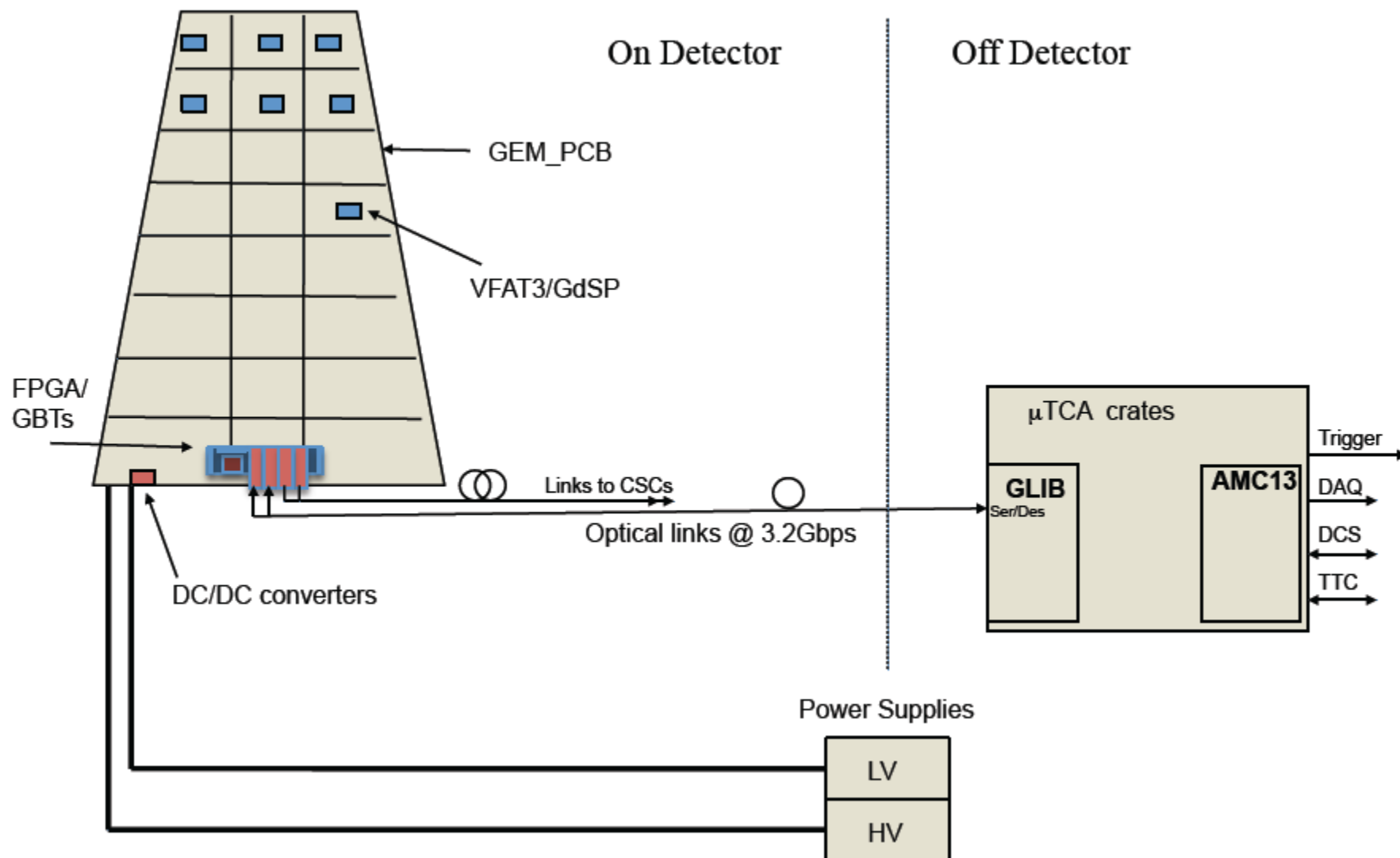
# KLOE IT – GASTONE64





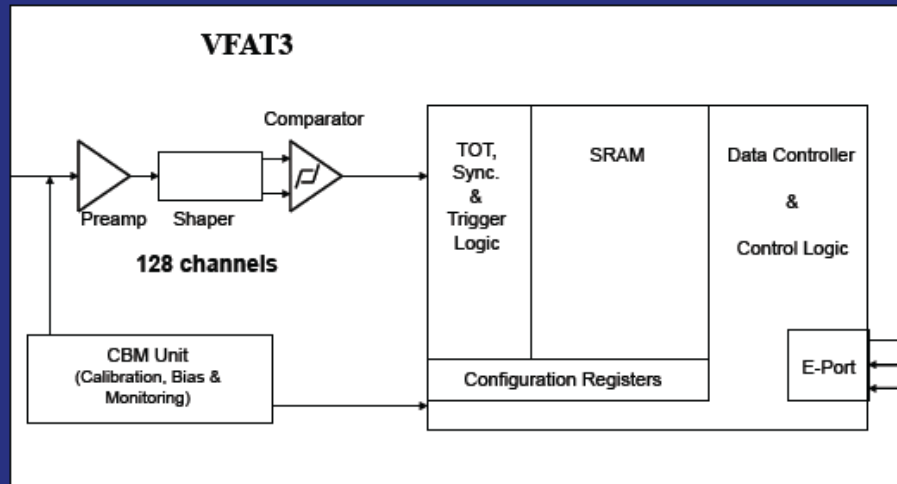
Shaper response for 20 fC input pulse

# VFAT3: a proposed FE chip for CMS GEM

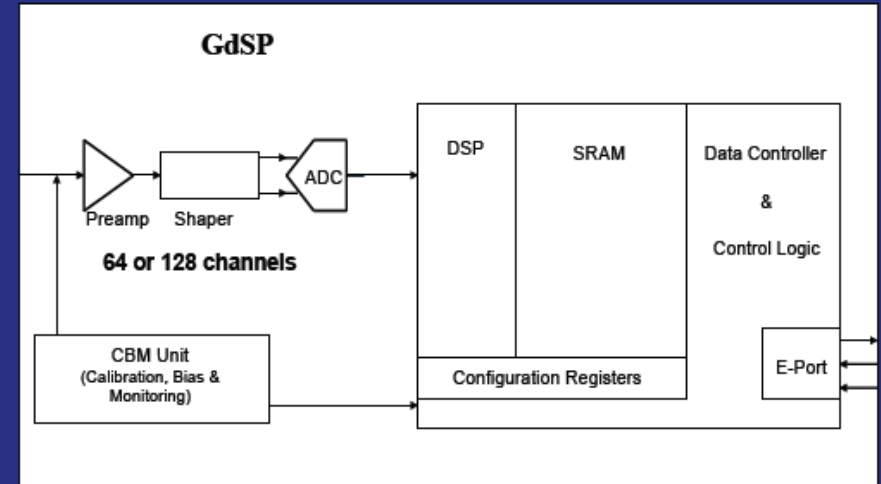


# VFAT3

2 Trigger & Tracking Front-end architectures considered.



OR



## VFAT3 :

Front-end with programmable shaping time.

Internal calibration.

Binary memory

Interface directly to GBT @ 320Mbps.

Designed for high rate  
(10kHz/cm<sup>2</sup> depending on segmentation)

Tech.: CMOS 130 nm

## GdSP :

Similar to VFAT3 except has an ADC / channel instead of a comparator.

Internal DSP allows subtraction of background artifacts enabling a clean signal discrimination.



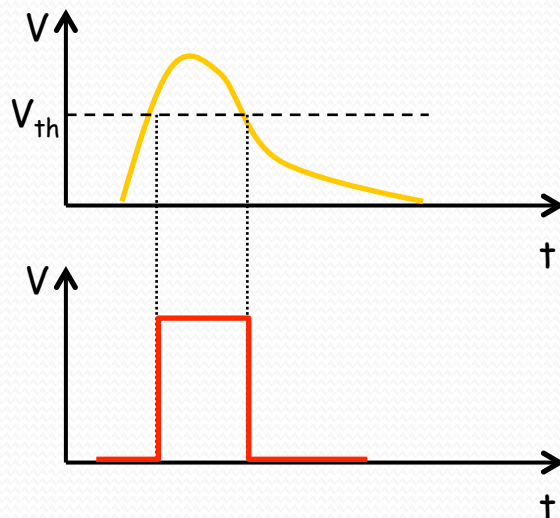
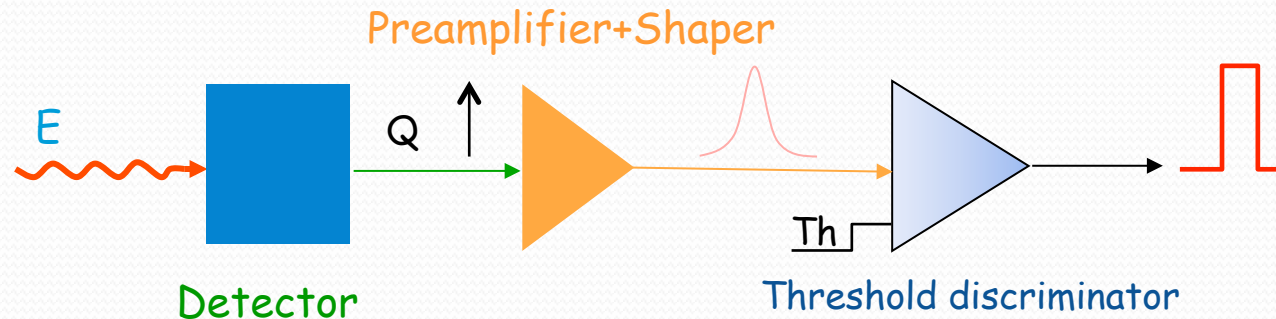


# Time measurement

# Time measurement

Often the purpose of the system **Detector + Front End Electronics** is time measurements

The simplest scheme is the following:



Leading edge or Threshold discriminator (comparator):  
when the signal crosses a threshold, the output goes from "low" to "high" level: we get a "time tag"

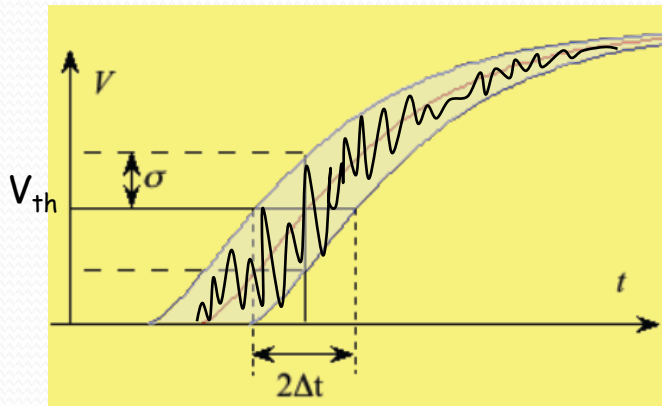
Accuracy of timing measurement is limited by:

1. Jitter
2. Time walk

# Time measurement and noise: the jitter

Noise has an impact also in time measurements:

uncertainty in the time of crossing threshold  $\rightarrow$  Jitter



$$\Delta t = \frac{\sigma_{noise}}{\underbrace{dV/dt}_{\text{slope}}}$$

How to decrease jitter?  $\rightarrow$  Conflicting conditions:

- decrease  $\sigma_{noise} \rightarrow$  decrease bandwidth
- increase slope  $\rightarrow$  increase bandwidth

As usual ... find compromise

Example:  $V = V_{max} (1 - e^{-t/\tau})$

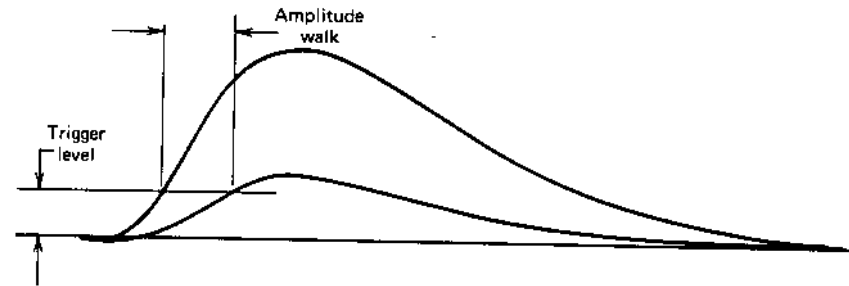
(If  $t \ll \tau$ )  $\rightarrow$

$$\Delta t \approx \frac{\sigma_{noise} \tau}{V_{max}}$$

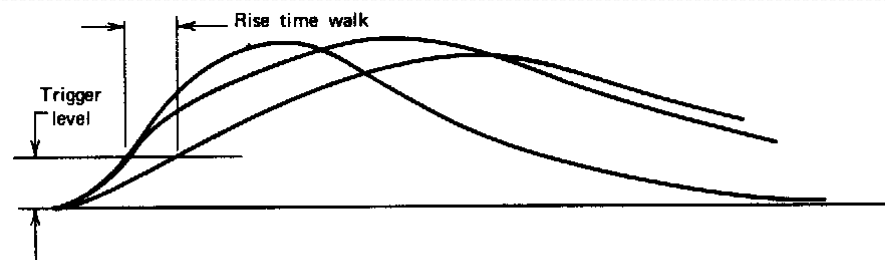
$$\text{Rise time (10\%-90\%)} = 2.2\tau$$

# 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** ( $\Delta T = \text{time walk}$ )



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)



The sensitivity of leading edge triggering to time walk is minimized by setting the threshold as low as possible but it must be compatible with noise level and the discrimination point should be in a region where the slope is steep to minimize jitter

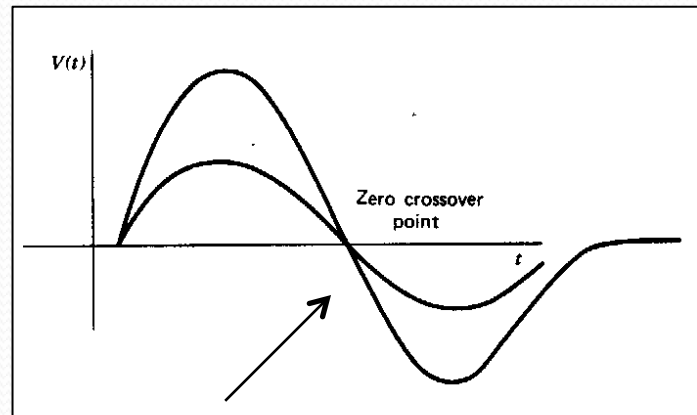
- Time walk correction:**
- Software: measure the pulse amplitude and apply correction to timing
  - Hardware: instead of leading edge triggering, use
    1. **Crossover timing**
    2. **Constant Fraction timing**

# Crossover timing

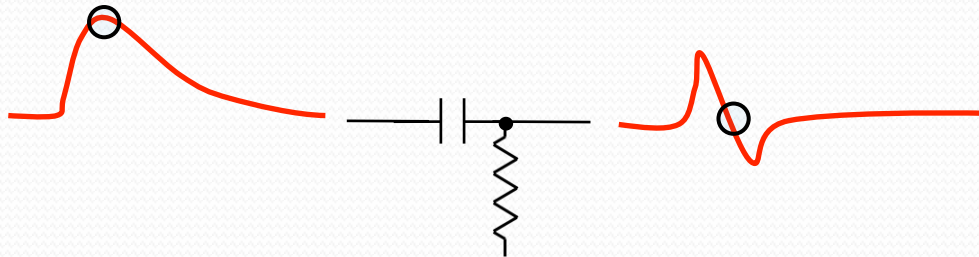
The crossover timing can greatly reduce the magnitude of the amplitude time walk

## Hypothesis:

- the output of the shaper is a bipolar pulse and the time of crossing from the positive to the negative side of the axis (**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

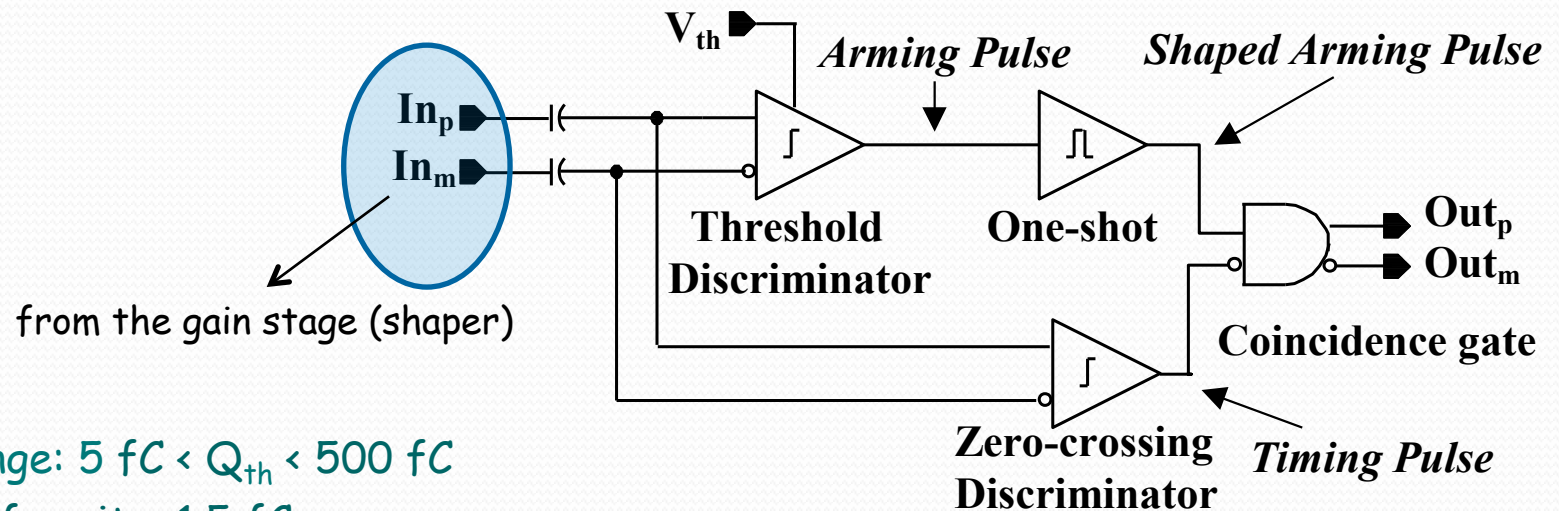
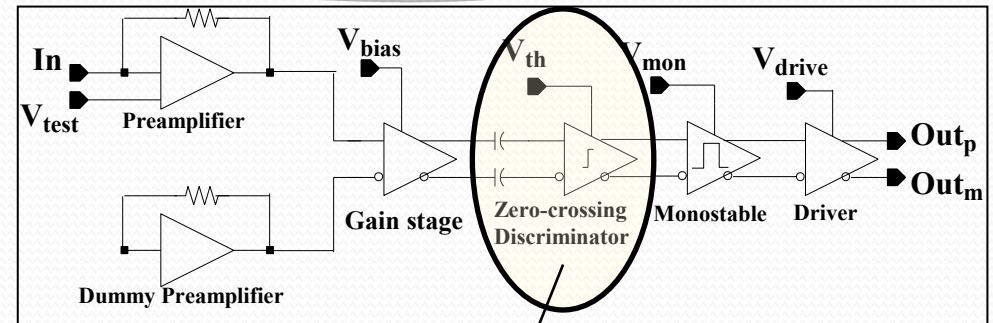


This method reduce amplitude walk, but usually jitter is larger than leading edge triggering

# Example of crossover timing: CMS RPC Front-End

## Zero-crossing discriminator:

- Threshold Discriminator (Charge selection)
- Zero-Crossing Discriminator (Time reference)
- One-shot
- Coincidence gate



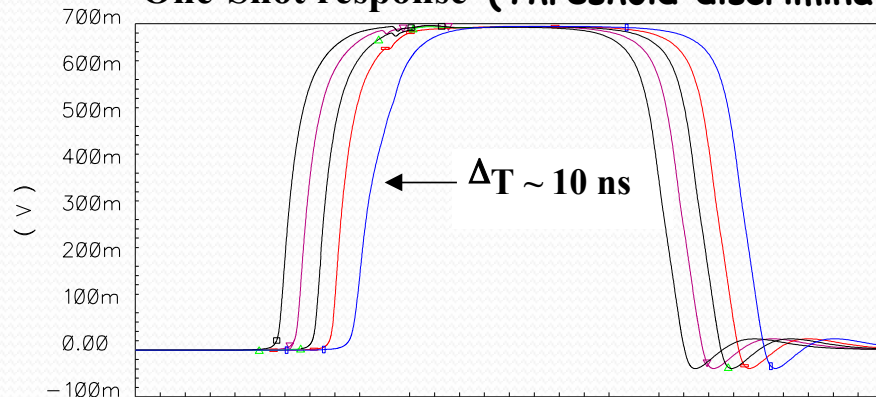
- Threshold range:  $5 \text{ fC} < Q_{th} < 500 \text{ fC}$
- Threshold uniformity:  $1.5 \text{ fC rms}$
- Shaped Arming pulse width:  $20 \text{ ns}$
- Power:  $8 \text{ mW}$

# Example of crossover timing: CMS RPC Front-End

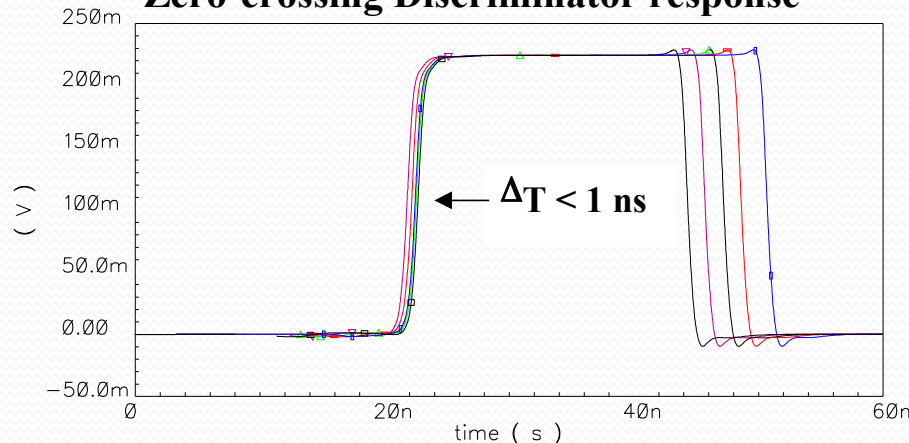
Zero-crossing discriminator response in the dynamic range

$$Q_{th} = 20 \text{ fC} \quad 1 \text{ fC} < Q_{ov} < 20 \text{ pC}$$

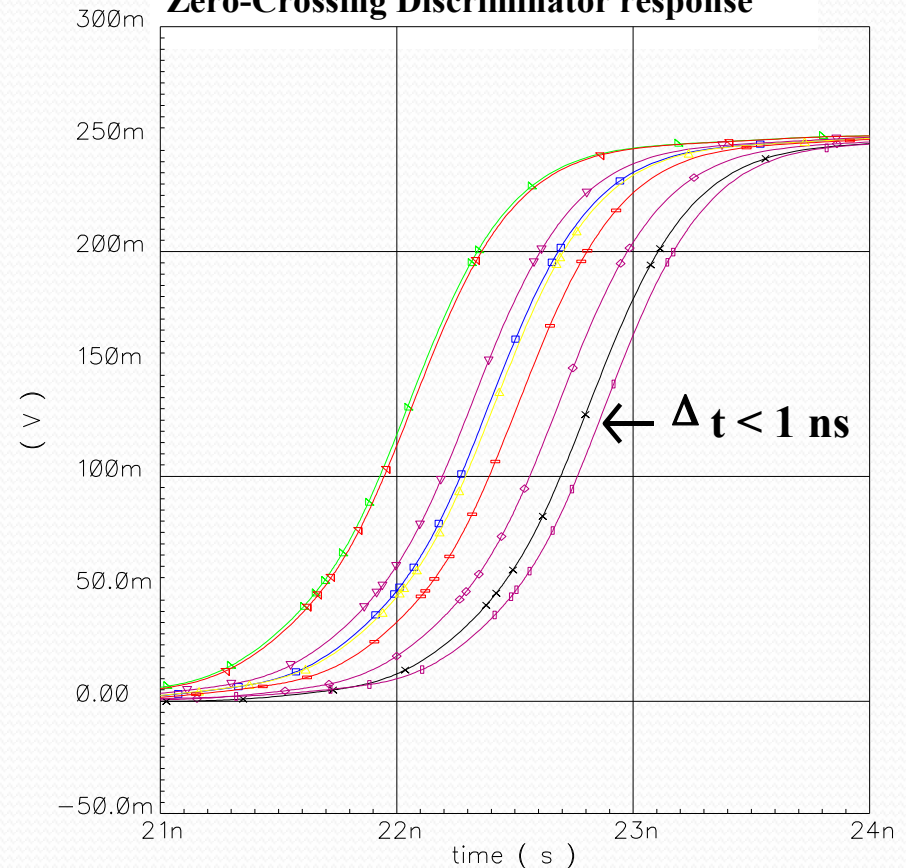
One-Shot response (Threshold discriminator)



Zero-crossing Discriminator response



Zero-Crossing Discriminator response



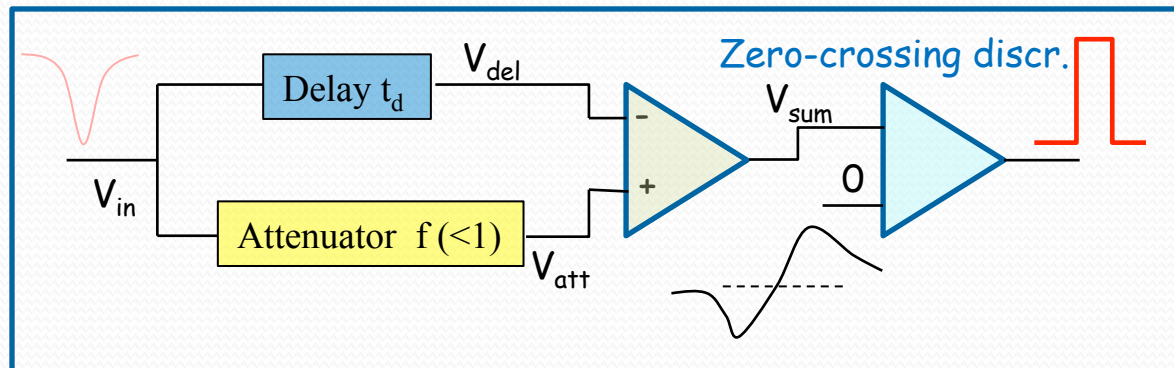
# The Constant Fraction Timing

- It is empirically found that the best leading edge timing characteristics are obtained when the threshold is set at about 10-20% of the pulse amplitude
- These observations have led to the development of the **Constant Fraction Timing**, that produce an output timing pulse a fixed time after the leading edge of the pulse has reached a constant fraction of the peak amplitude
- This point is independent of pulse amplitude for all pulses of constant shape, but with lower jitter than zero-crossing

## Making the sum of

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

we get a bipolar signal, whose zero-crossing time is independent of pulse amplitude and corresponds to the time at which the pulse reaches the fraction  $f$  of its final amplitude

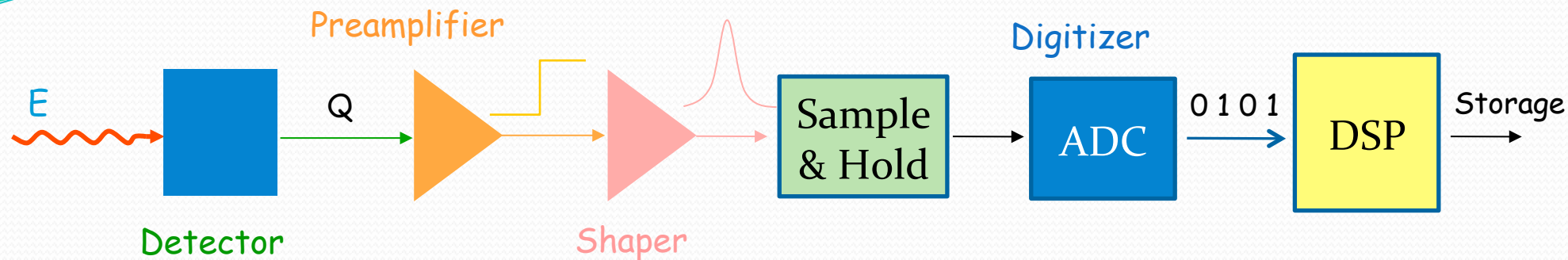






# Digital signal processing

# Digital signal processing

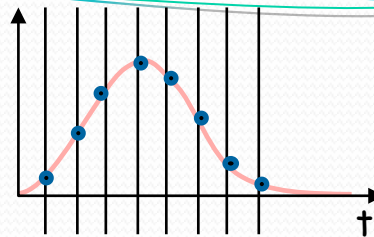
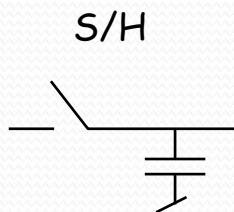
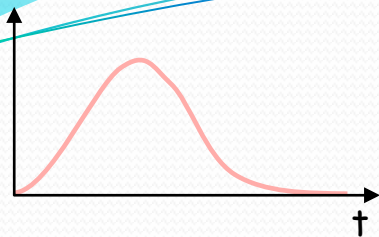


Another approach is to perform operations on the signal in the digital domain

- Baseline restoration
- Tail cancellation
- Filtering
- Zero suppression
- .....

1. Sample the amplifier output (at a proper sampling rate)
2. Convert the analog value of the samples into digital word
3. Make operations on digital words using processors or Digital Signal Processors

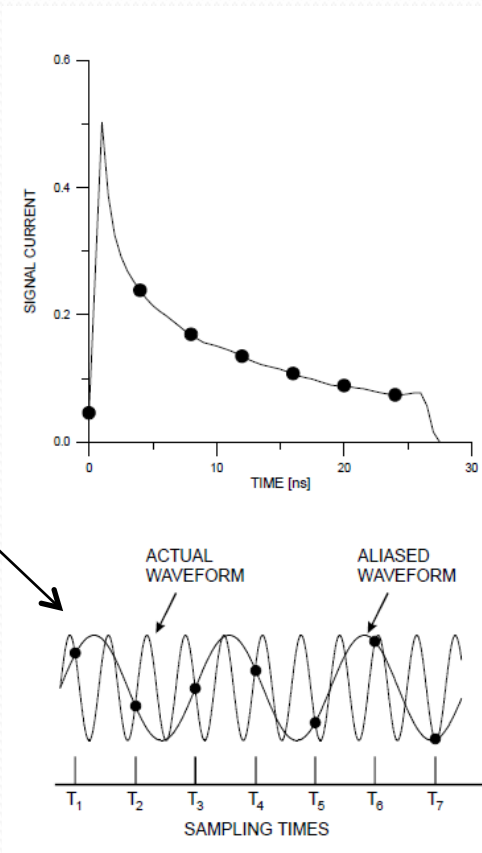
# Digital signal processing



If the sampling rate is too low, we lose information: high frequency components are *aliased* to lower frequencies and the reconstructed signal differs from the original one

## Nyquist condition

The sampling rate must be  $\geq 2 \times$  highest signal frequency



# Digital signal processing

## Advantages:

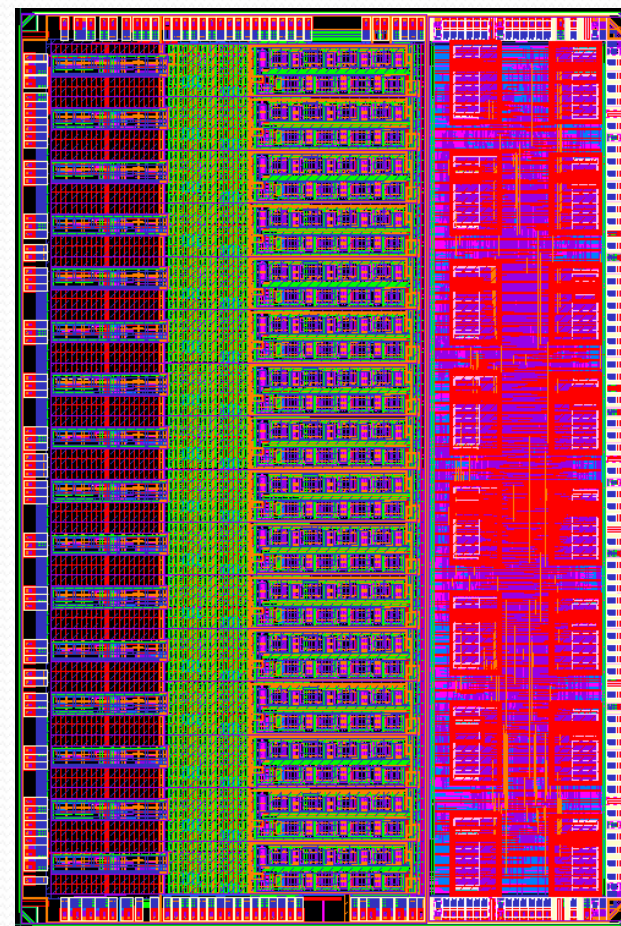
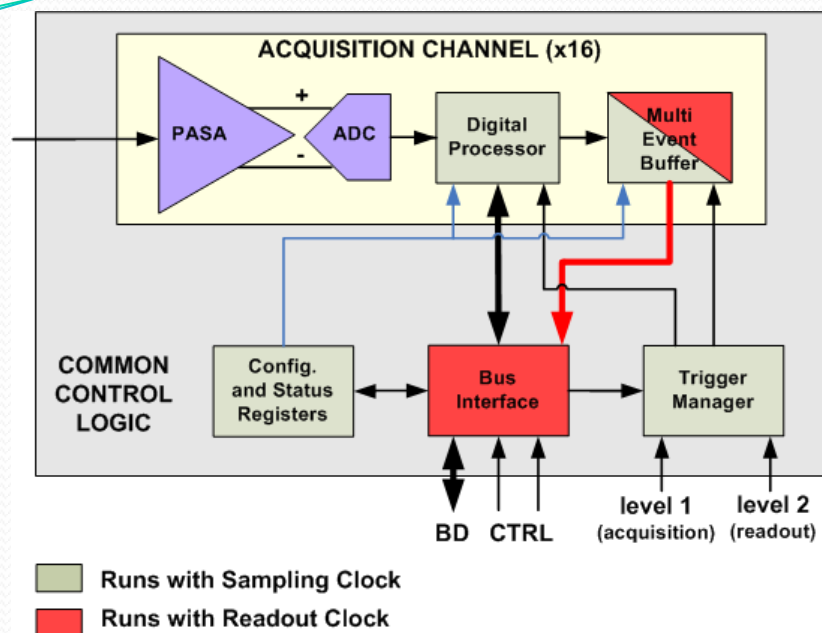
- great flexibility in implementing filter functions
- possibility to implement complex functions (impractical in hardware)
- easy to change filter parameters
- adaptive filtering can be used to compensate for pulse shape variations

## Drawbacks:

- increased complexity
- require fast and precise ADC
- large silicon area required
- power consumption

Currently, there are many progress in the development of fast, low-power ADC, so the interest in DSP applications is growing and becoming more and more attractive

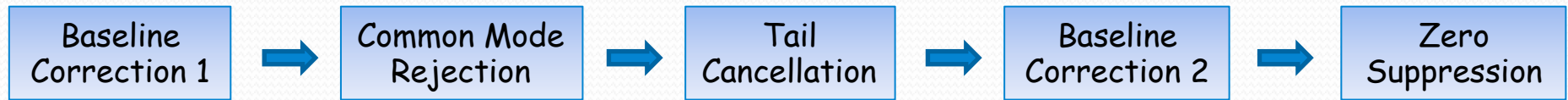
# Example of Front-End + DSP: SAltro16



- Front-end for TPC, GEM, Micromega ...
- Technology: CMOS 130 nm
- 16 channel demonstrator developed in 2009-2010
- Specification & architecture: Luciano Musa
- Design coordination: Paul Aspell
- Designers: Massimiliano De Gaspari, Hugo França-Santos, Eduardo Garcia

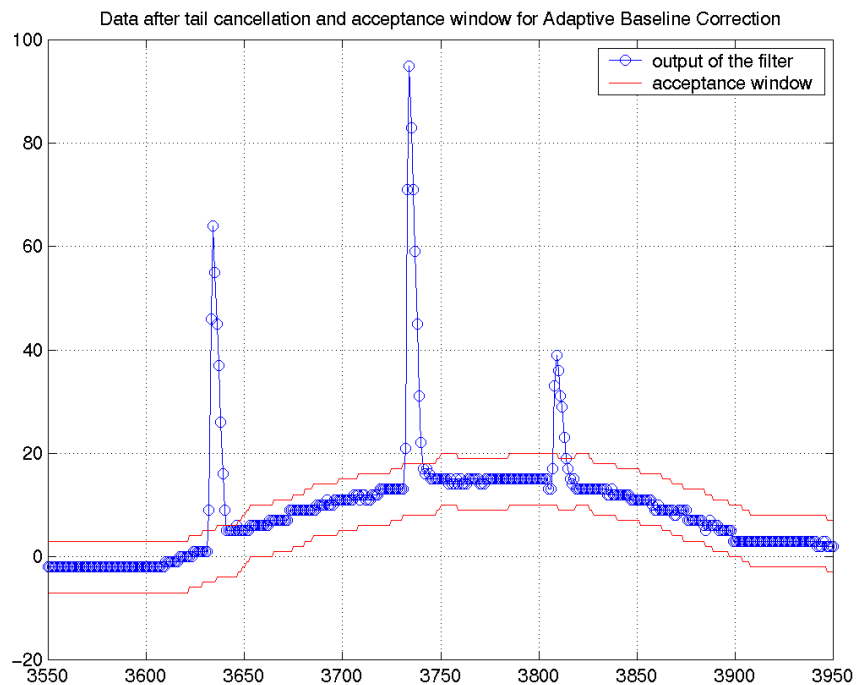
# Example of Front-End + DSP: SAltro16

## The Digital Signal Processing operations

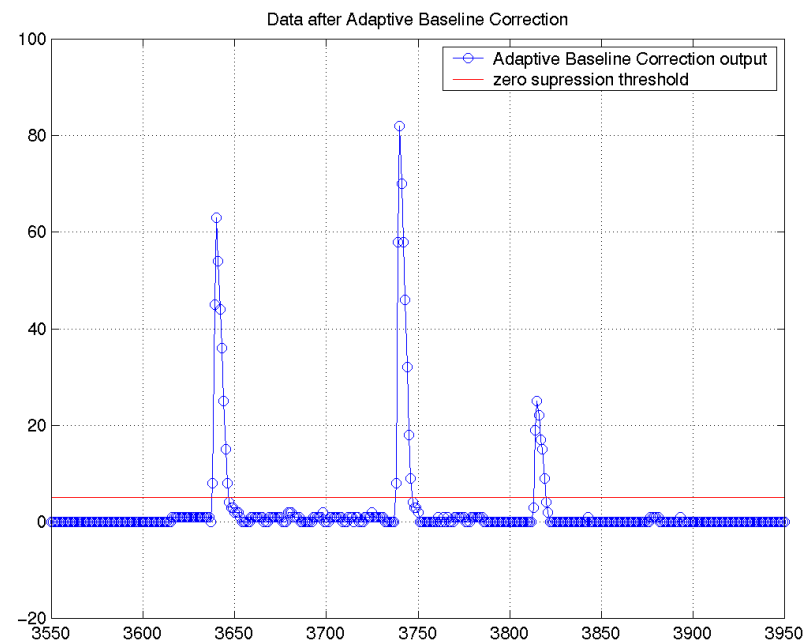


- **Baseline correction 1:** removes systematic offsets that may have been introduced by noise pickup, clock interferences etc
- **Common Mode Rejection:** calculates the common mode pickup across all the channels and subtracts it
- **Tail cancellation:** removes the distortion of the signal shape caused by undershoot
- **Baseline correction 2:** reduce low frequency baseline shifts
- **Zero suppression:** removes from the data flow the samples smaller than a programmable threshold

# Example of Front-End + DSP: SAltro16



BC 2



After baseline correction, a fixed threshold can be applied safely



# Summary

- The choice and design of Front-End electronics is crucial to obtain the desired energy and/or time resolution
- The technology strongly depends on the radiation environment
- ENC increases with detector capacitance and can be minimized matching  $C_{\text{det}}$  with preamplifier input capacitance
- The choice of pulse shape (and peaking time) comes out as a compromise between S/N optimization and double pulse resolution
- The shapers are built commonly with CR-RC<sup>n</sup> filters
- Depending on the event rate, baseline restoration may be needed
- Also digital shapers (using DSP) can be used, to enhance the flexibility and improve some optimizations (common mode rejections, baseline restoration ...) but require fast and precise ADC
- 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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*Thank you!*