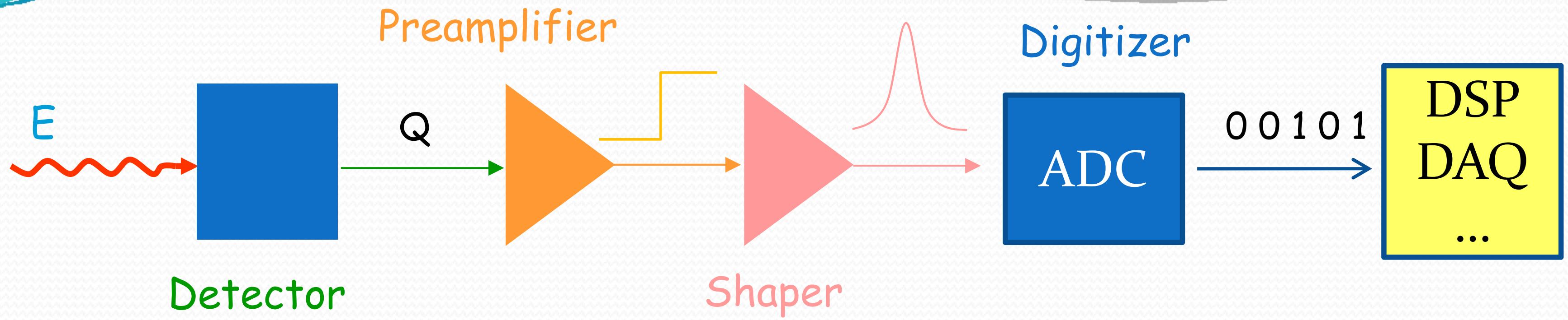
Basic electronic concepts for particle physics and beyond

Flavio Loddo
Istituto Nazionale di Fisica Nucleare – Sez.di Bari (Italy)
flavio.loddo@ba.infn.it

Outline

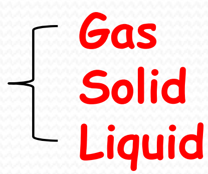
- Introduction
- Detectors and signals
- Noise basic principles
- Front-end schemes
- Charge-sensitive amplifier
- Shapers
- Time measurement
- Examples

Introduction



General readout architecture

- **The particle deposits energy in a detecting medium**



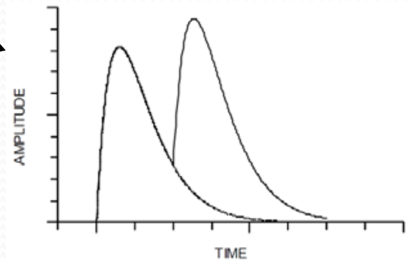
- **Energy is converted into an electrical signal: Q = KE**
- **The charge Q is typically small and must be amplified, in order to be measured and processed**
- **The preamplifier converts Q into a voltage**
- **The shaper provides gain and shape, according to the application and trying to optimize S/N**
- **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**



Example of pile-up

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 …)**

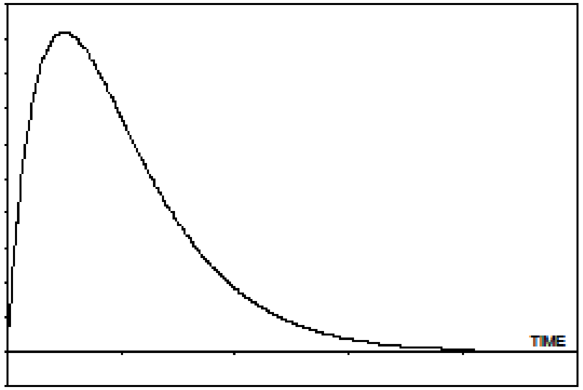
NOISE BASIC PRINCIPLES

Noise

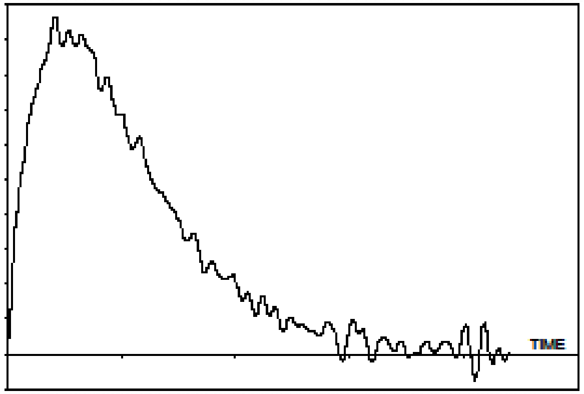
The precision of amplitude and timing measurements is limited by the NOISE

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**

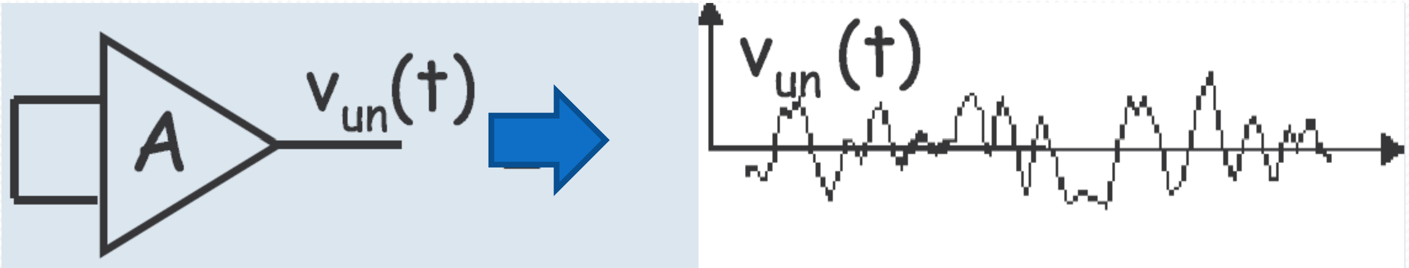
   It is a property of detector and/or electronics

   Can be reduced by proper design of front-end electronics

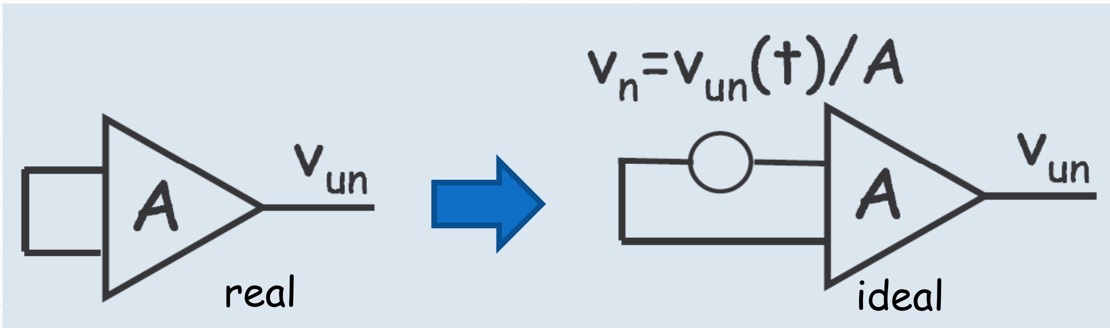
Intrinsic noise

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

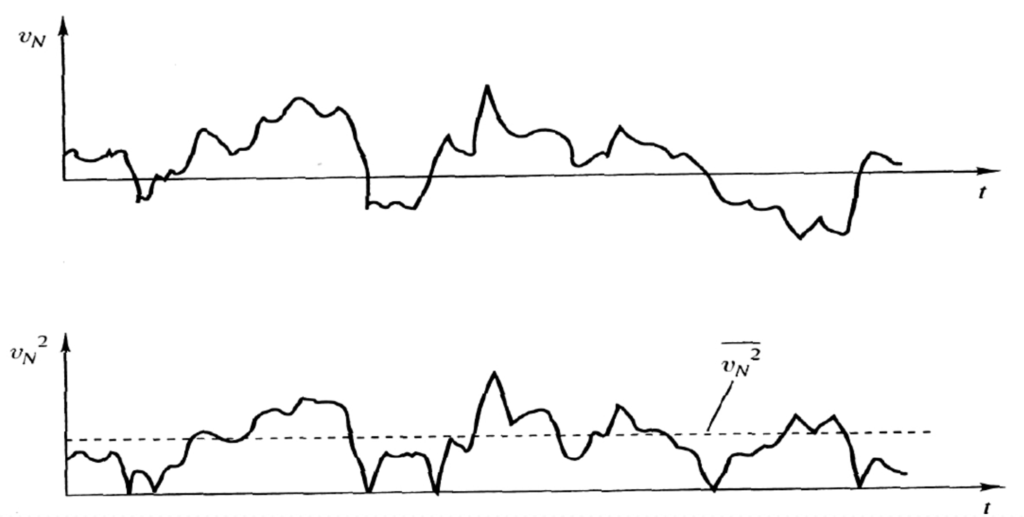
The fluctuations of $V_{un}(t)$ when $V_{in} = 0$ correspond to the noise of amplifier



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 ≠ 0

⬇

$V^2{}_n$

We can define:

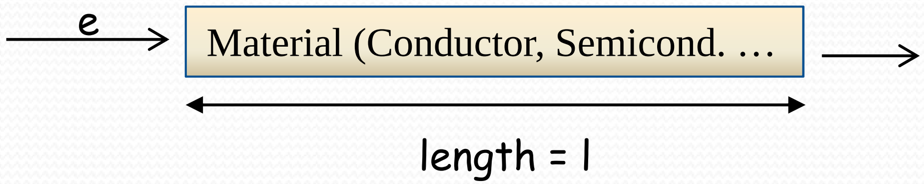
- Source of **voltage noise:** $$v_n = \sqrt{v_n^2(f)}$$
- Source of **current noise:** $$i_n = \sqrt{i_n^2(f)}$$

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

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

If Power Spectral Density is constant → White Noise

Basic noise mechanisms



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

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:

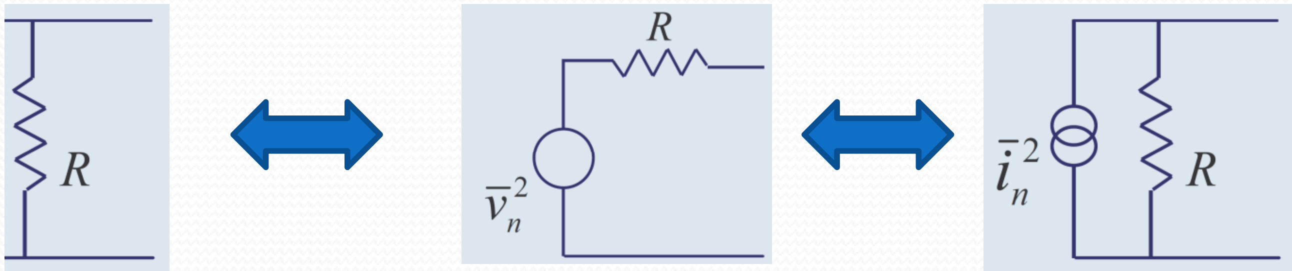
Velocity fluctuations → Thermal noise

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

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



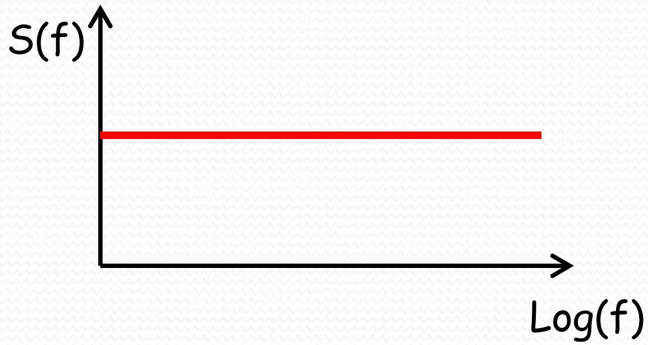
A real (noisy) resistor is equivalent to an ideal (noiseless) resistor + noise source (voltage or current)

Power spectral density:

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

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

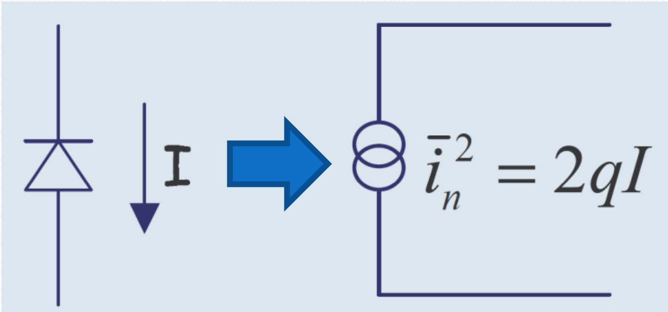
k = Boltzmann constant = $1.3806503 \times 10^{-23}$ J/K
T = absolute temperature
R = resistance



Does not depend on f → Thermal noise is a white noise

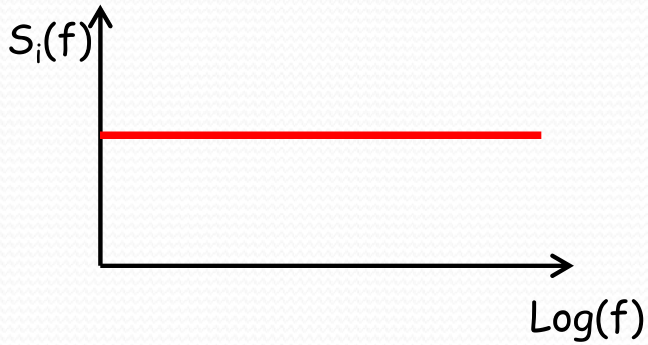
2. Shot noise

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



Power spectral density: $S_i(f) = \dfrac{d\bar{i}_n^2}{df} = 2qI$

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



Example: consider a reversed-biased diode, with leakage I = 1 nA

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

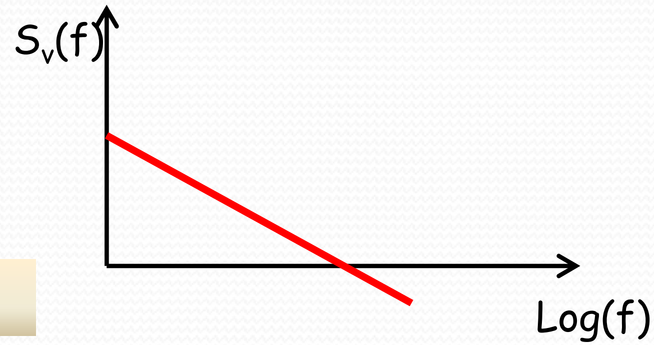
3. Flicker noise (1/f noise)

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:

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

I is dc current
K_f is a constant (vary from device to device)
a ~ 0.5 ÷ 2
b ~ 1



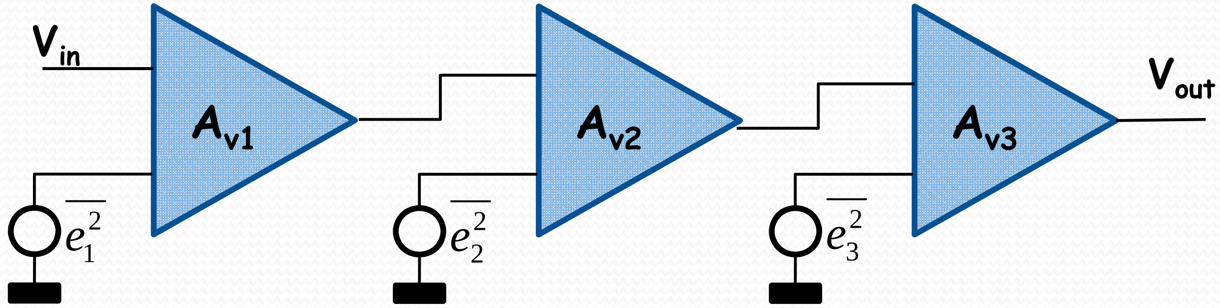
It depends on f and clearly it is important at low frequencies

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).

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

Intrinsic noise: importance of first stage



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

$$
\left(\frac{Noise}{Signal}\right)^2 = \left(\frac{\overline{e^2}_{out}}{V^2_{out}}\right) = \frac{\overline{e_1^2} + \frac{\overline{e_2^2}}{A_{V1}^2} + \frac{\overline{e_3^2}}{A_{V1}^2 * A_{V2}^2}}{V_{in}^2}
$$

1. It is necessary to decrease as much as possible the noise contribution 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

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

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

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 μprocessors

- Very low power consumption

- The new deep submicron CMOS tech. (< 130 nm) are **rad-hard** and suitable for S-LHC, ILC, Space applications

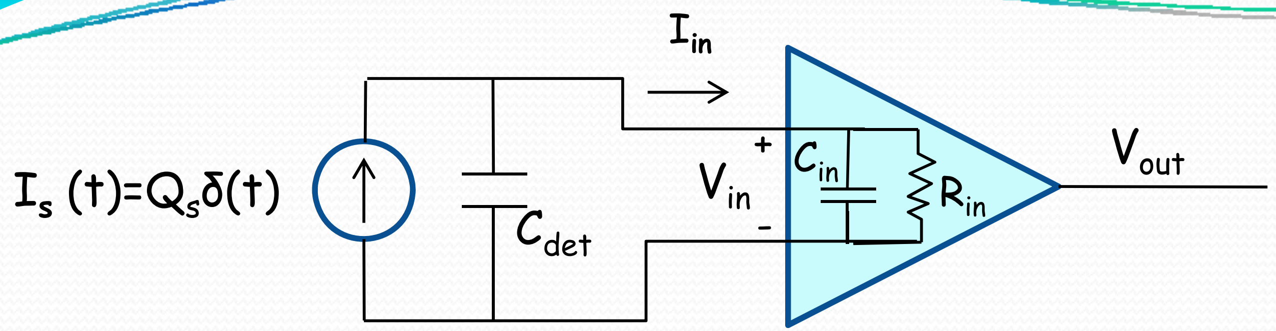
Monolithic technologies for custom applications

- 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
- SiGe
  - high speed
  - expensive
- GaAs
  - not suitable for analog applications with high bandwidth

Most used technology is CMOS

FRONT-END SCHEMES

Signal integration



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 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

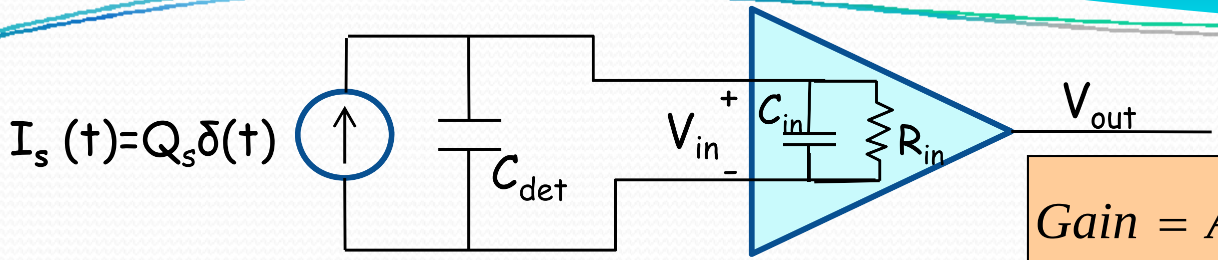
- We must integrate I to have a measurement of E: $$E \propto Q_S = \int I_S(t)dt$$

WHERE to integrate?

OPTIONS (depending on charge collection time t_c and input time constant R_iC_T:

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

1. 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 $C_t = C_{det} + C_{in}$

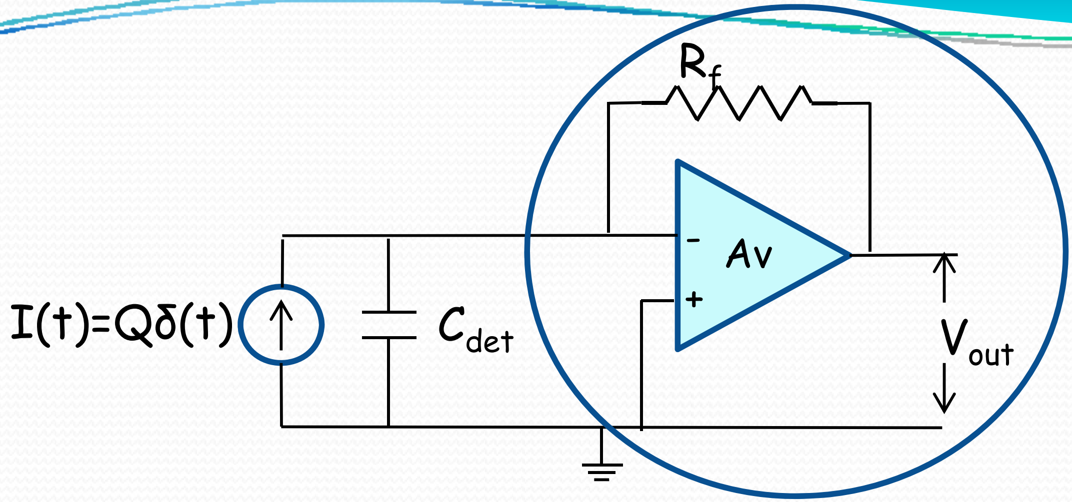
$$V_{in} = \frac{1}{C_t}\int I_s dt = \frac{Q_s}{C_{det} + C_{in}}$$

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

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
- ....

2. 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

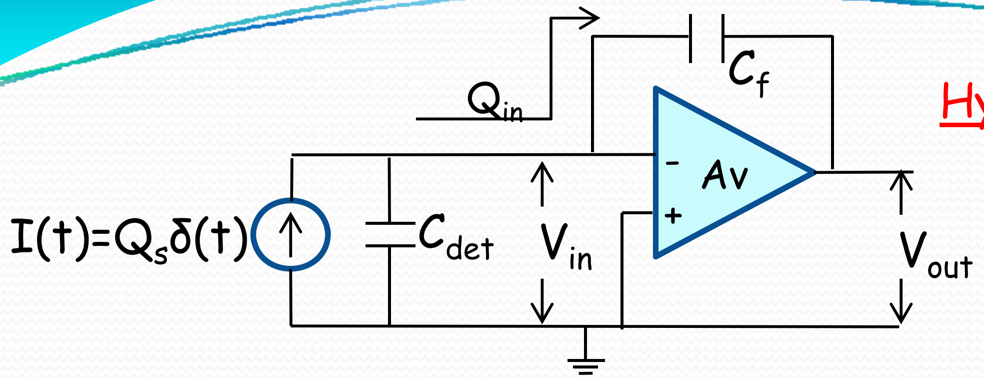
Using a transresistance amplifier (high gain operational amplifier with resistive feedback) we have:

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

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

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



Hypothesis:

1. input impedance of op-amp is ∞ (i.e. MOS gate) → 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: $Q_f = C_f V_f = C_f (A_v+1)V_{in} = Q_{in}$ (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 remains on C_{det} !!!

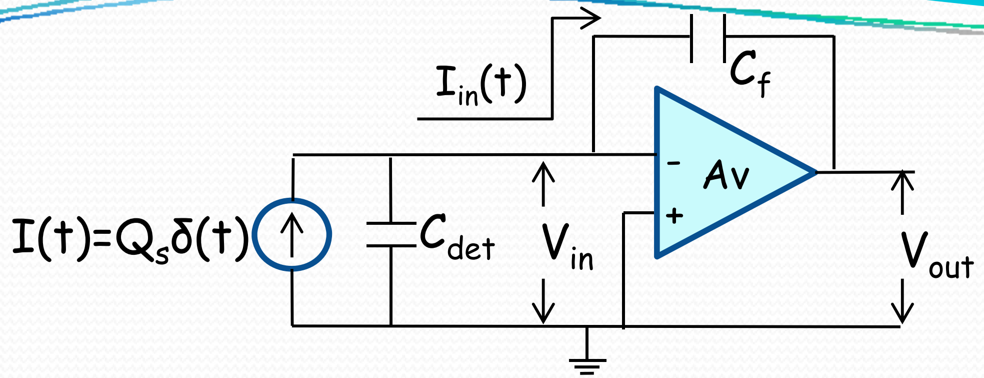
Charge transfer efficiency:

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

(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:

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

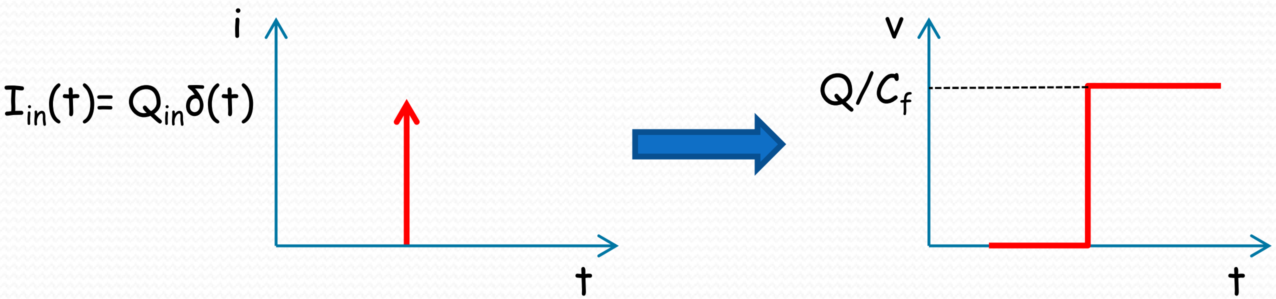
(infinite bandwidth)

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

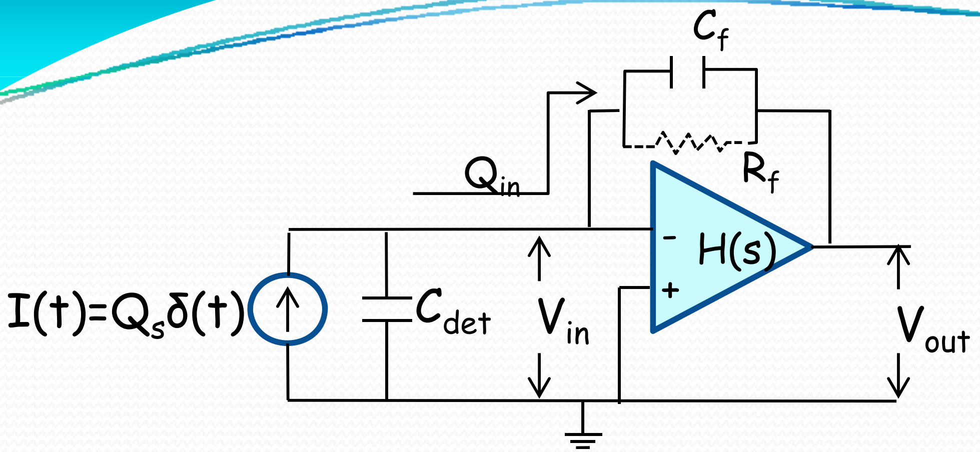
$$\longrightarrow \boxed{V_{out}(\omega) = -\frac{I_{in}(\omega)}{j\omega C_f}\left(\frac{1}{1+\frac{1}{A_v}}\right) \approx -\frac{I_{in}(\omega)}{j\omega C_f}}$$

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

Step function



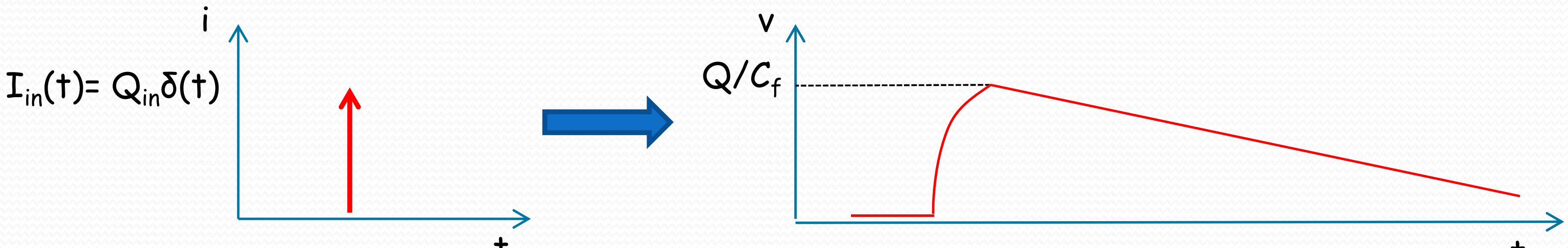
Charge-sensitive amplifier: the realistic case



Two elements to be considered:

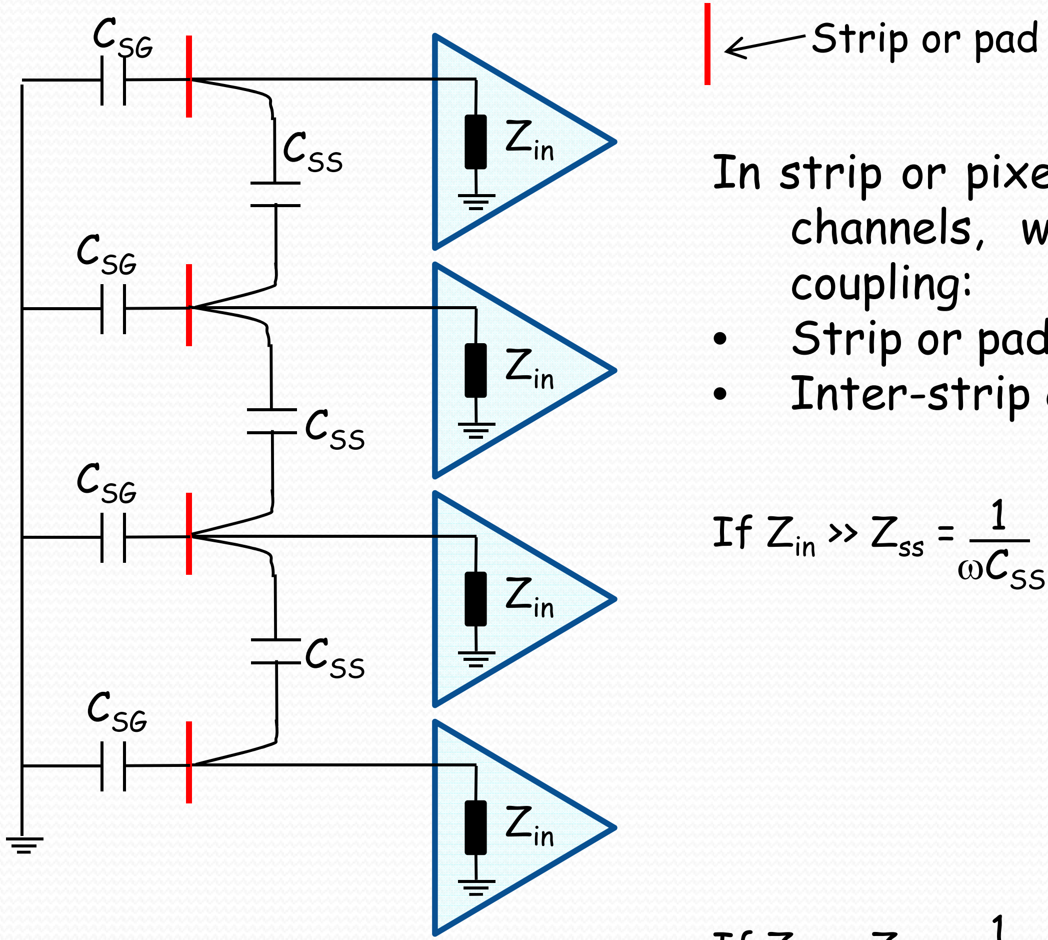
1. Resistor R_f used to discharge C_f. Since it is a source of parallel noise (inject noise current into input noise), it must made very large to decrease its contribution to noise. Typical values are several tens or hundreds of MΩ
2. Real amplifier (finite bandwidth and gain)

$$\frac{V_{out}(s)}{I(s)} = \frac{\dfrac{-g_m}{C_L C_t}}{\left(s+\dfrac{1}{R_f C_f}\right)\left(s+\dfrac{1}{R_i C_T}\right)} \xrightarrow{\text{2 poles}} \begin{cases} p_1 = \dfrac{1}{R_f C_f} & \text{(Low freq)} \\ p_2 = \dfrac{1}{R_i C_T} = \dfrac{\omega_0 C_f}{C_T} & \text{(High freq)} \end{cases} \Rightarrow \begin{cases} \tau_1 = R_f C_f & \text{Fall time constant} \\ \tau_2 = R_i C_T = \dfrac{C_T}{\omega_0 C_f} & \text{Rise time constant} \end{cases}$$



- 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

Input impedance vs crosstalk



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

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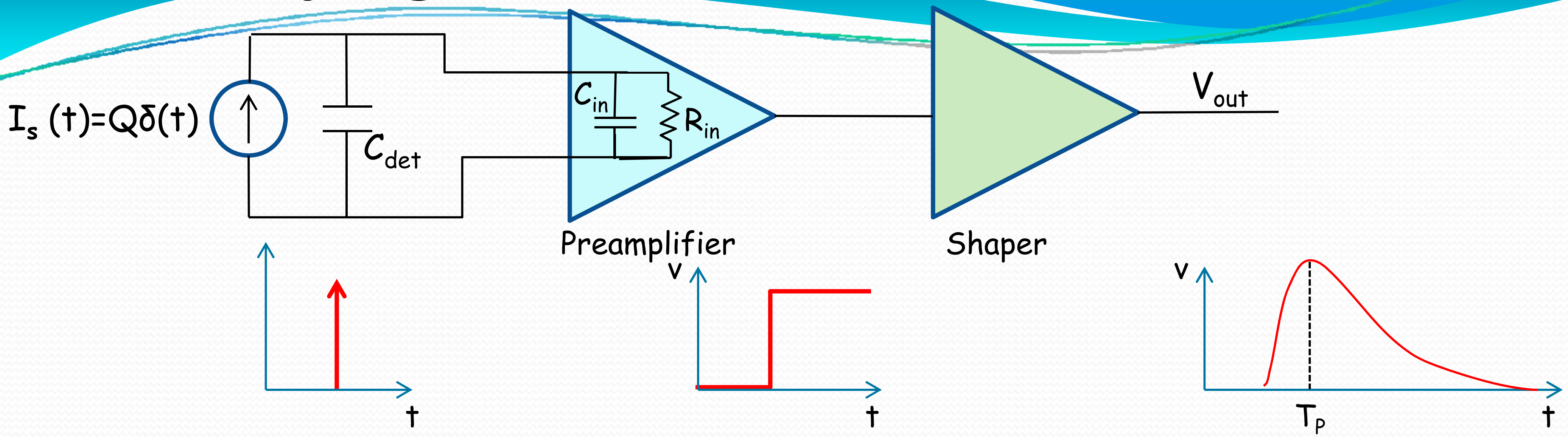
Pulse shaping

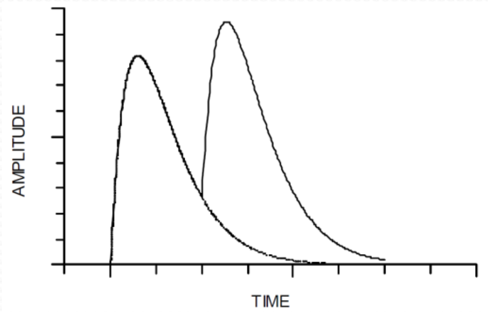


Preamplifier = input amplifier   It is usually located close to detector and must have enough gain to make negligible the effects of induced noise. Typical example: Charge Sensitive Amplifier

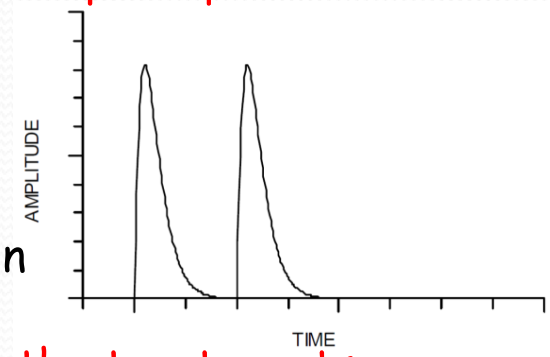
Shaper Two conflicting objectives:

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

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

$\omega = 2\pi f$

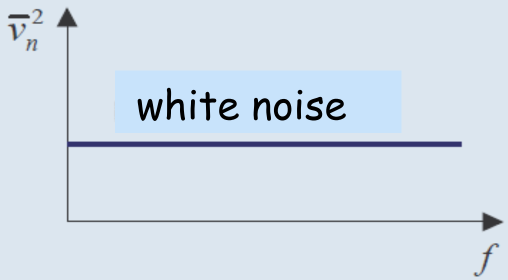
Noise power spectrum at output of a filter with transfer function H(jω) is equal to input power spectrum multiplied by squared transfer function

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

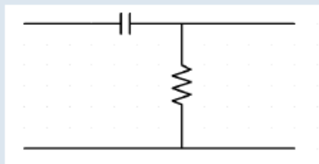
$$v_{on}^2 = \int_0^\infty \overline{v_{un}^2}\, d\omega = \int_0^\infty \overline{v_n^2} * |H(\omega)|^2 d\omega$$

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

<u>Example: white noise source connected to high-pass filter</u>



$\rightarrow \overline{v}_n^2 \rightarrow$

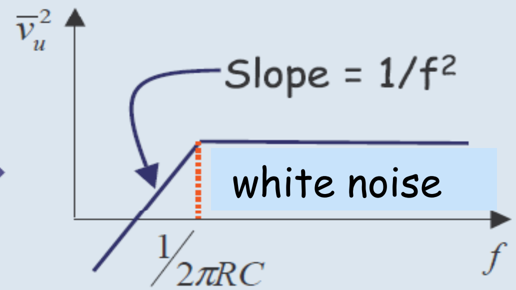


$\overline{v}_u^2 = ?$

$$H(j\omega) = \frac{j\omega RC}{1 + j\omega RC}$$

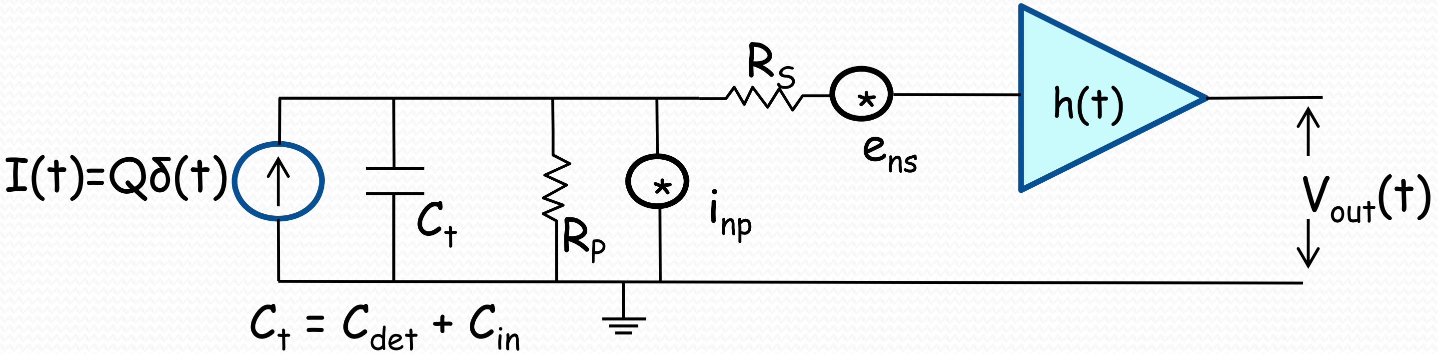
$$|H(j\omega)|^2 = \frac{(\omega RC)^2}{1 + (\omega RC)^2}$$

$$\overline{v}_u^2 = \frac{(\omega RC)^2}{1 + (\omega RC)^2} \cdot \overline{v}_n^2$$



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 → $R_S = \dfrac{a_n}{g_m}$ $\quad a_n = \begin{cases} 0.5 \text{ in BJT} \\ 0.7 \text{ in Mosfet} \end{cases}$

g_m = conductance = $\dfrac{\partial I}{\partial V}$

It is possible to demonstrate that:

$$ENC^2 = \underbrace{2KTR_S C_t^2 \int \left[\frac{d}{dt} h(t)\right]^2 dt}_{\text{series noise}} + \underbrace{\frac{2KT}{R_p} \int [h(t)]^2 dt}_{\text{parallel noise}} \Rightarrow ENC^2 = 2KTR_s C_t^2 \left[\int [h'(t)]^2 dt + \frac{1}{\tau_C^2} \int [h(t)]^2 dt\right]$$

where $\tau_C = C_t \sqrt{R_P R_S}$

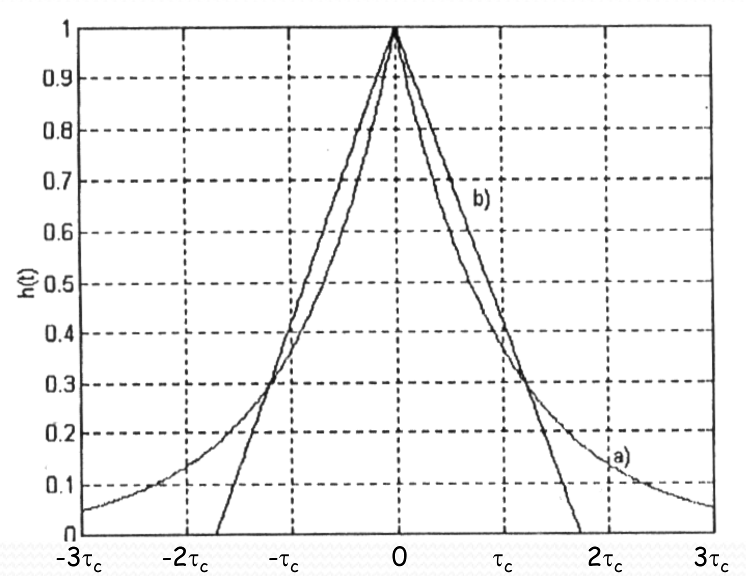
<u>noise corner time constant</u>

Optimum filter

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

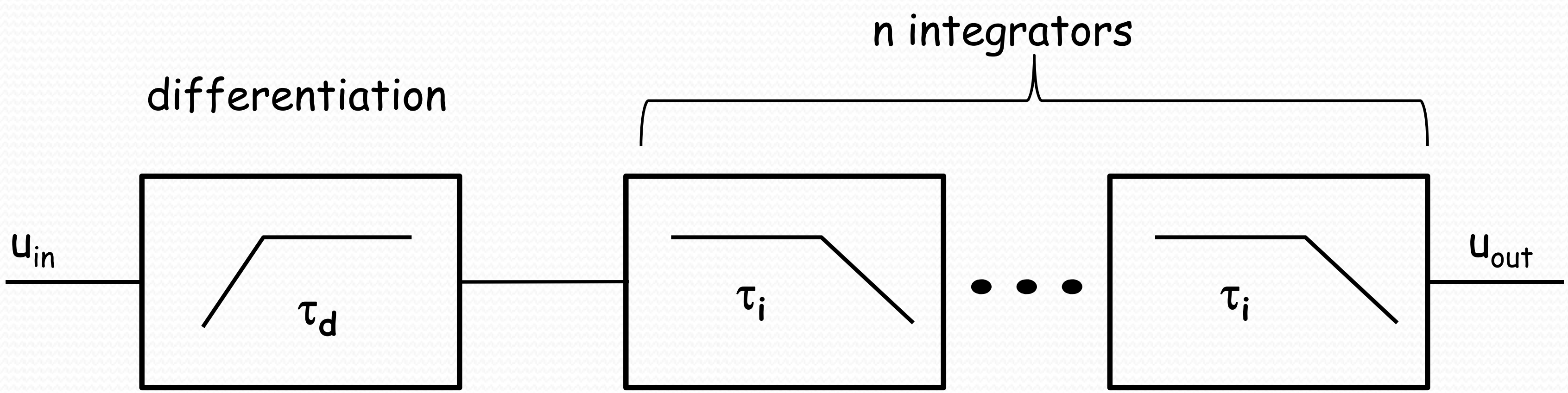
It is possible to demonstrate that $h_{opt}(t) = \exp\left(-\frac{|t|}{\tau_c}\right)$ $\longrightarrow$ $ENC^2_{opt} = 2KTR_S \frac{C_t^2}{\tau_c} = 2KTC_t\sqrt{\frac{R_s}{R_p}}$

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



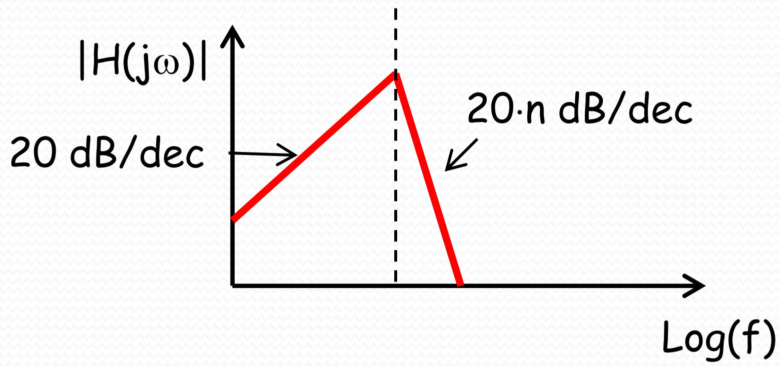
The cusp filter is not practically feasible, but can be approximated by triangular shapers (curve b) or Pseudo-Gaussian shaper

Pseudo-Gaussian (or Semi-Gaussian) shaper



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

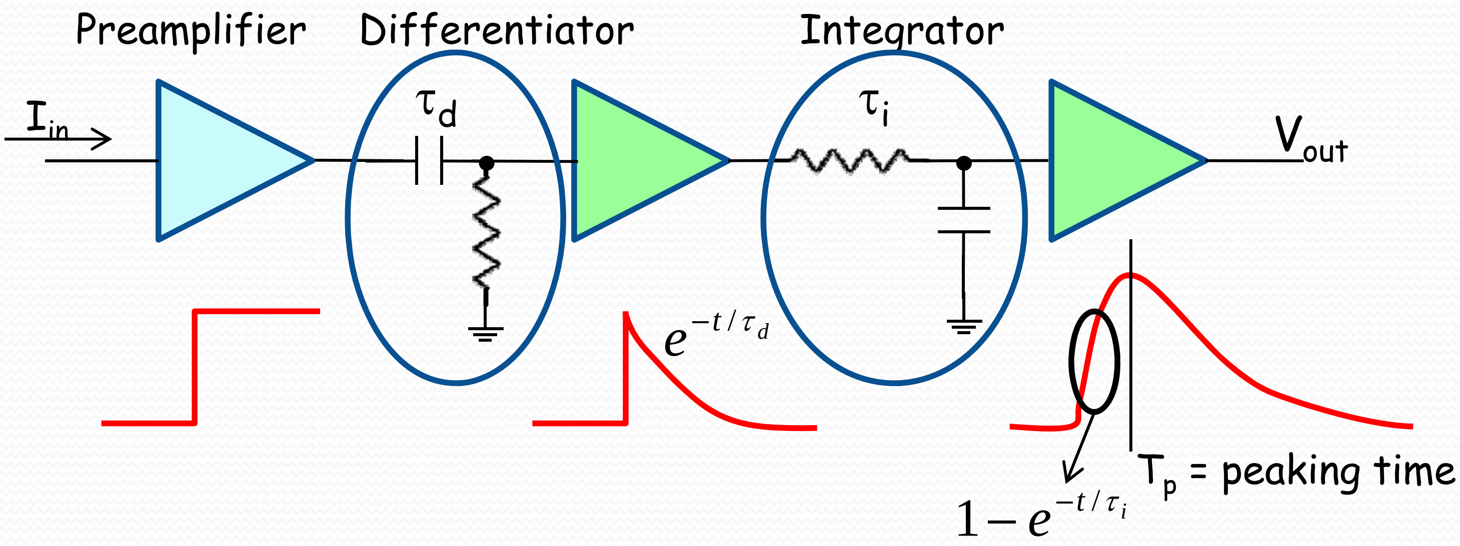
$$H(s) = \frac{u_{out}(s)}{u_{in}(s)} = \frac{s\tau_d}{(1+s\tau_d)} \frac{1}{(1+s\tau_i)^n}$$



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



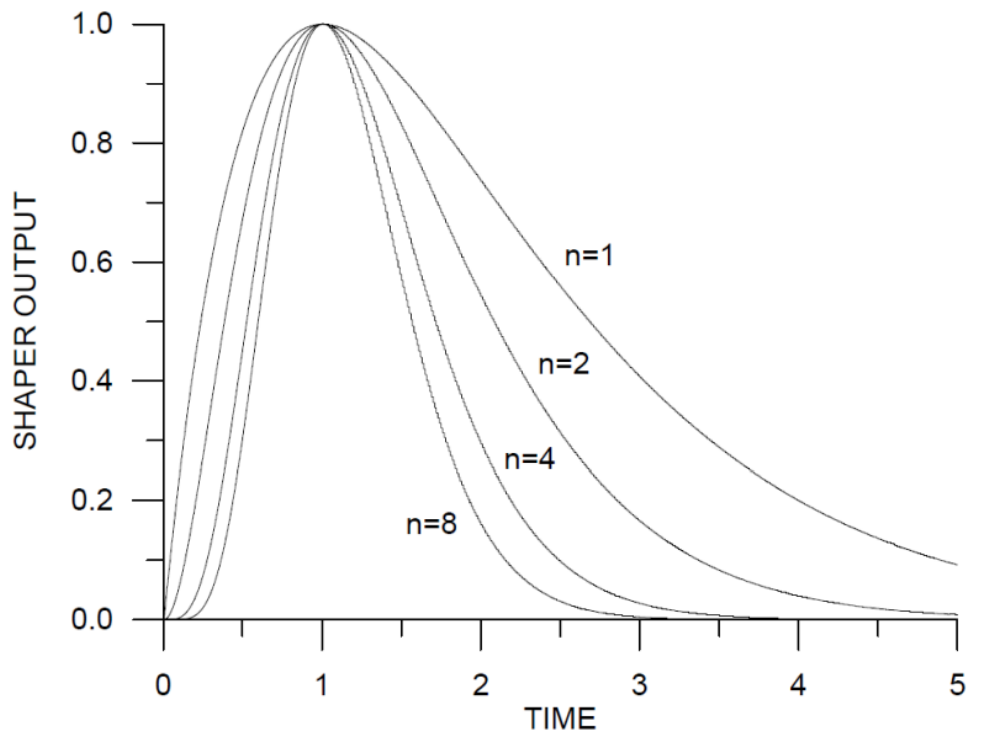
• This shaper is called CR-RC because the high-pass filter is made with CR network, while the low-pass filter with a RC network

• The noise is 36% worse than “optimum filter” with the same time constants

Shaper: CR-RC^n

The shapers are often more complicated, with multiple (n) integrators → CR-RC^n

- 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^2	0.84	0.63
CR-RC^3	0.95	0.51
CR-RC^4	0.99	0.45
CR-RC^5	1.11	0.4
CR-RC^6	1.16	0.36
CR-RC^7	1.27	0.34

- Current noise decreases with shaping order

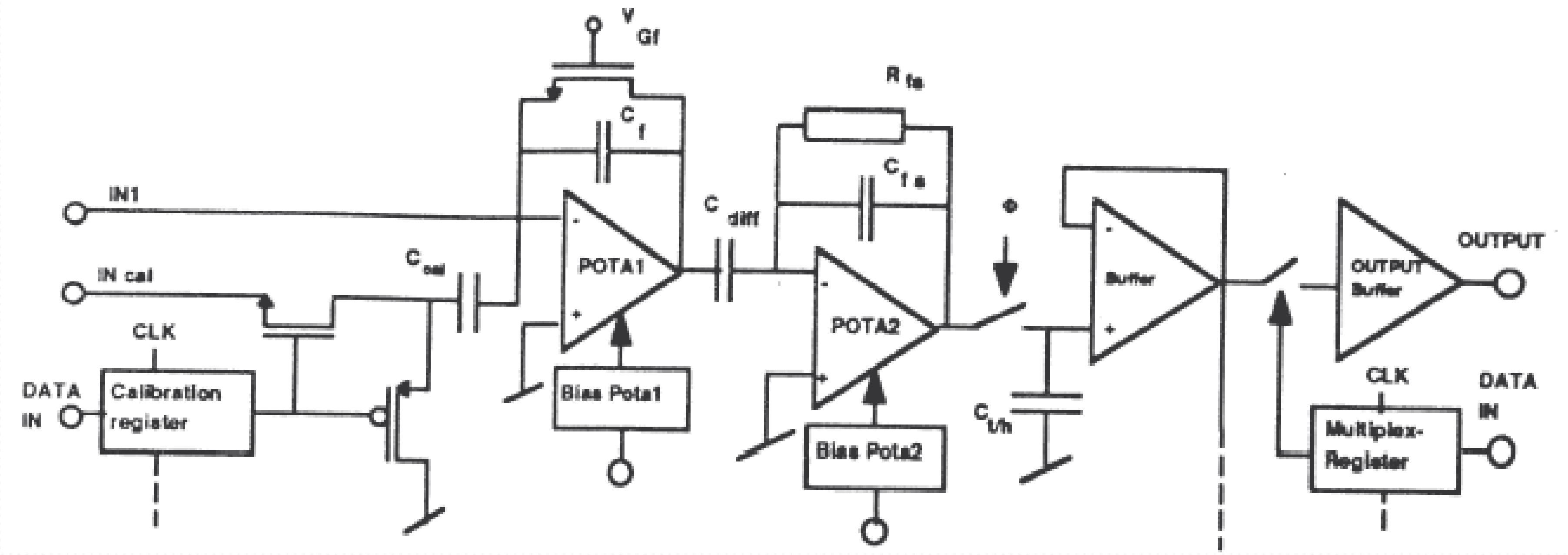
Some examples of Front-End amplifier

1. AMPLEX

2. Current sensitive amplifier: Front-end chip of Resistive Plate Chambers (RPC) for the CMS Experiment, at CERN

3. Charge sensitive amplifier: Front-end chip of Cylindrical GEM (CGEM) for the KLOE Experiment, at Frascati INFN LAB

1. Amplex – SiCAL (1992)



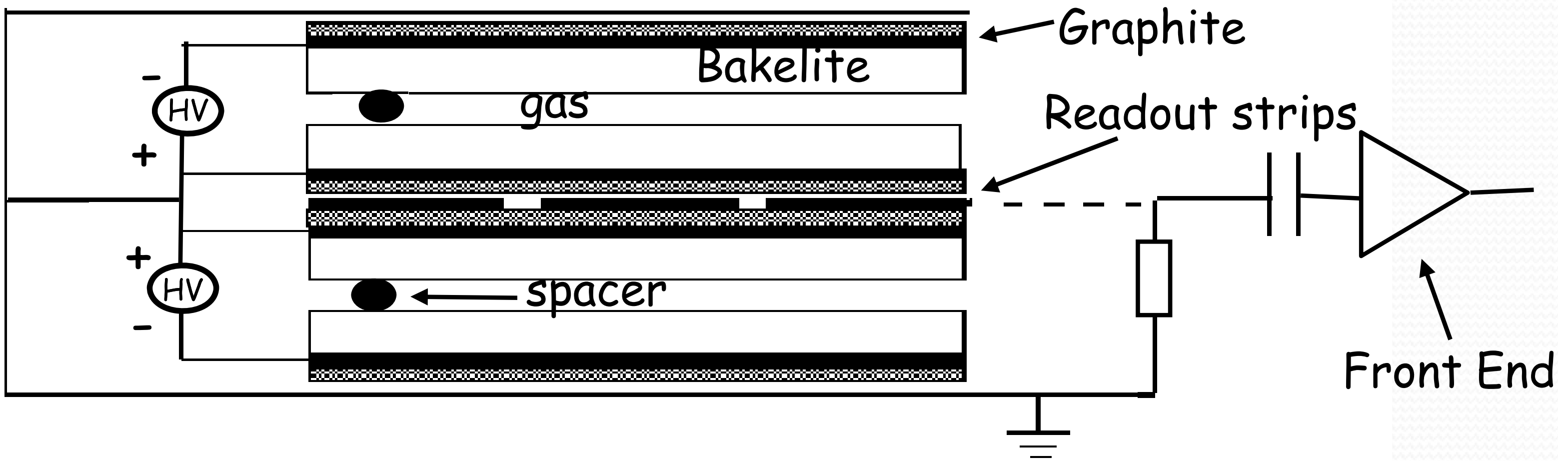
- Readout of silicon luminosity calorimeter of ALEPH Experiment
- 3 μm CMOS
- 16 channels (CSA + CR-RC Shaper + track & hold + multiplexer)
- Power = 100 mW
- Noise: 800 e^- rms + 38 e^- rms/pF

2. 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 (~ ns)

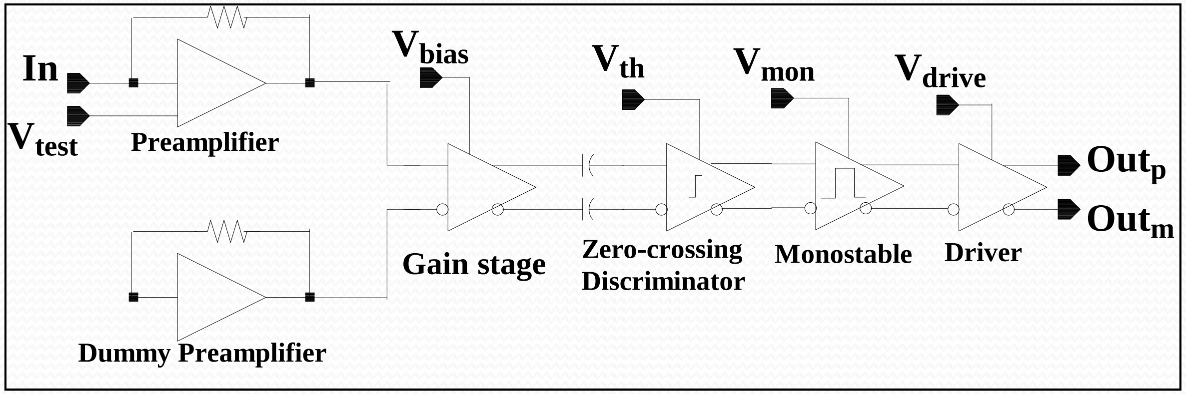
- They are suitable for fast space-time particle tracking as required for the muon trigger at the LHC experiments



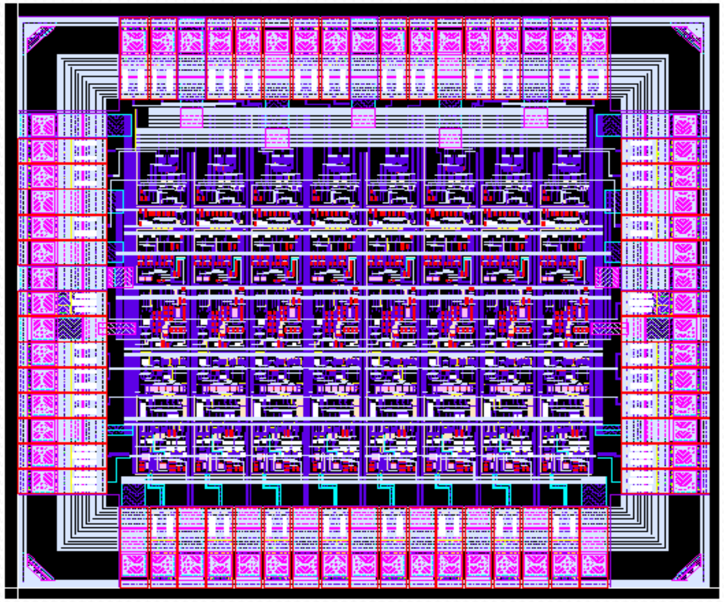
Cross sectional view of a double-gap RPC

CMS RPC Front-End

- RPC timing information is crucial for unambiguous assignment of the event to the related bunch crossing (T = 25 ns)
- We have developed an 8-channel front-end chip in BiCMOS technology



Single channel block diagram



Technology: 0.8 μm BiCMOS of AMS
8 channels
Power supplies: +5 V; GND
Power consumption: ~ 45 mW/channel

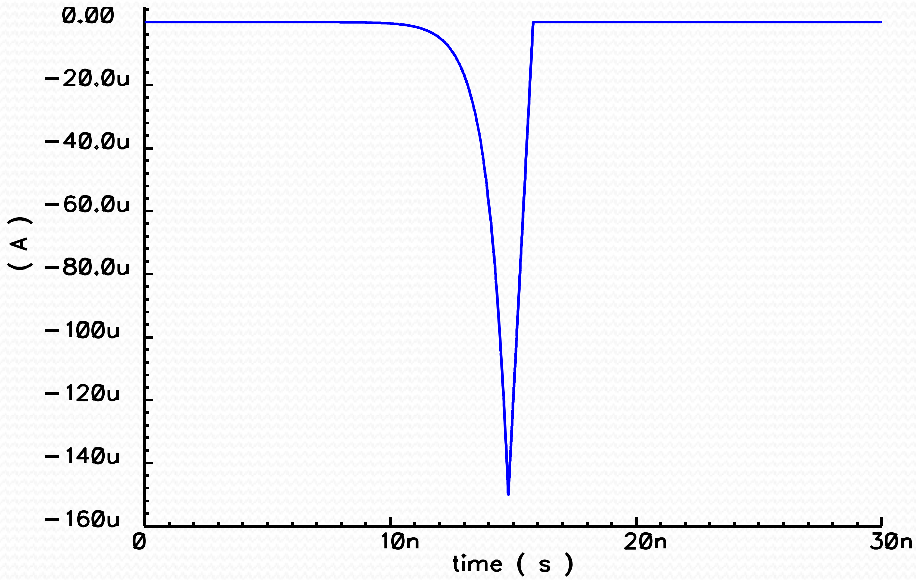
Dimensions: 2.9 mm X 2.6 mm
64 I/O pads
Package: PQFP 80

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

CMS RPC_FE Preamp

Shape of the signal: $I(t) = I_0 \exp(t/\tau)$ $(0 \leq t \leq 15\ \text{ns})$

$\tau \sim 1$ ns is the gas time constant

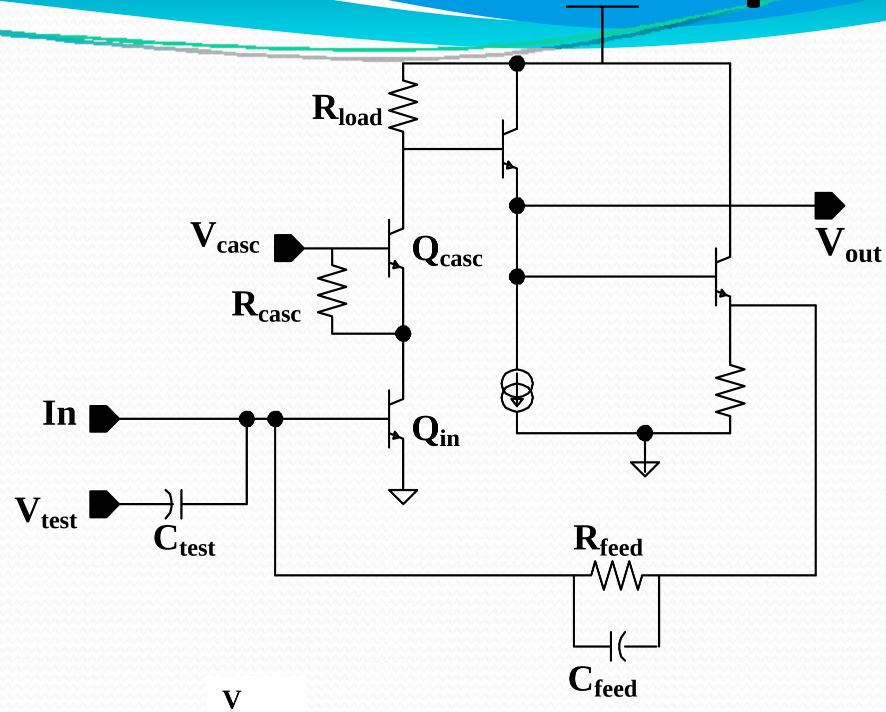


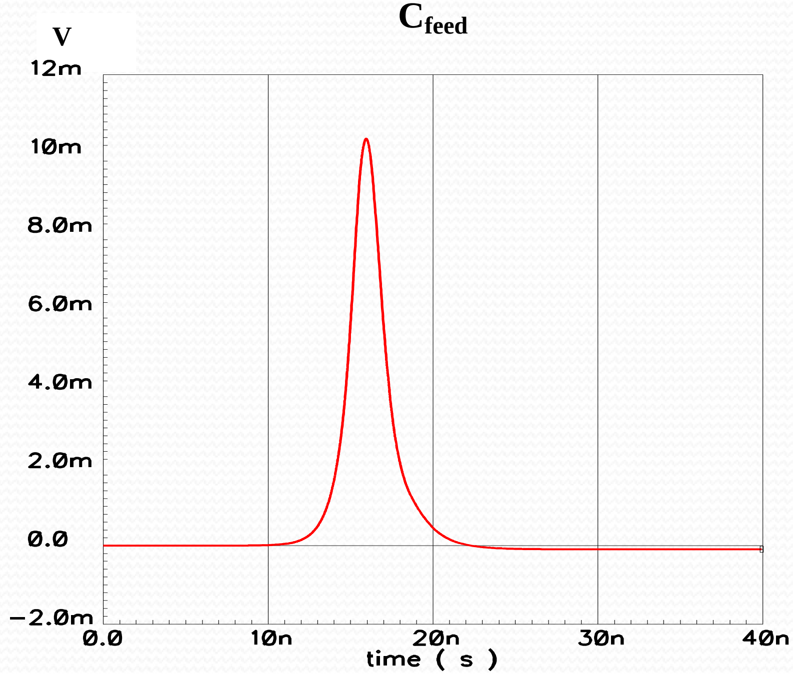
- Strip line 1.3 m long and 2 - 4 cm wide: $15\ \Omega \leq R_0 \leq 40\ \Omega$
  $160\ \text{pF} \leq C_{strip} \leq 350\ \text{pF}$
  Propagation delay ~ 5.5 ns/m
- Input dynamic range: $20\ \text{fC} \leq Q_{in} \leq 20\ \text{pC}$
- Expected rate: < 400 kHz/strip

CMS RPC_FE: The Current Sensitive Preamplifier

- Common emitter transimpedance stage
- Low input impedance: $R_{in} \sim 15\ \Omega$
- Preserve the shape of input signal
- Bandwidth: 116 MHz
- $I_{in} = 700\ \mu A$
- $g_m \sim 25$ mA/V
- $R_{feed} \sim 1\ K\Omega$
- $C_{feed} \sim 1$ pF (added for stability)
- $C_{test} \sim 1$ pF (1V → 1 pC)
- Charge sensitivity: 0.5 mV/fC
- ENC ~ 2 fC → No semi-gaussian shaper needed

  (only gain stage)

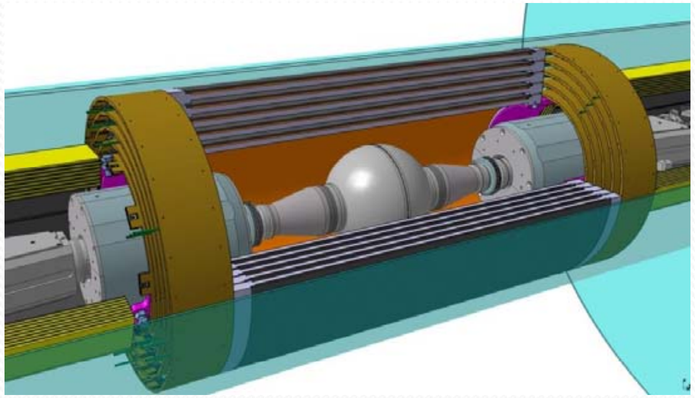




3. KLOE Inner Tracker Front-End

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)

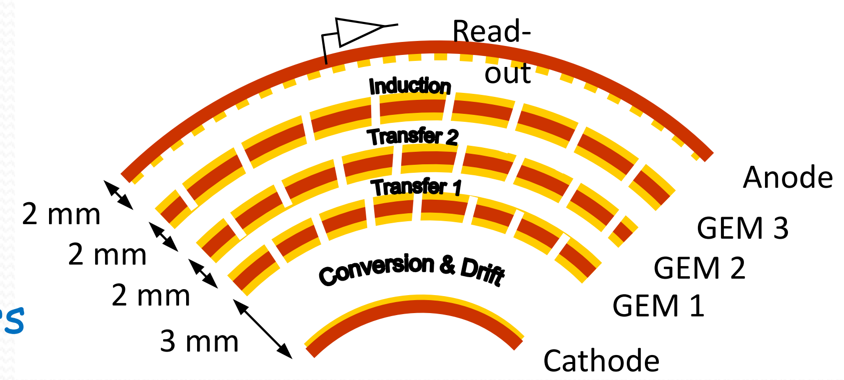


Gas Electron Multiplier (GEM) technology has been adopted, for its low material budget

- 4 independent tracking layers for a fine vertex reconstruction of K_S and η
- 200 μm $\sigma_{r\varphi}$ and 500 μm σ_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^2 rate capability

Realized with Cylindrical TRIPLE_GEM detectors

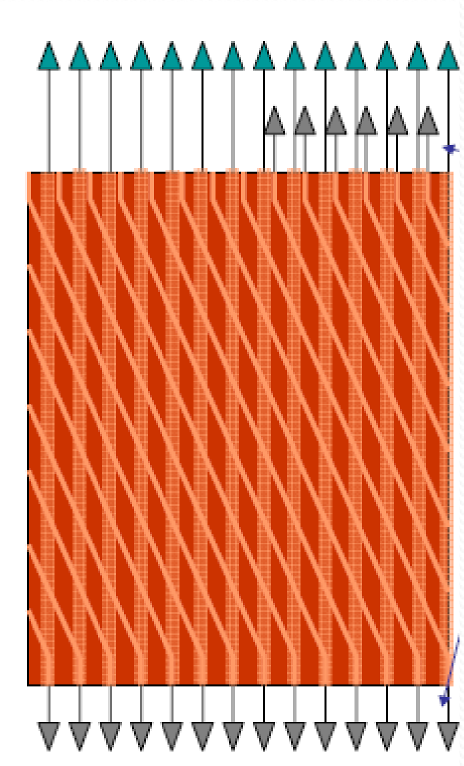
Cylindrical Triple GEM



KLOE IT Readout

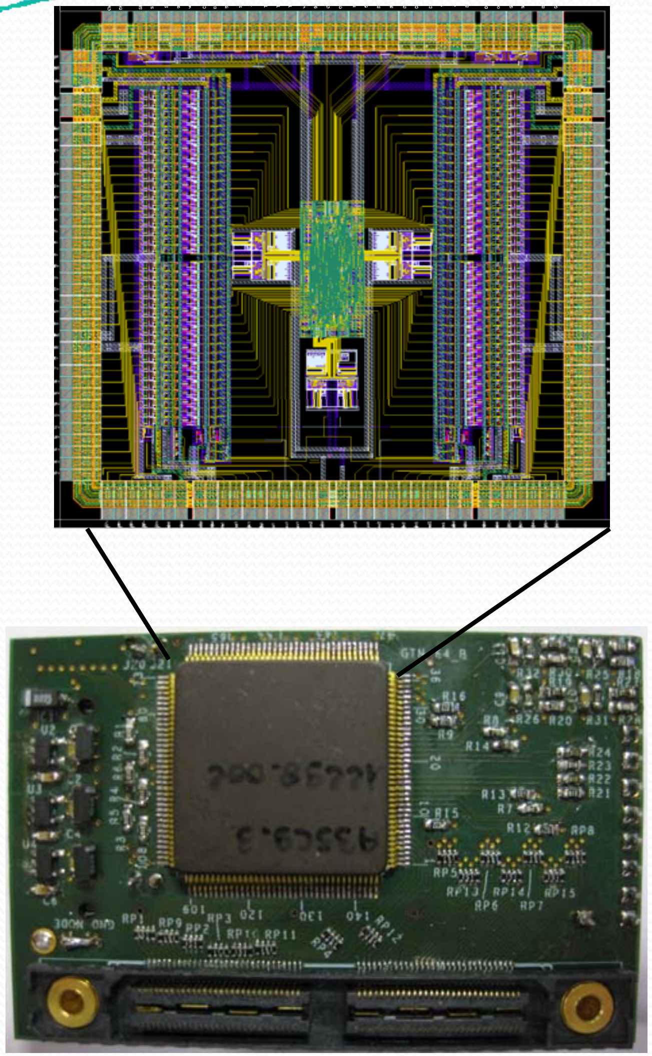
The Anode is shared out in a 2 dimensions layout having a XV geometry, X and V strips read out is shared out at both ends.

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



Prototpye of Layer #2

KLOE IT Front-End: GASTONE64

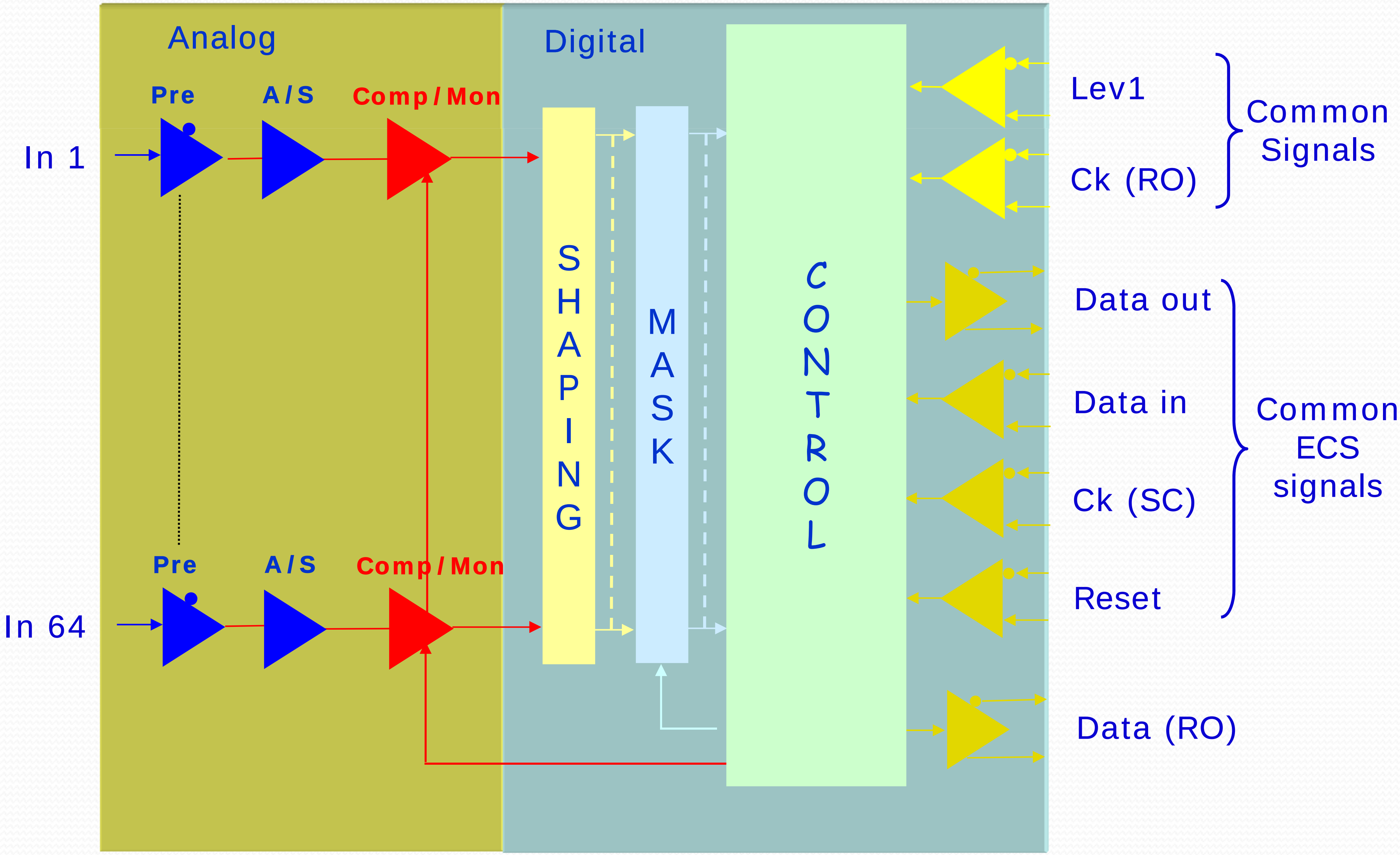


N. channels	64
Technology	CMOS 0.35 μm
Chip dimensions	4.5 X 4.5 mm^2
Input impedance	400 Ω
Charge sensitivity	22 mV/fC (Cdet = 0 pF)
Peaking time	80 ns÷150 ns (Cdet=0 ÷100 pF)
Crosstalk	< 3%
ENC	800 e- + 40 e-/pF
Power consumption	~ 7.5 mW/ch
Readout	Serial LVDS (100 MBps)

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

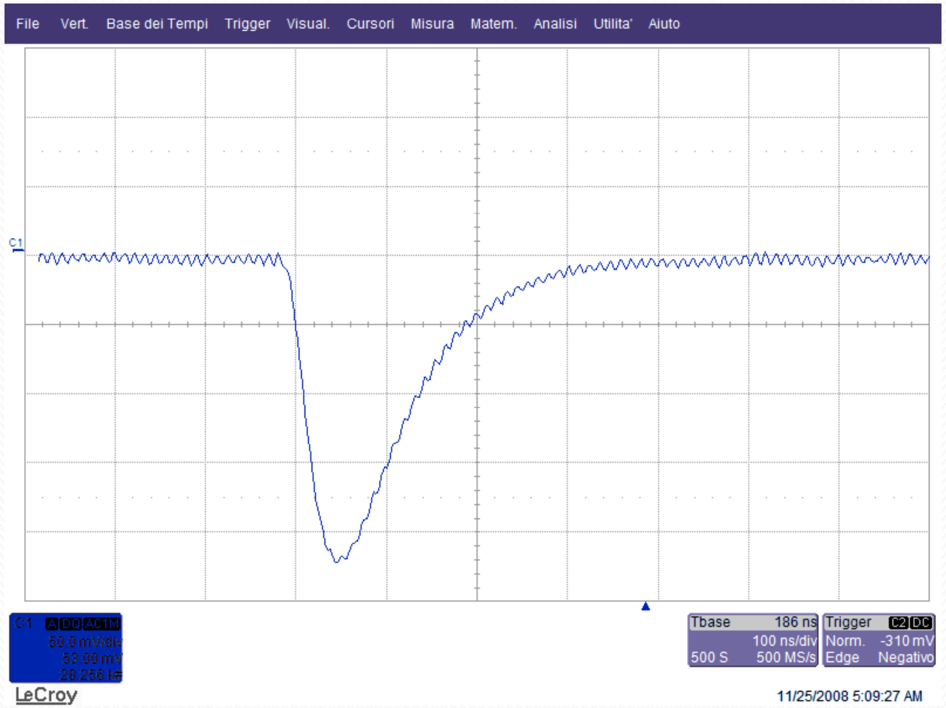
A. Balla et al., *GASTONE: A new ASIC for the cylindrical GEM inner tracker of KLOE experiment at DAFNE*, Nucl. Instr. & Meth. A 604 (2009) 23-25

KLOE IT – GASTONE64

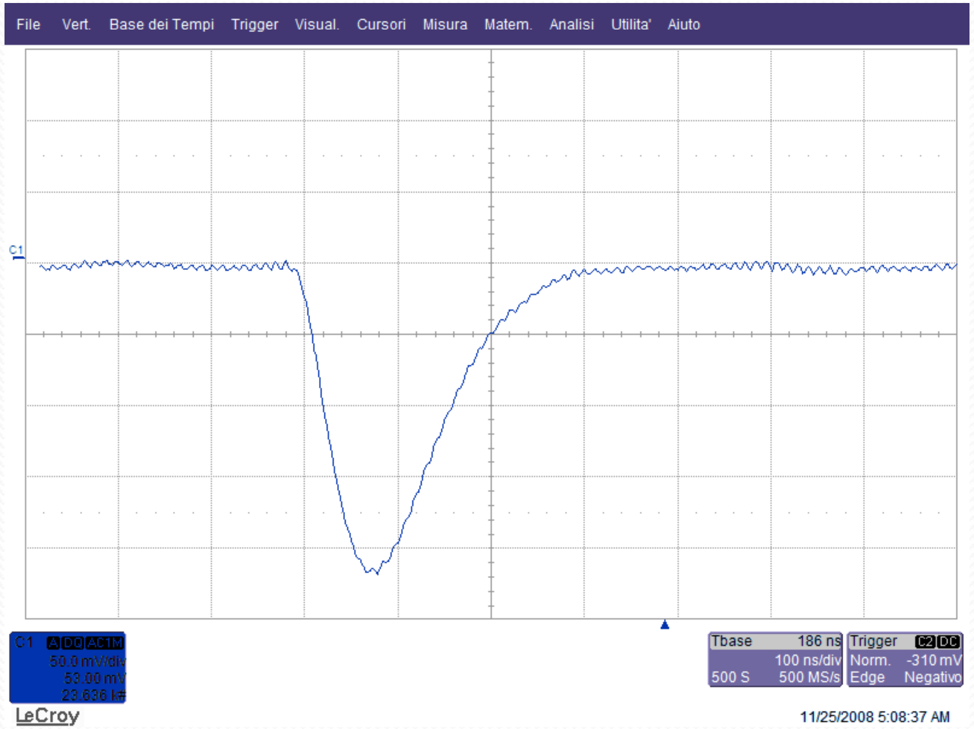


KLOE IT – GASTONE64

Shaper output (Q_{in} = 10 fC)



C_{in} = 10 pF



C_{in} = 50 pF

50 mV/div 100 ns/div

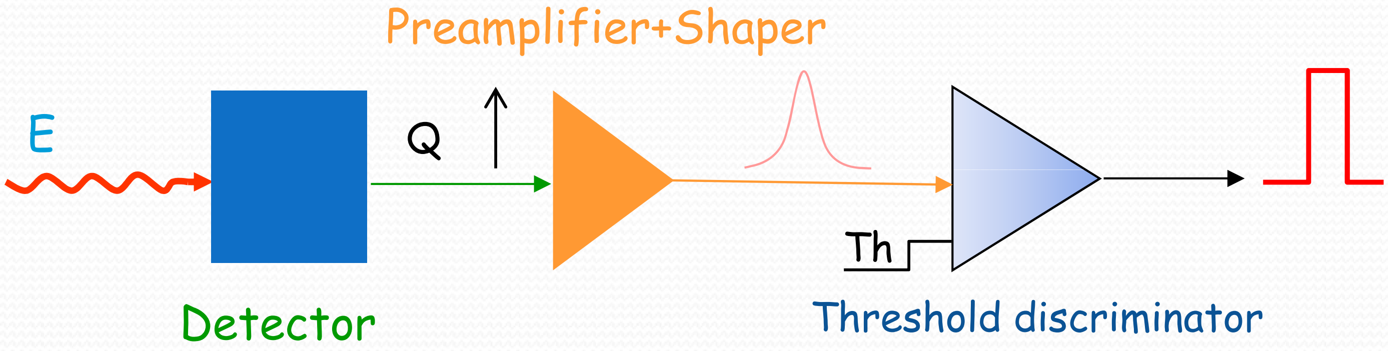
Time measurement

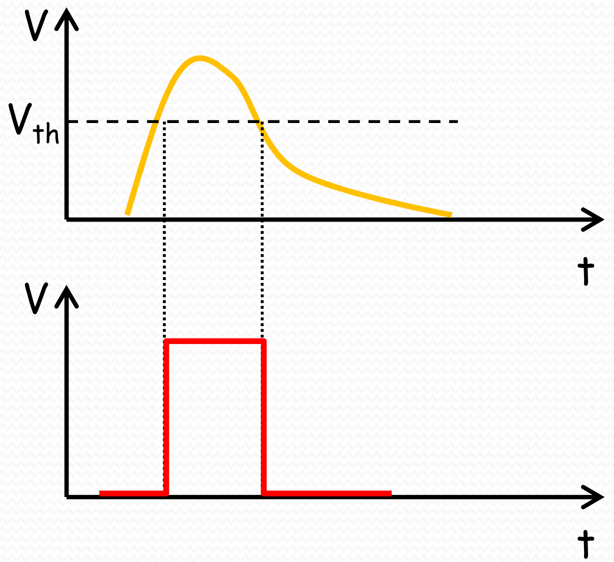
†

Time measurement

Often the purpose of the system Detector + Front End Electronics is the 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"

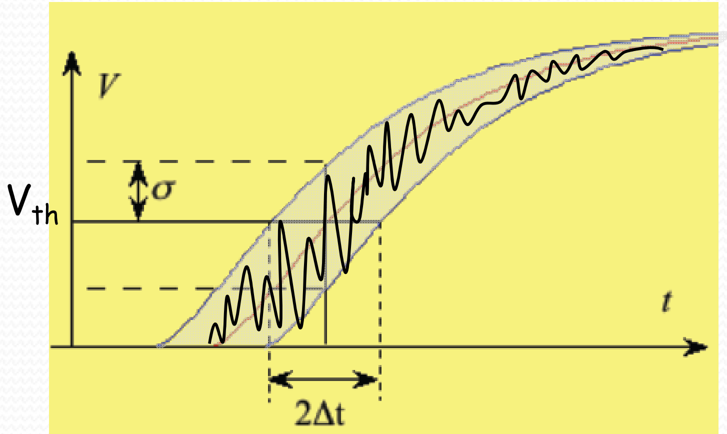
Accuracy of timing measurement is limited by:

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 → Jitter



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

→ slope

How to decrease jitter? → Conflicting conditions:

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

As usual … find compromise

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

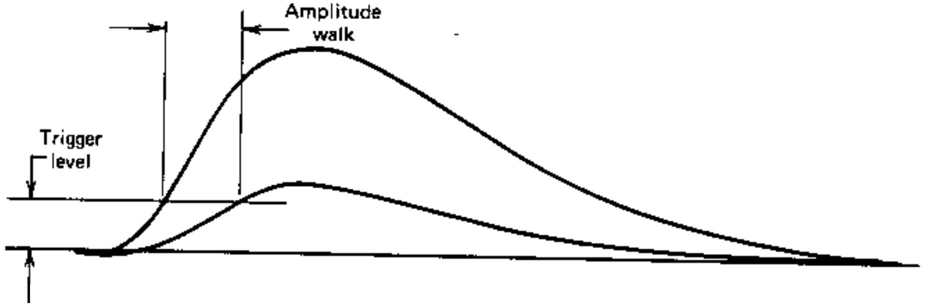
(If t « τ)

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

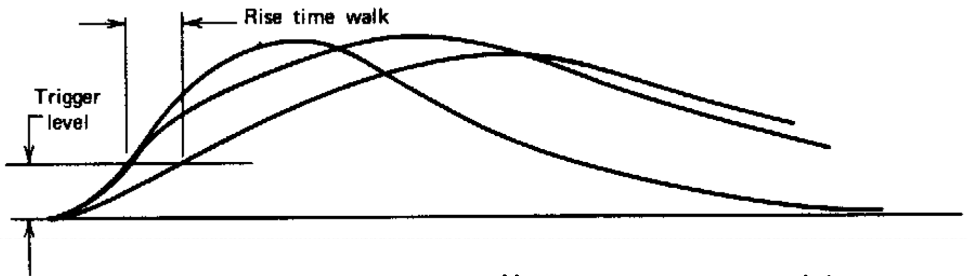
Rise time (10%-90%) = 2.2τ

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 (ΔT = 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:

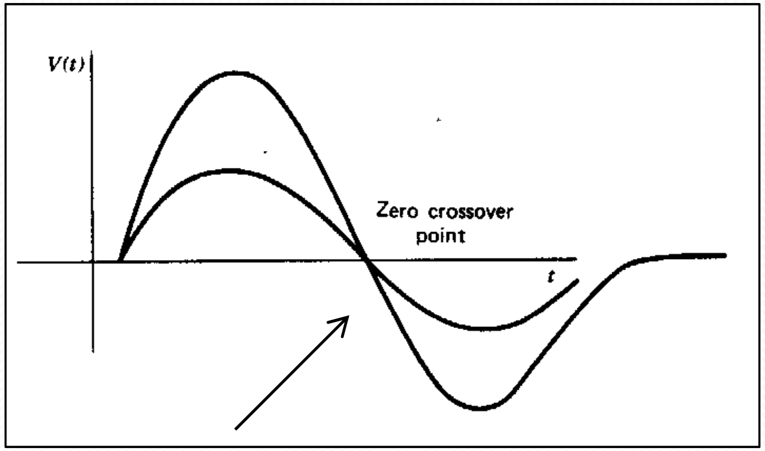
- Software: measure the pulse amplitude and apply correction to timing
- Hardware: instead of leading edge triggering, use
    1. Crossover timing
    2. Constant Fraction timing

Crossover timing

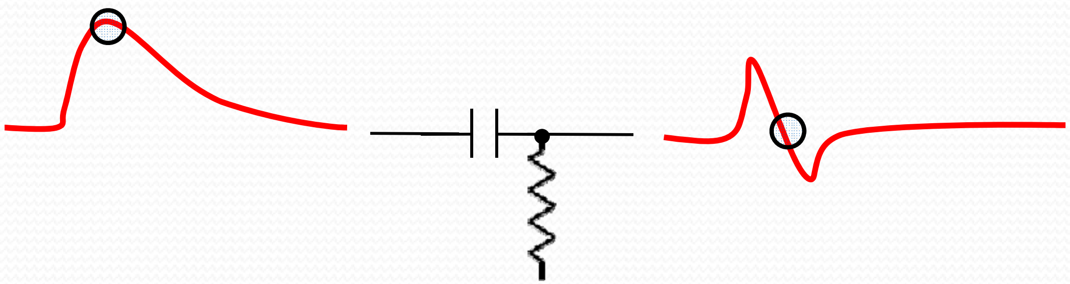
The crossover timing can greatly reduce the magnitude of the amplitude time walk

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

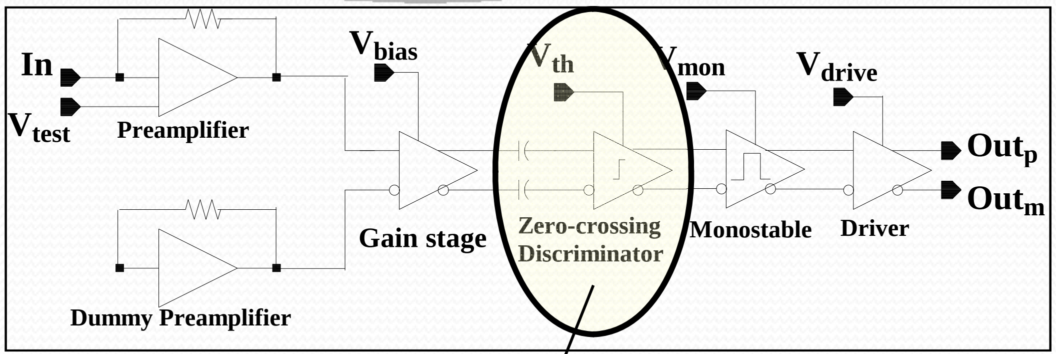


If the output of shaper is unipolar, but the peaking time is constant, adding a differentiator (C-R network) we get a bipolar pulse crossing the zero in correspondence of the signal peak



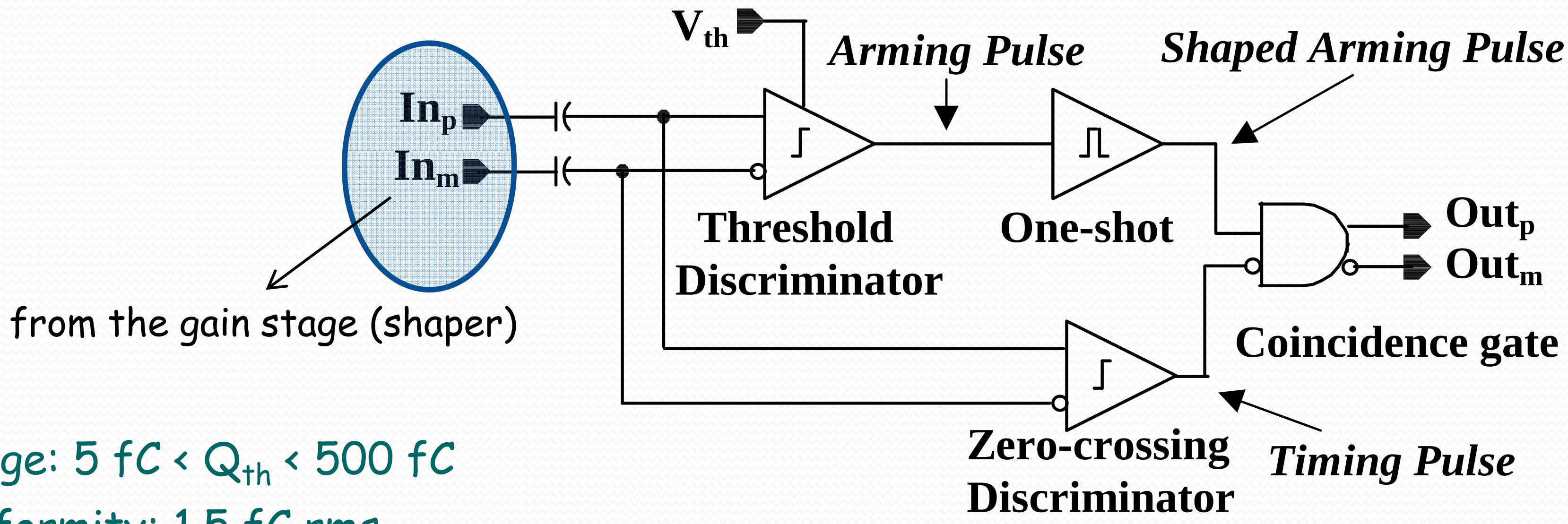
This method reduce amplitude walk, but usually jitter is larger than leading edge triggering

Example of crossover timing: CMS RPC Front-End



Zero-crossing discriminator:

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

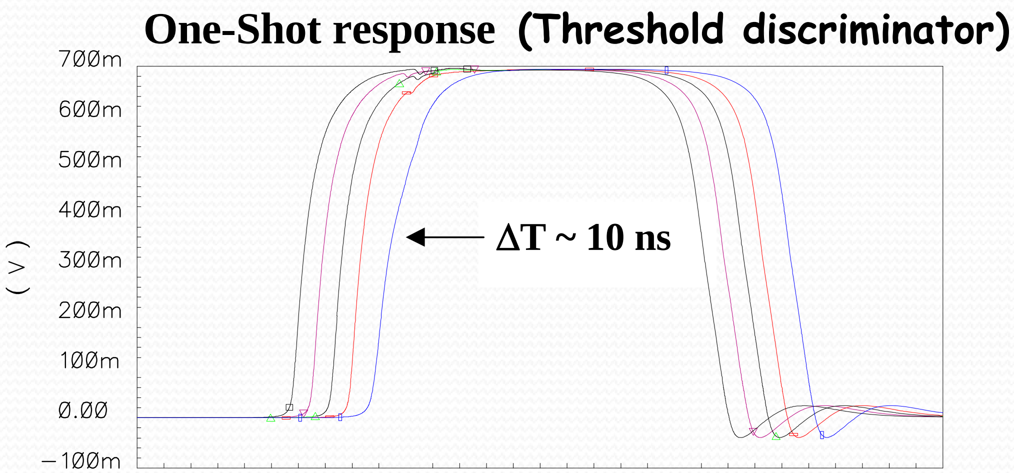


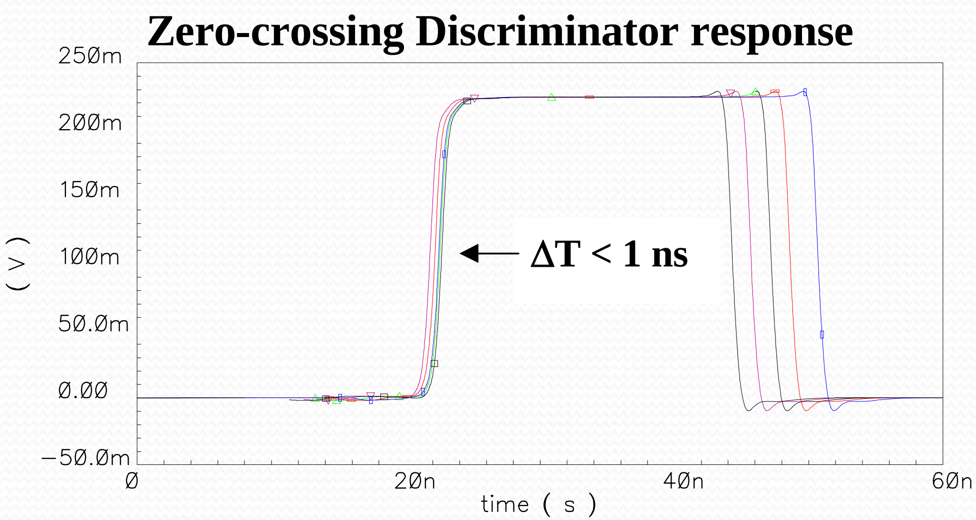
- Threshold range: 5 fC < Q_{th} < 500 fC
- Threshold uniformity: 1.5 fC rms
- Shaped Arming pulse width: 20 ns
- Power: 8 mW

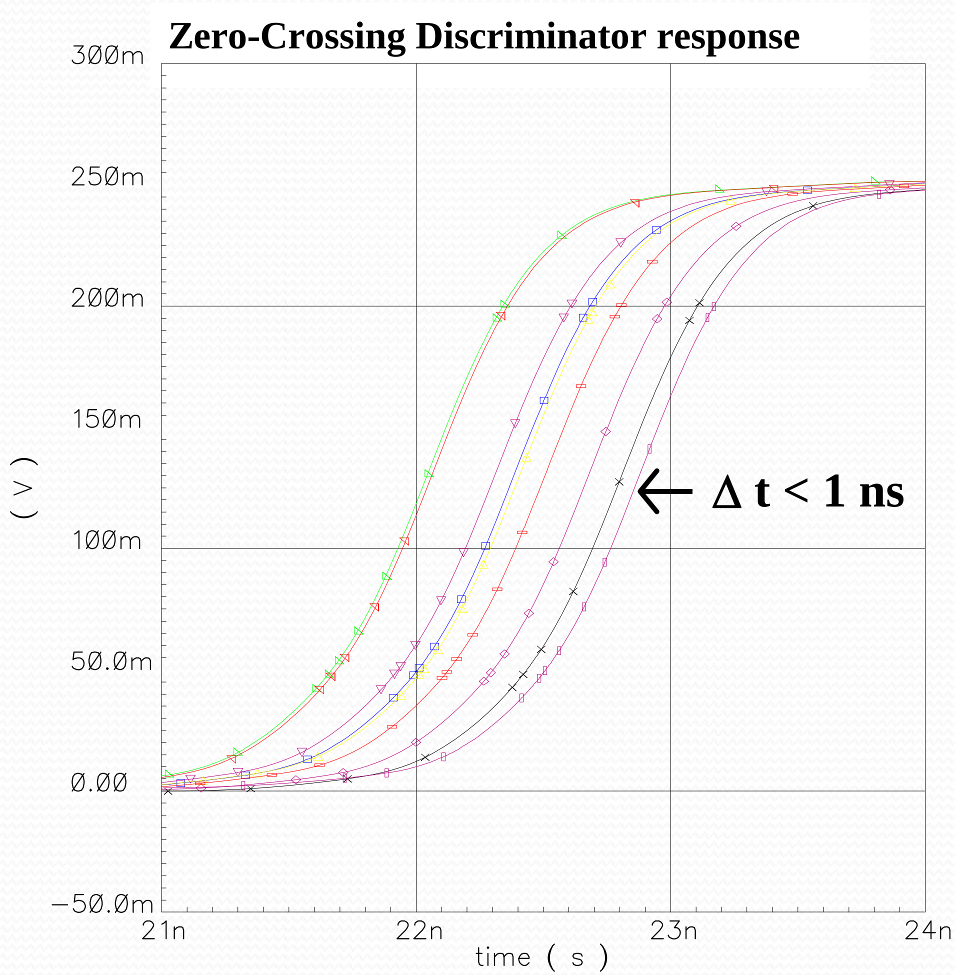
Example of crossover timing: CMS RPC Front-End

Zero-crossing discriminator response in the dynamic range

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



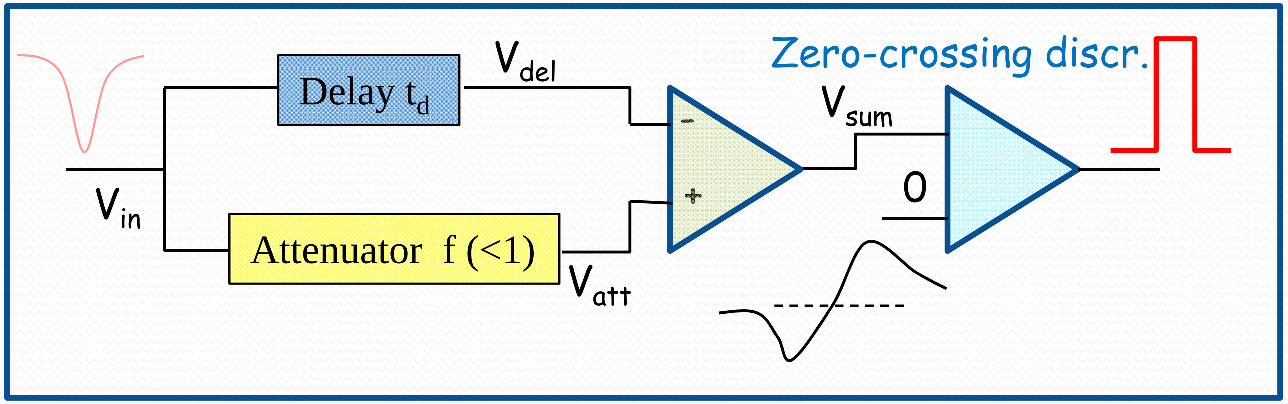




The Constant Fraction Timing

• 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



Summary

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

- 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^n filters

- 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