FRONT-END ELECTRONICS FOR GAS DETECTORS

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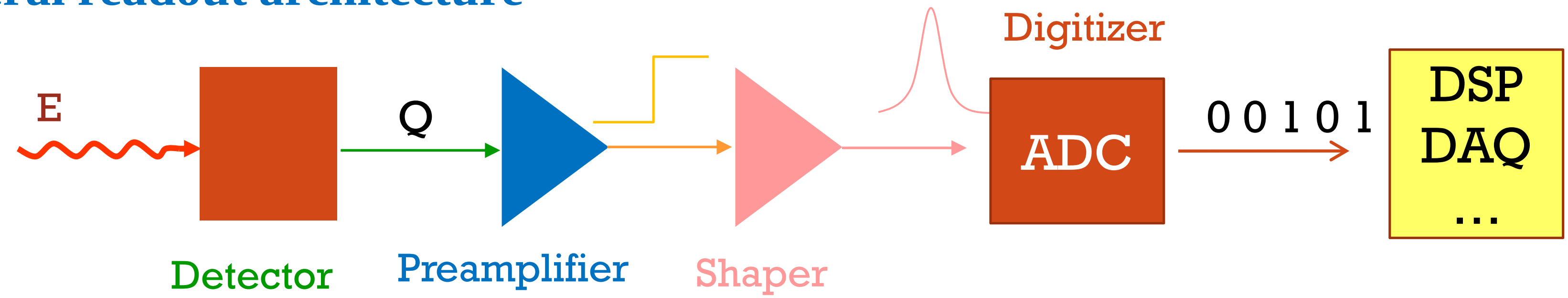
INTRODUCTION

General readout architecture



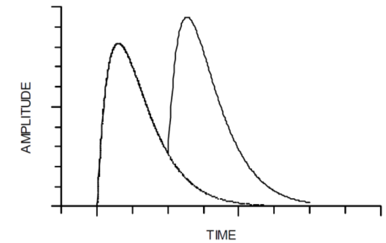
- The particle deposits energy in a detecting medium { **Gas** / **Solid** / **Liquid**
- Energy is converted into an electrical signal: $Q = KE$
- The charge Q is typically small and must be amplified, in order to be measured and processed
- The preamplifier converts Q into a voltage
- The shaper provides 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

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 detectable signal)
- Energy measurement (focus on linearity, dynamic range …)

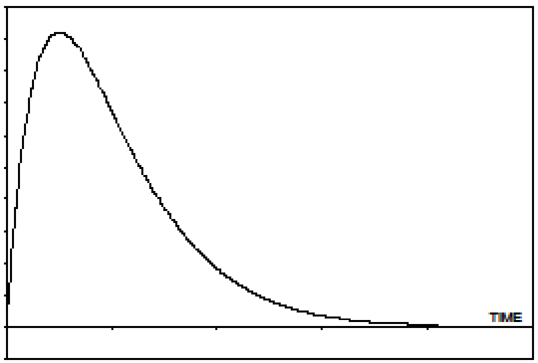
NOISE BASIC PRINCIPLES

NOISE

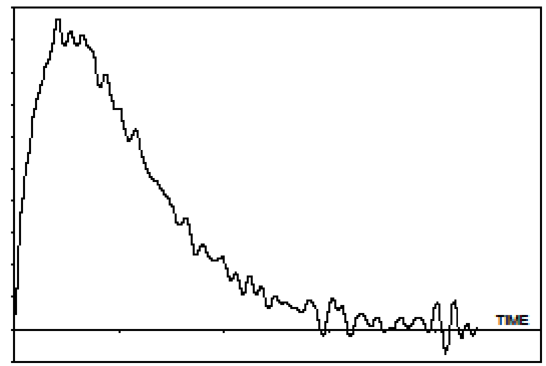
The precision of amplitude and timing measurements is limited by the NOISE

Definition

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



Signal of ideal system



Signal + Noise

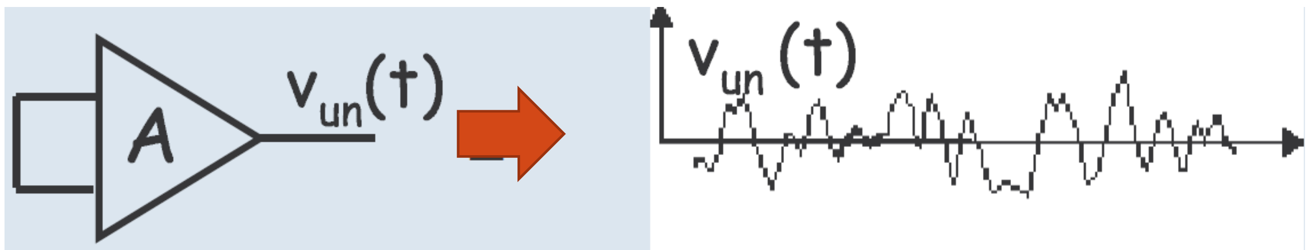
1. **External noise (interference)**
   It is generated by external sources (RF, ripple of power lines, ground loops …)
   Can be minimized by proper shielding, cabling …

2. **Intrinsic noise**
   It is a property of detector and/or electronics
   Can be reduced by proper design of front-end electronics

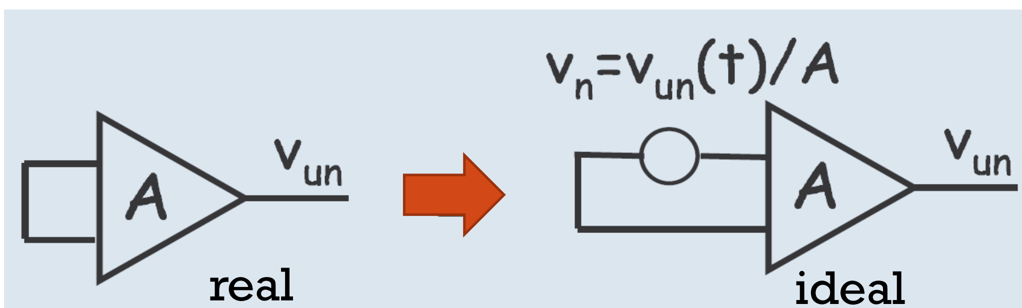
INTRINSIC NOISE

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

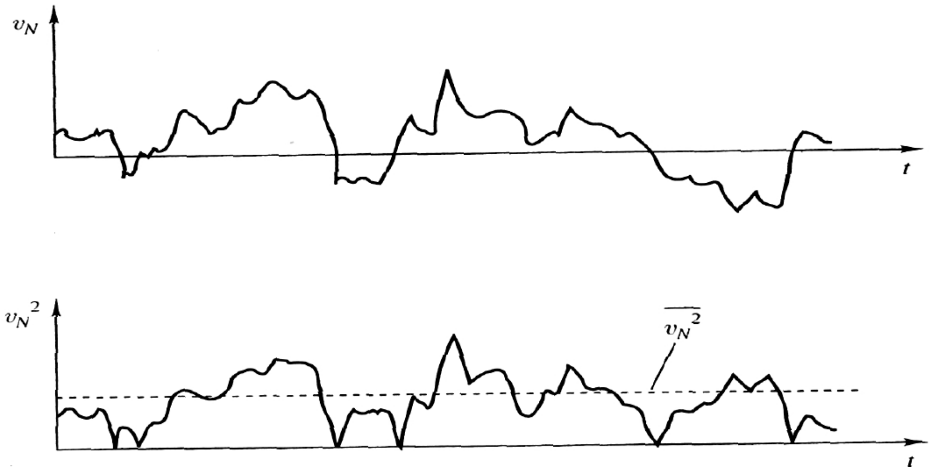
The fluctuations of $V_{un}(t)$ when $V_{in} = 0$ correspond to the noise of amplifier



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

$\Downarrow$

V^2_n

We can define:

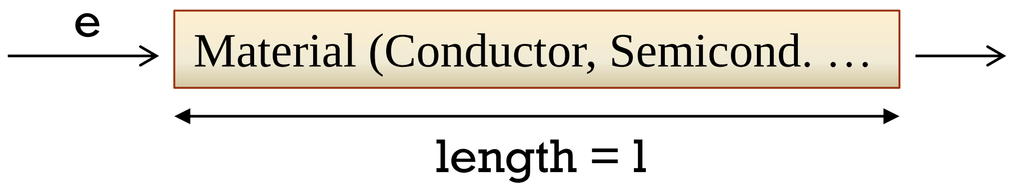
- Source of **voltage noise:** $$v_n = \sqrt{v_n^2(f)}$$
- Source of **current noise:** $$i_n = \sqrt{i_n^2(f)}$$

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: $$<di>^2 = \left(\frac{ne}{l}<dv>\right)^2 + \left(\frac{ev}{l}<dn>\right)^2$$

There are two basic mechanism contributing to noise:

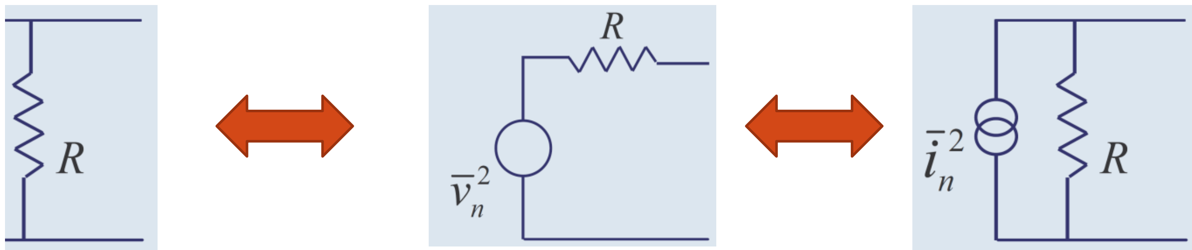
Velocity fluctuations → Thermal noise

Number fluctuations
- Shot noise
- Excess (or flicker, or "1/f") noise

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



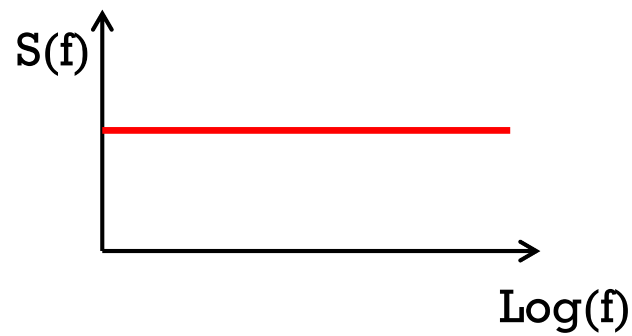
A real (noisy) resistor is equivalent to an ideal (noiseless) resistor + noise source (voltage or current)

Power spectral density:

$$\boxed{S_v(f) = \frac{d\overline{v}_n^2}{df} = 4kTR}$$

$$\boxed{S_i(f) = \frac{d\overline{i}_n^2}{df} = \frac{4kT}{R}}$$

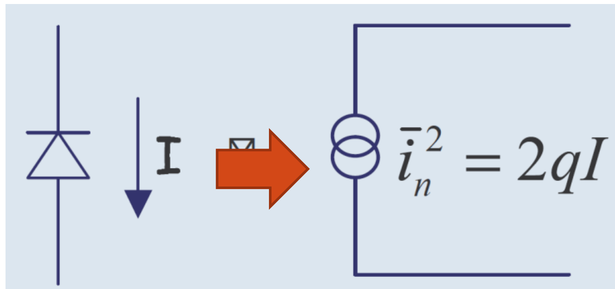
k = Boltzmann constant = $1.3806503 \times 10^{-23}$ J/K
T = absolute temperature
R = resistance



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

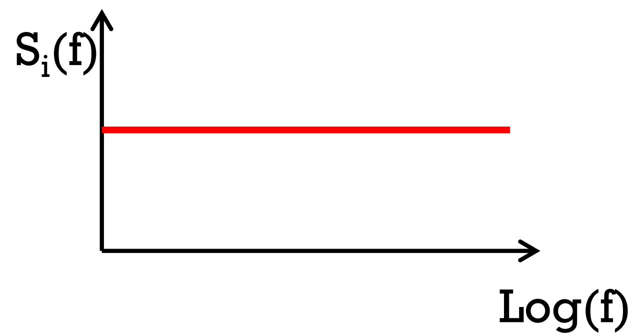


Power spectral density: $$S_i(f) = \frac{d\bar{i}_n^2}{df} = 2qI$$

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

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



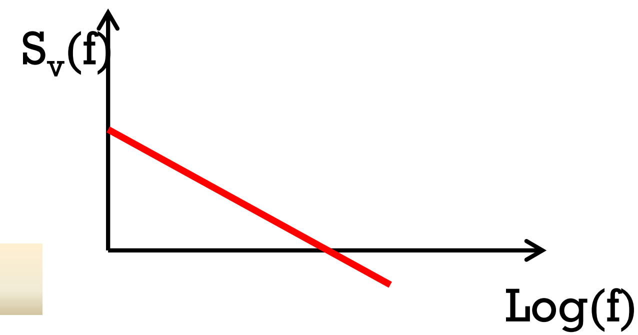
3. Flicker noise (1/f noise)

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



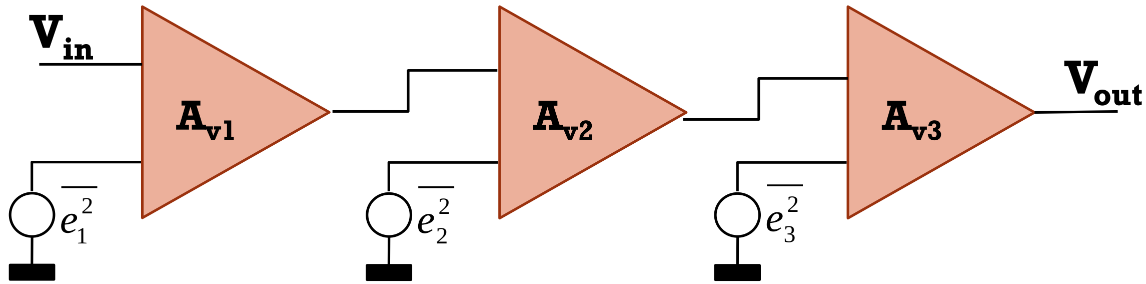
<u>It depends on f</u> and clearly it is important at <u>low frequencies</u>

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

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 <u>first stage</u> dominates

In first 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: <u>in this way the noise can be directly compared with input signal</u>

4. In the case of particle detection systems, where the input is a charge Q, we use ENC: <u>Equivalent Noise Charge</u> : 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

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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 (latch-up)

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

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

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

✓ **Single Event Effects (SEE)** → temporary or permanent errors

CMOS TECHNOLOGIES FOR FRONT-END ELECTRONICS

- **Most used technology for FEE is CMOS**

- Relatively "cheap" if recent/"old" technologies (nodes) are used

- Using the "multiproject foundry runs", prototyping and small productions are very affordable

- Suitable to combine on the same chip analog section, digital part and μprocessors

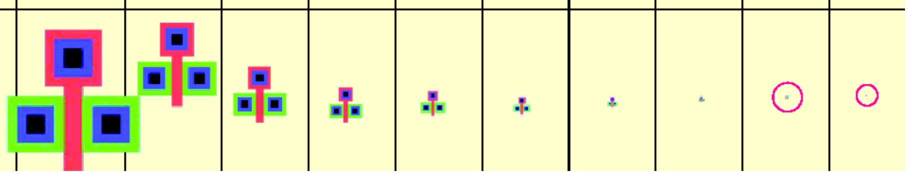
- Very low power consumption

- The deep submicron CMOS tech. (< 130 nm) are rad-tolerant and suitable for HL-LHC, ILC, space applications

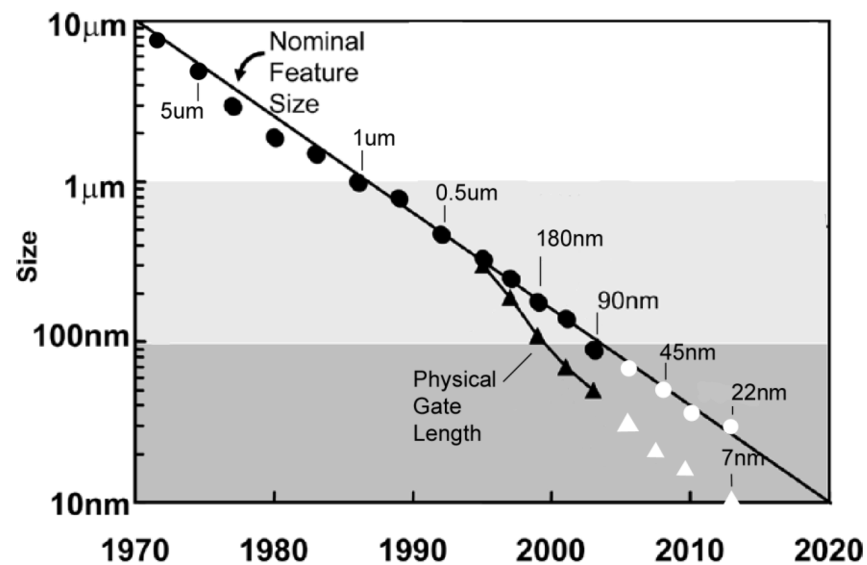
INDUSTRY SCALING ROADMAP FOR CMOS

- Industrial CMOS scaling is entirely driven by commercial digital electronics.
- Front-end electronics may benefit from scaling in terms of functional density (small pitch pixels) and digital performance and density
- Analog design is a challenge (reduced supply voltage and dynamic range …) without density improvement

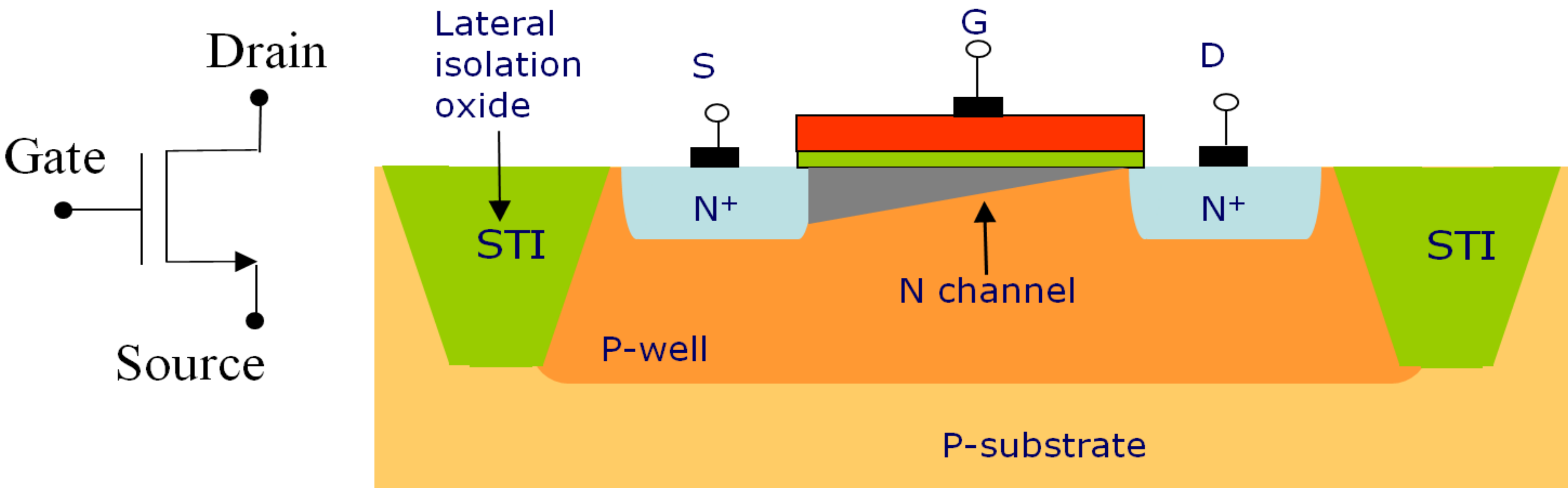
Feature Size [nm]	2000	1200	800	500	350	250	130	65	35	20
Minimum NMOS										



- New generation every ~2 years
- L_g (1970) 8 μm → L_g (2007) 18 nm



BASIC ELEMENT: THE MOSFET

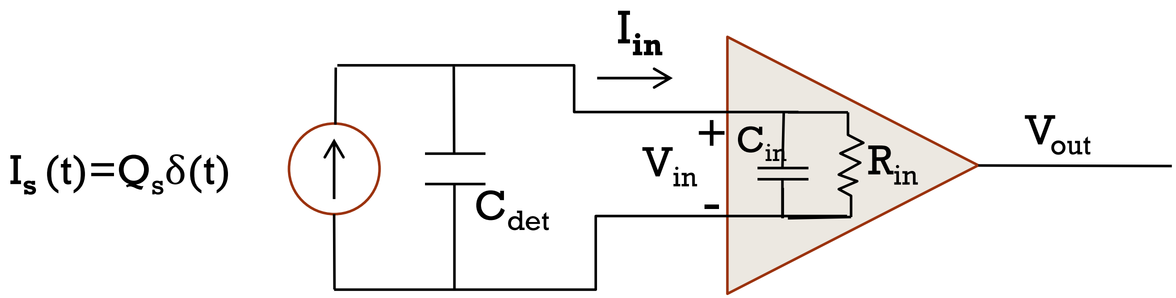


- Three-terminal device: an electrode controls the current flow between two electrodes at the end of a conductive channel

- The transconductance $g_m = dI_D/dV_{GS}$ is the ratio of change in the output (drain) current and of the change in the potential of the control (gate) electrode

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, SiPM and Resistive Plate Chambers to tens of μs, as in inorganic scintillators

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

- We must integrate I to have a measurement of E:

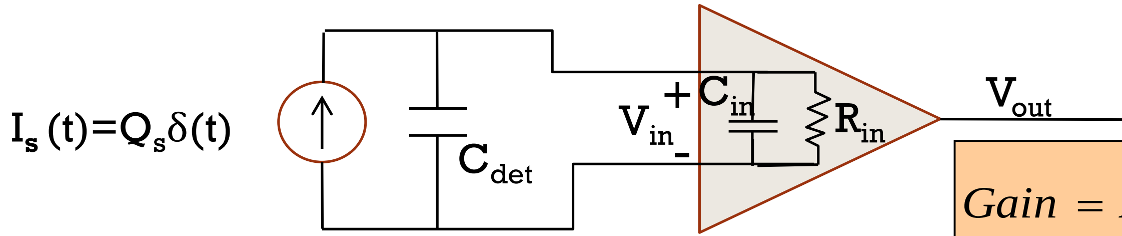
$$E \propto Q_S = \int I_S(t)dt$$

WHERE to integrate?

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

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

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

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

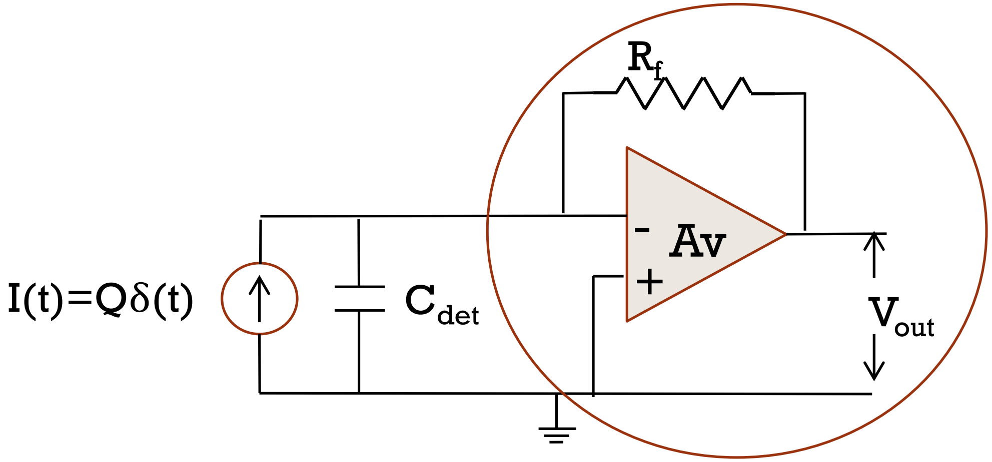
$$V_{out} = A_v \cdot V_{in} = A_v \cdot \frac{Q_s}{C_{det} + C_{in}}$$

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

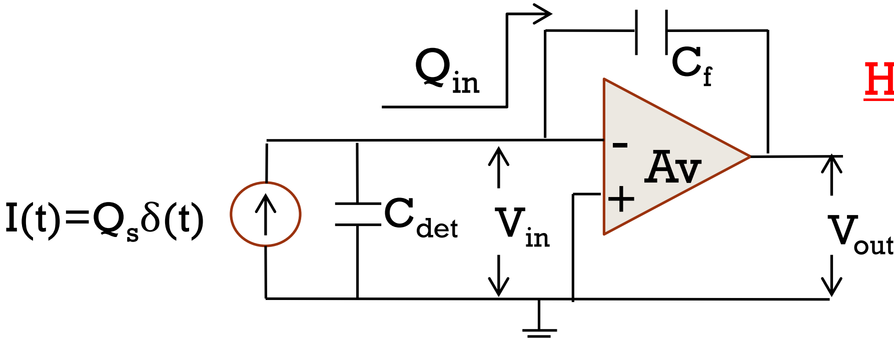
Using a transresistance amplifier (high gain operational amplifier with resistive feedback):

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

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

An integrating stage can follow the amplifier to provide a signal proportional to Q

3. CHARGE-SENSITIVE AMPLIFIER (CSA)



Hypothesis:

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

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

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

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

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

GAIN (Charge Sensitivity):

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

$(A_v >> 1)$

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

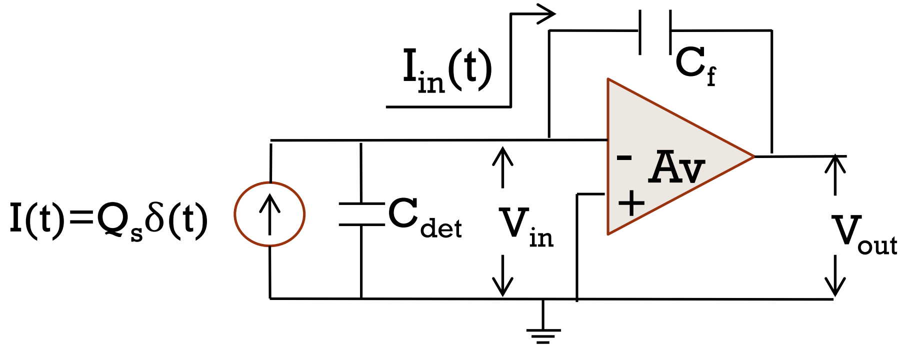
Charge transfer efficiency:

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

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

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

CHARGE-SENSITIVE AMPLIFIER: THE TIME RESPONSE



In the frequency domain:

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

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

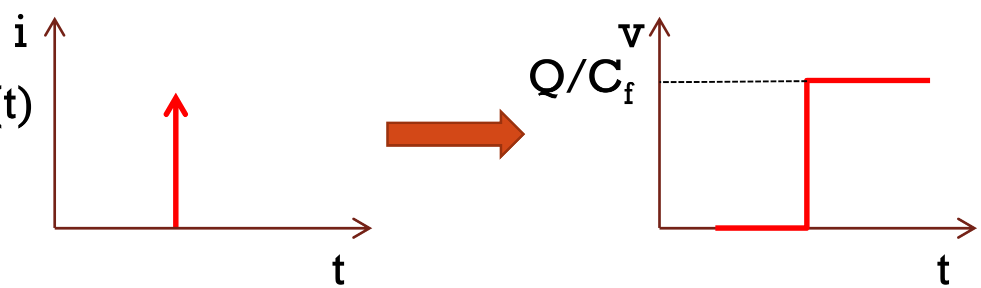
→

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

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

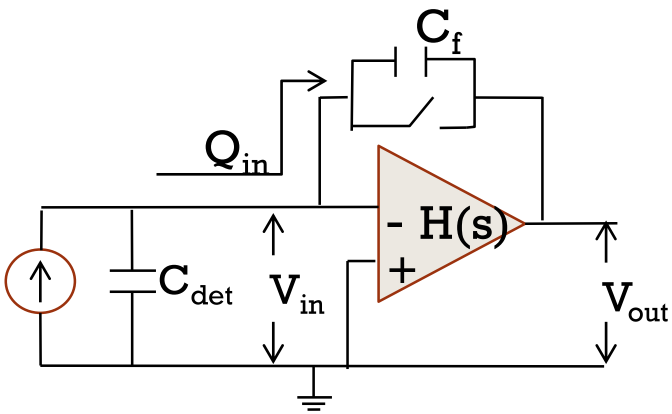
Step function

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



CHARGE-SENSITIVE AMPLIFIER: THE RESET

Pulsed RESET

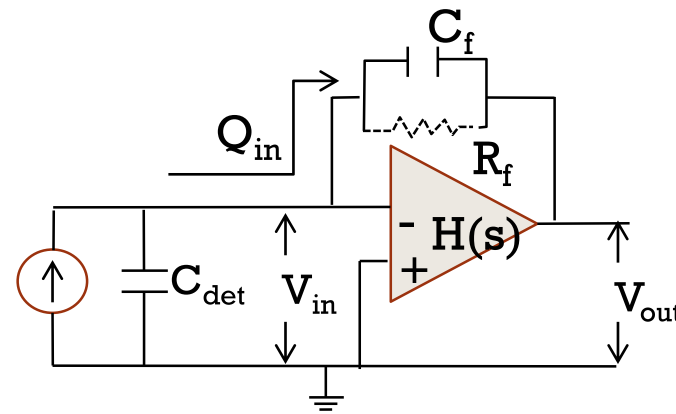


- The reset switch allows the removal of charge stored in C_f
- The switch can be closed periodically or driven by some control signal

Drawbacks:

- Dead time
- Switch noise
- Leakage current

Continuous RESET

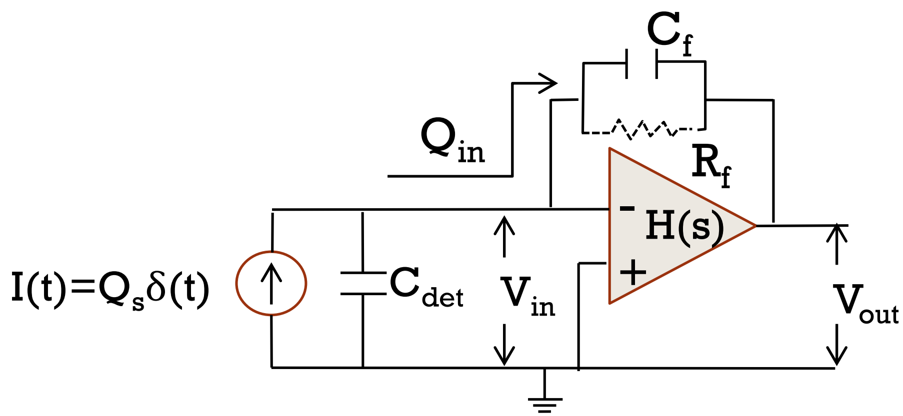


- The resistor R_f continuously discharges C_f after the pulse
- Discharge time constant R_fC_f

Drawbacks:

- Additional parallel noise
- Long tail → Risk of pile-up

CHARGE-SENSITIVE AMPLIFIER: THE REALISTIC CASE



Two elements to be considered:

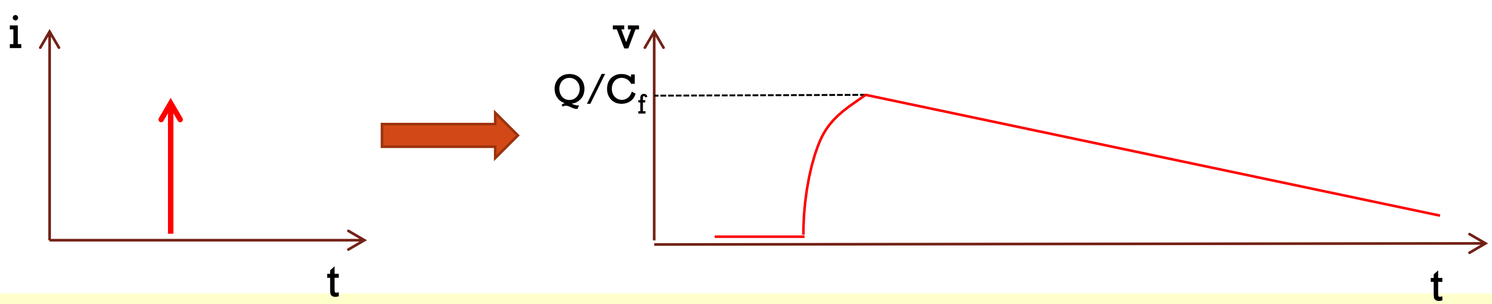
1. Resistor R_f used to discharge C_f. Since it is a source of parallel noise (inject noise current into input noise), it must made very large to decrease its contribution to noise. Typical values are several tens or hundreds of MΩ
2. Real amplifier (finite bandwidth and gain)

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

2 poles →

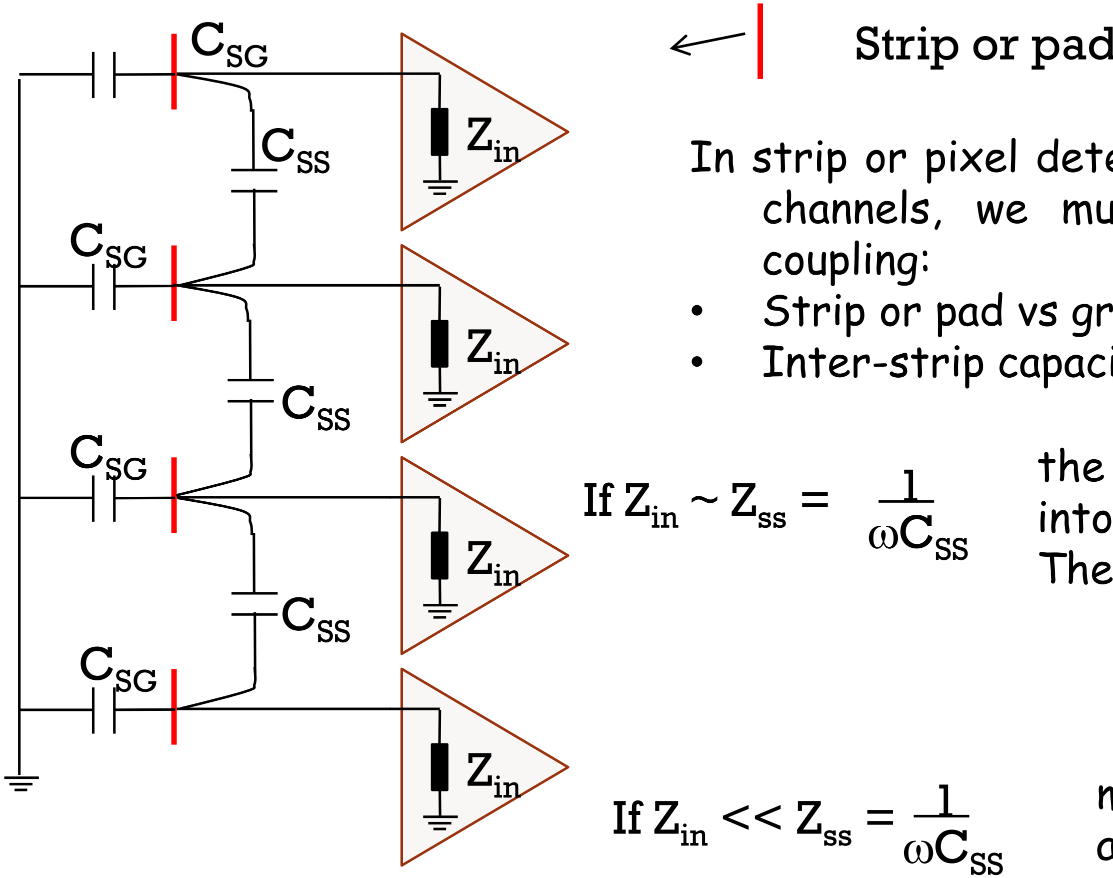
$$\begin{cases} p_1 = \dfrac{1}{R_f C_f} & \text{(Low freq)} \\ p_2 = \dfrac{1}{R_i C_T} = \dfrac{\omega_0 C_f}{C_T} & \text{(High freq)} \end{cases} \longrightarrow \begin{cases} \tau_1 = R_f C_f & \text{Fall time constant} \\ \tau_2 = R_i C_T = \dfrac{C_T}{\omega_0 C_f} & \text{Rise time constant}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

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



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

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 $Z_{in} \sim Z_{ss} = \dfrac{1}{\omega C_{SS}}$ — the charge induced on one strip is coupled into the adjacent channels through C_{SS}. The number of affected strips depends on $\dfrac{C_{SS}}{C_{SG}}$

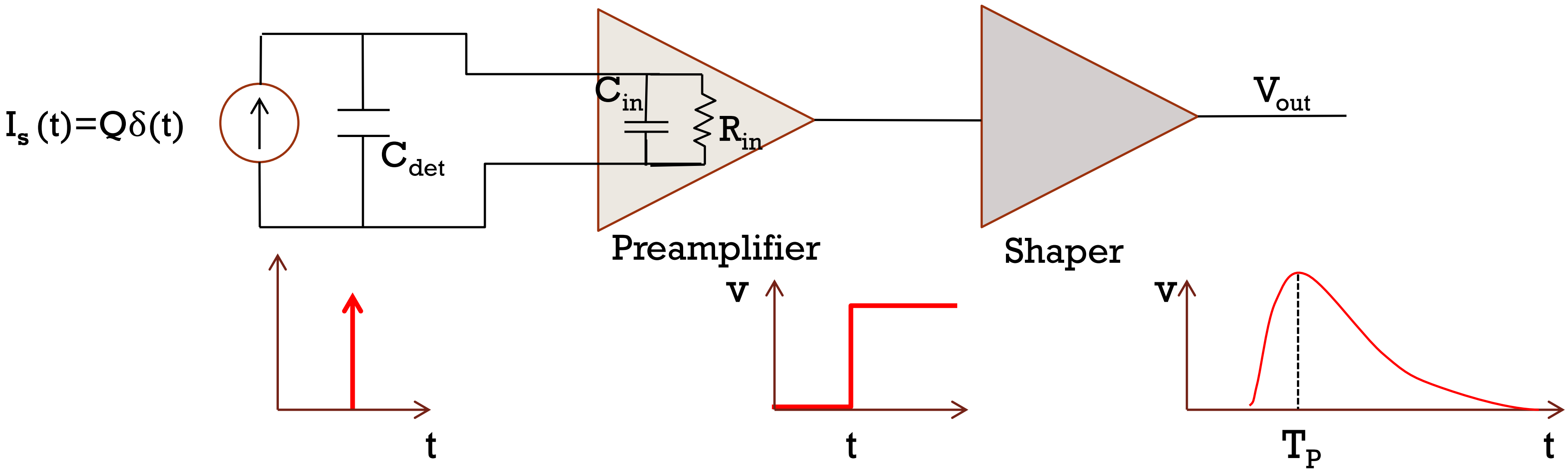
If $Z_{in} << Z_{ss} = \dfrac{1}{\omega C_{SS}}$ — most part of the charge flows into the amplifier and only small part is coupled into the adjacent channels through C_{SS}

Summary:
low input impedance →

- Short rise time
- Small cross-talk

NOISE FILTERING: SHAPERS

PULSE SHAPING



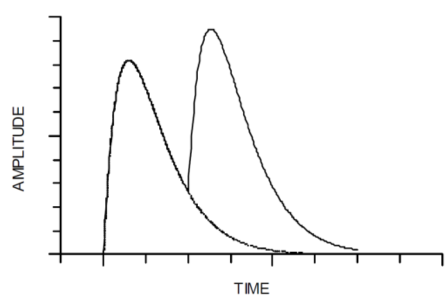
Preamplifier = input amplifier It is usually located close to detector and must have enough gain to make negligible the effects of induced noise. Typical example: Charge Sensitive Amplifier

Shaper = a **filter** with two main purposes:

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

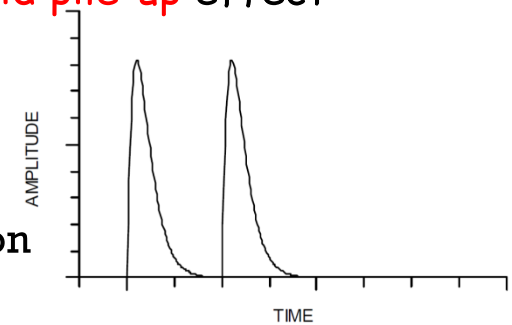
Slower pulse:

- 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) ——

$\omega = 2\pi f$

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

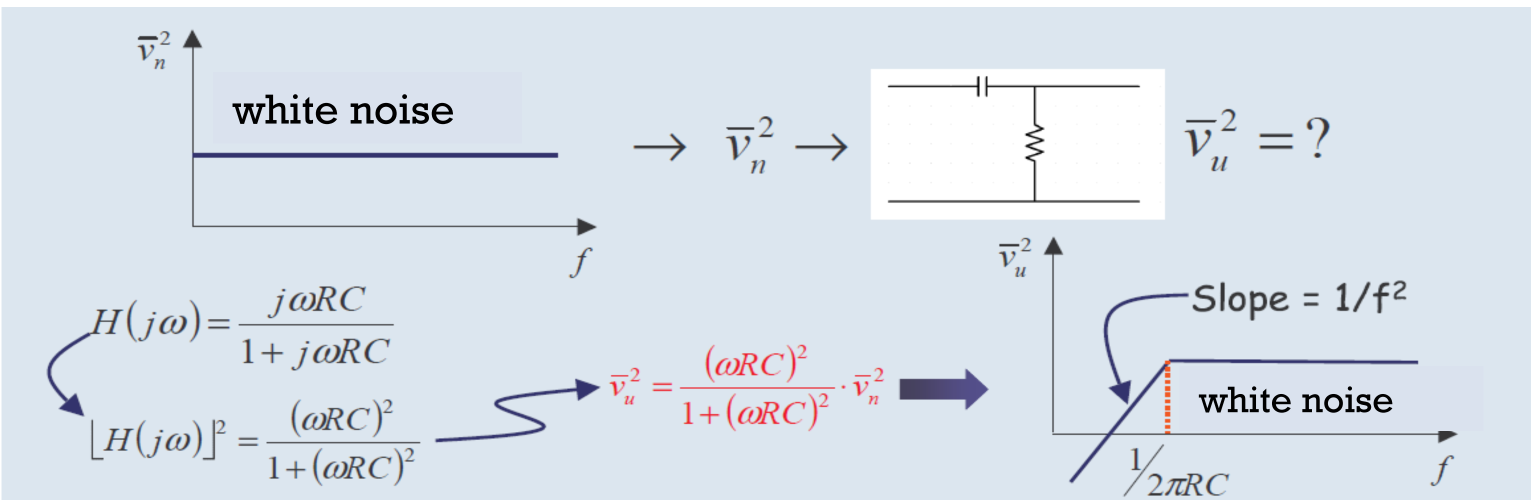
Noise power spectrum at output of a filter with transfer function H(jω) is equal to input power spectrum multiplied by squared transfer function

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

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

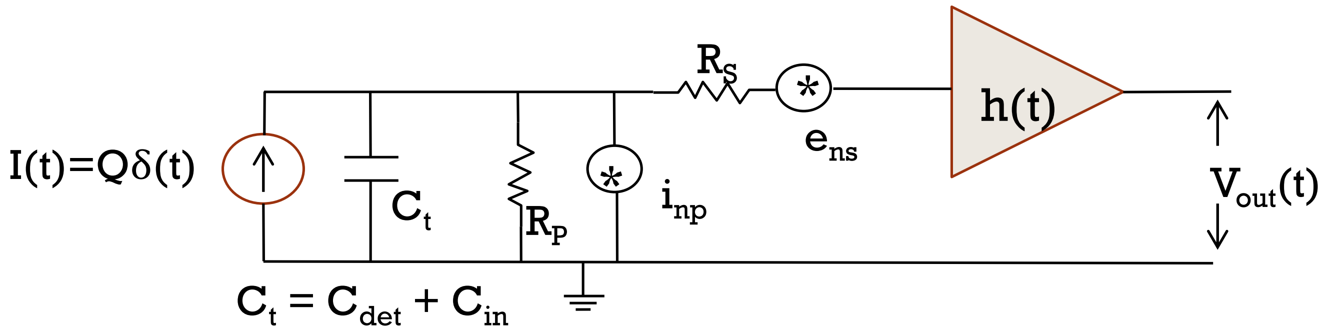
- The total noise increases with bandwidth
- Small bandwidth → large rise-times → less noise
- High bandwidth → fast pulse → more noise

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



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 ~ ENC_0 + K•C_{det}**

CAPACITIVE MATCHING

- Parallel noise depends mainly by "external" factors (Feedback resistor, detector bias and leakage)
- Series noise depends on amplifier characteristics (Rs → g_m, C_{in})

with proper design and dimensioning of preamp we can optimize ENC_S

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

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

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

Input transistor capacitance must be matched to detector capacitance

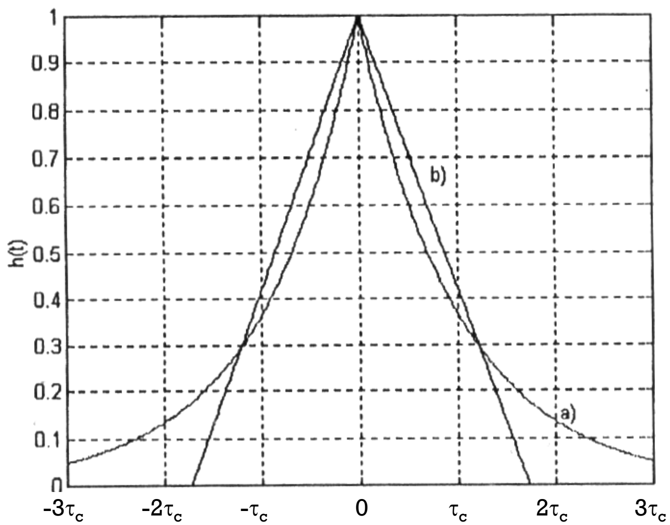
$$\boxed{ENC^2_{s\_Opt} = 16KTa_n C_{det} \frac{\tau_A}{t_m}}$$

OPTIMUM FILTER

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

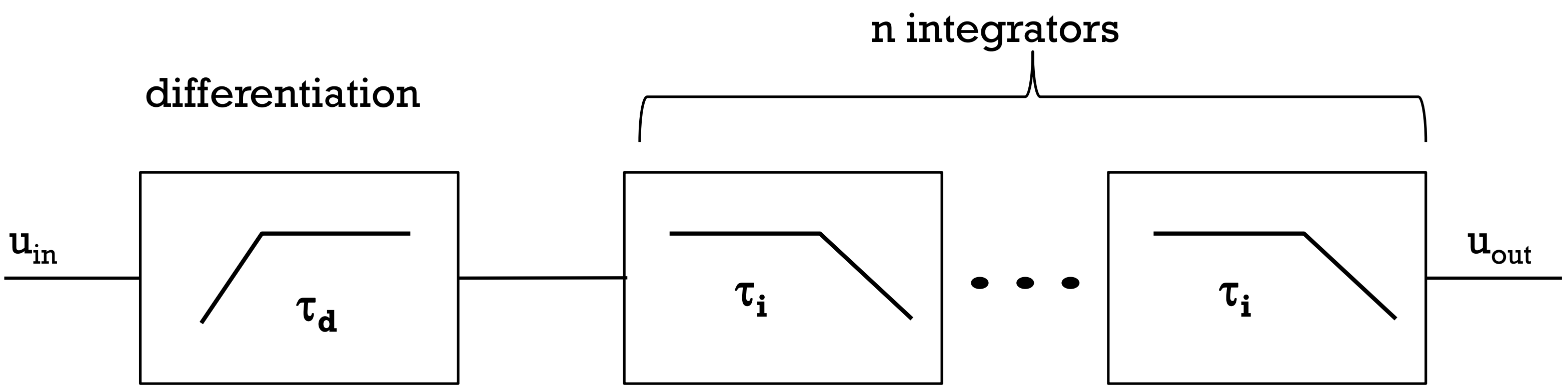
It is possible to demonstrate that $h_{opt}(t) = \exp\left(-\frac{|t|}{\tau_c}\right) \longrightarrow ENC_{opt}^2 = 2KTR_S \frac{C_t^2}{\tau_c} = 2KTC_t\sqrt{\frac{R_s}{R_p}}$

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



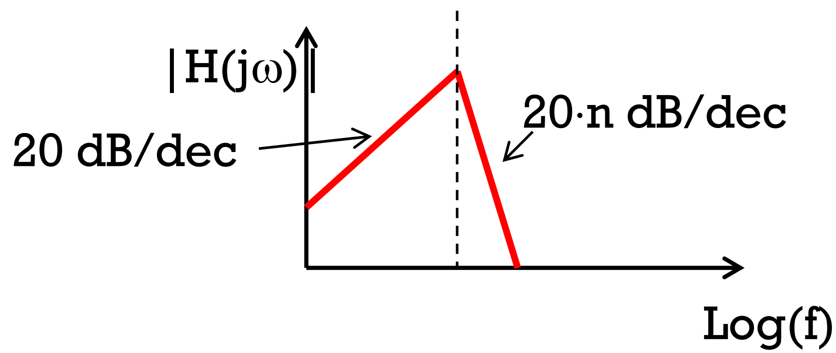
The cusp filter is not practically feasible, but can be approximated by triangular shapers (curve b) or Pseudo-Gaussian shaper

PSEUDO-GAUSSIAN (OR SEMI-GAUSSIAN) SHAPER



1. A high-pass filter, that makes the derivative of the input pulse and introduces the decay time 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)

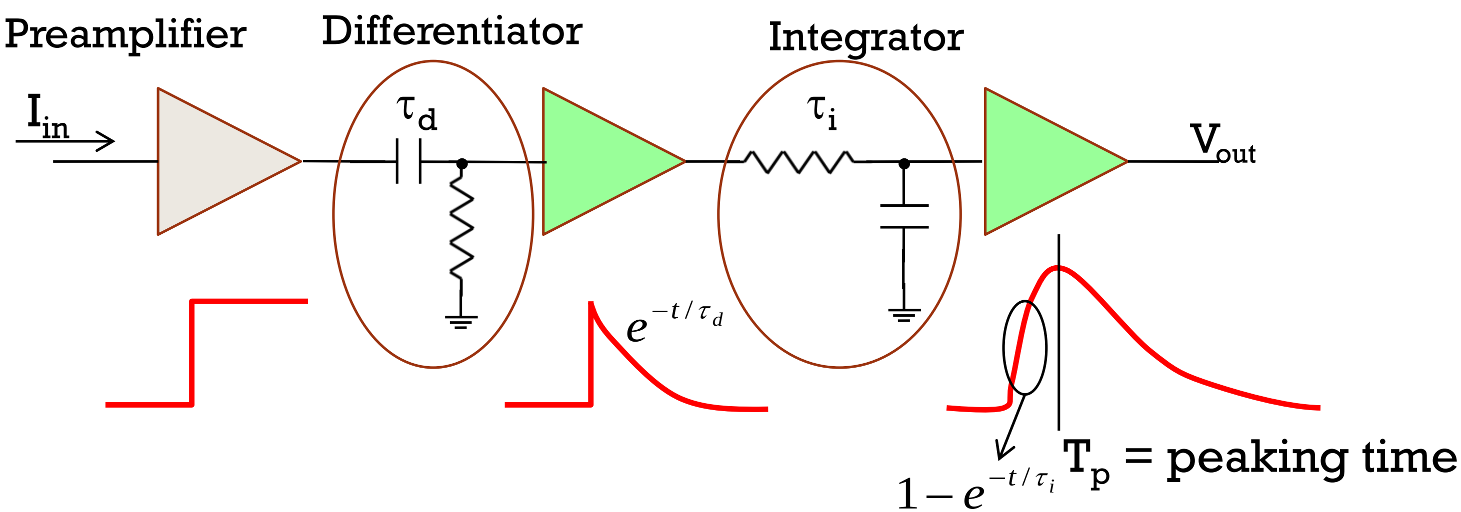
$$H(s) = \frac{u_{out}(s)}{u_{in}(s)} = \frac{s\tau_d}{(1+s\tau_d)} \frac{1}{(1+s\tau_i)^n}$$



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



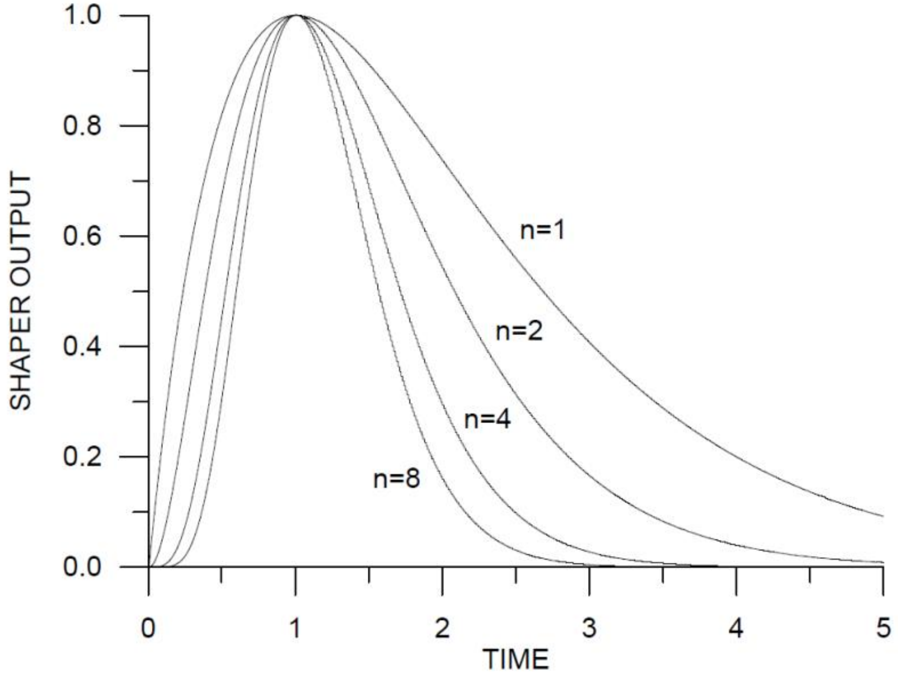
- This shaper is called CR-RC because the high-pass filter is made with CR network, while the low-pass filter with a RC network

- 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

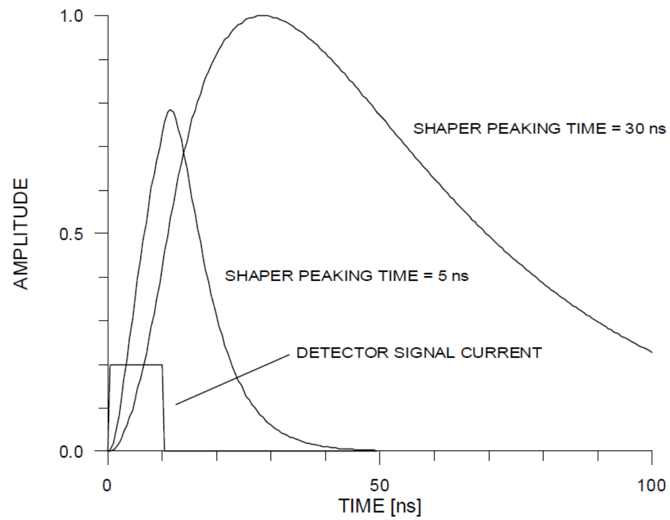
- Same peaking time if $\tau_n = \tau_{(n=1)}/n$
- With same peaking time
  1. More symmetrical
  2. Faster return to baseline
  3. Improved rate capability



2nd order shapers are commonly used

BALLISTIC DEFICIT

Ballistic Deficit is a <u>Loss in Pulse Height</u> if the peaking time T_p of the shaper is shorter than the detector collection time or, more in general, shorter than 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

<u>Consequences:</u>

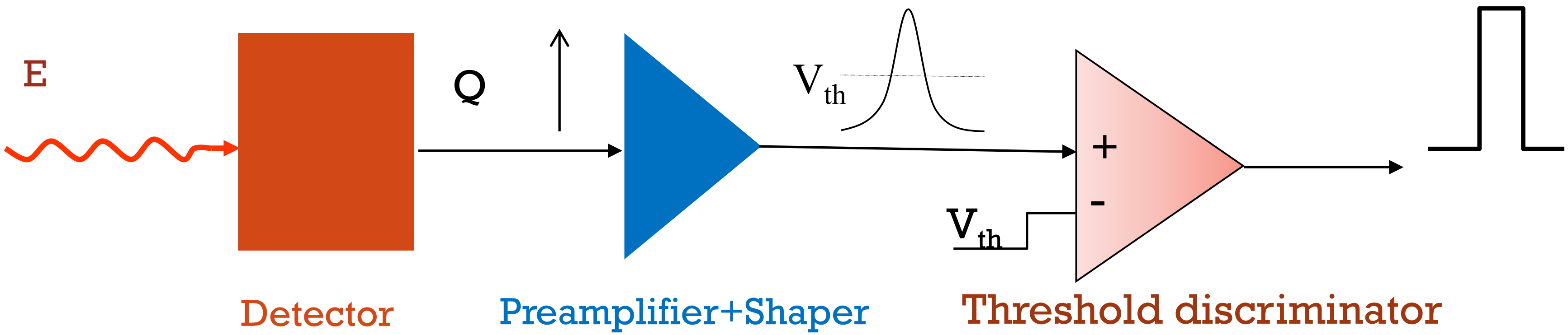
- Loss of useful signal
- Increase of ENC (or decrease of S/N)

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

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

HIT DISCRIMINATION AND TIME MEASUREMENT

HIT DISCRIMINATION



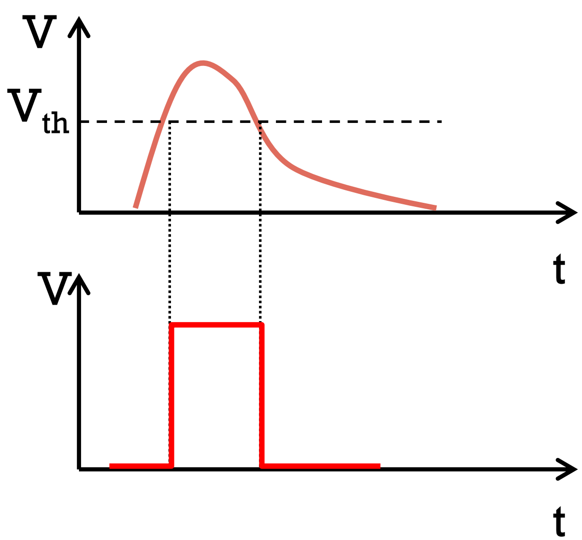
- Binary readout: hit/no-hit information from a discriminator

- In a multichannel readout chip, channel-to-channel threshold variations due to device mismatch may degrade detection efficiency and spurious hit rate

TIME MEASUREMENT

- The hit discrimination technique contributes to the **timing resolution** and **timing accuracy**
- Sometimes the purpose of the system is precise **time measurements** (using Time to Digital Converters)

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



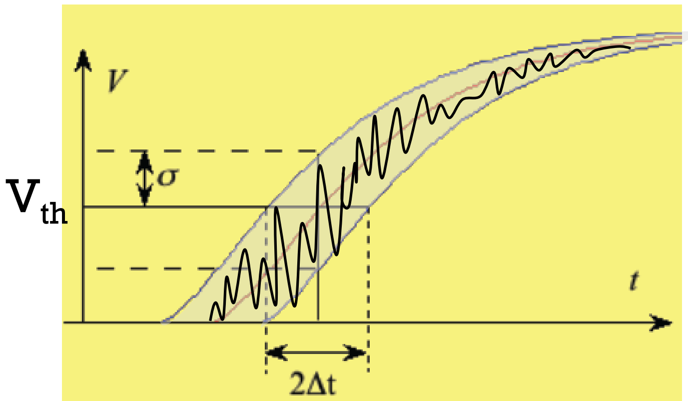
Timing measurement is limited by:

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

THE JITTER

Noise has an impact in time measurements:

uncertainty in the time of crossing threshold → Jitter



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

slope

How to decrease jitter? → Conflicting conditions:

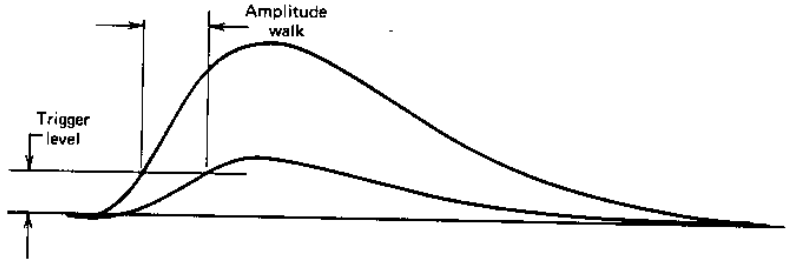
- decrease σ_{noise} → decrease bandwidth
- increase slope → increase bandwidth

As usual … find compromise

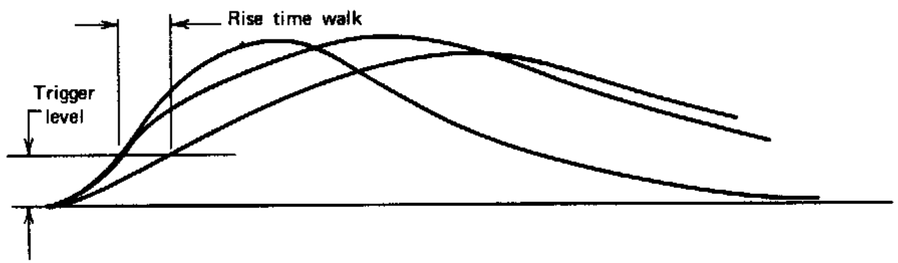
To minimize jitter, the best threshold level is where the slope is maximum

THE TIME WALK

In the leading edge discriminators, two pulses with identical shape and time of occurrence, but different amplitude cross the same threshold in different times (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 discriminator to time walk is minimized by setting the threshold as low as possible but it must be compatible with noise level

Time walk correction:

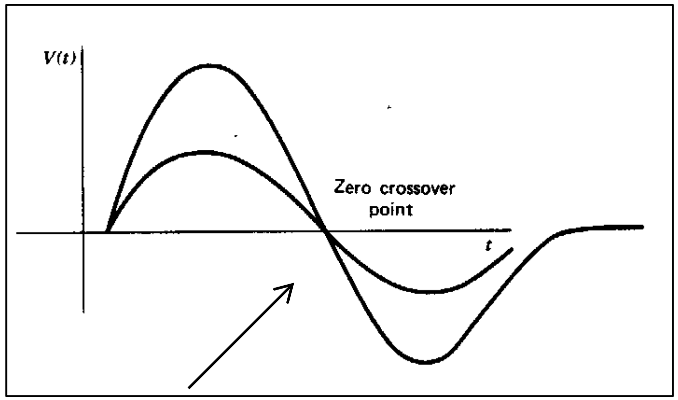
- Software: measure the pulse amplitude and apply correction to timing
- Hardware: instead of leading edge discriminator, use
  1. Crossover timing
  2. Constant Fraction timing

CROSSOVER TIMING

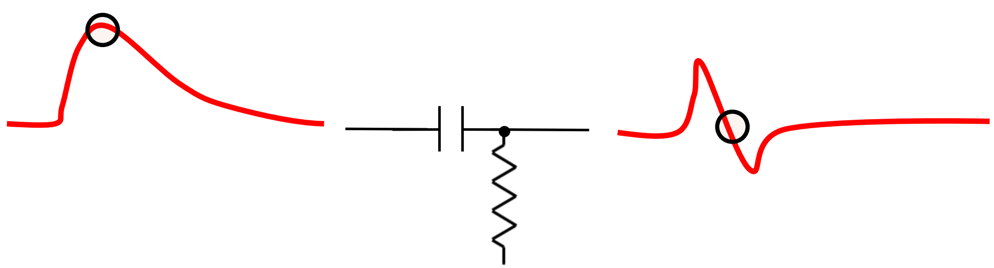
The crossover timing can greatly reduce the magnitude of the amplitude time walk

Hypothesis:

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

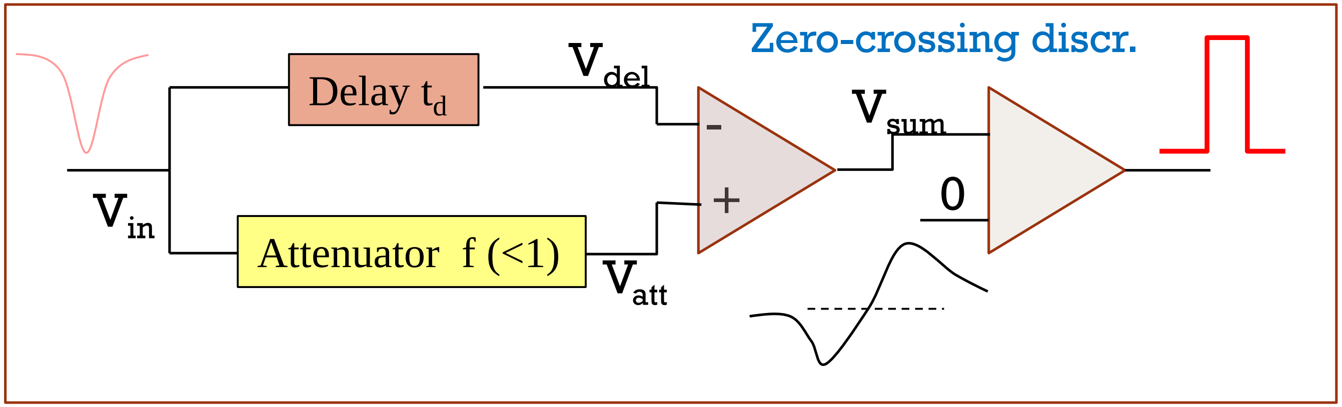


If the output of shaper is unipolar, but the peaking time is constant, adding a differentiator (C-R network) we get a bipolar pulse crossing the zero in correspondence of the signal peak



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

THE CONSTANT FRACTION TIMING



Summing:

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

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

SOME EXAMPLES OF FRONT-END ELECTRONICS FOR GAS DETECTORS

1. GASTONE64 : Front-end chip of Cylindrical GEM (CGEM) for the KLOE-2 Experiment, at Frascati INFN LAB

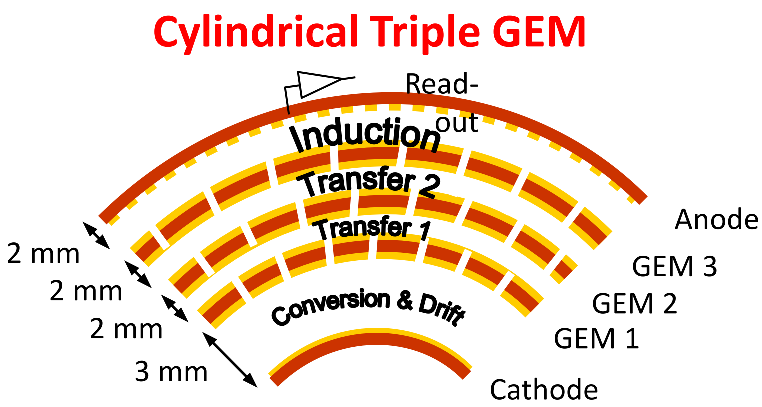
2. VFAT3: Front-end chip of GE1/1 detector for the CMS Experiment, at CERN

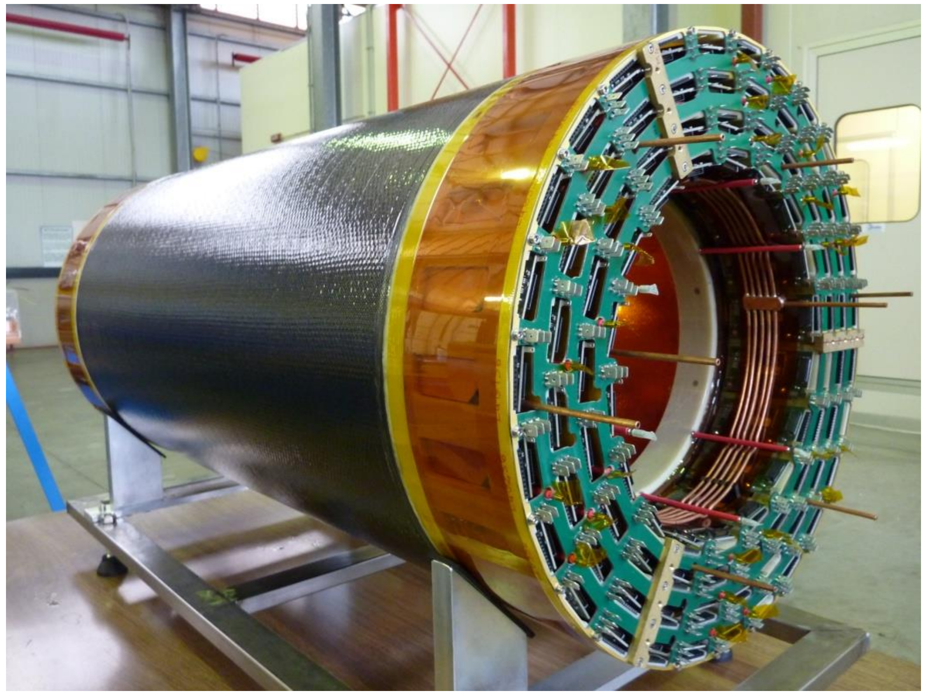
KLOE-2 INNER TRACKER FRONT-END

KLOE-2 is an experiment at DAFNE accelerator, in Frascati INFN National Laboratories

An Inner Tracker GEM-based was inserted around the interaction point to improve the vertex resolution

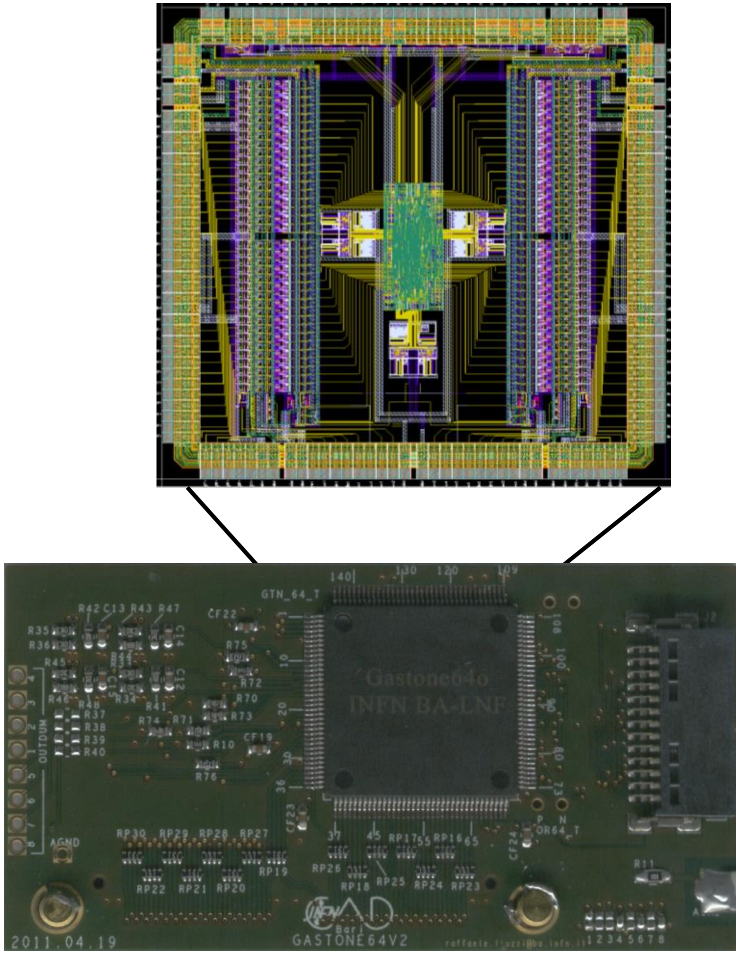
Realized with **Cylindrical_TRIPLE_GEM** detectors





KLOE-2 IT FRONT-END: GASTONE64

Developed by:
INFN-BARI
INFN-LNF

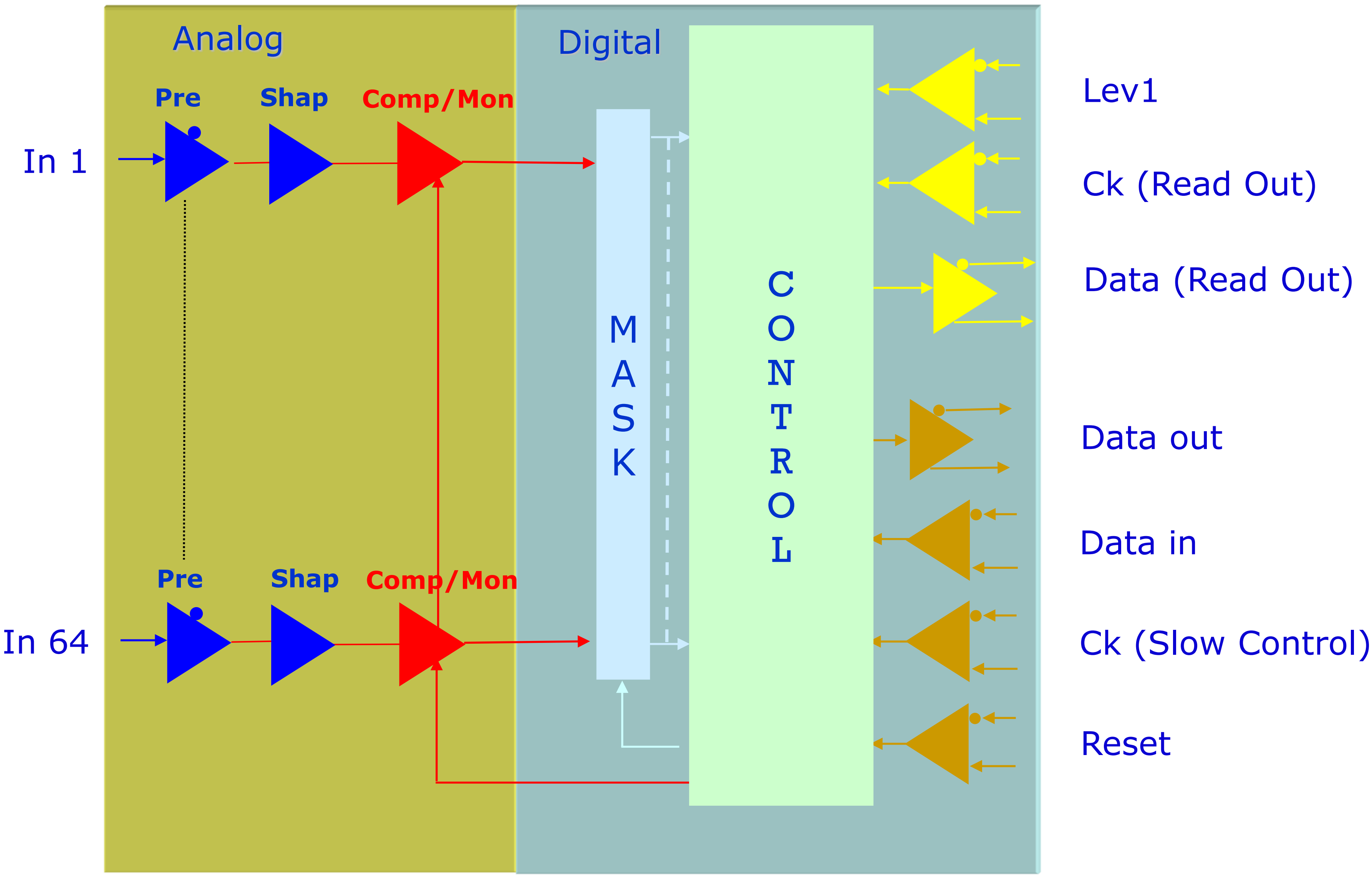


N. channels	64
Technology	CMOS 0.35 μm
Chip dimensions	4.5 X 4.5 mm^2
Input impedance	120 Ω
Charge sensitivity	16 mV/fC (Cdet = 100 pF)
Peaking time	~90 ns (Cdet=100 pF)
Crosstalk	< 3%
ENC	800 e- + 40 e-/pF
Power consumption	~ 6 mW/ch
Readout	Serial LVDS (100 MBps)

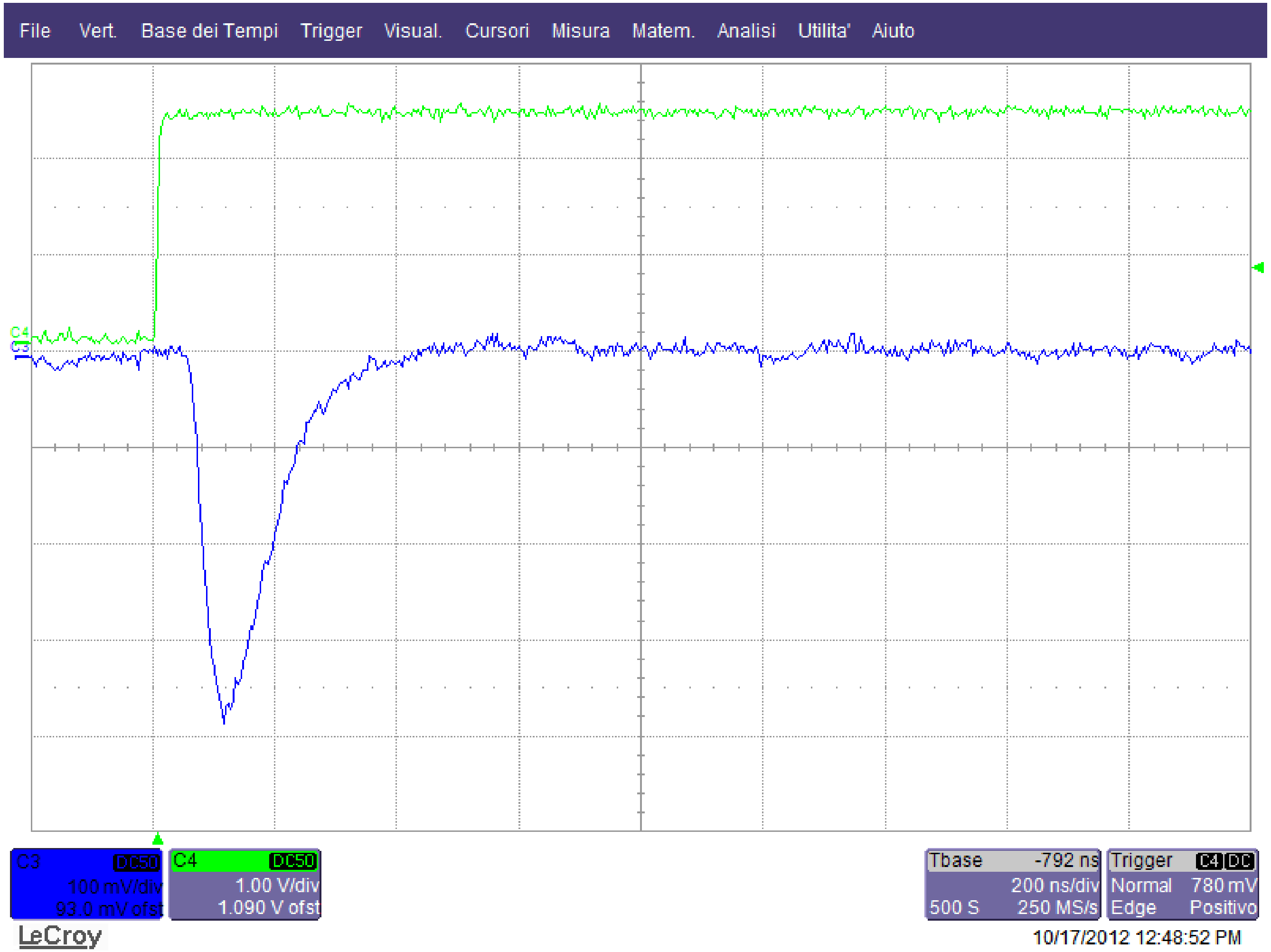
A. Balla et al., *A new cylindrical GEM inner tracker for the upgrade of the KLOE experiment*, Nucl. Phys. Proc. Suppl. 215:76-78,2011

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

GASTONE64: BLOCK DIAGRAM

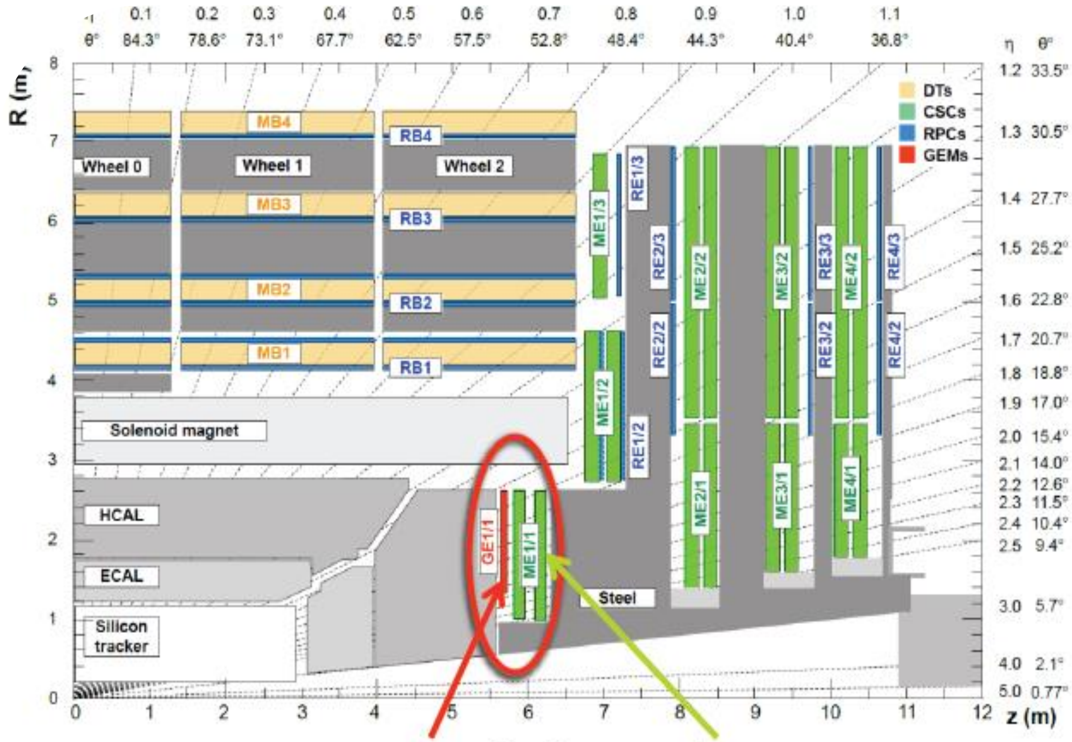


GASTONE64: SHAPER OUTPUT

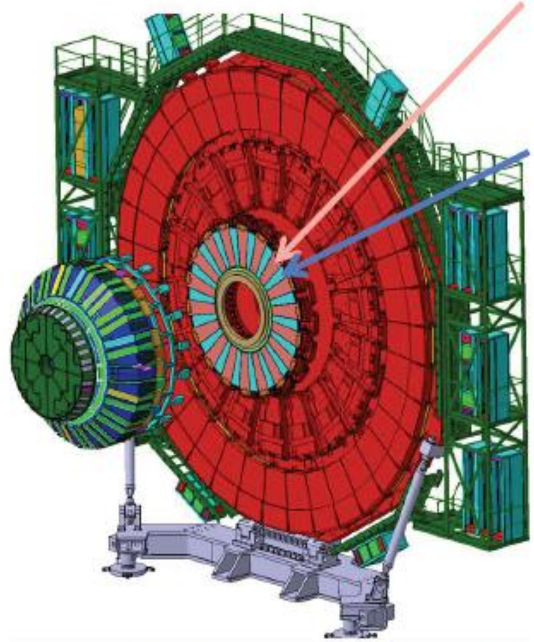


Shaper response for 20 fC input pulse

VFAT3: FE CHIP FOR CMS GE1/1 GEM DETECTOR

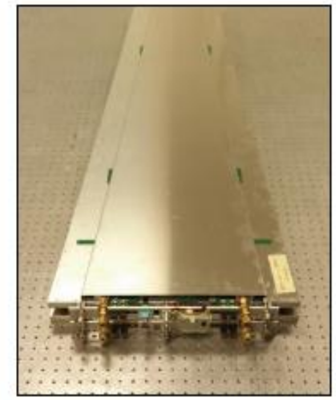


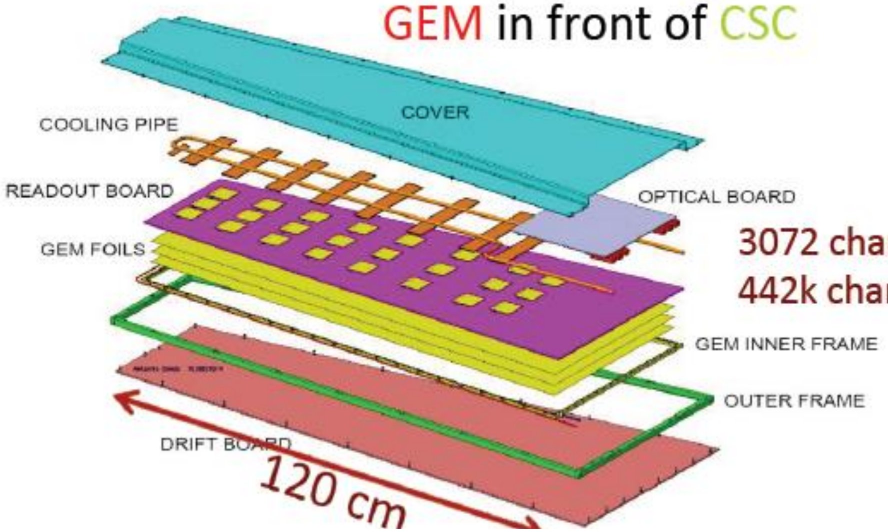
GEM in front of CSC



36 Long Super-chambers
1.55 < |η| < 2.18

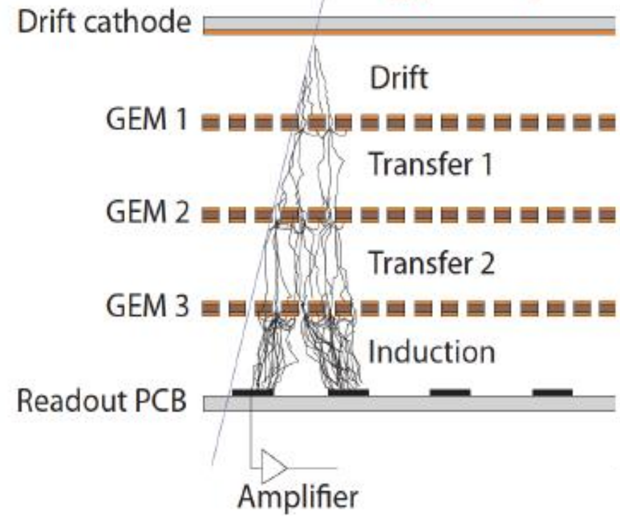
36 Short Super-chambers
1.61 < |η| < 2.18





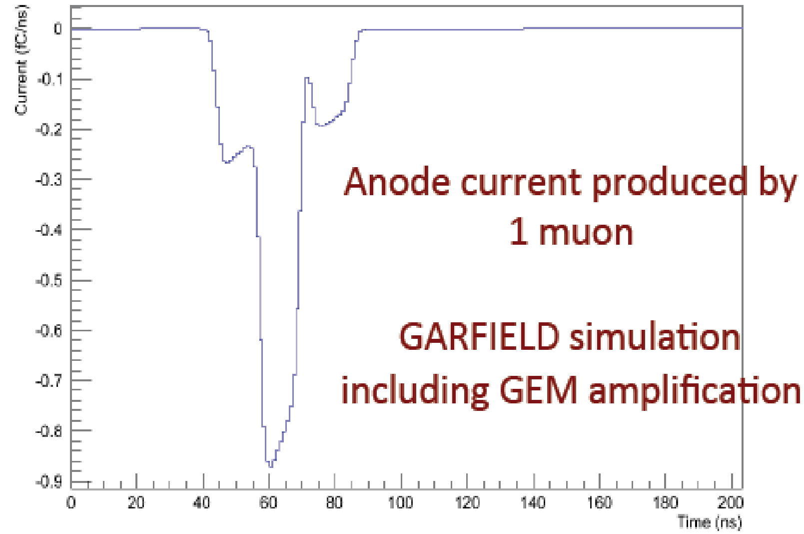
3072 channels/detector
442k channels in GE1/1
10° in ϕ

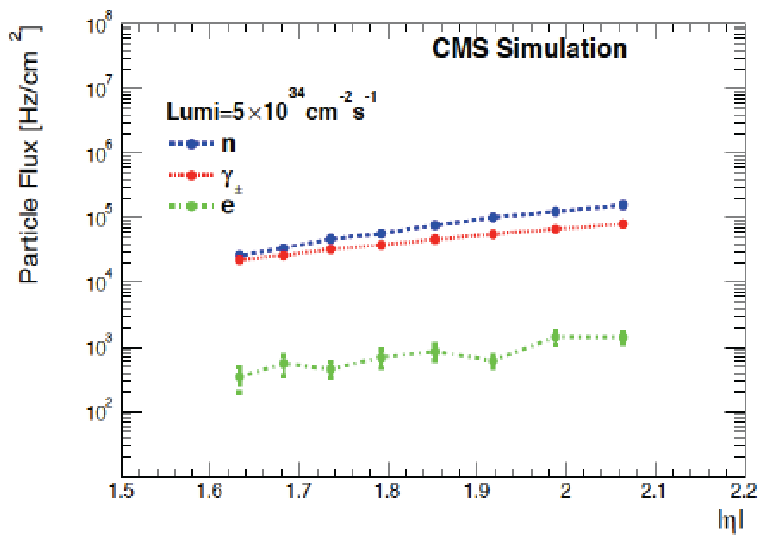
Detector technology: Triple-GEM



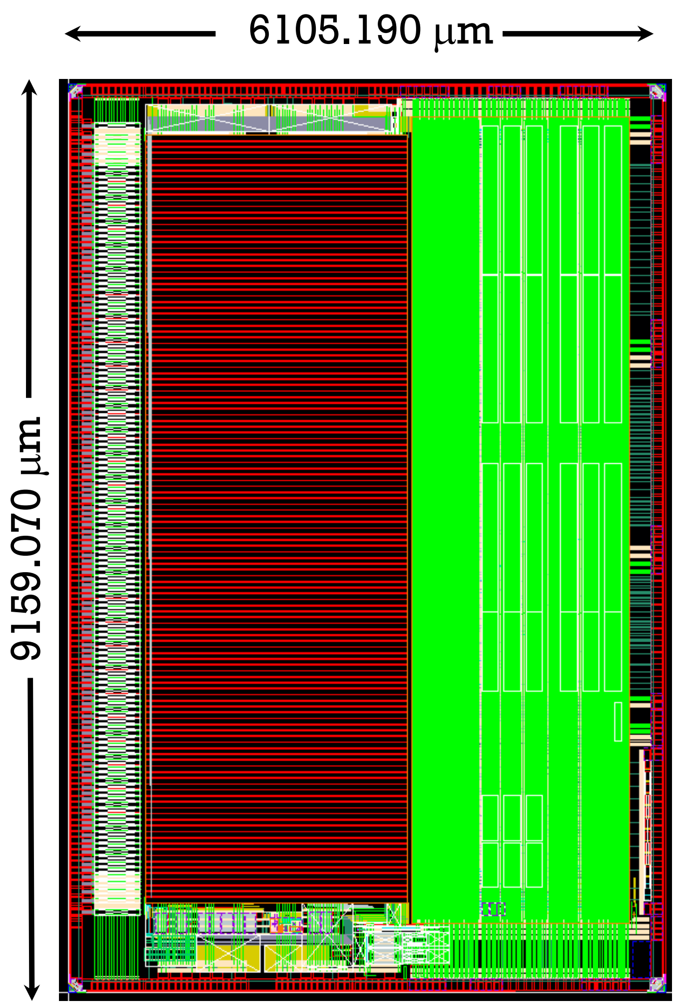
VFAT3: FE CHIP FOR CMS GE1/1 GEM DETECTOR

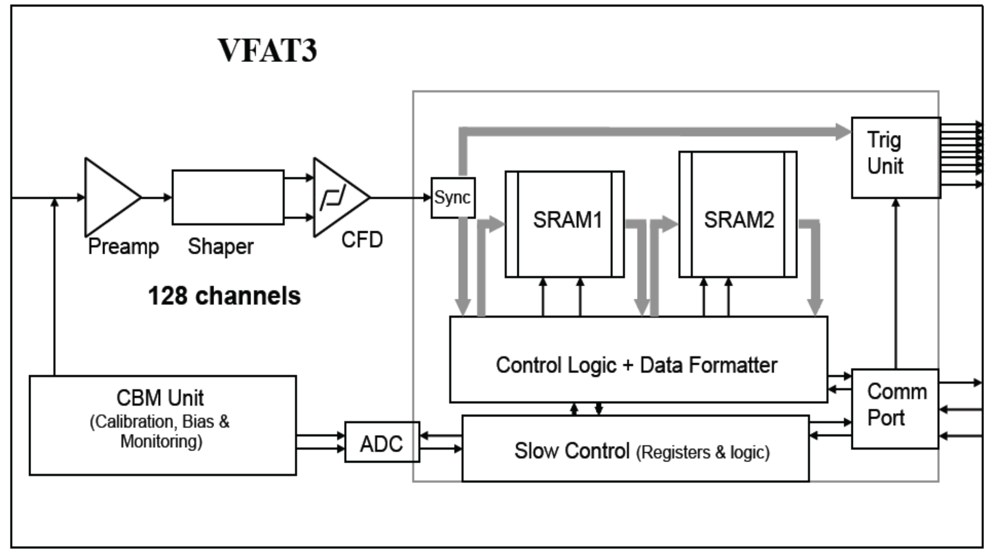
- CMS Triple-GEM
  - signal length: ~60 ns
  - detector capacitance: 10-30 pF
  - charge range (MIP): 4-110 fC
- Expected particle rate
  - up to 2 x 10^5 Hz/cm^2
  - mainly neutron background
- CMS Level-1 latency: 12.5 µs
- CMS Level-1 Accept rate: 750 kHz
- Total irradiation dose: up to 10 krad





VFAT3: FE CHIP FOR CMS GE1/1 GEM DETECTOR

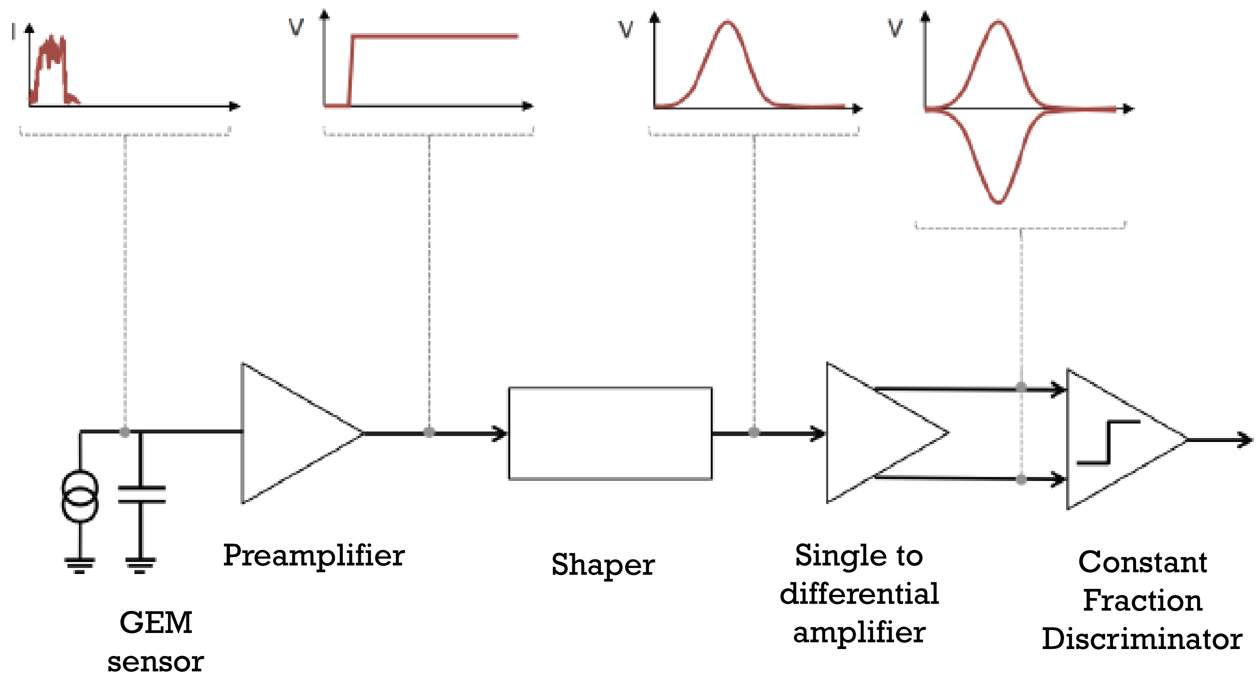




- **Tech.: CMOS 130 nm**

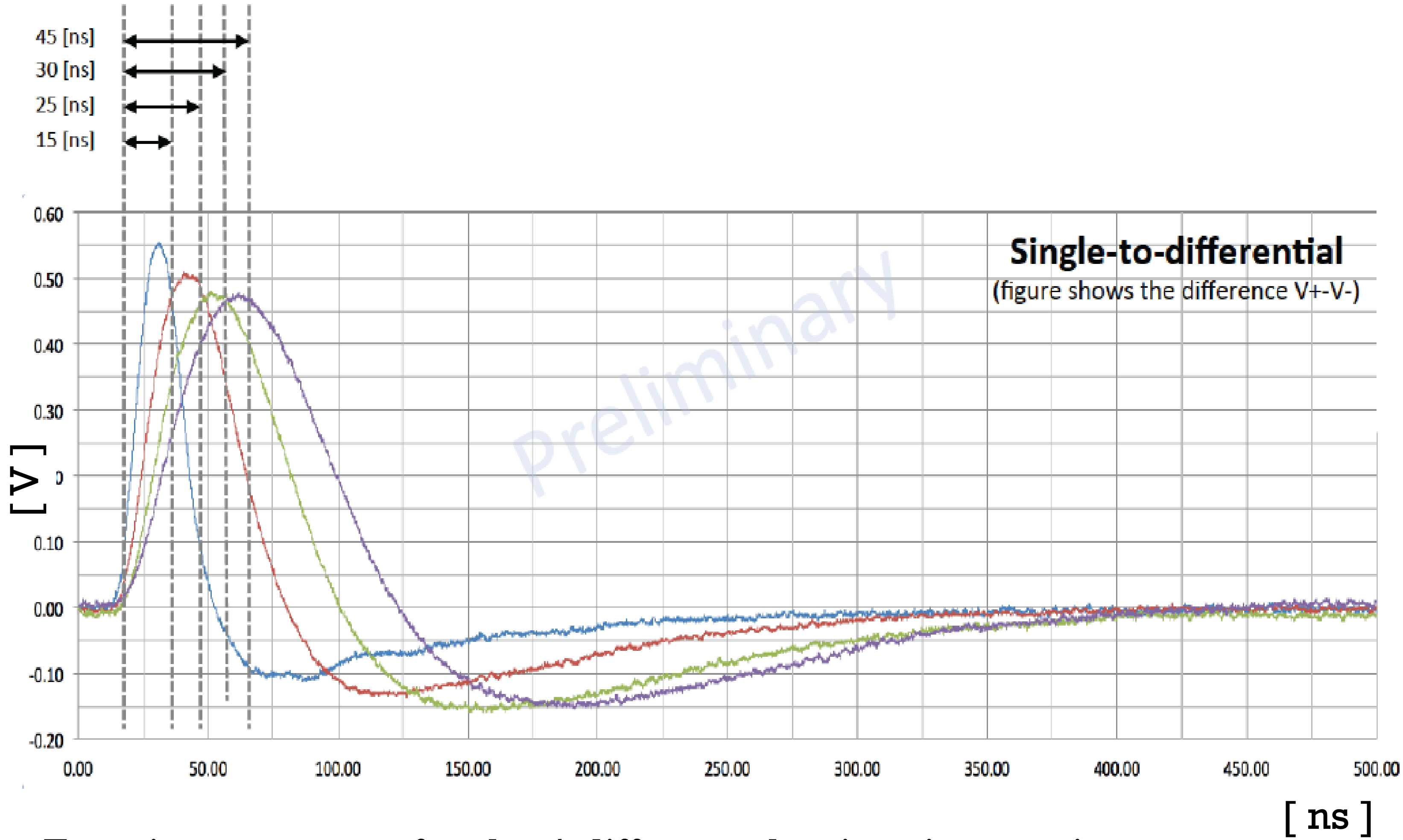
Developed by:
CERN
INFN-BARI
Lappenranta University of Technology
Universite' Libre De Bruxelles

VFAT3: FRONT-END CHANNEL



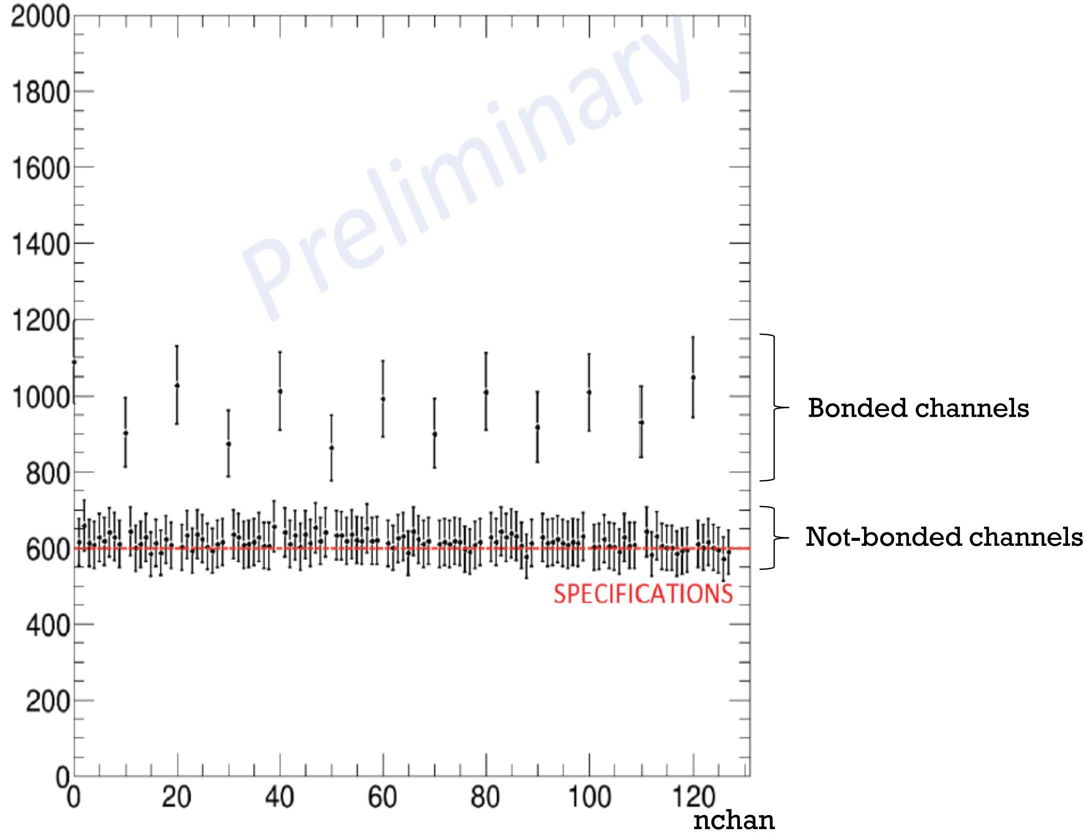
Detector capacitance	[pF]	10-90
Polarity		-/+
Shaping time	[ns]	15,25,35,50
ENC	[e^-]	< 2000 (@20 pF, 50 ns)
Dynamic range	[fC]	10,30,60
Power	[mW]	< 2.2
Crosstalk	[%]	< 2

VFAT3: MEASUREMENTS ON FE

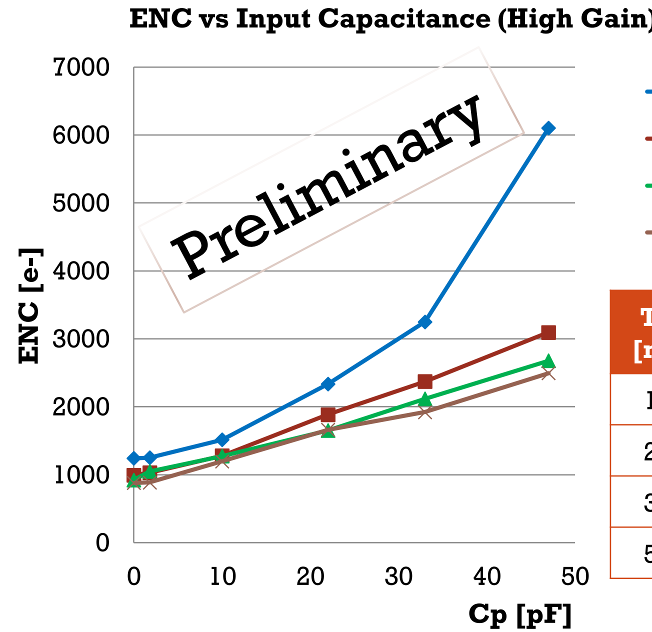


Transient response for the 4 different shaping time settings
Measured on the test channel with an oscilloscope

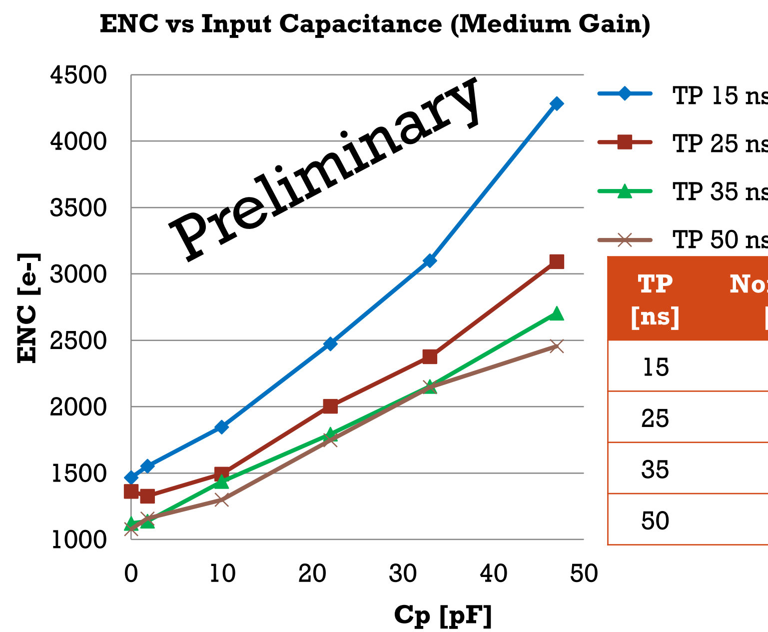
VFAT3: MEASURED ENC



Measurements done through internal calibration circuit
Noise extracted from S-curves

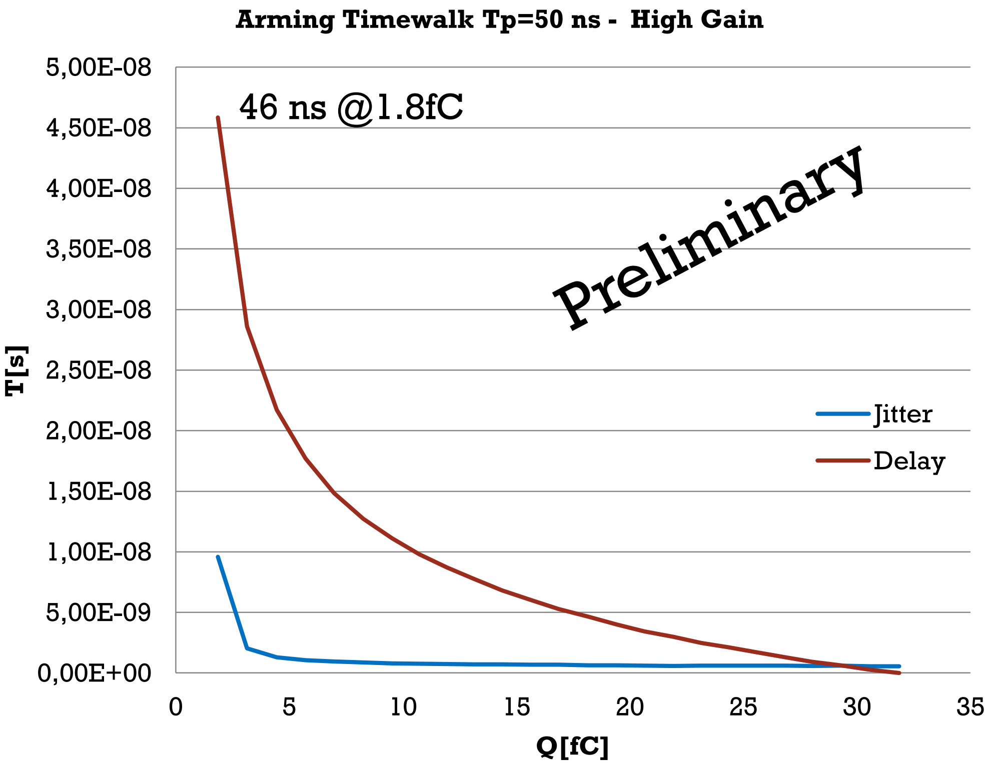


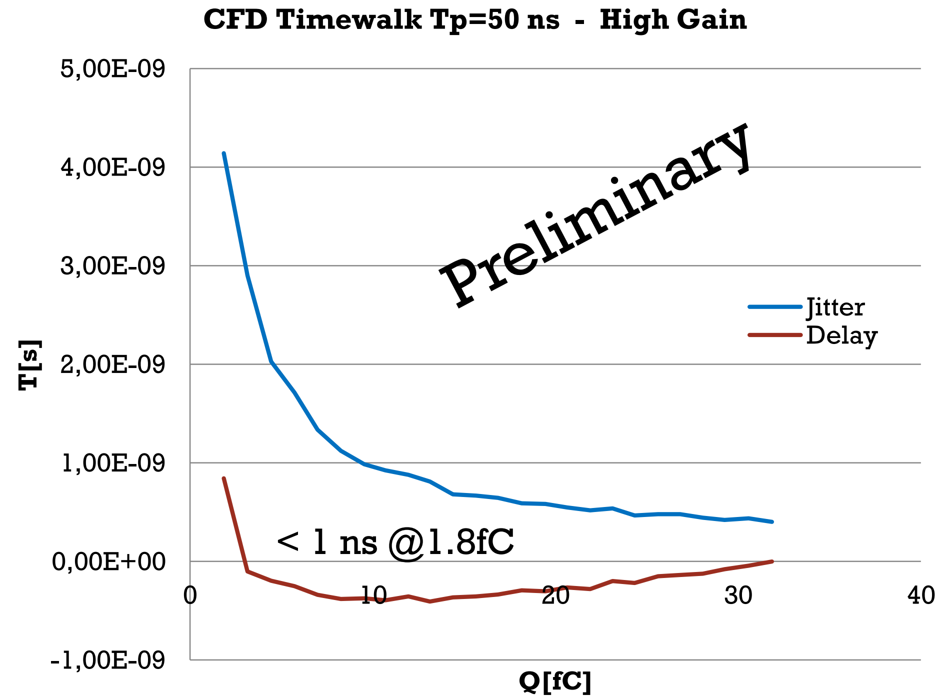
TP [ns]	Noise slope [e/pF]
15	75
25	47
35	36
50	33



TP [ns]	Noise slope [e/pF]
15	57
25	40
35	31
50	28

VFAT3: TIMING RESOLUTION





- Arming comparator mode (leading edge)
- Delay shows the time-walk
- Relatively high time-walk, as expected

- CFD comparator mode: time walk drastically reduced
- Timing precision is dominated by jitter

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

• 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

• Depending on the event rate, baseline restoration may be needed

• 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

• GASTONE64 and VFAT3: two examples of FEE for GEM detectors

BIBLIOGRAPHY

1. Helmut Spieler, "Front-End Electronics and Signal Processing" (presented at the ICFA Instrumentation Center in Morelia, Mexico, 2002 http://www-physics.lbl.gov/~spieler

2. Helmut Spieler, "Front-End Electronics for Detectors" (presented at 2007 IEEE Nuclear Science Symposium in Honolulu, Hawaii, 2007 http://www-physics.lbl.gov/~spieler

3. V. Radeka, "Low-Noise techniques in detectors", Ann. Rev. Part. Sci. 1988. 38: 217..277

4. Goulding,F.S.(1972). "Pulse shaping in low noise nuclear amplifiers: a physical approach to noise analysis" Nucl.Instr.andMeth.100(1972)493–504

5. Goulding,F.S. and Landis,D.A.(1982). "Signal processing for semiconductor detectors." IEEE Trans. Nucl. Sci. NS-29/3(1982)1125–1141

6. Radeka,V.(1974). "Signal, noise and resolution in position-sensitive detectors". IEEE Trans. Nucl. Sci. NS-21(1974)51–64