CPAN NETWORK
ON INSTRUMENTATION
AND DETECTORS

Ayudas ED2024-153684-T,
CEX2023-001292-S financiadas por

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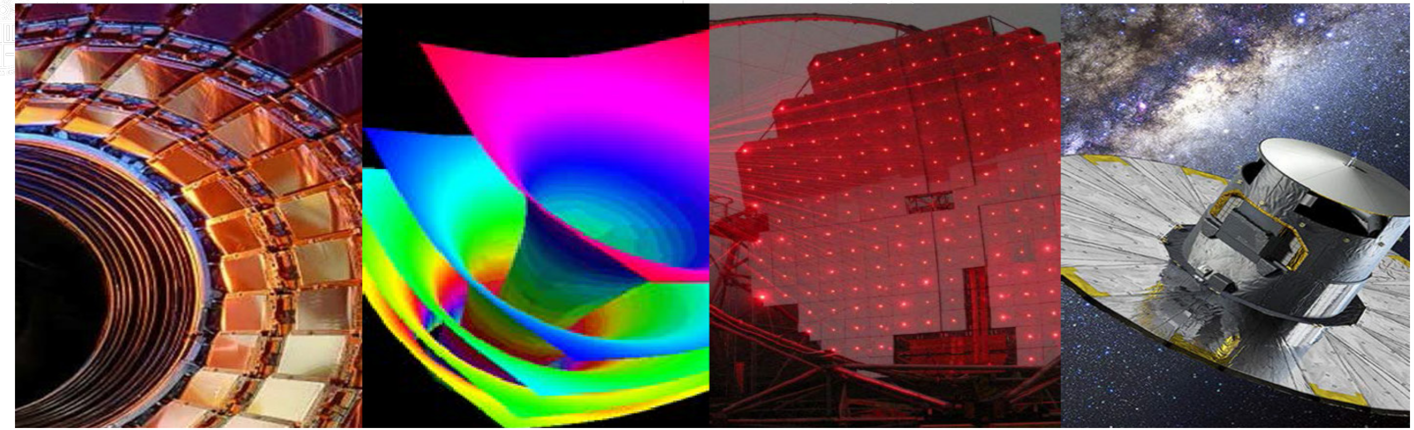

Electronics

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TIDPAN 2026 - Taller de Instrumentación y Detectores en Física de Particulas, Astropartículas y Nuclear

17 June 2026

ICCUB Institute of Cosmos Sciences
EXCELENCIA MARÍA DE MAEZTU
UNIVERSITAT DE BARCELONA

Introduction

- Electronics domains in modern particle, astroparticle and nuclear physics experiments are vast:
  - Detector Front End Electronics
  - Acquisition and trigger
  - Links (optical, copper, wireless)
  - Power distribution and cooling
  - Grounding and Electromagnetic Compatibility (EMC)

See active R&D in these domains:
↗TWEPP 26: https://indico.cern.ch/event/1643631/
↗DRD program: https://drd7.web.cern.ch/

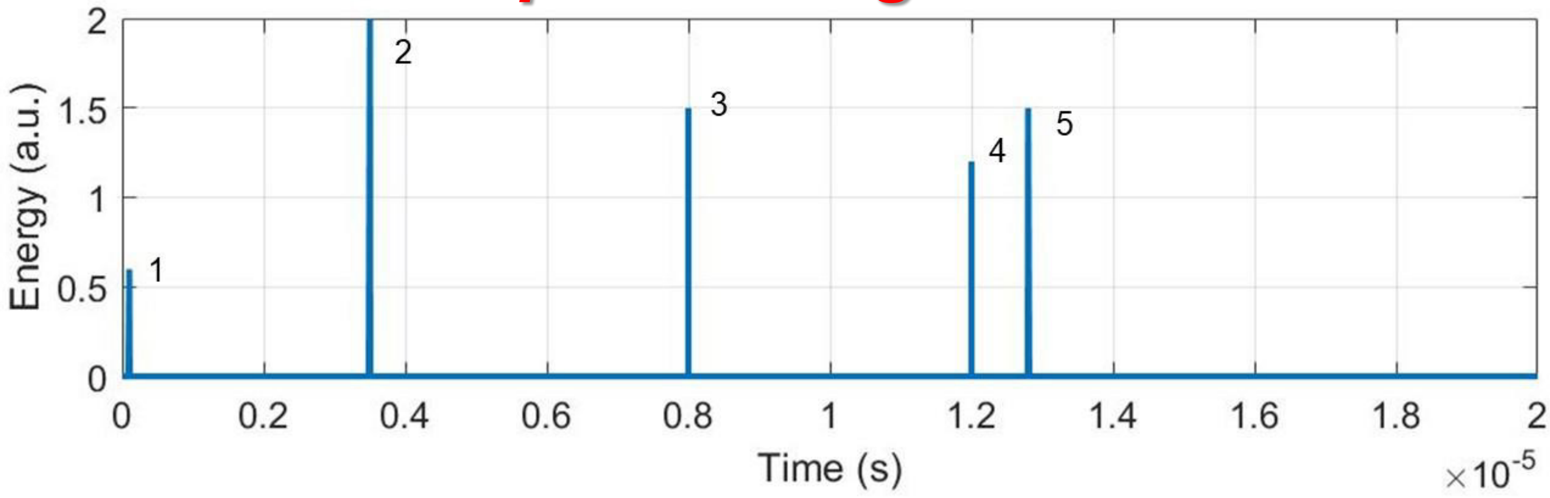
Introduction

- Even so it is a broad topic and it's difficult to cover it in 1h
- We will try to present the key concepts and provide references
- For a general introduction to electronics in our field
  - Grupen C, Shwartz B. Electronics in "Particle Detectors". 2nd ed. Cambridge University Press; 2023.
  - Spieler, Helmut , Semiconductor Detector Systems, Clarendon Press, Oxford, 2005.
- For details about ASIC design for detectors:
  - Rivetti, Angelo. CMOS : front-end electronics for radiation sensors . Boca Raton : CRC Press, 2015. ISBN 113882738X.
- For a general introduction to circuit theory and analog circuit design
  - https://www.youtube.com/@AliHajimiriChannel/courses

Introduction

- A particle interaction (photons, electrons, protons, etc) generates a current **pulse** in the detector

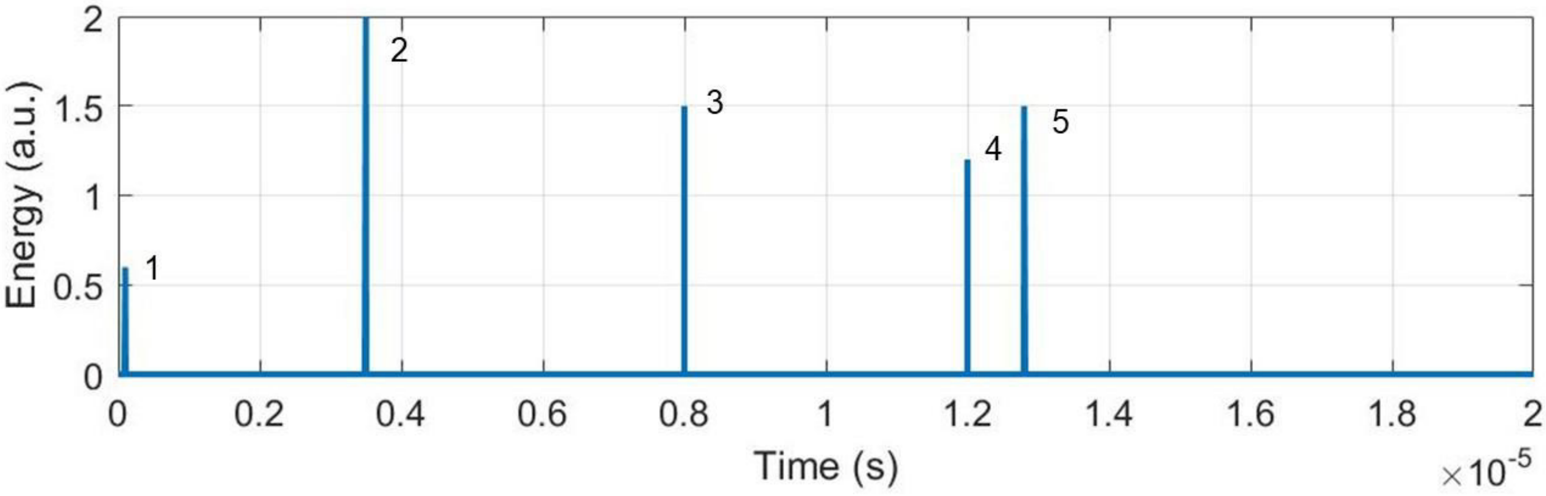
Pulse processing electronics



↗ R. Ballabriga, "Introduction to hybrid pixel detectors", Barcelona Techno Week, 2023

Introduction

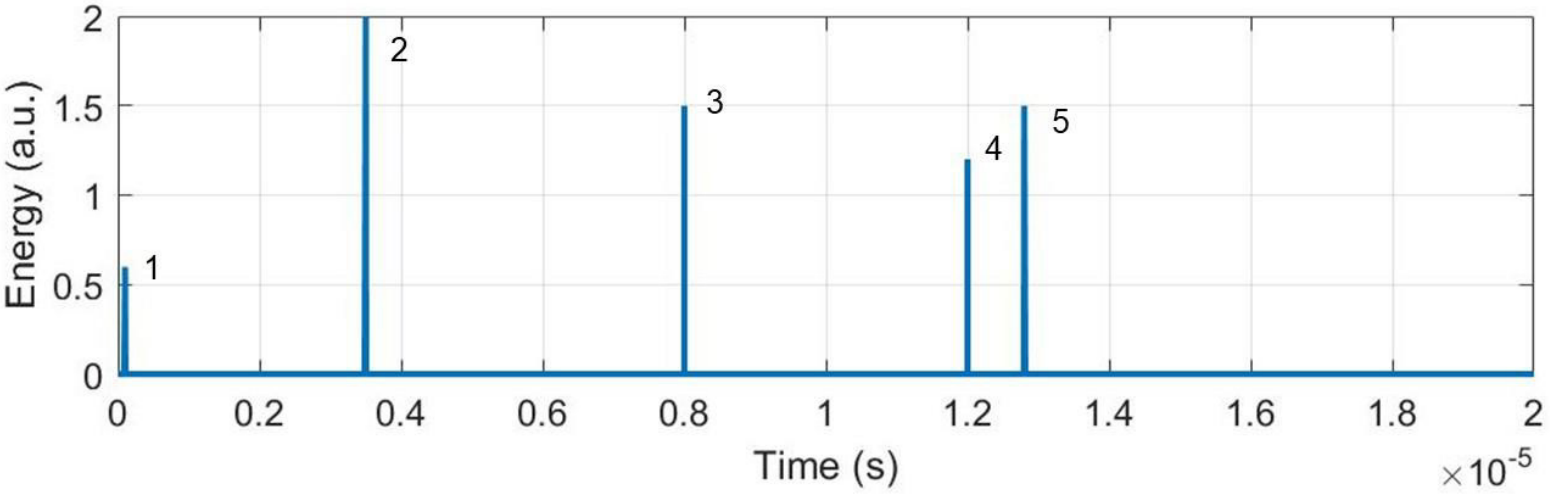
- The readout electronics must pick-up the relevant characteristics of the signal



↗ R. Ballabriga, "Introduction to hybrid pixel detectors", Barcelona Techno Week, 2023

Introduction

▶ What is relevant?



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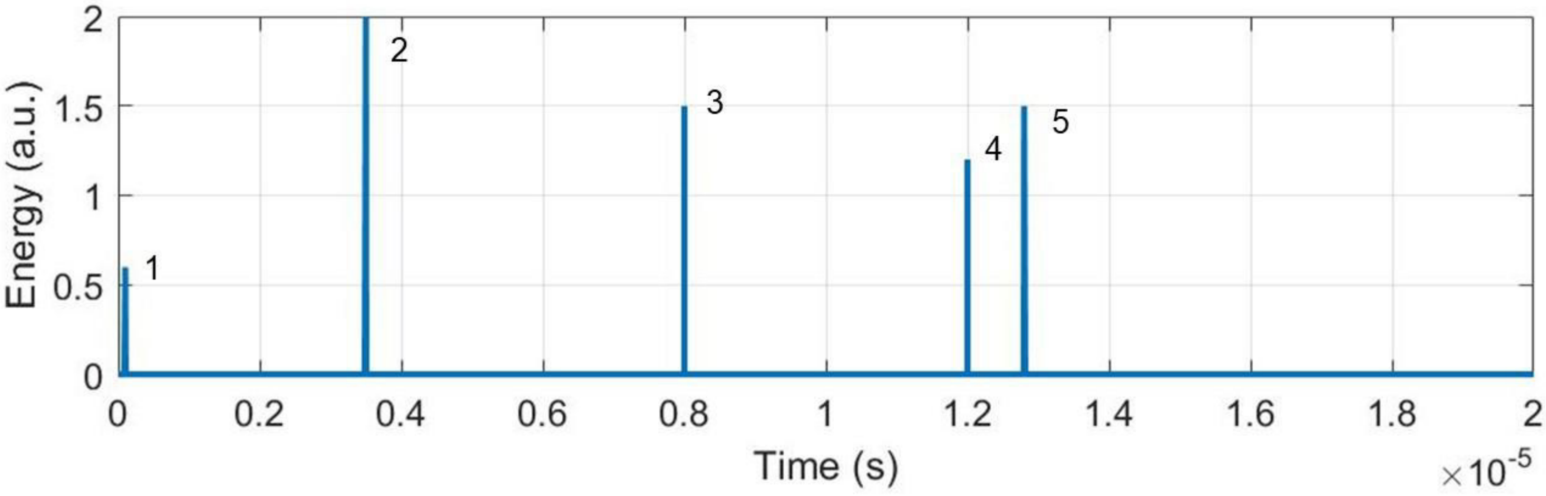
Introduction

▶ What is relevant?

The detection (presence) of a particle: here we have 5



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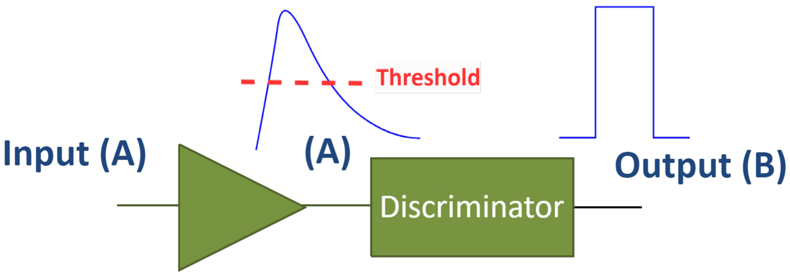
Introduction

- What kind of readout electronics?

Pulse processing electronics: detection (presence) of a particle



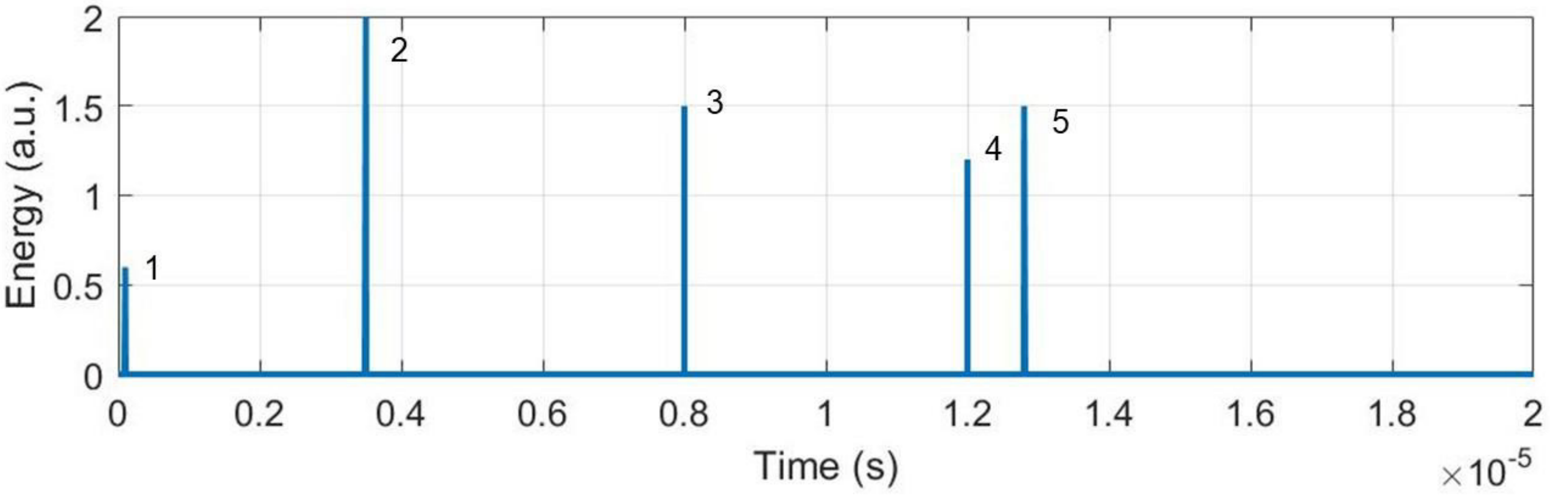
Domains: (A): analog // (B): binary // (D): digital

- The input stage (or preamplifier) senses the analog detector signal
  - A filter might be used after preamplifier to shape the signal and to improve the signal to noise ratio
- The output is a binary signal:
  - The amplitude is digital (“1”/“0”) but the time of the edges preserves information

Introduction

▶ What is relevant?

The rate: 5 in 20 μs (250 kHz on average)



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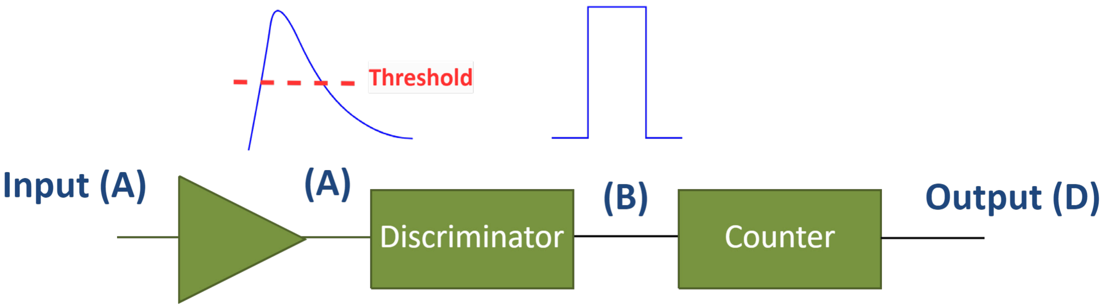
Introduction

- What kind of readout electronics?

Pulse processing electronics: computing the rate

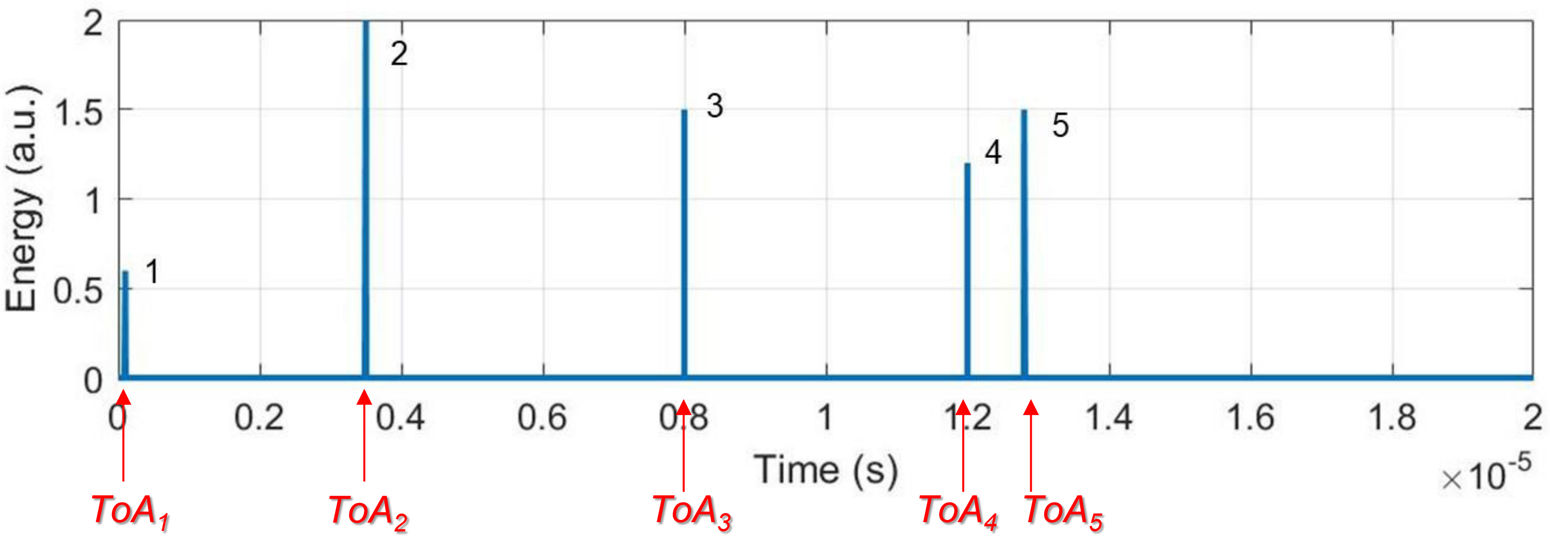


- The output is a digital value encoding the rate

Introduction

- What is relevant?

The Time of Arrival (ToA)



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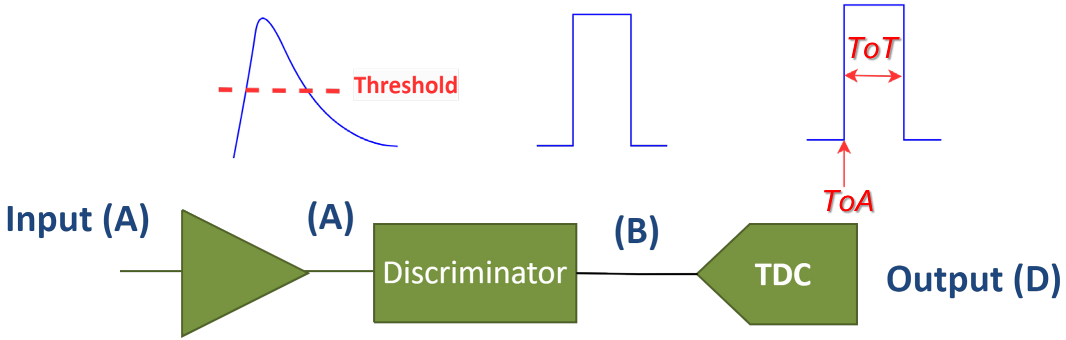
Introduction

- What kind of readout electronics?

Pulse processing electronics: Time of Arrival (ToA)

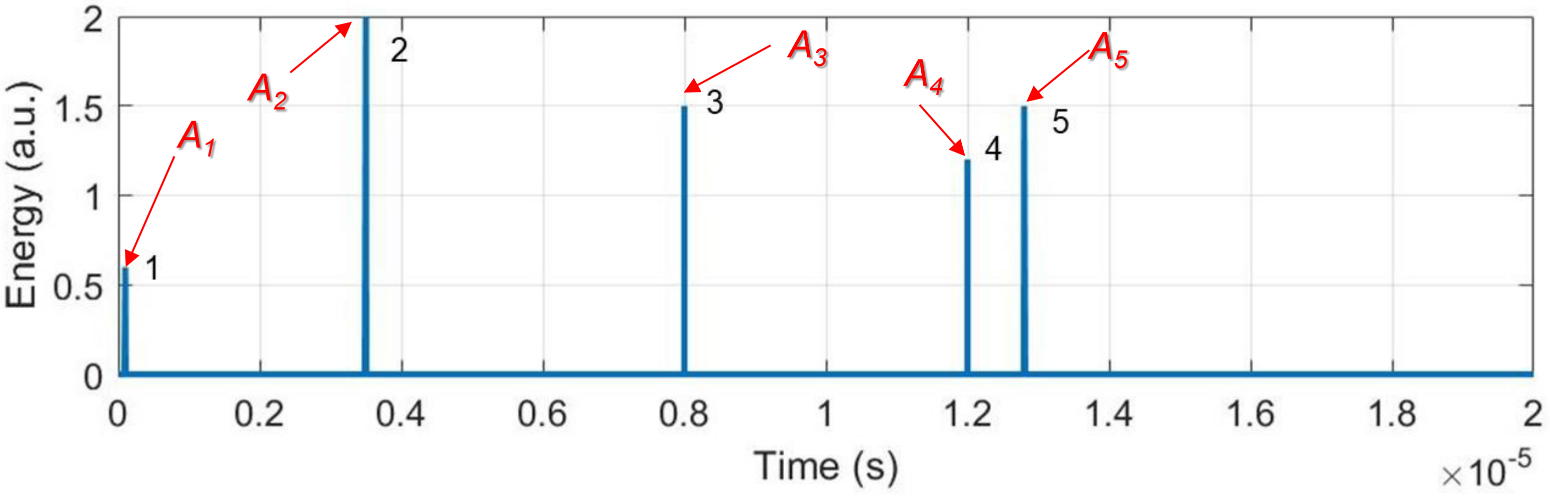


- The output is a digital value encoding the time of the leading edge of pulse at the discriminator output (a measurement of the ToA)
  - A Time to Digital converter is required to digitize the edges of the disc. Output
  - Also the trailing edge can be digitized, to encode the Time-Over-Threshold (ToT)
    - Provides some information on energy deposition

Introduction

▶ What is relevant?

The Amplitude (or area) which might be proportional to the deposited energy



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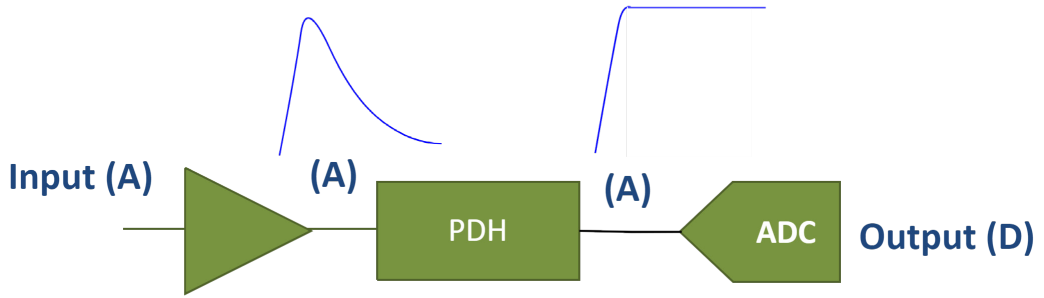
Introduction

- What kind of readout electronics?

Pulse processing electronics: Amplitude (deposited energy)



- A Peak Detector and Hold (PDH) circuit samples de peak value of the preamplifier output and holds it for digitization by the ADC
- A filter (or shaper) is usually inserted after preamplifier to shape the signal and to improve the signal to noise ratio

Introduction

- Do you imagine an alternative to do everything with a single readout electronics?

Introduction

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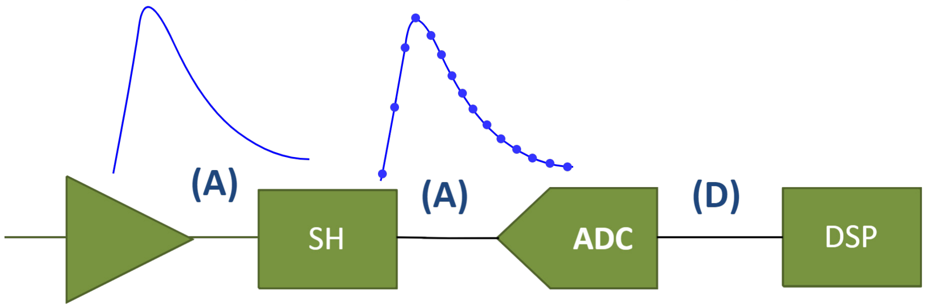
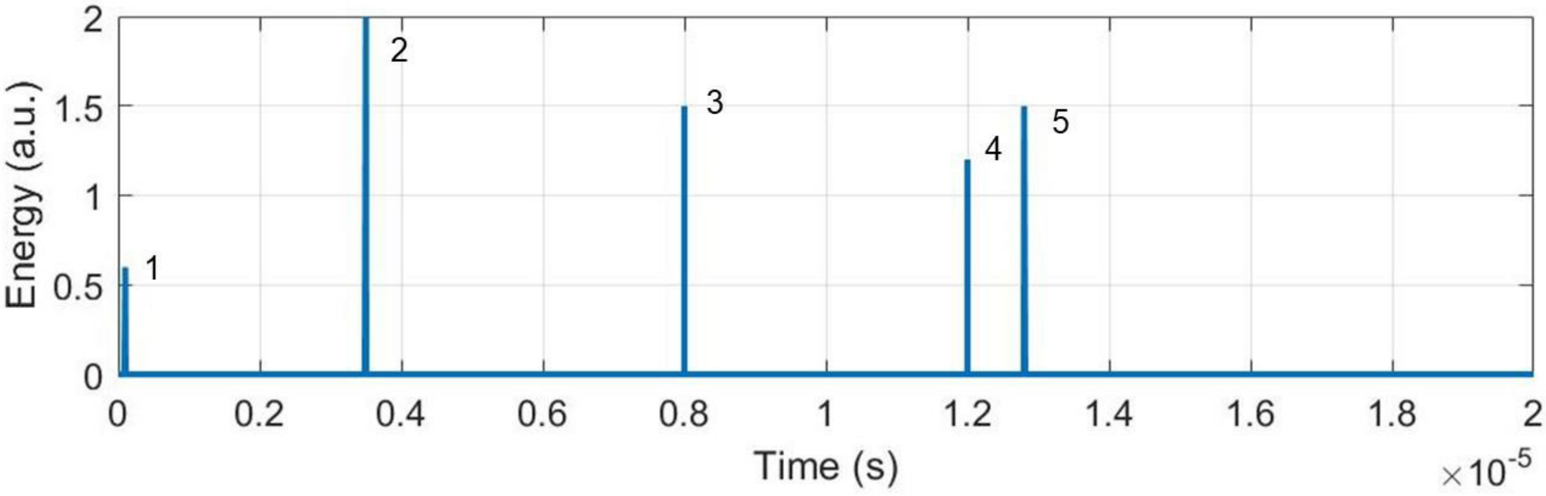
Waveform sampling electronics: everything !



- Sampling and digitization at sufficient rate (>> Signal BW)
  - ADC: typically, SAR or Flash ADC or analog memory (SCA)
- Digital Signal Processing (DSP): rate, ToA, energy, etc
  - Also in some detectors, pulse shape analysis techniques to identify different particles

Introduction

- Back to the beginning: a particle interaction generates a current pulse in the detector
  - What happens if the detector or the FE electronics is veeeeery slow ?



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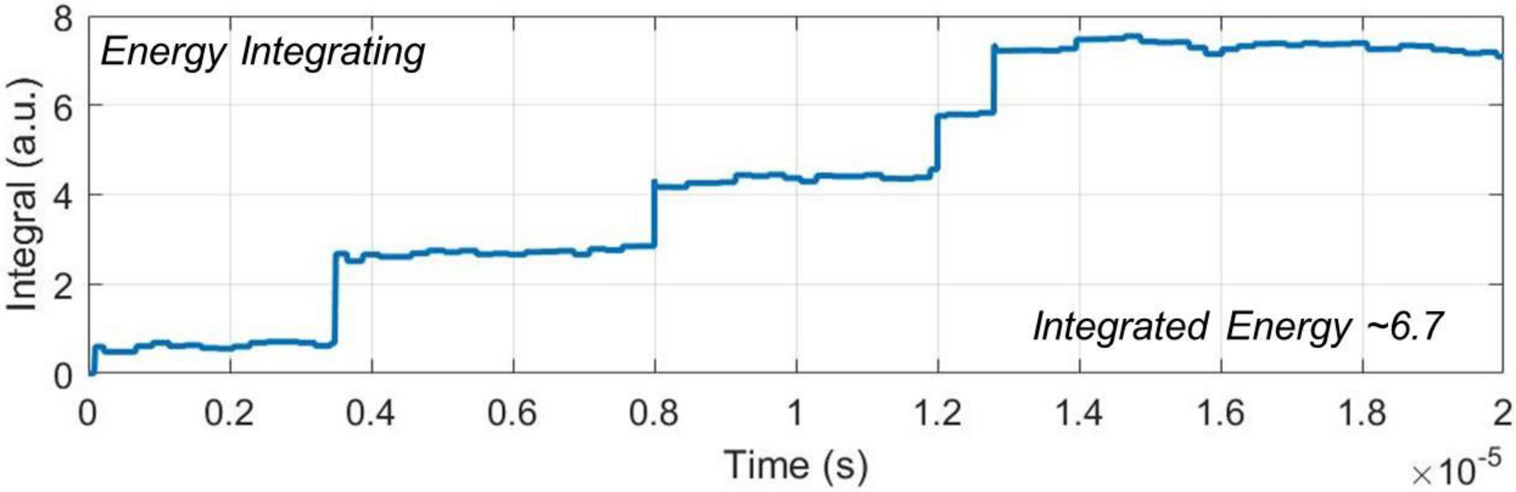
Introduction

▶ What happens if the detector or the FE electronics is veeeeery slow ?

Integrating mode (not pulse!): many CCD or CMOS imagers, dosimeters, etc



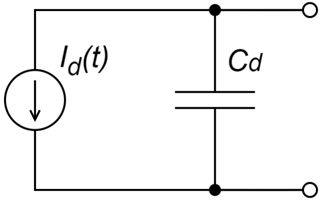
↗ R. Ballabriga, "Introduction to hybrid pixel detectors", Barcelona Techno Week, 2023

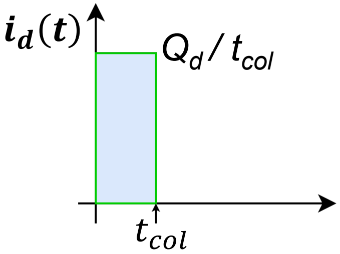
Outlook

1. **Detector electrical model**
2. Input stage (preamplifier)
3. Shaper (filter)
4. Noise analysis and optimization
5. Applications: ASICs

Introduction: detector model

- To design front-end circuits, we need a realistic electrical model of the sensor
- The simplest model will include:
  - A current source, $I_d(t)$, representing the sensor signal as a function of time
    - The pulse of a duration is often called “collection time”(t_{col})
    - The pulse conveying the input charge Q_d
    - C_d models the capacitive load the sensor presents to the FE electronics (source impedance)
    - The direction of the current source must be chosen to properly reflect the signal polarity

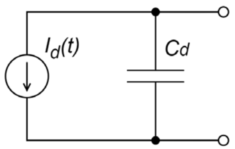


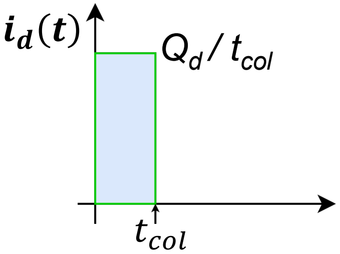


↗ Rivetti, Angelo. CMOS : front-end electronics for radiation sensors . Boca Raton : CRC Press, 2015.

Detector model: time domain characteristics

- Different approaches can be taken to model the transient behavior of $I_d(t)$
  - When the impulse response of the FE is much slower than t_{col}, the detector response can be represented by a Dirac impulse ($t_{col} \rightarrow 0$) conveying the input charge Q_d
  - In some cases, additional signal features need to be introduced:
    - More than one exponential signal to model different collection times of different carriers
    - Some sensors as the SiPMs require in principle a more complex model



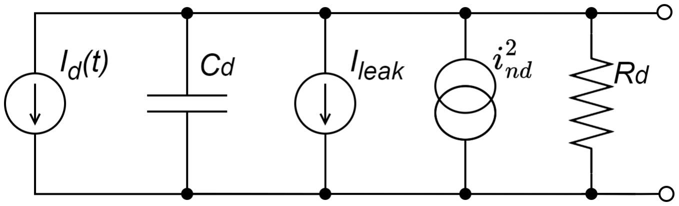


sensor	C_d	~ current pulse duration
MAPS, SDD, hybrid pixel	10 – 50 fF	Few ns
Si strip	10 pF	Few ns
MWPC	50 pF	10 – 500 ns
SPAD/SiPM	Few/hundred pF	20 ps
CZT planar 1cm	20 pF	1 µs
Liquid Argon calorimeter	1 nF	500 ns

↗ Paul O'Connor, "Pulse Processing Electronics Techno Week", Barcelona Techno Week, 2021

Detector model

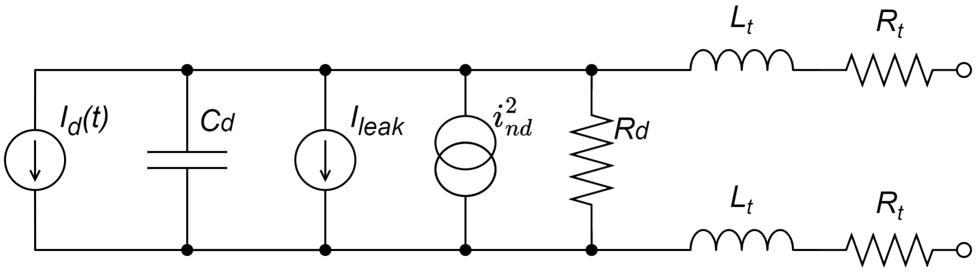
- We can also include the non-idealities of the sensor in the model:
  - The current source I_{leak} takes into account the device leakage current
    - In many semiconductors it is modelled as a DC current source
    - In sensors with gain (as SPADs or SiPMs) it can a pulse current source modeling DCR and correlated noise pulses (afterpulses an crosstalk)
  - i_{nd}^2 is the noise generator representing the “shot noise” of the dark current
  - R_d models the resistive component of the sensor output impedance (usually negligible) or physical resistors that might be used in the sensor biasing network
    - If it models a physical resistor the corresponding current noise contribution should be added to i_{nd}^2



↗ Rivetti, Angelo. CMOS : front-end electronics for radiation sensors . Boca Raton : CRC Press, 2015.

Detector model

- Finally, in high speed and very low noise applications we may need to model the interconnection
  - The simplest approach is a lumped parameter model:
    - Inductors L_t and resistor R_t describe respectively the parasitic inductance and resistance of the connections
  - If this is not precise enough RF models and simulator should be used



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Outlook

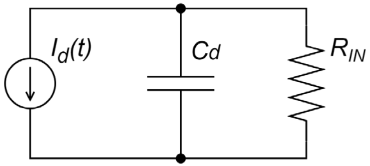
1. Detector electrical model
2. **Input stage (preamplifier)**
3. Shaper (filter)
4. Noise analysis and optimization
5. Applications: ASICs

Input stage

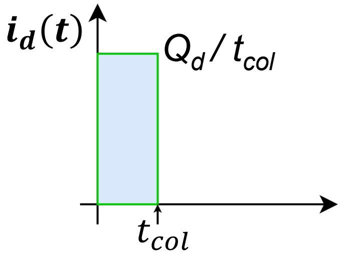
- Very often they are preamplifiers (or low noise amplifiers (LNAs))
  - But in some cases, they are just current or voltage buffers
- In general, the main purpose of an input stage is to properly sense the detector signal
- What does mean properly? According to specific requirements mainly on:
  - Signal to noise ratio
  - Dynamic range
  - Impedance adaptation
  - Pulse shape
  - Event rate
  - Power
  - Area

Input stage: input impedance

▶ Let's begin by examining the role of the input impedance R_{IN} of the input stage in sensing the detector signal



$\tau_{DET} = C_d R_{IN}$



- Simplified detector model
  - C_d and R_{IN} form a current divider
  - Current sensing
    - Current though R_{IN}

$$I_{R_{IN}}(s) = I_d(s)\frac{1/sC_d}{1/sC_d + R_{IN}} = I_d(s)\frac{1}{s\tau_{DET} + 1}$$

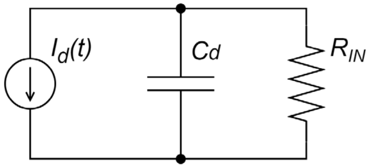
First order response

  - Voltage sensing
    - Voltage across R_{IN}

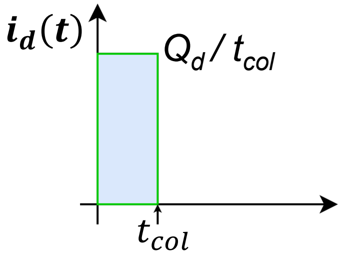
$$V_{R_{IN}}(s) = I_d(s)\frac{R_{IN}}{s\tau_{DET} + 1}$$

Input stage: input impedance

- We can compute the response by inverse Laplace Transform ($\mathcal{L}^{-1}$)
  - With : $i_{d(t)} = \frac{Q_{in}}{t_{col}}(u(t) - u(t - t_{col}))$



$\tau_{DET} = C_d R_{IN}$



- Simplified detector model
  - C_d and R_{IN} form a current divider
  - Current sensing
    - Current though R_{IN}

$$I_{R_{IN}}(t) = \frac{Q_{in}}{t_{col}}\left[\left(1 - e^{-\frac{t}{\tau_{DET}}}\right)u(t) - \left(1 - e^{-\frac{t - t_{col}}{\tau_{DET}}}\right)u(t - t_{col})\right]$$

  - Voltage sensing
    - Voltage across R_{IN}

$$V_{R_{IN}}(t) = R_{IN}\frac{Q_{in}}{t_{col}}\left[\left(1 - e^{-\frac{t}{\tau_{DET}}}\right)u(t) - \left(1 - e^{-\frac{t - t_{col}}{\tau_{DET}}}\right)u(t - t_{col})\right]$$

Input stage: input impedance

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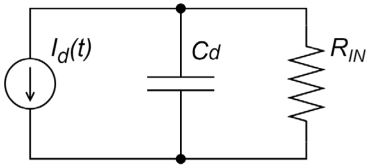
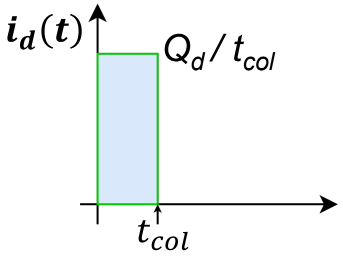


$\tau_{DET} = C_d R_{IN}$



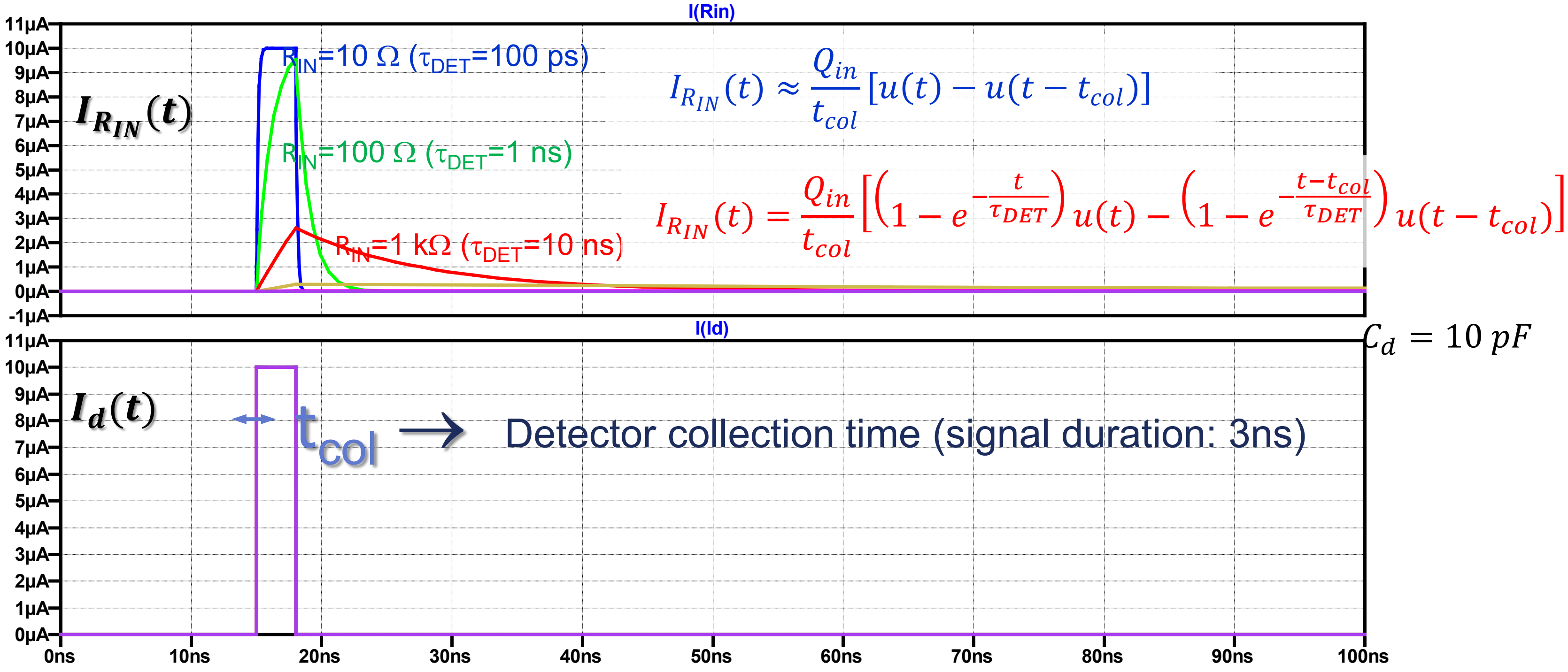
- Simplified detector model
  - C_d and R_{IN} form a current divider
  - Current sensing
    - Current though R_{IN} $\quad I_{R_{IN}}(t) \approx \frac{Q_{in}}{t_{col}}[u(t) - u(t - t_{col})]$

$t_{col} >> \tau_{DET}$

  - Voltage sensing
    - Voltage across R_{IN} $\quad V_{R_{IN}}(t) \approx R_{IN}\frac{Q_{in}}{t_{col}}[u(t) - u(t - t_{col})]$

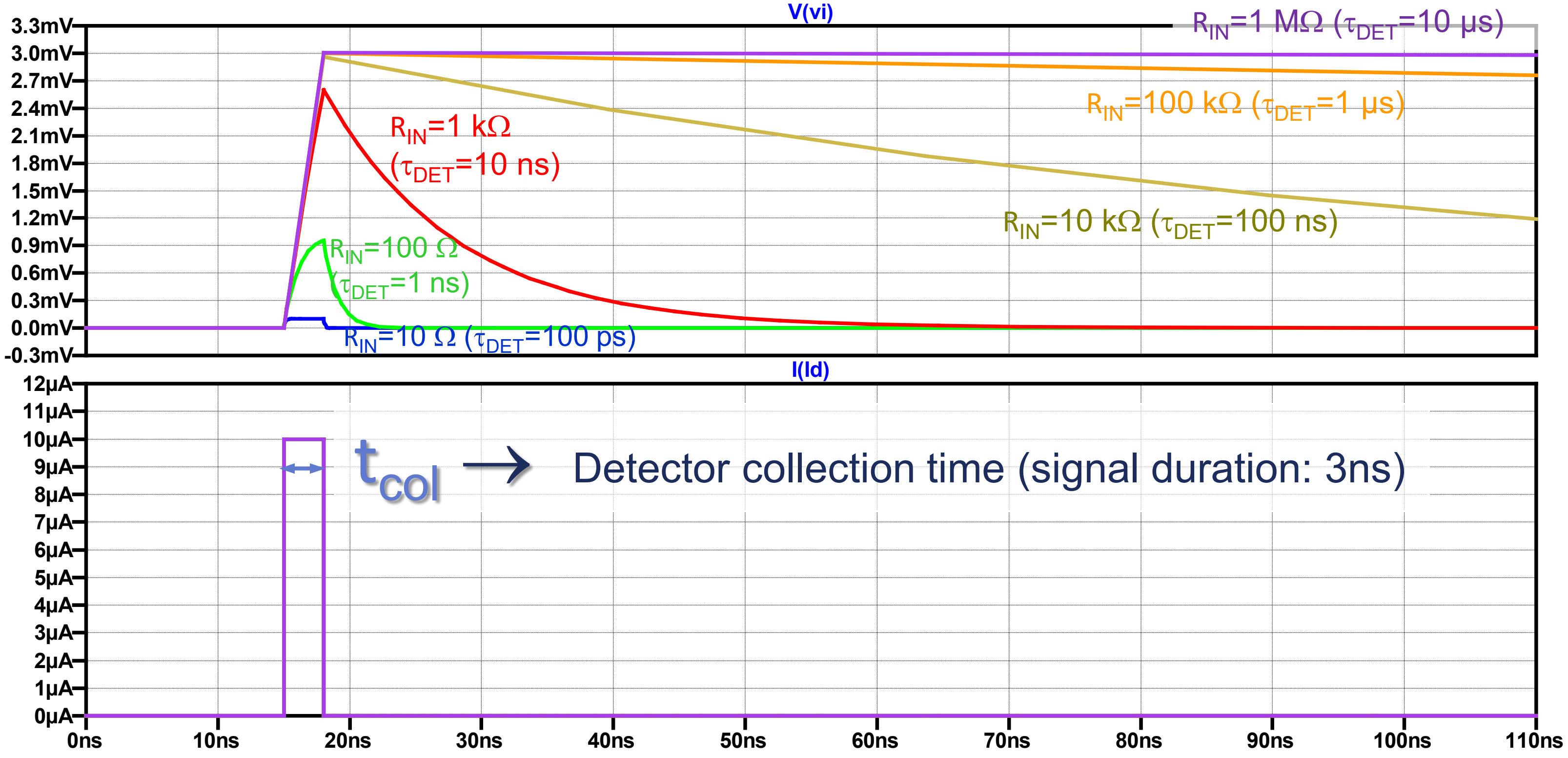
Input stage: input impedance and time constants

- Current sensing: the smaller the R_{IN}, the larger the input signal
  - If $R_{IN} \approx 0$ then $I_{R_{IN}} \approx I_d$



Input stage: input impedance and time constants

- Opposite situation in voltage mode: the larger the R_{IN}, the larger the input signal
  - But note that R_{IN} (more precisely τ_{DET}) affects the shape of the signal!



Input stage: input impedance and time constants

$t_{col} \gg \tau_{DET} \rightarrow \boldsymbol{I_{R_{IN}}(s) \approx I_d(s)}$

$V_{R_{IN}}(s) \approx I_d(s)\, R_{IN}$

$V_{R_{IN}}(t) \approx I_d(t) R_{IN}$

I/V sensing. Shape is preserved

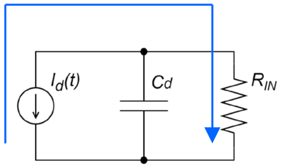
$\xrightarrow{\tau_{DET} \uparrow}$

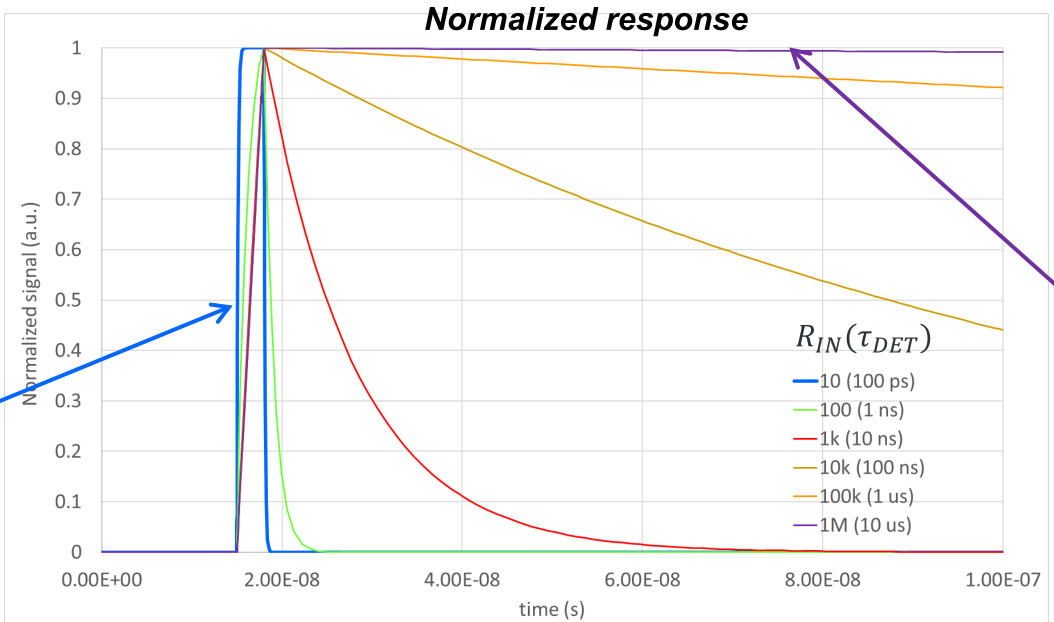
$t_{col} \ll \tau_{DET} \rightarrow \boldsymbol{I_{R_{IN}}(s) \approx I_d(s) \frac{1/sC_d}{R_{IN}}}$

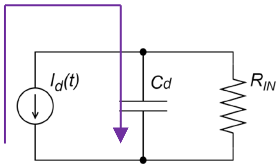
$V_{R_{IN}}(s) \approx I_d(s)\, {}^{1}/_{sC_d}$

$V_{R_L}(t) \approx \frac{\int I_d(t)dt}{C_d} \rightarrow V_{R_L}(\tau_{DET} \gg t > t_{col}) \approx \frac{Q_d}{C_d}$

Integration: charge sensitive







“Fast” front-end response

“Slow” front-end response

Input stage: topologies

- Open loop
  - Common source, common gate, etc
    - Sometimes local feedback is employed
  - Preferred for
    - High speed
    - Very compact implementation (e.g.: pixels in CMOS imagers)

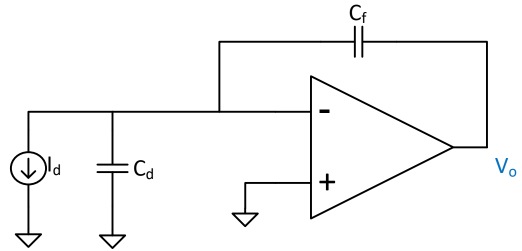
- Closed loop
  - An amplifier with negative feedback
  - Transimpedance amplifiers, charge amplifiers, etc
  - Preferred for
    - Linearity
    - Accuracy

Input stage: closed loop topologies: CSA

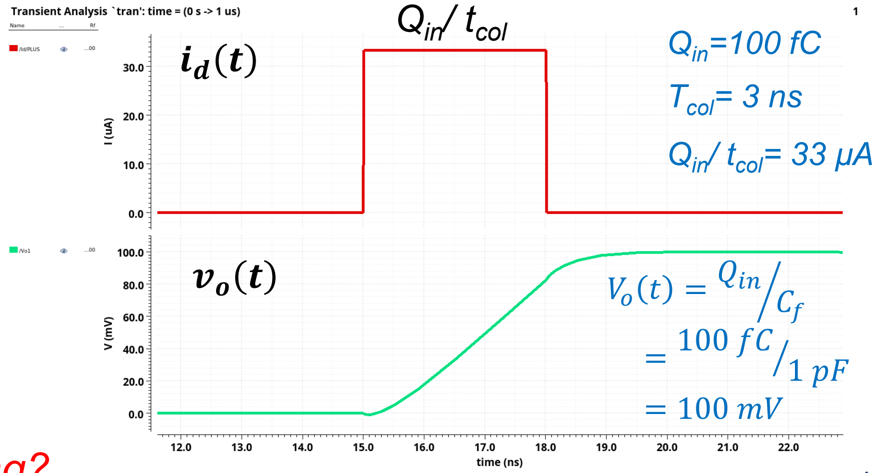
$$Z_f(s) = {}^{1}/_{sC_f}$$

Charge Sensitive Amplifier (CSA)

- Capacitive feedback
- High gain and low noise
- **The most popular front end for semiconductors with no intrinsic gain**
  - **Pixels, strips, etc**
- Integration: long tail
  - Pulse pile-up (see next section)
  - Problems for high-rate detectors



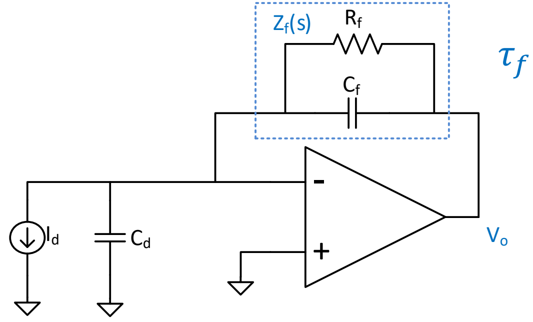
$$\frac{V_o}{I_i}(s) = \frac{1}{sC_f} \xrightarrow{\mathcal{L}^{-1}} V_o(t) = {}^{1}/_{C_f} \int I_i(t) = {}^{Q_{in}}/_{C_f}$$



Anything missing?

Input stage: closed loop topologies: CSA

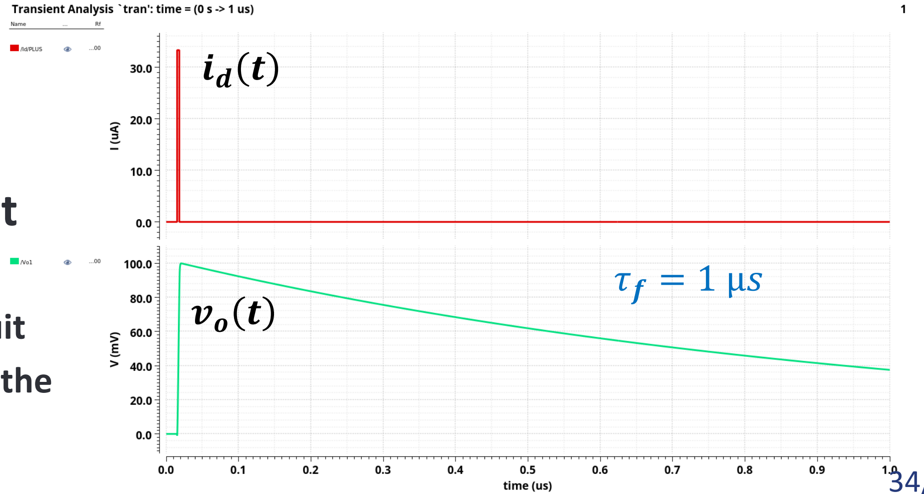
$$Z_f(s) = \frac{R_f}{s\tau_f + 1} \approx {}^{1}\!/_{sC_f}$$



$$\tau_f = C_f\, R_f$$

- Charge Sensitive Amplifier (CSA)
  - Capacitive feedback
  - High gain and low noise
  - The most popular front end for semiconductors with no intrinsic gain
    - Pixels, strips, etc
  - Integration: long tail
    - Pulse pile-up (see next section)
    - Problems for high-rate detectors
  - **We need a feedback resistor R_f to set the DC operating point**
    - **Can be a passive resistor or an active circuit**
    - **Must be large enough so $Z_f(s) \approx {}^{1}\!/_{sC_f}$ in the frequency range of interest**

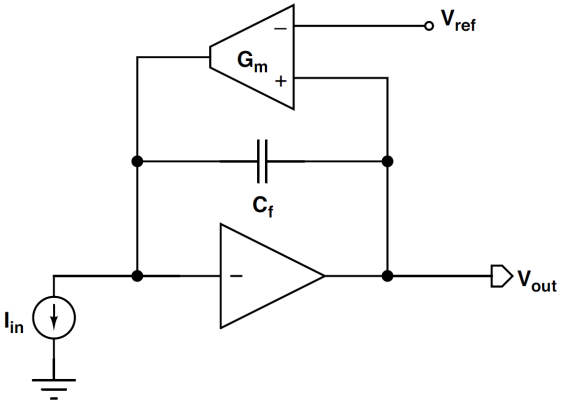
$$\frac{V_o}{I_i}(s) \approx \frac{1}{sC_f} \xrightarrow{\mathcal{L}^{-1}} V_o(t) \approx {}^{1}\!/_{C_f} \int I_i(t) \approx {}^{Q_{in}}\!/_{C_f}$$



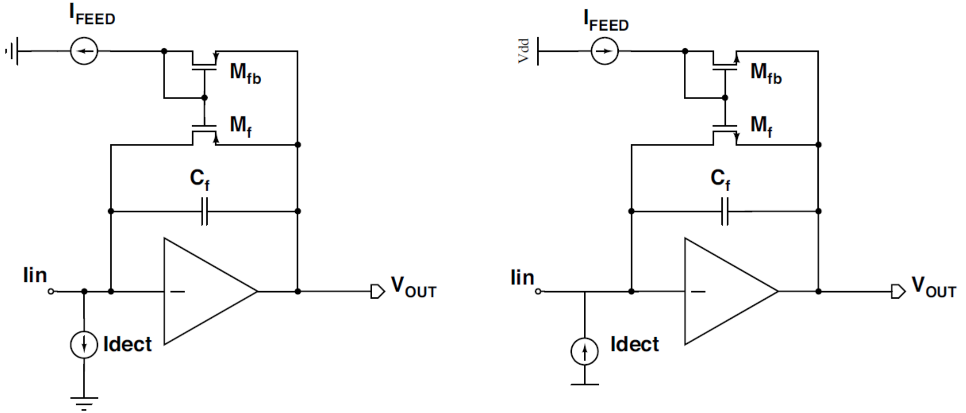
Input stage: closed loop topologies: CSA

- Examples of active reset circuits

Discharge based on transconductor



Constant current discharge



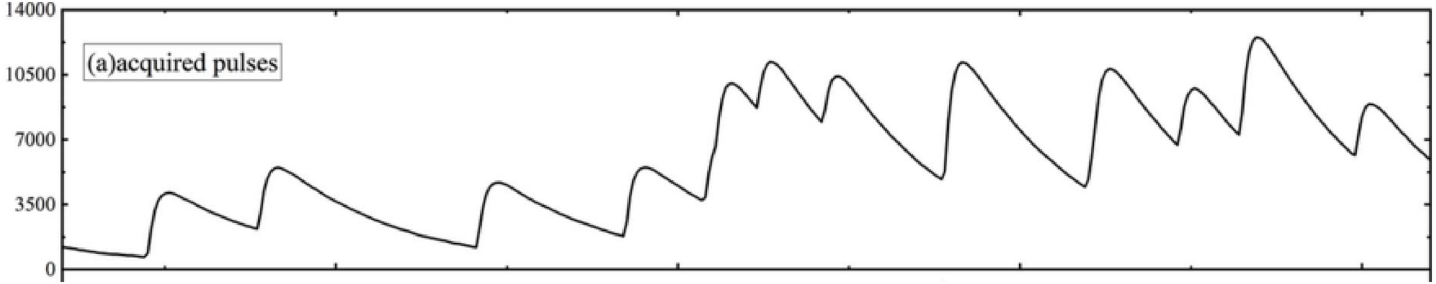
↗ Rivetti, Angelo. CMOS : front-end electronics for radiation sensors . Boca Raton : CRC Press, 2015

Outlook

1. Detector electrical model
2. Input stage (preamplifier)
3. **Shaper (filter)**
4. Noise analysis and optimization
5. Applications: ASICs

Shaper: Introduction

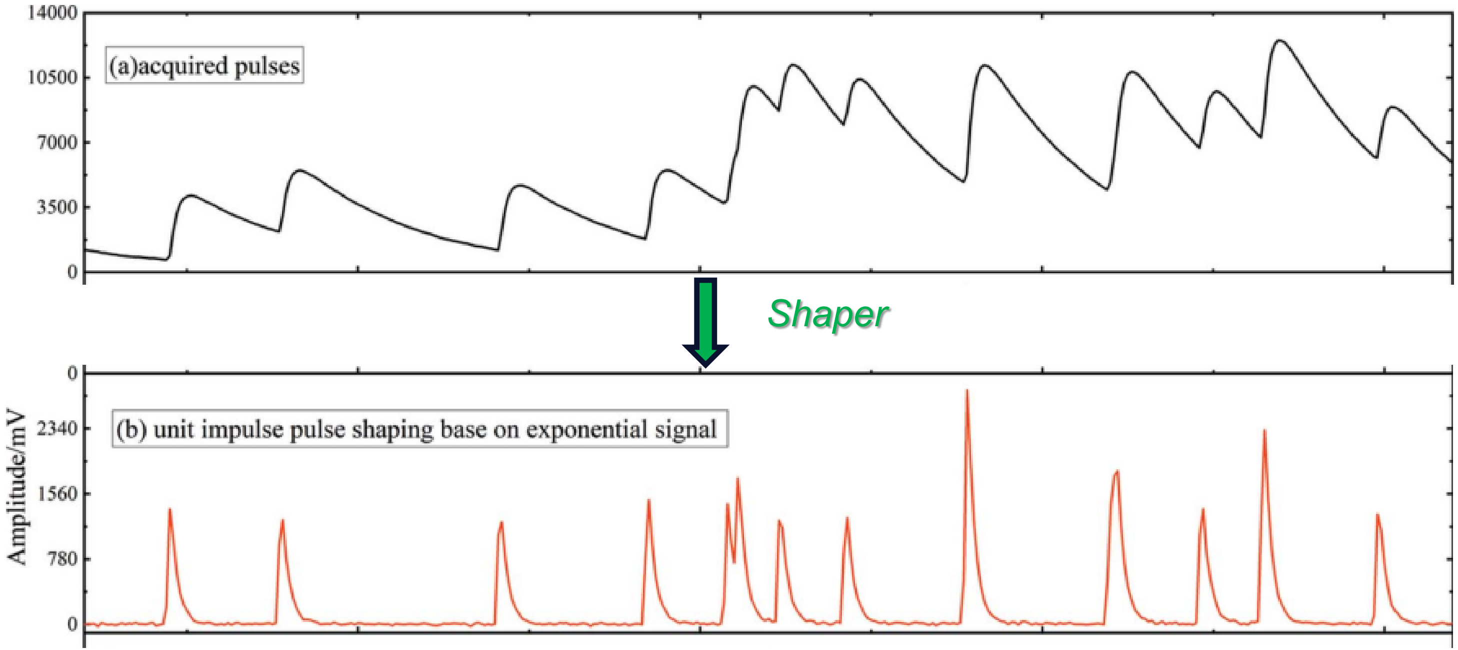
- When the event rate is high enough, the output pulses of the preamplifier will pile up
  - This is particularly important for CSA because of the long tails (large τ_f is required)
- This leads to random measurement errors (the distance between pulses is random)



↗ J. Synchrotron Rad. (2020). 27, 1609–1613

Shaper: Introduction

- The solution is to "shape" the signal to shorten the long tails
  - The shaper is nothing else than a filter analyzed in the time domain



↗ J. Synchrotron Rad. (2020). 27, 1609–1613

Shaper: Introduction

- Shaper is characterized by:
  - Impulse or step response in the time domain: **peaking time t_p**
  - Transfer function in the frequency domain

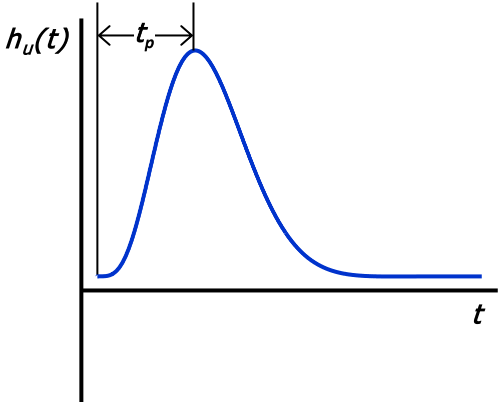
- Shapers serve mainly 2 purposes:
  1) Shape the signal
     - Minimize pile-up
     - Minimize baseline fluctuations
     - Reduce amplitude and timing error
  2) Optimize signal to noise ratio
     - Shaper modifies the spectrum of both signal and noise
     - The goal is to optimize the SNR
     - More on the next section

Unipolar CR-RC^n impulse response Shaper(t)

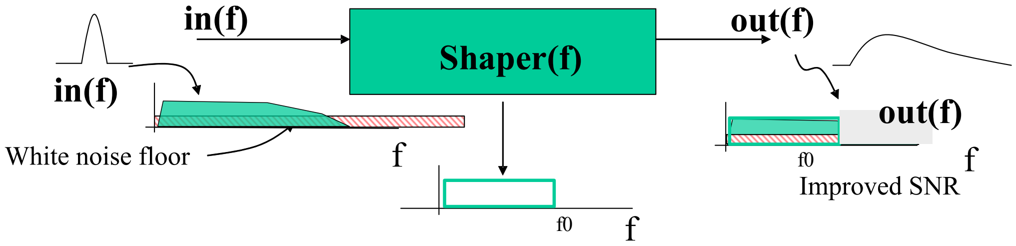
$$H_u(s) = \frac{s\tau}{(1+s\tau)^{n+1}}$$

$$h_u(t) = \frac{1}{n!}\left(\frac{t}{\tau}\right)^n e^{-t/\tau}$$

$$t_p = n\tau$$



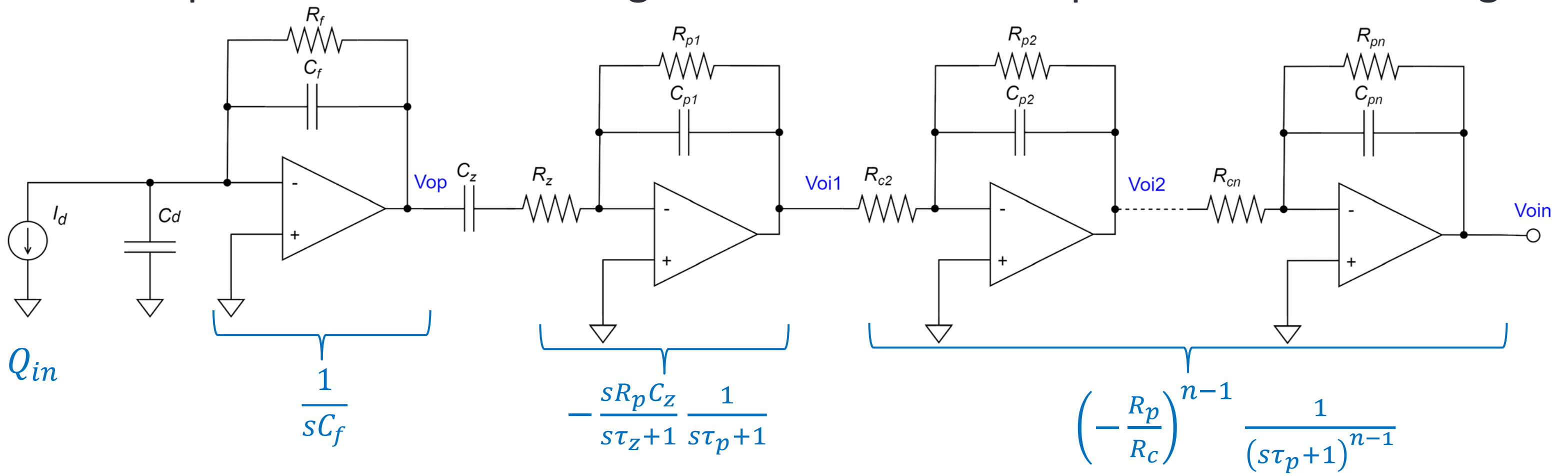
↗ P. O'Connor, "Pulse Processing Electronics", Third Barcelona Techno Week. Course on semiconductor detectors. 2018



↗ From F. Anghinolfi, "ANALOG SIGNAL PROCESSING OF PARTICLE DETECTOR SIGNALS", CERN Technical Training, 2005

Shaper: practical implementation of Semi-Gaussian shapers (CR-RC^n)

- Schematic representation: CSA + Shaper
- Combines impedance isolation and gain to recover the amplitude lost in filtering



$\tau_p = \tau_{pi}$

$R_p = R_{pi}$

$C_p = C_{pi}$

$R_c = R_{ci}$

$\tau_z = C_z R_z$

$\tau_p = C_p R_p$

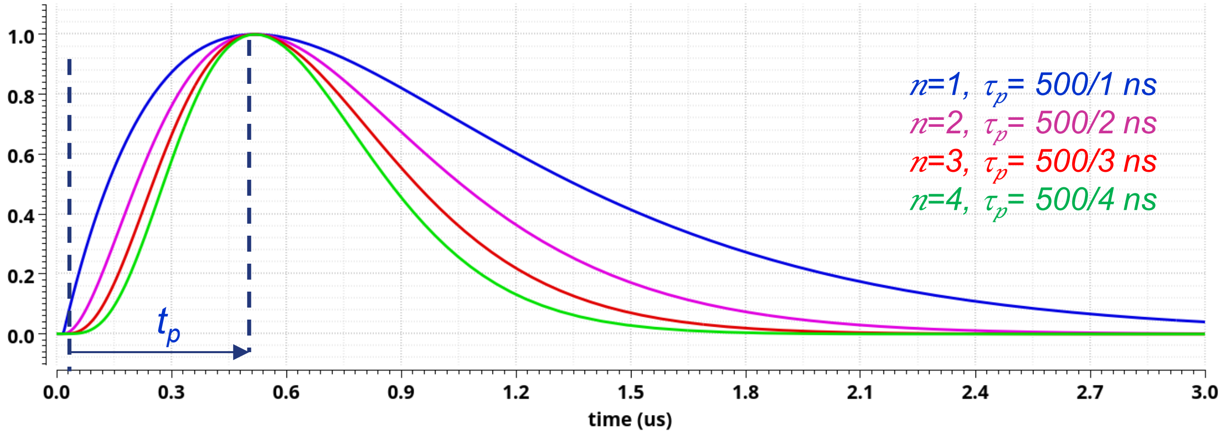
Impulse response assuming $\tau_f \rightarrow \infty$:

$$V_{oin}(s) \approx -\frac{Q_{in}}{C_f}\frac{\tau_p}{(1+s\tau_p)^{n+1}}\frac{C_z}{C_p}\left(-\frac{R_p}{R_c}\right)^{n-1} \xrightarrow{\mathcal{L}^{-1}} V_{oin}(t) \approx -\frac{Q_{in}}{C_f}\frac{1}{n!}\left(\frac{t}{\tau_p}\right)^n e^{-\frac{t}{\tau_p}} \cdot \frac{C_z}{C_p}\left(-\frac{R_p}{R_c}\right)^{n-1} u(t)$$

Shaper: Semi-Gaussian shapers (CR-RC^n)

- To keep the same peaking time, the time constants must be reduced

Normalized CR-RC^n shaper impulse response for a fixed peaking time t_p =500 ns



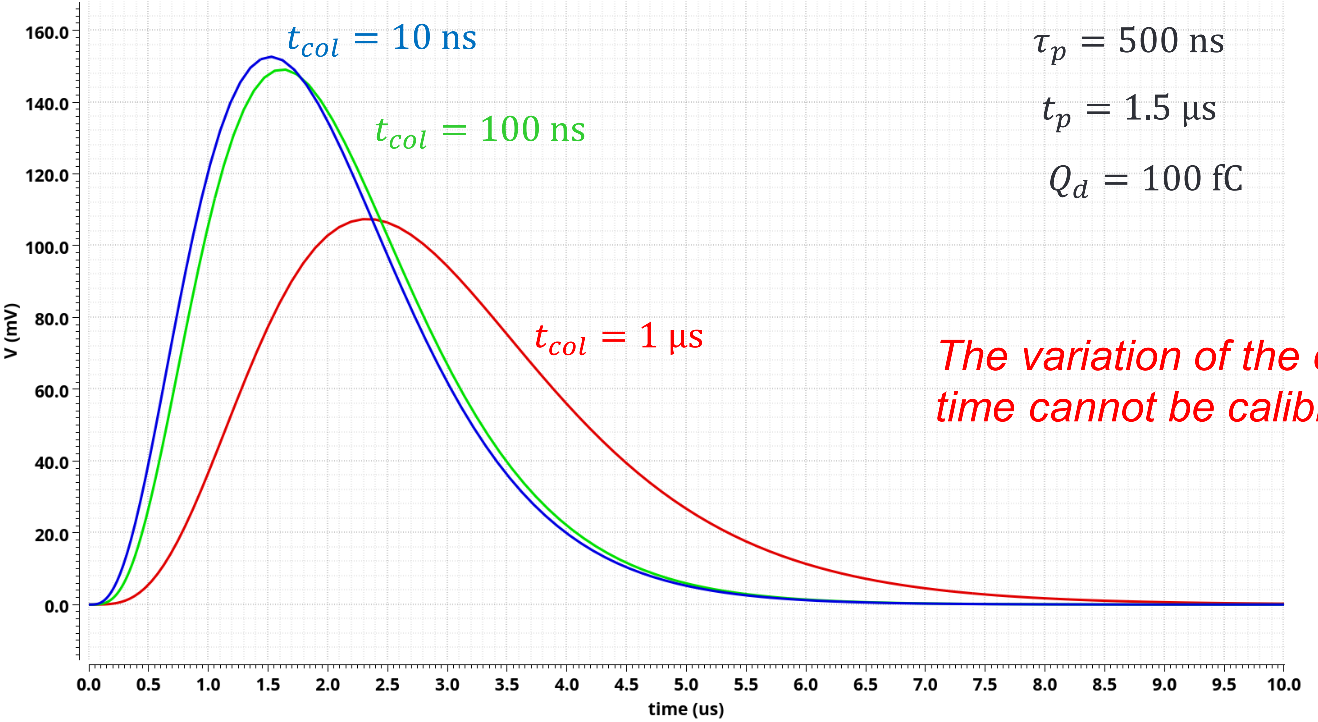
For a given peaking time t_p ***higher order*** *shapers provide shorter pulse duration:* ***lower pile-up*** *!*

Peak output voltage: $V_{oin,p} \approx -\dfrac{Q_{in}}{C_f}\, k\, e^{-n}$

$$\underbrace{k = \frac{C_z}{C_p}\left(-\frac{R_p}{R_c}\right)^n \frac{n^n}{n!}}_{\text{Usually normalized to 1}}$$

Shaper: ballistic deficit

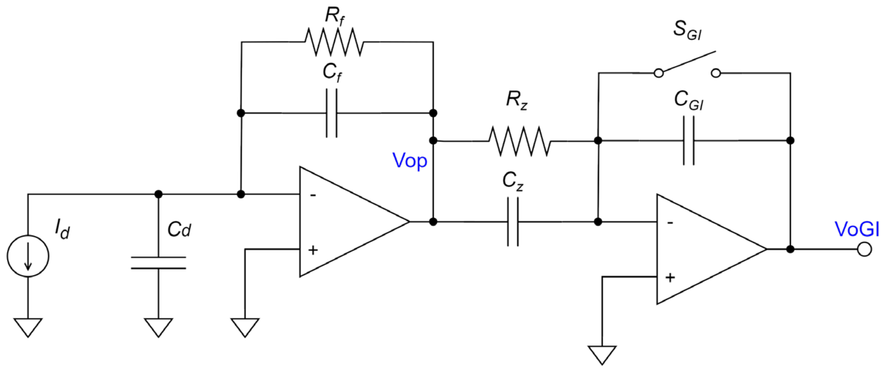
- Effect of finite collection time in CRRC[4] response
  - Attenuation when signal duration t_{col} is comparable to shaper peaking time t_p
  - Also peaking time t_p is delayed

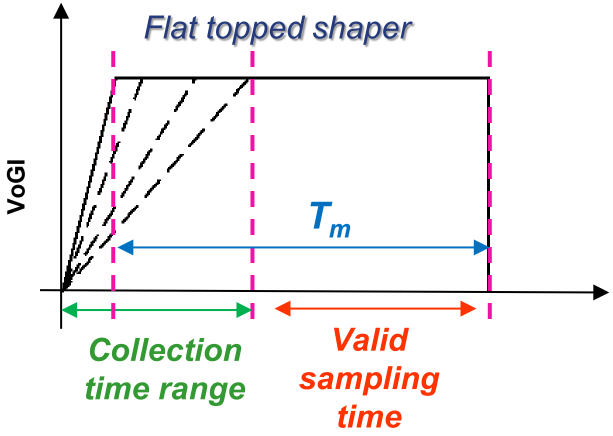


The variation of the collection time cannot be calibrated !!!

Shaper: time variant networks (Gated Integrator)

- **Gated integrator**: time variant network with flat top for ballistic deficit correction
  - The signal is integrated in C_{GI} during the measurement time T_m (while switch S_{GI} is open)
  - When the measurement is complete switch S_{GI} is closed to reset t C_{GI}
  - The measurement time T_m should be longer than the fluctuations of the collection time and provide some additional time to sample the output (S&H, ADC, etc)





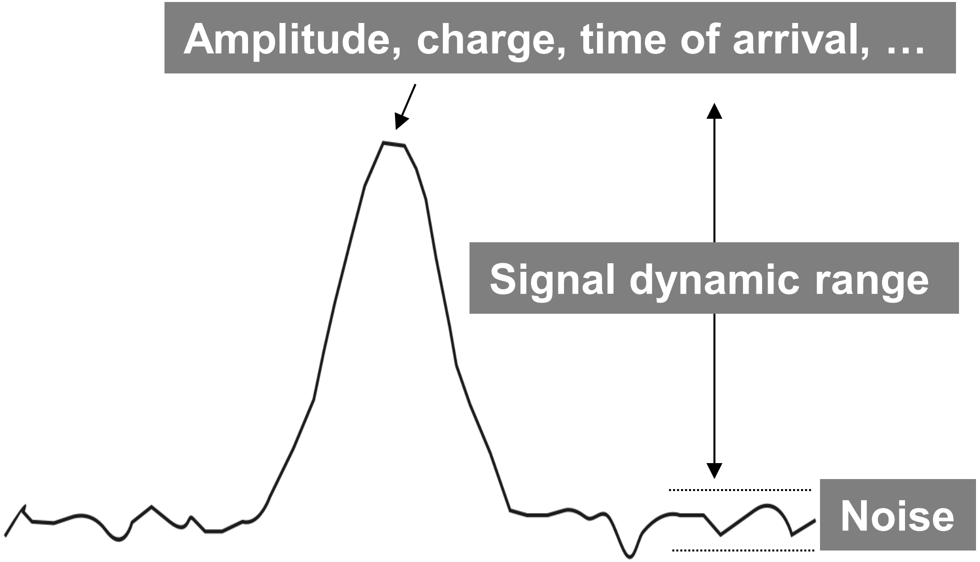
↗ D. Gascón et alt., "Noise Analysis of Time Variant Shapers in Frequency Domain," in IEEE Transactions on Nuclear Science, vol. 58, no. 1, pp. 177-186, Feb. 2011
↗ Spieler, Helmut , Semiconductor Detector Systems, Clarendon Press, Oxford, 2005.
↗ Rivetti, Angelo. CMOS : front-end electronics for radiation sensors . Boca Raton : CRC Press, 2015

Outlook

1. Detector electrical model
2. Input stage (preamplifier)
3. Shaper (filter)
4. **Noise analysis and optimization**
5. Applications: ASICs

Noise: Introduction

- Noise is a perturbation of the expected response of our system
- Two main different types
  - Fundamental noise (this lecture)
  - Interference from other systems (Electromagnetic Interference, EMI)
- Noise limits the resolution and dynamic range of the electronics systems



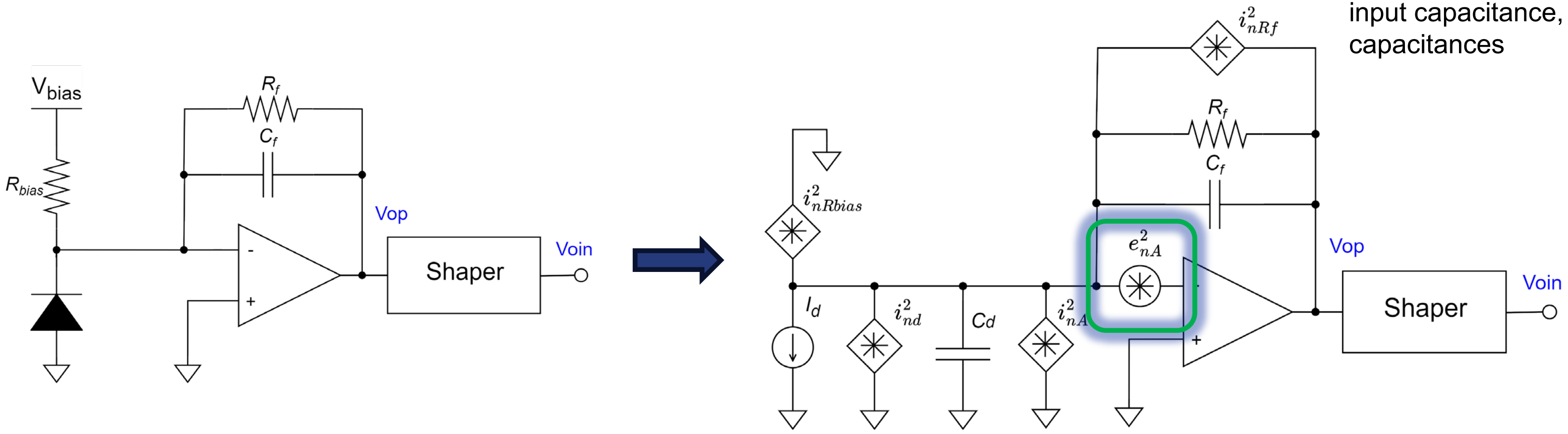
[See this video tutorial to review basic concepts on noise and how to deal with it in circuit analysis]

↗ R. Pallás-Areny and J. G. Webster, "Analog Signal Processing", John Wiley & Sons, New York, 1999
↗ C.D. Motchenbacher, "Low Noise Electronic System Design", Wiley Interscience, 1993
↗ H. W. Ott "Noise reduction techniques in electronic systems", Second Edition, John Wiley & Sons, New York, 1988

Noise: Shaping

- The main noise contributors in a well-designed FE are related to the input stage and the detector (remember Friis)

C_d also embeds amplifier's input capacitance, parasitic capacitances



Series noise contributors

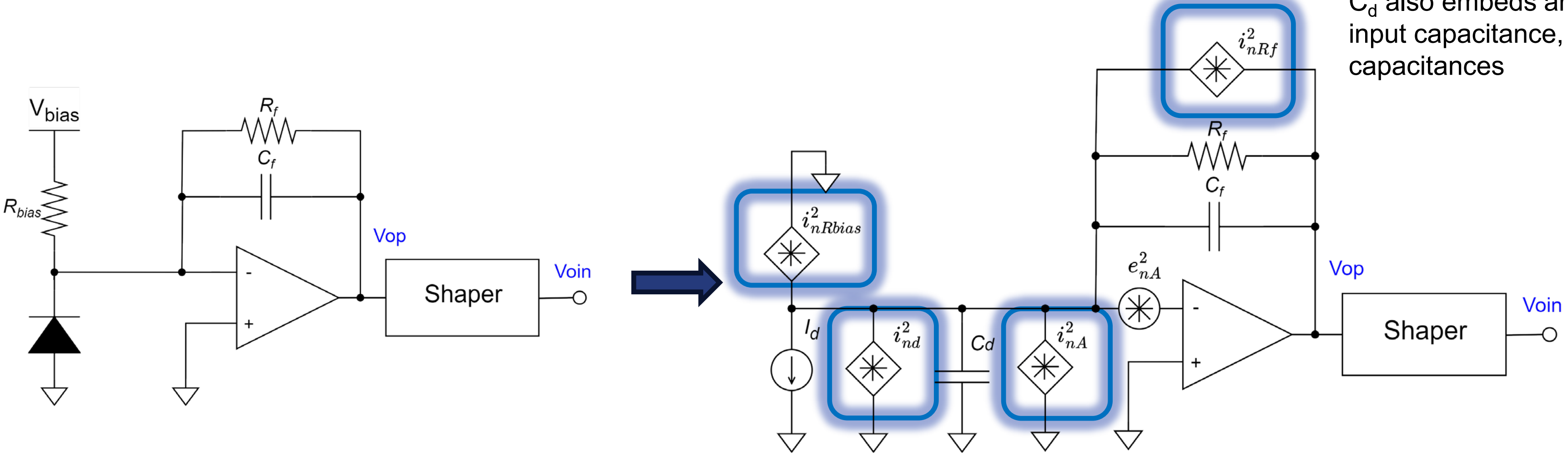
e^2_{nA}: Amplifier equivalent input referred voltage (series) noise. Includes:

1) Thermal white noise Power Spectral Density (PSD): e^2_{nwA}
2) Flicker (1/f) noise contributions PSD: $e^2_{nfA}(f) = \frac{A_f}{f}$

Noise: Shaping

- The main noise contributors in a well-designed FE are related to the input stage and the detector (remember Friis)

C_d also embeds amplifier's input capacitance, parasitic capacitances



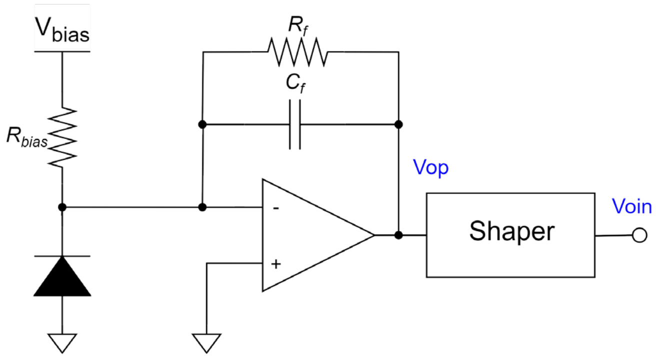
Parallel noise contributors PSD

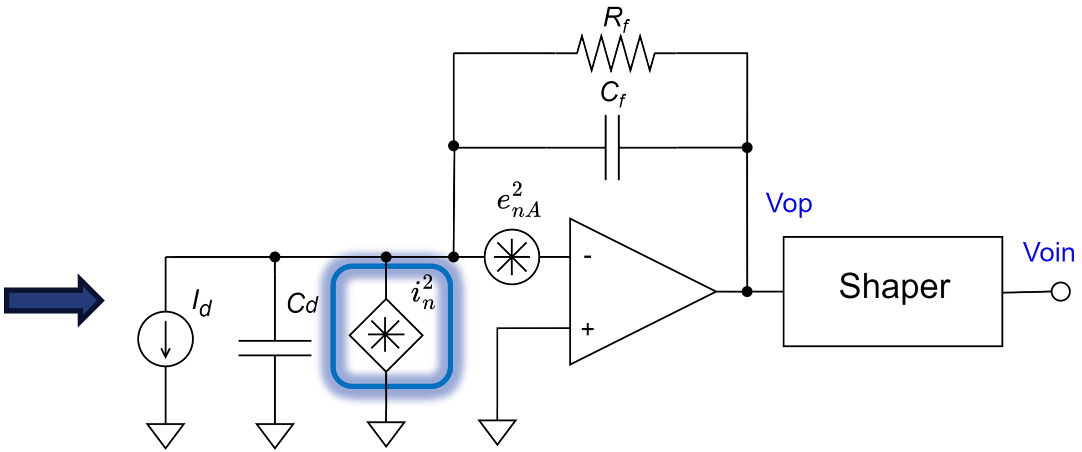
$$i^2_{nd} = 2qI_d \qquad i^2_{nRbias} = \frac{4kT}{R_{bias}} \qquad i^2_{nRf} = \frac{4kT}{R_f}$$

i^2_{nA}: Amplifier equivalent input referred current (parallel) noise

Noise: Shaping

- The main noise contributors in a well-designed FE are related to the input stage and the detector (remember Friis)





Parallel noise contributors can be combined in a single parallel noise generator

$$i_n^2 = i_{nd}^2 + i_{nRbias}^2 + i_{nRf}^2 + i_{nA}^2$$

Noise: Shaping

- To obtain the output noise σ_{no}^2 we will add (in quadrature) all the weighted noise contributions

- Output noise σ_{no}^2 variance for CR-RC shaper

$$\sigma_{no\_tot}^2 = e_{nwA}^2 \left(\frac{C_d}{C_f}\right)^2 \boxed{\frac{1}{8}} \frac{1}{t_p} + A_f \left(\frac{C_d}{C_f}\right)^2 \boxed{\frac{1}{2}} + i_{nwA}^2 \left(\frac{1}{C_f}\right)^2 \boxed{\frac{1}{8}} t_p$$

The value of the coefficients depend on shaping function (CR-RC here)

Noise: Shaping

- Equivalent Noise Charge (ENC)
  - Noise can be referred to the input and normalized to the signal generated by an input charge of 1 e-

$$ENC\ [e^{-}\ rms] = \frac{\sigma_{no\_tot}}{V_{o,p}(Q_{in} = q)}$$

$$V_{o,p}(Q_{in} = q) = \frac{q}{C_f}\,\frac{1}{e} \qquad \textit{Peak shaper output voltge for } q\ (1\ e^{-} \textit{charge})$$

$$\boldsymbol{ENC^2 = \frac{1}{q^2}\left(e_{nwA}^2 \boxed{\frac{e^2}{8}} C_d^2 \boxed{\frac{1}{t_p}} + A_f C_d^2 \boxed{\frac{e^2}{2}} + i_{nwA}^2 \boxed{\frac{e^2}{8}} \boxed{t_p}\right)}$$

$$ENC_S \propto 1\Big/\sqrt{t_p} \qquad ENC_F \cancel{\propto} t_p \qquad ENC_P \propto \sqrt{t_p}$$

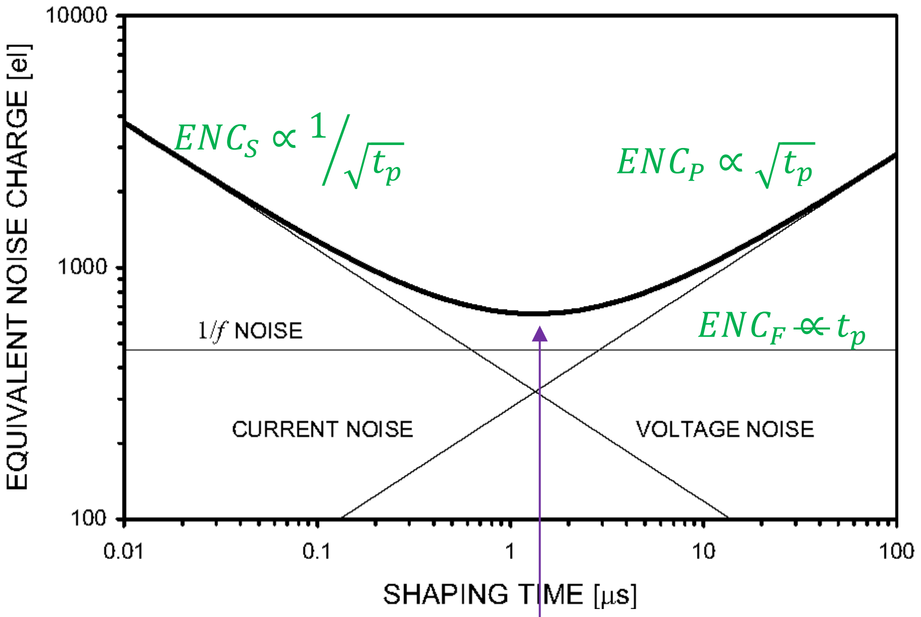
The coefficients depend on shaping function (CR-RC here)

↗ Rivetti, Angelo. CMOS : front-end electronics for radiation sensors . Boca Raton : CRC Press, 2015

Noise: Shaping

- We can provide a general expression for CR-RC^n shapers

$$\boldsymbol{ENC^2 = \frac{1}{q^2}\left(e_{nwA}^2 C_d^2 \frac{N_W}{t_p} + A_f C_d^2 N_f + i_{nwA}^2 N_i t_p\right)}$$



$$\frac{\partial ENC^2}{\partial t_p} = 0$$

$$\Downarrow$$

$$t_{p,opt} = \frac{e_{nwA}}{i_{nwA}} C_d \sqrt{\frac{N_w}{N_i}}$$

Optimal peaking time $t_{p,opt}$

↗ V. Radeka, Low-Noise Techniques in Detectors;
Ann. Rev. Nucl. Part. Sci. 38 (1988) 217

Noise: Shaping

- We can provide a general expression for CR-RC^n shapers

$$ENC^2 = \frac{1}{q^2}\left(e_{nwA}^2 C_d^2 \frac{N_W}{t_p} + A_f C_d^2 N_f + i_{nwA}^2 N_i t_p\right)$$

Noise indexes of CR-RC^n shapers. n indicates the number of integrators in the chain.

	n=1	n=2	n=3	n=4	n=5	n=6	n=7
N_w	0.92	0.85	0.93	1.02	1.10	1.19	1.27
N_f	3.69	3.42	3.32	3.27	3.25	3.23	3.22
N_i	0.92	0.63	0.52	0.45	0.40	0.36	0.34

Noise indexes of shapers with complex conjugate poles. n is the total number of poles in the transfer function.

	n=2	n=3	n=4	n=5	n=6	n=7
N_w	0.93	0.85	0.91	0.96	1.10	1.04
N_f	3.70	3.39	3.33	3.27	3.26	3.21
N_i	0.88	0.61	0.51	0.45	0.42	0.40

$$t_{p,opt} = \frac{e_{nwA}}{i_{nwA}} C_d \sqrt{\frac{N_w}{N_i}}$$

High order shapers are advantageous when long t_p are required because:

1) Smaller N_i

2) Faster return to the baseline (for a fixed t_p)

↗ Rivetti, Angelo. CMOS : front-end electronics for radiation sensors . Boca Raton : CRC Press, 2015

Outlook

1. Detector electrical model
2. Input stage (preamplifier)
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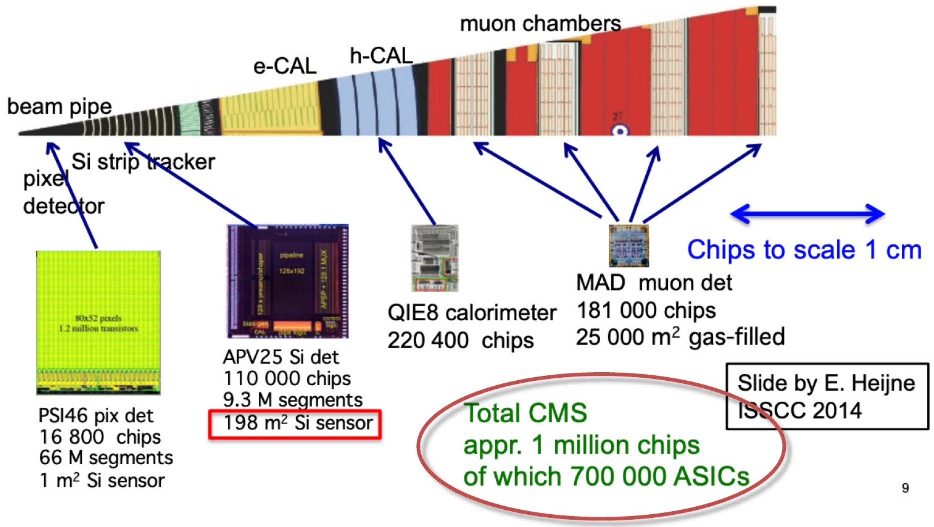
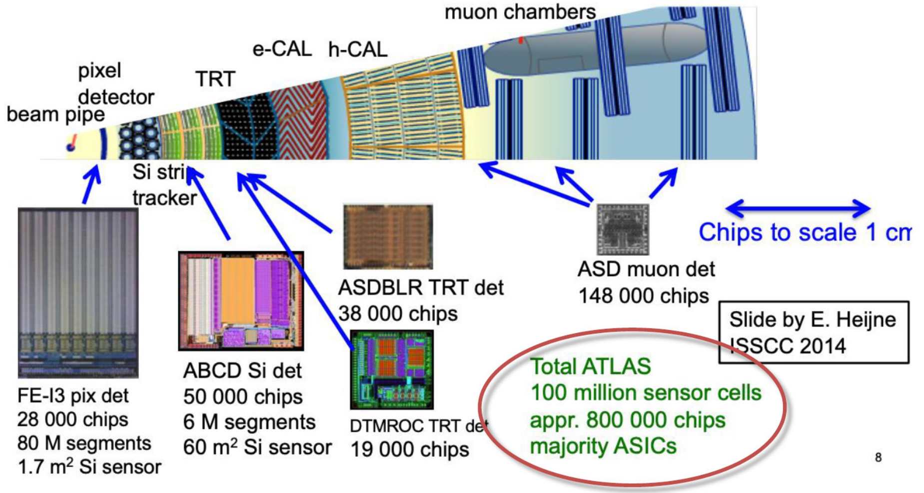
4. Noise analysis and optimization
5. **Applications: ASICs**

ASICs

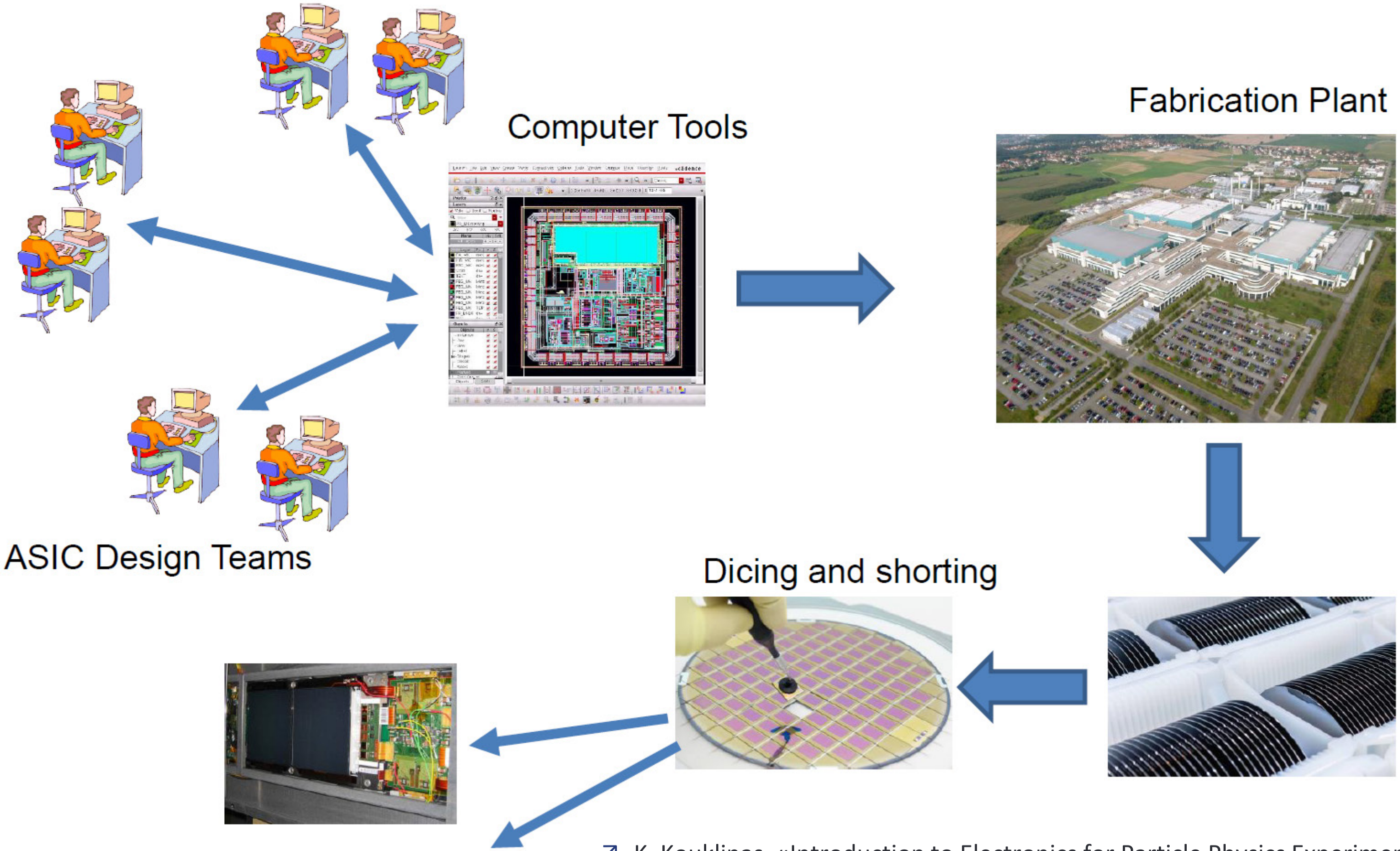
- Modern experiments in nuclear, particle and astro-particle physics require:
  - Large number of channels
  - Low noise and high speed processing
  - High channel density
  - Tailored processing: analog, mixed and digital

Application Specific Integrated Circuits ASICs



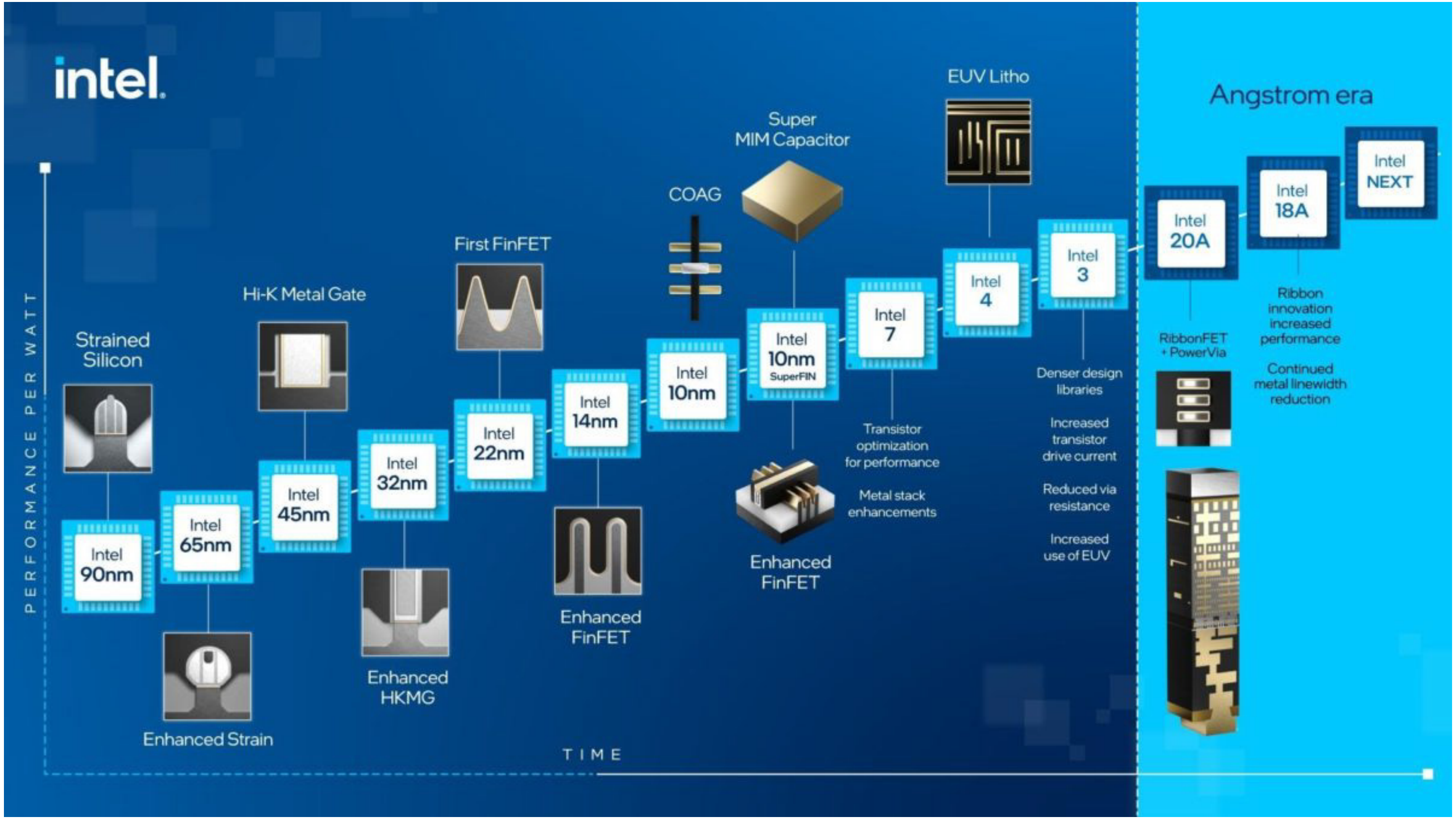


ASICs



↗ K. Kouklinas, «Introduction to Electronics for Particle Physics Experiments », 2023

ASICs

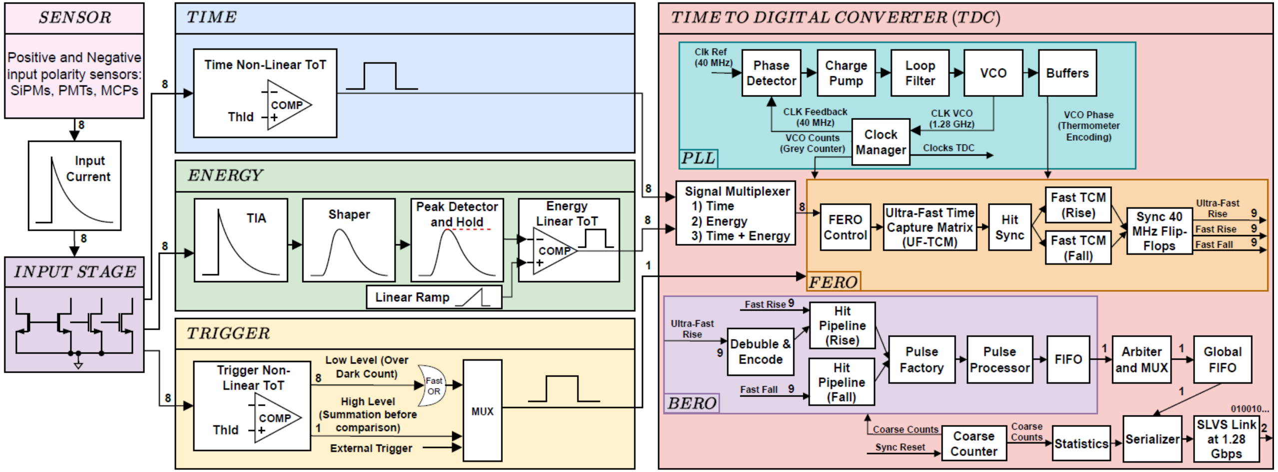


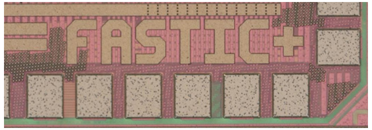
↗ https://semiwiki.com/semiconductor-manufacturers/intel/309633-euv-and-intel-foundry-services/

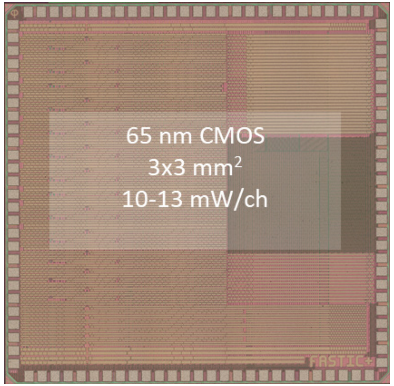
ASICs: planar integration

- Connected to many different types of sensors: silicon or gaseous sensors for tracking, photosensors for calorimetry or particle ID, etc

Example of ASIC for Fast Timing Applications with sensors that present intrinsic gain (SiPM, PMT, MCP, etc)







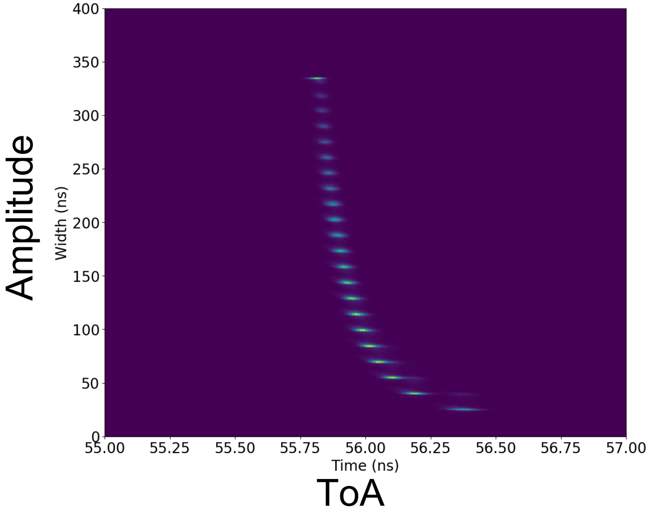
↗ Mazzanti, D. Gascon, D. et al « Below 100 ps CTR using FastIC+, an ASIC including on-chip digitization for ToF-PET and beyond ». IEEE Transactions on Radiation and Plasma Medical Sciences. PP. 1-1. 10.1109/TRPMS.2026.3678341.

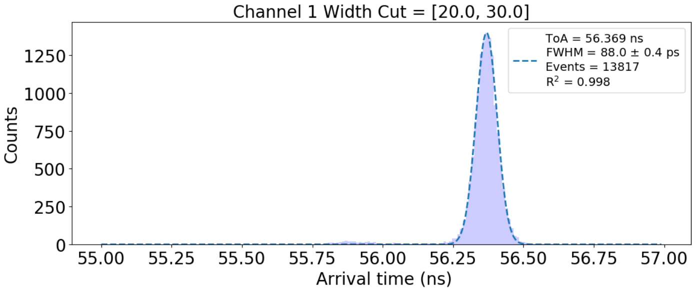
ASICs: planar integration

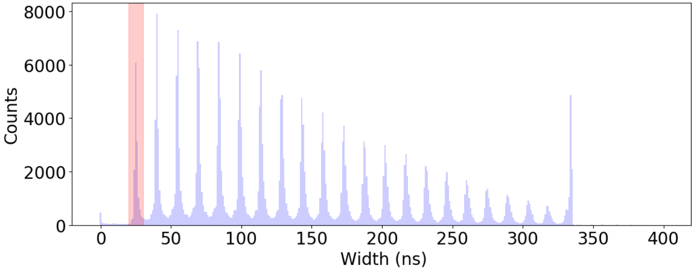
- Connected to many different types of sensors: silicon or gaseous sensors for tracking, photosensors for calorimetry or particle ID, etc

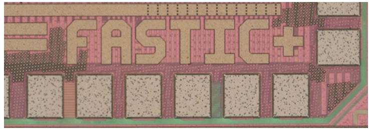
Example of ASIC for Fast Timing Applications with sensors that present intrinsic gain (SiPM, PMT, MCP, etc)

Single photon identification
with FBK-NUV 3x3 mm² SiPM









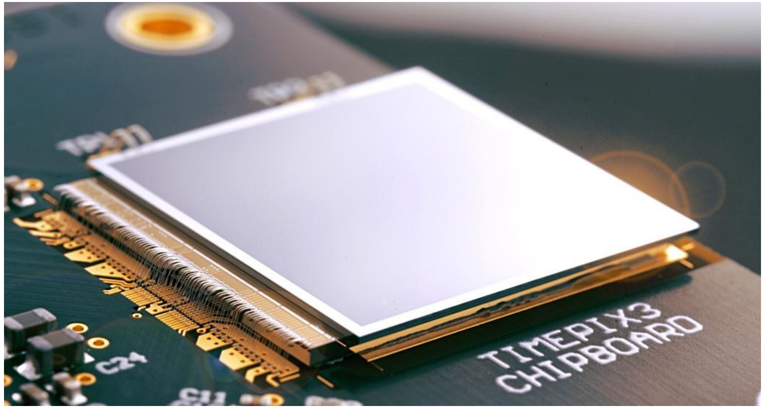
Single Photon Time Resolution

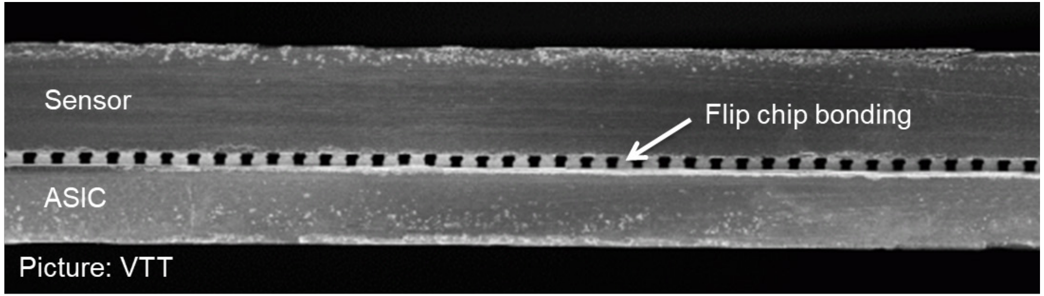
< 100 ps FWHM

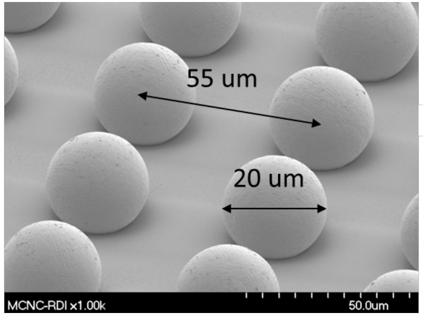
< 50 ps rms

↗ Mazzanti, D. Gascon, D. et al « Below 100 ps CTR using FastIC+, an ASIC including on-chip digitization for ToF-PET and beyond ». IEEE Transactions on Radiation and Plasma Medical Sciences. PP. 1-1. 10.1109/TRPMS.2026.3678341.

ASICs: vertical integration





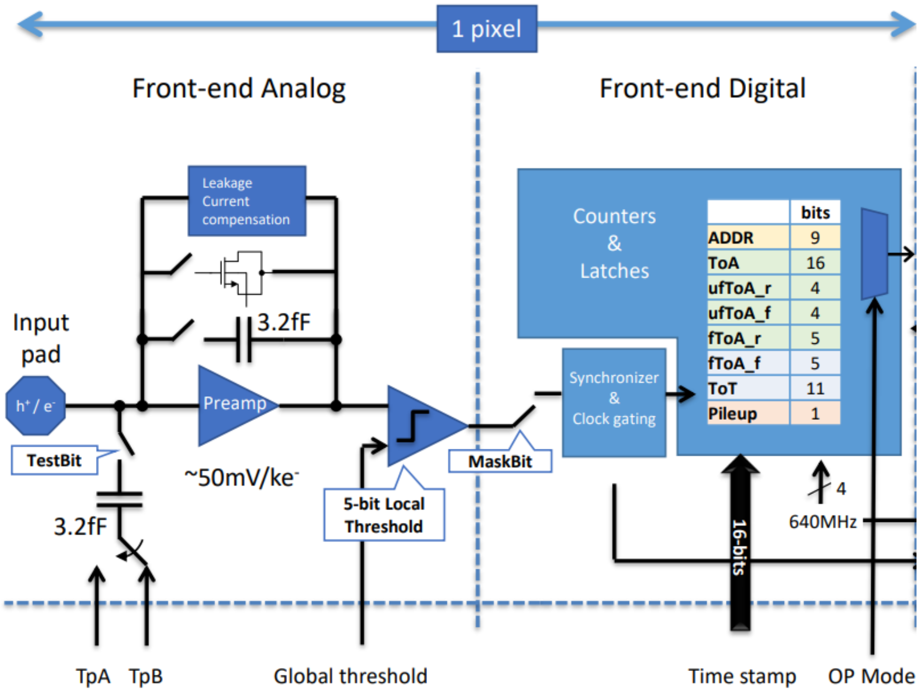


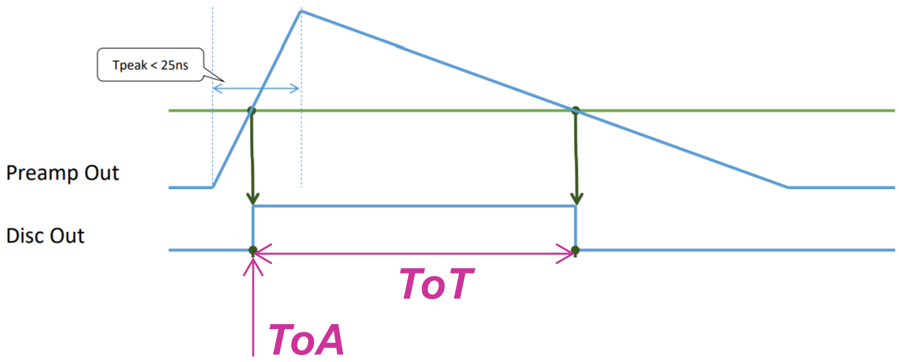
- The sensor and the readout electronics are connected using fine pitch flip chip technology
- Flip chip process consist of (1.) sensor and readout ASIC wafer bumping, (2.) dicing of wafers and (3.) flip chip bonding of good known dies.

↗ R. Ballabriga "Signal Processing and Front End Electronics". Short Course EDIT School, 2026

ASICs: vertical integration

- Hybrid Pixel Detectors: Medipix/Timepix/Velopix chips family (CERN)
  - Based on CSA + discriminator + counter or TDC



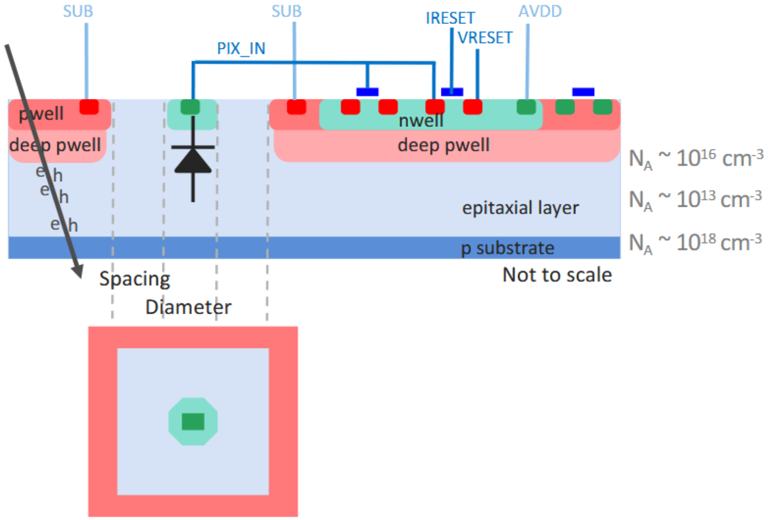


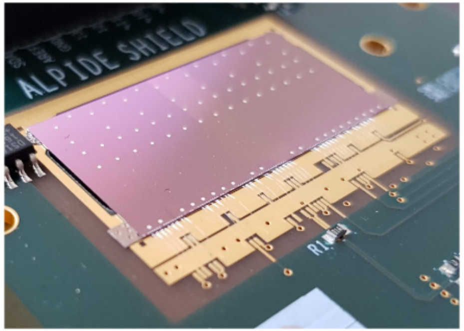
- CSP with linear discharge
- Energy by Time over Threshold (ToT)
  - Linear in limited dynamic range
- Time by Time of Arrival (ToA)
  - On-chip TDC with 200 ps time bin
- Segmentation allows high rate despite “slow” CSP based pixel
  - Max rate (data driven)> 3 Mhits/mm^2/s

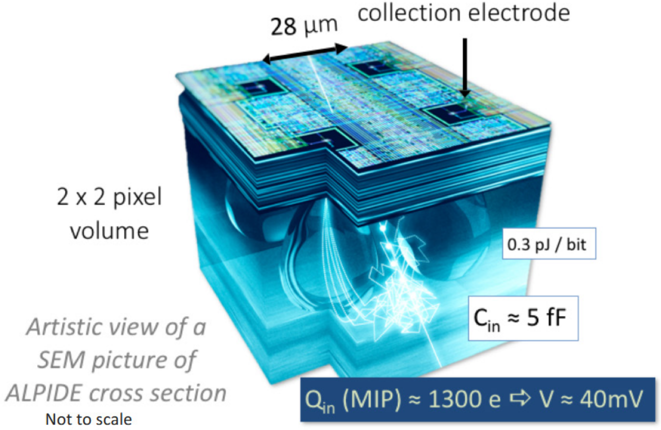
↗ X. Llopart “Timepix4, a pixel detector readout chip with sub-200 ps timestamp binning”. CERN Detector Seminar, February 2022

ASICs: monolithics

- Sensor and FE implemented in the same substrate







↗ https://indico.cern.ch/event/863068/contributions/3752479/attachments/1996261/3330551/20200302-Aglieri-ALPIDE-Overview.pdf

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CPAN NETWORK
ON INSTRUMENTATION
AND DETECTORS

Ayudas ED2024-153684-T,
CEX2023-001292-S financiadas por

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AGENCIA
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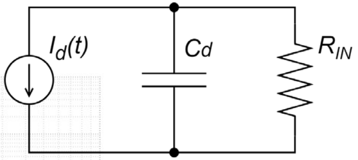
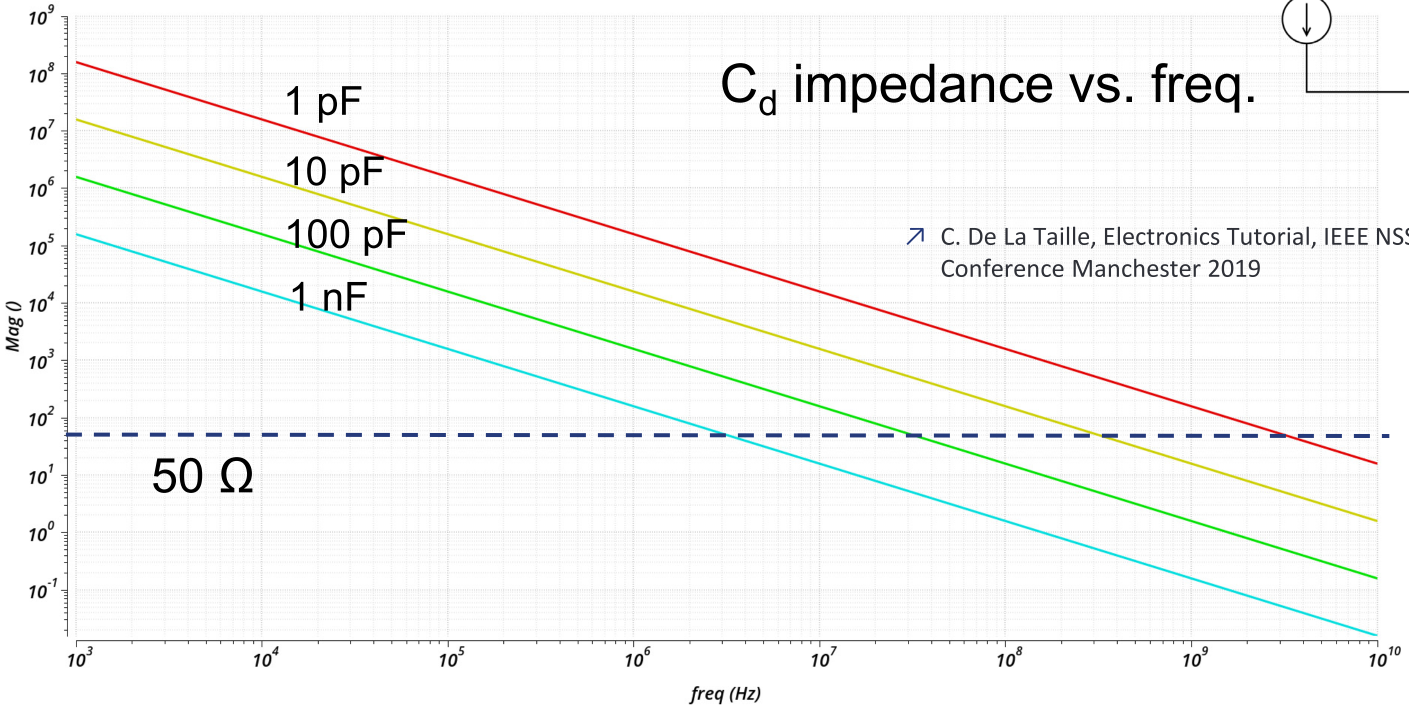
Input stage: input impedance

- C_{DET} and R_L form a current divider
- Low R_{IN} better for I sensing
  - Difficult at HF for large sensors: C_d is hard to beat !!!

$$I_{R_{IN}}/I_d \approx \frac{Z_d}{Z_d + R_{IN}} \approx \frac{\frac{1}{sC_d}}{\frac{1}{sC_d} + R_{IN}}$$

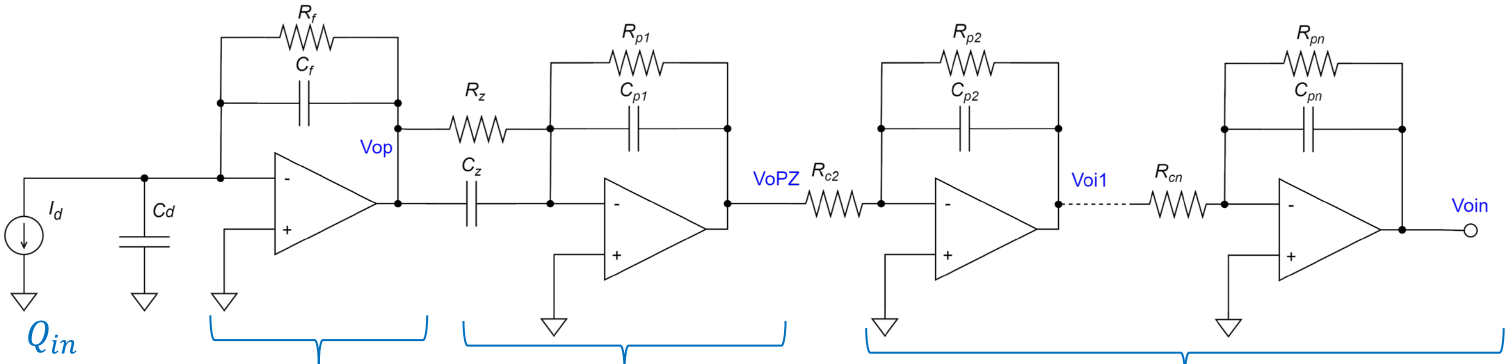




↗ C. De La Taille, Electronics Tutorial, IEEE NSS-MIC Conference Manchester 2019

Shaper: Pole-Zero cancellation

- Circuit is modified to shift the position of the zero introduced by C_z from 0 to $^1/_{\tau_f}$
  - The $^1/_{\tau_f}$ pole is cancelled with $^1/_{\tau_z}$ zero



$$\frac{R_f}{s\tau_f + 1} \qquad -\frac{R_p}{R_z}\frac{s\tau_z+1}{s\tau_p+1} \qquad \left(-\frac{R_p}{R_c}\right)^n \frac{1}{(s\tau_p+1)^n}$$

$\tau_p = \tau_{pi}$

$R_p = R_{pi}$

$C_p = C_{pi}$

$R_c = R_{ci}$

$\tau_z = C_z R_z$

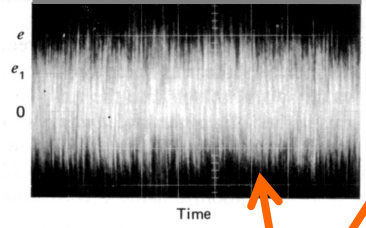
$\tau_p = C_p R_p$

$$V_{oin}(s) \approx -Q_{in}\frac{R_f R_p}{R_z}\frac{1}{\tau_p^n}\frac{1}{\left(\frac{1}{\tau_p}+s\right)^{n+1}}\left(-\frac{R_p}{R_c}\right)^n \xrightarrow{\mathcal{L}^{-1}} V_{oin}(t) \approx -Q_{in}\frac{R_f R_p}{R_z \tau_p}\frac{1}{n!}\left(-\frac{R_p}{R_c}\frac{t}{\tau_p}\right)^n e^{-\frac{t}{\tau_p}}u(t)$$

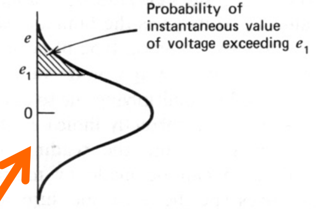
Noise: Introduction

- Noise signal x(t), like any random signal, can be described in different domains:
  - Amplitude (Probability Density Function, PDF)
  - Time
  - Frequency (Power Spectral Density, PSD)
- The time-varying part of the mean-square value is the signal variance
  - Defined as the mean-square of x(t) about is mean value
- Power dissipated by a random voltage on a resistor is proportional to mean-square voltage
  - The signal intensity σ^2 is also termed signal power.
    - Its units are squared V: V^2 (or I^2)
    - Standard deviation σ unit is V rms (or I rms)

Time representation of x(t)



PDF: Amplitude distribution of x(t)



Variance of X(t)

$$\sigma_x^2 = \lim_{T\to\infty} \frac{1}{T} \int_0^T [x(t) - \mu_x]^2 dt$$

$$\Psi_x^2 = \lim_{T\to\infty} \frac{1}{T} \int_0^T x(t)^2 dt \qquad \psi_x^2 = \mu_x^2 + \sigma_X^2$$

↗ R. Pallás-Areny and J. G. Webster, "Analog Signal Processing", John Wiley & Sons, New York, 1999
↗ C.D. Motchenbacher, "Low Noise Electronic System Design", Wiley Interscience, 1993
↗ H. W. Ott "Noise reduction techniques in electronic systems", Second Edition, John Wiley & Sons, New York, 1988

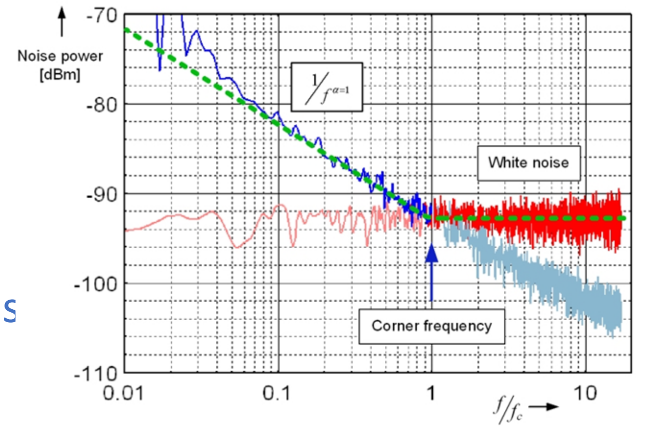
Noise: Introduction

- The PDF does not take into account the time when the different amplitudes of random signal appear.
  - Very different amplitude sequences (or waveforms) can lead to the same Gaussian distribution.
- Better characterization: consider the distribution of its power in different frequency bands.
  - The power spectral density (PSD) of a random signal x(t) is

$$G_{xx}(f) \doteq \lim_{\Delta f \to 0} \frac{\Psi_x^2(f, \Delta f)}{\Delta f} = \lim_{\Delta f \to 0} \frac{1}{\Delta f}\left[\lim_{T \to \infty} \frac{1}{T} \int_0^T x^2(t, f, \Delta f)\, dt\right]$$

  - White spectrum: same power density at all frequencies (flat spectrum)
  - The power of a signal can be obtained by integrating its PSD over the entire frequency range

Frequency representation (PSD)



$$\boldsymbol{\sigma_x^2 = \int_0^\infty G_{xx}(f)\, df}$$

Noise: Introduction

- Response of a linear system to a deterministic signal:
  - Fourier transform of the input signal: X(f) $Y(f) = H(f)\,X(f) \quad Units\ [V\ or\ A]$
  - Fourier transform of the output signal: Y(f)
  - Fourier (Laplace) transform of the impulse response: H(f)
- Response of a linear system to noise :
  - Input PSD: $G_{xx}(f)$
  - Output PSD: $G_{yy}(f)$
  - Fourier (Laplace) transform of the impulse response: H(f)

$$G_{yy}(f) = |H(f)|^2 G_{xx}(f) \quad Units{:}\ \left[{}^{V^2}/_{Hz}\ or\ {}^{A^2}/_{Hz}\right]$$

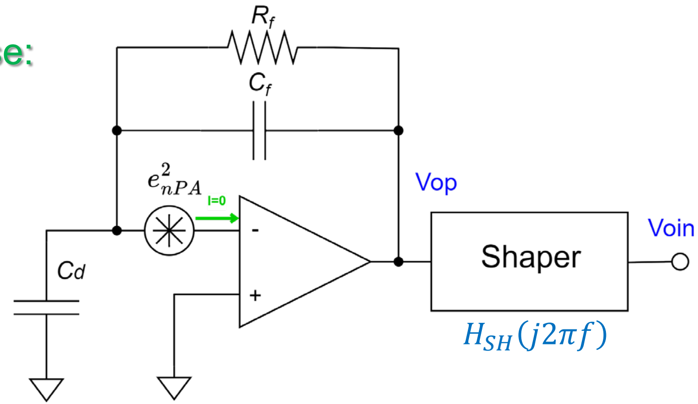
- The output integrated noise $\sigma_y^2 = \int_0^\infty G_{yy}(f)\,df \quad Units\ [V^2 or\ A^2]$

↗ R. Pallás-Areny and J. G. Webster, “Analog Signal Processing”, John Wiley & Sons, New York, 1999
↗ C.D. Motchenbacher, “Low Noise Electronic System Design”, Wiley Interscience, 1993
↗ H. W. Ott “Noise reduction techniques in electronic systems”, Second Edition, John Wiley & Sons, New York, 1988

Noise: Shaping

- To obtain the output noise σ_{no}^2 we will:
  1) **Obtain transfer function $H_{x_n}(j2\pi f)$ for any noise source x_n to the output (V_{oin})**
  2) Calculate output PSD $e_{no\_x_n}^2$ due to a noise source x_n by multiplying its PSD by $\left|H_{x_n}(j2\pi f)\right|^2$
  3) Integrate it to obtain the variance of the output signal $\sigma_{no\_x_n}^2$ due to noise source x_n
  4) Add in quadrature (assuming no correlation) to obtain the total integrated noise variance $\sigma_{no\_tot}^2$

For the series noise:



$$H_{e_n}(j2\pi f) = \left(j2\pi f C_d\, Z_f(j2\pi f) + 1\right) H_{SH}(j2\pi f) \approx \left(\frac{C_d}{C_f} + 1\right) H_{SH}(j2\pi f) \approx \frac{C_d}{C_f} H_{SH}(j2\pi f)$$

For a CSA

Usually $C_d >> C_f$

Noise: Shaping

- To obtain the output noise σ_{no}^2 we will:
  1) **Obtain transfer function $H_{x_n}(j2\pi f)$ for any noise source x_n to the output (V_{oin})**
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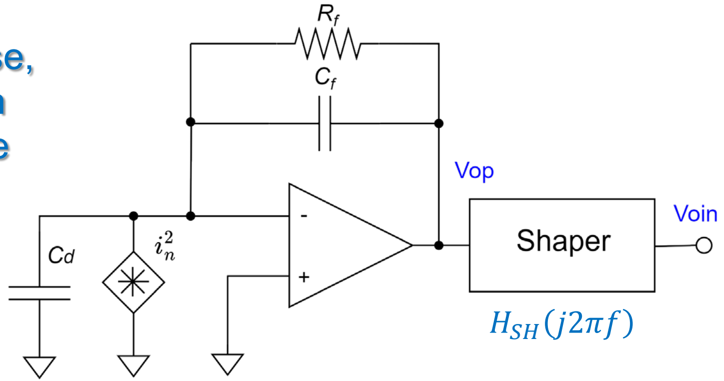
  4) Add in quadrature (assuming no correlation) to obtain the total integrated noise variance $\sigma_{no\_tot}^2$

For the parallel noise, the transfer function $H_{i_n}(f)$ is exactly the same as for the detector signal



$$H_{i_n}(j2\pi f) = Z_f(j2\pi f)H_{SH}(j2\pi f) \approx \frac{1}{j2\pi f C_f} H_{SH}(j2\pi f)$$

For a CSA

Noise: Shaping

- To obtain the output noise σ_{no}^2 we will:
  1) Obtain transfer function $H_{x_n}(j2\pi f)$ for any noise source x_n to the output (V_{oin})
  2) **Calculate output PSD $e_{no\_x_n}^2$ due to a noise source x_n by multiplying its PSD by $\left|H_{x_n}(j2\pi f)\right|^2$**
  3) Integrate it to obtain the variance of the output signal $\sigma_{no\_x_n}^2$ due to noise source x_n
  4) Add in quadrature (assuming no correlation) to obtain the total integrated noise variance $\sigma_{no\_tot}^2$

Series white noise $\longrightarrow$ $e_{no\_en\_wA}^2 = e_{nwA}^2 \left|H_{e_n}(j2\pi f)\right|^2 = e_{nwA}^2 \left(\frac{C_d}{C_f}\right)^2 |H_{SH}(j2\pi f)|^2$

Series flicker noise $\longrightarrow$ $e_{no\_en\_fA}^2 = e_{nfA}^2 \left|H_{e_n}(j2\pi f)\right|^2 = \frac{A_f}{f}\left(\frac{C_d}{C_f}\right)^2 |H_{SH}(j2\pi f)|^2$

Parallel white noise $\longrightarrow$ $e_{no\_in\_wA}^2 = i_{nwA}^2 \left|H_{i_n}(j2\pi f)\right|^2 = i_{nwA}^2 \left(\frac{1}{2\pi f C_f}\right)^2 |H_{SH}(j2\pi f)|^2$

Series and parallel white noise PSD contributions are equal at $\tau_c = \frac{1}{2\pi f_c} = \frac{e_{nwA}}{i_{nwA}} C_d$

Noise: Shaping

- To obtain the output noise σ^2_{no} we will:
  1) Obtain transfer function $H_{x_n}(j2\pi f)$ for any noise source x_n to the output (V_{oin})
  2) Calculate output PSD $e^2_{no\_x_n}$ due to a noise source x_n by multiplying its PSD by $\left|H_{x_n}(j2\pi f)\right|^2$
  3) **Integrate it to obtain the variance of the output signal $\sigma^2_{no\_x_n}$ due to noise source x_n**
  4) Add in quadrature (assuming no correlation) to obtain the total integrated noise variance $\sigma^2_{no\_tot}$

Series white noise $\longrightarrow$ $\sigma^2_{no\_en\_wA} = \int_0^\infty e^2_{nwA}\left|H_{e_n}(j2\pi f)\right|^2 df = e^2_{nwA}\left(\frac{C_d}{C_f}\right)^2 \boxed{\int_0^\infty |H_{SH}(j2\pi f)|^2 df}$

Series flicker noise $\longrightarrow$ $\sigma^2_{no\_en\_fA} = \int_0^\infty e^2_{nfA}\left|H_{e_n}(j2\pi f)\right|^2 df = A_f\left(\frac{C_d}{C_f}\right)^2 \boxed{\int_0^\infty \frac{1}{f}|H_{SH}(j2\pi f)|^2 df}$

Parallel white noise $\longrightarrow$ $\sigma^2_{no\_in\_wA} = \int_0^\infty i^2_{nwA}\left|H_{i_n}(j2\pi f)\right|^2 df = i^2_{nwA}\left(\frac{1}{C_f}\right)^2 \boxed{\int_0^\infty \left(\frac{1}{2\pi f}\right)^2 |H_{SH}(j2\pi f)|^2 df}$

The results of the integrals depend only on the parameters that define the shaper: function and peaking time t_p

Noise: Shaping

▶ To obtain the output noise σ_{no}^2 we will:

1) Obtain transfer function $H_{x_n}(j2\pi f)$ for any noise source x_n to the output (V_{oin})
2) Calculate output PSD $e_{no\_x_n}^2$ due to a noise source x_n by multiplying its PSD by $\left|H_{x_n}(j2\pi f)\right|^2$
3) **Integrate it to obtain the variance of the output signal $\boldsymbol{\sigma_{no\_x_n}^2}$ due to noise source $\boldsymbol{x_n}$**
4) Add in quadrature (assuming no correlation) to obtain the total integrated noise variance $\sigma_{no\_tot}^2$

Series white noise ⟶ $$\sigma_{no\_en\_wA}^2 = e_{nwA}^2\left(\frac{C_d}{C_f}\right)^2\int_0^\infty |H_{SH}(j2\pi f)|^2 df = e_{nwA}^2\left(\frac{C_d}{C_f}\right)^2 \boxed{\frac{1}{8}\frac{1}{t_p}}$$

Series flicker noise ⟶ $$\sigma_{no\_en\_fA}^2 = A_f\left(\frac{C_d}{C_f}\right)^2\int_0^\infty \frac{1}{f}|H_{SH}(j2\pi f)|^2 df = A_f\left(\frac{C_d}{C_f}\right)^2 \boxed{\frac{1}{2}}$$

Parallel white noise ⟶ $$\sigma_{no\_in\_wA}^2 = i_{nwA}^2\left(\frac{1}{C_f}\right)^2\int_0^\infty \left(\frac{1}{2\pi f}\right)^2 |H_{SH}(j2\pi f)|^2 df = i_{nwA}^2\left(\frac{1}{C_f}\right)^2 \boxed{\frac{t_p}{8}}$$

For a CR-RC shaper

↗ Rivetti, Angelo. CMOS : front-end electronics for radiation sensors . Boca Raton : CRC Press, 2015

Noise: Shaping

- To obtain the output noise σ_{no}^2 we will:
  1) Obtain transfer function $H_{x_n}(j2\pi f)$ for any noise source x_n to the output (V_{oin})
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  4) **Add in quadrature (assuming no correlation) to obtain the total integrated noise variance $\sigma_{no\_tot}^2$**

$$\sigma_{no\_tot}^2 = e_{nwA}^2\left(\frac{C_d}{C_f}\right)^2\frac{1}{8}\frac{1}{t_p} + A_f\left(\frac{C_d}{C_f}\right)^2\frac{1}{2} + i_{nwA}^2\left(\frac{1}{C_f}\right)^2\frac{t_p}{8}$$

For a CR-RC shaper

Equivalent Noise Charge (ENC)

Noise can be referred to the input and normalized to the signal generated by an input charge of 1 e-

$$\text{ENC [e− rms]} = \frac{\sigma_{no\_tot}}{V_{o,p}(Q_{in} = q)}$$

$$V_{o,p}(Q_{in} = q):\ Peak\ shaper\ output\ voltge\ for\ q\ (1\ e^{-} charge)$$

↗ Rivetti, Angelo. CMOS : front-end electronics for radiation sensors . Boca Raton : CRC Press, 2015

Noise: Shaping

- To obtain the output noise σ_{no}^2 we will:
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$$\sigma_{no\_tot}^2 = e_{nwA}^2 \left(\frac{C_d}{C_f}\right)^2 \frac{1}{8}\frac{1}{t_p} + A_f \left(\frac{C_d}{C_f}\right)^2 \frac{1}{2} + i_{nwA}^2 \left(\frac{1}{C_f}\right)^2 \frac{t_p}{8}$$

$$V_{oi,p} = \frac{q}{C_f}\frac{1}{e}$$

For a CR-RC shaper

$$\boldsymbol{ENC^2 = \frac{1}{q^2}\left(e_{nwA}^2 \boxed{\frac{e^2}{8}} C_d^2 \boxed{\frac{1}{t_p}} + A_f C_d^2 \boxed{\frac{e^2}{2}} + i_{nwA}^2 \boxed{\frac{e^2}{8}} \boxed{t_p}\right)}$$

$ENC_S \propto 1/\sqrt{t_p}$ $\quad ENC_F \cancel{\propto} t_p$ $\quad ENC_P \propto \sqrt{t_p}$

The coefficients depend on shaping function (CR-RC here)

↗ Rivetti, Angelo. CMOS : front-end electronics for radiation sensors . Boca Raton : CRC Press, 2015

Noise: Shaping

- The impulse response of the optimal shaper is $h_{opt}(t) = e^{-\frac{|t|}{\tau_C}}$
- It is called the "cusp" or as the "matched filter”
  - Shapers with triangular or trapezoidal shape are a good approximation
  - High order semi-gaussian shapers are a good compromise (ENC versus complexity)

Shaper	$t_{p,opt}$	$ENC_{tp,opt}/ENC_{CUSP}$
Cusp	∞	1
Triangular	$\sqrt{3}\,\tau_c$	1.07
Trapezoidal	$\sqrt{3}\,\tau_c$	$1.07\left[1+0.43\frac{T_{FT}}{\tau_c}\right]$
Semi-Gaussian 4th order (CR-(RC)^4)	$1.5\,\tau_c$	1.16
CR-RC	τ_c	1.36
Bipolar SG ((CR)^2-(RC)^4)	$0.74\,\tau_c$	1.38
Bipolar: (CR)^2-RC		1.41

↗ G. F. Knoll, “Radiation Detection and Measurement”, Third Edition, John Wiley & Sons, New York, 2000.
↗ E. Gatti and P. F. Manfredi, “Processing the Signals from Solid-State Detectors in Elementary-Particle Physics”, Rivista del Nuovo Cimento, Vol. 9, no. 1, 1985.
↗ V. Radeka, “Low-noise techniques in detectors”, Ann. Rev. Nucl. Part. Sci, 38: 217-77, 1988.
↗ D. Gascón et alt., "Noise Analysis of Time Variant Shapers in Frequency Domain," in IEEE Transactions on Nuclear Science, vol. 58, no. 1, pp. 177-186, 2011

Appendix

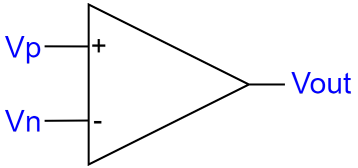
Discriminators

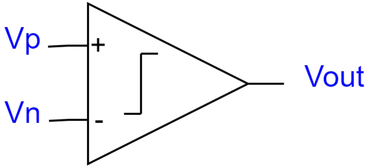
Discriminators: principle

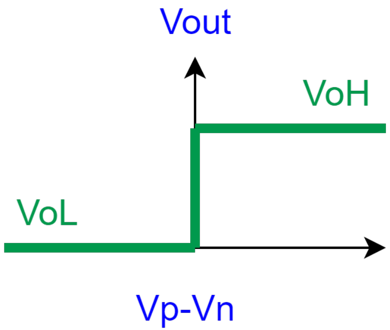
- A decision-making circuit
- Compares the magnitude two input analog signals: Vp and Vn
- Provides two possible output signal Vout levels…
  - VoH (logic 1) or VoL (logic 0)
- …depending on whether Vp is greater or smaller than Vn
- Interface between analog and binary (digital) work
  - Output amplitude (voltage or current) information is quantized (binary)
  - But the time is not quantized !
    - This is exploited in many circuits, as some ADCs or TDCs

Vout = VoH if Vp > Vn

Vout = VoL if Vp < Vn



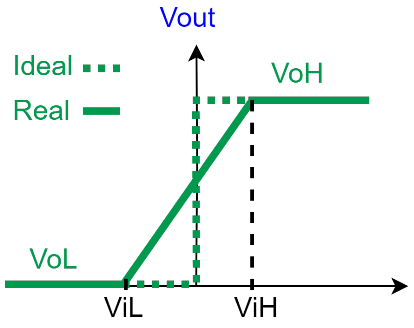




Discriminators: static characteristics

- **Resolution**
  - In an ideal comparator any small difference of the input voltages $\Delta V = Vp - Vn$ is sufficient to achieve either VoH or VoL levels
  - In real comparators the gain A_V is finite
    - Can be modeled as VCVS whose output is limited between VoH and VoL
    - For small ΔV the comparator can be modelled as a linear circuit
  - ViH and ViL represent the minimum ΔV required to saturate the output to VoH or VoL $\rightarrow$ this minimum ΔV is called resolution

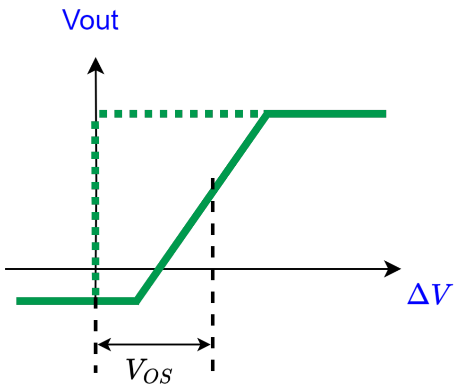
$$Resolution = \Delta V_{min} = ViH - ViL = \frac{VoH - VoL}{A_V}$$



$$\Delta V = Vp - Vn$$

$$A_V = \lim_{\Delta V \to 0} \frac{VoH - VoL}{\Delta V}$$

- **Offset (V_{OS})**
  - When the output does not change for $\Delta V = 0$ but for a different a voltage $\Delta V = V_{OS}$
- **Input impedance**
- **Input common mode range**



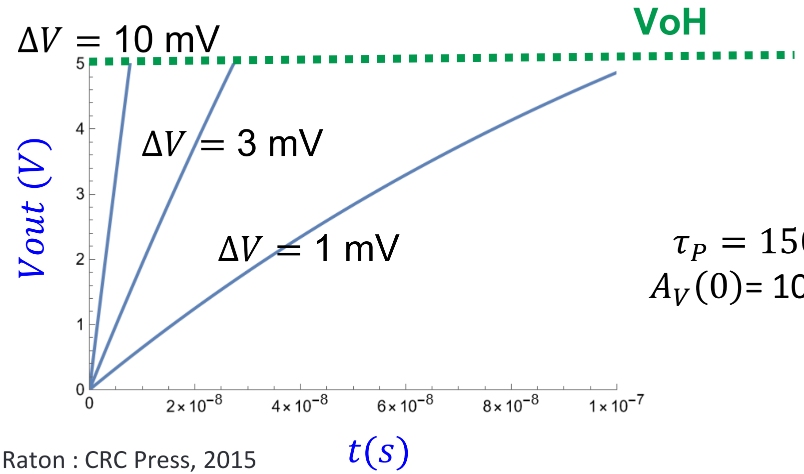
Discriminators: dynamic characteristics

- **Propagation delay (t_{delay})**
  - Time to respond to a variation in the differential input
    - Critical parameter that limits the conversion speed of many systems (ADCs, digital comms, etc)
  - Depends on the small signal and large signal characteristics of the circuit
  - In small ($\Delta V \sim 0$) signal the comparator is modeled as a linear circuit
    - Simplest model is a first order system with a cutt-off frequency (dominant pole) $f_P = {}^{1}/_{2\pi\tau_P}$
    - The higher ΔV (also called overdrive), the sooner Vout reaches VoH (or VoL)

$$A_V(s) = \frac{A_V(0)}{1 + s\tau_P}$$

$$\Big\Downarrow \; Step\ response$$

$$V_{out}(t) = \Delta V A_V(0)\left(1 - e^{-t/\tau_P}\right)u(t)$$



$\tau_P = 150\ ns$
$A_V(0)$= 10k V/V

↗ Rivetti, Angelo. CMOS : front-end electronics for radiation sensors . Boca Raton : CRC Press, 2015

Discriminators: dynamic characteristics

Propagation delay (t_{delay})

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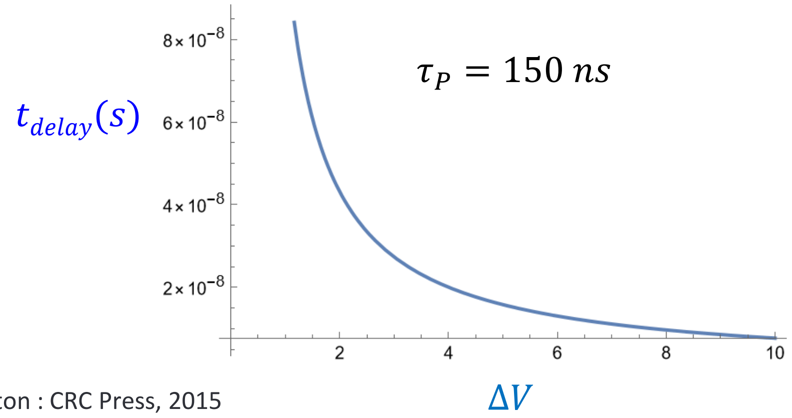
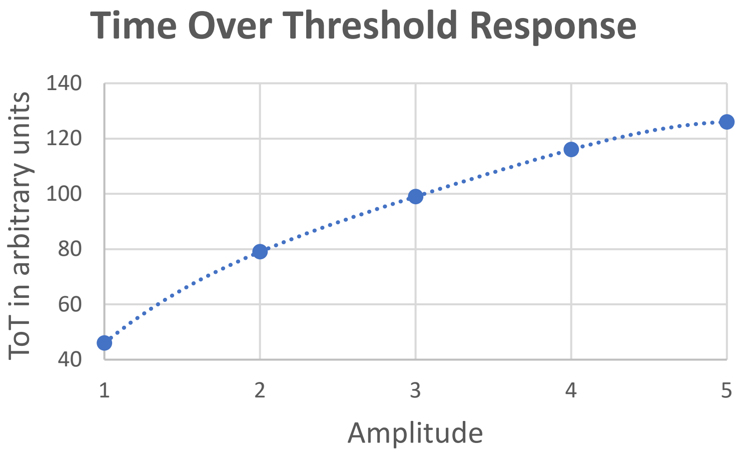
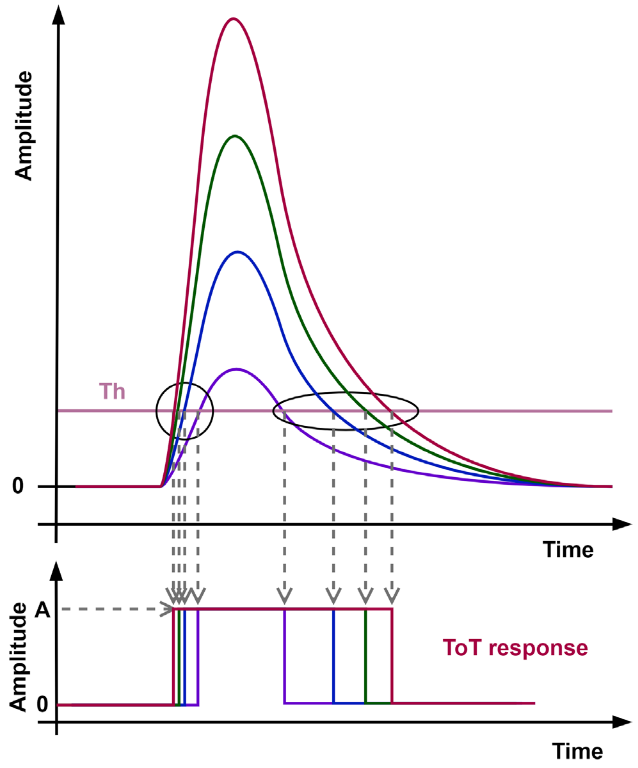


↗ Rivetti, Angelo. CMOS : front-end electronics for radiation sensors . Boca Raton : CRC Press, 2015

Discriminators: time walk

- Pulses of the same shape and different amplitude crosses the threshold at different mean arrival time
- Time walk degrades the electronic jitter in those systems where each event has a different amplitude
- The Time Over Threshold is related to the amplitude
  - Non linear relation
  - “Poor man” ADC

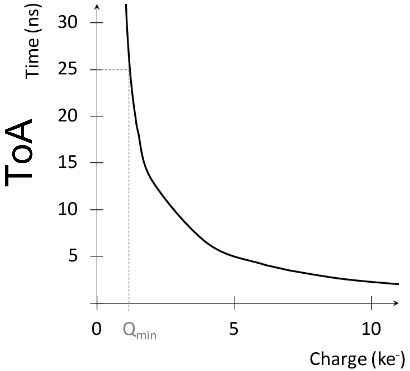




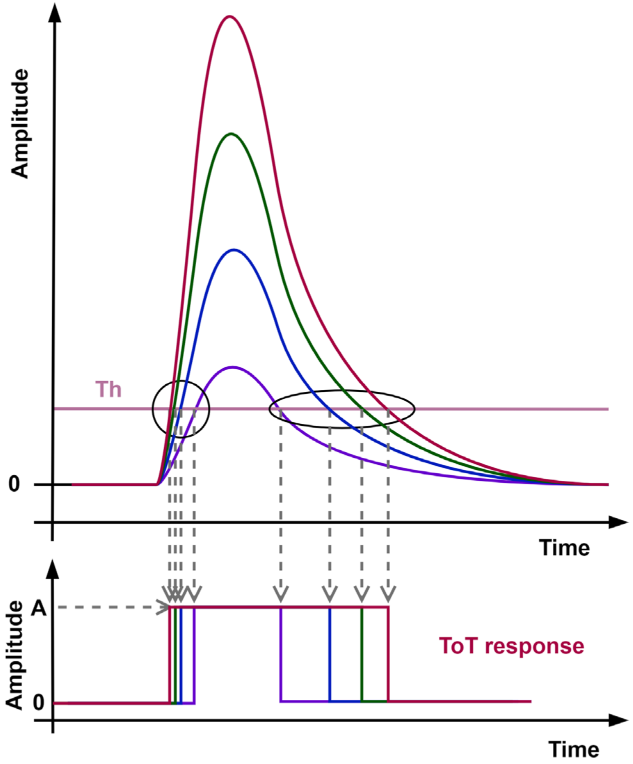
↗ S. Gomez, “Integrated Sensors and Circuits for Imagers and Radiation Detectors”, Master in Semiconductor Engineering and Microelectronic Design, 2025

Discriminators: time walk

- The Time Walk causes a dependence of ToA with amplitude



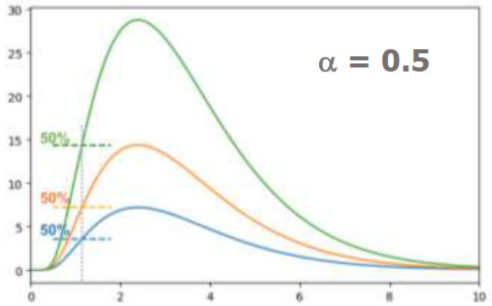
- Similar behaviour as propagation delay but different origin
  - A discriminator with BW→∞ will suffer from Time Walk !



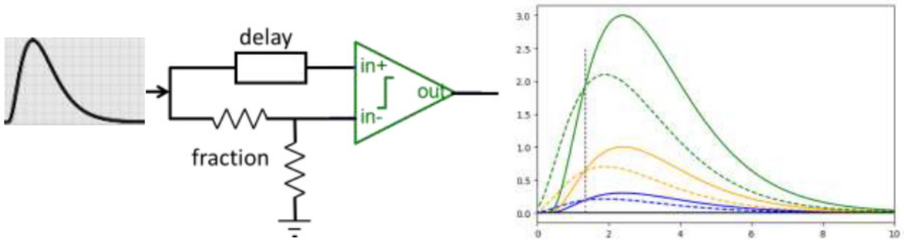
↗ S. Gomez, “Integrated Sensors and Circuits for Imagers and Radiation Detectors”, Master in Semiconductor Engineering and Microelectronic Design, 2025

Discriminators: Constant Fraction Discriminator

- For a pulse waveform of amplitude $Vmax$, the time of the first crossing of $\alpha \cdot Vmax$ (0<a<1) is constant
- Comparison between a delayed copy of the signal and an attenuated one
- Insensitive to time-walk
- Less sensitive to pulse shape variations



IMPLEMENTATION



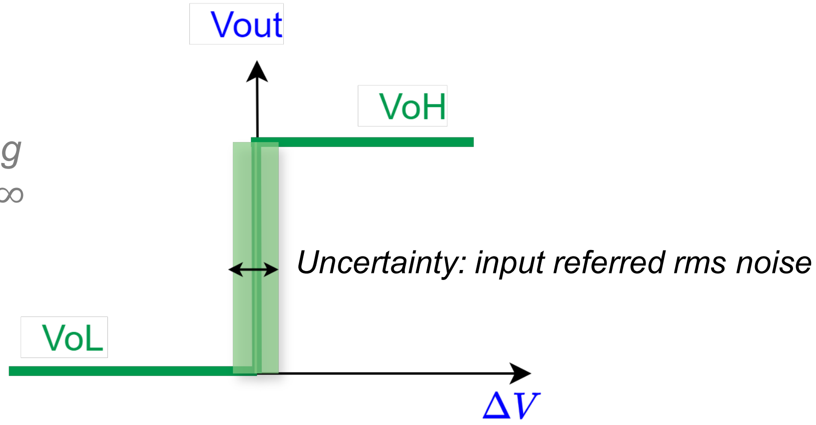
↗ Courtesy: P. O'Connor, "Pulse Processing Electronics", Third Barcelona Techno Week. Course on semiconductor detectors. 2018

Discriminators: electronic noise

Electronic noise: limit on resolution

- Even for $A_V(0) \rightarrow \infty$ electronic noise will limit the resolution of the comparator
- Noise is relevant for small ΔV: we will model again the comparator as a linear circuit
- The noise will lead an uncertainty in the transition region
    - Uncertainty is determined by the input referred noise of the comparator

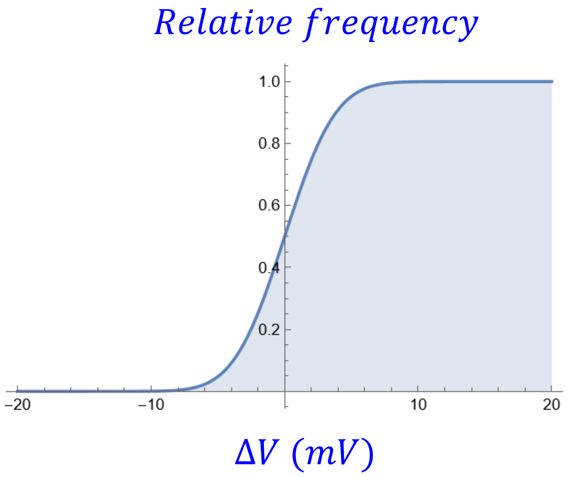
Comparator resolution considering electronic noise and for $A_V(0) \rightarrow \infty$



Discriminators: electronic noise

Electronic noise: estimation

- For small ΔV the response of the comparator is not deterministic
    - If many measurements are performed, we can compute relative frequencies (estimation of probability)
    - This is called S-curve and it corresponds to a cumulative distribution



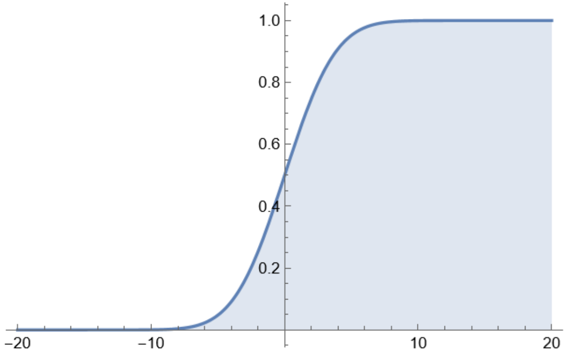
- Relative frequency of Vout=VoH as function of $\Delta V = Vp - Vn$

Discriminators: electronic noise

Electronic noise: estimation

- For small ΔV the response of the comparator is not deterministic
  - If many measurements are performed, we can compute relative frequencies (estimation of probability)
  - This is called S-curve and it corresponds to a cumulative distribution
  - An estimation of the effective noise of the comparator can be obtained by differentiation of the cumulative distribution

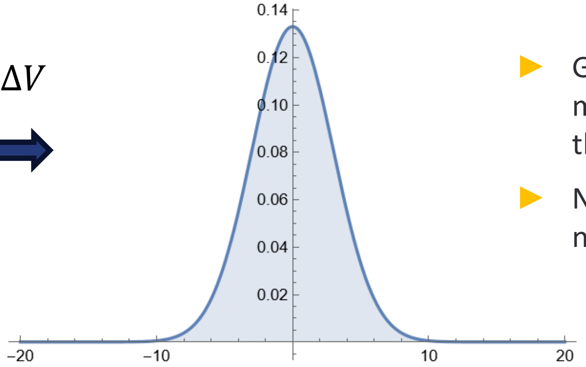
$Relative\ frequency\ of\ Vout = VoH$



$\Delta V\ (mV)$

$$\xrightarrow{\partial/\partial\Delta V}$$

$\sigma = 3\ mV\ rms$

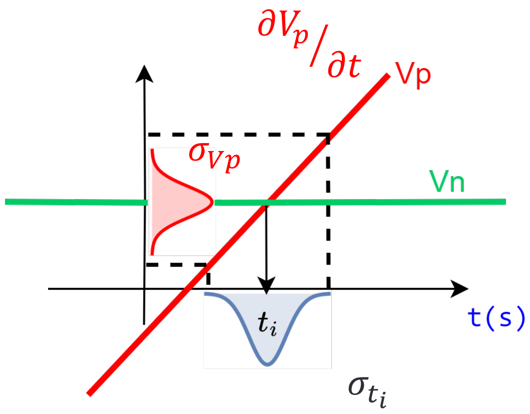


$\Delta V\ (mV)$

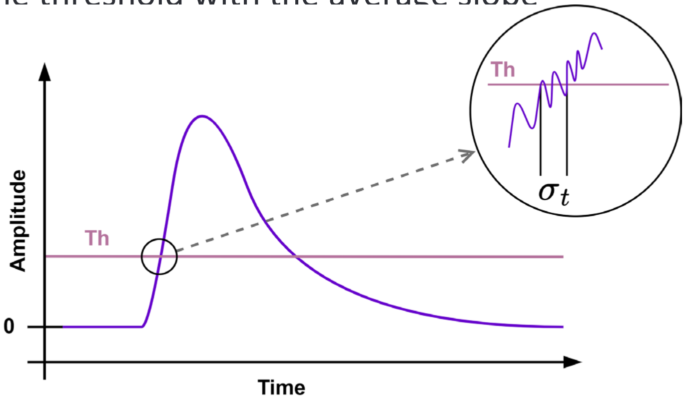
- Gaussian if the noise is determined by most usual intrinsic noise sources: thermal, shot, flicker, etc
- Non-gaussian if dominated by pick noise or EMI

Discriminators: electronic noise

- **Conversion of the electronic noise to timing jitter**
  - Detection of the **crossing time (t_i)** of a signal (e.g.:Vp) that crosses a threshold (e.g.: Vn)
  - The signal crosses the threshold at the output of the preamplifier at some average rate ${}^{\partial V_p}/_{\partial t}$ or ${}^{\partial \Delta V}/_{\partial t}$
  - The signal (and the threshold) has a noise voltage distribution at the average crossing time (e.g.: $\sigma_{\Delta V}^2 = \sigma_{Vp}^2 + \sigma_{Vn}^2$)
  - The standard deviation of the noise voltage projects back to the threshold with the average slope



$$\sigma_{t_i}^2 = \frac{\sigma_{V_p}^2 + \sigma_{V_n}^2}{\left|{}^{\partial V_p}/_{\partial t}\right|_{V_p=V_n}^2}$$



↗ R. Manera. Design of SiPM Front-End Readout Integrated Circuits for large area single-photon imagers, Thesis, 2024

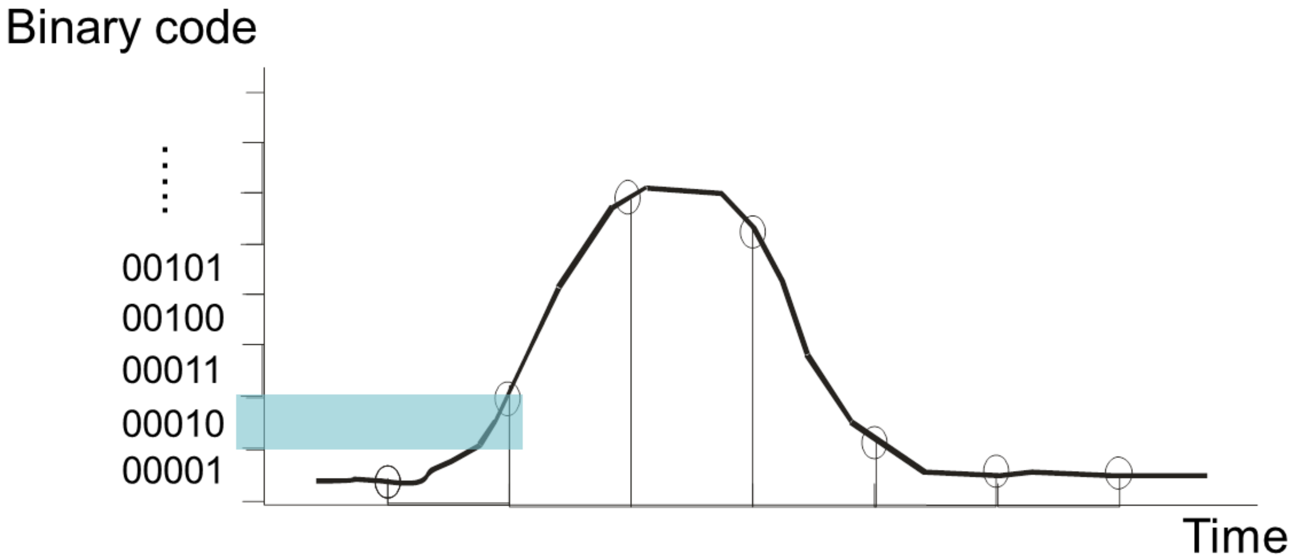
↗ Sepke, Todd & Holloway, Peter & Sodini, Charles & Lee, H.-S. (2009). Noise Analysis for Comparator-Based Circuits. Circuits and Systems I: Regular Papers, IEEE Transactions on. 56. 541 - 553. 10.1109/TCSI.2008.2002547.

Appendix

Digitization: ADCs and TDCs

ADCs

- Interface between analogue and digital domain
- Two alterations of the signal
  1) Signal is sampled at given instants (sampling time)
  2) Continuous amplitude is encoded to a limited number of binary word, i.e. a binary word represents an interval of amplitude (quantization)



ADCs and DACs

- Interface between analogue and digital domain
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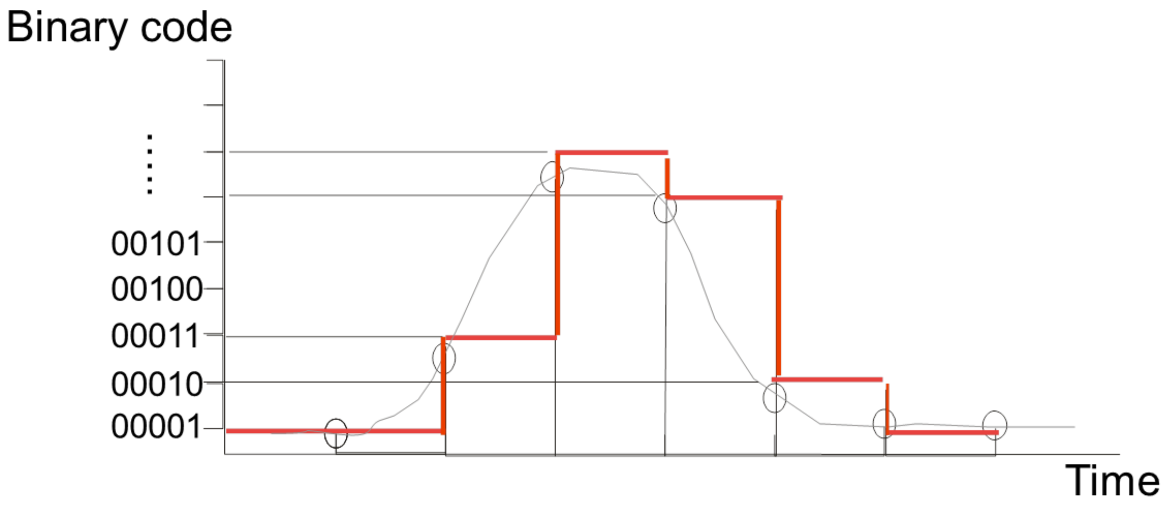
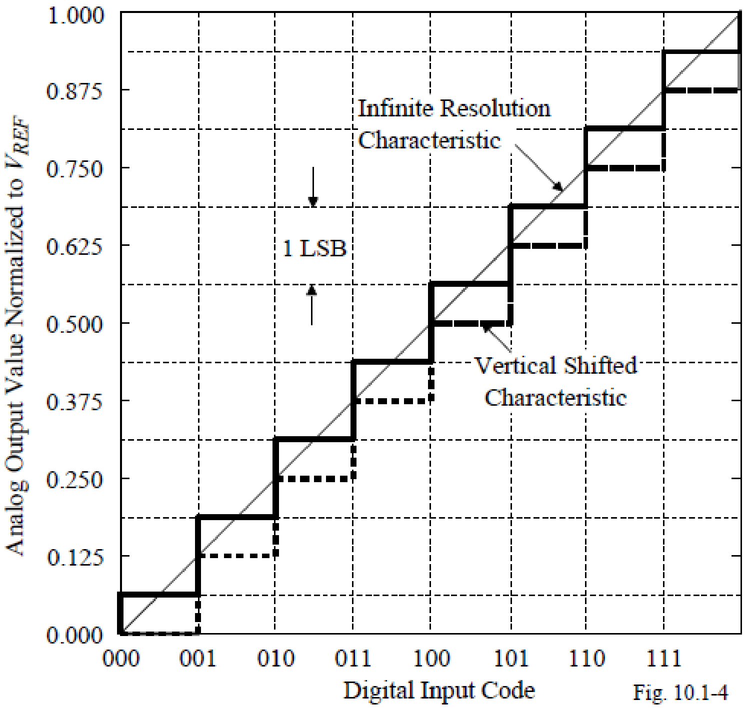


ADCs and DACs

- Resolution
  - DAC: number of bits in the applied digital input word
  - ADC: the smallest analog change that distinguishable by an ADC
- Full Scale range (FS): maximum analog output (DAC) or input (ADC)
  - Corresponds to all bits to 1
- Least Significant Bit (LSB): smallest signal difference that can be resolved
- Dynamic Range (DR) is the ratio of the FSR to the smallest difference that can be resolved (i.e. an LSB)

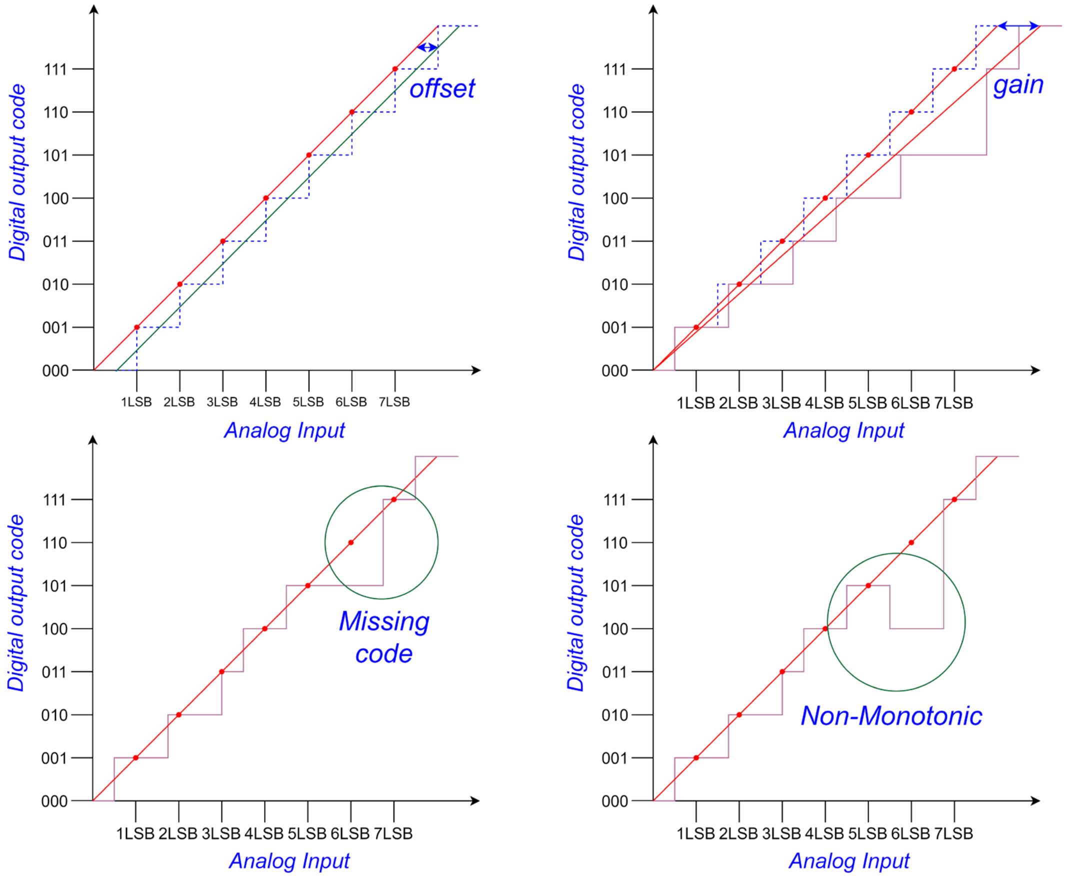
$$DR = \frac{FSR}{LSB} = \frac{FSR}{\left(FSR/_{2^N}\right)} = 2^N$$



↗ CMOS Analog Circuit Design, Lectures 34-40 Prof P. Allen. EECT 7327

ADCs and DACs

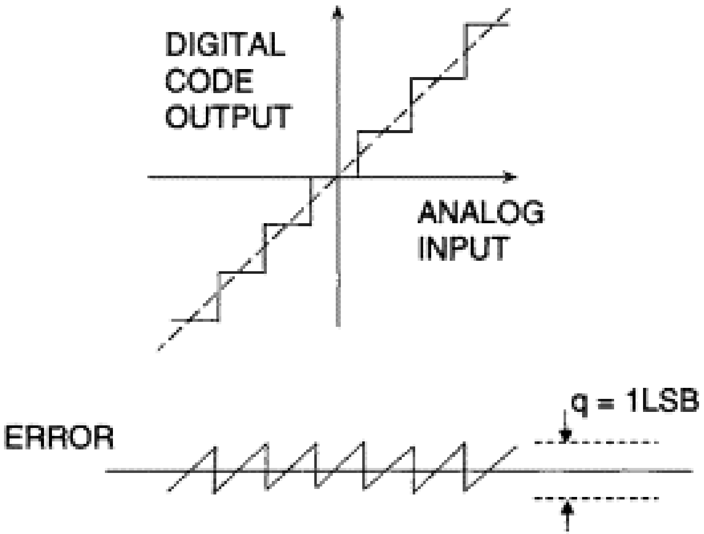
- Static errors

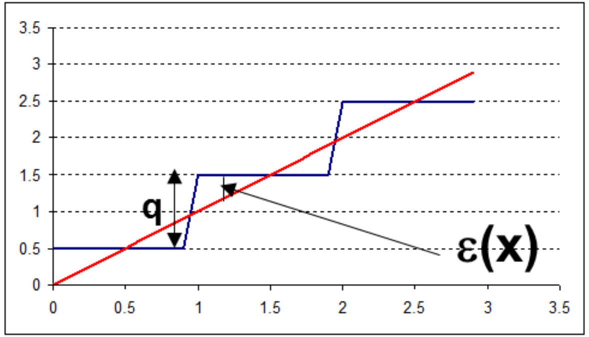


↗ S. Gomez, "Integrated Sensors and Circuits for Imagers and Radiation Detectors", Master in Semiconductor Engineering and Microelectronic Design, 2025

ADCs and DACs

- Quantization Noise is the inherent uncertainty in digitizing an analog value with a finite resolution converter
- Signal-to-noise ratio (SNR) is the ratio of the full scale value to the rms value of the quantization noise



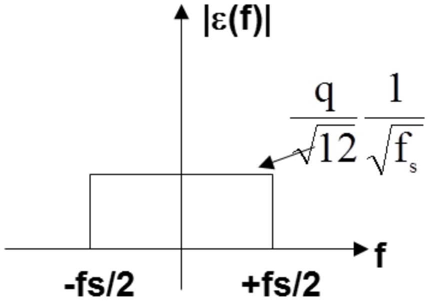


$$-\frac{q}{2} \leq \varepsilon \leq +\frac{q}{2}$$

Variance of uniform distribution:

$$\varepsilon^2 = \frac{1}{q}\int_{-\frac{q}{2}}^{+\frac{q}{2}} x^2\, dx = \frac{q^2}{12}$$

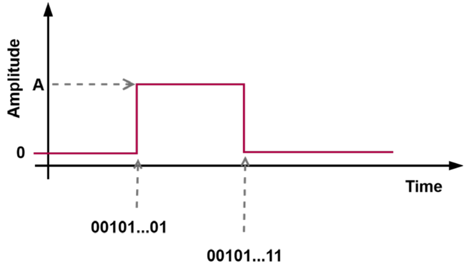
Power Spectral Distribution (PSD)



Types of ADCs

Type	Resolution Number of Bits	Bandwidth	Sampling Rate	Latency	Applications
Integrating Dual-Ramp ADC	$> 16\text{bits}$	$< f_S/2$	$\rightarrow 1\text{kS/s}$	2^{N+1}cycles	DVM, Instrumentation, Sensors
Incremental ADC	$> 16\text{bits}$	$< f_S/2$	$\rightarrow 1\text{kS/s}$	2^{N}cycles	Instrumentation
Sigma-Delta Oversampled	$\sim (10\text{bits}, 18\text{bits})$	$\rightarrow 5\text{MHz}$	$\rightarrow 400\text{MS/s}$	N.A.	Sensors, Audio CODECs, XDSL, Wireless Trans., ...
Algorithmic ADC	$\sim 12\text{bits}$	$< f_S/2$	$\rightarrow 10\text{MS/s}$	$> N\text{cycles}$	Wide Range Low-power medical to telecom
Successive Approximation	$\sim 10\text{bits}$	$< f_S/2$	$\rightarrow 10\text{MS/s}$	$> N\text{cycles}$	Wide Range Low-power medical to telecom
Pipeline (*M* stages)	$\sim 10\text{bits}$	$< f_S/2$	$\rightarrow 100\text{MS/s}$	$> M\text{cycles}$	Telecom, Video
Two-Step Flash	$\sim 10\text{bits}$	$< f_S/2$	$\rightarrow 500\text{MS/s}$	$> 2\text{cycles}$	Video, Multi-Channel Base Stations
Flash Parallel	$\sim 8\text{bits}$	$< f_S/2$	$\rightarrow GS/s$	$> 1\text{cycles}$	Video

Time to Digital Coverters (TDCs)

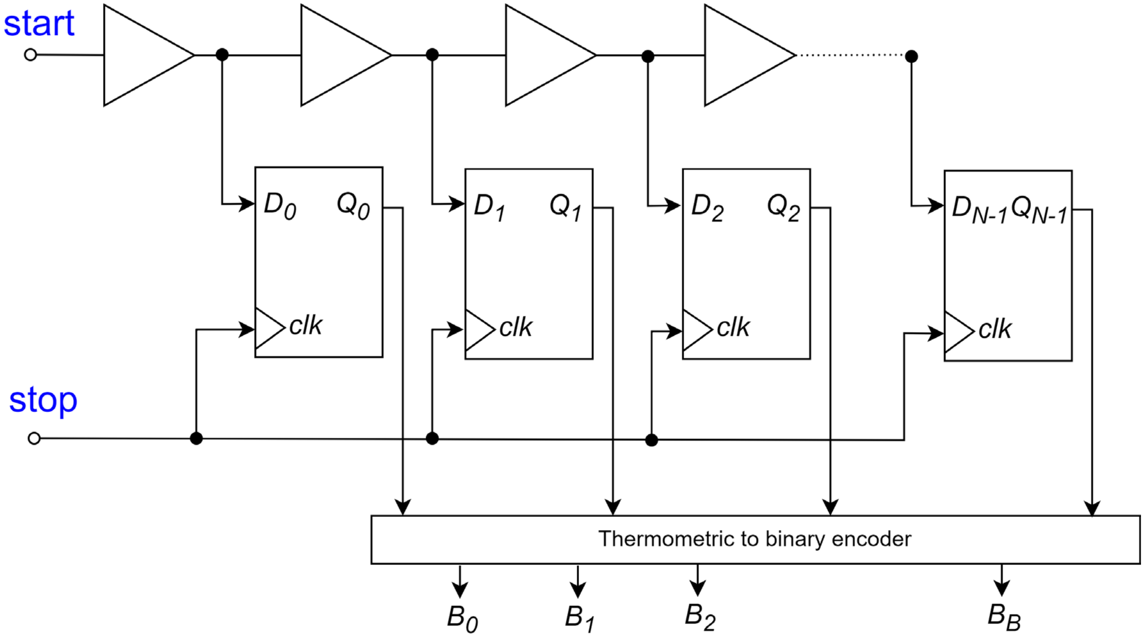


- Digitizes the time of occurrence of signal edges (ToA and ToT)
- Two important concepts:
  - Full scale range T_ref: maximum time difference that can be measured.
  - Time Binning (LSB, least significant bit): smallest time step measured as: LSB= $T\_ref/2^N$
- Similar metrics as ADCs:
  - Conversion Rate (samples/sec)
  - DNL and INL
  - Quantization error (rms): $\sigma_q = LSB/\sqrt{12}$ (sigma of a uniform distribution with an interval size LSB)
- **Single-shot accuracy or jitter** (precision in measuring an individual hit): Evaluated as the accumulated uncertainty or jitter of all blocks in the TDC that cause inaccuracies in the generation of the timestamp.

↗ Rivetti, Angelo. CMOS : front-end electronics for radiation sensors . Boca Raton : CRC Press, 2015

Example of Time to Digital Converters: TDC with Digital Delay Lines

- This architecture exploits the intrinsic **delay** of digital **gates** to measure **time** intervals.

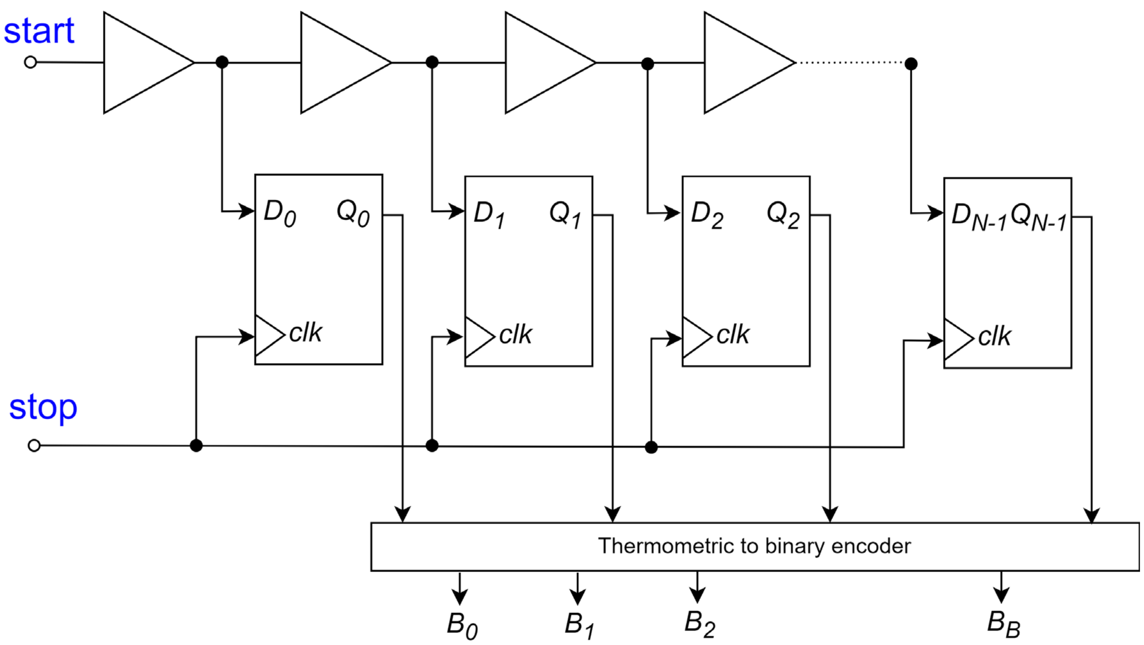


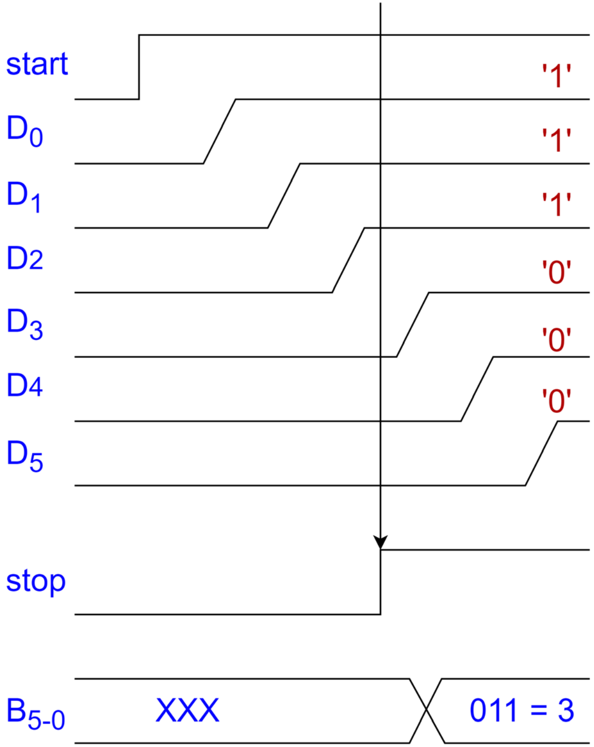
- **Start**: signal is fed to the **buffer** (two inverters) chain.
- **Stop**: time reference (e.g. clock) simultaneously reaches each register and samples the **buffer** outputs.
- The bit pattern is processed by a **thermometric** to **binary** encoder.
- Cell delay is **twice** the **minimum gate delay** of the **technology**.
- **Resolution** determined by the delay of a **single cell**.

↗ Rivetti, Angelo. CMOS: front-end electronics for radiation sensors . Boca Raton : CRC Press, 2015

Example of Time to Digital Converters: TDC with Digital Delay Lines

▶ This architecture exploits the intrinsic **delay** of digital **gates** to measure **time** intervals.





▲ Many other architectures

▲ **But this one can be implemented in FPGAs !**

↗ Rivetti, Angelo. CMOS: front-end electronics for radiation sensors . Boca Raton : CRC Press, 2015