SIGNAL PROCESSING FOR RADIATION DETECTORS

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ATLAS / ILC etc..

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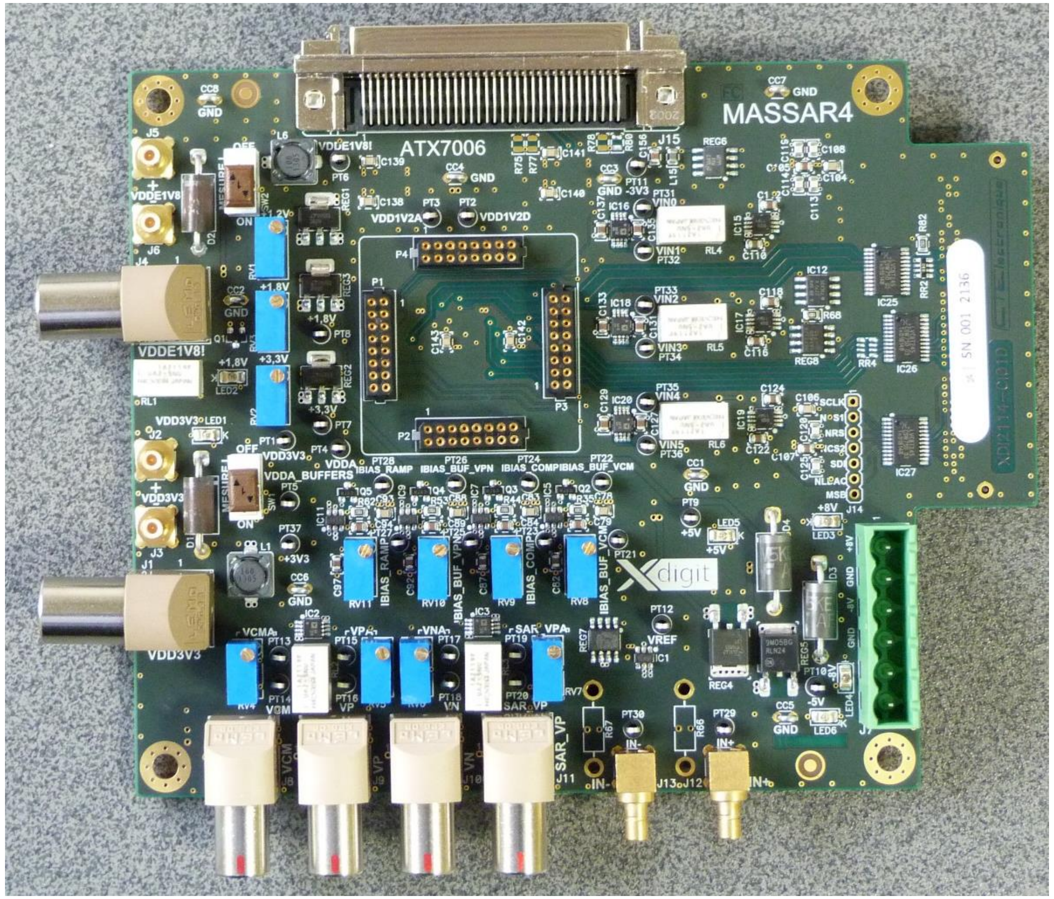
Xdigit is a spin-off on specific ADC design

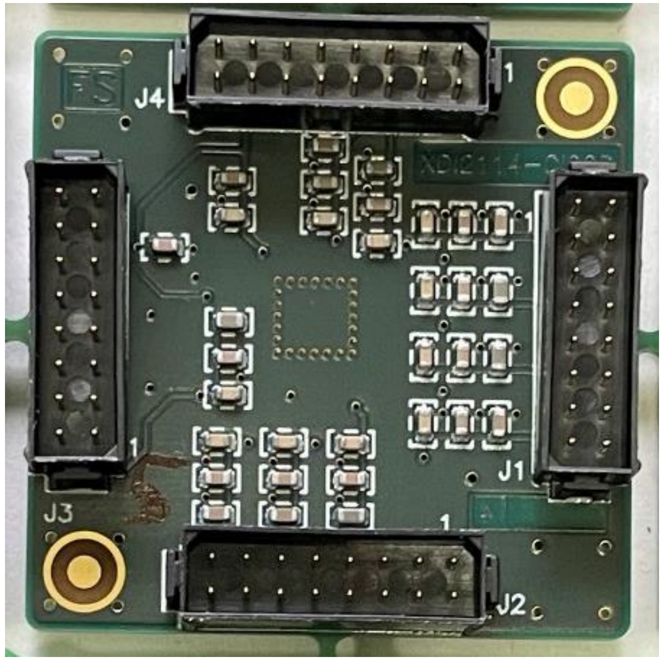
Mainly for array like pixels

www.xdigit.fr

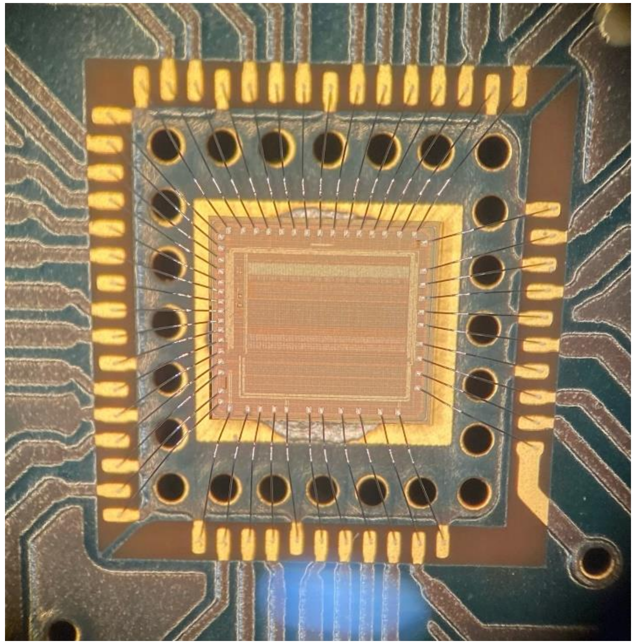
Example of MASSAR 4 chip designed by Xdigit



Mother board



Daughter board

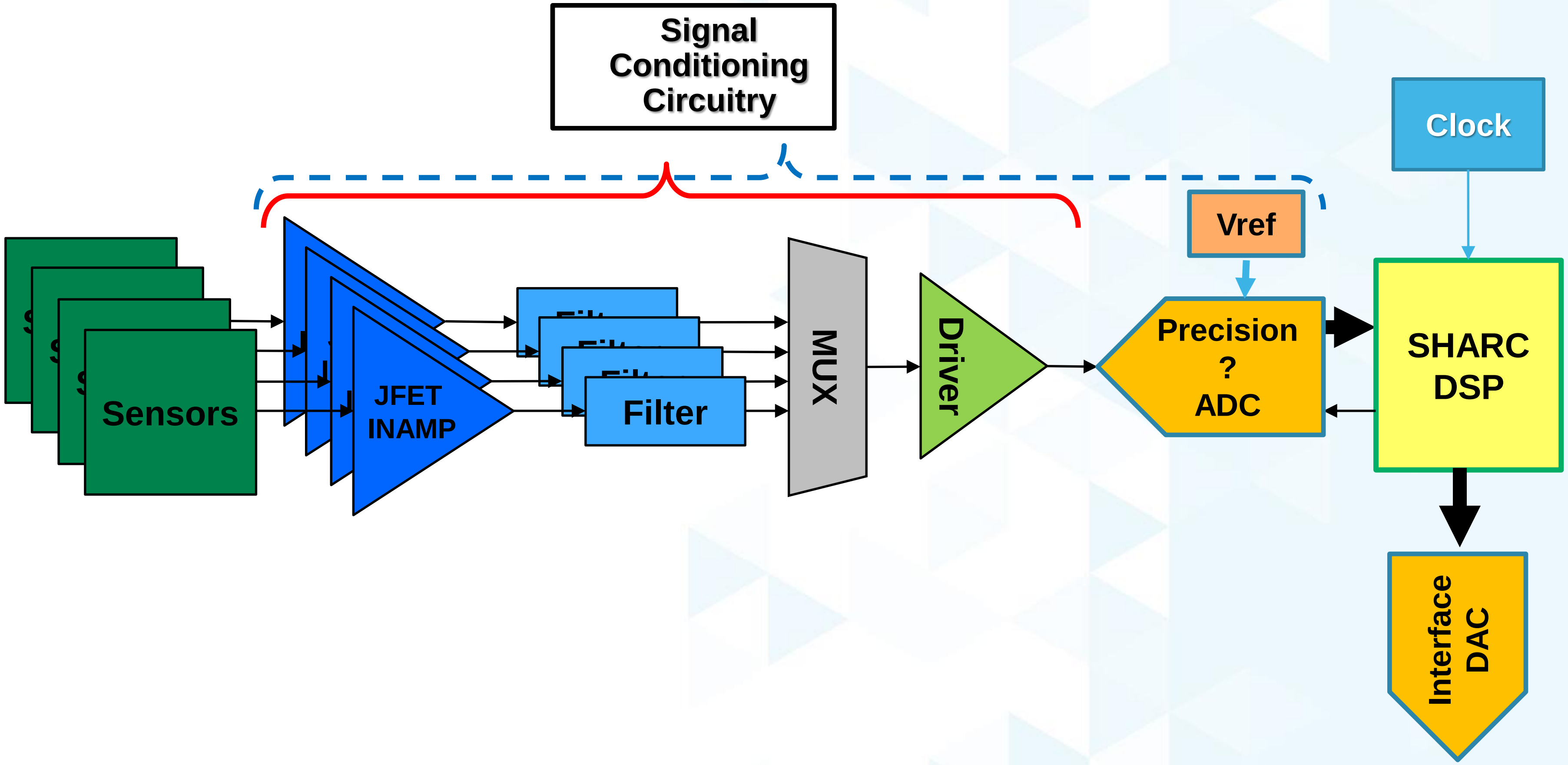


Chip on board
128 channels
of 14 bits ADC
@ 150KS/s

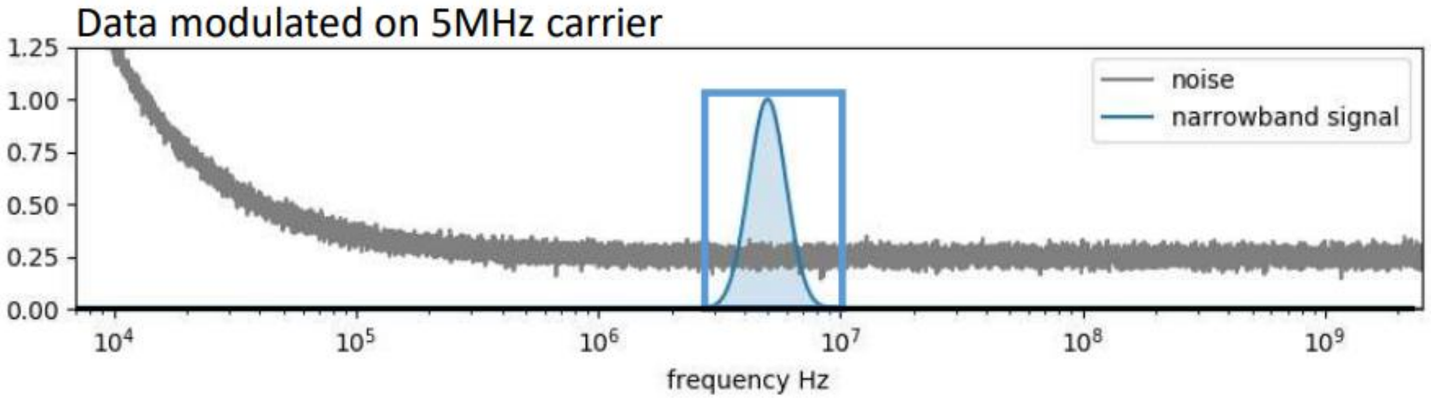
THANKS to many who provide slides or other documents online that I used for this lecture

- Angelo Rivetti: *Front end electronics for radiation sensors (book)*
- *De Geremino Gianluigi (BNL)*
- *Emilio Gatti & Manfredi (INFN)*
- Yan Kaplon (CERN)
- Christophe de la Taille (Omega lab)
- Helmuth Spieler (Lawrence Berkeley National Laboratory)
- Glenn F. Knoll: *Radiation detection & measurement (book)*
- Chiara Guazzoni; http://home.dei.polimi.it/guazzoni
- Paul O'Connor (Brookhaven National Lab Upton, NY USA)

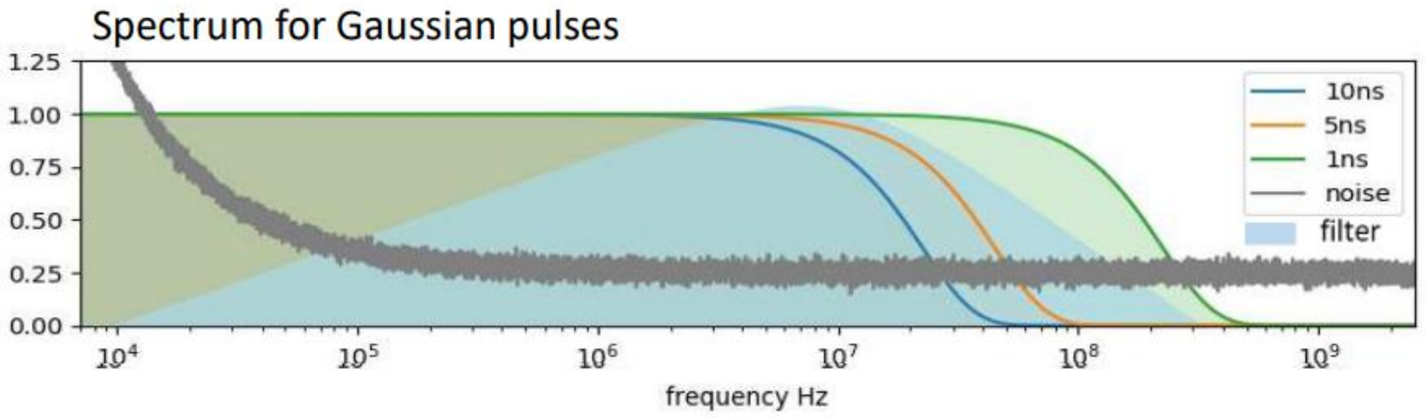
Industrial Analog Signal Chain -> Digital



Noise filtering (frequency domain)



Communications, radar, etc: signal energy is concentrated in narrow band of frequencies

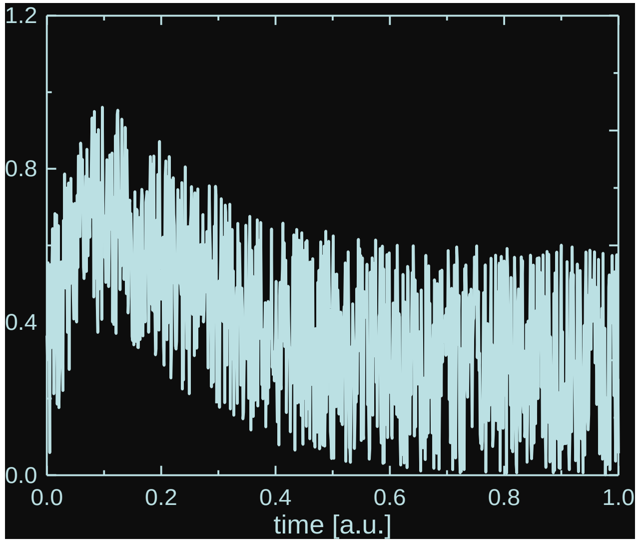


Radiation detection: signal is delta-like in time domain, has energy over a broad range of frequencies

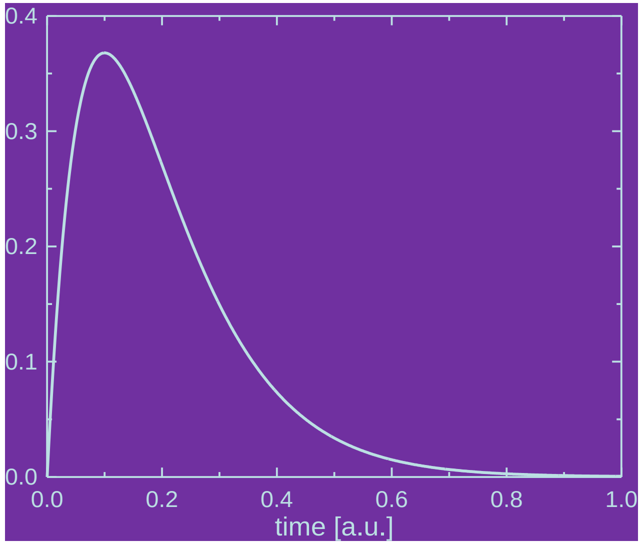
What pulse signal processing means?

Sculpturing!

Designing?



signal processing

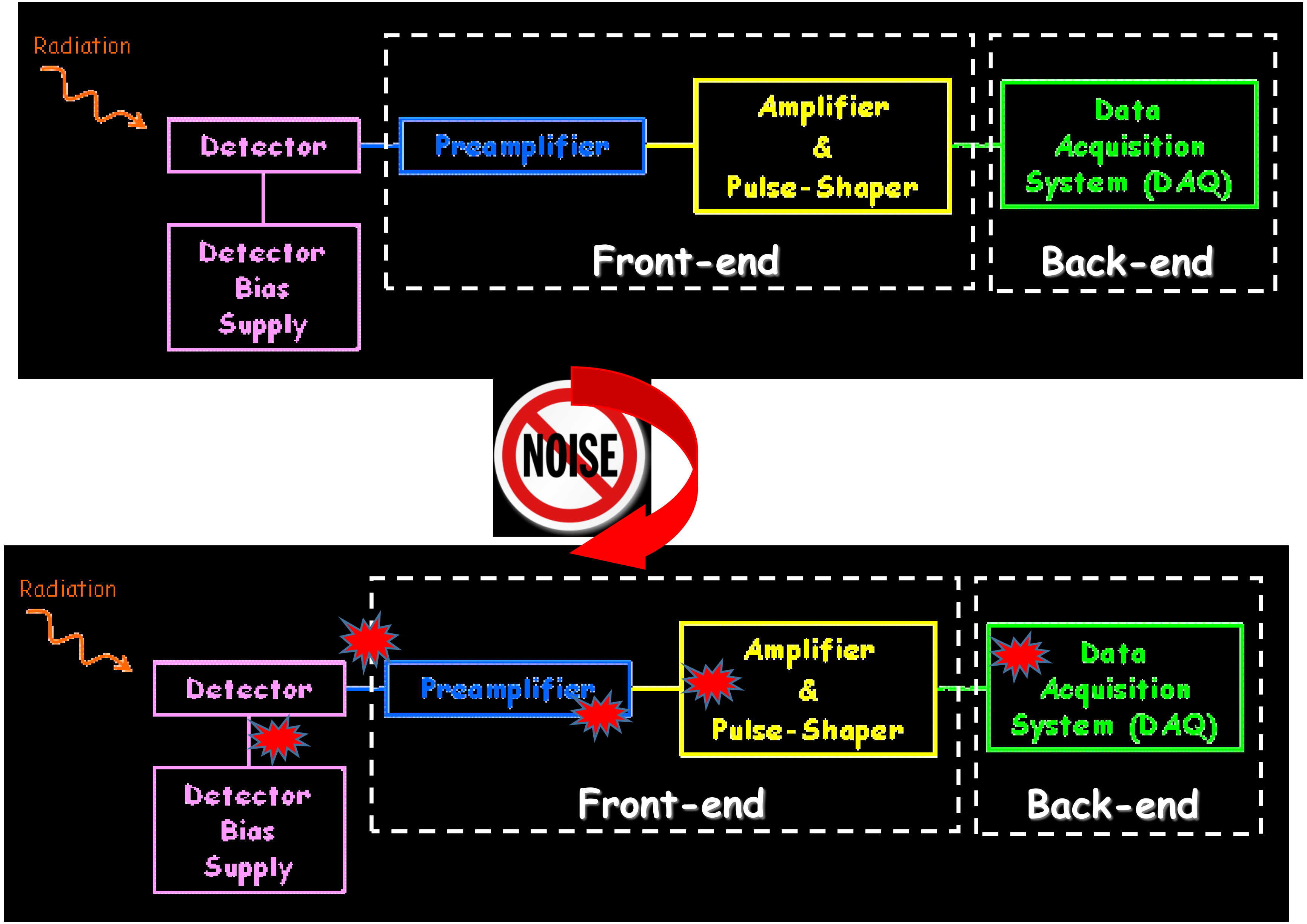


How noisy are flowers growing in your garden

If plants can grow without ?? noise,

Why can't I amplify a signal without adding noise?

Actual Read-out system is unfortunately noisy



Many <u>different types of detectors</u> are used for radiations detection.

Almost all rely on electronics readout.

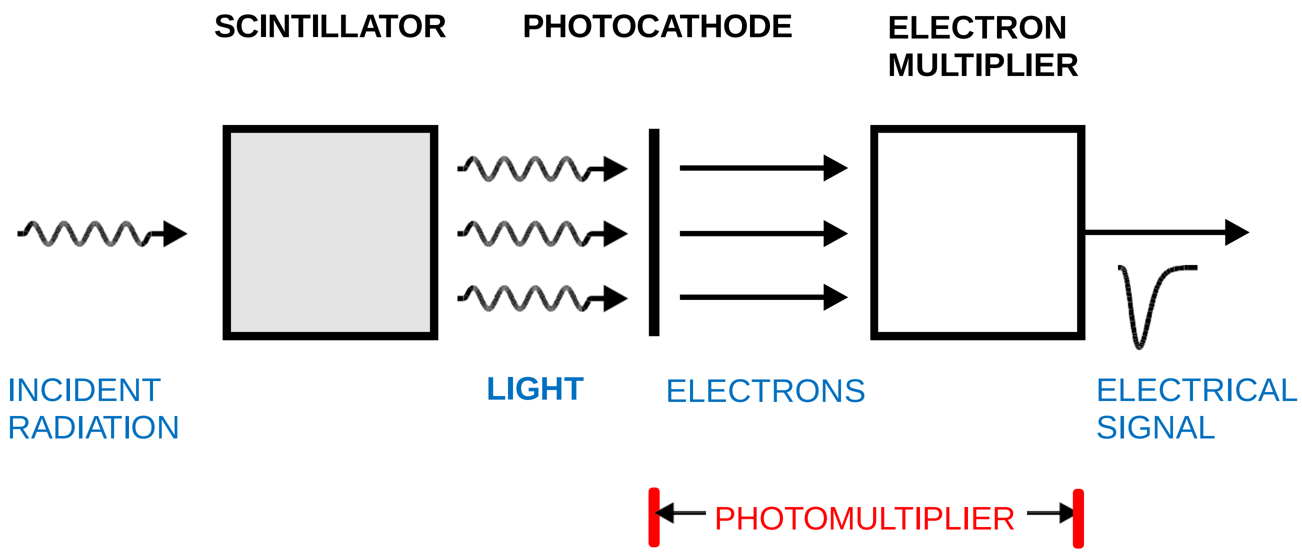
Although detectors appear to be very different, **basic principles** of the readout apply to all.

- The sensor signal is a series of **current** pulses.
- The integrated current $Q_S = \int i_S(t)dt$ yields the signal **charge**.
- The total charge is proportional to the absorbed **energy**.

Readout systems include usually the following functions:

- Signal acquisition (amplification)
- Pulse shaping
- Digitization
- Data Readout

Example: Scintillation Detector



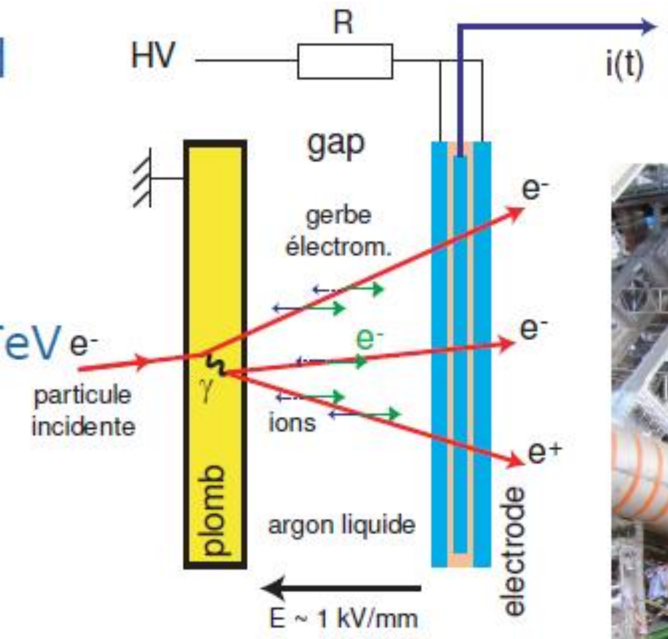
number of photons
∝ absorbed energy

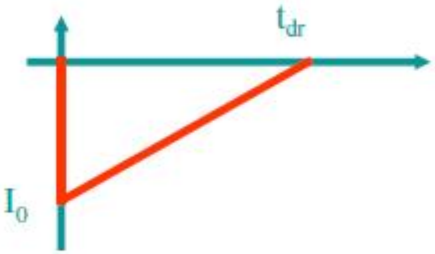
number of photoelectrons
∝ absorbed energy

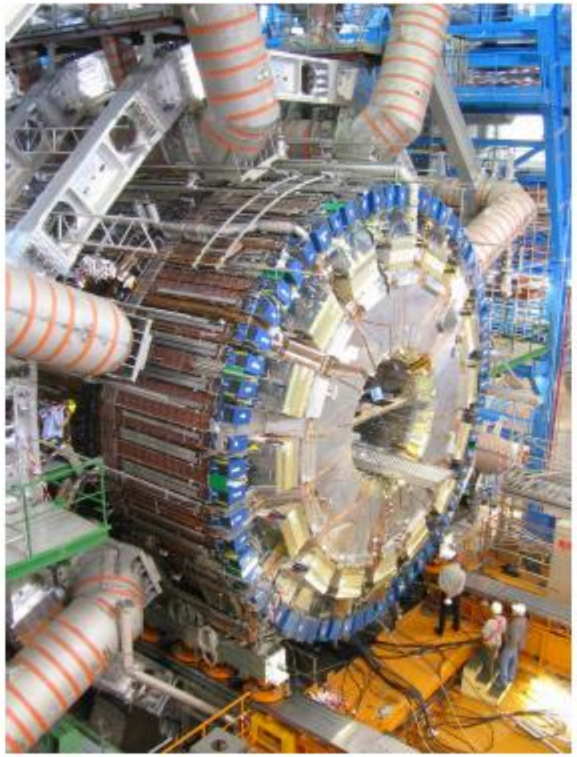
charge in pulse
∝ abs. energy

ATLAS: signal creation in the Calorimeter

- Lead plates in Liquid Argon
- Dynamic range :
  - 16 bits 50 MeV-3 TeV
- Energy resolution :
  - 10%/√E
- Triangular ionisation signal
  - I0 = 2.5 μA/GeV
  - tdr= 450 ns







Classification of detector systems

Detector system is:

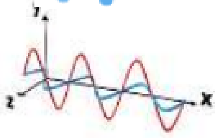
Physics domain

A ***collection of sensor elements*** designed to be sensitive to either:

- Discrete events: arrival of short bursts of energy (may or may not be single quanta) *tracker, calorimeter, PET, XRF, LIDAR*

- Continuous or slowly-varying flux of energy *dosimeter, camera*

- Wave properties of the incident EM field (frequency, phase, polarization) treated classically *UHECR, Dark Ages cosmology, axion search*



This lecture

EE domain

Readout electronics

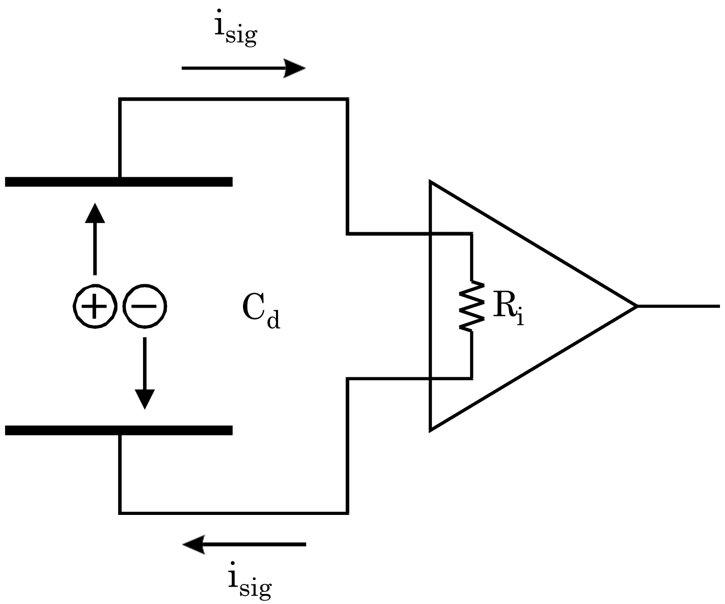
- Amplification, filtering
- Sampling, peak detection
- Discrimination
- Amplitude- or time-to-digital conversion
- Power conditioning

Analysis domain

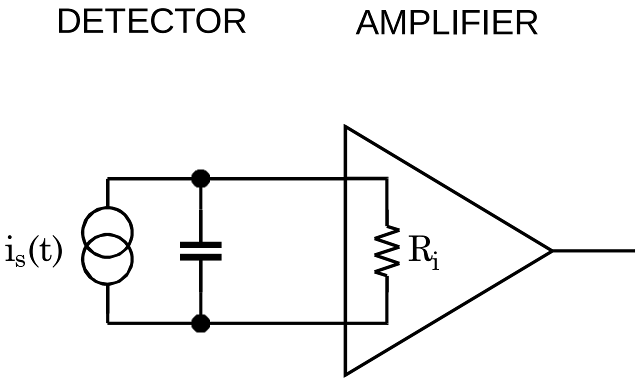
Digital processing to correct sensor + electronics nonidealities and extract features of interest

SIGNAL FORMATION

DETECTOR **AMPLIFIER**



EQUIVALENT CIRCUIT

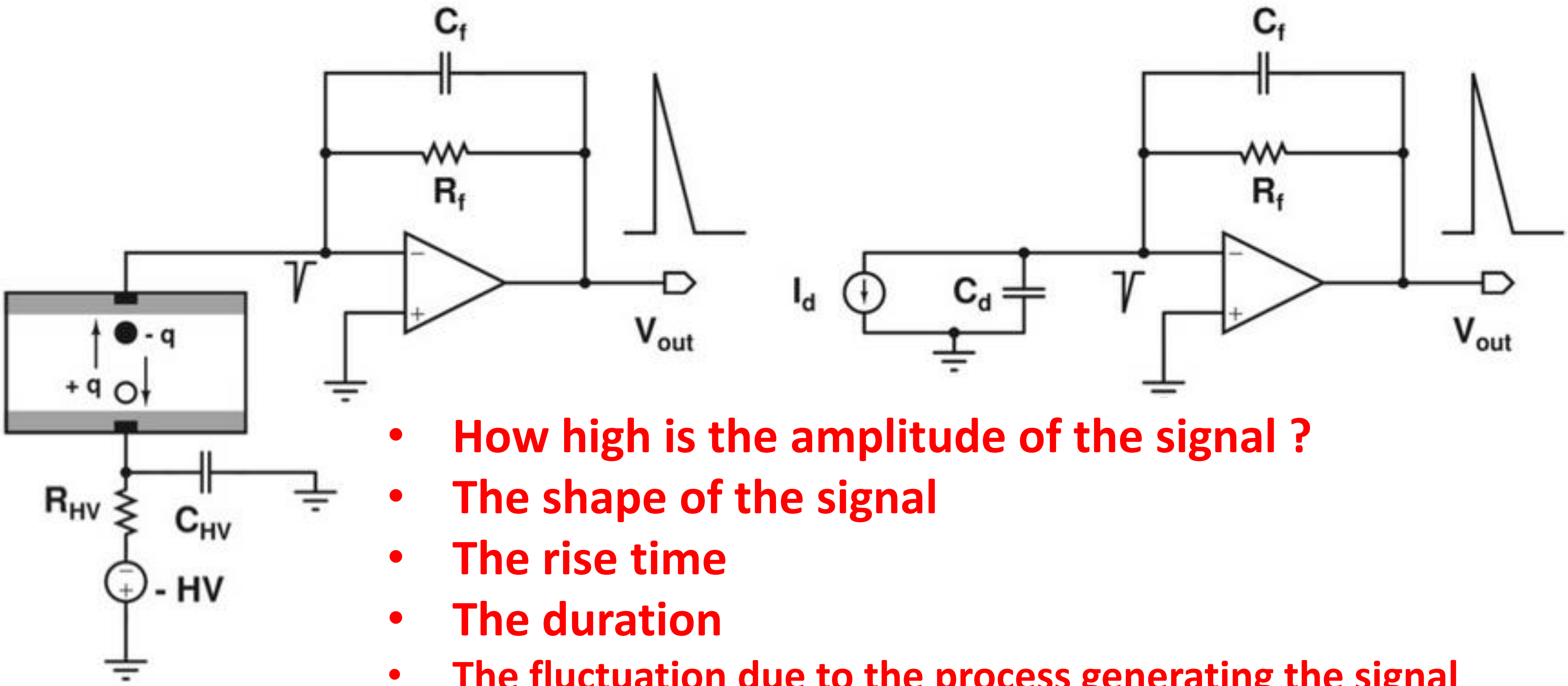


Detector signal **duration**: how short?

- Generally a detector signal is a short current pulse:
  - **thin silicon detector** (10 –300 μm): 100 ps–30 ns
  - **thick (~cm)** Si or Ge detector: 1 –10 μs
  - **proportional chamber**: 10 ns –10 μs
  - **Microstrip Gas Chamber**: 10 –50 ns
  - **Scintillator+ PMT/APD**: 100 ps–10 μs

$$\boldsymbol{Energy \sim \int i(t)\, dt}$$

Signal polarity with negative High voltage



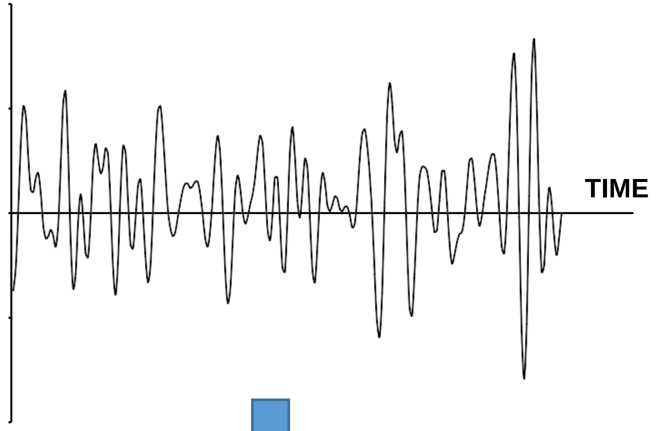
- **How high is the amplitude of the signal ?**
- **The shape of the signal**
- **The rise time**
- **The duration**
- **The fluctuation due to the process generating the signal**

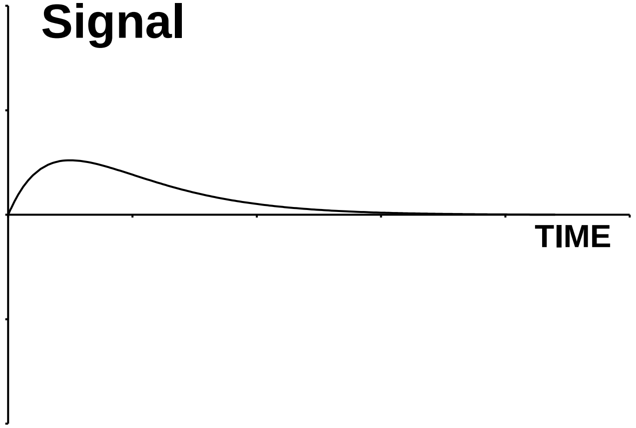
Electronic Noise from the Read out Chain

Choose a time when no signal is present.

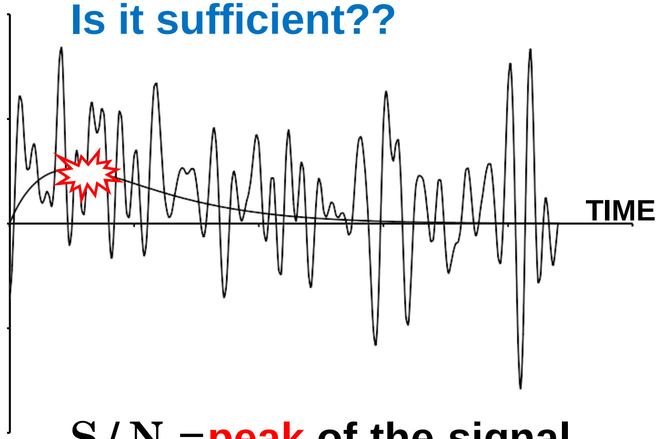
Amplifier's quiescent output level (**baseline**):

In the presence of a signal, noise mix with signal





Signal+Noise (S/N = 1)
Is it sufficient??



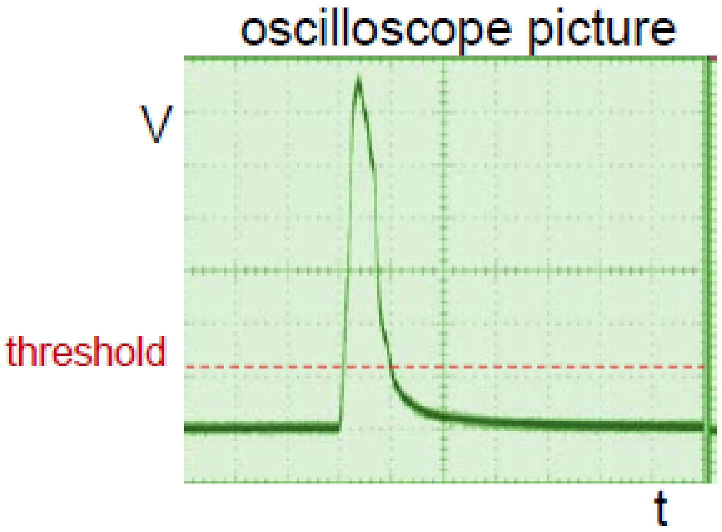
S / N ≡ peak of the signal compared to rms noise

What Information can one extract from a pulse?

Various measurements of this signal are possible
Depending on information required:

- Signal above threshold
  digital response / event count
- Integral of current = charge
  ➔ energy deposited
- Time of leading edge
  ➔ time of arrival (ToA) or time of flight (ToF)
- Time of signal above threshold
  ➔ energy deposited by TOT

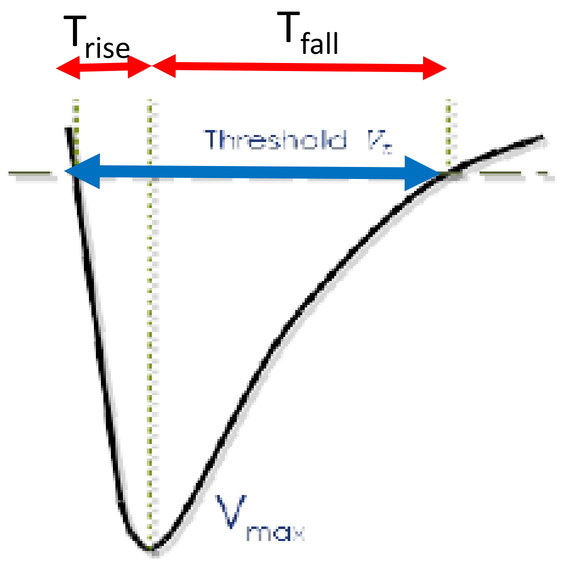
and many more …

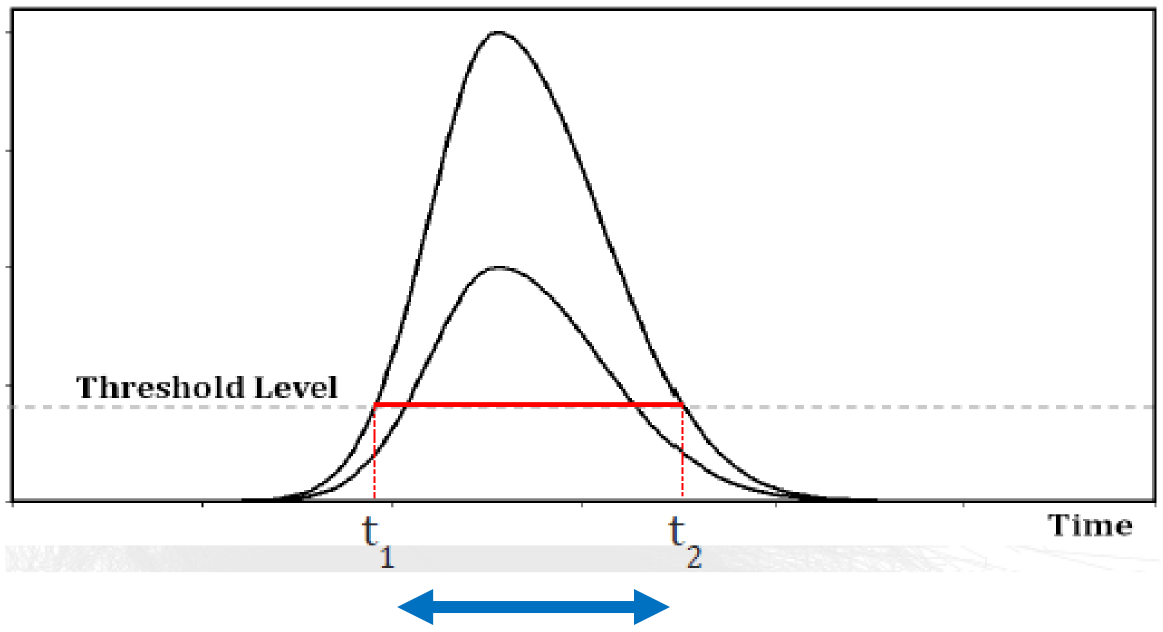


Counting and time over threshold

For **digital imaging** a counter is used after the comparator
Then one proceed by counting the number of events in a frame rate.

The comparator system could be used also to quatify the amplitude of an incoming signal.
The time spent over threshold by the amplifier output is somehow proportional to the amplitude of the incoming signal: TOT





Parameter impacting the pulse **amplitude**:

Ei = Minimum **ionization** energy (depends on the detector cristal, gaz, or liquide)
Ep (> Ei): *average* energy to generate a charges **pair**
E : Energy lost by an incoming particle =>
Np: *Average* **Number** of generated **pairs**

Np = E/Ep => an *average* number

But instantly, the number follows a probabilistic law with a fluctuation from one event to another displaying a ***standard deviation*** $\sigma_{Np} = \sqrt{F * Np}$; F is the Fano factor

In many material F<1 then σ_{Np} is better than one could expect from the Poisson statistics ($\sqrt{Np}$);

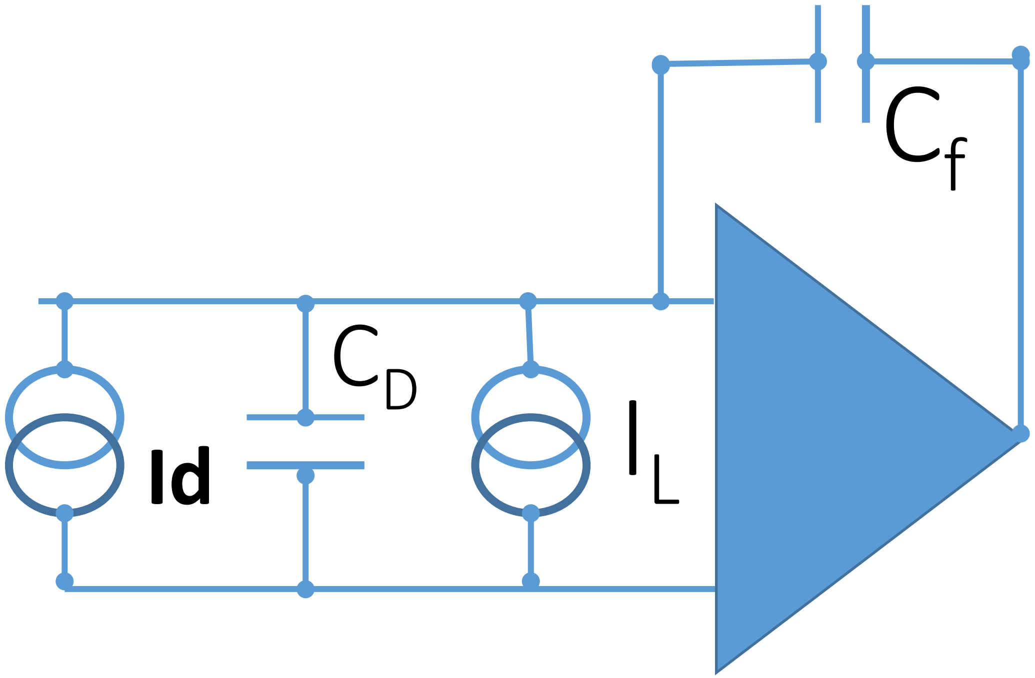
Detector's equivalent circuit: C_D an $I_{L\,(leakage)}$

Detector = capacitance Cd

- Pixels : 0.1-10 pF
- PMs : 3-30 pF
- Ionization chambers: 10-1000 pF
- Sometimes effect of transmission line

Signal : current source

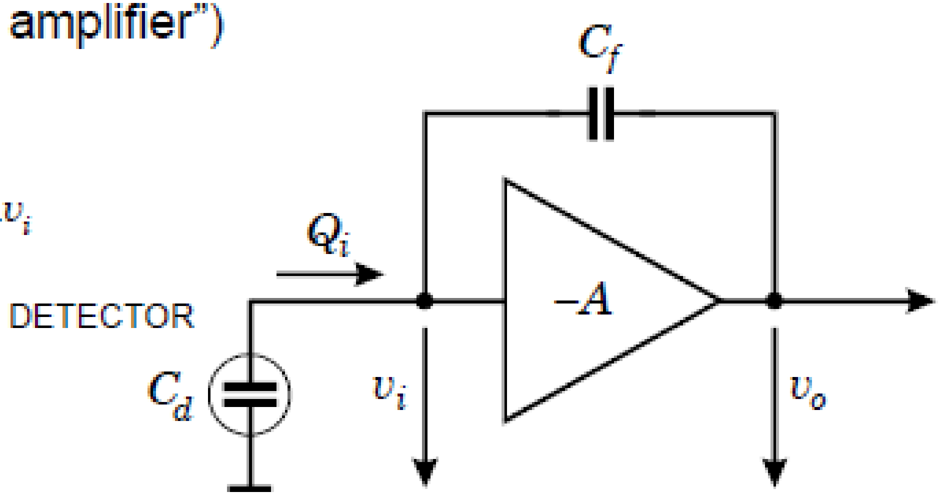
- Pixels : ~100 $e^-/\mu m$
- PMs : 1 photoelectron -> 10^5-10^7 e^-
- Modeled as an impulse (Dirac) :
  $i(t)=Q_0\delta(t)$



- C_D: Impacts on **speed** and **noise** figures
- I_L: impacts on output **DC level**, and on **noise**

Charge sensitive preamplifier: open loop gain

Active Integrator ("charge-sensitive amplifier")



Start with inverting voltage amplifier

Voltage gain $dv_o / dv_i = -A \quad \Rightarrow v_o = -Av_i$

Input impedance = ∞
(no signal current flows
into amplifier input)

Connect feedback capacitor C_f
between output and input.

so $v_0 - v_i = -Av_i - v_i = -(A+1)v_i = -vf$

Voltage difference across C_f: $v_f = (A+1)v_i$

⇒ Charge deposited on C_f: $Q_f = C_f v_f = C_f (A+1) v_i$

$Q_i = Q_f$ (since $Z_i = \infty$)

⇒ Effective input capacitance $C_i = \frac{Q_i}{v_i} = C_f (A+1)$ ("dynamic" input capacitance)

Gain $A_Q = \frac{dV_o}{dQ_i} = \frac{A \cdot v_i}{C_i \cdot v_i} = \frac{A}{C_i} = \frac{A}{A+1} \cdot \frac{1}{C_f} \approx \frac{1}{C_f} \quad (A >> 1)$

Set by a well-controlled quantity, the feedback capacitance.

Helmuth Spieler
LBNL

Charge preamplifier: **open loop gain** and CD typical values

So finally the fraction of charge signal measured by the amplifier is:

$$\frac{Q_i}{Q_s} = \frac{C_i v_i}{v_i(C_i + C_{det})} = \frac{1}{1 + C_{det}/C_i}$$

Must be high enough compared to Cdet

It is like a capacitive divider:
Signal over Cdet, then shared with
a virtual input capacitor Ci= Cf*(A+1)
If Ci is big then only limited charges will
remain on C_{det}.

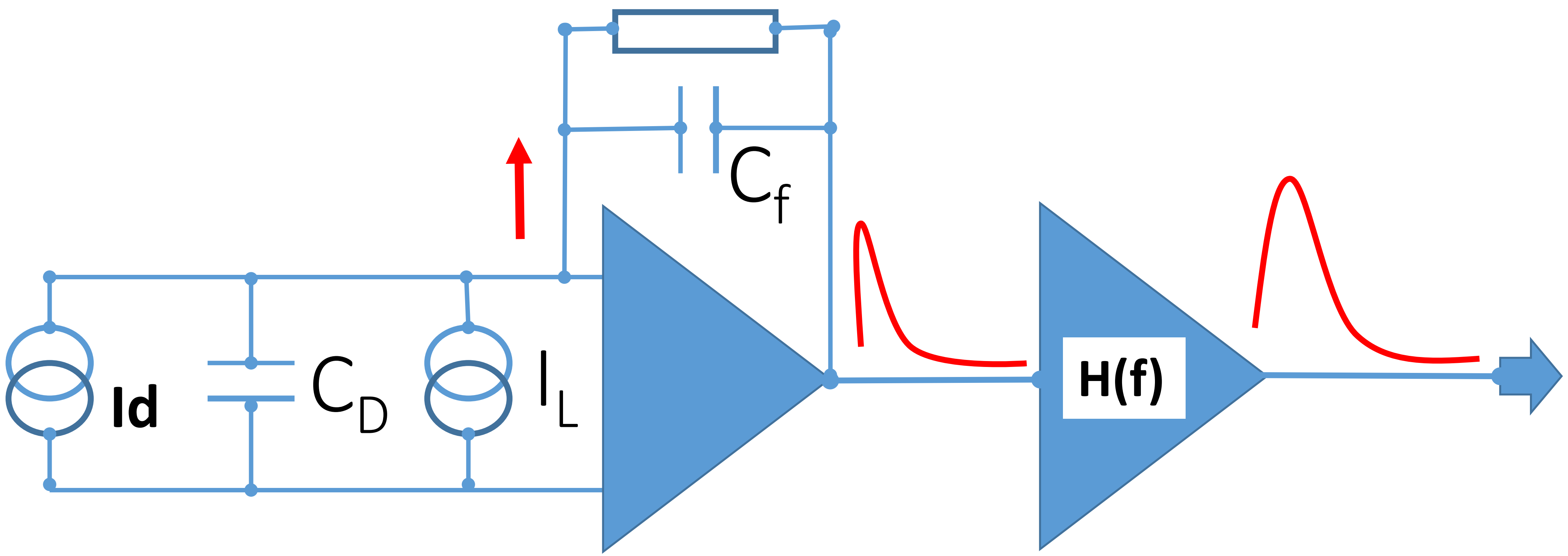
Example: $A = 10^3$, $C_f = 1pF$ ➔ $C_i = 1nF$

So if we consider $C_{det} = 10pF$ ➔ $Q_i/Q_s = 0.99$ $(C_i >> C_{det})$

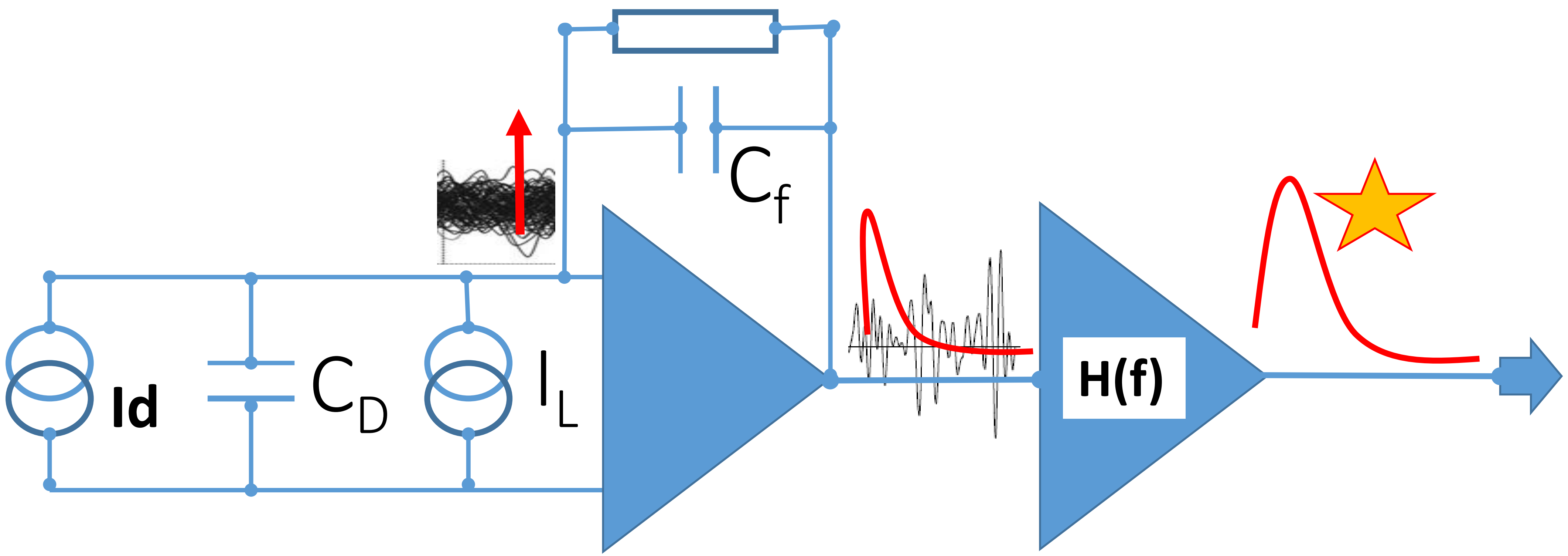
But if we consider $C_{det} = 500pF$ ➔ $Q_i/Q_s = 0.67$ $(C_i \sim C_{det})$

Si det: 50um thick, 500mm^2 area

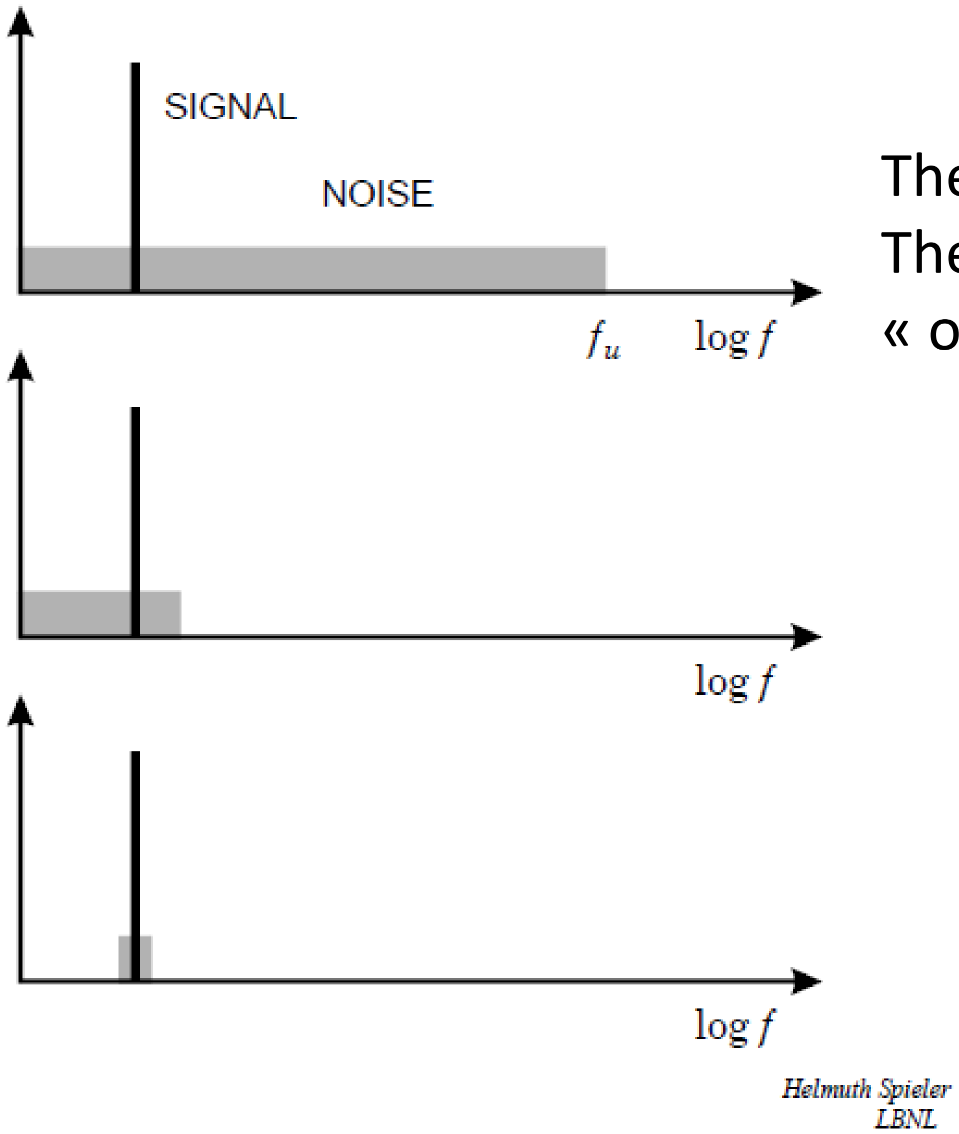
Front end amplifier and shaper circuit



Noise and Front end amplifiers

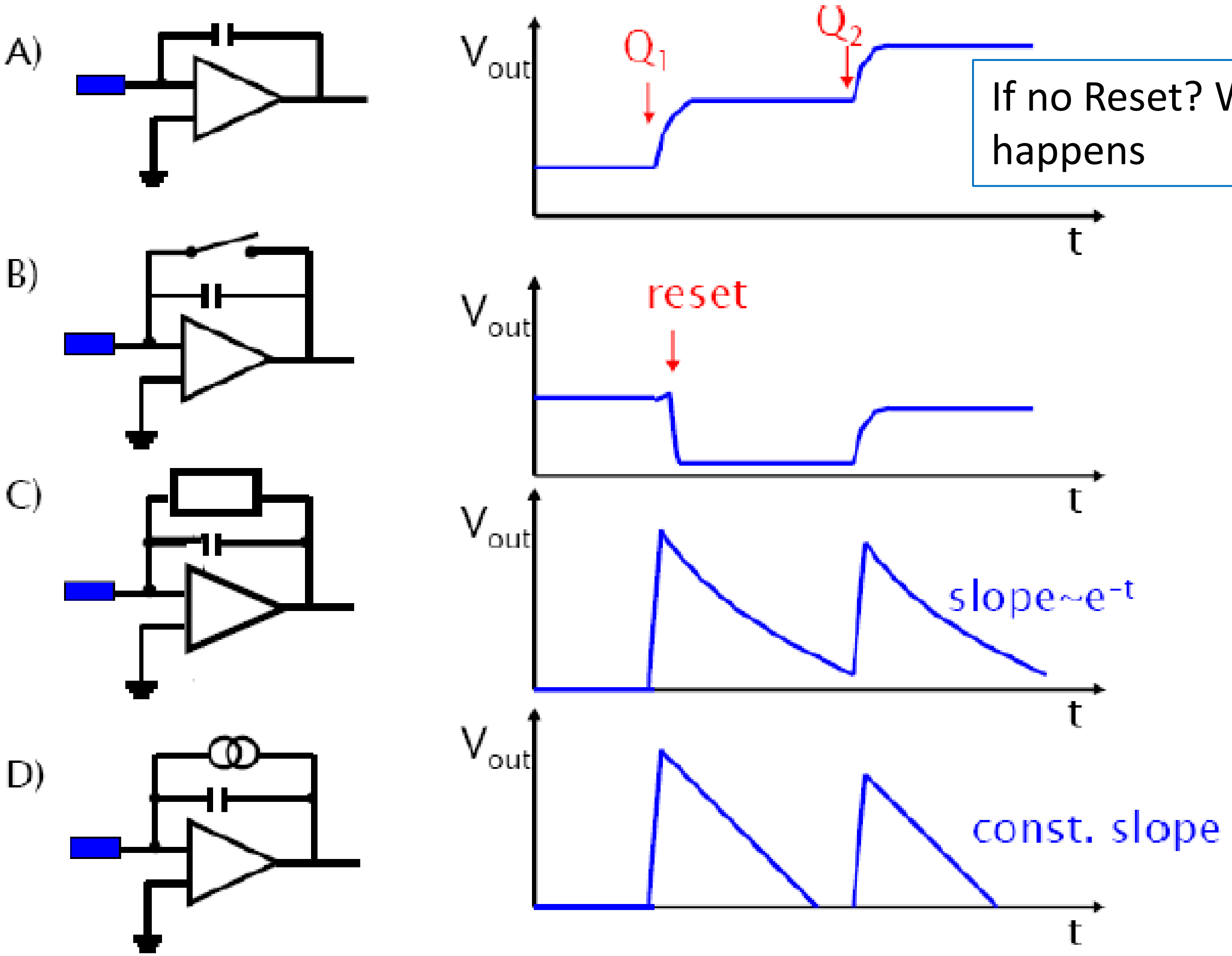


Signal/Noise optimization with bandwidth



The total noise will be the integral over the bandwidth
Then Signal/Noise could be improved by
« optimizing » the noise bande close to the signal's.

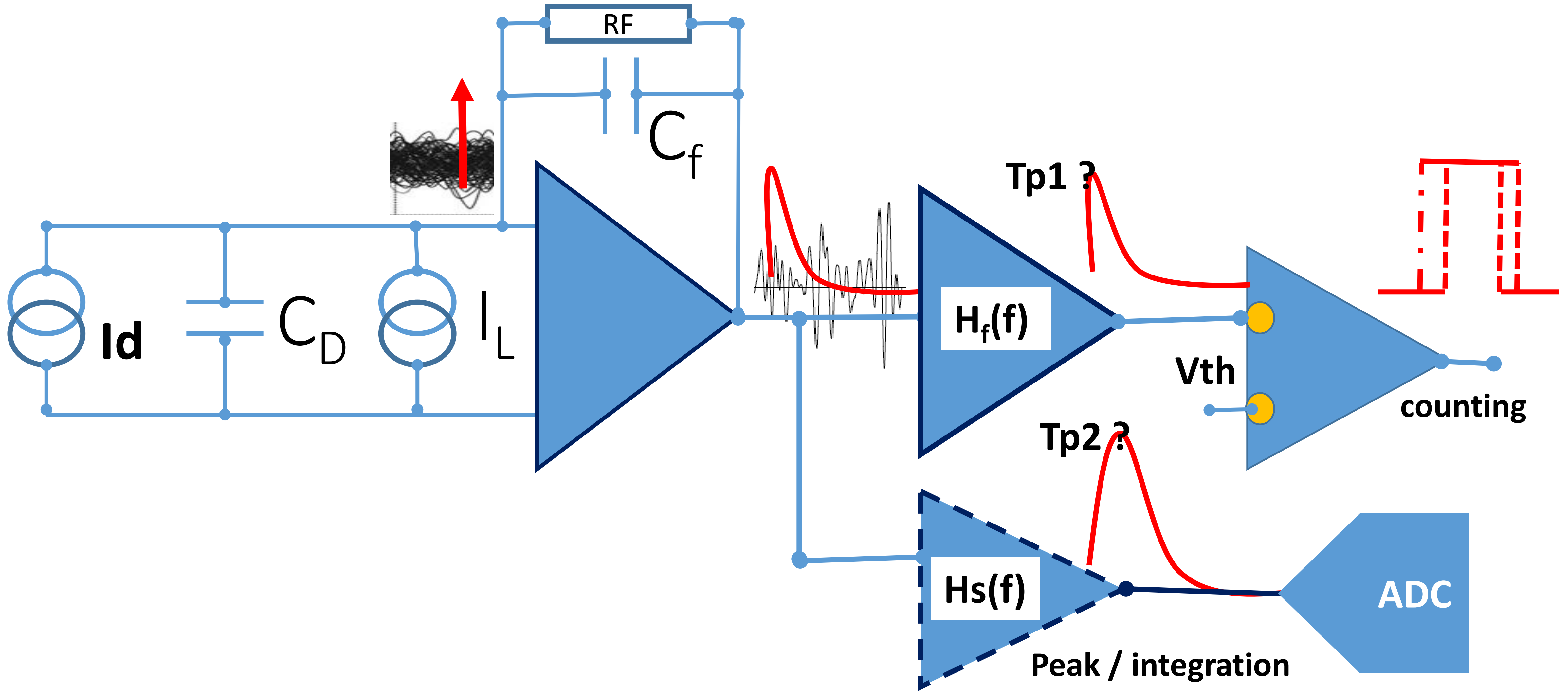
Different options of reset systems



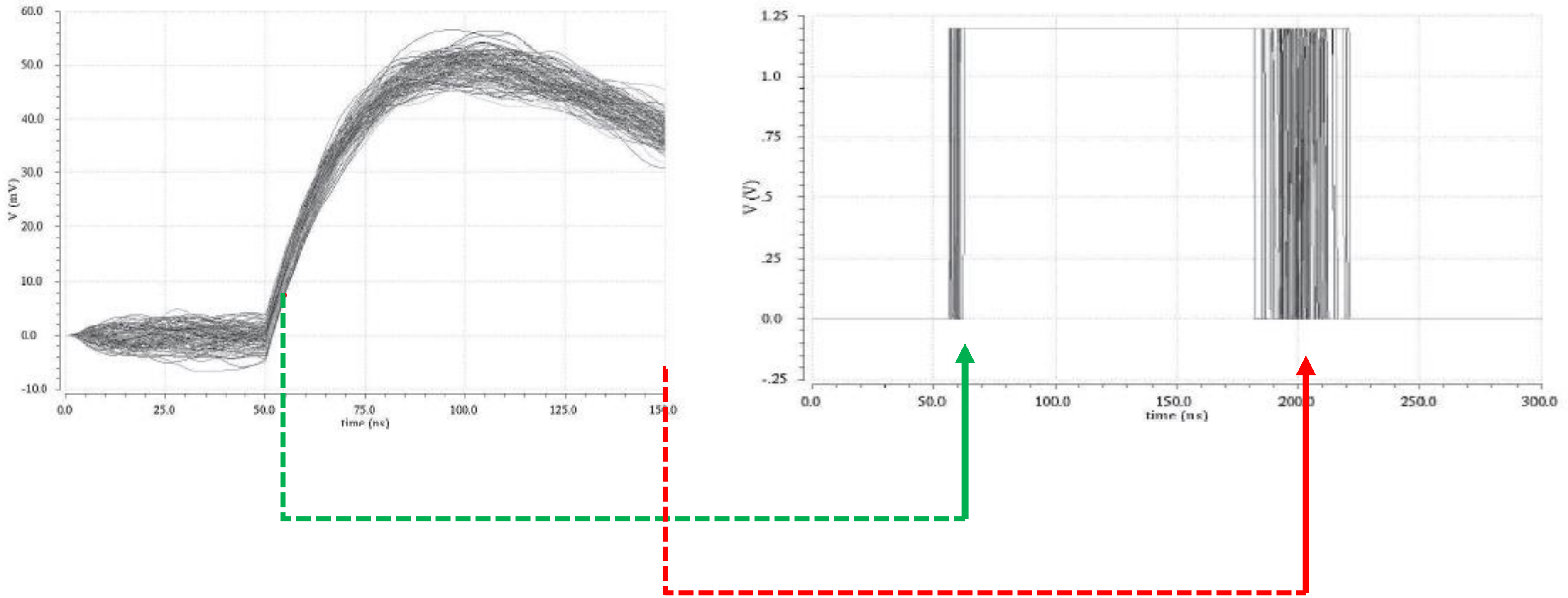
If no Reset? What happens

Ref LBL + de La Taille

Noise in integration stage versus counting



Even for counting, beware of the noise: why using a shorter peaking time for counting readout?

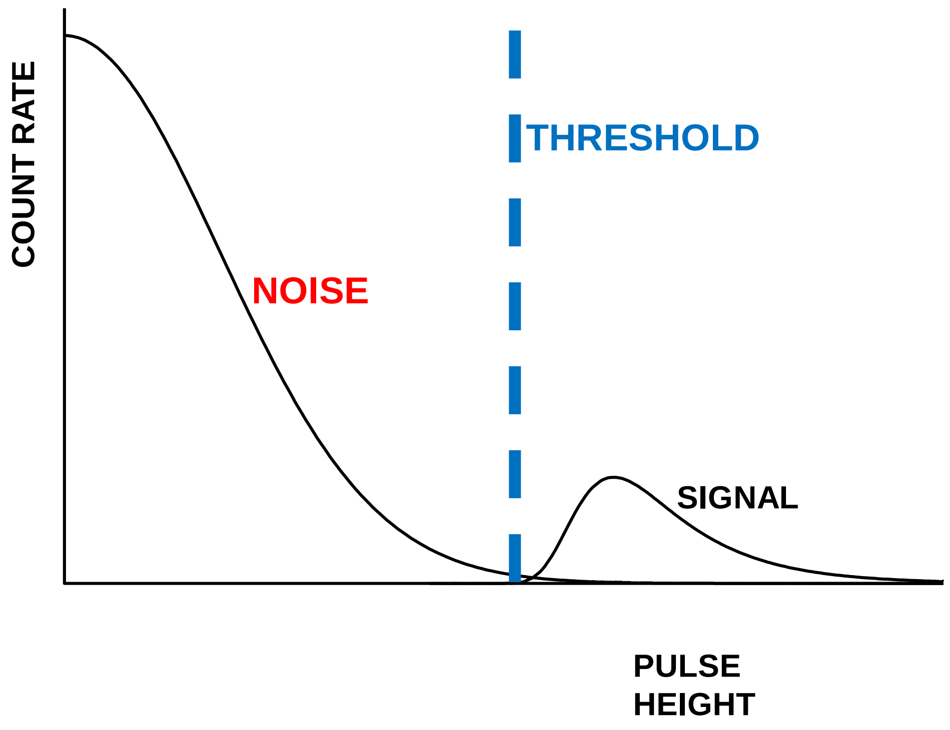


How setting the threshold of a comparator for counting purpose?

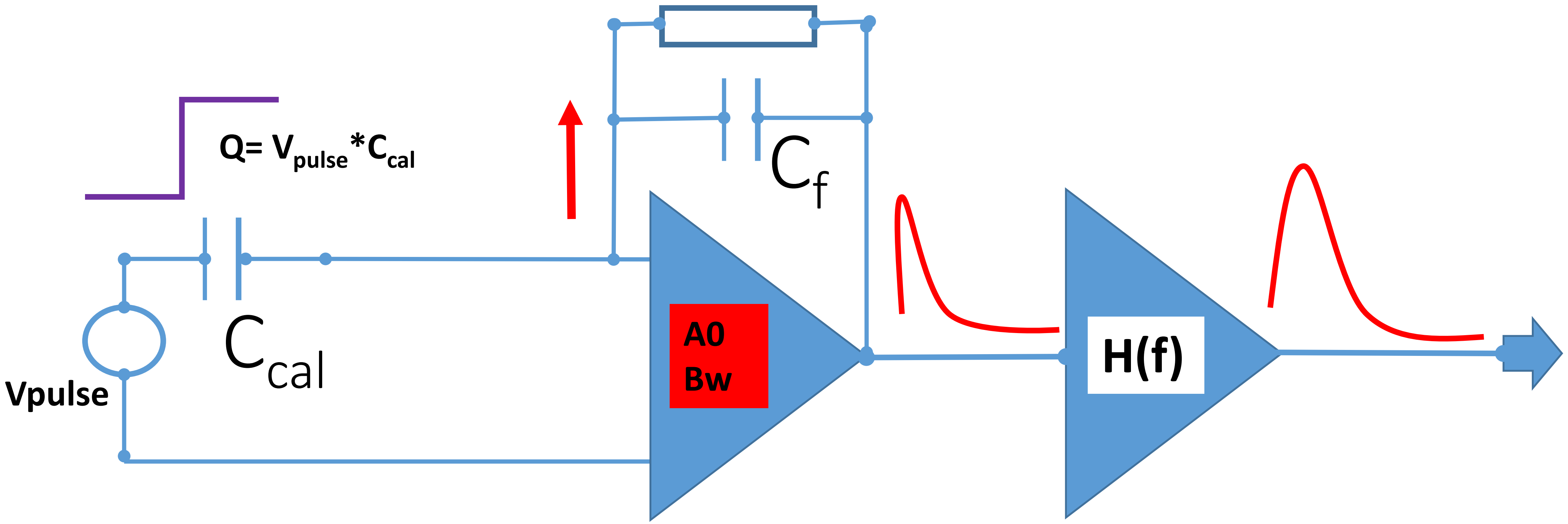
It must be set:

1. High enough to reduce noisy hits

2. Low enough to capture the minimum signal

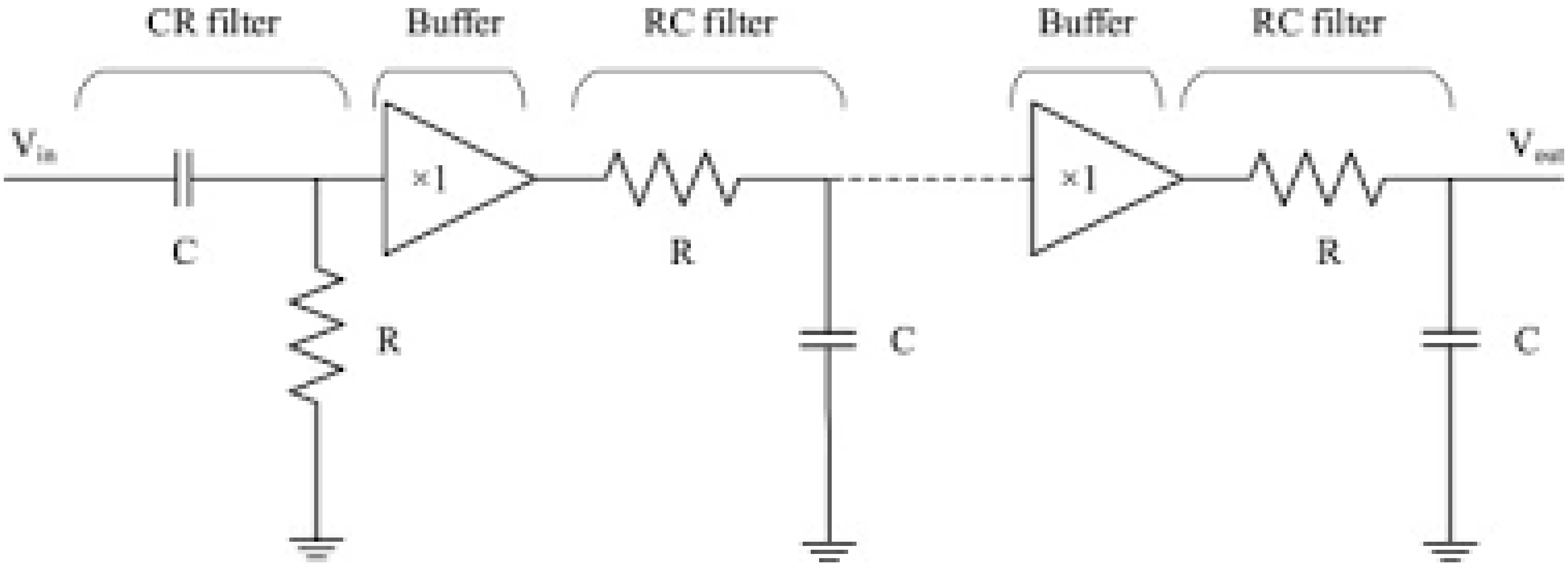


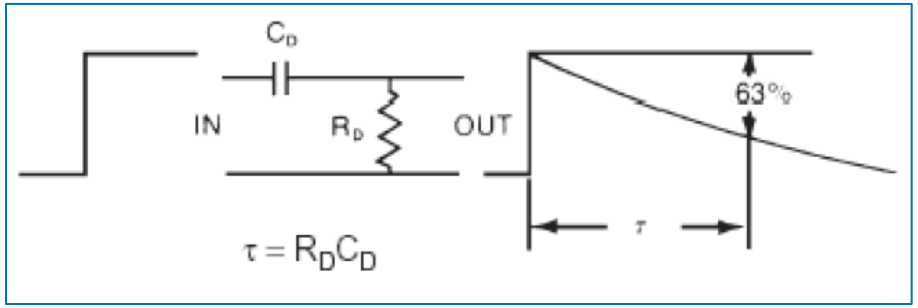
Calibration and simulation of Front end amplifiers



Before been so happy that you find a noise very low!!,
make sure your circuit is still amplifying the signal

CR*RC$_n$ shapers (filters overview)





CR stage after the preamp

$$V_{in}(t) = \frac{Q(t)}{C} + V_{out}(t) \;\Rightarrow\; \frac{dV_{in}(t)}{dt} = \frac{i(t)}{C} + \frac{dV_{out}(t)}{dt}$$

by $V_{out}(t) = i(t)R$ and $\tau = RC$, $\quad \tau\frac{dV_{in}(t)}{dt} = V_{out}(t) + \tau\frac{dV_{out}(t)}{dt}$

Assuming the zero initial condition, taking Laplace transform leads to

$$V_{out}(s) = \frac{\tau s}{1+\tau s} V_{in}(s) = G_{CR}(s) V_{in}(s)$$

For the step function input

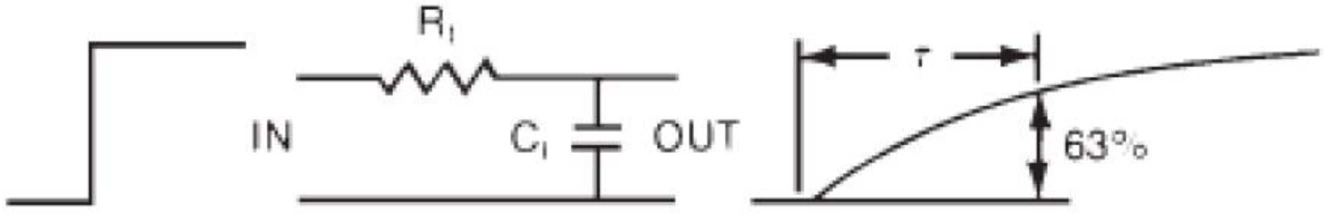
$$V_{in}(t) = \begin{matrix} V_0 \;\; (t>0) \\ 0 \;\; (t \le 0) \end{matrix} \;\Rightarrow\; V_{in}(s) = L[V_{in}(t)] = \frac{V_0}{s}$$

the output signal becomes

$$V_{out}(s) = \frac{\tau}{1+\tau s} V_0 \;\Rightarrow\; V_{out}(t) = V_0 e^{-t/\tau}$$

$$G_{CR}(i\omega) = \frac{i\omega\tau}{1+i\omega\tau} \;\Rightarrow\; \left|G_{CR}(i\omega)\right| = \frac{\omega\tau}{\sqrt{1+\omega^2\tau^2}}$$

RC stage of the shaper



$\tau = R_I C_I$

$$V_{in}(t) = i(t)R + V_{out}(t) \text{ and } i(t) = \frac{dQ(t)}{dt} = C\frac{dV_{out}(t)}{dt}$$

$$\Rightarrow V_{in}(t) = \tau\frac{dV_{out}(t)}{dt} + V_{out}(t) \Rightarrow V_{out}(s) = \frac{1}{1+\tau s}V_{in}(s) = G_{RC}(s)V_{in}(s)$$

Output signal for the step function input:

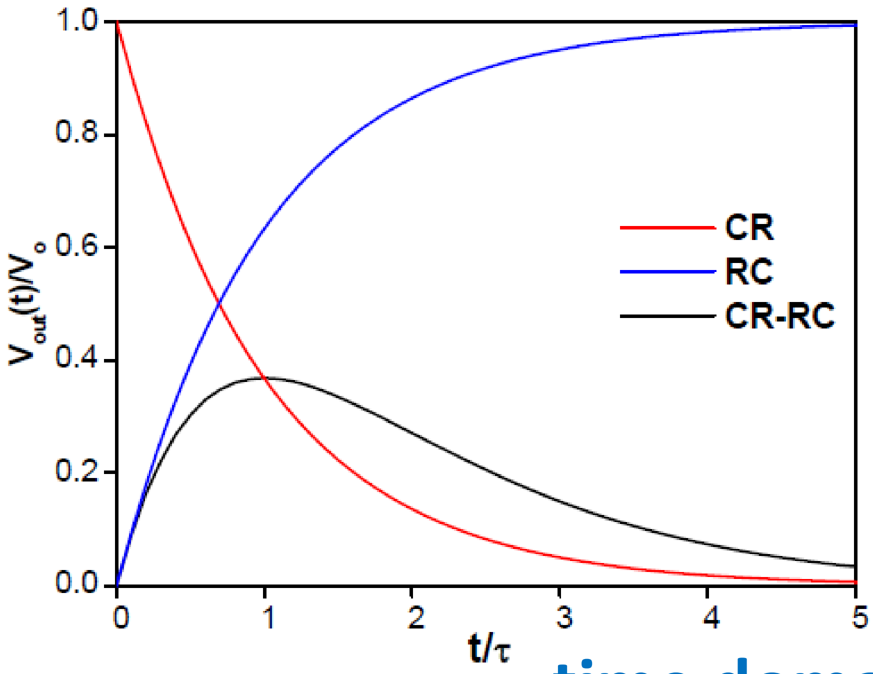
$$V_{out}(s) = \frac{1}{1+\tau s}\frac{V_0}{s} \Rightarrow V_{out}(t) = V_0(1 - e^{-t/\tau})$$

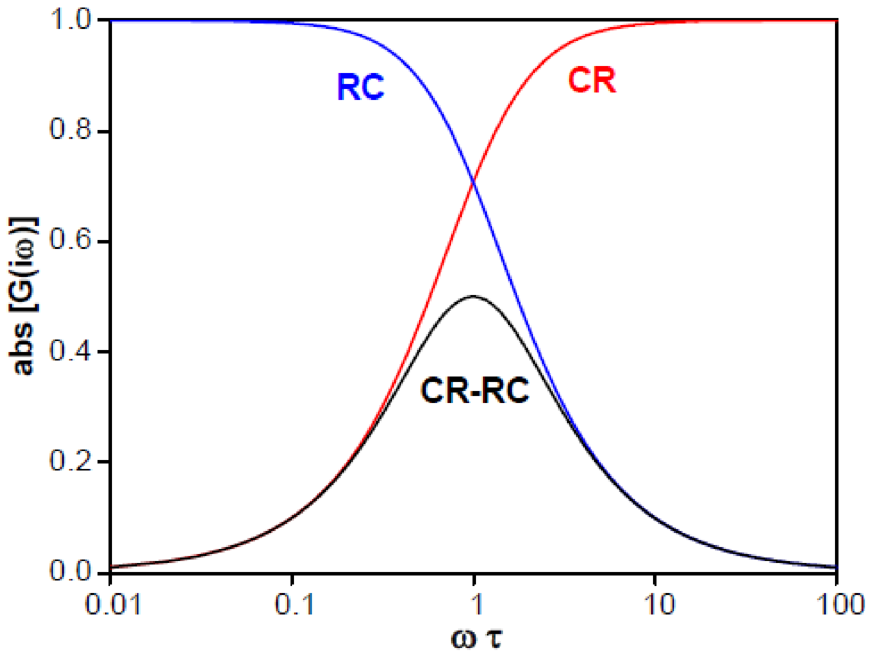
Frequency domain transfer function:

$$G_{RC}(i\omega) = \frac{1}{1+i\omega\tau} \Rightarrow |G_{RC}(i\omega)| = \frac{1}{\sqrt{1+\omega^2\tau^2}}$$

CR+RC transfert functions for a step input

This step stand for the integrator output signal





time domaine

$$V_{out}(t) = \frac{V_0 \tau_1}{\tau_1 - \tau_2}\left(e^{-t/\tau_1} - e^{-t/\tau_2}\right)$$

if $\tau_1 = \tau_2$ ⇒ $V_{out}(t) = \frac{v_0}{\tau} t e^{-t/\tau}$

Frequency domaine:

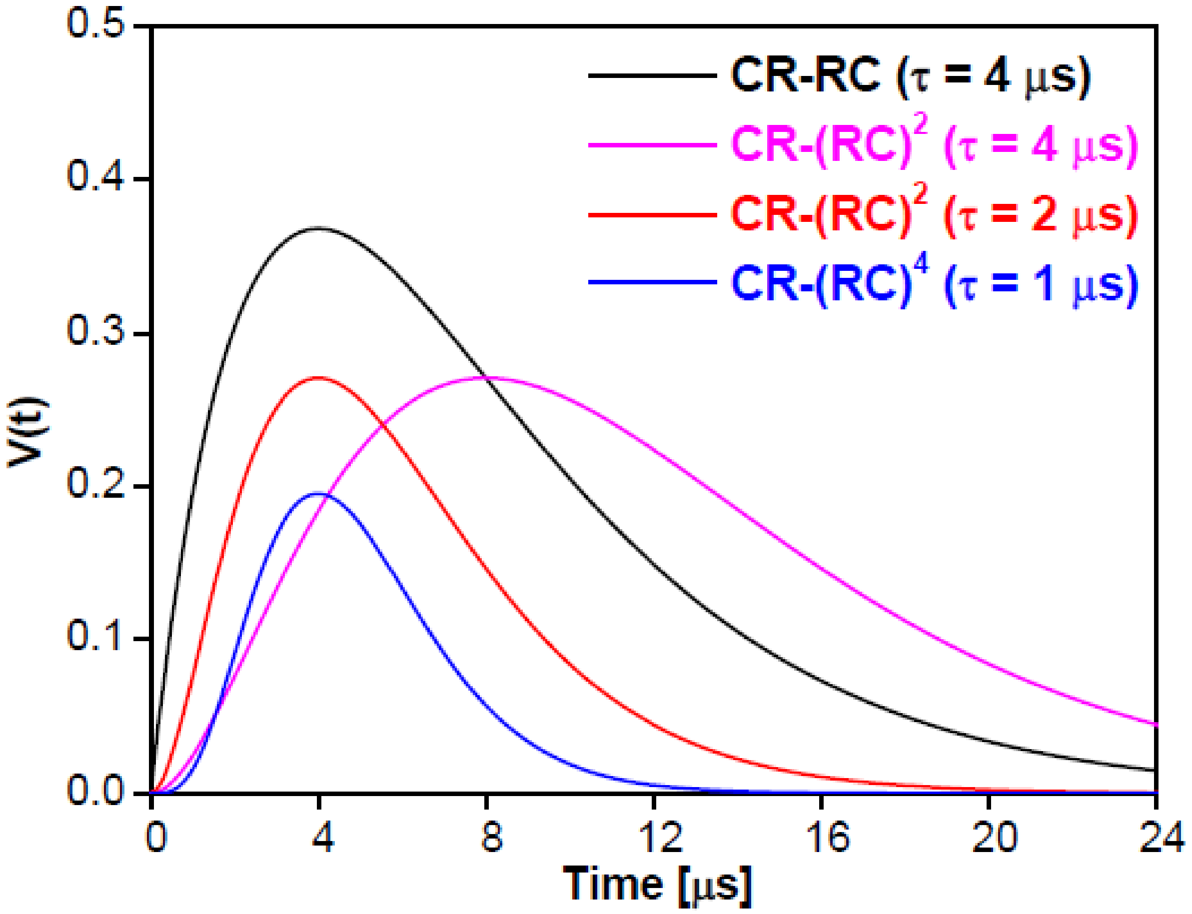
Pass high * pass low

$$G_{CR-RC}(s) = \frac{1}{(1+\tau_2 s)} \frac{\tau_1 s}{(1+\tau_1 s)}$$

Filter / Shaper

first order? second? Why?

CR*RC^n filters or Semi-Gaussian pulse shaping

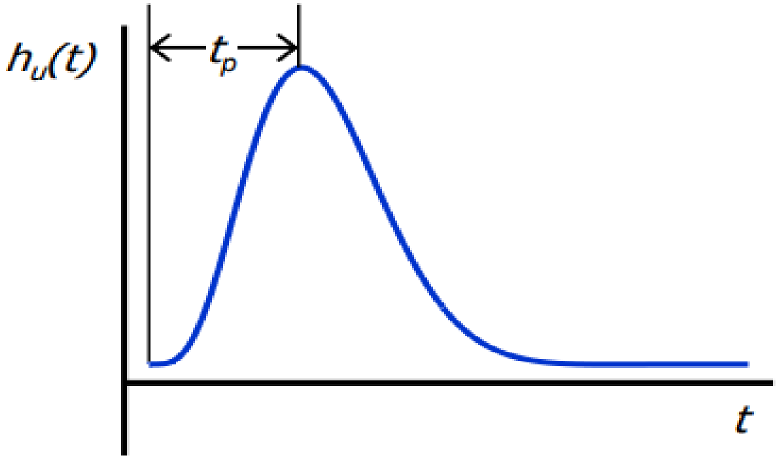


If a single CR high-pass filter is followed by several stages of RC integration, the output pulse shape becomes close to Gaussian amplifiers shaping, in this way are called **semi-Gaussian shaper**. Its output pulse is given by:

$$V_{out}(t) \propto \left(\frac{t}{\tau}\right)^n e^{-t/\tau}$$

The peaking time in this case is equal to *n**τ.

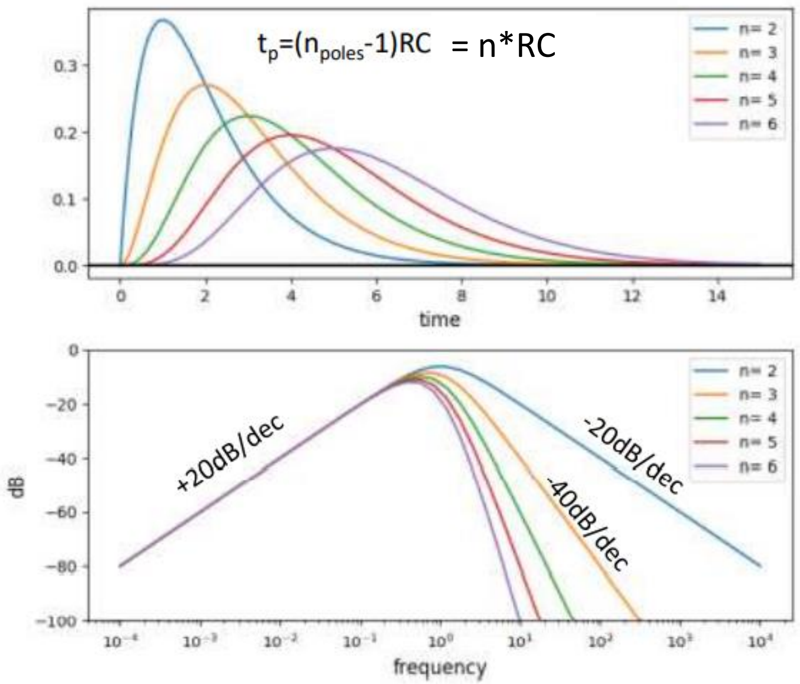
CR*RC^n filters impulse response (summary)



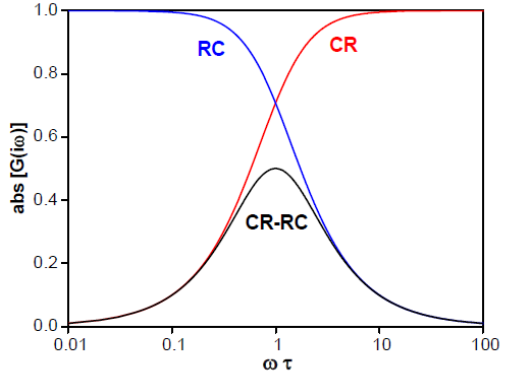
impulse response: $h_u(t) = \frac{1}{n!}\left(\frac{t}{\tau}\right)^n e^{-t/\tau}$

peaking time: $t_p = n\tau$

CR + n times RC low pass



CR+RC low pass

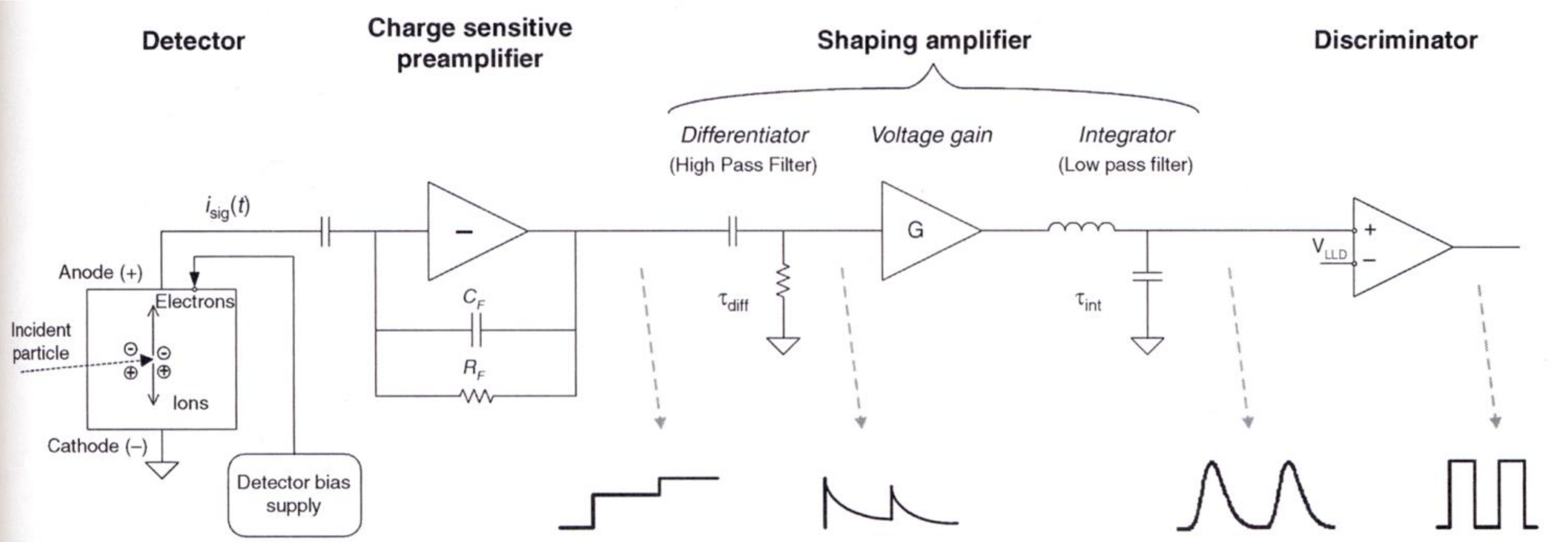


transfer function: $H_u(s) = \frac{s\tau}{(1+s\tau)^{n+1}}$

- CR=> 1 zero + 1 pole
- n*RC => n poles

So (n+1) poles in total

Signal shape following read out steps when 2 successive pulses (II)

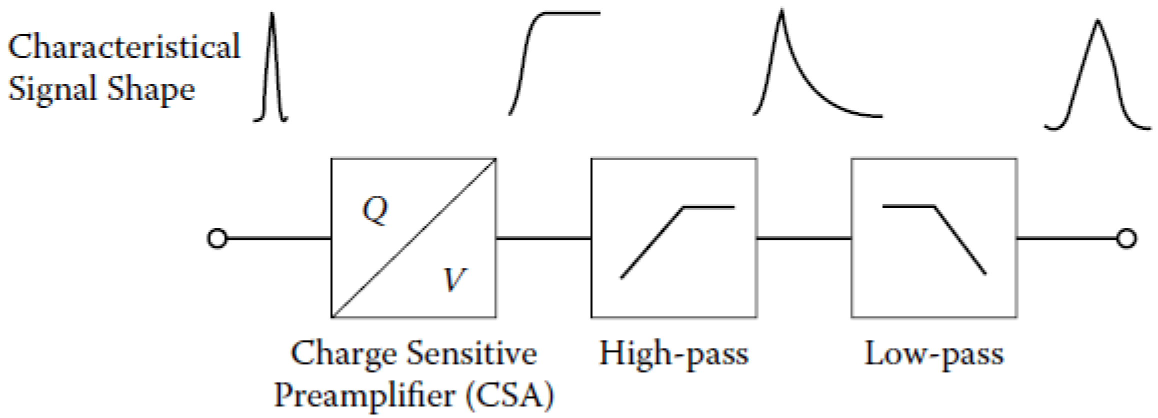


Schematic of simple signal processing electronics. This circuit is suitable for use, as shown, in many applications and is conceptually similar to more complex circuits. These elements are discussed in greater detail in Chapter 17. (Courtesy of R. Redus, Amptek, Inc.)

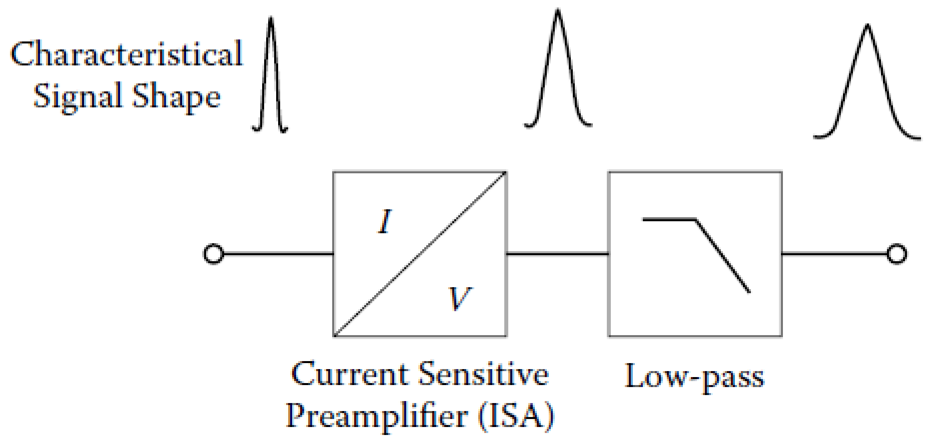
Courtesy, Glenn F. « radiation detection »

Simplified Signal shape throughout electronic readout

Charge sensitive Signal conditioning



Current sensitive Signal conditioning



Up til now we describe more the signal's amplitude and it's shaping

LET'S CONSIDER NOW THE NOISE ISSUE

- **FIRST: FOR AN AMPLIFIER IN GENERAL**
- **SECOND: PULSE PREAMPLIFIERS**

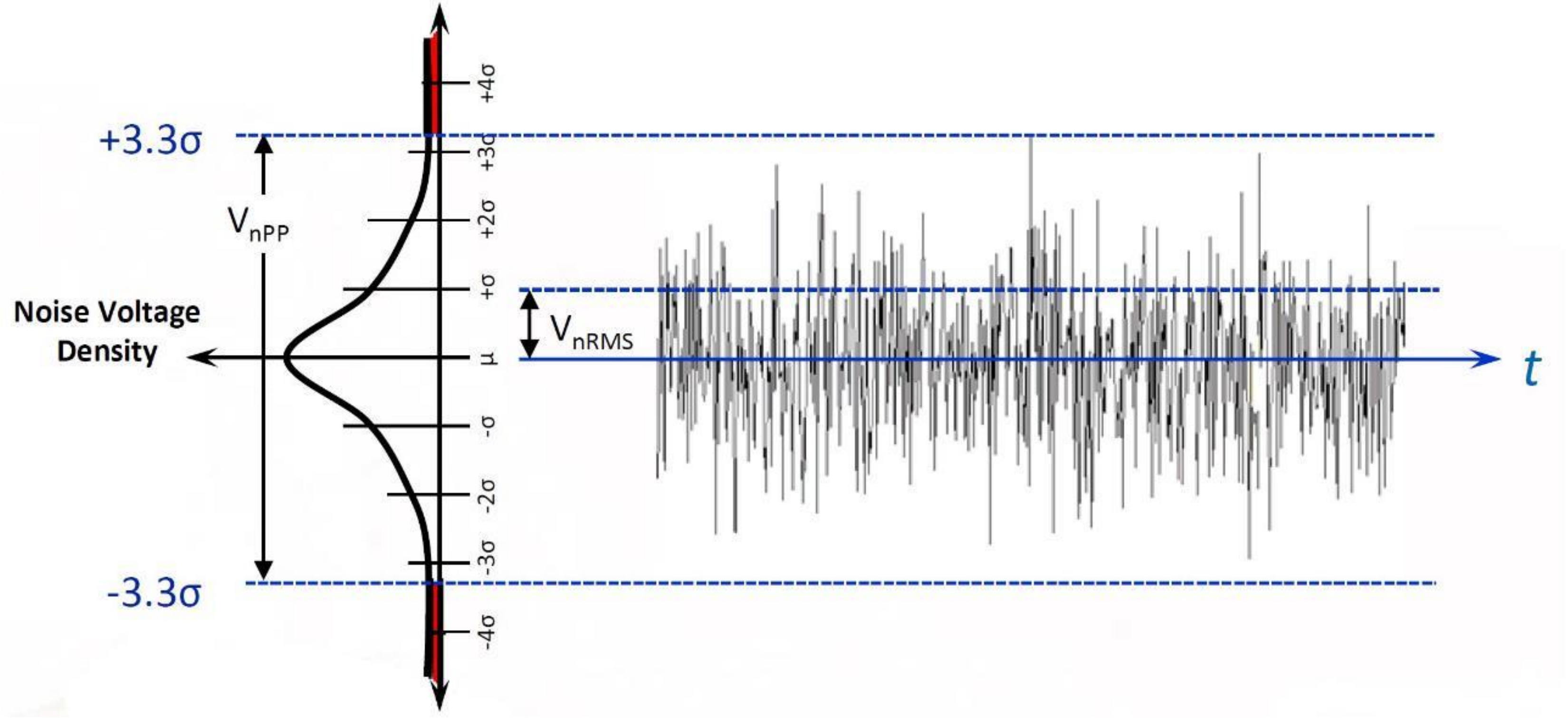
WHAT IS NOISE ?

- **What is noise?**
  - **Noise is any undesired signal that masks the signal of interest.**
    - Unwanted disturbance that interferes with a desired signal
    - External: power supply & substrate coupling, crosstalk, EMI, etc.
    - Internal: random fluctuations that result from the physics of the devices or materials
    - Smallest detectable signal, signal-to-noise ratio (SNR), and dynamic range are determined by noise

$$SNR = \frac{P_{signal}}{P_{noise}} = \frac{V^2_{rms,\,signal}}{V^2_{rms,\,noise}}$$

We will look at internal noise sources and how they affect key performance metrics.

How a noise is quantified: amplitude versus time

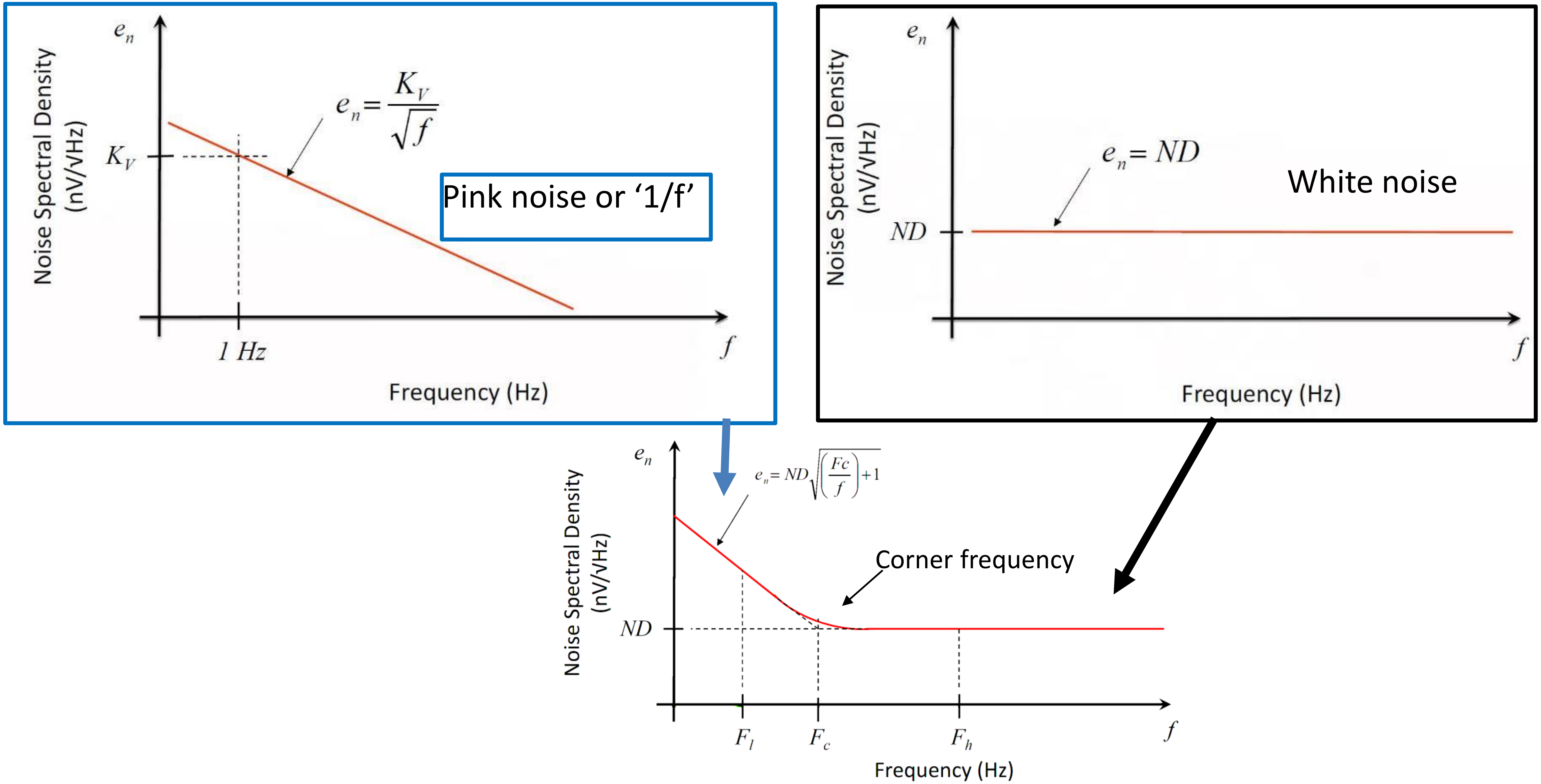


$Vn_{pp} = 6.6 \times Vn_{RMS}$

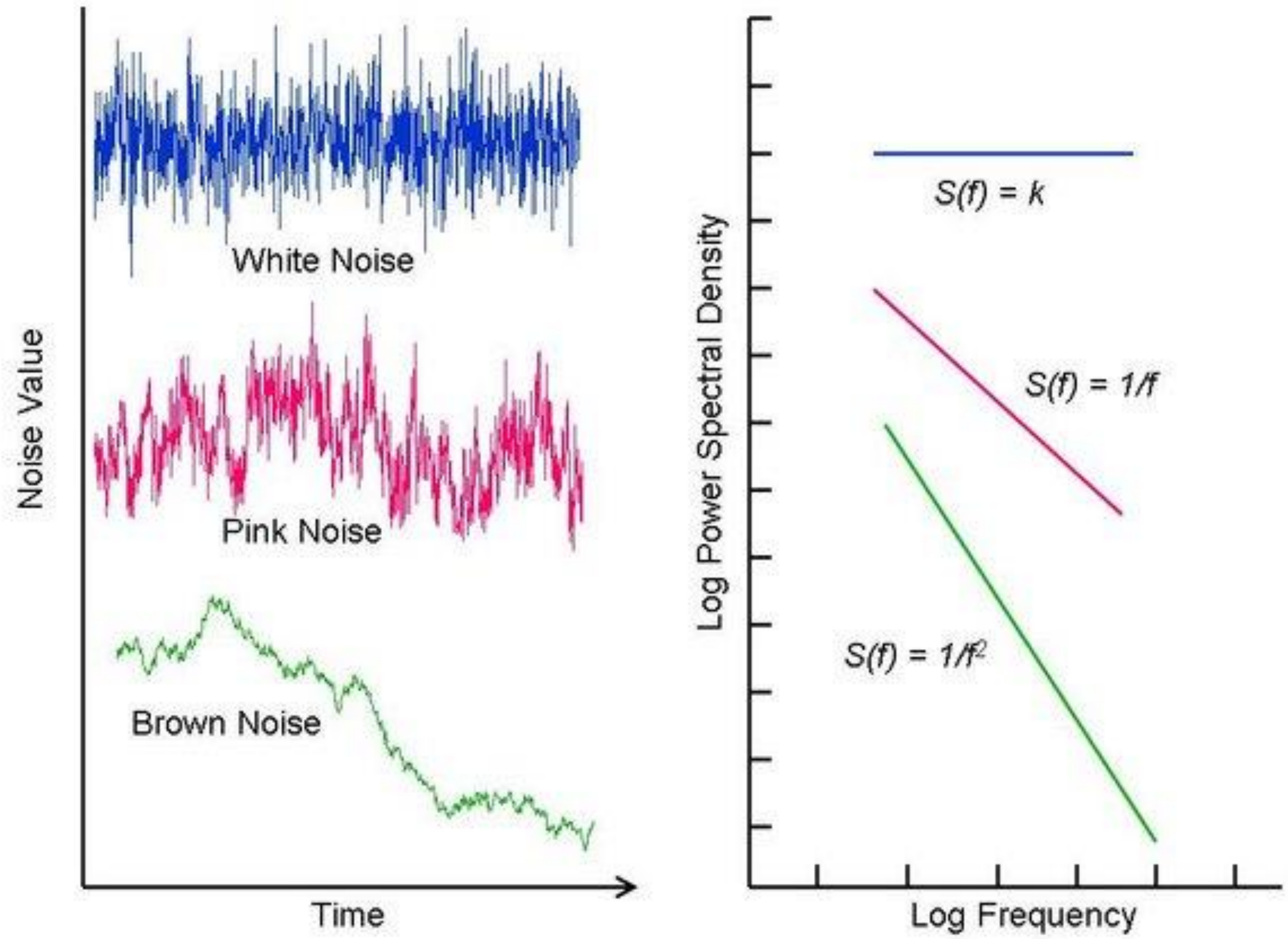
Root Mean Square (RMS)

Peak-to-Peak (PP)

Noise specifications: **spectral density**(frequency domain)

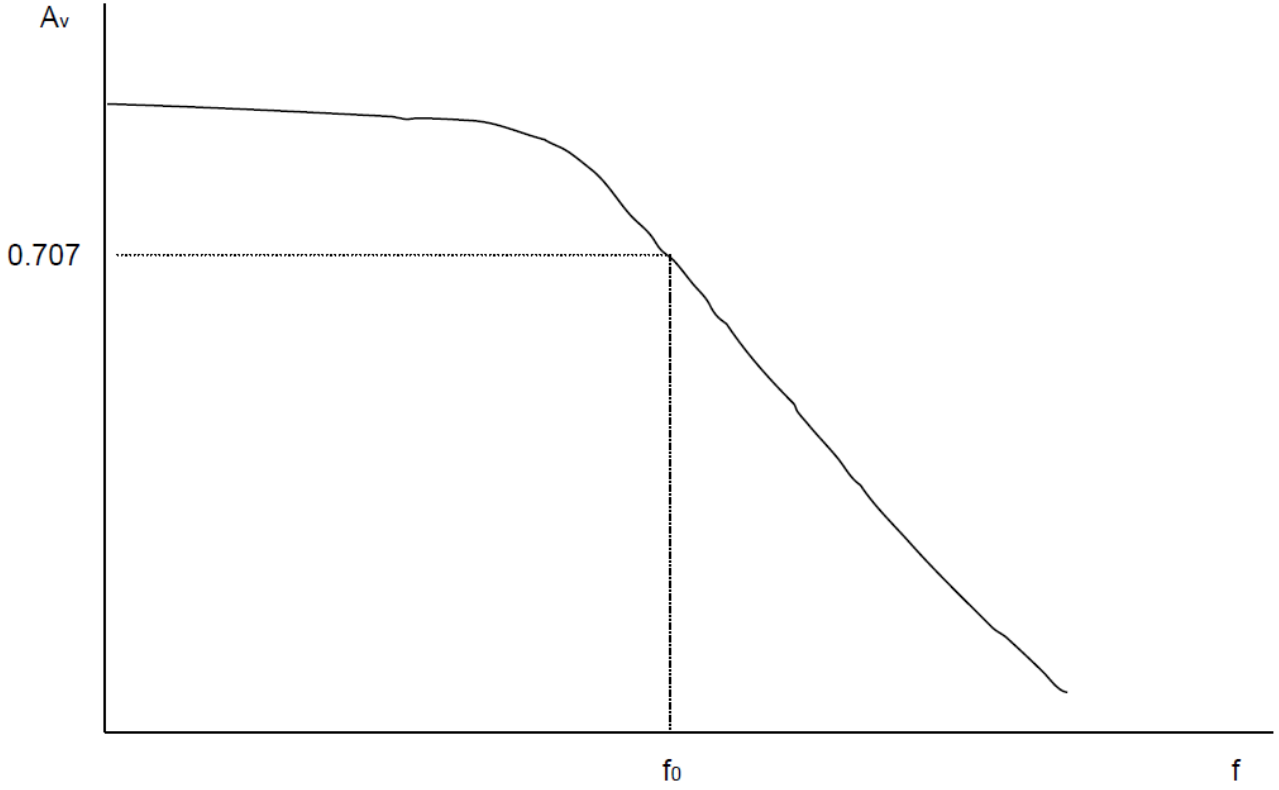


Noise in general: time & frequency domain



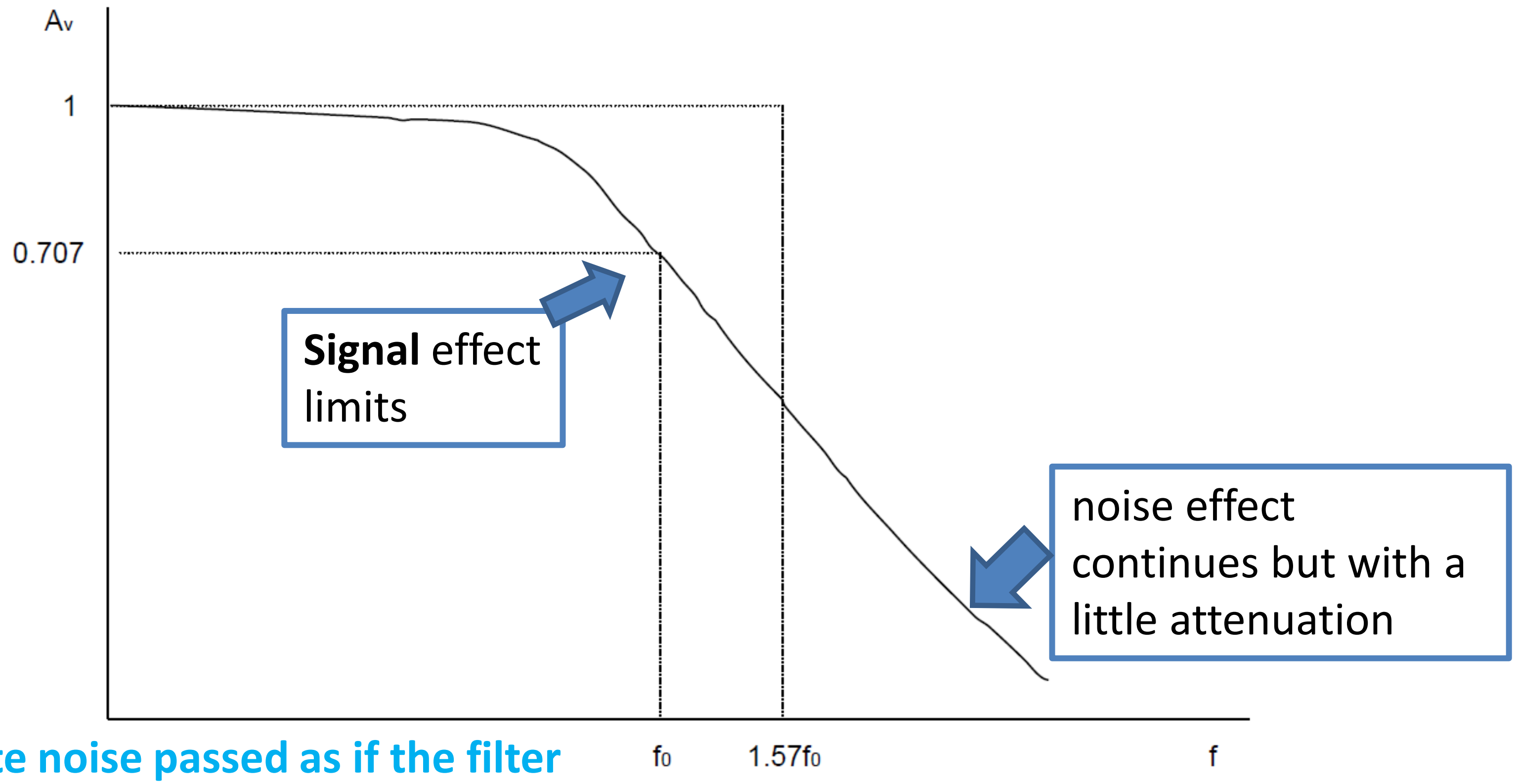
Noise Effective BW (NEB) of an amplifier (filter)

Lets consider a general *low pass* amplification (filter) system;
What happen to a white noise located at the input of such amplifier?
Is it amplified exactly as the signal?



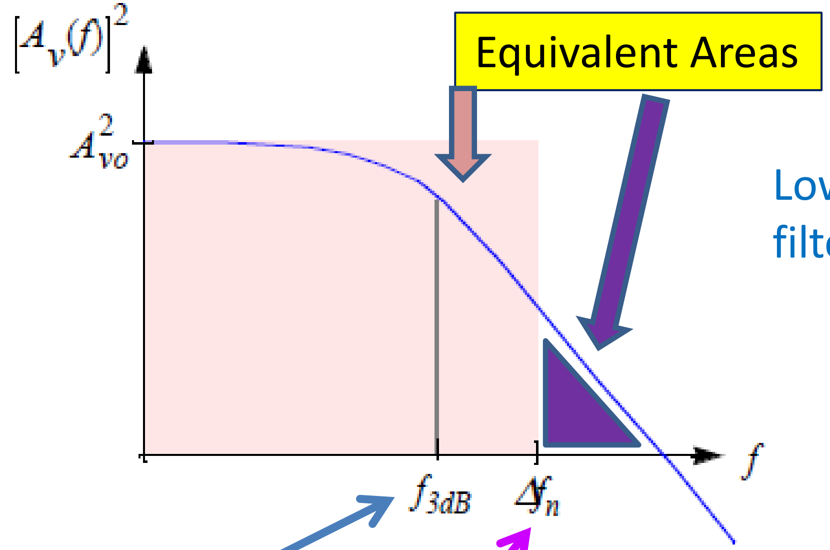
Noise bandwidth

Lets consider a low pass amplification system;
What is the effect of a larger bandwidth white noise located at the input of such amplifier?



The white noise passed as if the filter were a brick wall type but with a cutoff freq 1.57 times larger

Equivalent Noise Bandwidth # signal bandwidth



Signal concern

Noise concern up to 0db

$$A_{vo}^2 \cdot \Delta f_n = \int_0^\infty [A_v(f)]^2 df$$

No. Poles	NOISE BW
1	$\Delta f_n = \frac{\pi}{2} \cdot \Delta f_{3dB}$
2	$\Delta f_n = 1.22 \cdot \Delta f_{3dB}$
3	$\Delta f_n = 1.16 \cdot \Delta f_{3dB}$
2nd-order BPF	$\Delta f_n = \frac{\pi}{2} \cdot \Delta f_{3dB}$

Low pass filters (poles 1, 2, 3)

This case refers to a Band-Pass Filters (1 zero + 2poles)

- Noise bandwidth is defined for a brickwall transfer function
- Noise bandwidth is not the same as 3dB bandwidth

Noise Bandwidth improves when number of poles increase

Keep it in mind and make the link later with $CR*RC_n$ filtering

Optimizing Filtering <=> Optimize NEB

For Maximally flat (Butterworth) where $f_0=f_{3dB}$

$$NEB = \left(\int_0^\infty \frac{df}{1+(f/f_0)^{2n}}\right)$$

n	NEB
1	$1.57f_0$
2	$1.11f_0$
3	$1.05f_0$
4	$1.025f_0$

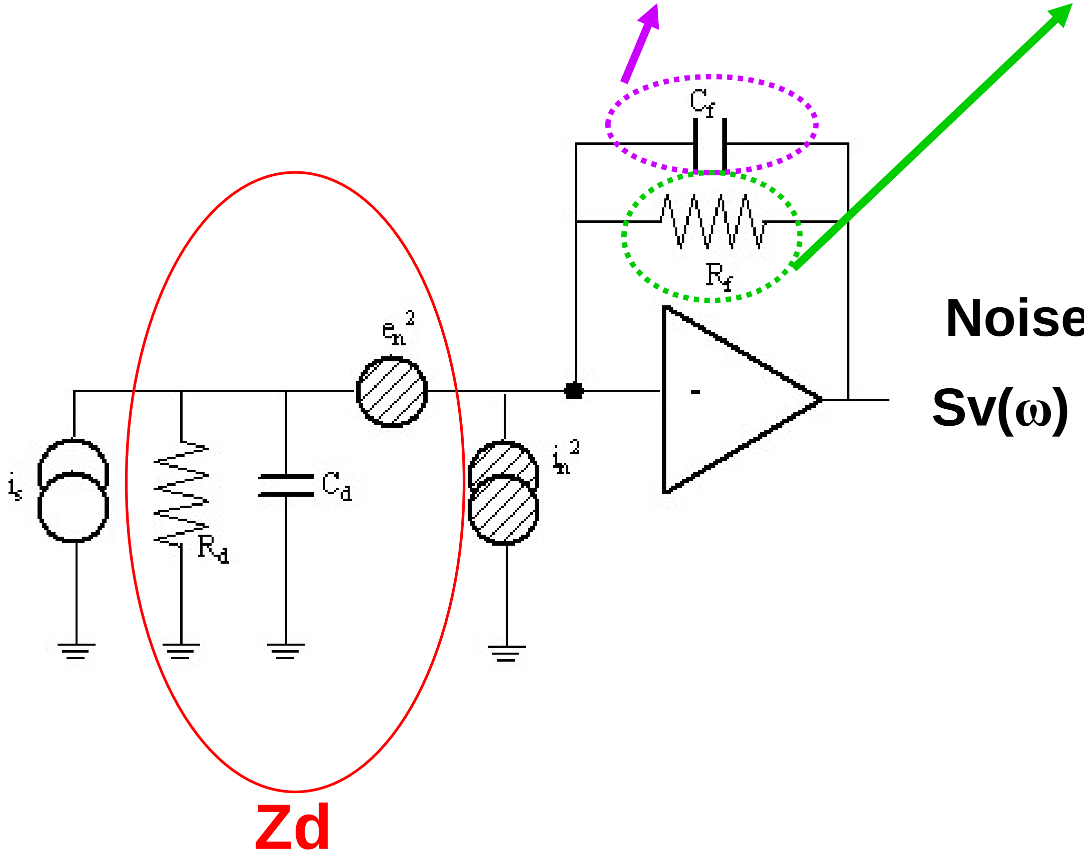
But higher order n means: More complicated and/or more power budget

In spectroscopie, one considers often **n=2** as a compromise

High order filter is good(0.5dB improvement)

Charge / Current preamplifiers

Preamplifier: charges / Current output noise



Noise spectrum at the output

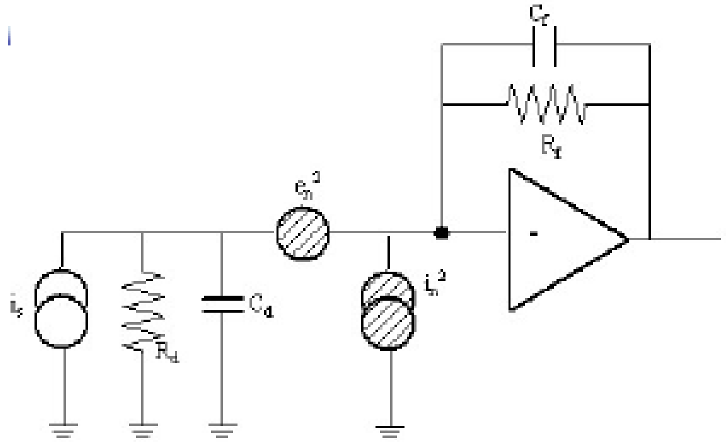
$$Sv(\omega) = (in^2 + en^2/|Zd|^2) / \omega^2 Cf^2$$

$$= in^2 /\omega^2 C_f^2 + en^2 C_d^2 / Cf^2$$

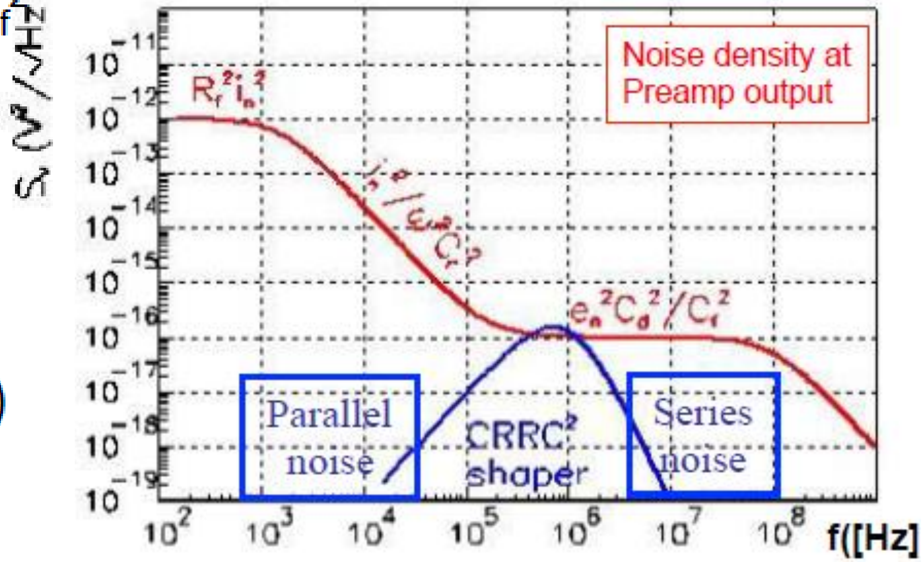
If we neglect Rd and
we translate en to a current
By doing en/Zd

Noise issues for charge preamp: frequency domaine

- 2 noise generators at the input
  - Parallel noise : (i_n^2) (leakage currents)
  - Series noise : (e_n^2) (preamp)
- Output noise spectral density :
  - $S_v(\omega) = (i_n^2 + e_n^2/|Z_d|^2) / \omega^2 C_f^2$
    $= i_n^2 / \omega^2 C_f^2 + e_n^2 C_d^2/C_f^2$
  - Parallel noise in $1/\omega^2$
  - Series noise is flat, with a « noise gain » of C_d/C_f
- *rms* noise V_n
  - $V_n^2 = \int Sv(\omega)\, d\omega/2\pi \rightarrow \infty$ (!)
  - Benefit of shaping…



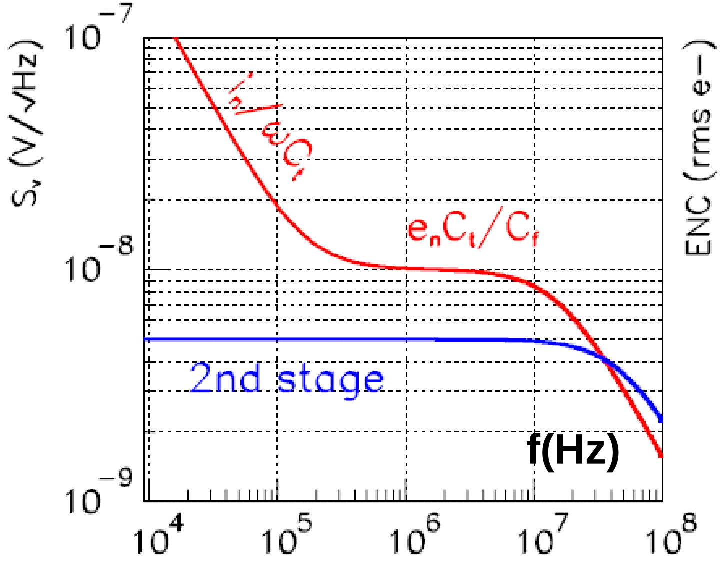
Noise generators in charge preamp

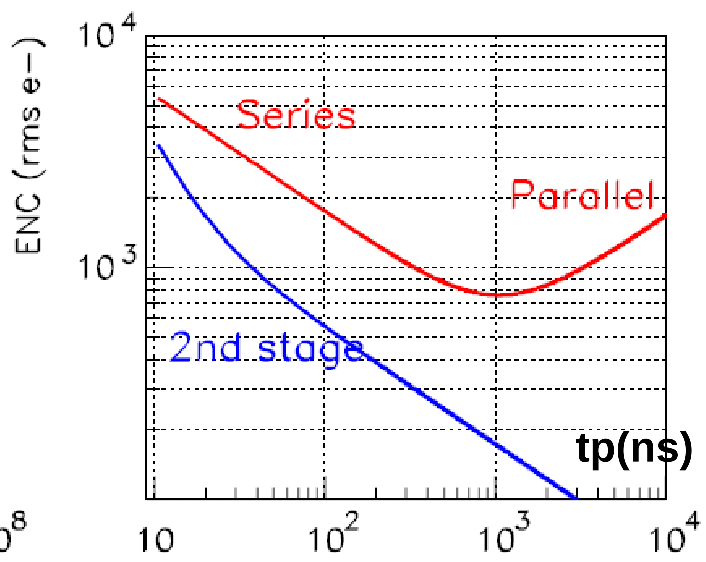


The methode here is to Transfert each input To current domain, then Multiply by the feedback

Ref. C. de la Taille

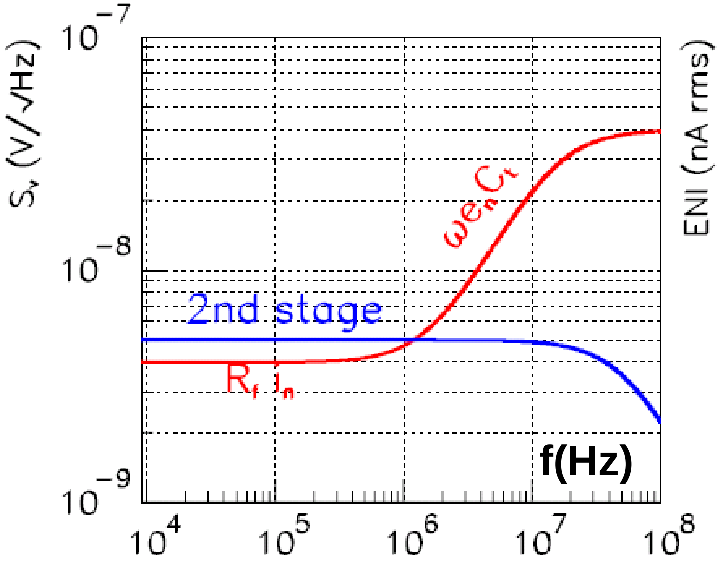
Output spectrum and it's equivalent input

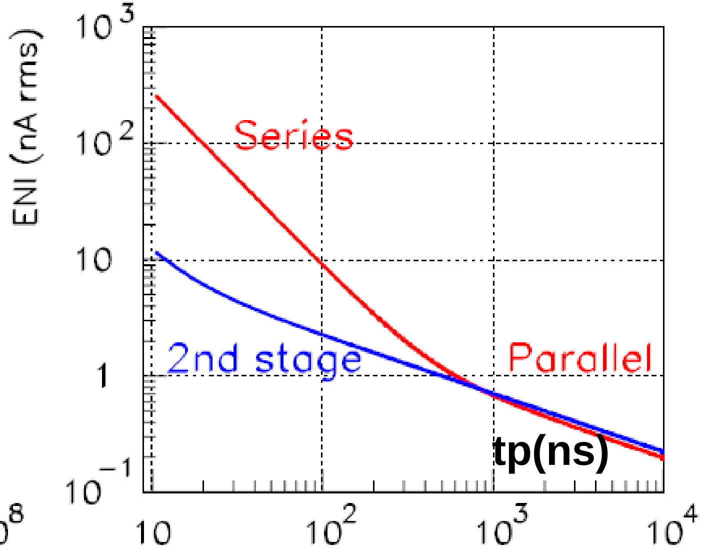




CHARGE Preamp

Output spectrum **(f)** — — — — Equivalent input noise **(tp)**

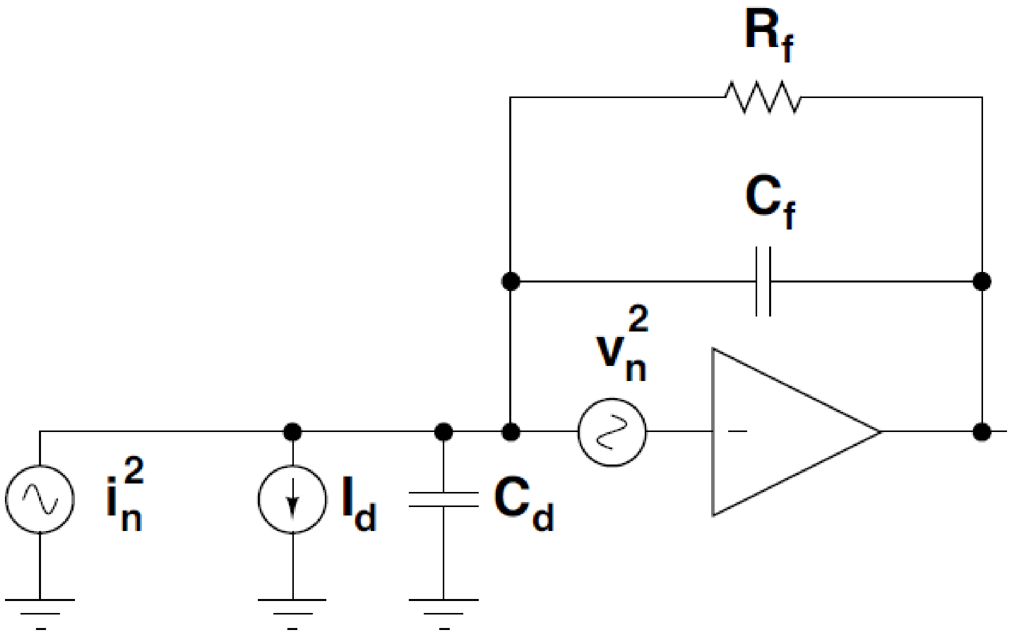


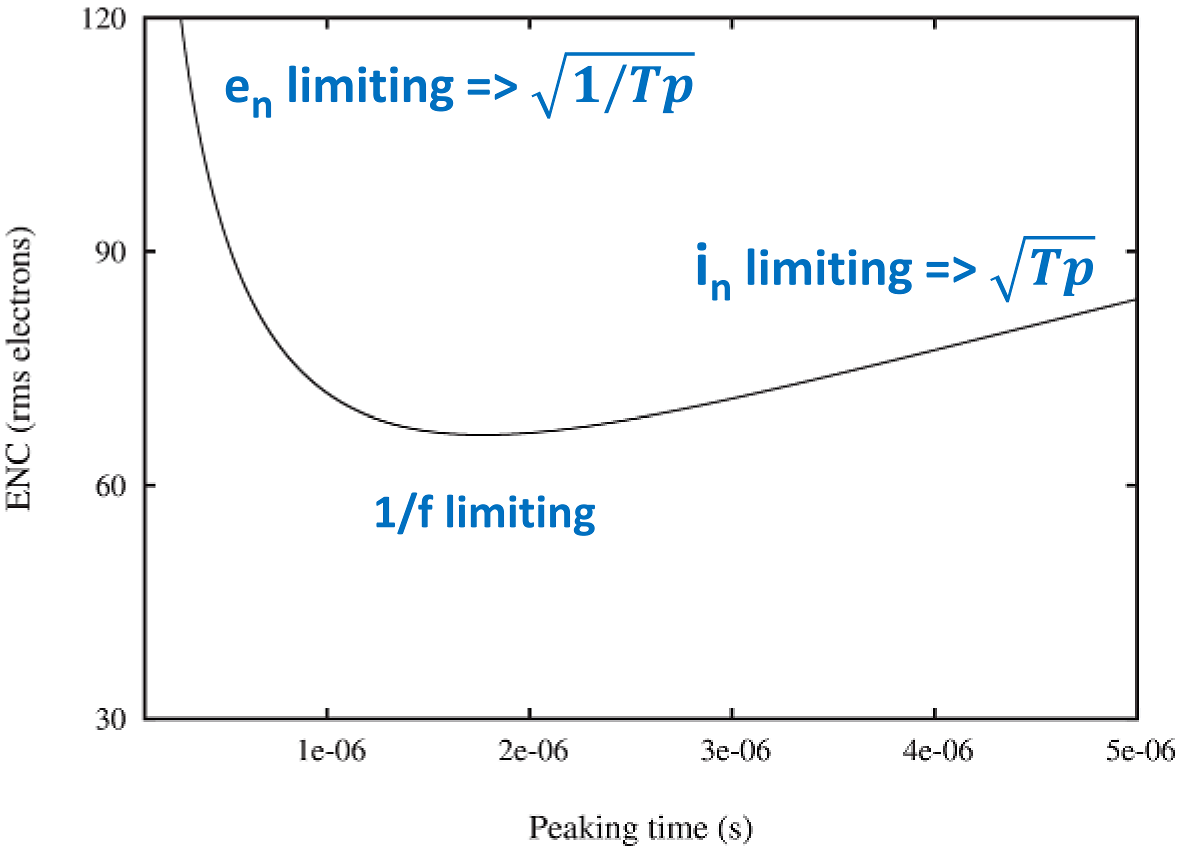


See next how tp Appears in noise equations of ENC

CURRENT Preamp

Noise issues in charge preamp: *time domaine*

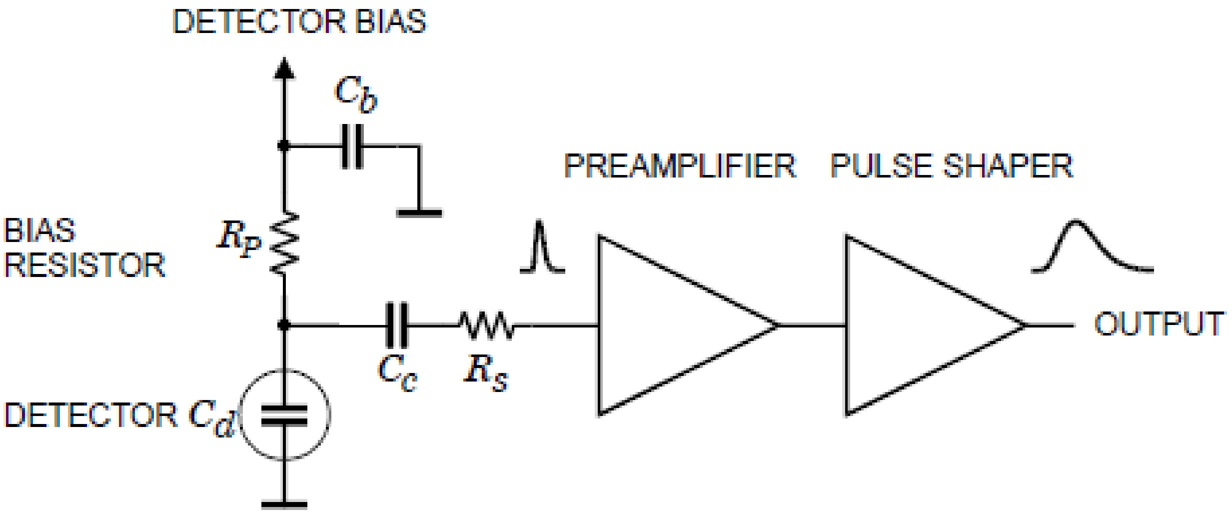


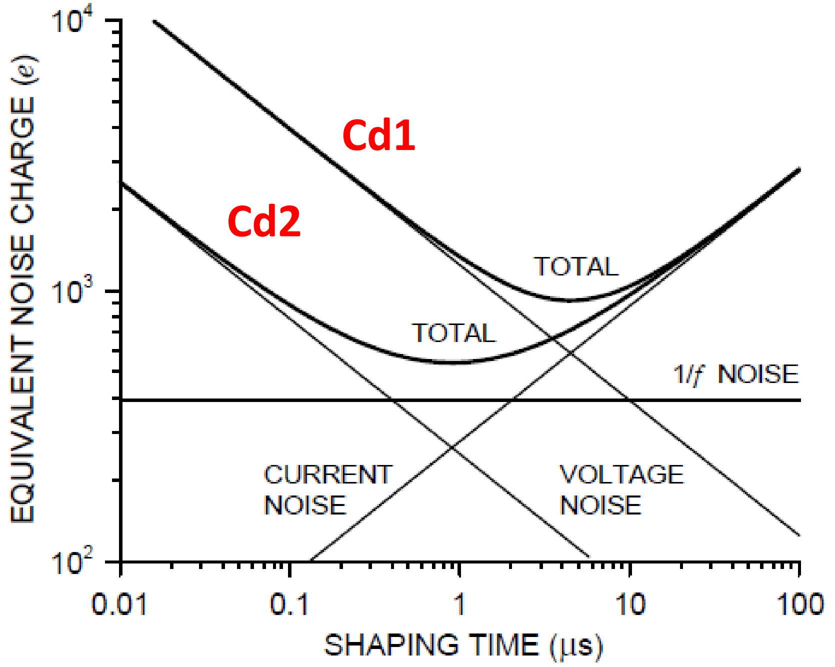


$$ENC^2 = (C_d + C_{in})^2 \left(A_w v_n^2 \frac{1}{T_p} + A_f K_f \right) + A_p i_n^2 T_p$$

Increase with Cd

Practical summary about noise in charge integrator





ENC using a simple CR-RC shaper with peaking time T:

$$Q_n^2 \approx \left[\left(2q_e I_d + \frac{4kT}{R_P} + i_{na}^2\right)\cdot T \;+\; \left(4kTR_S + e_{na}^2\right)\cdot\frac{C_d^2}{T} \;+\; 4A_f C_d^2\right]$$

current noise	voltage noise	$1/f$ noise
$\propto \tau$	$\propto 1/T$	independent of T
independent of C_d	$\propto C_d^2$	$\propto C_d^2$

Helmuth Spieler
LBNL

Preamp non idealities: GBW limitations

General case: feedback impedance is combination of resistor and capacitor

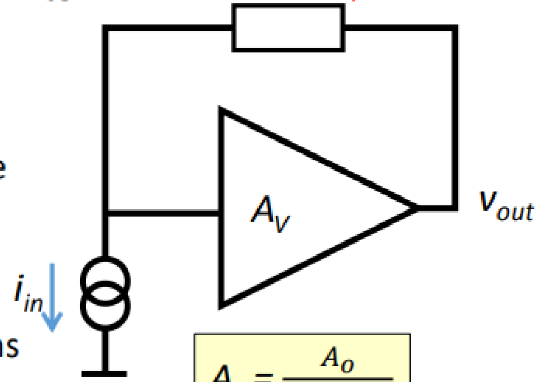
"Real" amplifier = ideal voltage amplifier with finite gain-bandwidth product

Feedback ***lowers input impedance*** to minimize voltage swing on input node:

- Improve linearity and speed
- Avoid destabilizing sensor
- Mitigate crosstalk in multi-electrode systems

$$Z_F = \frac{R_F}{1+s/\omega_F}$$

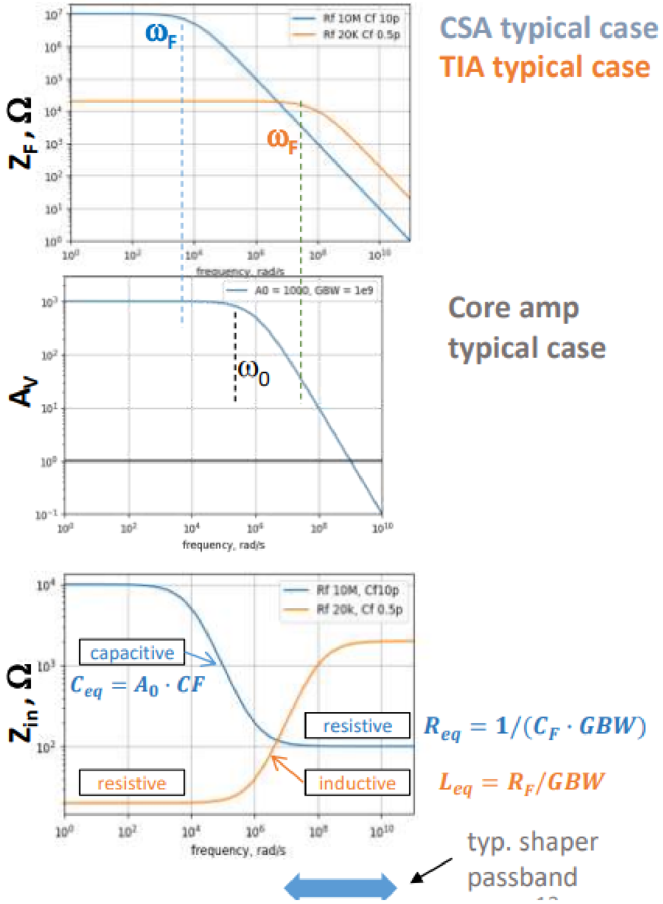
$$\omega_F = 1/R_F C_F$$



$$A_V = \frac{A_0}{1+s/\omega_0}$$

$$\omega_0 = GBP/A_0$$

$$Z_{in} = \frac{Z_F}{1+A_V}$$

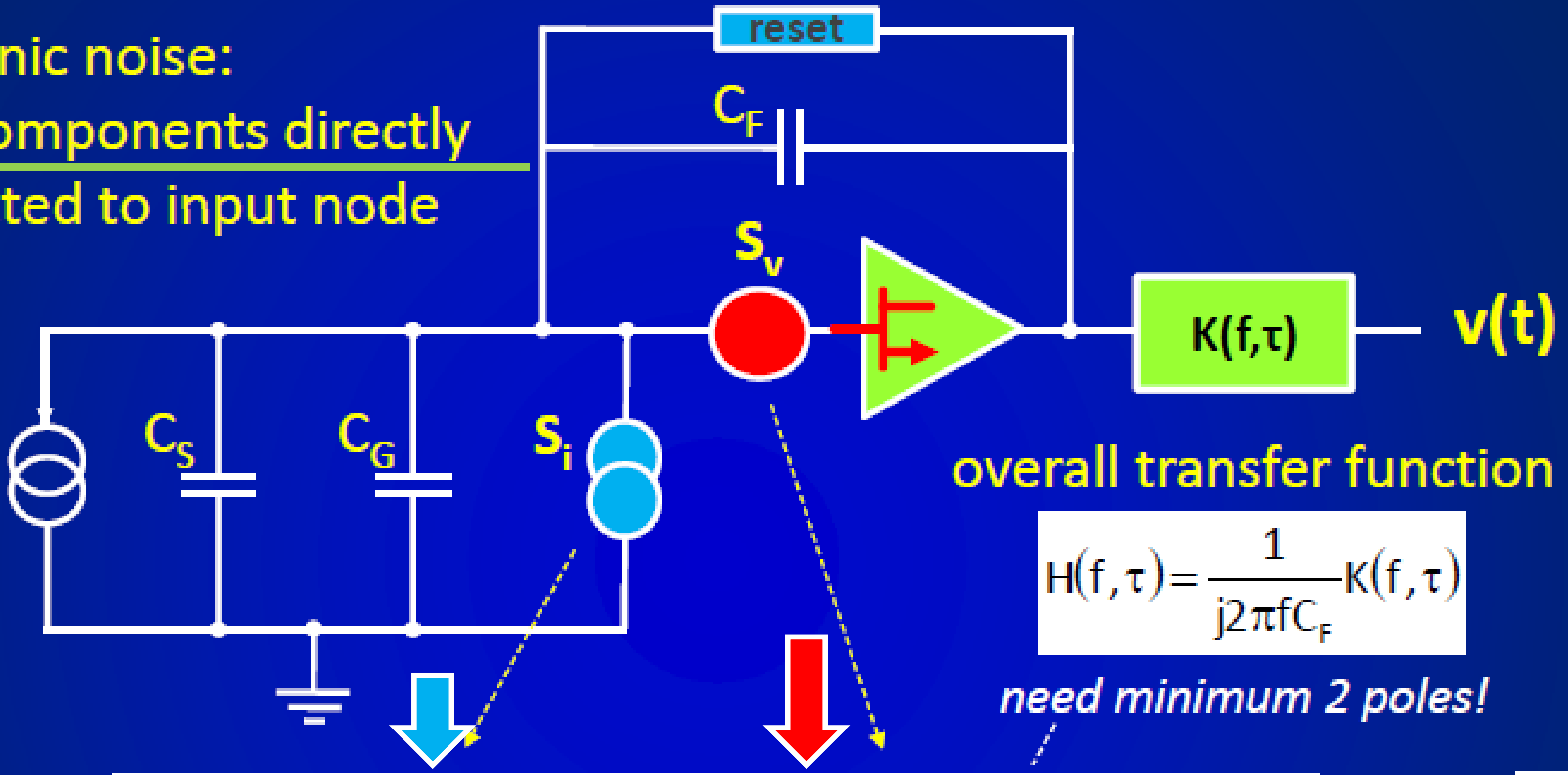


One stage preamplifier scheme

- The main contributor to the total noise is the **preamp input transistor**. We consider next the contribution of this transistor to the equivalent noise $\mathbf{e_n}$ and $\mathbf{i_n}$

Sources of Electronic Noise

Electronic noise:
from components directly
connected to input node



overall transfer function

$$H(f,\tau)=\frac{1}{j2\pi fC_F}K(f,\tau)$$

need minimum 2 poles!

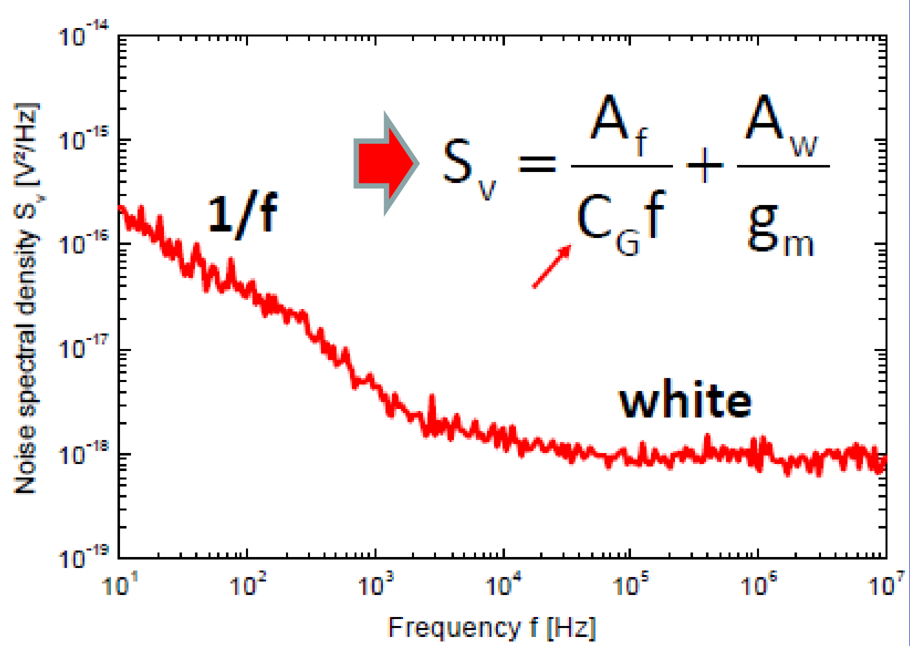
$$ENC^2 \cong \frac{\int_0^\infty S_i\left|H(f,\tau)\right|^2 df + \int_0^\infty S_v\omega^2\left(C_S+C_G\right)^2\left|H(f)\right|^2 df}{\left.h(t)\right|^2_{max}}$$

Time-variant → time-domain analysis (noise weighting function)

Here, time appears in equations

De Geromino

Noise from Input Transistor



$$S_v = \frac{A_f}{C_G f} + \frac{A_w}{g_m}$$

C_G intrinsic gate capacitance

proportional to the gate size

$C_G = C_S$ (capacitive matching)

From input transistor:

$$ENC_v^2 = a_f A_f \frac{(C_S + C_G)^2}{C_G} + \frac{a_w}{\tau} \frac{A_w}{g_m / C_G} \frac{(C_S + C_G)^2}{C_G}$$

ASIC: power constraints

f_{Tmax} f_T (max current)

Input Transistor in CMOS

From transistor's white noise:

$$ENC^2_{vw} \approx \frac{a_w}{\tau} \frac{A_w}{g_m(I_D)/C_G} \frac{(C_S + C_G)^2}{C_G}$$

Fix power = fix drain current I_D

→ **size (W,L) ?**

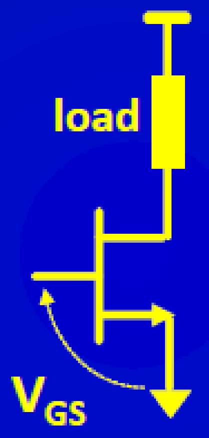
$V_{GS} >> V_{th}$ (strong inversion)

$$g_m(I_D) \approx \sqrt{\frac{2\mu C_{ox}}{n}\frac{W}{L} I_D} \propto \sqrt{\frac{C_G I_D}{L^2}}$$

↓

$$ENC^2_{vw} \propto \frac{L}{\sqrt{I_D}} \frac{(C_S + C_G)^2}{\sqrt{C_G}}$$

- minimum L
- $C_G = C_S/3$



$V_{GS} << V_{th}$ (weak inversion)

$$g_m(I_D) \approx \frac{I_D}{nV_T} \propto I_D$$

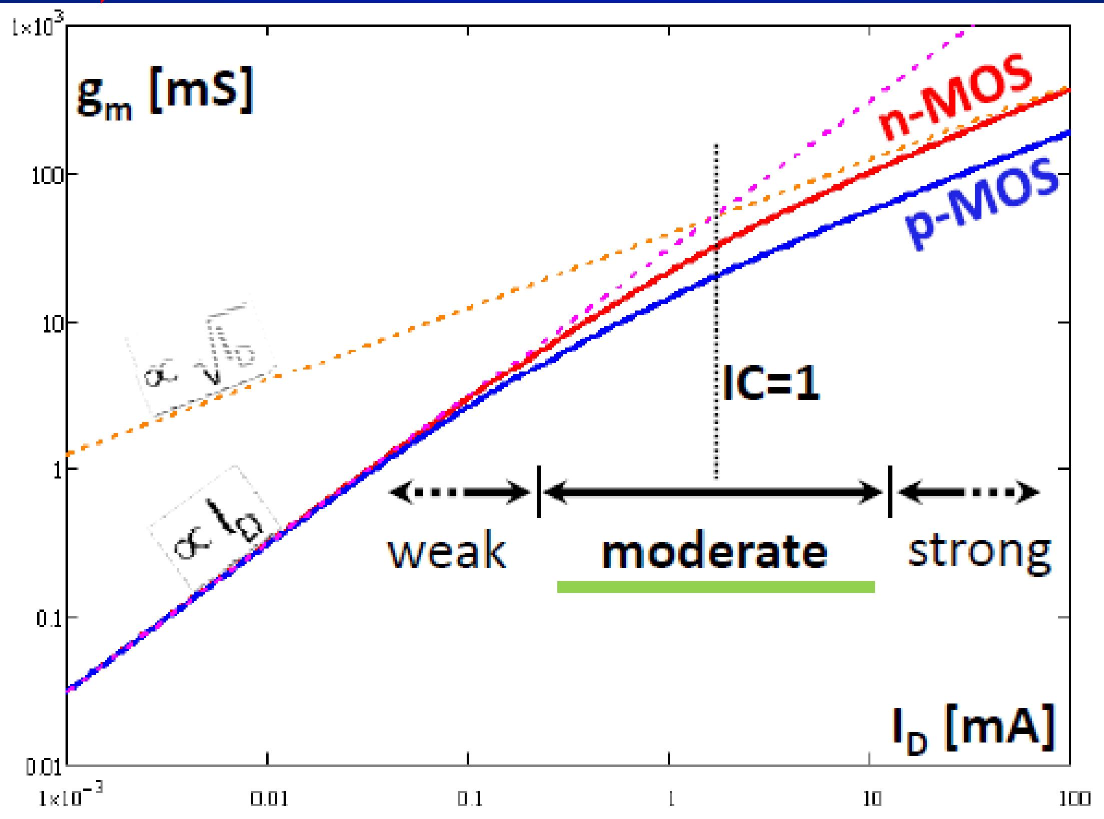
↓

$$ENC^2_{vw} \propto \frac{(C_S + C_G)^2}{I_D}$$

- independent of L
- $C_G = 0$ pushes back towards strong inversion

→ **$V_{GS} \approx V_{th}$ (moderate inversion): model?**

Moderate Inversion



Notice gm/Id efficiency

From EKV model

$$g_m(I_D) \approx \frac{I_D}{nV_T} \frac{\sqrt{1+4\cdot IC}-1}{2\cdot IC}$$

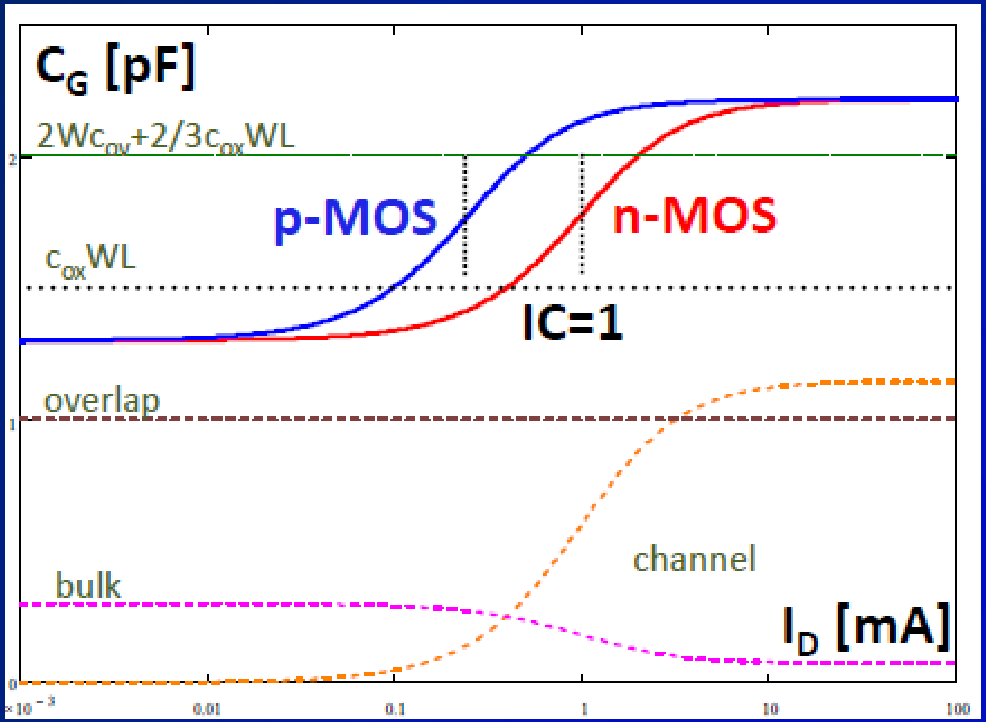
$$IC = \frac{L}{W} \frac{I_D}{2nV_T^2 \mu c_{ox}}$$

inversion coefficient

alternative: extract from simulators (BSIM)

De Geronimo, IEEE TNS 52, 2005

Gate Capacitance



~~$\propto C_{ox}WL$~~

$$C_G(I_D) \approx 2c_{ov}W + C_{ox}WL\left(\gamma_C(IC) + \frac{n-1}{n}\left[1-\gamma_C(IC)\right]\right)$$

$$\gamma_c(IC) \approx \left(\frac{3}{2} + \frac{1}{3}\frac{\sqrt{1+4\cdot IC}+1}{IC^2}\right)^{-2/3}$$

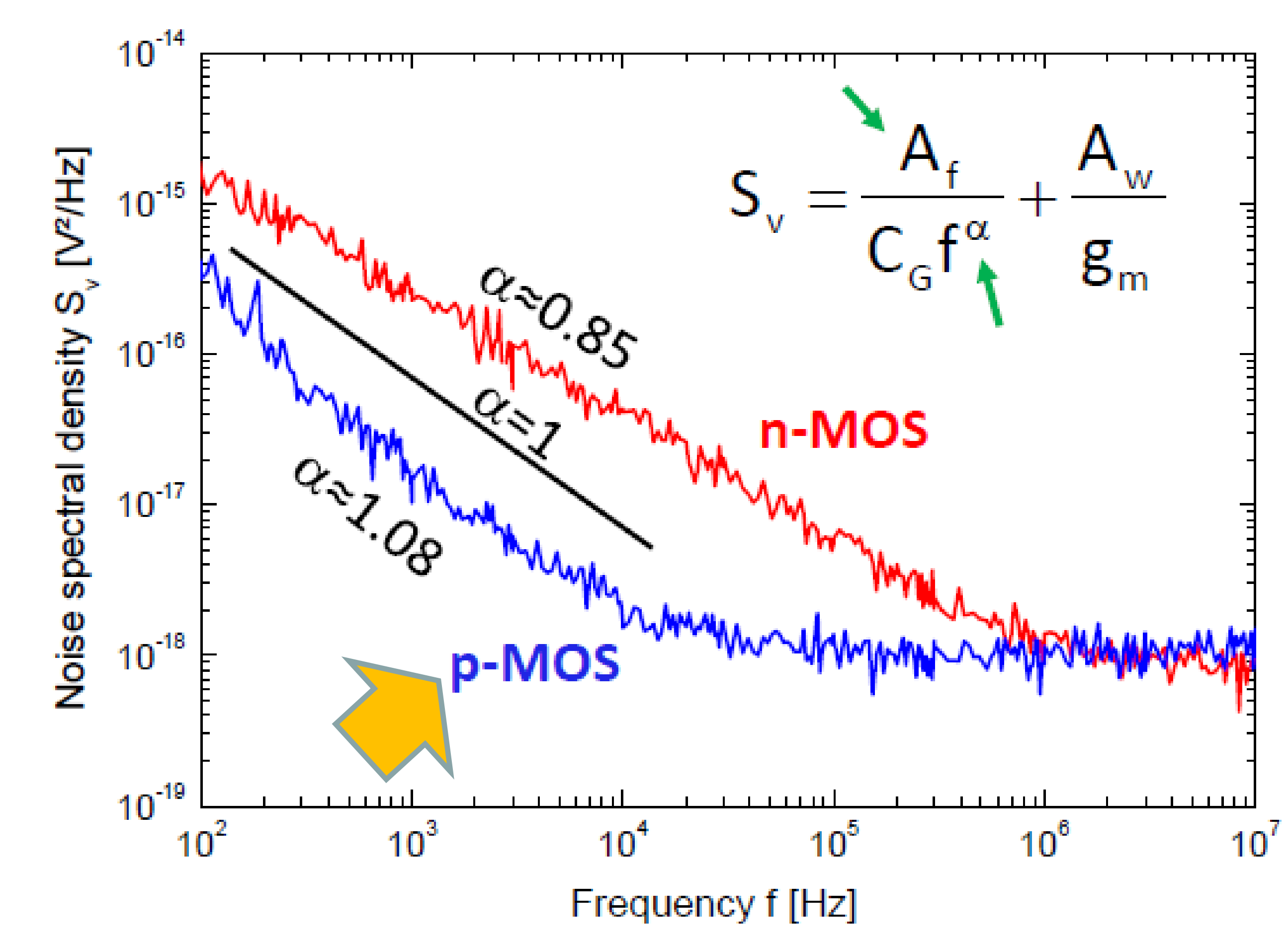
Both g_m and C_G push towards using n-channel and **$L = L_{min}$**

Conclusion 1: considering (g_m and Cg)

De Geronimo, IEEE TNS 52, 2005

BUT,
now let add the 1/f noise contribution

Low-Frequency Noise



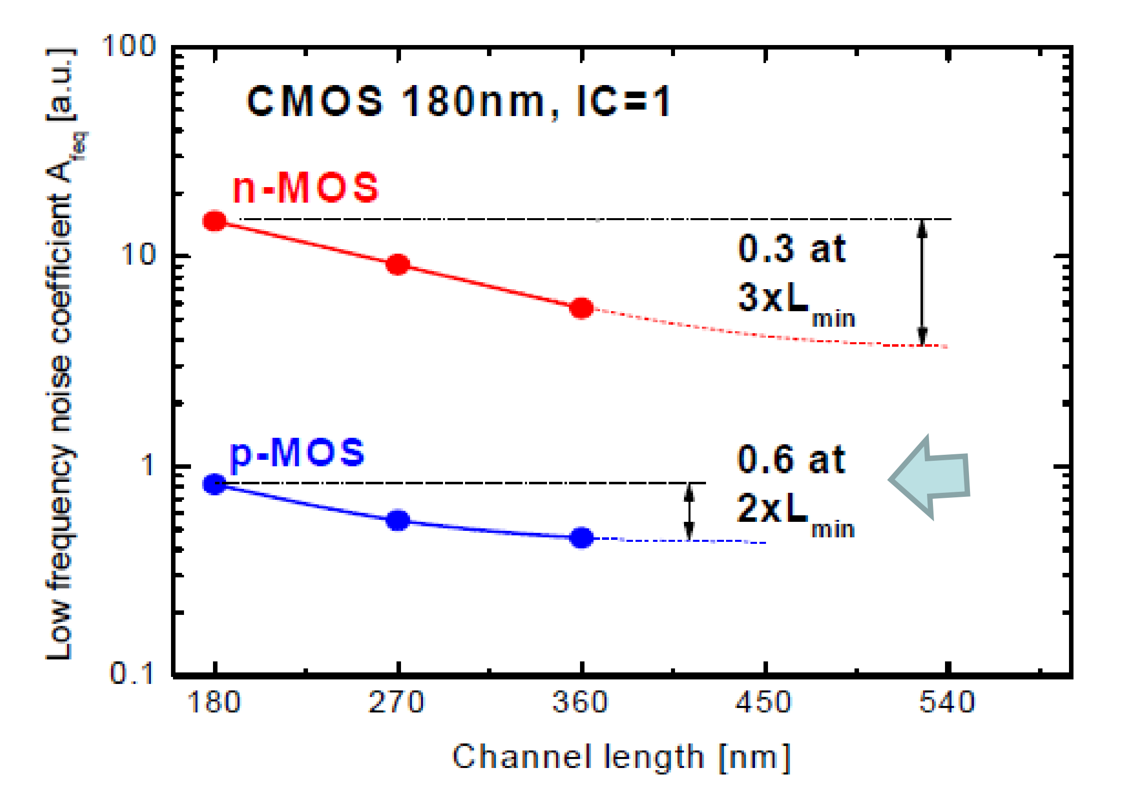
From transistor's low-freq. noise:

$$ENC_{vf}^2 = a_f(\alpha)\frac{A_f}{\tau^{1-\alpha}}\frac{(C_S + C_G)^2}{C_G}$$

depends on τ

De Geromino

Low-Frequency Noise vs L



$$S_v = \frac{A_f(L)}{C_G f^{\alpha}} + \frac{A_w}{g_m}$$

1/f equivalent, IEEE TNS 58, 2011

From transistor's low-freq. noise:

$$ENC_{vf}^2 = a_f(\alpha)\frac{A_f(L)}{\tau^{1-\alpha}}\frac{(C_S + C_G)^2}{C_G}$$

Conclusion 2

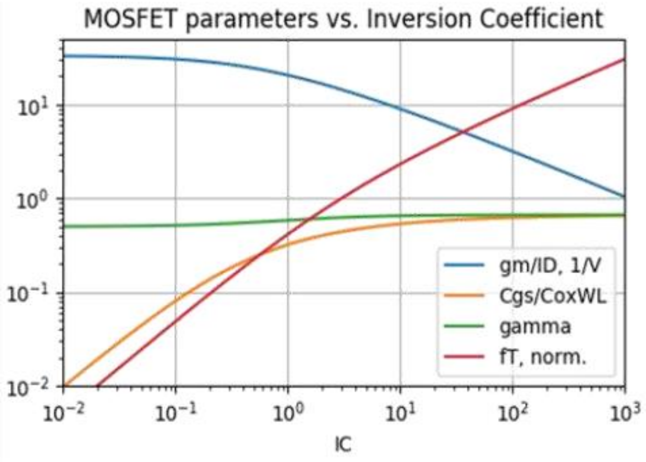
LF noise pushes towards p-channel & L > L_{min}

MOSFET parameters (summary)

Parameter	Symbol	Strong inversion value	Weak inversion value
Transconductance	g_m	$\sqrt{2n\mu C_{ox}\frac{W}{L}I_D}$	$\frac{qI_D}{nkT}$
Input capacitance	C_{GS}	$\frac{2}{3}C_{ox}WL$	$\ll C_{ox}WL$
Cutoff frequency	ω_T	g_m/C_{GS}	
Noise (white)	$e_{n,w}^2$	$\frac{8kTn}{3g_m}$	$\frac{2kTn}{g_m}$
Noise (flicker)	$e_{n,f}^2$	$\frac{K_f}{fC_{ox}WL}$	
Mismatch	$\sigma(V_T)$	$\frac{\alpha}{\sqrt{WL}}$	
Intrinsic voltage gain	A_{V0}	g_m/g_d	

short-channel effects and parasitics not modeled

- Select *normalized current density* and gate length as primary design variables
- Interpolation formula* models the MOSFET's key parameters continuously from weak to strong inversion

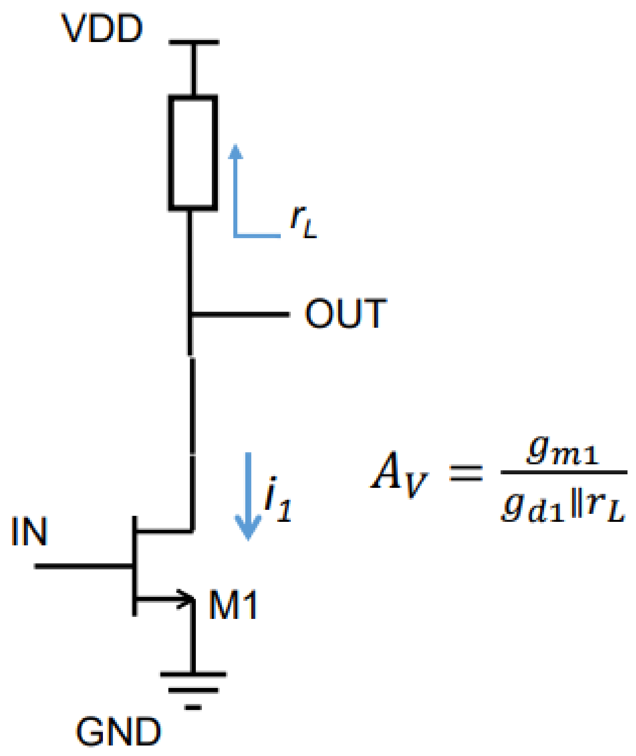


** C. Enz, F. Krummenacher, E. Vittoz, Analog Integrated Circuits and Signal Processing (1995)*
Binkley, D. (2007) 14th Intl. Conf.on Mixed Design of Integrated Circuits and Systems; Binkley et al. (2006) Analog ICs and Si Processing

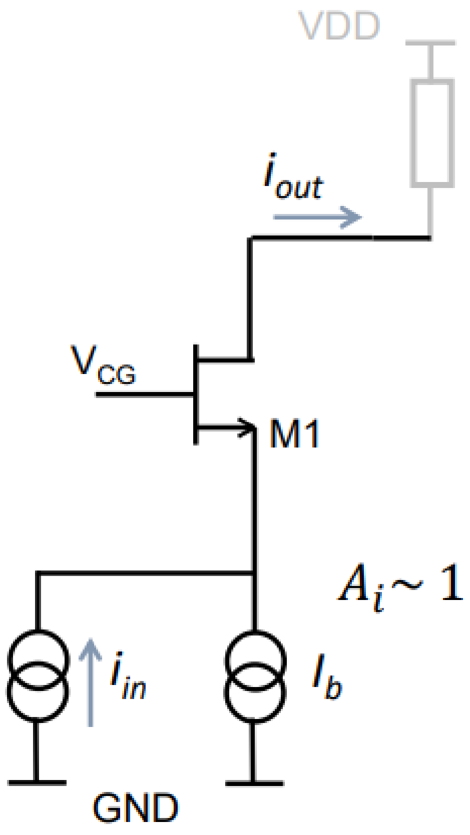
Paul O'connor

Topologies for preamp

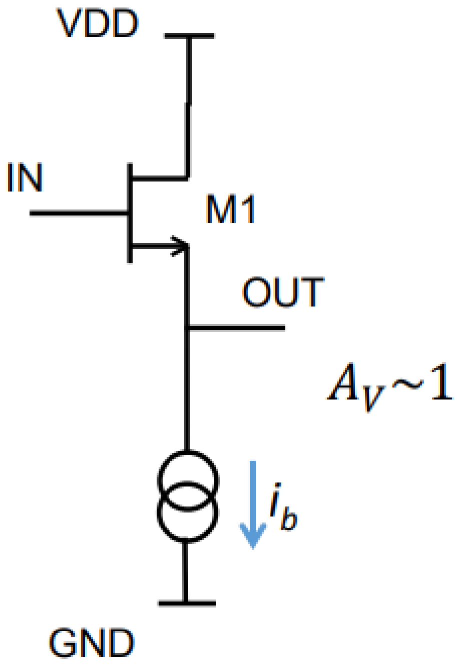
basic common-source stage
voltage-in, voltage-out



basic common-gate stage
current-in, current-out



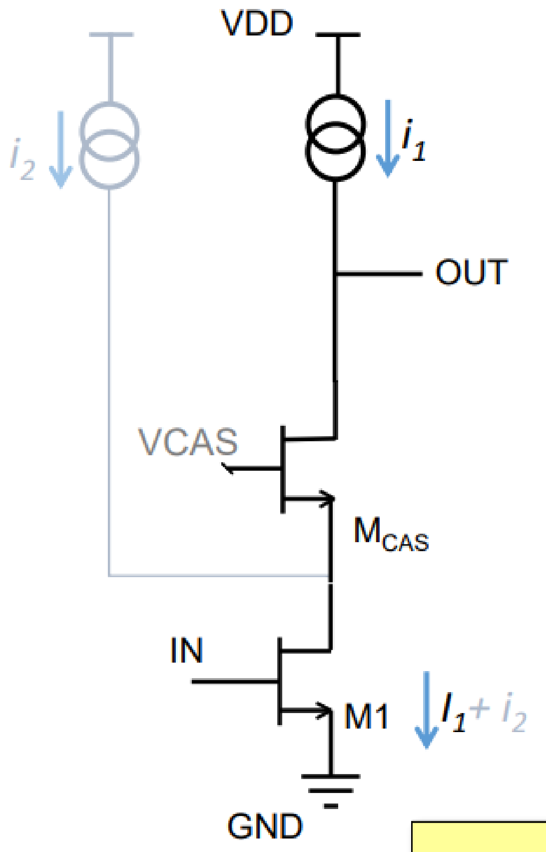
basic common-drain stage
voltage-in, voltage-out
(source follower)



Paul O'connor (BNL)

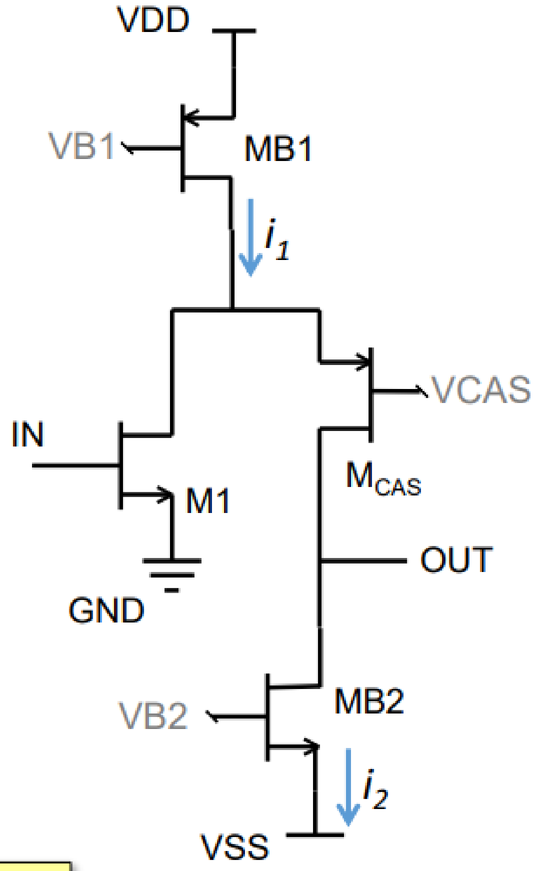
Common CSA/TIA preamplifiers topology

NMOS-input common-source cascode
(with supplementary current source i_2)

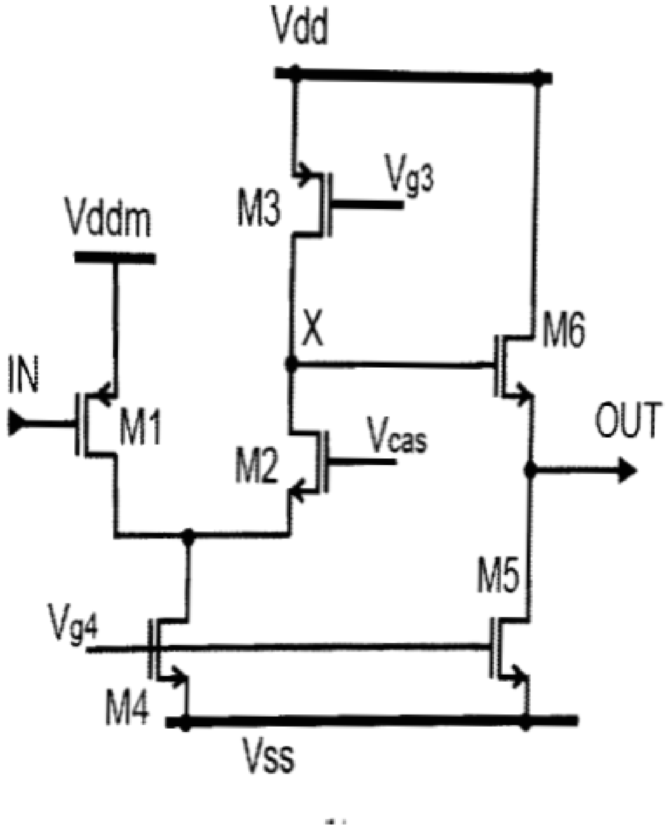


$$A_{V0} \cong \frac{g_{M1}\, g_{MCAS}}{g_{d1}\, g_{dCAS}}$$

NMOS-input folded cascode



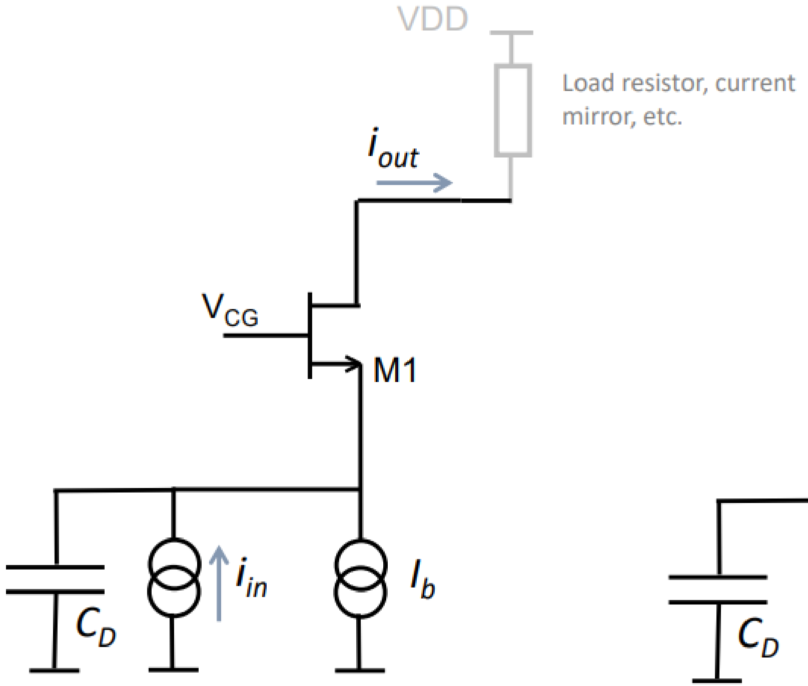
PMOS-input folded cascode
with source follower



Paul O'connor (BNL)

Current preamplifiers (Common gate) topologies

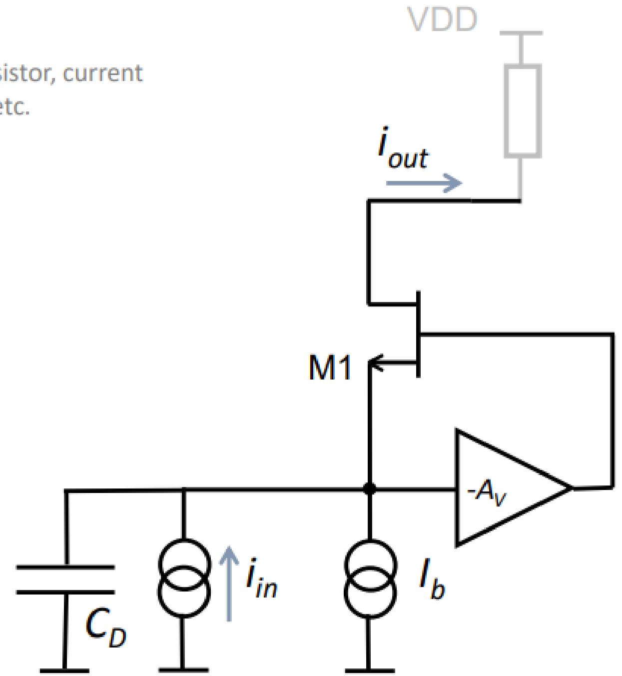
common gate



$$Z_{in} = 1/g_{M1}$$

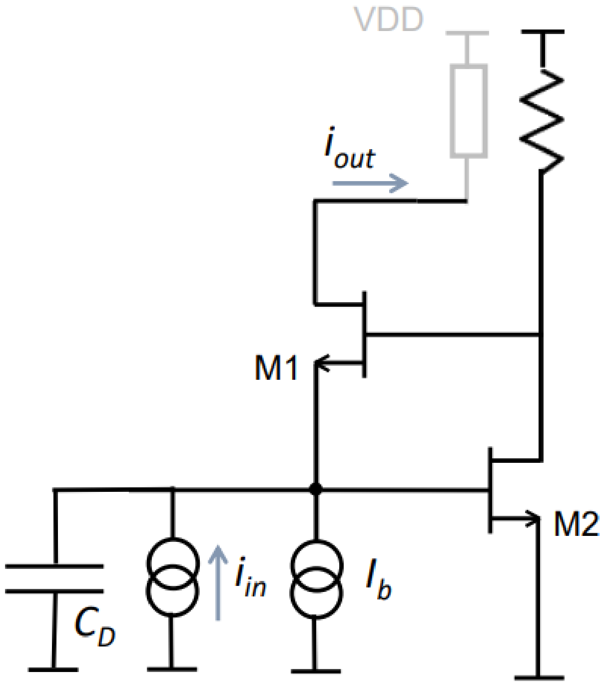
ENC dominated by ***parallel*** noise of M1
→ tradeoff of Z_{in}, ENC

regulated common gate



$$Z_{in} = \frac{1}{g_{M1} A_V}$$

RCG realization



ENC dominated by ***series*** noise of M2
→ more options for tradeoff of Z_{in}, ENC

Paul O’connor (BNL)

Preamp trends with agressive process

Preamp design & 'Scaling'

One may consider 2 prospection studies (old now !!)

1) Paul O'Connor => Brookhaven Lab, Upton , New York: 1,5µ au 180n

At constant power one save 23% in term of noise per generation: λ=0.7

$$ENC' = \lambda^{3/4} * ENC$$

At constant noise: one save 60% of power per generation

$$P' = \lambda^{3} * P$$

But in dynamique range, one lose 10% of SNR per generation

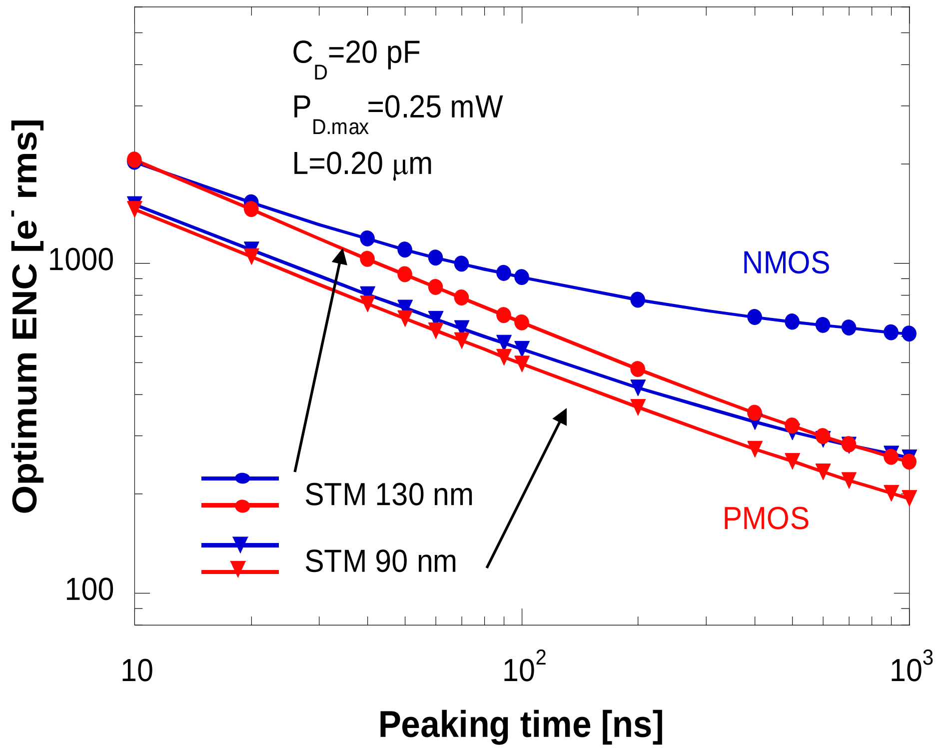
$$\mathbf{SNR' = \lambda^{1/4} * SNR}$$

2) L. Rattia NSS 2007 => Università degli Studi di Pavia : 100n et 90n

Next slides show some of L. Rattia' work

ENC *vs* peaking time, @ Pd=cte

L. Rattia NSS 2007 =>The 90nm leads to less ENC (noise) than the 130nm

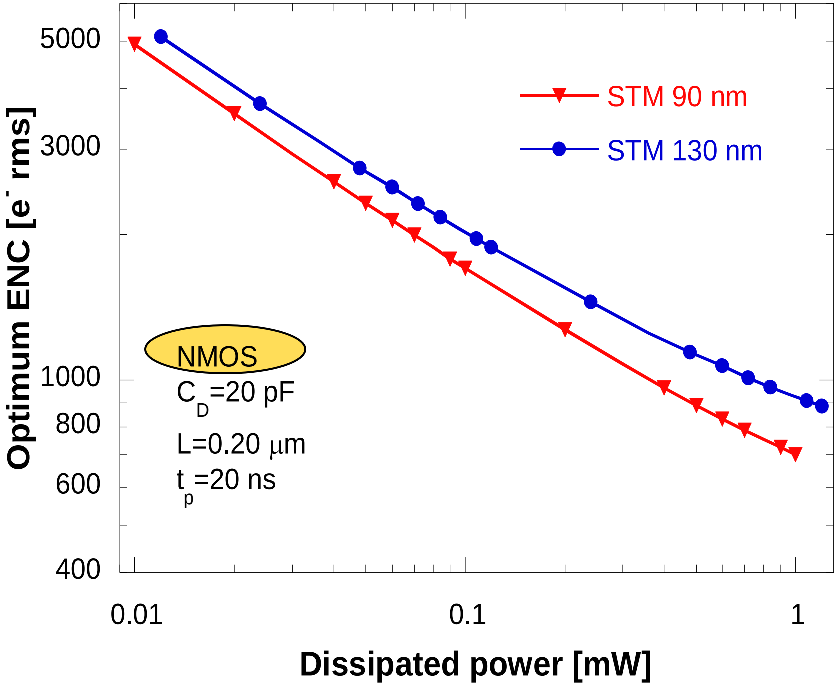


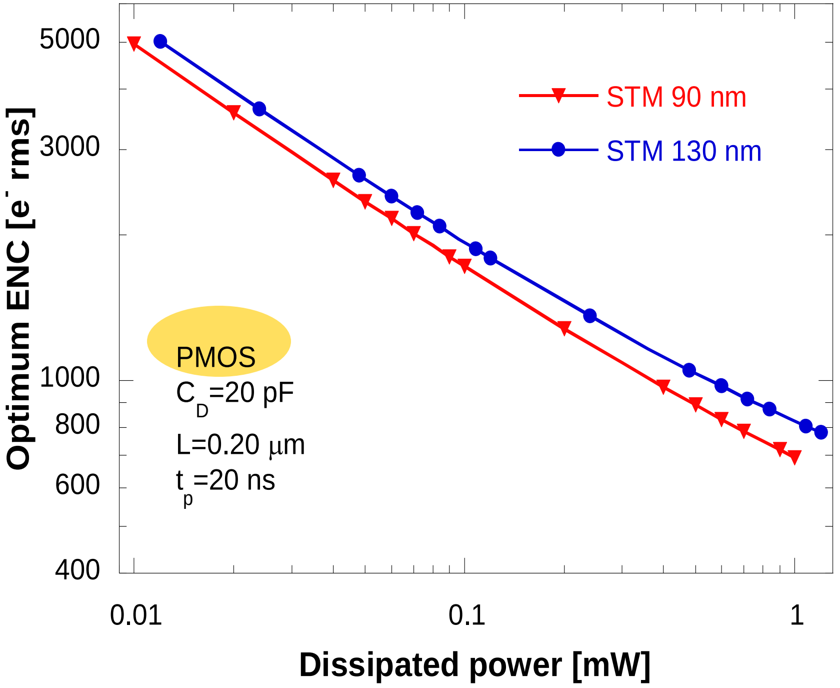
ENC was evaluated in the case of a second order, unipolar (RC^2-CR) shaping processor

- In the explored peaking time and power range, **PMOS input device** always provides better noise performances than NMOS input (*except for the 130 nm process at* t_p *close to 10 ns*)
- Using the 90 nm process may yield quite significant improvement with respect to the 130 nm technology, especially when NMOS input charge preamplifiers are considered

ENC *as function of power necessary (P et N)* ***for given tp***

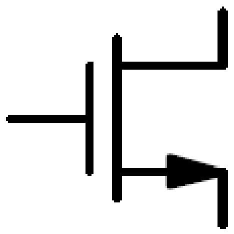
L. Rattia NSS 2007 **=> The 90nm needs less power at a constant noise level**





- At t_p=20 ns, noise performances provided by NMOS and PMOS input devices in the 90 nm technology are comparable
- Better noise-power trade-off can be achieved by using the 90 nm technology

Figure of Merit for a MOS Process



- Transit Frequency

$$\omega_T = \frac{g_m}{C_{gs}}$$

- Transconductor Efficiency

$$\frac{g_m}{I_D}$$

- Intrinsic Gain

$$g_m r_o$$

DO NOT FORGET: You may optimize the *bias point of input transistor* (moderate inversion, for a gm/Id efficiency)

Topics for later sessions

How can one characterize a comparator for counting accurately?

In couting flow: Frequency of noise hits (**fn**); Threshold **Vth**; input noise (**vn**) …

S. O. Rice *Mathematical analysis of random noise* [1945] Bell System Technical journal, 24; 46-156

- How often are noisy events counted?
- Noise at your comparator input?
- Threshold value above the baseline?
- Counting rate and so your bandwidth or τ?

$$f_n = \frac{1}{2\pi\tau} e^{-(Vth^2/2Vn^2)}$$

Or Threshold over noise ratio

$$\frac{Vth}{Vn} = \sqrt{-2 * \ln(fn * 2\pi\,\tau)}$$

EXO: estimate f_n for V_{th}/V_n =7 for different τ

HELP for LTSPICE

Parameters

Input Resistor

```
;param RINP=1
.step param RINP list 1 10MEG
```

RE sets Drain current

```
.param RE=249
;step param RE 80 300 10
```

RD sets Drain-Source voltage

```
.param RD=2.49K
;step param RD 2500 5500 100
```

Temperature

```
.param temp=25
;step temp -50 125 25
```

Miscellaneous

```
.param PSV=15
```

Analysis

```
;op
;tf V(out) V3
;dc V3 -0.1 0.1 0.01
;tran 0 0.1 0
;ac dec 10 0.1 100MEG
.noise V(out) V3 dec 10 0.1 10K
```

Voltage Noise Measurements

```
.meas NOISE end1_10_RMS FIND V(inoise)@1 AT 10
.meas NOISE en1_10_RMS PARAM en1_10_RMS*0.707

.meas NOISE end1_1K_RMS FIND V(inoise)@1 AT 1K
.meas NOISE en1_1K_RMS PARAM  end1_1K_RMS*0.707

.meas NOISE en1_LF_RMS INTEG V(inoise)@1*0.707 FROM 0.1 TO 10
.meas NOISE en1_LF_PP PARAM en1_LF_RMS*5
```

Current Noise Measurements

```
.meas NOISE en21_100_RMS FIND V(inoise)@2-V(inoise)@1 AT 100
.meas NOISE in_100_RMS PARAM en21_100_RMS/1e7

.meas NOISE en21_LF_RMS INTEG (V(inoise)@2-V(inoise)@1) FROM 0.1 TO 10
.meas NOISE in_LF_PP PARAM (en21_LF_RMS/1e7)*5
```

LTSPICE help!

However, let's look at one to get the input voltage noise at 1kHz.

.meas NOISE en1_1k_RMS FIND V(inoise)@1 AT 1K

NOISE - Apply the measurement to a noise simulation

en1_1k_RMS - Just a name for the result. Used in the log file.

FIND - Specifies the measurement, which in this case is just getting a data value

V(inoise)@1 - The data set to use in the measurement. Details are below.

AT 1K - Selects the frequency of the data