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Advanced Photodetectors Shortcourse

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Signal Processing and Electronics

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These course notes are posted together with additional tutorials at
http://www-physics.lbl.gov/~spieler

or simply websearch "spieler detectors"

More detailed discussions in
H. Spieler: Semiconductor Detector Systems, Oxford University Press, 2005

Course Contents

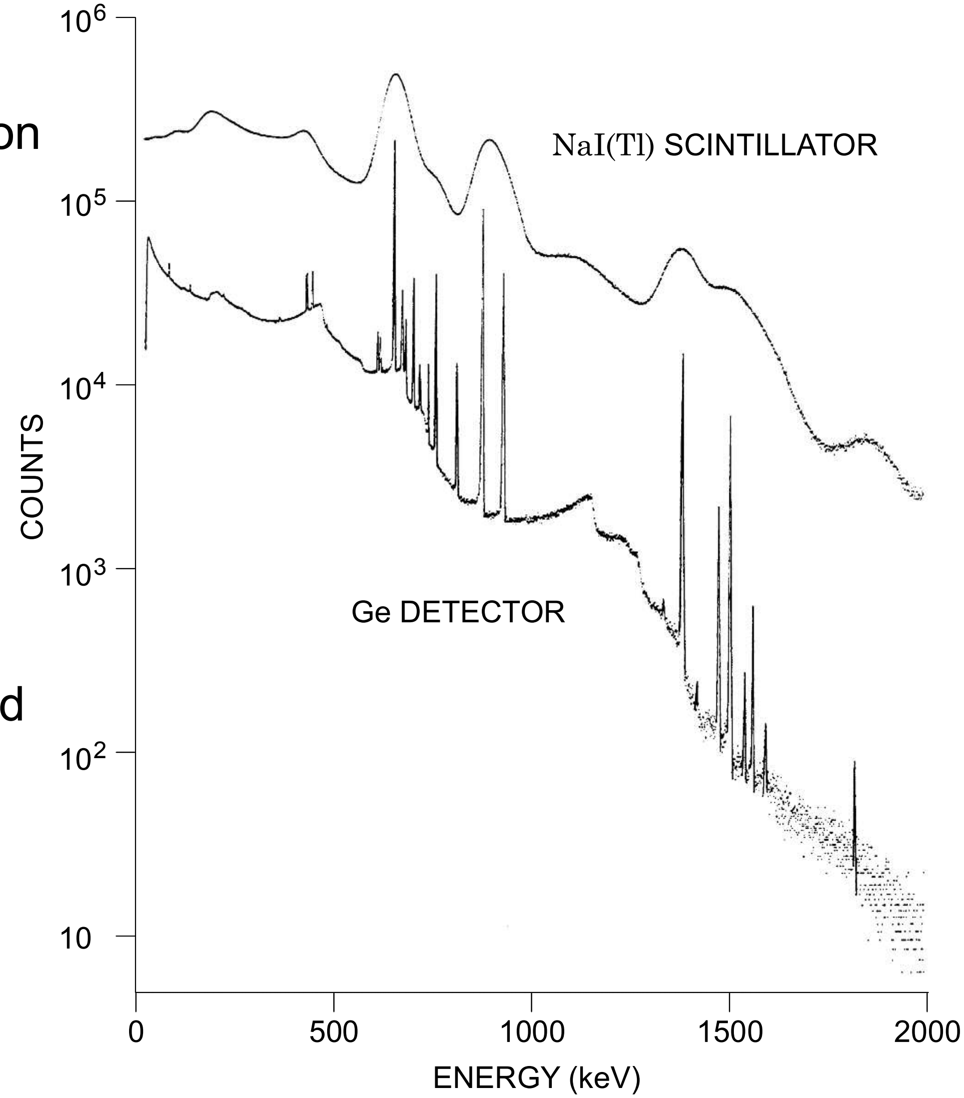
Why understand front-end electronics?

Energy resolution enables recognition of structure in energy spectra.

Optimizing energy resolution often depends on electronics.

Comparison of NaI(Tl) scintillation detector and Ge semiconductor diode detector.

- Resolution in NaI(Tl) is determined by the scintillator.
- Resolution of the Ge detector depends significantly on electronics.



J.Cl. Philippot, IEEE Trans. Nucl. Sci. **NS-17/3** (1970) 446

Energy resolution is also important in experiments that don't measure energy.

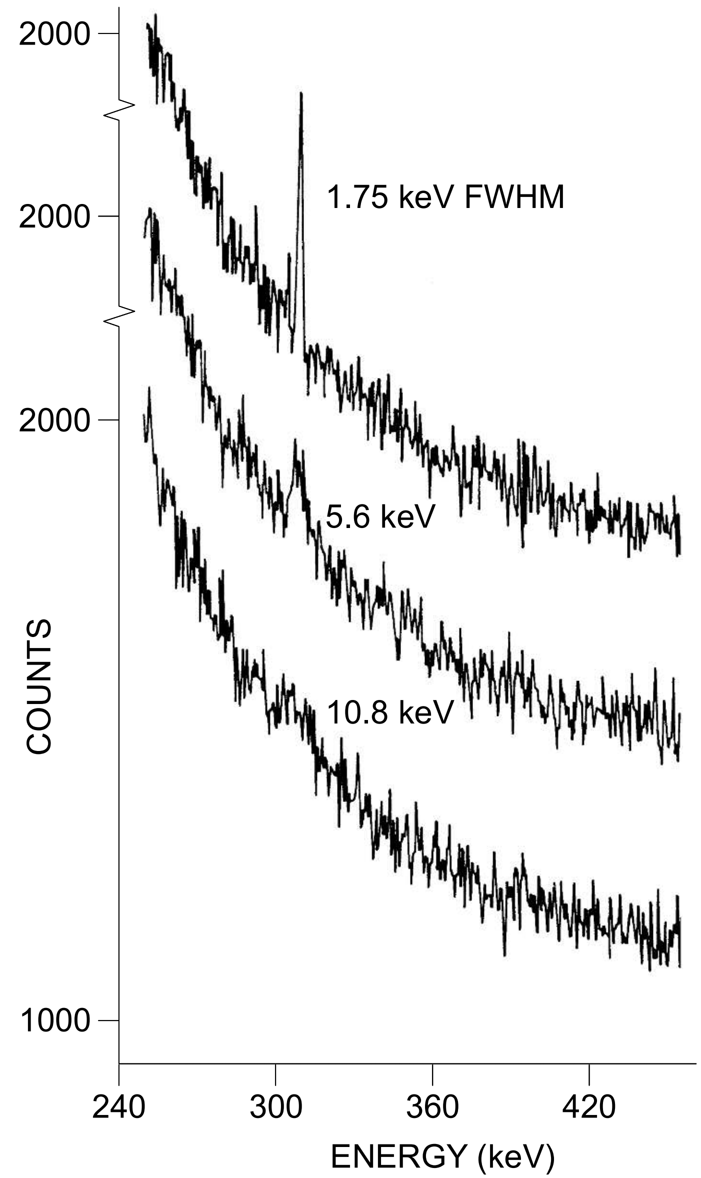
Energy resolution improves sensitivity because

signal-to-background ratio improves with better resolution.

(signal counts in fewer bins compete with fewer background counts)

In tracking detectors a minimum signal-to-background ratio is essential to avoid fake hits.

Achieving the required signal-to-noise ratio with minimized power dissipation is critical in large-scale tracking detectors.



G.A. Armantrout *et al.*, IEEE Trans. Nucl. Sci. **NS-19/1** (1972) 107

Recognizing overall contributions to signal sensitivity does not require detailed knowledge of electronics engineering.

It does require a real understanding of basic classical physics.

i.e. recognize which aspects of physics apply in
practical situations

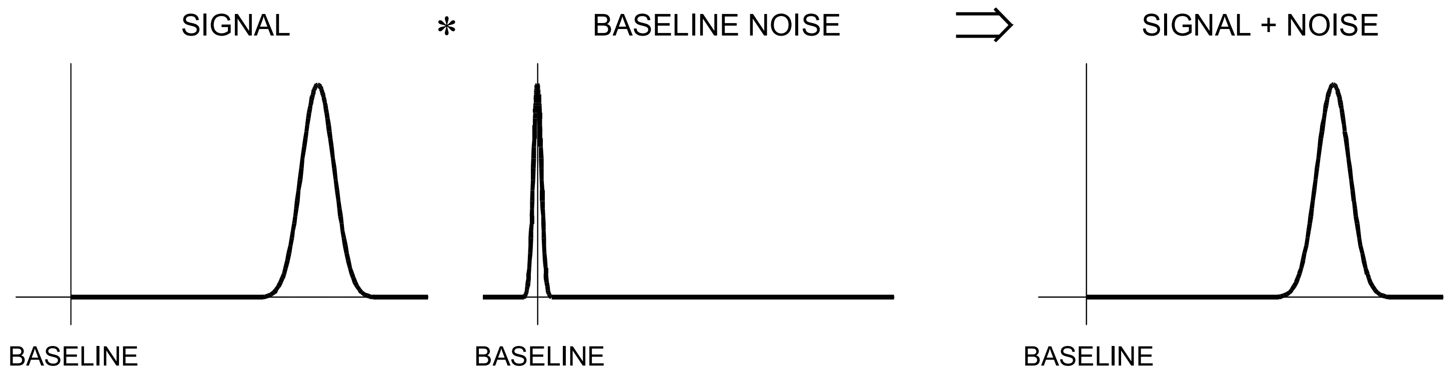
… nope, real life doesn't tell you which chapter recipes to follow!

For physicists and electronics engineers to work together efficiently it is necessary that physicists understand basic principles so that they don't request things that cannot work.

A common problem is "wouldn't it be nice to have this ...", which often adds substantial effort and costs
– without real benefits.

What Determines Sensitivity or Resolution?

1. Signal variance (e.g. statistical fluctuations) >> Baseline Variance



$\Rightarrow$ Electronic (baseline) noise not important

Examples:
- High-gain proportional chambers
- Scintillation Counters with High-Gain PMTs

e.g. 1 MeV γ-rays absorbed by NaI(Tl) crystal

Number of photoelectrons: $N_{pe} \approx 8 \cdot 10^4$ [MeV^{-1}] $\times E_\gamma \times QE \approx 2.4 \cdot 10^4$

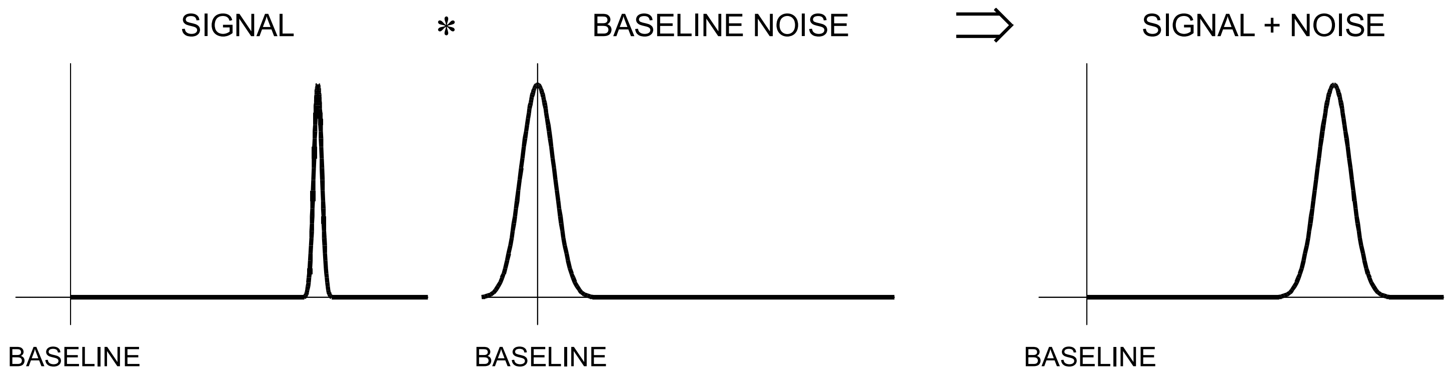
Variance typically: $\sigma_{pe} = N_{pe}^{1/2} \approx 160$ and $\sigma_{pe} / N_{pe} \approx 5$ - 8%

Signal at PMT anode (assume Gain= 10^4): $Q_{sig} = G_{PMT}\, N_{pe} \approx 2.4 \cdot 10^8$ el

$\sigma_{sig} = G_{PMT}\, \sigma_{pe} \approx 1.2 \cdot 10^7$ el

whereas electronic noise easily $< 10^4$ el

2. Signal Variance << Baseline Variance



⇒ Electronic (baseline) noise critical for resolution

Examples:
- Gaseous ionization chambers (no internal gain)
- Semiconductor detectors

e.g. in Si : Number of electron-hole pairs $N_{ep} = \frac{E_{dep}}{3.6 \text{ eV}}$

Variance $\sigma_{ep} = \sqrt{F \cdot N_{ep}}$ (where F = Fano factor ≈ 0.1)

For 50 keV photons: $\sigma_{ep} \approx 40$ el ⇒ $\sigma_{ep} / N_{ep} = 7.5 \cdot 10^{-4}$

Obtainable noise levels are 10 to 1000 el.

Baseline fluctuations can have many origins ...

pickup of external interference

artifacts due to imperfect electronics

… etc.,

but the (practical) fundamental limit is electronic noise.

Depends on noise sources and signal processing.

Sources of electronic noise:

- Thermal fluctuations of carrier motion
- Statistical fluctuations of currents

Both types of fluctuations are random in amplitude and time

$\Rightarrow$ Power distributed over wide frequency range

$\Rightarrow$ Contribution to energy fluctuations depends on signal processing

Many different types of detectors are used for radiation detection.

Nearly all rely on electronics.

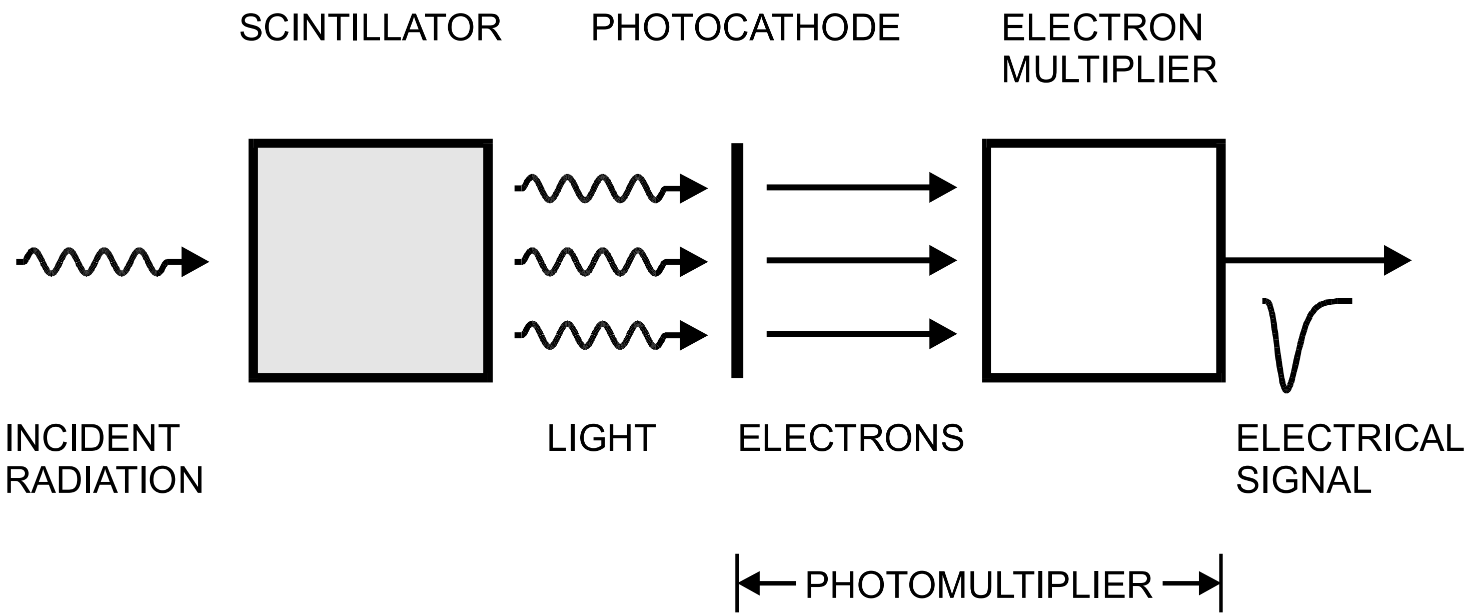
Although detectors appear to be very different, basic principles of the readout apply to all.

- The sensor signal is a current.
- 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 the following functions:

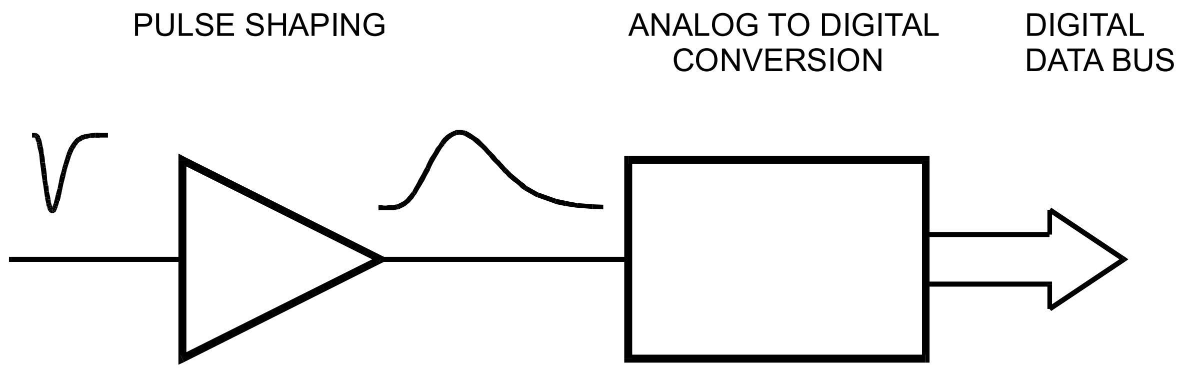
- Signal acquisition
- Pulse shaping
- Digitization
- Data Readout

Example: Scintillation Detector



number of photons	number of photoelectrons	charge in pulse
$\propto$ absorbed energy	$\propto$ absorbed energy	$\propto$ abs. energy

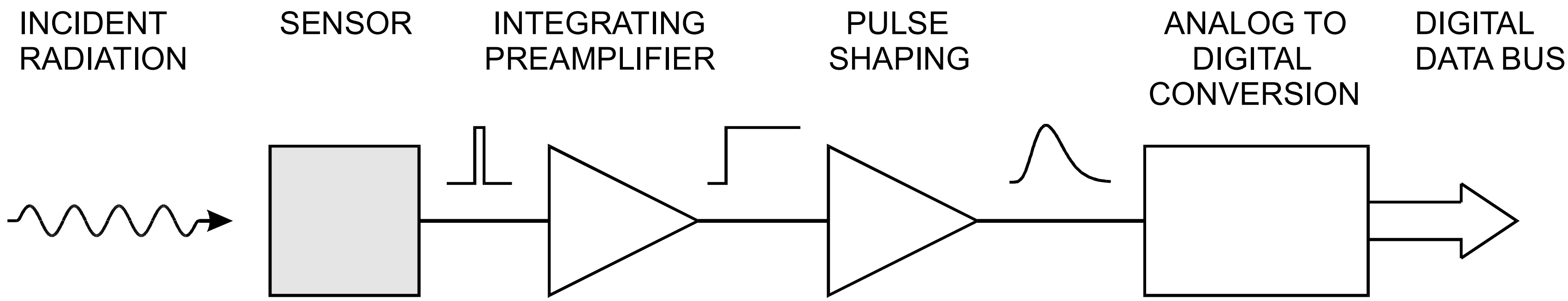
Readout



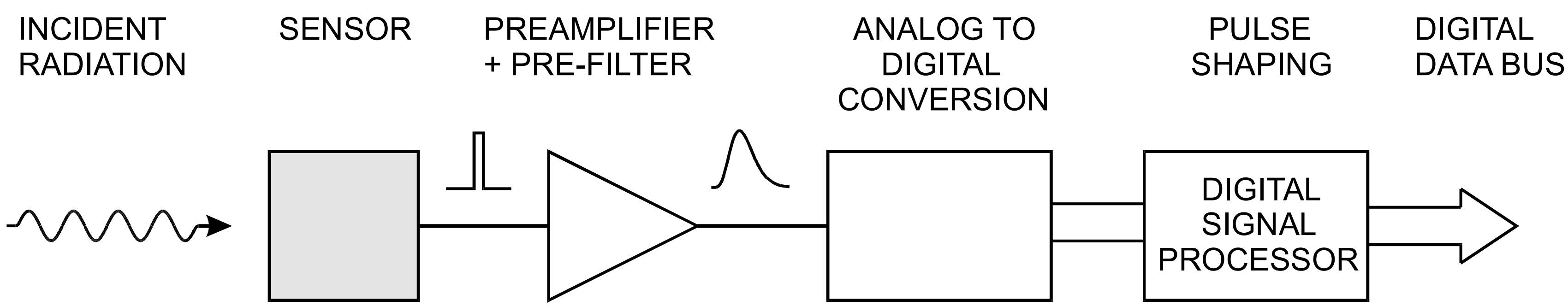
charge in pulse
$\propto$ absorbed energy

pulse height
$\propto$ absorbed energy

Basic Functions of Front-End Electronics

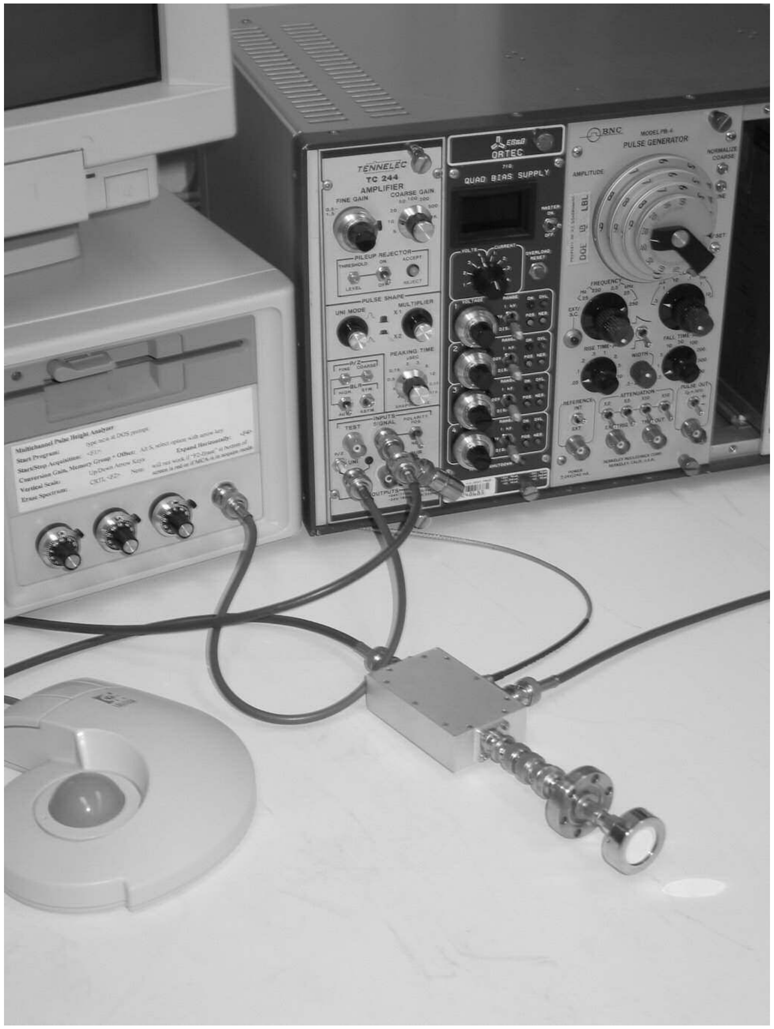


Pulse shaping can also be performed with digital circuitry:



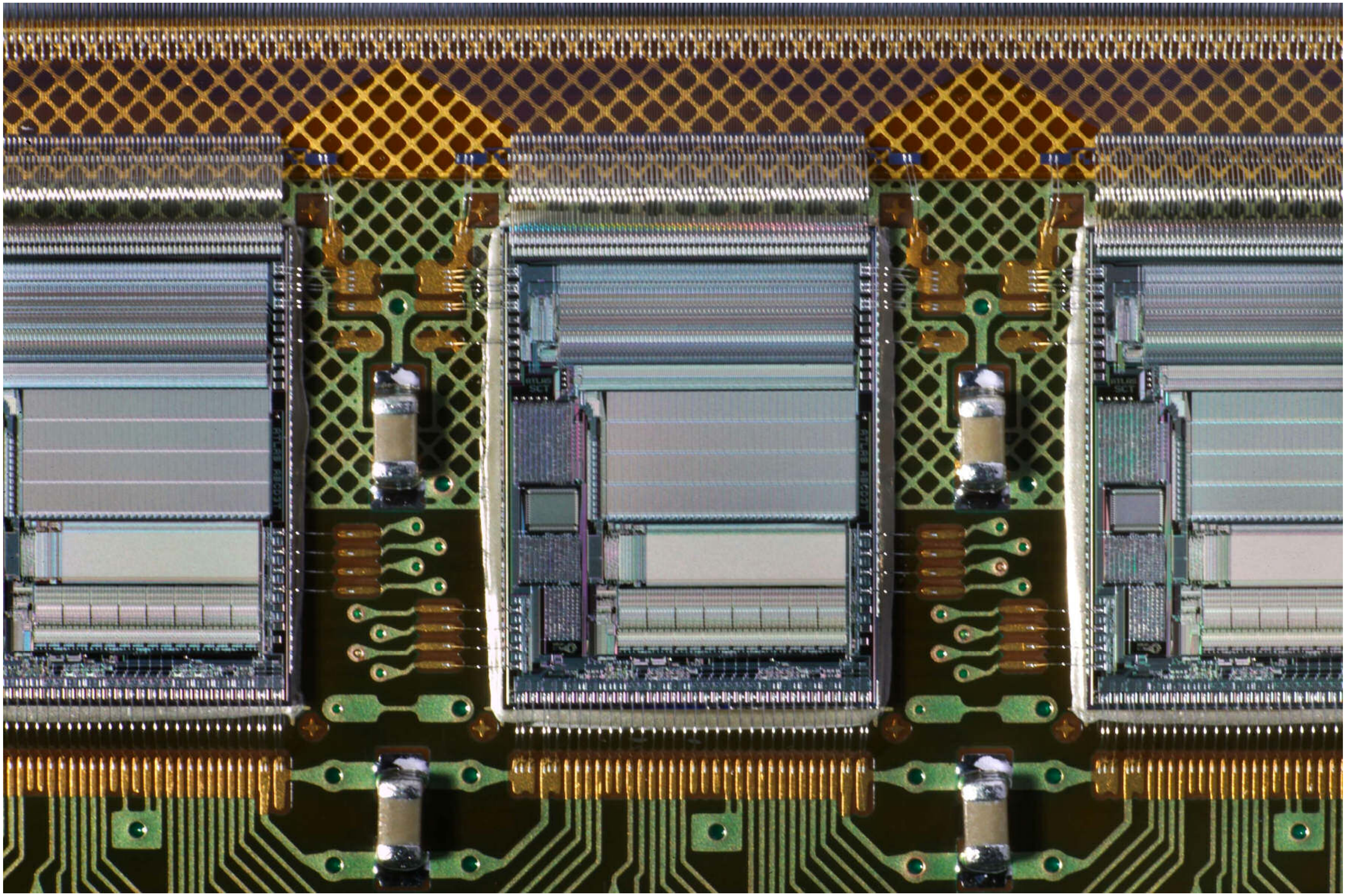
Many Different Implementations

“Traditional” Si detector system
for charged particle measurements



Tracking Detector Module (CDF SVX)
512 electronics channels on 50 μm pitch

ATLAS Silicon Strip system (SCT): 128-Channel chips mounted on hybrid



Design criteria depend on application

1. Energy resolution
2. Rate capability
3. Timing information
4. Position sensing

Large-scale systems impose compromises

1. Power consumption
2. Scalability
3. Straightforward setup + monitoring
4. Cost

Technology choices

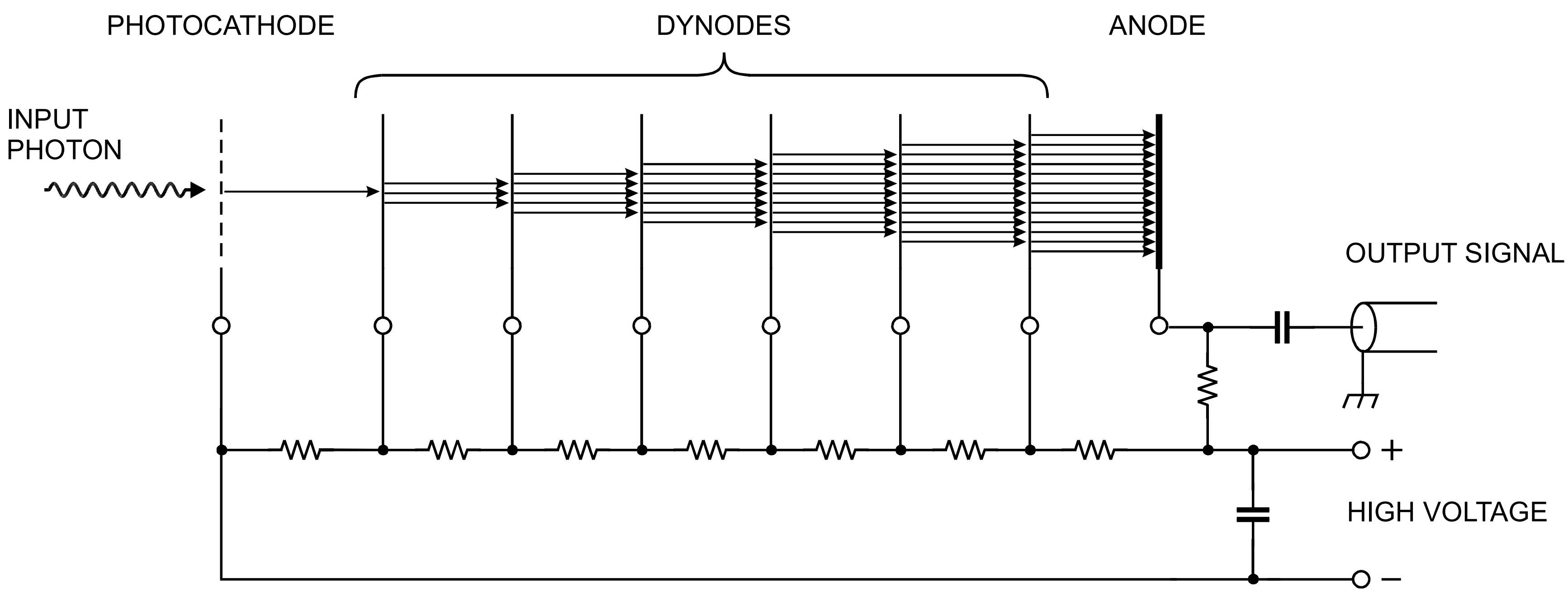
1. Discrete components – low design cost
fix "on the fly"
2. Full-custom ICs – high density, low power, but
better get it right!

Successful systems rely on many details that go well beyond "headline specs"!

II. Signal Formation and Acquisition

1. Photomultipliers

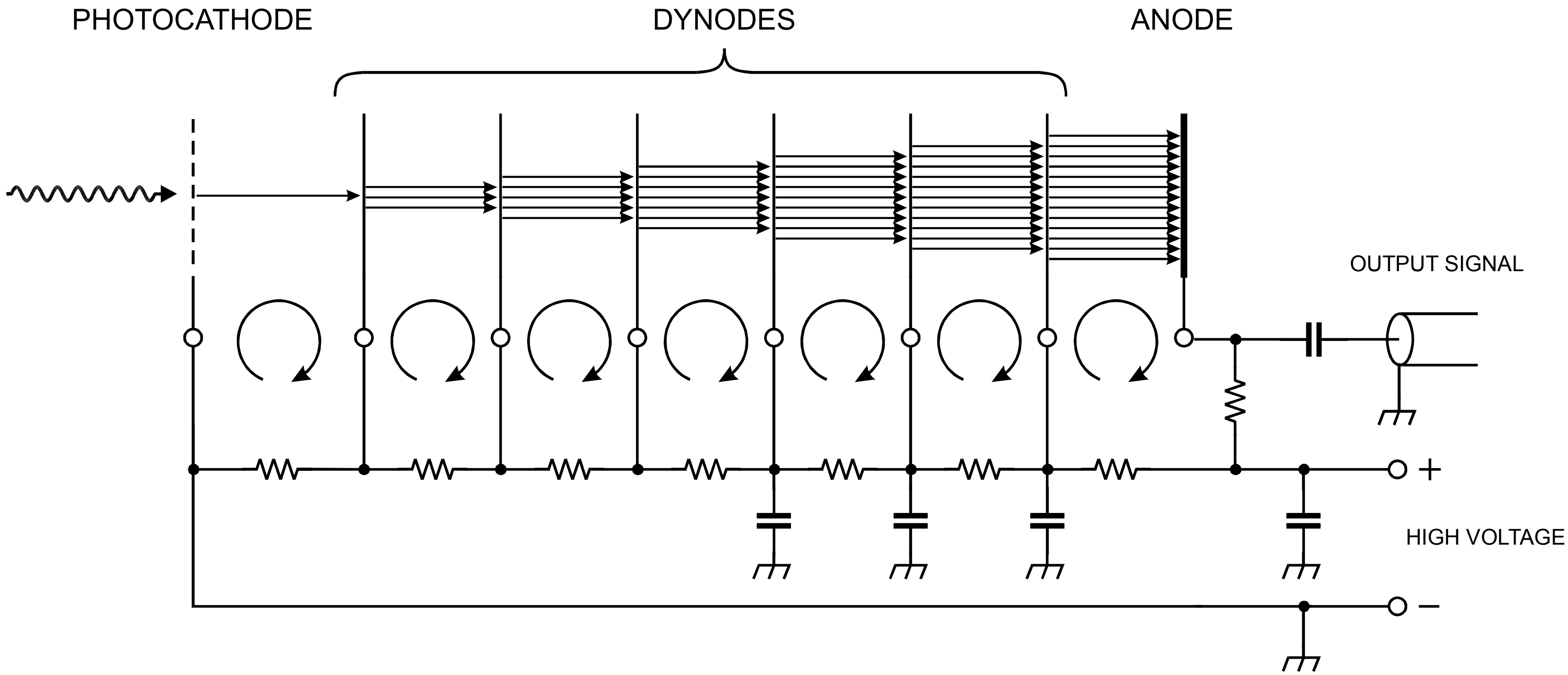
Basic Function



Voltage applied between dynodes accelerates the electrons to yield a significant gain.

The final set of electrons is collected by the anode and yields the output signal.

A common configuration:

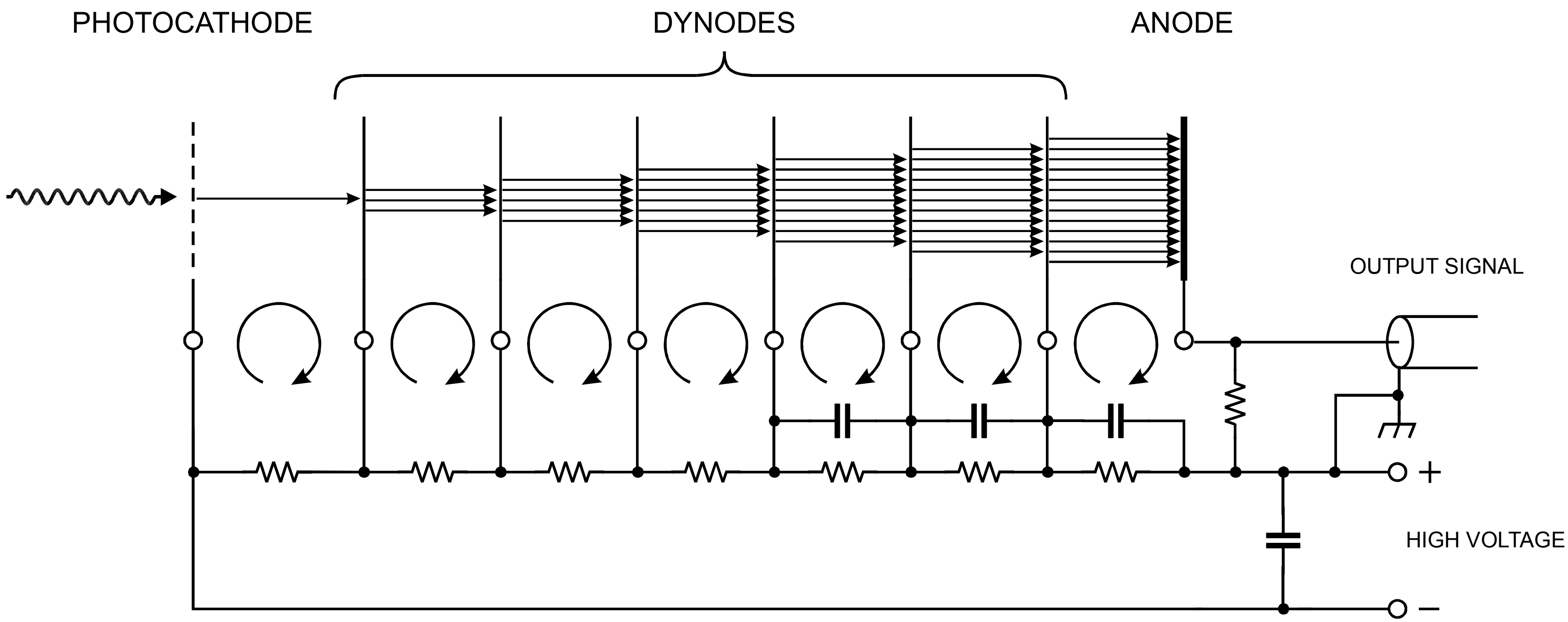


It is important to follow the individual currents

In the final dynodes the signal current can be so large that the voltage drop across the resistor is excessive.

The capacitors connected to the final dynodes are to store sufficient charge to avoid a significant voltage change.

The previous arrangement is typical, but it can be done better:



Connecting the storage capacitors directly between the dynodes provides a direct current path.

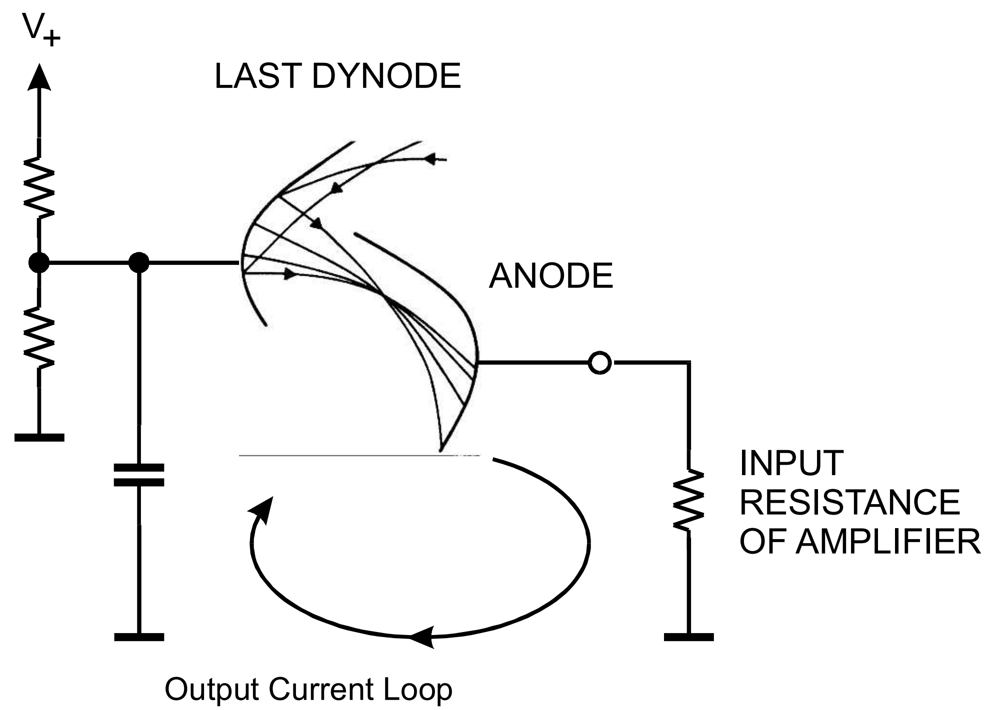
The smaller voltage across the capacitors also makes larger capacitances more practical, e.g. 1 μF multi-layer ceramic chip capacitors.

$\Rightarrow$ Less dynode-dynode voltage drop per pulse

Having the anode at ground level allows direct connection to the signal cable, often referred to as "DC coupling". However, it really isn't ...

Detail of output circuit

The output current loop always passes through the capacitor to the last dynode.

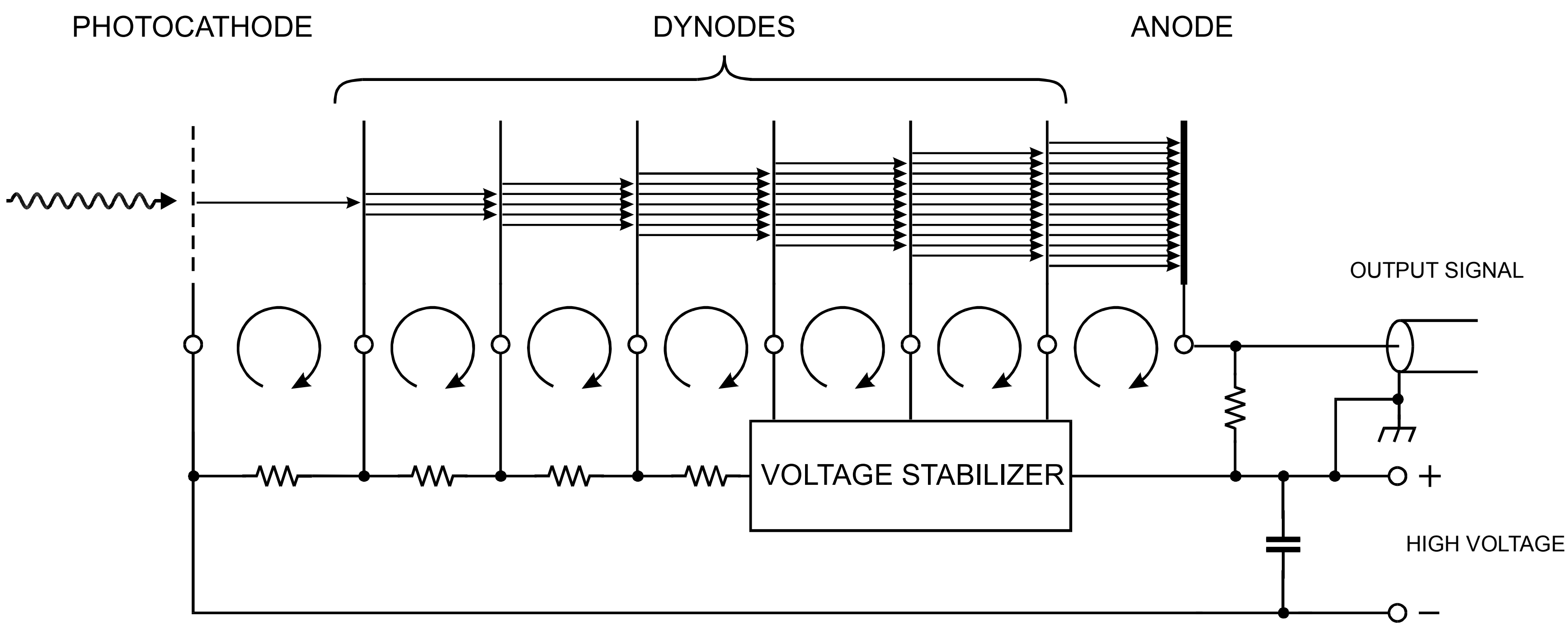


Placing the anode at ground level reduces the voltage on the final dynode relative to ground.

This allows a larger capacitance, increasing the time constant of the output coupling.

The larger signal currents at the last dynodes can reduce the dynamic range at high rates.

Electronic circuits that stabilize the voltage across individual dynodes can maintain the gain at high rates and can also provide true DC coupling of the output signal.



A simpler solution is to run the PMT at lower overall gain and adding a preamplifier.

- Fast low-noise amplifiers are quite practical.

The first dynode gain should still be kept high, but especially the final dynodes can run at much lower gain. PMTs with fewer dynode stages would be best.

Signal Fluctuations in a Scintillation Detector

Example: Scintillation Detector - a typical NaI(Tl) system

(from Derenzo)

Resolution of energy measurement is determined by the statistical variance of produced signal quanta.

$$\frac{\Delta E}{E} = \frac{\Delta N}{N} = \frac{\sqrt{N}}{N} = \frac{1}{\sqrt{N}}$$

Resolution is determined by the smallest number of quanta in the chain, i.e. number of photoelectrons arriving at the first dynode.

In this example

$$\frac{\Delta E}{E} = \frac{1}{\sqrt{3000}} = \text{ 2\% rms = 5\% FWHM}$$

511 keV gamma ray

$\Downarrow$

25000 photons in scintillator

$\Downarrow$

15000 photons at photocathode

$\Downarrow$

3000 photoelectrons at first dynode

$\Downarrow$

$3 \cdot 10^9$ electrons at anode

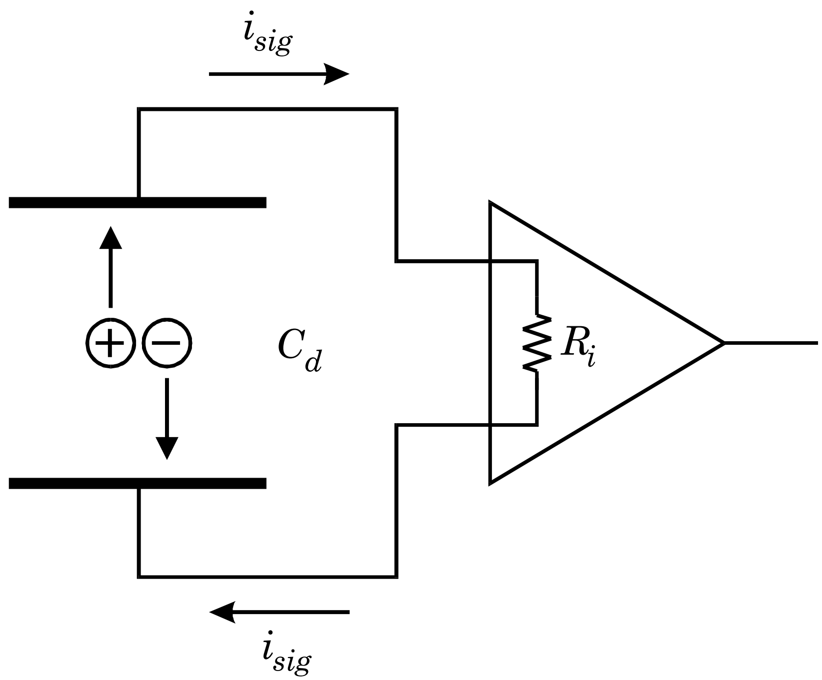
2 mA peak current

Typically 7 – 8% obtained, due to non-uniformity of light collection and gain.

2. Semiconductor Detectors

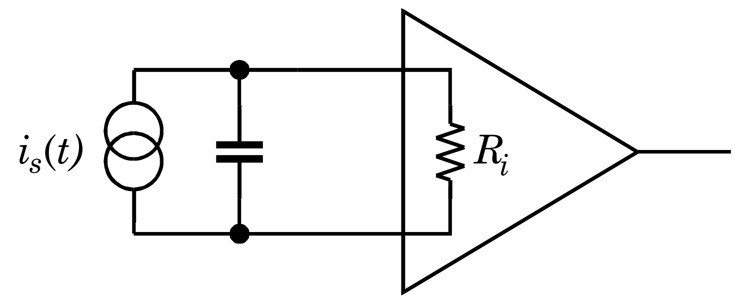
Signal Formation

DETECTOR AMPLIFIER



EQUIVALENT CIRCUIT

DETECTOR AMPLIFIER



When does the signal current begin?

a) when the charge reaches the electrode?

or

b) when the charge begins to move?

Although the first answer is quite popular (encouraged by the phrase “charge collection”), the second is correct.

When a charge pair is created, both the positive and negative charges couple to the electrodes.

As the charges move the induced charge changes, i.e. a current flows in the electrode circuit.

The electric field of the moving charge couples to the individual electrodes and determines the induced signal.

The following discussion applies to ALL types of structures that register the effect of charges moving in an ensemble of electrodes, i.e. not just semiconductor or gas-filled ionization chambers, but also resistors, capacitors, photoconductors, vacuum tubes, etc.

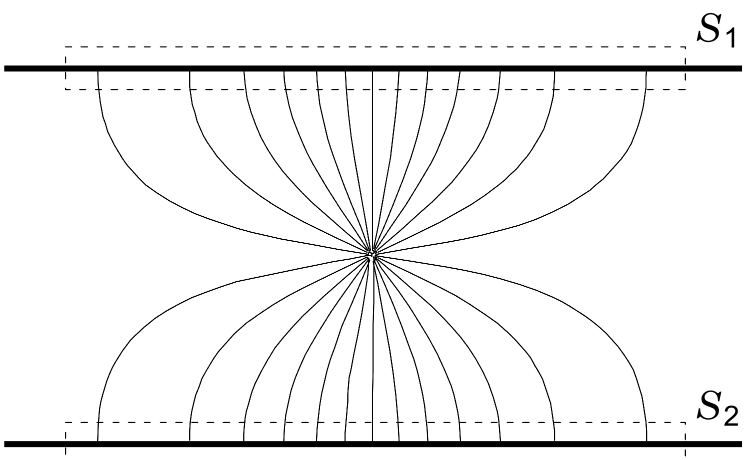
The effect of the amplifier on the signal pulse will be discussed in the Electronics part.

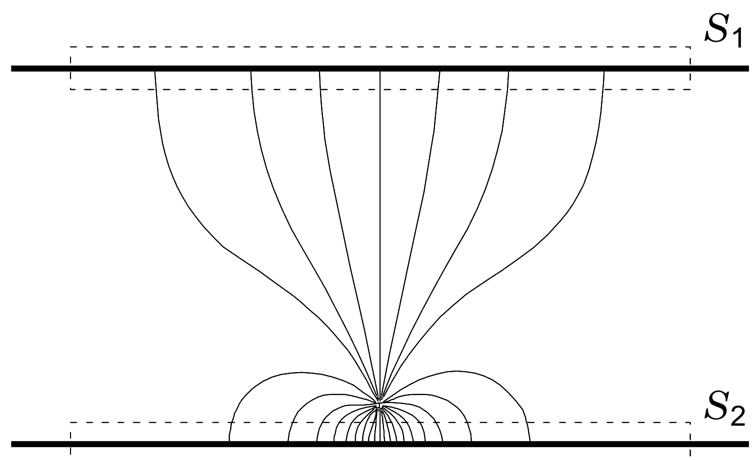
Induced Charge

Consider a charge q in a parallel plate capacitor:

When the charge is midway between the two plates, the charge induced on one plate is determined by applying Gauss' law. The same number of field lines intersect both S_1 and S_2, so equal charge is induced on each plate (= q / 2).

When the charge is close to one plate, most of the field lines terminate on that plate and the induced charge is much greater.

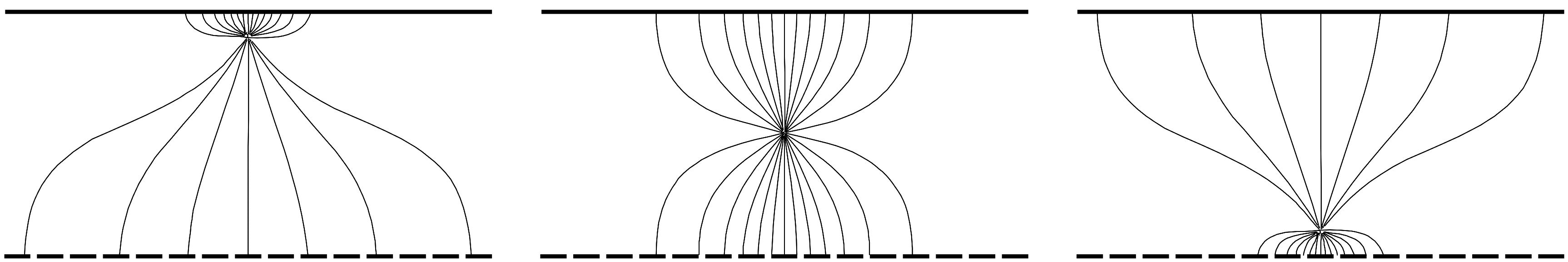




As a charge traverses the space between the two plates the induced charge changes continuously, so current flows in the external circuit as soon as the charges begin to move.

Induced Signal Currents in a Strip or Pixel Detector

Consider a charge originating near the upper contiguous electrode and drifting down towards the strips.



Initially, charge is induced over many strips.

As the charge approaches the strips, the signal distributes over fewer strips.

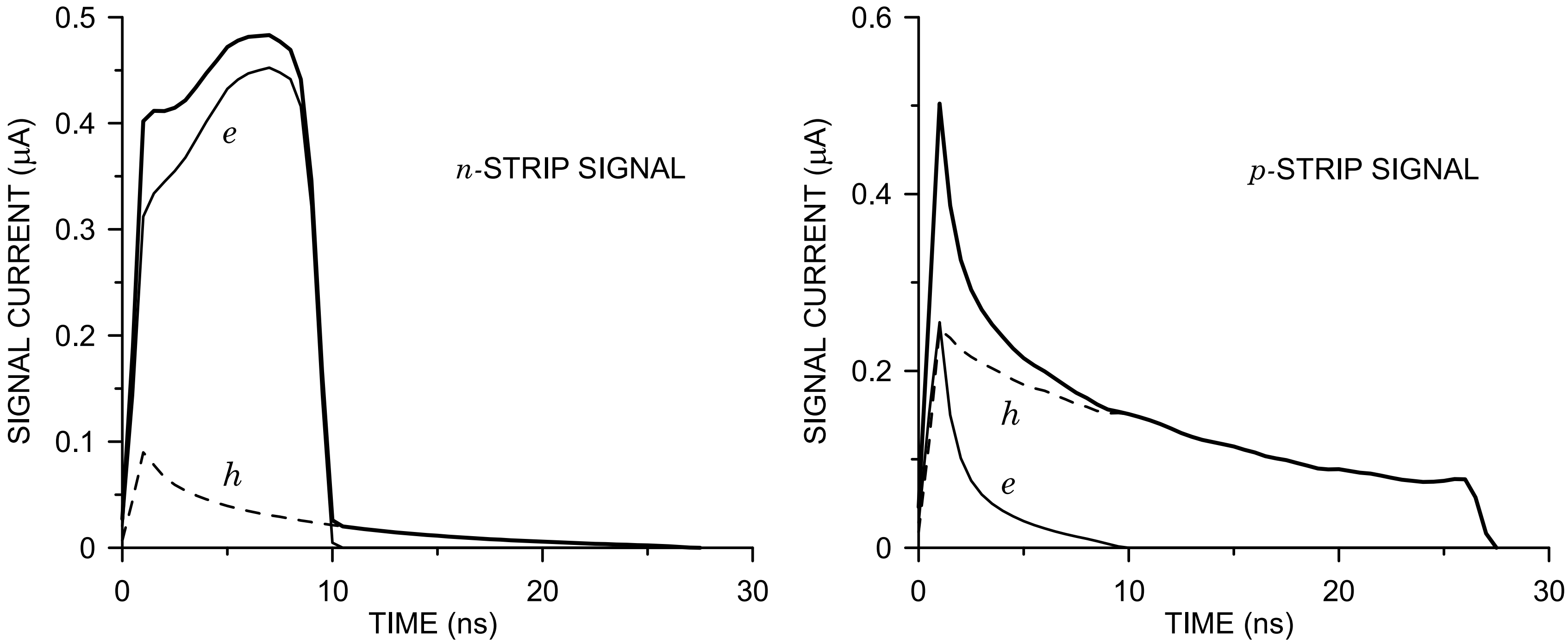
When the charge is close to the strips, the signal is concentrated over few strips

The magnitude of the induced current due to the moving charge depends on the coupling between the charge and the individual electrodes.

Mathematically this can be analyzed conveniently by applying Ramo's theorem.
(Chapter 2, pp 71-82)

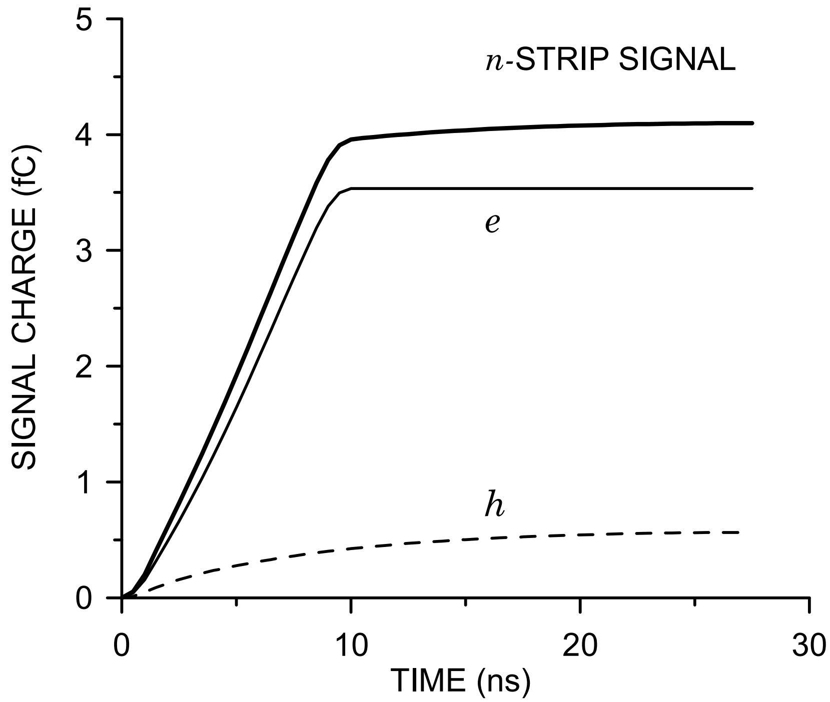
Note that deriving induced charge from "energy conservation" generally yields wrong results (it's typically not a theory based on the relevant physics).

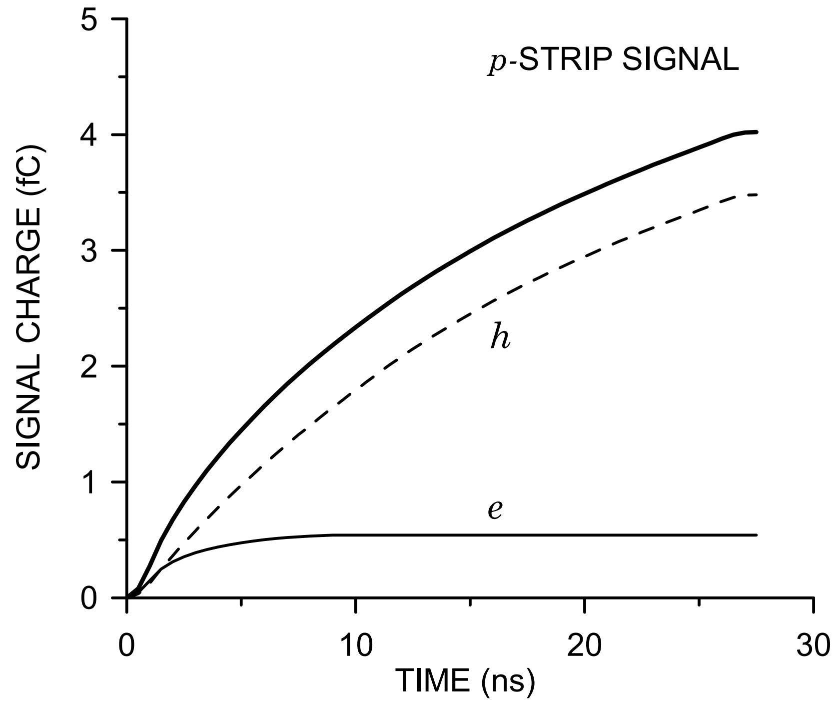
Current pulses in strip detectors (track traversing the detector)



The duration of the electron and hole pulses is determined by the time required to traverse the detector as in a parallel-plate detector, but the shapes are very different.

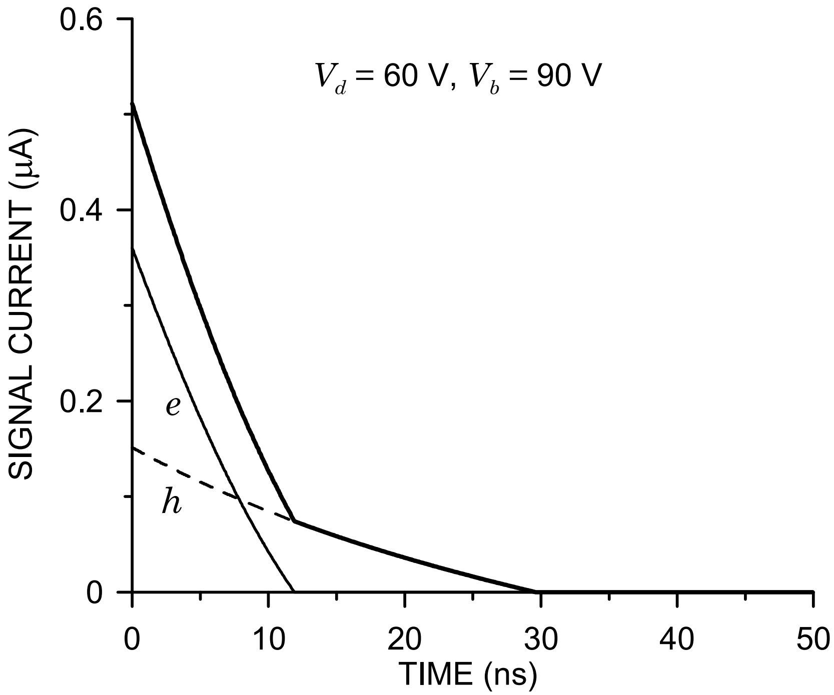
Strip Detector Signal Charge

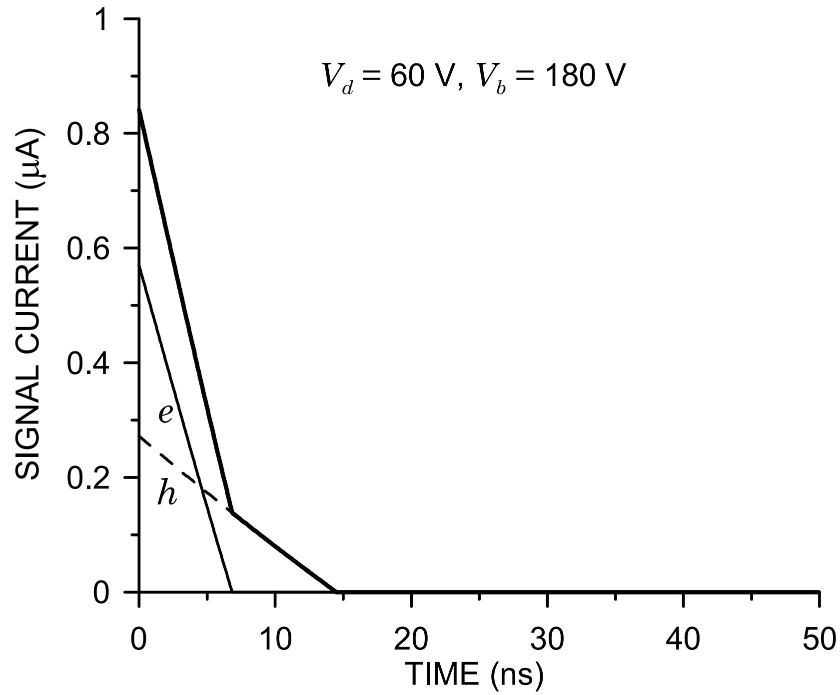




For comparison:

Current pulses in pad detectors (track traversing the detector)



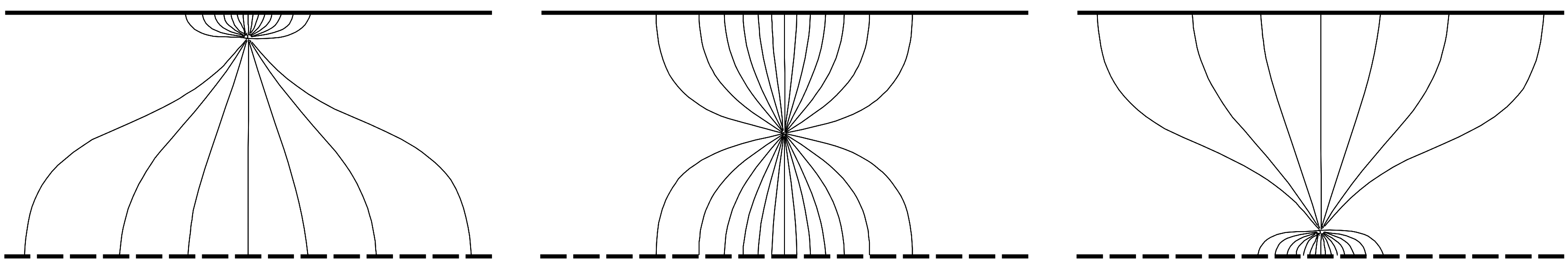


For the same depletion and bias voltages the pulse durations are the same as in strip detectors, although the shapes are very different.

Overbias decreases the collection time.

Varying time delays in signal amplitude can affect timing measurements.

The coupling of the charge increases greatly when the charge comes close to the electrode in strip or pixel detectors.

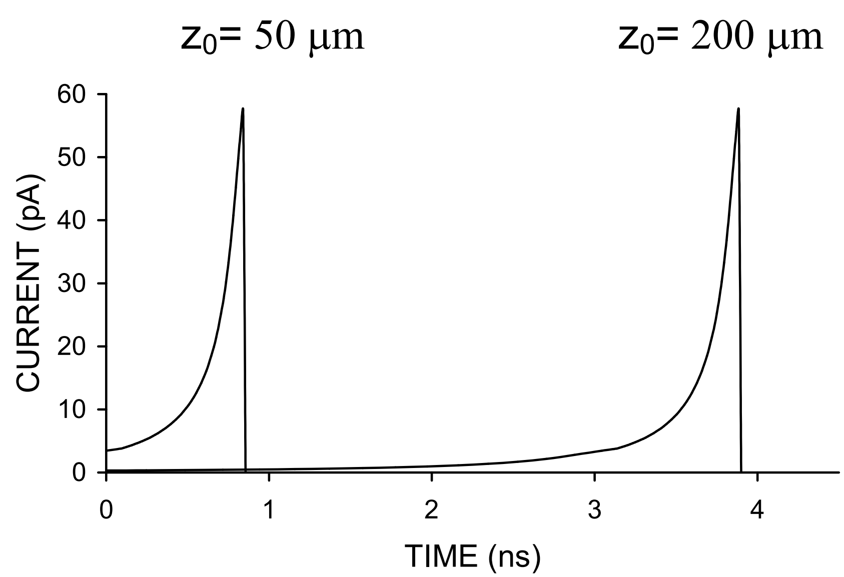


Initially, charge is induced over many strips.

As the charge approaches the strips, the signal distributes over fewer strips.

When the charge is close to the strips, the signal is concentrated over few strips

X-rays deposit localized charge, so this shifts the arrival of the peak amplitude and may shift the triggering time of the timing system.

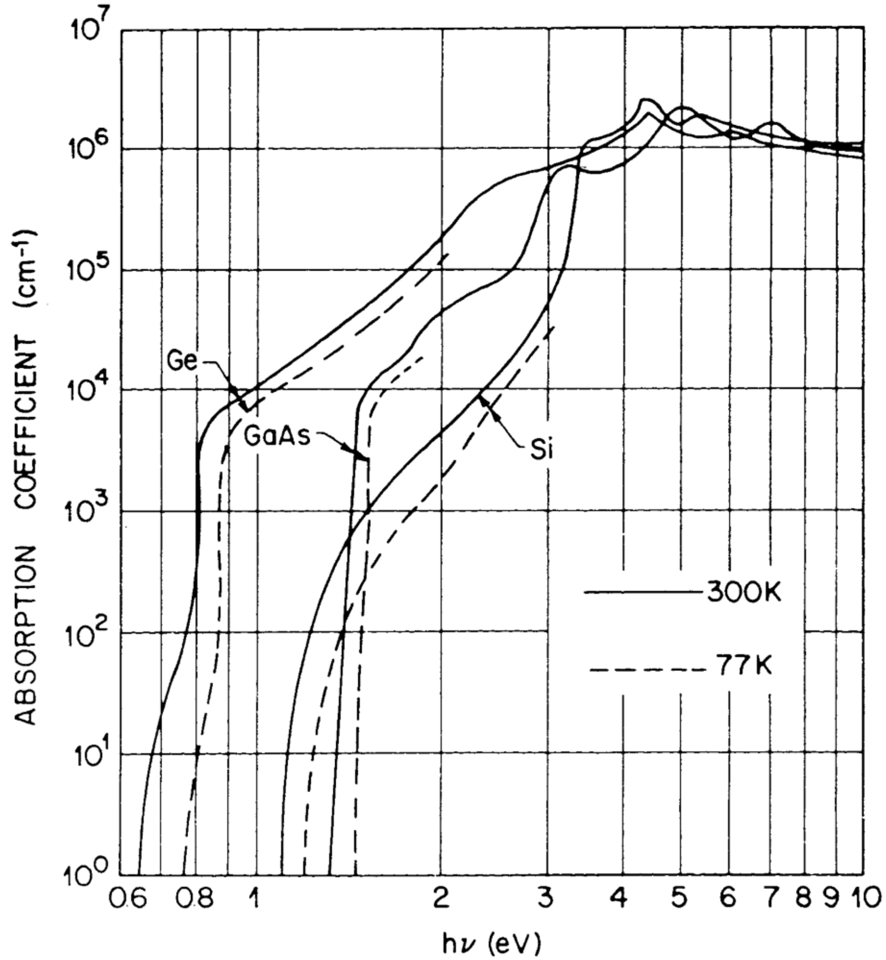


Detector Sensitivity in Semiconductor Detectors

a) Visible light (energies near band gap)

Detection threshold = energy required to produce an electron-hole pair
≈ band gap

In indirect bandgap semiconductors (Si), additional momentum required: provided by phonons



(from Sze, Semiconductor Physics)

b) High energy quanta ($E \gg E_g$)

It is experimentally observed that the energy required to form an electron-hole pair exceeds the bandgap.

In Si: $E_i = $ 3.6 eV ($E_g = $1.1 eV)

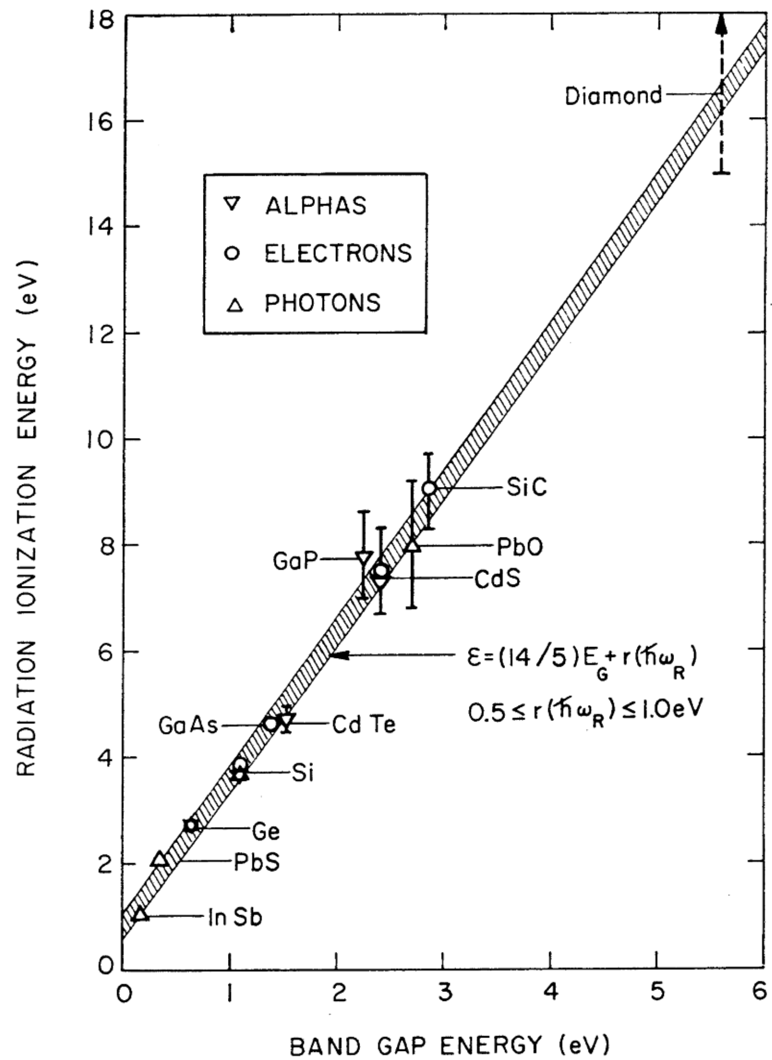
Why?

When particle deposits energy one must conserve both

energy and momentum

momentum conservation not fulfilled by transition across gap

$\Rightarrow$ excite phonons
(lattice vibrations, i.e. heat)



A. Klein, J. Applied Physics **39** (1968) 2029

Signal Fluctuations: Intrinsic Resolution of Semiconductor Detectors

$$\Delta E_{FWHM} = 2.35 \cdot E_i \sqrt{FN_Q} = 2.35 \cdot E_i \sqrt{F \frac{E}{E_i}} = 2.35 \cdot \sqrt{FEE_i}$$

F is the Fano factor, which inherently reduces statistical fluctuations.

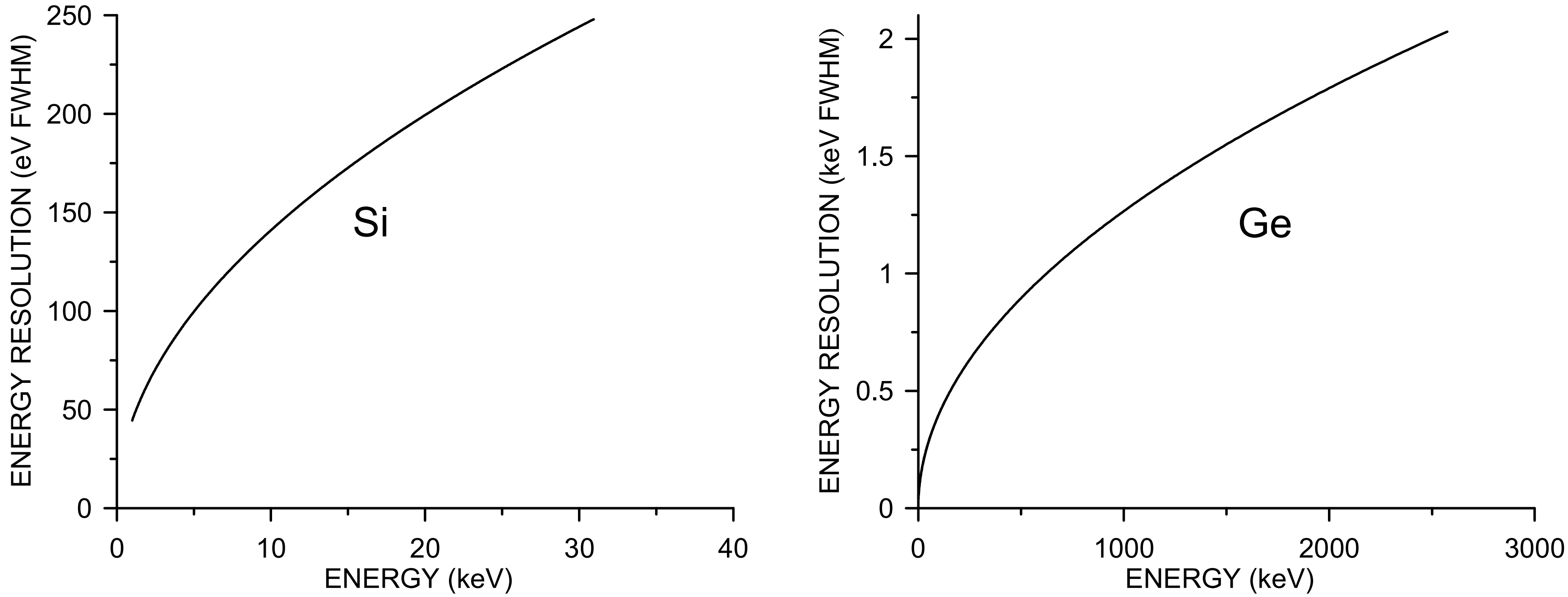
Si: $E_i = 3.6$ eV $F = 0.1$

Ge: $E_i = 2.9$ eV $F = 0.1$

The Fano factor is small in Si and Ge, because a large portion of the energy adds phonons with a much smaller (~meV) excitation level (Chapter 2, pp 52-55).

For comparison, in liquid Xenon $F \approx 20$, whereas in Xe gas $F \approx 0.15$.

Inherent Detector Energy Resolution



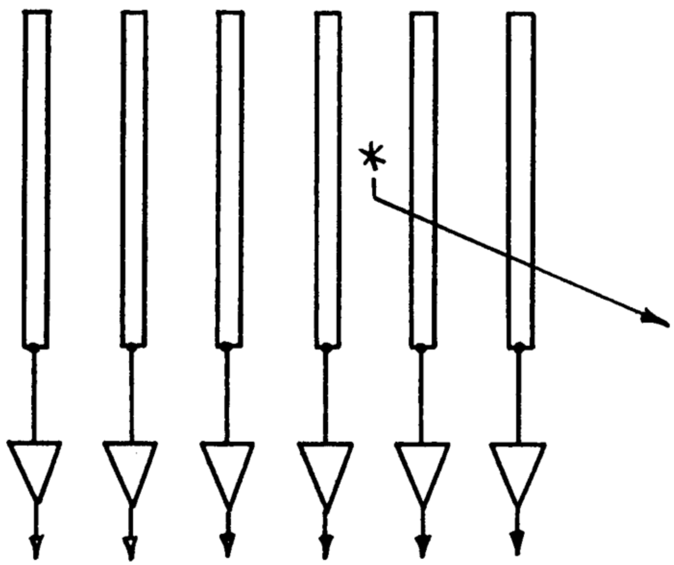
Detectors with good efficiency in the 10s of keV range can have sufficiently small capacitance to allow electronic noise of ~100 eV FWHM, so the variance of the detector signal is a significant contribution.

At energies >100 keV the detector sizes required tend to increase the electronic noise to dominant levels.

In addition to energy measurements, semiconductor detectors allow precision position sensing.

Resolution determined by precision of micron scale patterning of the detector electrodes.

Two options: Binary Readout Analog Readout



Interpolation yields resolution < pitch

Relies on transverse diffusion

$$\sigma_x \propto \sqrt{t_{coll}}$$

e.g. in Si: $t_c \approx$10 ns $\Rightarrow \sigma_x$= 5 μm

depends on S/N and p

p= 25 μm and S/N=50

$\Rightarrow$ 3 – 4 μm resolution

to discriminators

Position resolution determined directly by pitch p: $\sigma_x = p/\sqrt{12}$

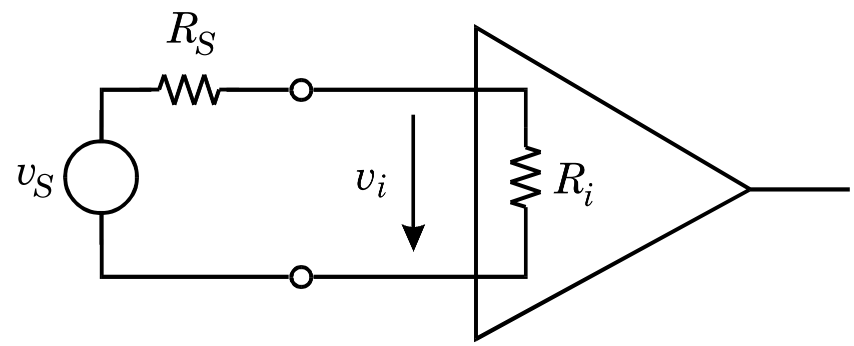
Pixel detectors yield 2-dimensional resolution.

More electronics required, but usually practical at the same total energy per unit area.

3. Signal Acquisition

Amplifier Types

a) Voltage-Sensitive Amplifier



The signal voltage at the amplifier input

$$v_i = \frac{R_i}{R_S + R_i} v_S$$

If the signal voltage at the amplifier input is to be approximately equal to the signal voltage

$$v_i \approx v_S \quad \Rightarrow \quad R_i \gg R_S$$

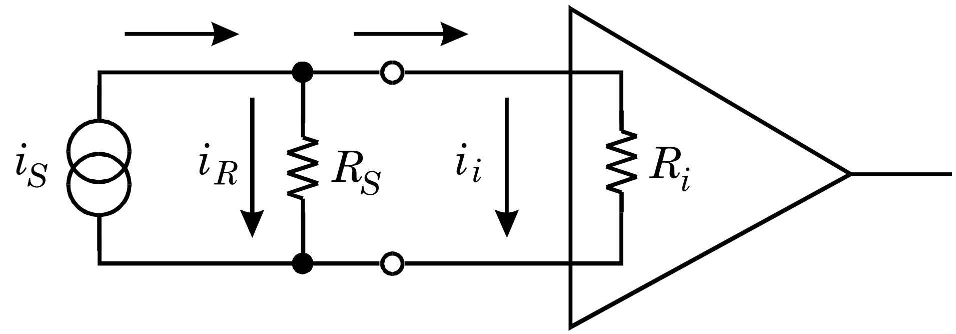
To operate in the voltage-sensitive mode, the amplifier's input resistance (or impedance) must be large compared to the source resistance (impedance).

In ideal voltage amplifiers one sets $R_i = \infty$, although this is never true in reality, although it can be fulfilled to a good approximation.

To provide a voltage output, the amplifier should have a low output resistance, i.e. its output resistance should be small compared to the input resistance of the following stage.

b) Current-Sensitive Amplifier

The signal current divides into the source resistance and the amplifier's input resistance. The fraction of current flowing into the amplifier



$$i_i = \frac{R_s}{R_s + R_i} i_S$$

If the current flowing into the amplifier is to be approximately equal to the signal current

$$i_i \approx i_S \qquad \Rightarrow \qquad R_i \ll R_S$$

To operate in the current-sensitive mode, the amplifier's input resistance (or impedance) must be small compared to the source resistance (impedance).

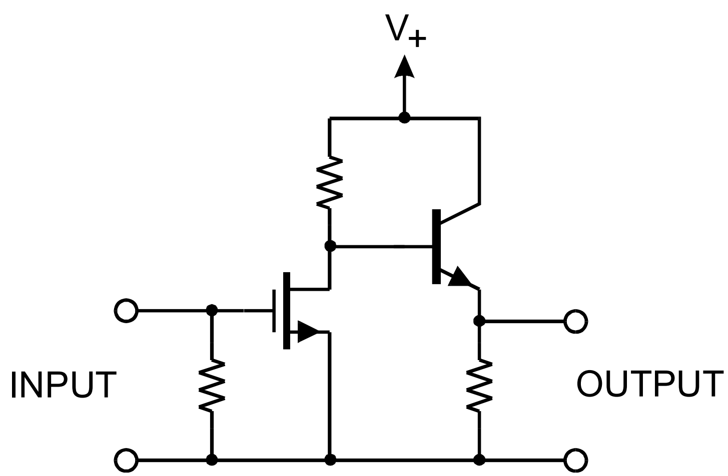
One can also model a current source as a voltage source with a series resistance. For the signal current to be unaffected by the amplifier input resistance, the input resistance must be small compared to the source resistance, as derived above.

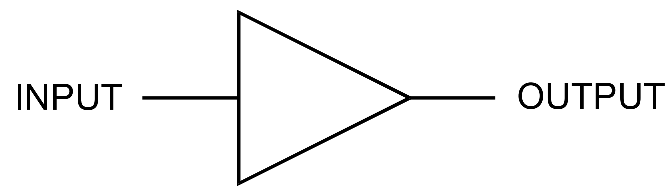
At the output, to provide current drive the output resistance should be high, i.e. large compared to the input resistance of the next stage.

- Whether a specific amplifier operates in the current or voltage mode depends on the source resistance.

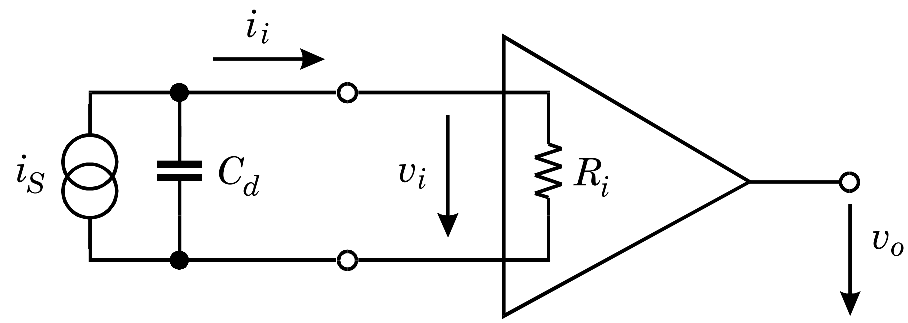
- Amplifiers can be configured as current mode input and voltage mode output or, conversely, as voltage mode input and current mode output. The gain is then expressed as V/A or A/V.

Although an amplifier has a pair of input and a second pair of output connections, since the two have a common connection a simplified representation is commonly used:





c) Voltage and Current Mode with Capacitive Sources



Output voltage:

v_o = (voltage gain A_v) × (input voltage v_i).

Operating mode depends on charge collection time t_c and the input time constant $R_i C_d$:

a) $R_i C_d \ll t_c$

detector capacitance discharges rapidly

$\Rightarrow \quad v_o \propto i_s(t)$

current sensitive amplifier

b) $R_i C_d \gg t_c$

detector capacitance discharges slowly

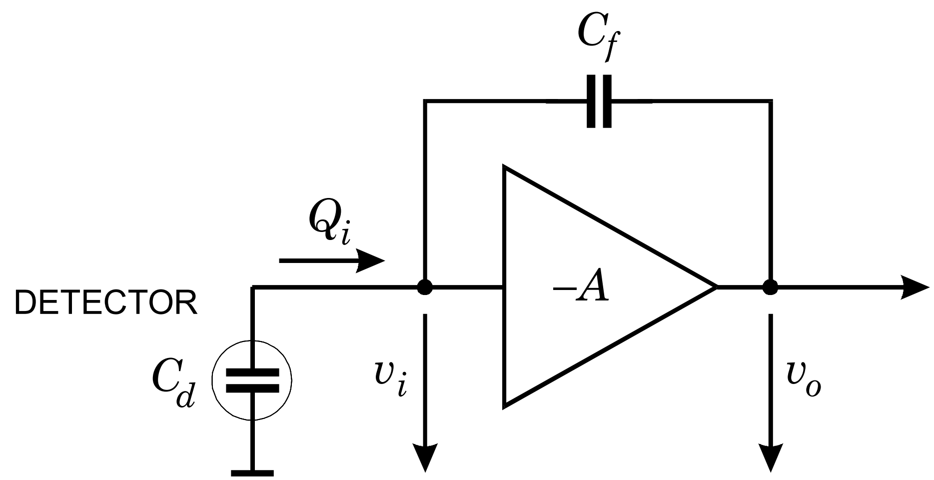
$\Rightarrow \quad v_o \propto \int i_s(t)dt$

voltage sensitive amplifier

Note that in both cases the amplifier is providing voltage gain, so the output signal voltage is determined directly by the input voltage. The difference is that the shape of the input voltage pulse is determined either by the instantaneous current or by the integrated current and the decay time constant.

Goal is to measure signal charge, so it is desirable to use a system whose response is independent of detector capacitance.

Active Integrator ("charge-sensitive amplifier")



Start with inverting voltage amplifier

Voltage gain $dv_o / dv_i = -A \Rightarrow$

$$v_o = -Av_i$$

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

Connect feedback capacitor C_f between output and input.

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

$\Rightarrow$ Charge deposited on C_f: $\quad Q_f = C_f v_f = C_f(A+1)v_i$

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

$\Rightarrow$ Effective input capacitance $\quad C_i = \dfrac{Q_i}{v_i} = C_f(A+1)$ ("dynamic" input capacitance)

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

Charge gain set by a well-controlled quantity, the feedback capacitance.

Q_i is the charge flowing into the preamplifier .... but some charge remains on C_d.

What fraction of the signal charge is measured?

$$\frac{Q_i}{Q_s} = \frac{C_i v_i}{Q_d + Q_i} = \frac{C_i}{Q_s} \cdot \frac{Q_s}{C_i + C_d}$$

$$= \frac{1}{1 + \dfrac{C_d}{C_i}} \approx 1 \quad (\text{if } C_i >> C_d)$$

Example: $A = 10^3$

$C_f = 1$ pF $\Rightarrow$ $C_i = 1$ nF

$C_{det} = 10$ pF: $Q_i / Q_s = 0.99$

$C_{det} = 500$ pF: $Q_i / Q_s = 0.67$

$\uparrow$

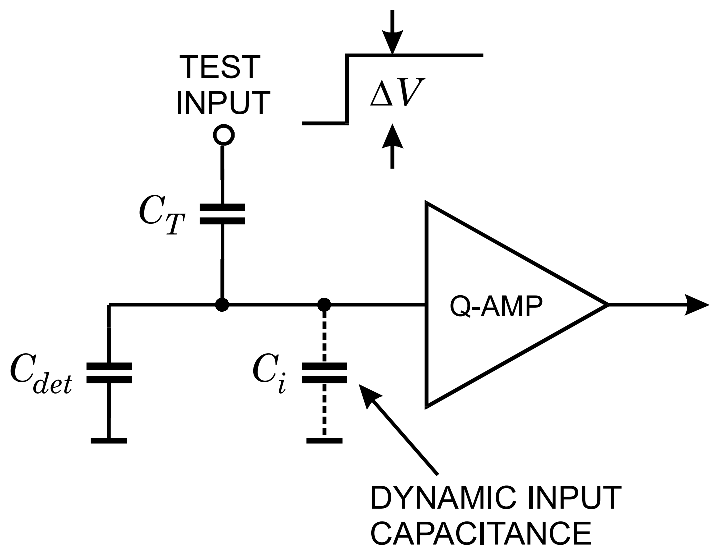
Si Det.: 50 μm thick, 250 mm^2 area

Note: Input coupling capacitor must be $\gg C_i$ for high charge transfer efficiency.

Calibration

Inject specific quantity of charge - measure system response

Use voltage pulse (can be measured conveniently with oscilloscope)



$C_i >> C_T \Rightarrow$ Voltage step applied to test input develops over C_T.

$$\Rightarrow \quad Q_T = \Delta V \cdot C_T$$

Accurate expression: $$Q_T = \frac{C_T}{1+\frac{C_T}{C_i}} \cdot \Delta V \approx C_T \left(1 - \frac{C_T}{C_i}\right) \Delta V$$

Typically: $C_T / C_i = 10^{-3} - 10^{-4}$

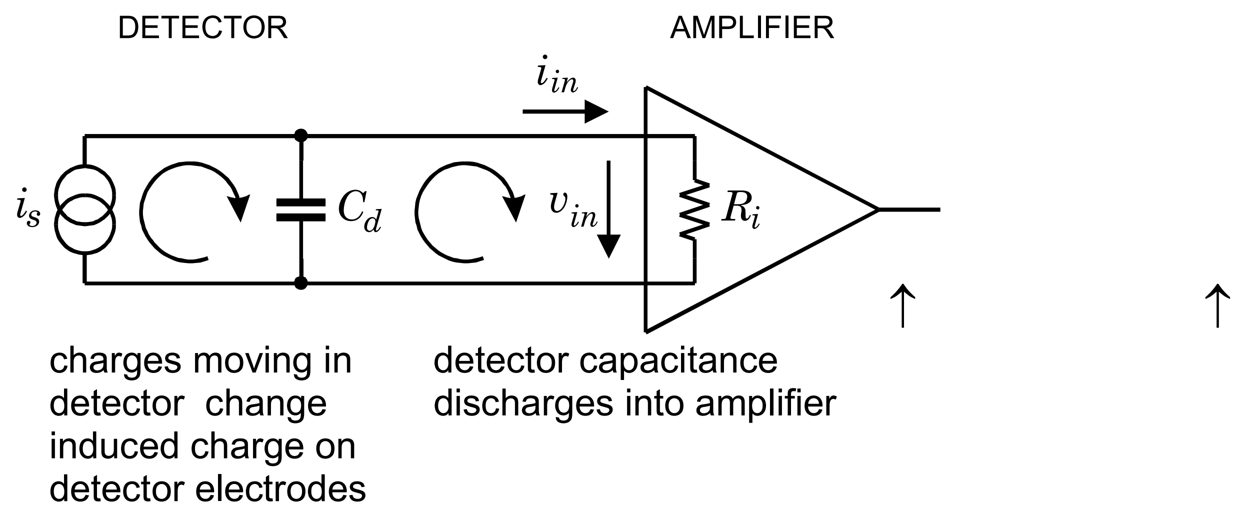
Realistic Charge-Sensitive Preamplifiers

The preceding discussion assumed idealized amplifiers with infinite speed.

In reality, amplifiers may be too slow to follow the instantaneous detector pulse.

Does this incur a loss of charge?

Equivalent Circuit:

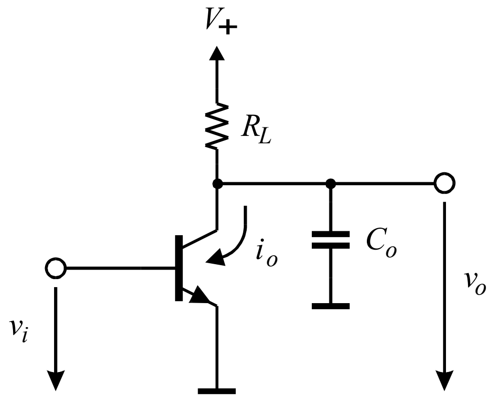


Signal is preserved even if the amplifier responds much more slowly than the detector signal.

However, the response of the amplifier affects the measured pulse shape.

- How do “real” amplifiers affect the measured pulse shape?
- How does the detector affect amplifier response?

A Simple Amplifier



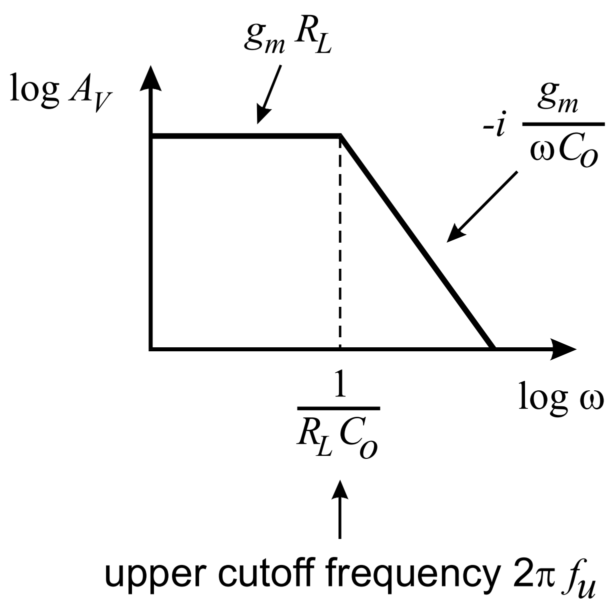
Voltage gain: $$A_V = \frac{dv_o}{dv_i} = \frac{di_o}{dv_i} \cdot Z_L \equiv g_m Z_L$$

$g_m \equiv$ transconductance

$$Z_L = R_L \, // \, C_o$$

$$\frac{1}{Z_L} = \frac{1}{R_L} + \mathbf{i}\omega \, C_o$$

Gain vs. Frequency

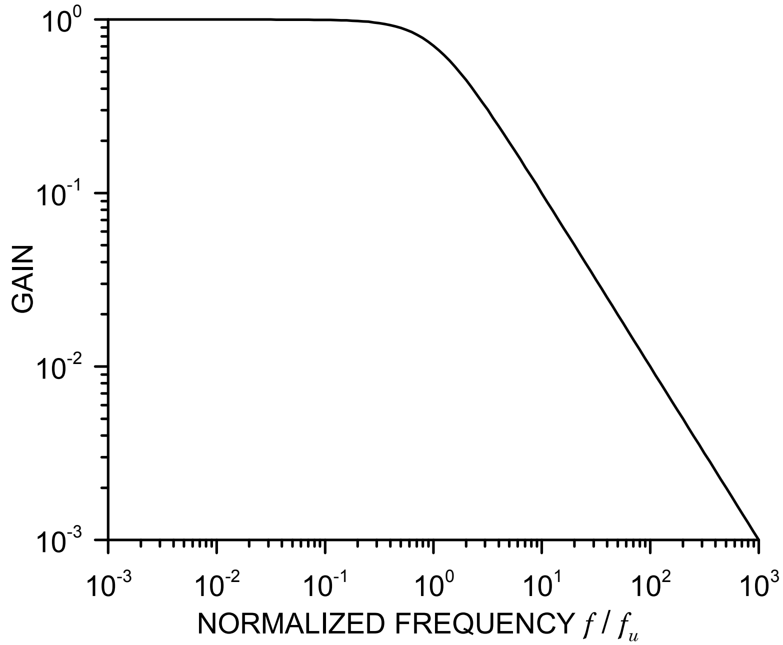


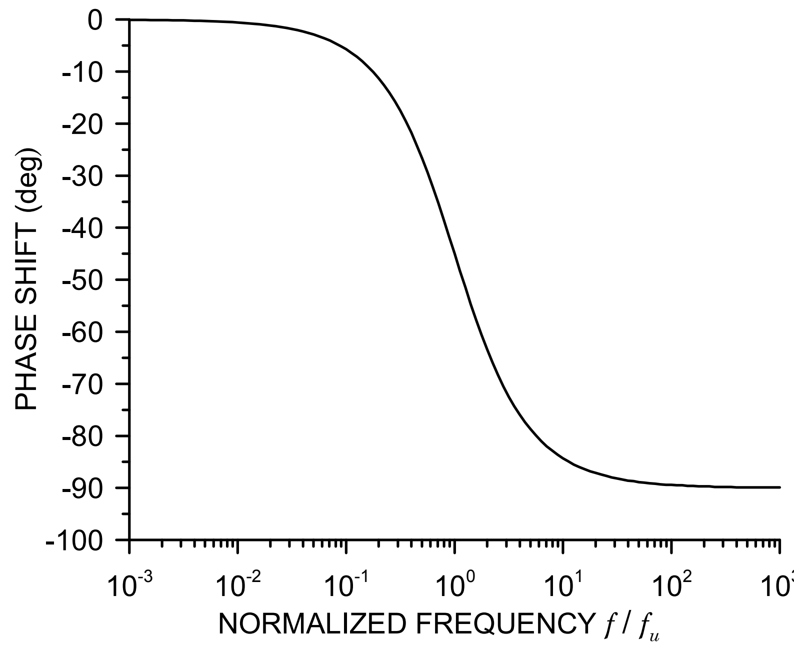
$$\Rightarrow \quad A_V = g_m \left(\frac{1}{R_L} \; + \; \mathbf{i}\omega \, C_o \right)^{-1}$$

↑ low freq. ↑ high freq.

Note: **i** indicates the 90° phase shift between capacitor voltage and current

Frequency and phase response:





Phase shows change from low-frequency response. For an inverting amplifier add 180°.

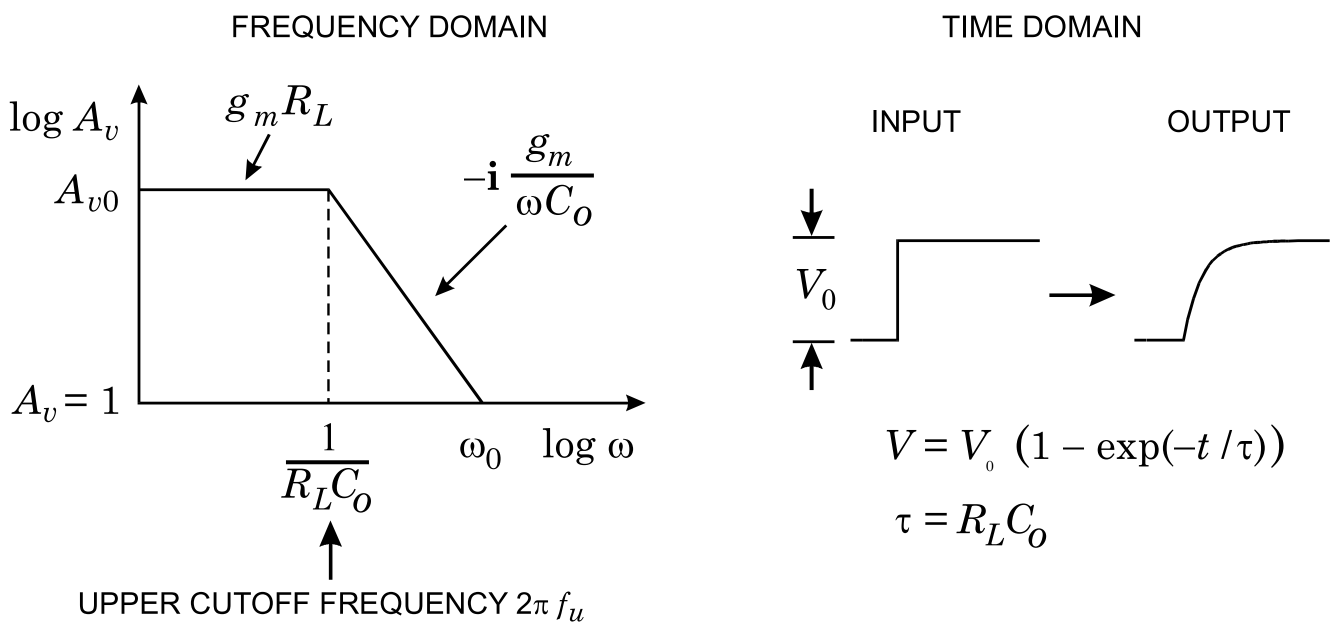
The corner (cutoff) frequency is often called a “pole”.

Pulse Response of the Simple Amplifier

A voltage step $v_i(t)$ at the input causes a current step $i_o(t)$ at the output of the transistor.

For the output voltage to change, the output capacitance C_o must first charge up.

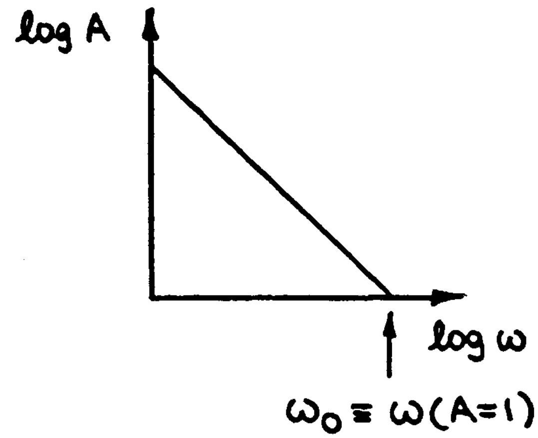
$\Rightarrow$ The output voltage changes with a time constant $\tau = R_L C_O$



The time constant τ corresponds to the upper cutoff frequency : $\tau = \dfrac{1}{2\pi f_u}$

Input Impedance of a Charge-Sensitive Amplifier

Input impedance $$Z_i = \frac{Z_f}{A+1} \approx \frac{Z_f}{A} \quad (A >> 1)$$



Gain-Bandwidth Product

Amplifier gain vs. frequency beyond the upper cutoff frequency

$$A = -\mathbf{i}\ \frac{\omega_0}{\omega}$$

Feedback impedance $$Z_f = -\mathbf{i}\ \frac{1}{\omega\ C_f}$$

⇒ Input Impedance $$Z_i = -\frac{\mathbf{i}}{\omega\ C_f} \cdot \frac{1}{-\mathbf{i}\,\frac{\omega_0}{\omega}} = \frac{1}{\omega_0 C_f}$$

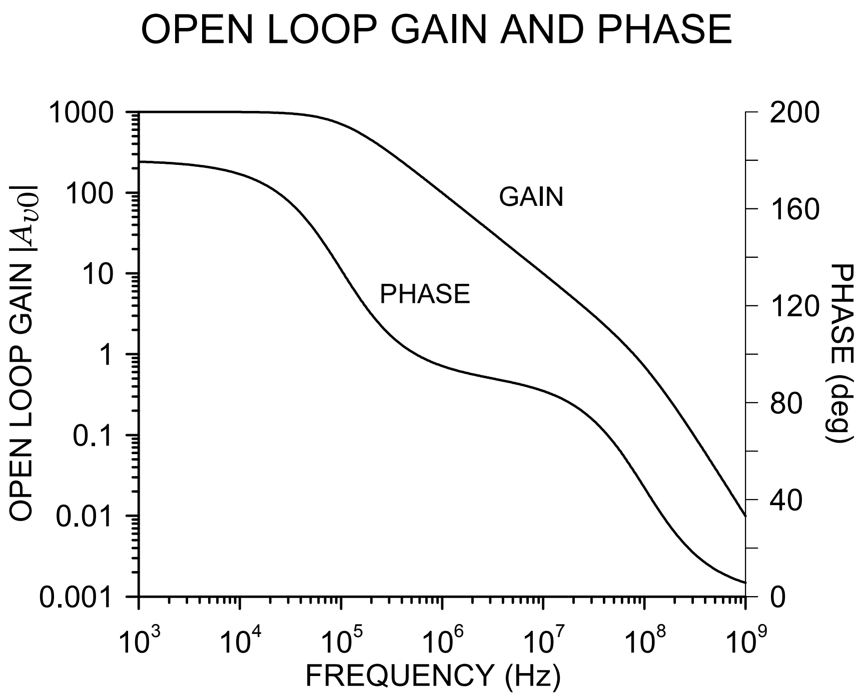
Imaginary component vanishes ⇒ *Resistance:* $Z_i \to R_i$

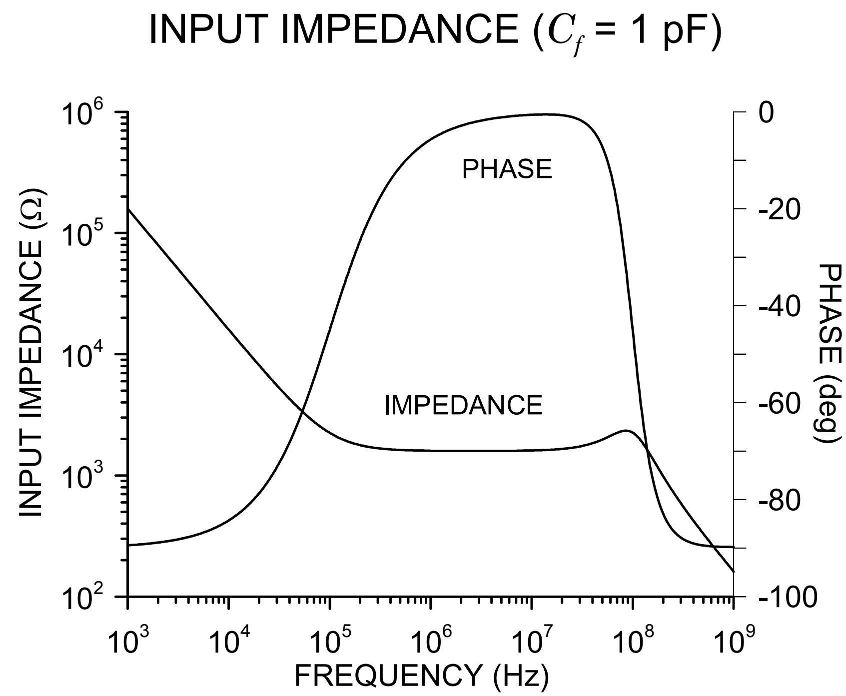
⇒ low frequencies ($f < f_u$): capacitive input
high frequencies ($f > f_u$): resistive input

Practically all charge-sensitive amplifiers operate in the 90° phase shift regime.
⇒ Resistive input

However ... Note that the input impedance varies with frequency.

Example: cutoff frequencies at 10 kHz and 100 MHz, low frequency gain = 10^3

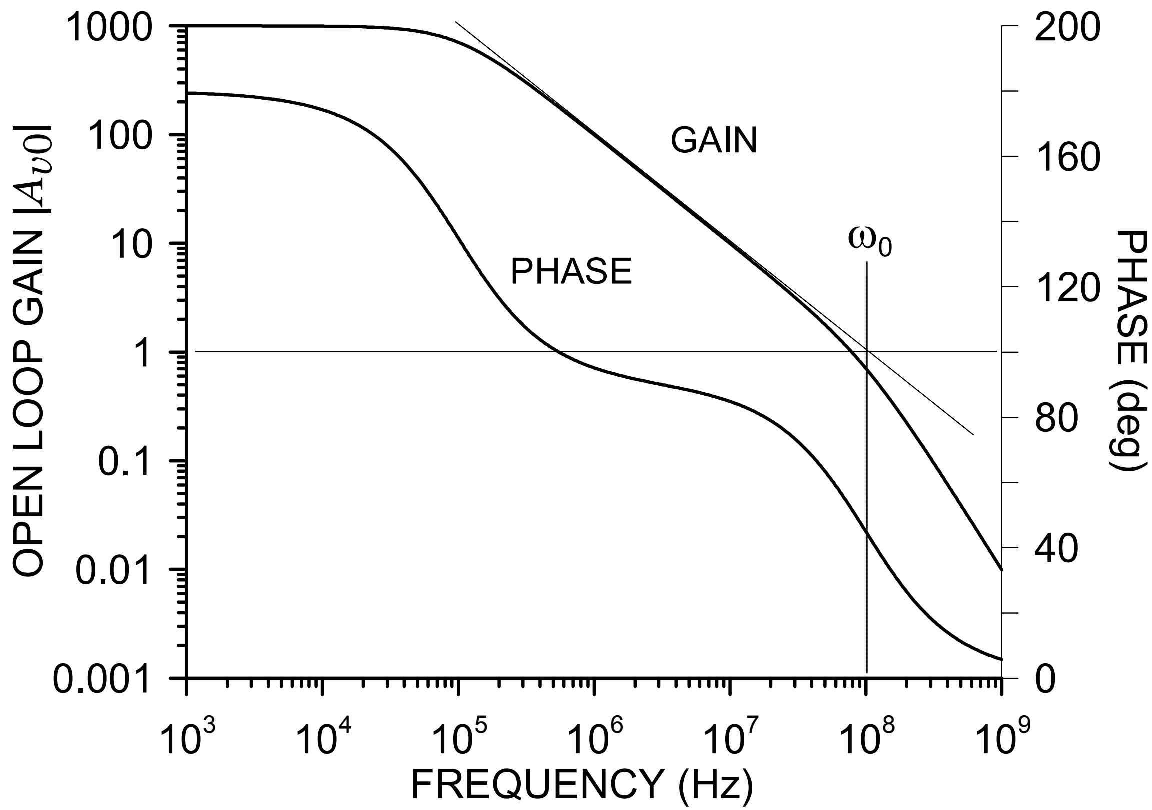




In the resistive regime the input impedance

$$Z_i = \frac{1}{\omega_0 C_f} ,$$

where C_f is the feedback capacitance and ω_0 is the extrapolated unity gain frequency in the 90° phase shift regime.



Time Response of a Charge-Sensitive Amplifier

Input resistance and detector capacitance form RC time constant:

$$\tau_i = R_i C_D$$

$$\tau_i = \frac{1}{\omega_0 C_f} \cdot C_D$$

$\Rightarrow$ Rise time increases with detector capacitance.

Or apply feedback theory:

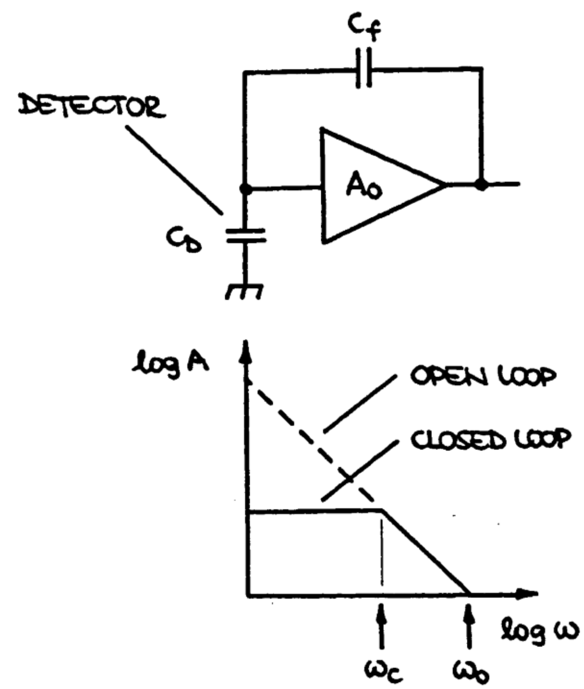
Closed Loop Gain $A_f = \frac{C_D + C_f}{C_f} \quad (A_f << A_0)$

$$A_f \approx \frac{C_D}{C_f} \quad (C_D >> C_f)$$

Closed Loop Bandwidth $\omega_C A_f = \omega_0$

Response Time $\tau_{amp} = \frac{1}{\omega_C} = C_D \frac{1}{\omega_0 C_f}$

Same result as from input time constant.



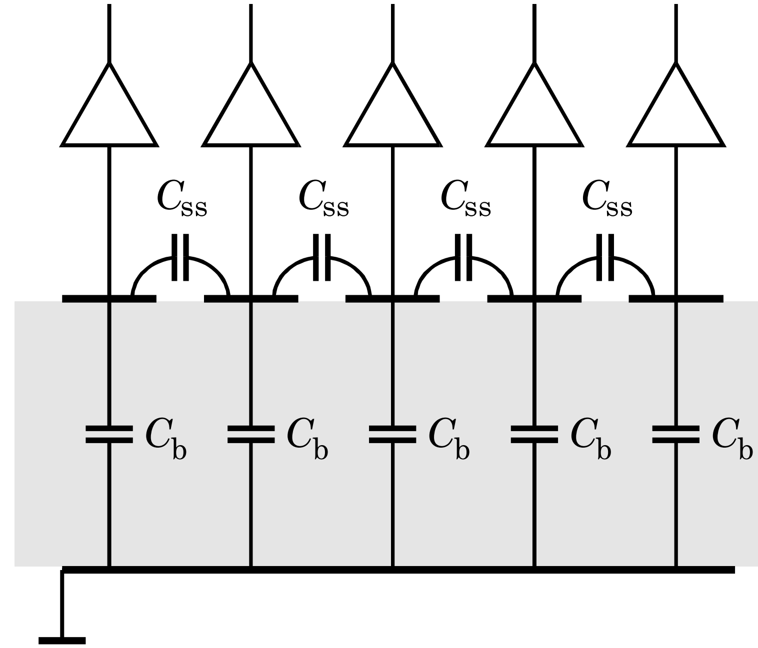
Importance of input impedance in strip and pixel detectors:

Amplifiers must have a low input impedance to reduce transfer of charge through capacitance to neighboring strips

In the previous example at
8 MHz ($\hat{=}$ ~20 ns peaking time)
$Z_i \approx 1.6\ \mathrm{k}\Omega$, corresponding to 12 pF

$\Rightarrow$ with 6 cm long strips about half of the signal current will go to the neighbors.

STRIP DETECTOR



For strip pitches that are smaller than the bulk thickness, the capacitance is dominated by the fringing capacitance to the neighboring strips C_{SS}.

Typically: 1 – 2 pF/cm for strip pitches of 25 – 100 μm on Si.

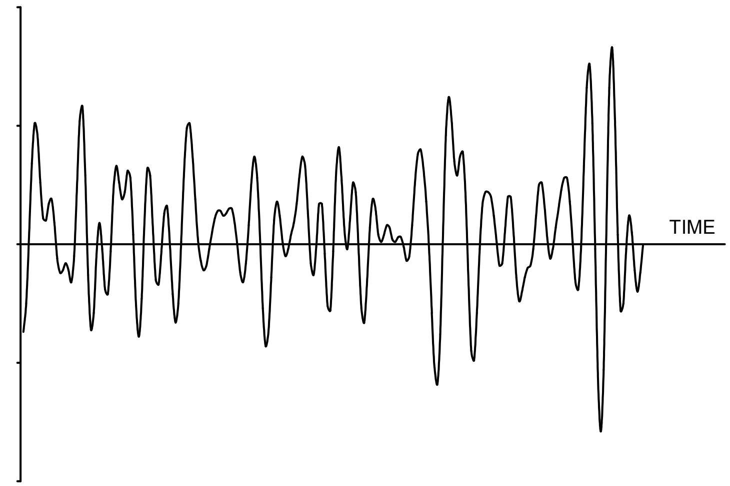
The backplane capacitance C_b is typically 20% of the strip-to-strip capacitance.

Negligible cross-coupling at shaping times $T_P > (2\ ...\ 3) \times R_i C_D$ and if $C_i \gg C_D$.

III. Electronic Noise

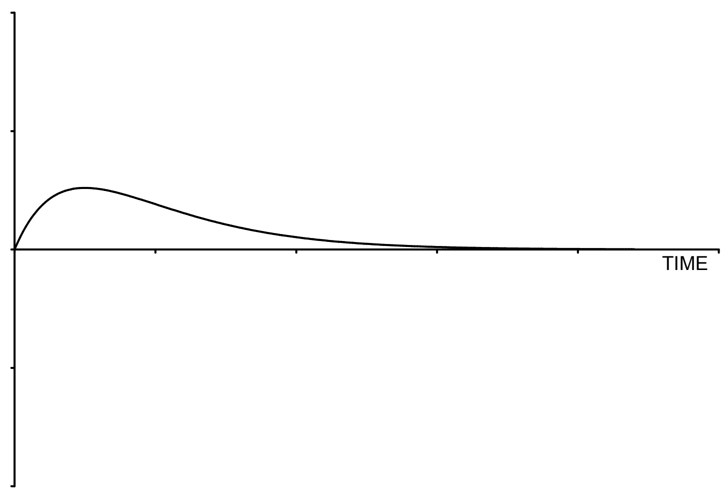
Choose a time when no signal is present.

Amplifier's quiescent output level (baseline):

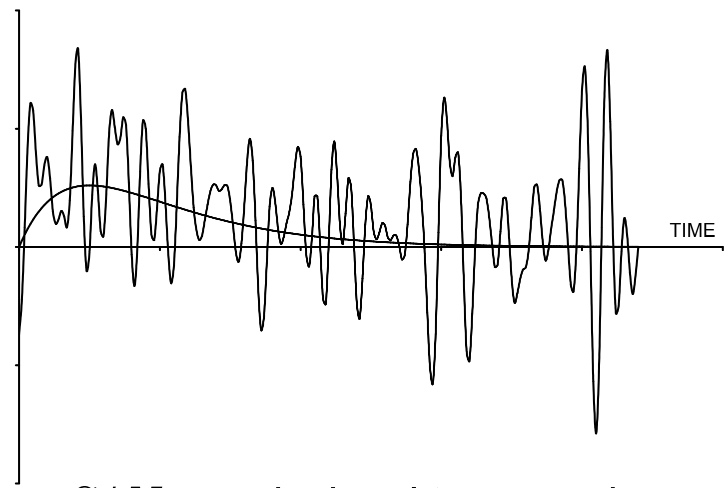


In the presence of a signal, noise + signal add.

Signal



Signal+Noise (S/N = 1)



$S/N \equiv$ peak signal to rms noise

Measurement of peak amplitude yields signal amplitude + noise fluctuation

The preceding example could imply that the fluctuations tend to increase the measured amplitude, since the noise fluctuations vary more rapidly than the signal.

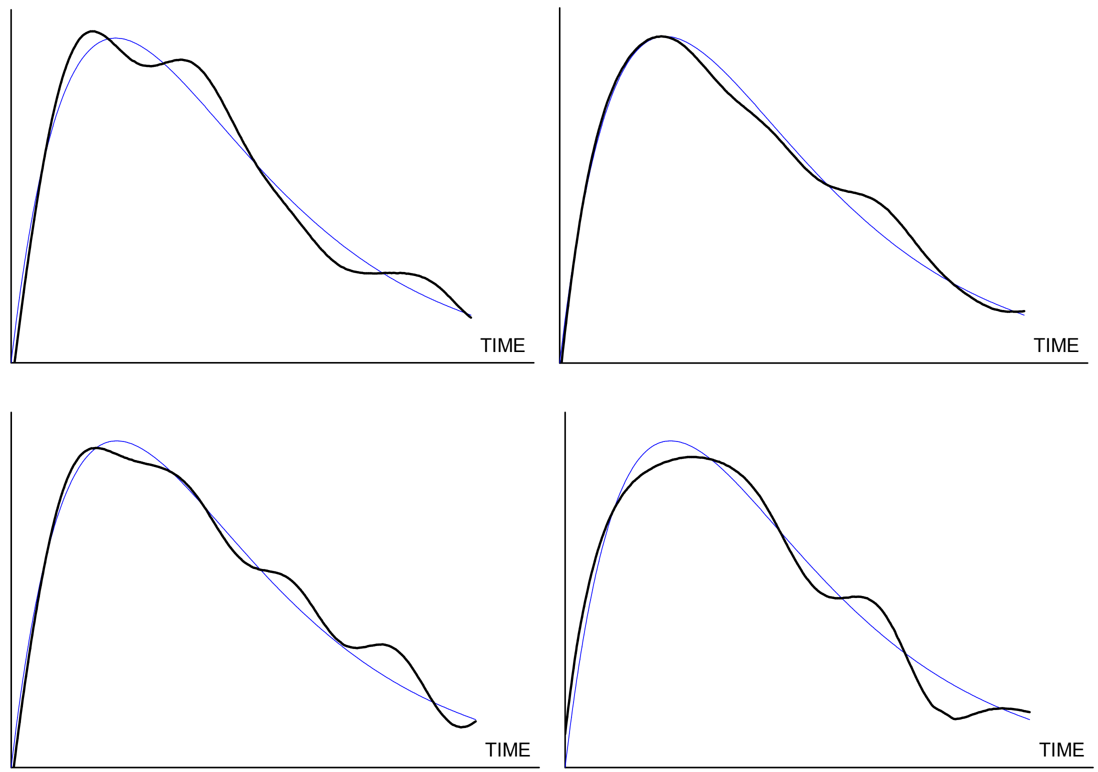
In an optimized system, the time scale of the fluctuation is comparable to the signal peaking time.

Then the measured amplitude fluctuates positive and negative relative to the ideal signal.

Measurements taken at 4 different times:

noiseless signal superimposed for comparison

$S/N = 20$



Noise affects

Peak signal

Time distribution

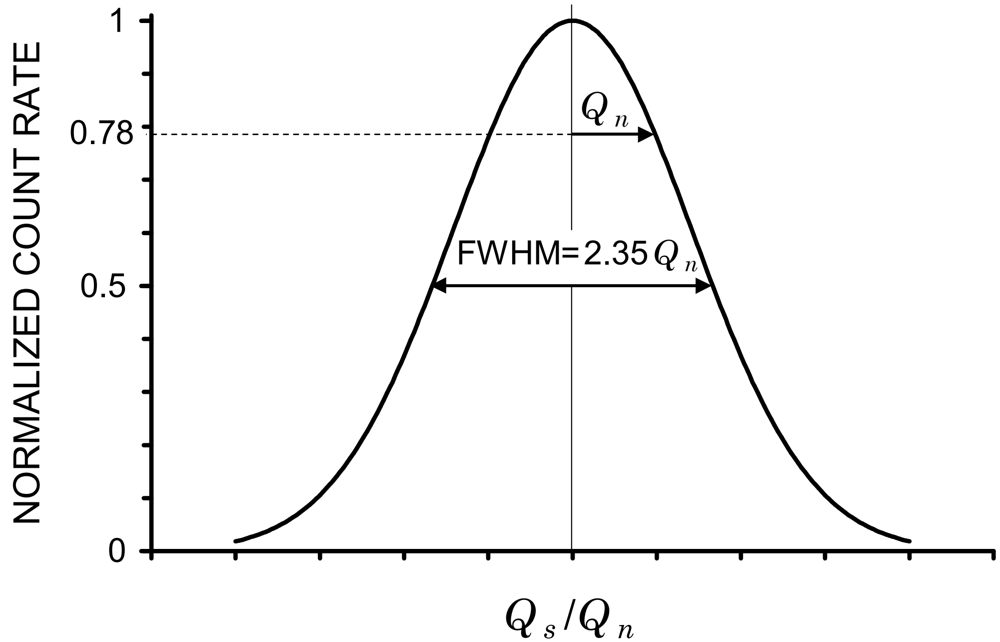
Electronic noise is purely random.

⇒ amplitude distribution is Gaussian

⇒ noise modulates baseline

⇒ baseline fluctuations superimposed on signal

⇒ output signal has Gaussian distribution



Measuring Resolution

Inject an input signal with known charge using a pulse generator set to approximate the detector signal shape.

Measure the pulse height spectrum.

peak centroid ⇒ signal magnitude
peak width ⇒ noise (FWHM= 2.35 Q_n)

Basic Noise Mechanisms and Characteristics

Consider n carriers of charge e moving with a velocity v through a sample of length l. The induced current i at the ends of the sample is

$$i = \frac{n\ e\ v}{l}$$

The fluctuation of this current is given by the total differential

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

where the two terms are added in quadrature since they are statistically uncorrelated.

Two mechanisms contribute to the total noise:

- velocity fluctuations, *e.g.* thermal noise
- number fluctuations, *e.g.* shot noise
excess or "1/f" noise

Thermal noise and shot noise are both "white" noise sources, i.e.

power per unit bandwidth (≡ spectral density) is constant: $$\frac{dP_{noise}}{df} = const.$$

1. Thermal Noise in Resistors

The most common example of noise due to velocity fluctuations is the thermal noise of resistors.

Noise power density vs. frequency f: $$\frac{dP_{noise}}{df} = 4kT$$

k = Boltzmann constant

T = absolute temperature

R = DC resistance

since $P = \frac{V^2}{R} = I^2 R$

the spectral noise voltage density $$\frac{dV^2_{noise}}{df} \equiv e_n^2 = 4kTR$$

and the spectral noise current density $$\frac{dI^2_{noise}}{df} \equiv i_n^2 = \frac{4kT}{R}$$

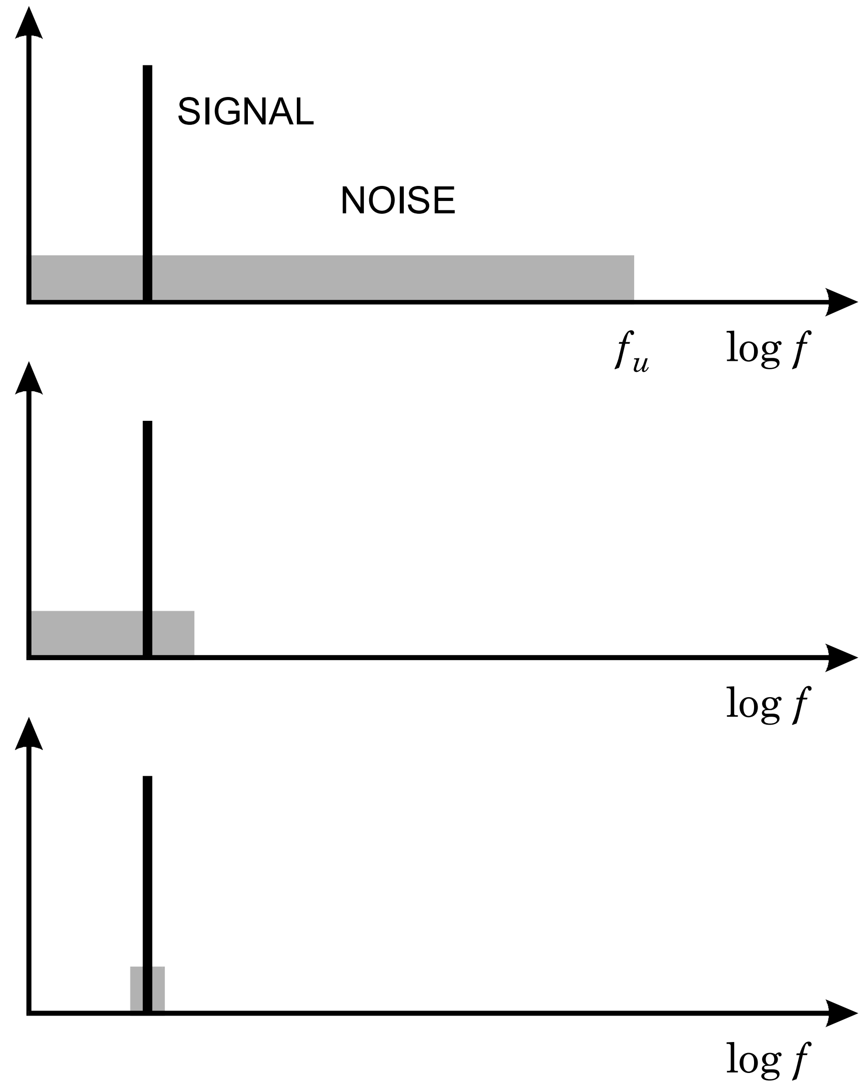
The total noise depends on the bandwidth of the system.
For example, the total noise voltage at the output of a voltage amplifier with the frequency dependent gain $A_v(f)$ is

$$v_{on}^2 = \int_0^\infty e_n^2 \; A_v^2(f) \; df$$

Note: Since spectral noise components are not correlated, one must integrate over the noise power (proportional to voltage or current squared).

Total noise increases with bandwidth.

Total noise is the integral over the shaded region.



S/N increases as noise bandwidth is reduced until signal components are attenuated significantly.

2. Shot noise

A common example of noise due to number fluctuations is “shot noise”, which occurs whenever carriers are injected into a sample volume independently of one another.

Example: current flow in a semiconductor diode
(emission over a barrier)

Spectral noise current density: $i_n^2 = 2eI$

e = electronic charge
I = DC current

When measuring signal charge the signal pulses are integrated over time. The shot noise then results from the statistical fluctuations of total number of injected carriers during the integration time, so the noise increases with the square root of shaping time.

Note: Shot noise does not occur in “ohmic” conductors. Since the number of available charges is not limited, the fields caused by local fluctuations in the charge density draw in additional carriers to equalize the total number.

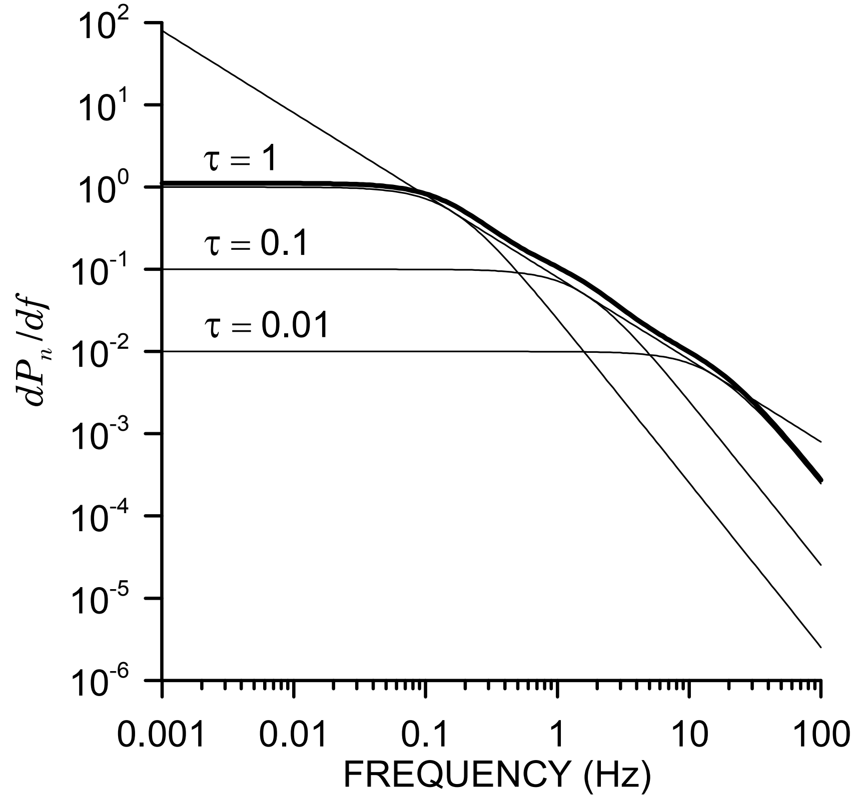
3. Low Frequency ("$1/f$") Noise

Charge can be trapped and then released after a characteristic lifetime τ.

The spectral density for a single lifetime

$$S(f) \propto \frac{\tau}{1+(2\pi f \tau)^2} \ .$$

For $2\pi f \tau \gg 1$: $S(f) \propto \dfrac{1}{f^2}$.



However,
several traps with different time constants can yield a "$1/f$" distribution:

Traps with three time constants of 0.01, 0.1 and 1 s yield a $1/f$ distribution over two decades in frequency.

Low frequency noise is ubiquitous – must not have $1/f$ dependence, but commonly called $1/f$ noise.

Spectral power density: $\dfrac{dP_{noise}}{df} = \dfrac{1}{f^{\alpha}}$ (typically α = 0.5 – 2)

4. Noise Bandwidth vs. Signal Bandwidth

Consider an amplifier with the frequency response $A(f)$. This can be rewritten

$A(f) \equiv A_0 G(f)$,

where A_0 is the maximum gain and $G(f)$ describes the frequency response.

For example, for the simple amplifier described above

$$A_v = g_m \left(\frac{1}{R_L} + \mathbf{i}\omega\, C_o \right)^{-1} = g_m R_L \, \frac{1}{1 + \mathbf{i}\omega R_L C_o}$$

and using the above convention $A_0 \equiv g_m R_L \quad \text{and} \quad G(f) \equiv \dfrac{1}{1 + \mathbf{i}\,(2\pi f\, R_L C_o)}$

If a "white" noise source with spectral density e_{ni} is present at the input, the total noise voltage at the output is

$$v_{no} = \sqrt{\int_0^\infty e_{ni}^2 \left| A_0 G(f) \right|^2 df} = e_{ni}\, A_0 \sqrt{\int_0^\infty G^2(f) df} \equiv e_{ni}\, A_0 \sqrt{\Delta f_n}$$

Δf_n **is the "noise bandwidth".**

Note that, in general, the noise bandwidth and the signal bandwidth are not the same.

If the upper cutoff frequency is determined by a single *RC* time constant, as in the "simple amplifier",

the signal bandwidth $$\Delta f_s = f_u = \frac{1}{2\pi\ RC}$$

and the noise bandwidth $$\Delta f_n = \frac{1}{4RC} = \frac{\pi}{2} f_u \quad .$$

Noise Bandwidth and Low Frequency (1/*f*) Noise

For a spectral noise density $$P_{nf} = \frac{S_f}{f}$$

and a corresponding voltage density $$e_{nf}^2 = \frac{A_f}{f}$$

the total noise integrated in a frequency band f_1 to f_2 is

$$v_{nf}^2 = \int_{f_1}^{f_2} \frac{A_f}{f}\, df = A_f \log\left(\frac{f_2}{f_1}\right)$$

Thus, for a 1/*f* spectrum the total noise depends on the ratio of the upper to lower cutoff frequency.

Since this is a power distribution, the voltage or current spectral density changes 10-fold over a 100-fold span in frequency.

Frequently, the 1/*f* noise corner is specified: frequency where 1/*f* noise intercepts white noise.

Higher white noise level reduces corner frequency, so lower noise corner does not equate to lower 1/*f* noise.

“Noiseless” Resistance Example – Dynamic Resistance

In many instances a resistance is formed by the slope of a device’s current-voltage characteristic, rather than by a static ensemble of electrons agitated by thermal energy.

Example: forward-biased semiconductor diode

Diode current vs. voltage $I = I_0 (e^{q_e V / kT} - 1)$

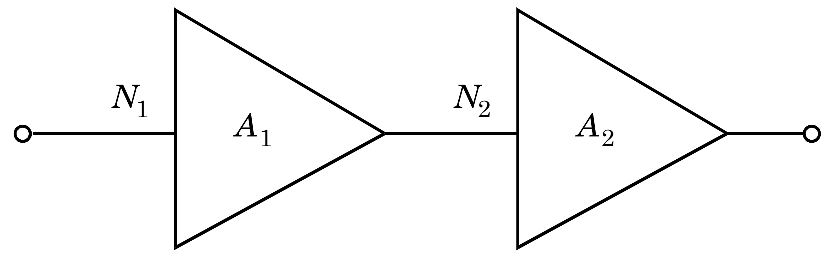
The differential resistance $r_d = \frac{dV}{dI} = \frac{kT}{q_e I}$

i.e. at a given current the diode presents a resistance, e.g. 26 Ω at I = 1 mA and T = 300 K.

Note that two diodes can have different charge carrier concentrations, but will still exhibit the same dynamic resistance at a given current, so the dynamic resistance is not uniquely determined by the number of carriers, as in a resistor.

There is no thermal noise associated with this “dynamic” resistance, although the current flow carries shot noise.

Noise in Amplifier Chains



Consider a chain of two amplifiers (or amplifying devices), with gains A_1 and A_1, and input noise levels N_1 and N_2 .

A signal S is applied to the first amplifier, so the input signal-to-noise ratio is S/N_1.

At the output of the first amplifier the signal is A_1S and the noise A_1N.

Uncorrelated noise components add in quadrature, e.g. $V_{n,tot} = \sqrt{\sum_i V_{ni}^2}$

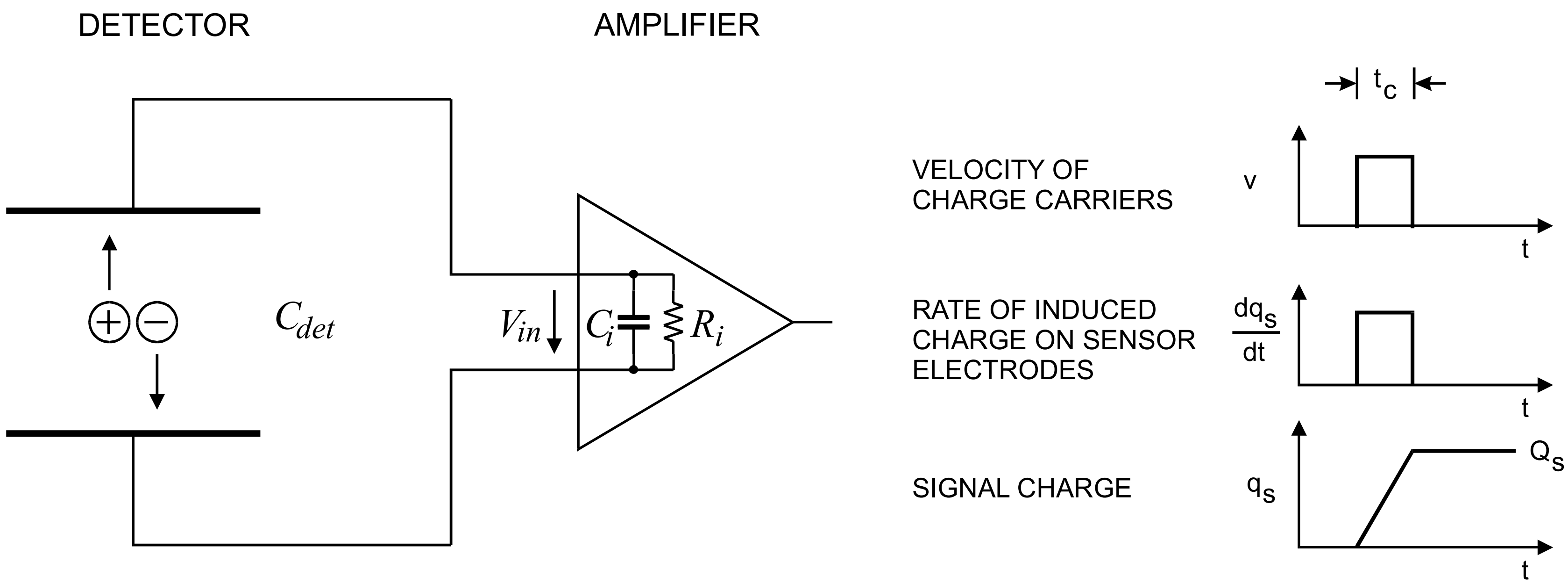
Both are amplified by the second amplifier, but in addition the second amplifier contributes its noise, so the signal-to-noise ratio at the output of the second amplifier

$$\left(\frac{S}{N}\right)^2 = \frac{(SA_1A_2)^2}{(N_1A_1A_2)^2 + (N_2A_2)^2} = \frac{S^2}{N_1^2 + \left(\frac{N_2}{A_1}\right)^2} = \left(\frac{S}{N_1}\right)^2 \frac{1}{1 + \left(\frac{N_2}{A_1N_1}\right)^2}$$

The overall sign-to-noise ratio is reduced, but the noise contribution from the second-stage can be negligible, provided the gain of the first stage is sufficiently high.

$\Rightarrow$ In a well-designed system the noise is dominated by the first gain stage.

5. Signal-to-Noise Ratio vs. Detector Capacitance



if $R_i \times (C_{det} + C_i) \gg$ collection time,

peak voltage at amplifier input $V_{in} = \dfrac{Q_s}{C} = \dfrac{\int i_s dt}{C} = \dfrac{Q_s}{C_{det} + C_i}$

↑

Magnitude of voltage depends on total capacitance at input!

The peak amplifier signal V_S is inversely proportional to the **total capacitance at the input**, i.e. the sum of

1. detector capacitance,
2. input capacitance of the amplifier, and
3. stray capacitances.

Assume an amplifier with a noise voltage v_n at the input.

Then the signal-to-noise ratio

$$\frac{S}{N} = \frac{V_S}{v_n} \propto \frac{1}{C}$$

- However, S/N does not become infinite as $C \to 0$
  (then front-end operates in current mode)
- The result that $S/N \propto 1/C$ generally applies to systems that measure signal charge.

Noise vs. Detector Capacitance – Charge-Sensitive Amplifier

In a voltage-sensitive preamplifier

- noise voltage at the output is essentially independent of detector capacitance,
- input signal decreases with increasing input capacitance, so signal-to-noise ratio depends on detector capacitance.

In a charge-sensitive preamplifier, the signal at the amplifier output is independent of detector capacitance (if $C_i \gg C_d$).

What is the noise behavior?

- Noise appearing at the output of the preamplifier is fed back to the input, decreasing the output noise from the open-loop value $v_{no} = v_{ni} A_v$.
- The magnitude of the feedback depends on the shunt impedance at the input, i.e. the detector capacitance.

Although specified as an equivalent input noise, the dominant noise sources are typically internal to the amplifier.

Only in a fed-back configuration is some of this noise actually present at the input. In other words, the primary noise signal is not a physical charge (or voltage) at the amplifier input to which the loop responds in the same manner as to a detector signal.

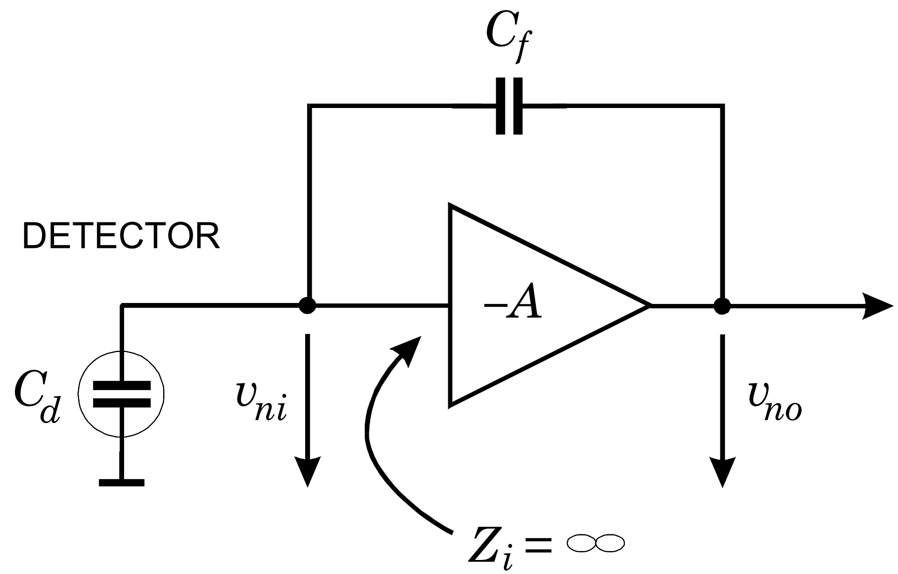
$\Rightarrow$ **S/N at the amplifier output depends on feedback.**

Noise in charge-sensitive preamplifiers

Start with an output noise voltage v_{no}, which is fed back to the input through the capacitive voltage divider $C_f - C_d$.

$$v_{no} = v_{ni} \frac{X_{C_f} + X_{C_d}}{X_{C_d}} = v_{ni} \frac{\dfrac{1}{\omega C_f} + \dfrac{1}{\omega C_d}}{\dfrac{1}{\omega C_d}}$$

$$v_{no} = v_{ni} \left(1 + \frac{C_d}{C_f} \right)$$



Equivalent input noise charge

$$Q_{ni} = \frac{v_{no}}{A_Q} = v_{no} C_f$$

$$Q_{ni} = v_{ni} \left(C_d + C_f \right)$$

Signal-to-noise ratio $$\frac{Q_s}{Q_{ni}} = \frac{Q_s}{v_{ni}(C_d + C_f)} = \frac{1}{C} \frac{Q_s}{v_{ni}}$$

Same result as for voltage amplifier, but here

- *the signal is constant and*
- *the noise grows with increasing C.*

As shown previously, the pulse rise time at the amplifier output also increases with total capacitive input load C, because of reduced feedback.

In contrast, the rise time of a voltage sensitive amplifier is not affected by the input capacitance, although the equivalent noise charge increases with C just as for the charge-sensitive amplifier.

Conclusion

In general

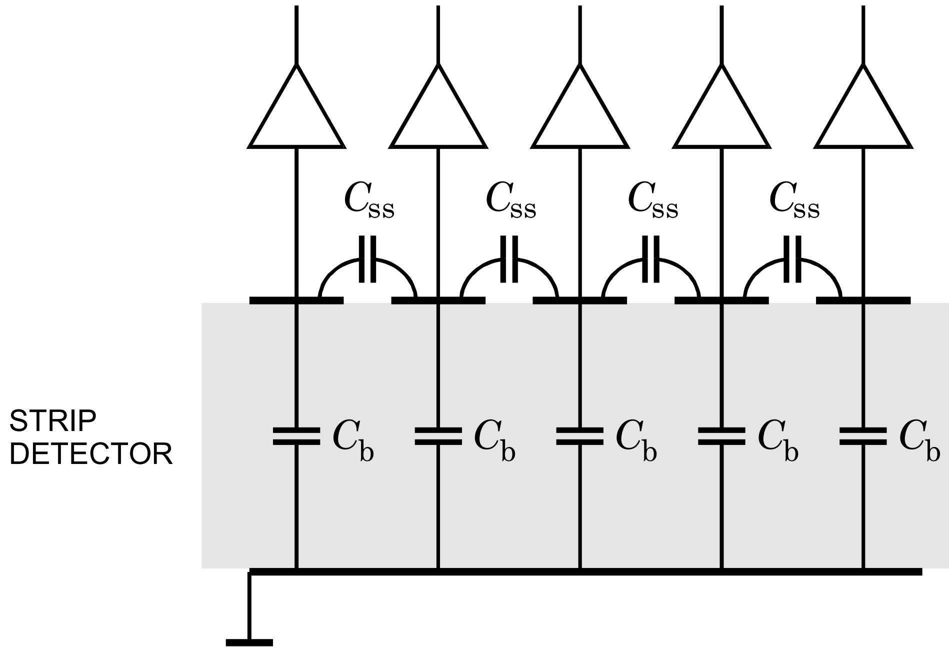
- optimum S/N is independent of whether the voltage, current, or charge signal is sensed.

- S/N cannot be *improved* by feedback.

Practical considerations, i.e. type of detector, amplifier technology, can favor one configuration over the other.

6. Complex Sensors

Cross-coupled noise



Noise at the input of an amplifier is cross-coupled to its neighbors.

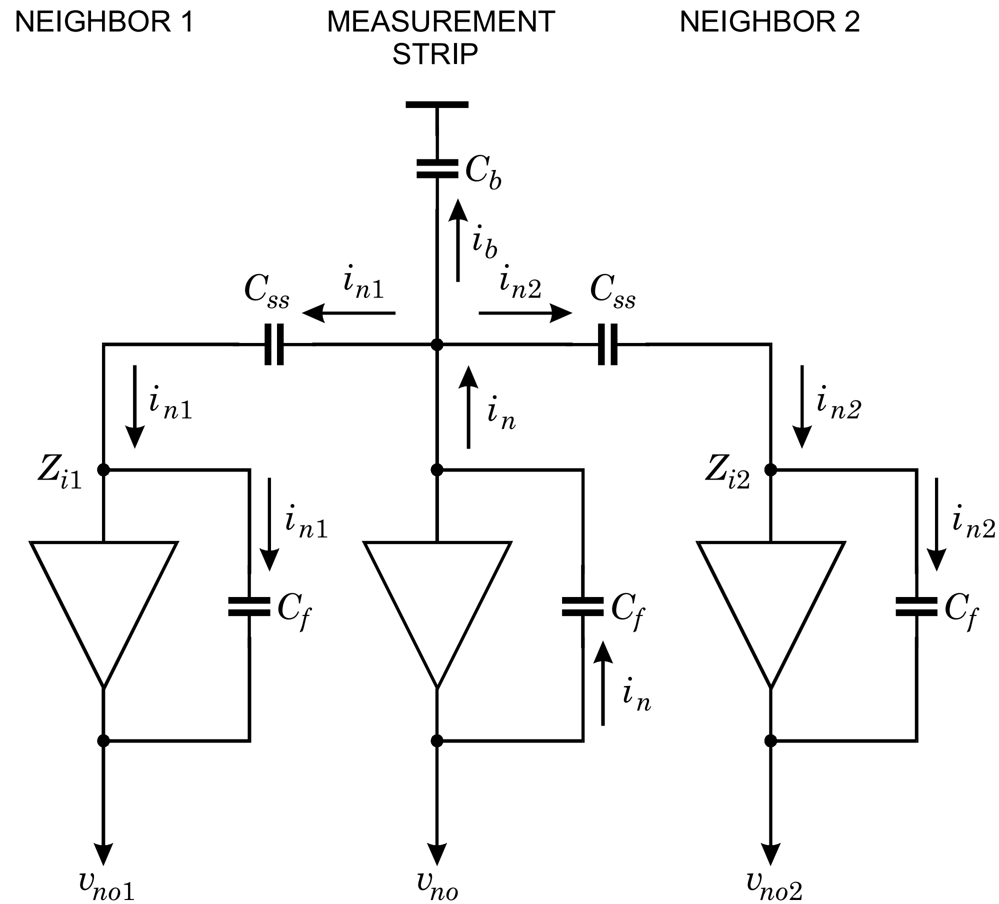
Noise Cross-Coupling Function in Strip and Pixel Detectors

The center amplifier's output noise voltage v_{no} causes a current noise i_n to flow through its feedback capacitance C_f and the inter-electrode capacitances into the neighboring amplifiers, adding to the other amplifiers' noise.

The backplane capacitance C_b attenuates the signal transferred through the strip-to-strip capacitance C_{ss}.

The additional noise introduced into the neighbor channels

$$v_{no1} = v_{no2} \approx \frac{v_{no}}{2} \frac{1}{1+2C_b/C_{ss}}$$



For a backplane capacitance $C_b = C_{ss}/10$ the amplifier's noise with contributions from both neighbors increases by 16%.

In pixel detectors additional paths must be included.
This requires realistic data on pixel-pixel capacitances (often needs tests).

IV. Signal Processing

1. Requirements – Two conflicting objectives:

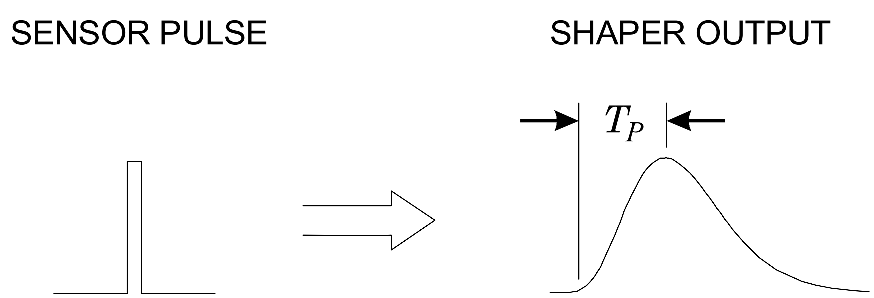
1. Improve Signal-to-Noise Ratio S/N

Restrict bandwidth to match measurement time $\Rightarrow$ Increase pulse width

Typically, the pulse shaper transforms a narrow detector current pulse to

a broader pulse (to reduce electronic noise),

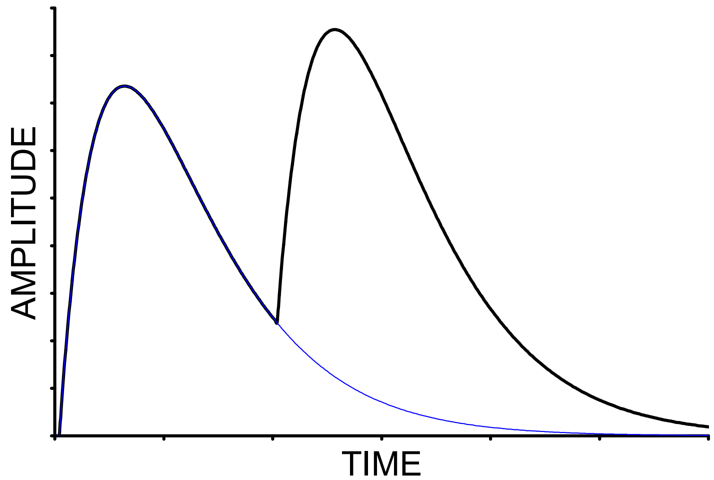
with a gradually rounded maximum at the peaking time T_P
(to facilitate measurement of the peak amplitude)



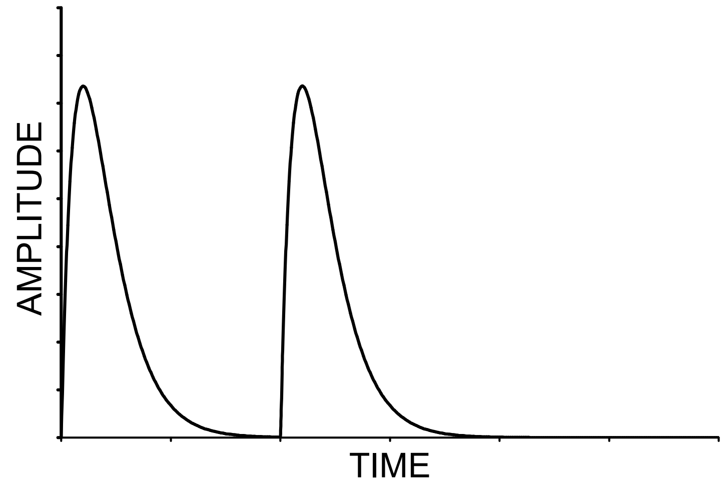
If the shape of the pulse does not change with signal level, the peak amplitude is also a measure of the energy, so one often speaks of pulse-height measurements or pulse height analysis. The pulse height spectrum is the energy spectrum.

2. Improve Pulse Pair Resolution ⇒ Decrease pulse width

Pulse pile-up distorts amplitude measurement.



Reducing pulse shaping time to 1/3 eliminates pile-up.



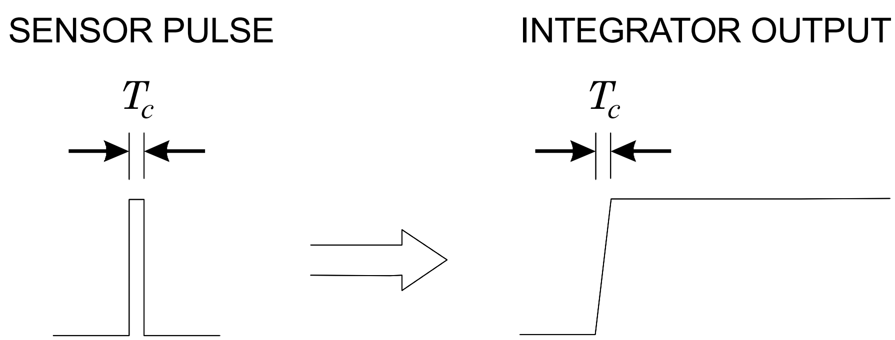
Necessary to find balance between these conflicting requirements. Sometimes minimum noise is crucial, sometimes rate capability is paramount.

Usually, many considerations combined lead to a "non-textbook" compromise.

- *"Optimum shaping" depends on the application!*
- Shapers need not be complicated – *Every amplifier is a pulse shaper!*

Goal: Improve energy resolution

Procedure: Integrate detector signal current $\Rightarrow$ Step impulse



Commonly approximated as "step" response (zero rise time).

Long "flat top" allows measurements at times well beyond the collection time T_C.

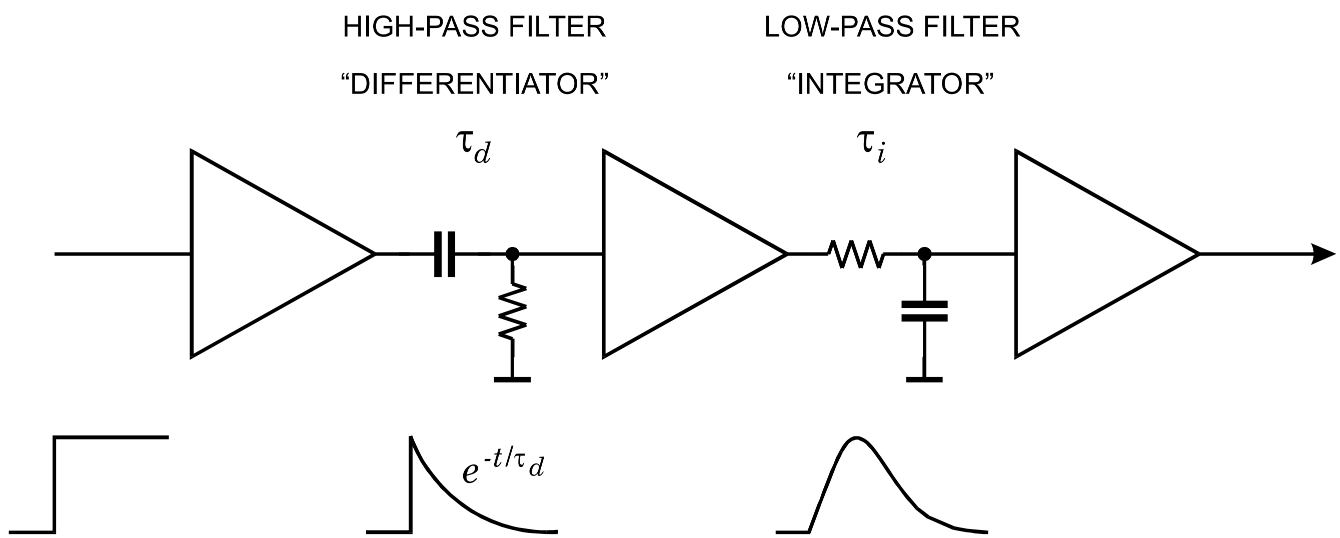
$\Rightarrow$ Allows reduced bandwidth and great flexibility in selecting shaper response.

Optimum for energy measurements, but not for fast timing!

"Fast-slow" systems utilize parallel processing chains to optimize both timing and energy resolution (see Timing Measurements in other tutorials).

2. Pulse Shapers

Simple Example: CR-RC Shaping



Simple arrangement: Noise performance only 36% worse than optimum filter with same time constants.

$\Rightarrow$ Useful for estimates, since simple to evaluate

Key elements:

- lower frequency bound ($\triangleq$ pulse duration)
- upper frequency bound ($\triangleq$ rise time)

are common to all shapers.

Pulse Shaping and Signal-to-Noise Ratio

Pulse shaping affects both the

- total noise

and

- peak signal amplitude

at the output of the shaper.

Equivalent Noise Charge

Inject known signal charge into preamp input
(either via test input or known energy in detector).

Determine signal-to-noise ratio at shaper output.

Equivalent Noise Charge $\equiv$ Input charge for which $S/N = 1$

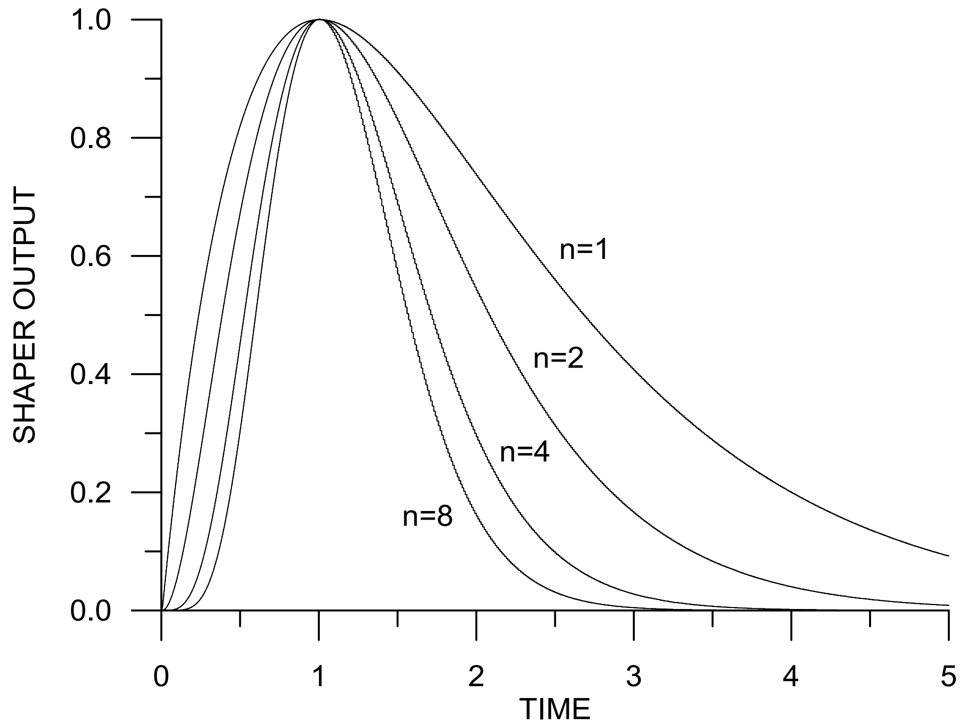
Shapers with Multiple Integrators

Start with simple CR-RC shaper and add additional integrators (n= 1 to n= 2, ... n= 8).

Change integrator time constants to preserve the peaking time $\tau_n = \tau_{n=1} / n$

Increasing the number of integrators makes the output pulse more symmetrical with a faster return to baseline.

$\Rightarrow$ improved rate capability at the same peaking time



Multiple integrators often do not require additional circuitry. Several gain stages are typically necessary to bring the signal to the level required for a threshold discriminator or analog-to-digital converter. Their bandwidth can be set to provide the desired pulse shaping.

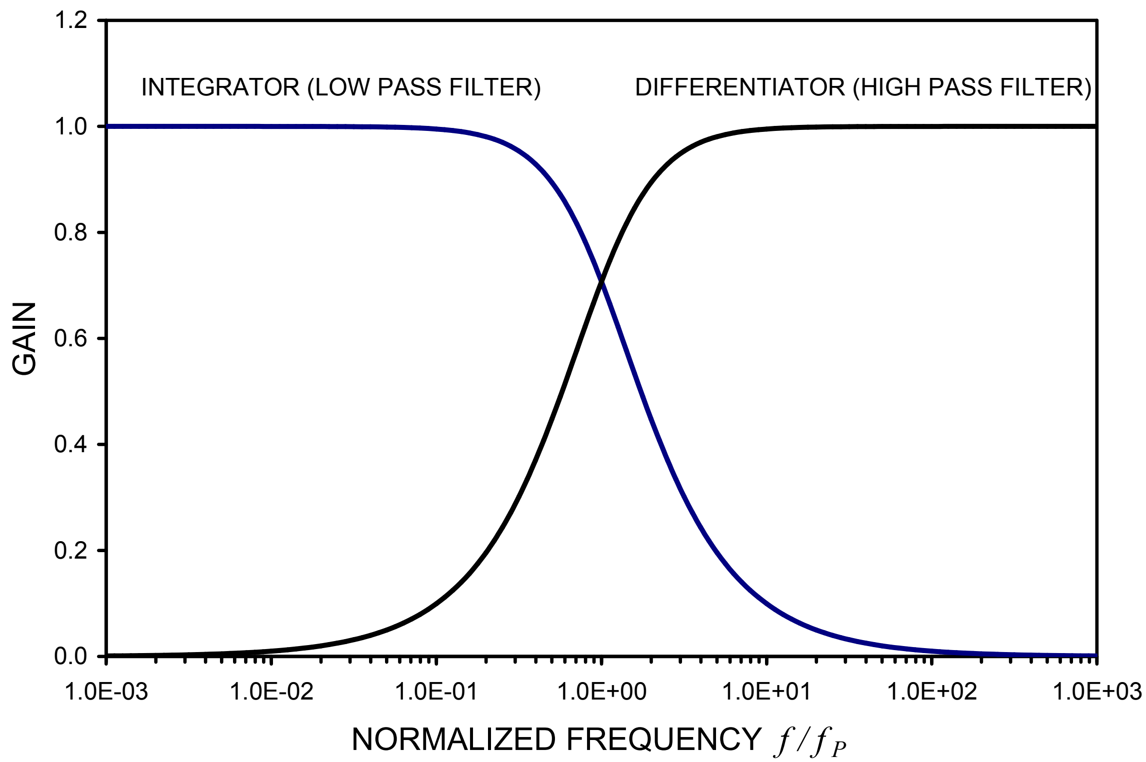
In γ-spectroscopy systems shapers with the equivalent of 8 RC integrators are common. Usually, this is achieved with active filters.

3. Noise Charge vs. Shaping Time

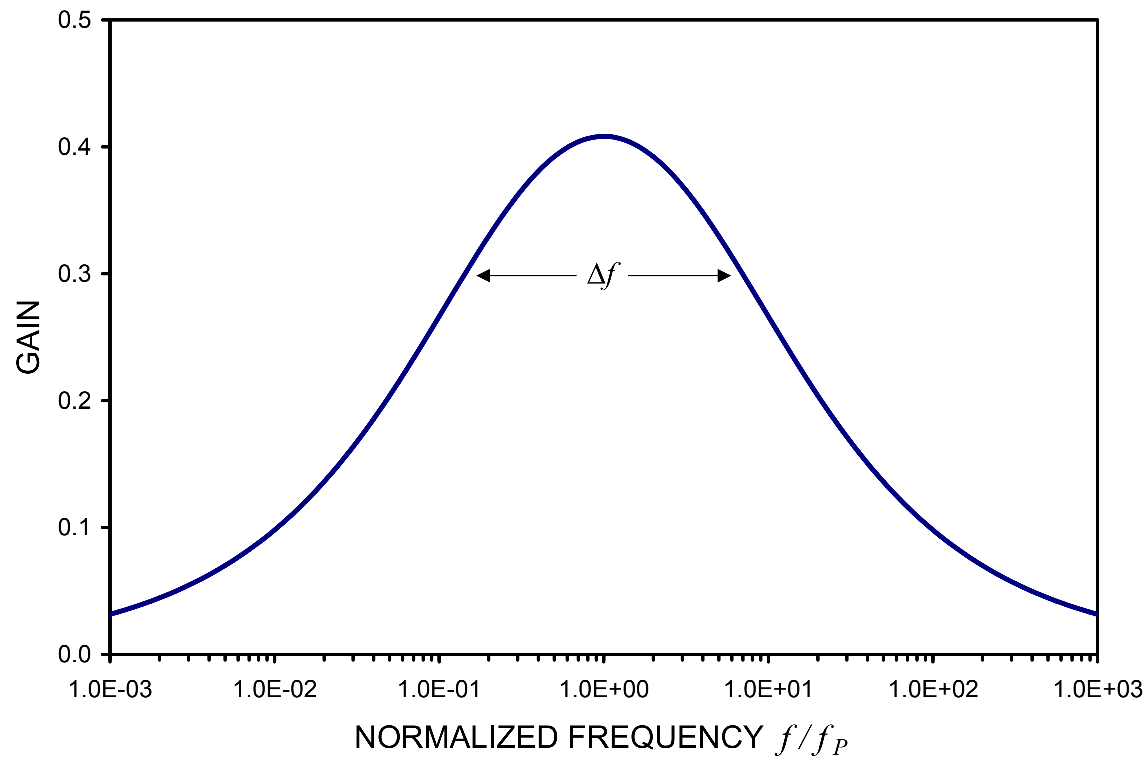
Assume that differentiator and integrator time constants are equal $\tau_i = \tau_d \equiv \tau$.

$\Rightarrow$ Both cutoff frequencies equal: $f_U = f_L \equiv f_P = 1/2\pi\tau$.

Frequency response of individual pulse shaping stages



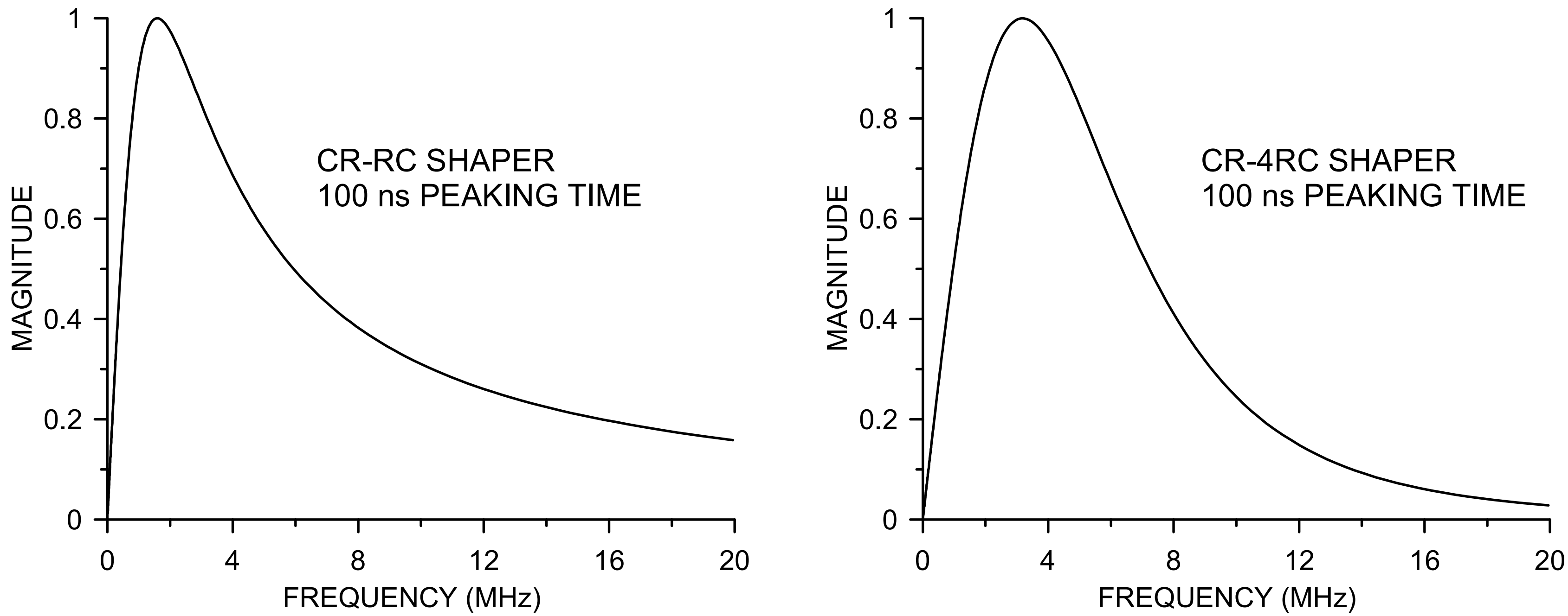
Combined frequency response



Logarithmic frequency scale $\Rightarrow$ shape of response independent of τ .

Bandwidth Δf decreases with increasing time constant τ.

Comparison with CR-$4RC$ shaper



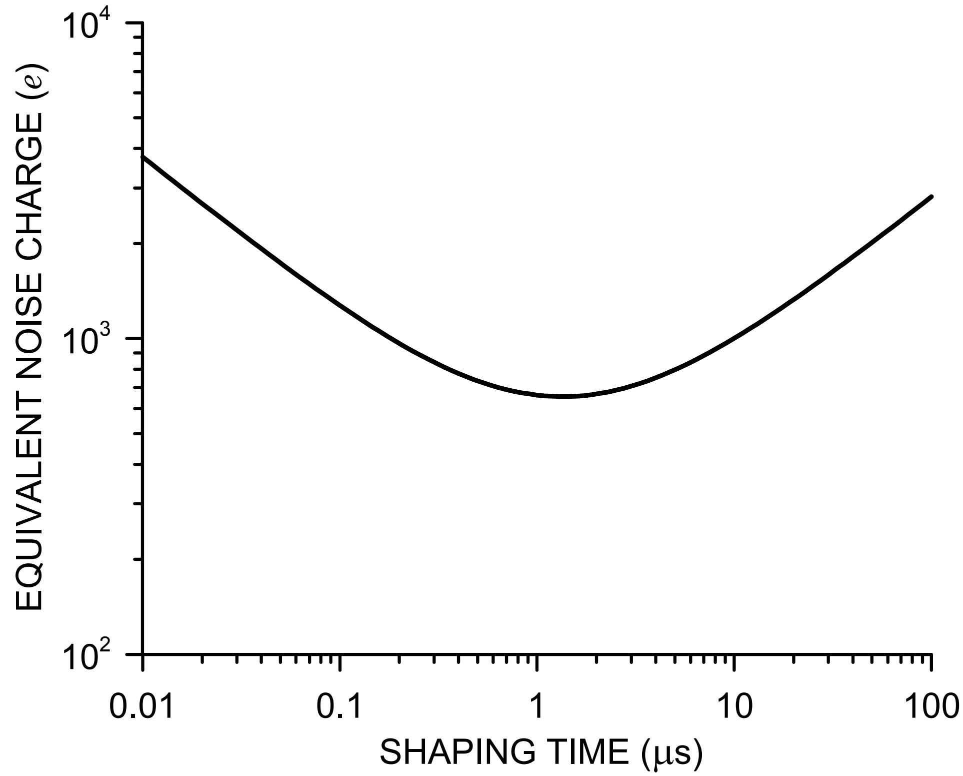
Both have a 100 ns peaking time.

The peaking frequencies are 1.6MHz for the CR-RC shaper and 3.2 MHz for the CR-4RC.

The bandwidth, i.e. the difference between the upper and lower half-power frequencies is 3.2 MHz for the CR-RC shaper and 4.3 MHz for the CR-4RC shaper.

The peaking frequency and bandwidth scale with the inverse peaking time.

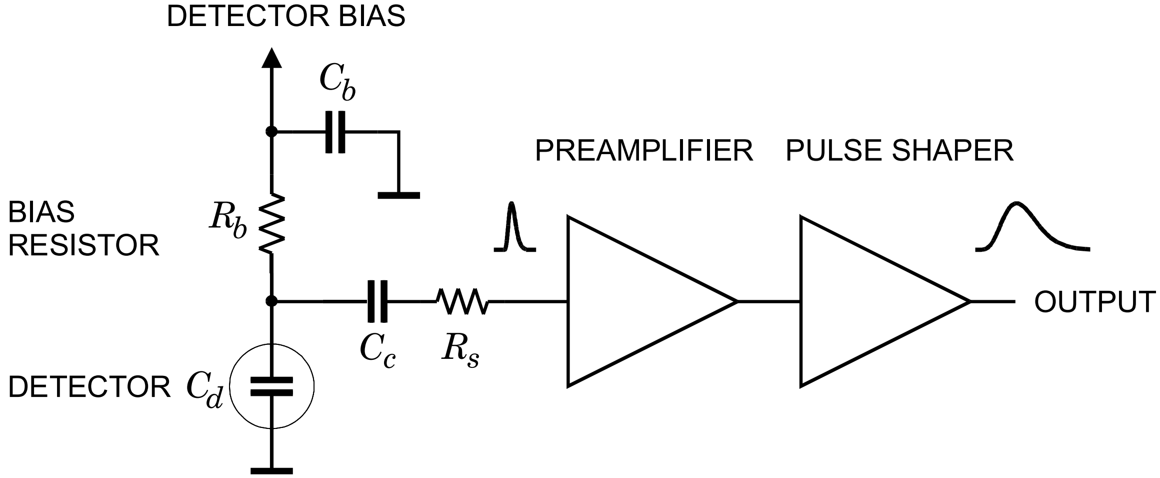
Result of typical noise measurement vs. shaping time



Noise sources (thermal and shot noise) have a flat ("white") frequency distribution.

Why doesn't the noise decrease monotonically with increasing shaping time (decreasing bandwidth)?

Analytical Analysis of a Detector Front-End



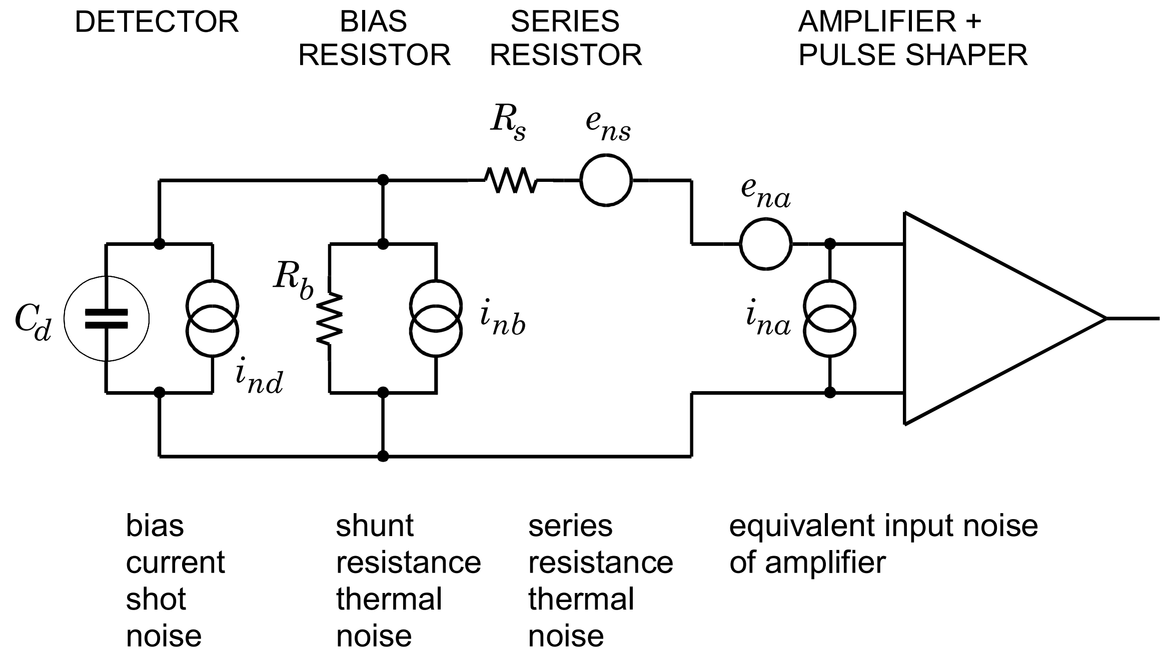
Detector bias voltage is applied through the resistor R_B. The bypass capacitor C_B serves to shunt any external interference coming through the bias supply line to ground. For AC signals this capacitor connects the "far end" of the bias resistor to ground, so that R_B appears to be in parallel with the detector.

The coupling capacitor C_C in the amplifier input path blocks the detector bias voltage from the amplifier input (which is why this capacitor is also called a "blocking capacitor").

The series resistor R_S represents any resistance present in the connection from the detector to the amplifier input. This includes

- the resistance of the detector electrodes
- the resistance of the connecting wires
- any resistors used to protect the amplifier against large voltage transients

Equivalent circuit for noise analysis



In this example a voltage-sensitive amplifier is used, so all noise contributions will be calculated in terms of the noise voltage appearing at the amplifier input.

Resistors can be modeled either as voltage or current generators.

- Resistors in parallel with the input act as current sources.
- Resistors in series with the input act as voltage sources.

Steps in the analysis:

1. Determine the frequency distribution of the noise voltage presented to the amplifier input from all individual noise sources

2. Integrate over the frequency response of a CR-RC shaper to determine the total noise output.

3. Determine the output signal for a known signal charge and calculate equivalent noise charge (signal charge for $S/N = 1$)

First, assume a simple CR-RC shaper with

equal differentiation and integration time constants $\tau_i = \tau_d = \tau$,

which in this special case is equal to the peaking time.

Equivalent Noise Charge

$$Q_n^2 = \left(\frac{\mathbf{e}^2}{8}\right) \left[\left(2q_e I_D + \frac{4kT}{R_P} + i_{na}^2 \right) \cdot \tau + \left(4kTR_S + e_{na}^2 \right) \cdot \frac{C_d^2}{\tau} + 4A_f C_d^2 \right]$$

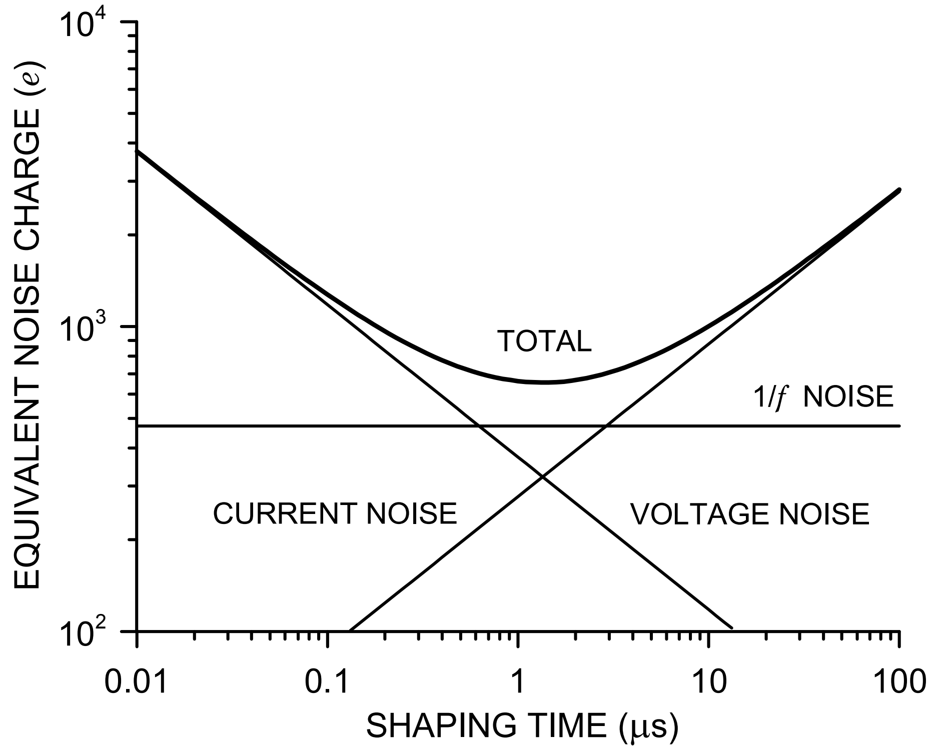
$\mathbf{e} = \exp(1)$	current noise	voltage noise	$1/f$ noise
	$\propto \tau$	$\propto 1/\tau$	independent of τ
	independent of C_d	$\propto C_d^2$	$\propto C_d^2$

- Current noise is independent of detector capacitance, consistent with the notion of "counting electrons", which is also why it increases with time constant τ.
- Voltage noise increases with detector capacitance (reduced signal voltage) and decreases with time constant τ.
- $1/f$ noise is independent of shaping time.

  In general, the total noise of a $1/f$ source depends on the ratio of the upper to lower cutoff frequencies, not on the absolute noise bandwidth. If τ_d and τ_i are scaled by the same factor, this ratio remains constant.
- Detector leakage current and FET noise decrease with temperature

  $\Rightarrow$ High resolution Si and Ge detectors for x-rays and gamma rays operate at cryogenic temperatures.

The equivalent noise charge Q_n assumes a minimum when the current and voltage noise contributions are equal. Typical result:



dominated by voltage noise current noise

For a given pulse width CR-RC shaper, the noise minimum obtains for $\tau_d = \tau_i = \tau$.

This criterion does not hold for more sophisticated shapers.

Time-Variant Shapers

Time variant shaper change the filter parameters during the processing of individual pulses.

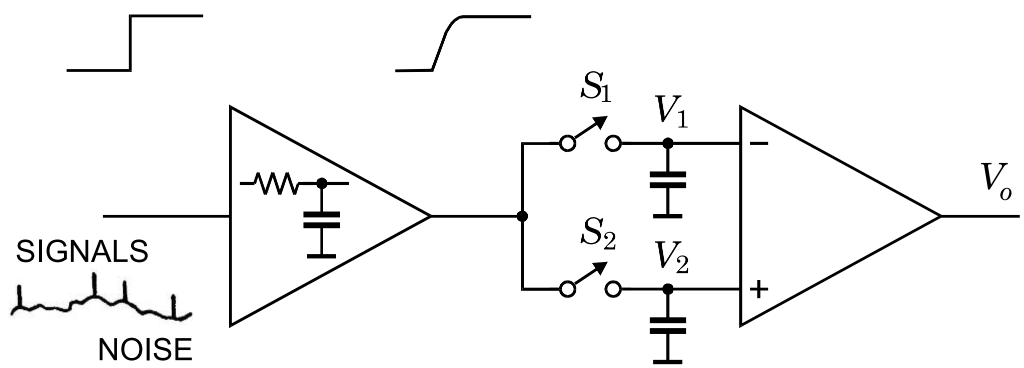
A commonly used time-variant filter is the correlated double-sampler.

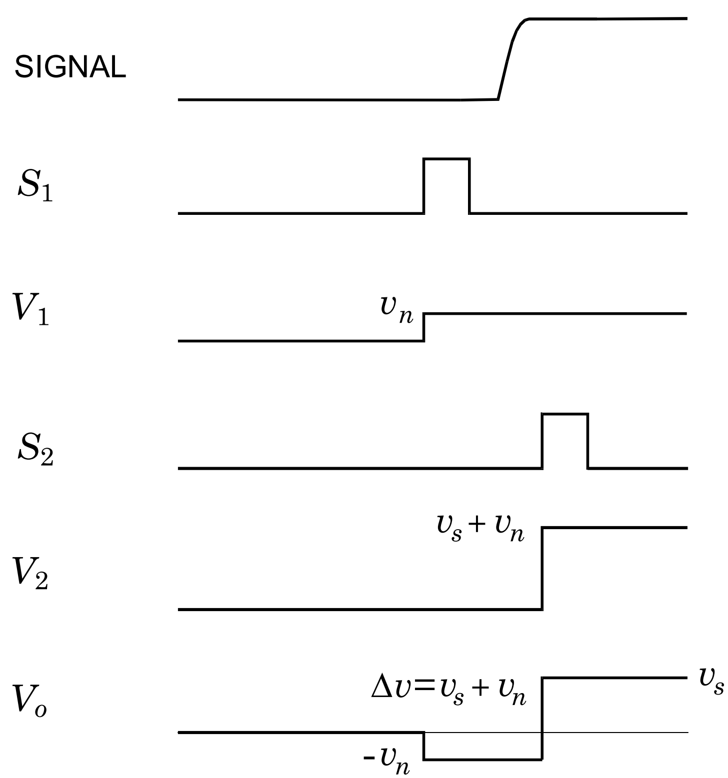
1. Signals are superimposed on a (slowly) fluctuating baseline
2. To remove baseline fluctuations the baseline is sampled prior to the arrival of a signal.
3. Next, the signal + baseline is sampled and the previous baseline sample subtracted to obtain the signal

S/N depends on

1. time constant of prefilter
2. time difference between samples

See "Semiconductor Detector Systems" for a detailed noise analysis. (Chapter 4, pp 160-166)





Numerical expression for the noise of a CR-RC shaper
(amplifier current noise negligible)

(note that some units are "hidden" in the numerical factors)

$$Q_n^2 = 12\, I_B\, \tau \; + \; 6 \cdot 10^5 \frac{\tau}{R_P} \; + \; 3.6 \cdot 10^4\, e_n^2 \frac{C^2}{\tau} \quad \text{[rms electrons}^2\text{]}$$

where

- τ shaping time constant [ns]
- I_B detector bias current + amplifier input current [nA]
- R_P input shunt resistance [kΩ]
- e_n equivalent input noise voltage spectral density [nV/√Hz]
- C total input capacitance [pF]

$Q_n = 1\, e$ corresponds to 3.6 eV in Si
2.9 eV in Ge

“Series” and “Parallel” Noise

For sources connected in parallel, currents are additive.

For sources connected in series, voltages are additive.

⇒ In the detector community voltage and current noise are often called “series” and “parallel” noise.

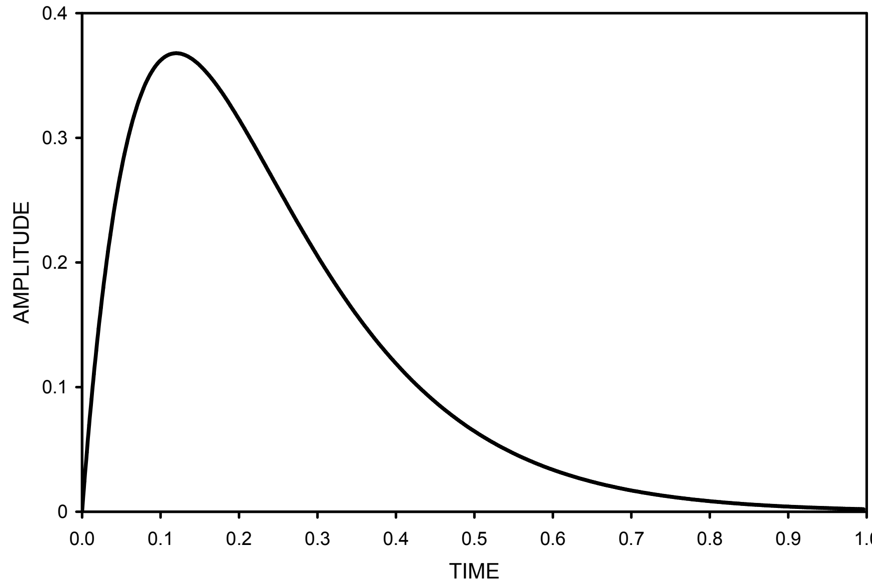
The rest of the world uses equivalent noise voltage and current.

Since they are physically meaningful, use of these widely understood terms is preferable.

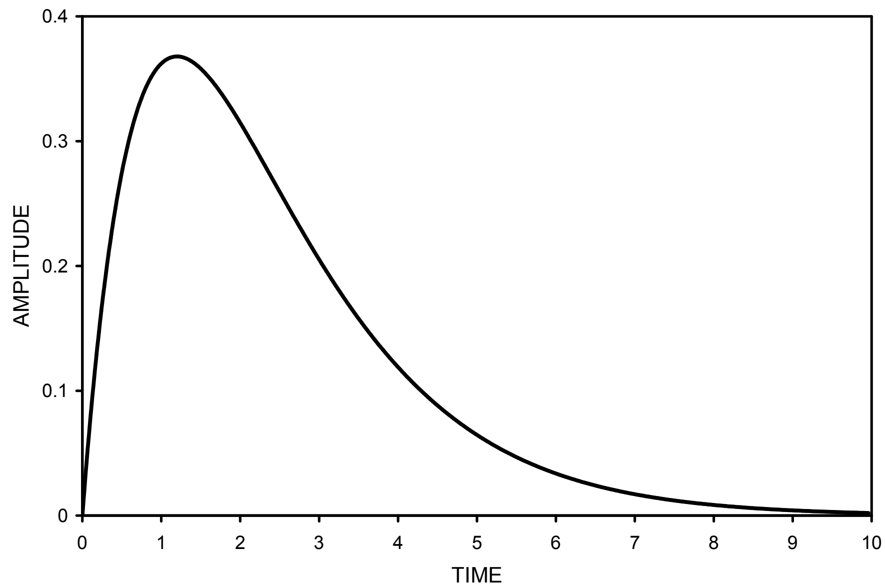
Scaling of Filter Noise Parameters

Pulse shape is the same when shaping time is changed.

shaping time = τ



shaping time = 10τ



Shaper can be characterized by a "shape factor" which multiplied by the shaping time sets the noise bandwidth.

The expression for the equivalent noise charge

$$Q_n^2 = \left(\frac{\mathbf{e}^2}{8}\right)\left[\left(2q_e I_D + \frac{4kT}{R_P} + i_{na}^2\right)\cdot\tau + \left(4kTR_S + e_{na}^2\right)\cdot\frac{C_d^2}{\tau} + 4A_f C_d^2\right]$$

$\mathbf{e} = \exp(1)$

	current noise	voltage noise	$1/f$ noise
	$\propto \tau$	$\propto 1/\tau$	independent of τ
	independent of C_d	$\propto C_d^{\,2}$	$\propto C_d^{\,2}$

can be put in a more general form that applies to all type of pulse shapers:

$$Q_n^2 = i_n^2\, T_s\, F_i + C^2\, e_n^2\, \frac{F_v}{T_s} + F_{vf}\, A_f\, C^2$$

- The current and voltage terms are combined and represented by i_n^2 and e_n^2.
- The shaper is characterized by a shape and characteristic time (e.g. the peaking time).
- A specific shaper is described by the “shape factors” F_i, F_v, and F_{vf}.
- The effect of the shaping time is set by T_S.
- The effective capacitance C exceeds the detector capacitance C_d, as it must include the preamplifier input capacitance and any other capacitances shunting the input.

Detector Noise Summary

Two basic noise mechanisms: input noise current i_n
input noise voltage e_n

Equivalent Noise Charge: $$Q_n^2 = i_n^2 \, T_s \, F_i + C^2 \, e_n^2 \, \frac{F_v}{T_s}$$

T_S Characteristic shaping time (*e.g.* peaking time)

F_i , F_v "Shape Factors" that are determined by the shape of the pulse.

C Total capacitance at the input (detector capacitance + input capacitance of preamplifier + stray capacitance + …)

Note that $F_i < F_v$ for higher order shapers.

Typical values of F_i , F_v

CR-RC shaper	F_i = 0.924	F_v = 0.924
CR-(RC)^4 shaper	F_i = 0.45	F_v = 1.02
CR-(RC)^7 shaper	F_i = 0.34	F_v = 1.27
CAFE chip	F_i = 0.4	F_v = 1.2

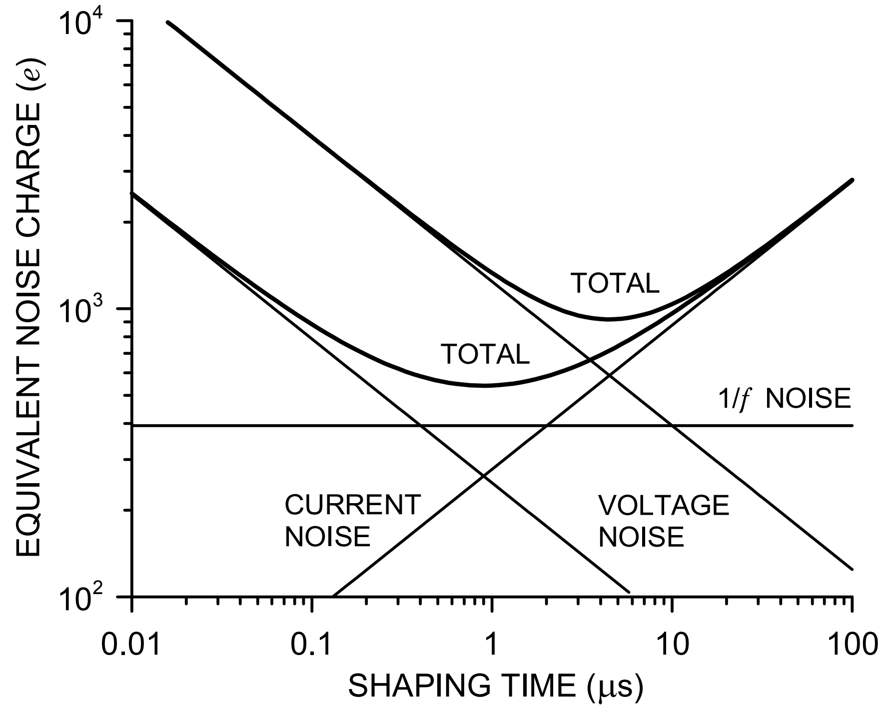
Shapers can be optimized to reduce current noise contribution relative to the voltage noise (mitigate radiation damage!).
Minimum noise obtains when the current and voltage noise contributions are equal.

Current noise

- detector bias current
  increases with detector size, strongly temperature dependent
- noise from resistors shunting the input
  increases as resistance is decreased
- input transistor –
  low for FET, higher for BJTs

Voltage noise

- input transistor –
  noise decreases with increased current



- series resistance, e.g. detector electrode, protection circuits

FETs commonly used as input devices – improved noise performance when cooled (T_{opt} ≈ 130 K)

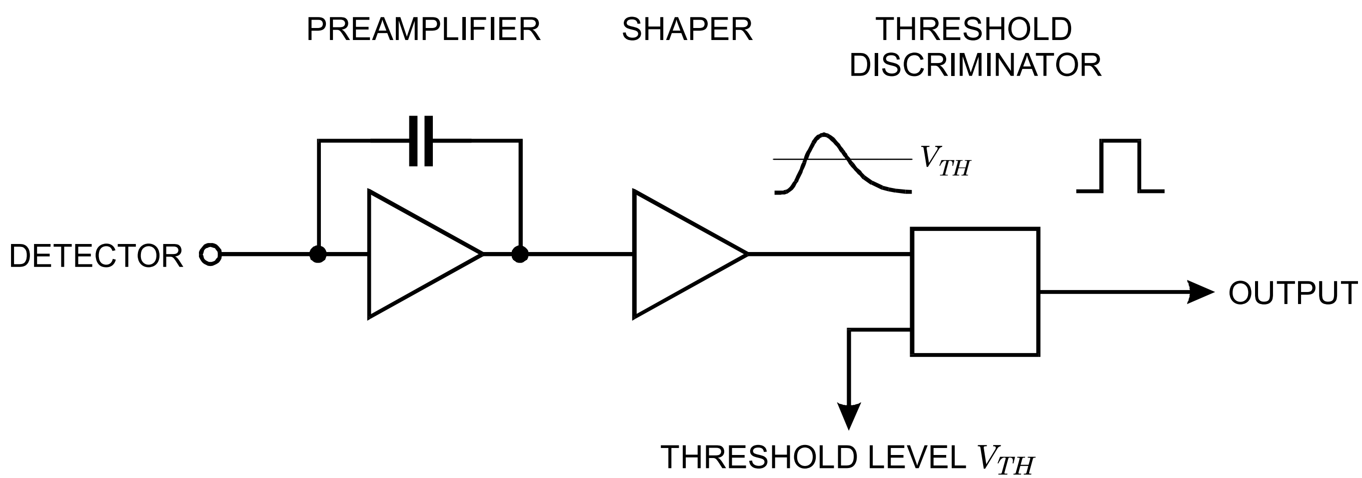
Bipolar transistors advantageous at short shaping times (<100 ns).
When collector current is optimized, bipolar transistor equivalent noise charge is independent of shaping time (see Chapter 6).

4. Threshold Discriminator Systems

The simplest form of a digitized readout is a threshold discriminator system, which produces a normalized (digital) output pulse when the input signal exceeds a certain level.

Noise affects not only the resolution of amplitude measurements, but also the determines the minimum detectable signal threshold.

Consider a system that only records the presence of a signal if it exceeds a fixed threshold.



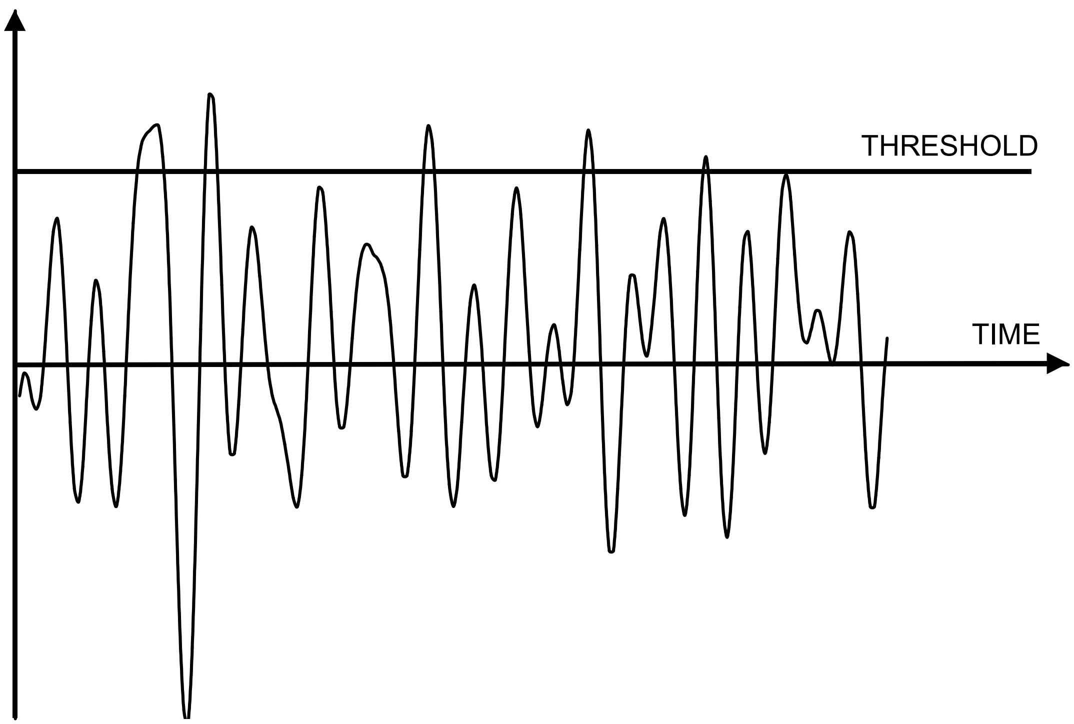
How small a detector pulse can still be detected reliably?

Consider the system at times when no detector signal is present.

Noise will be superimposed on the baseline.

Some noise pulses will exceed the threshold.

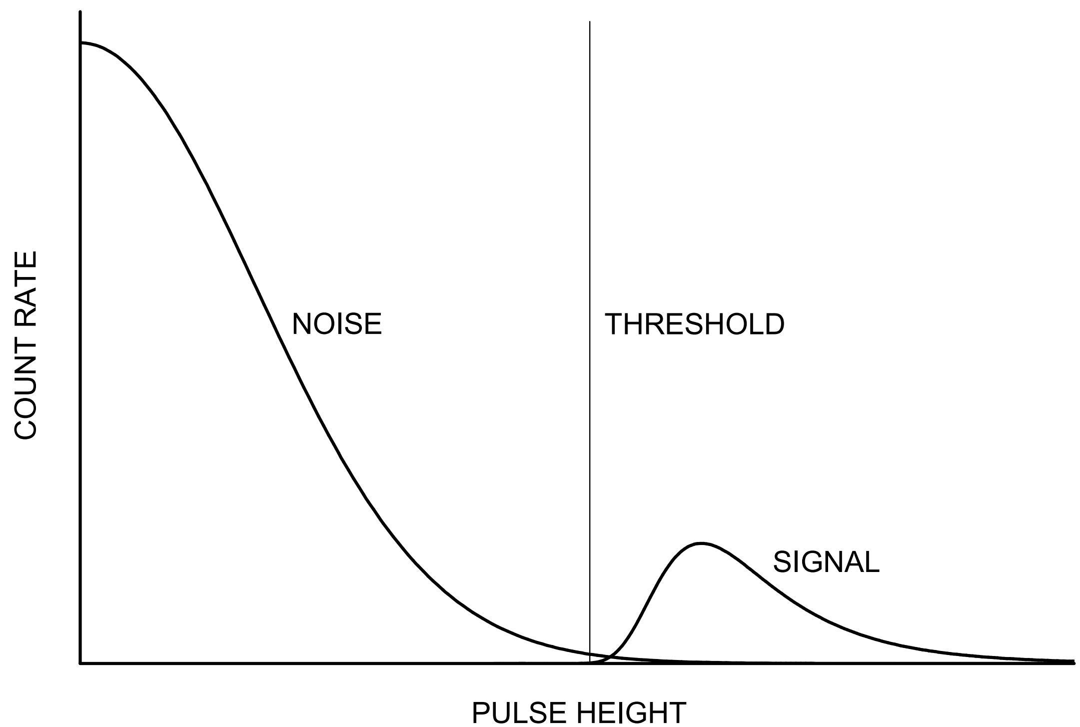
This is always true since the amplitude spectrum of Gaussian noise extends to infinity



The threshold must be set

1. high enough to suppress noise hits

2. low enough to capture the signal



Combined probability function

Both the amplitude and time distribution are Gaussian.

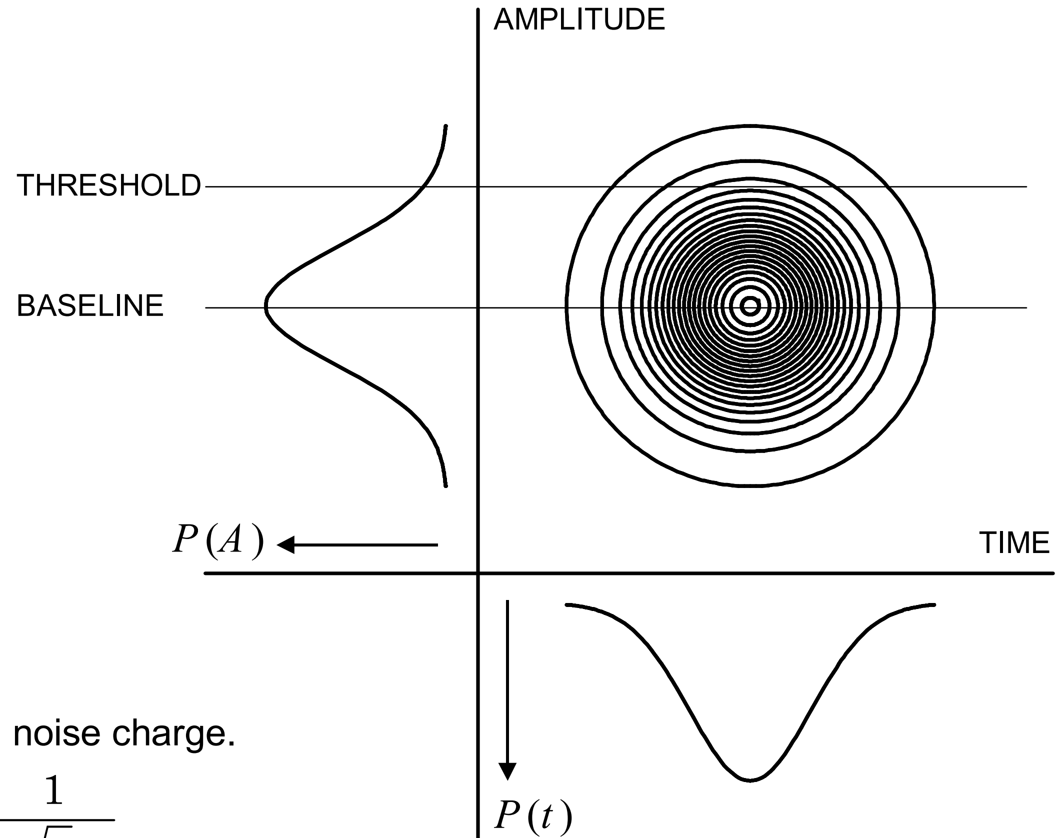
The rate of noise hits is determined by integrating the combined probability density function in the regime that exceeds the threshold.

This yields

$$f_n = f_{n0} \cdot e^{-Q_T^2/2Q_n^2}$$

where Q_T and Q_n are the threshold and noise charge.

The zero-threshold noise rate $f_{n0} \approx \dfrac{1}{2\sqrt{3}\ \tau}$



Thus, the required threshold-to-noise ratio for a given frequency of noise hits f_n is

$$\frac{Q_T}{Q_n} \approx \sqrt{-2\log\left(\frac{f_n}{f_P}\right)},$$

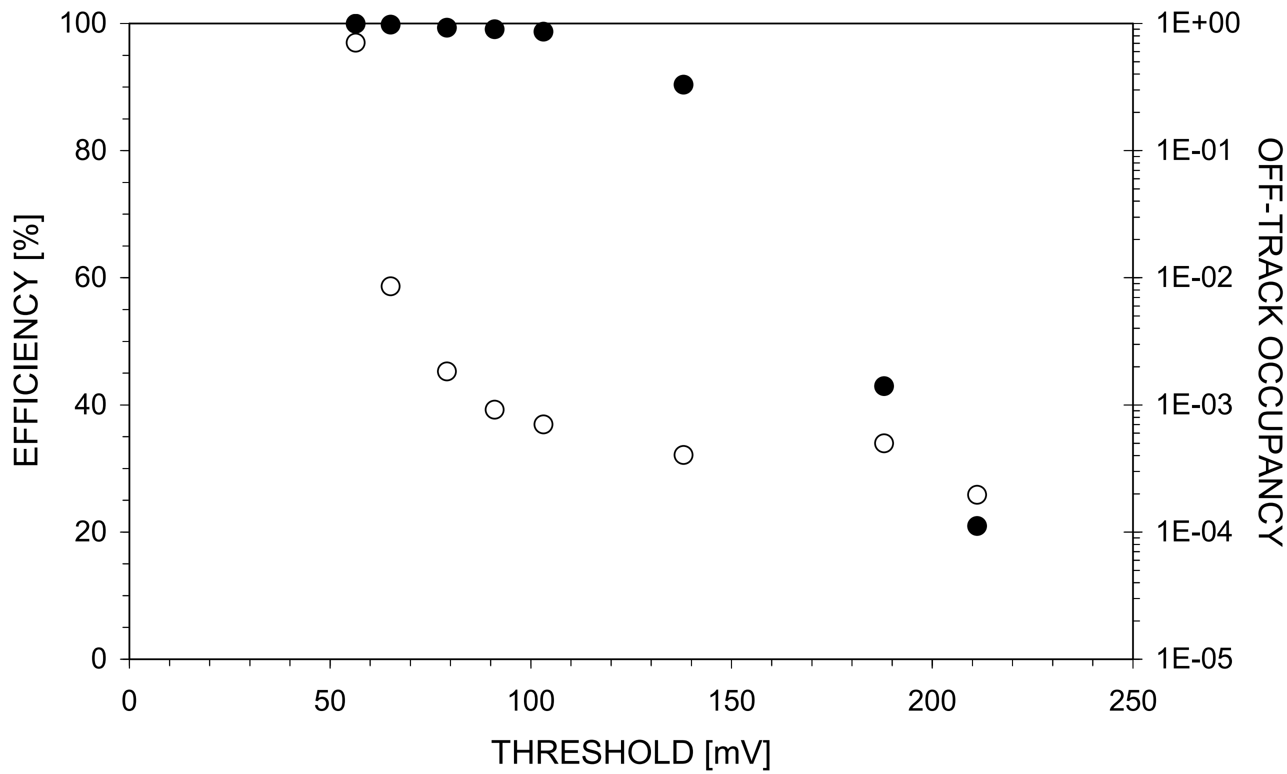
where f_P is the peaking frequency of the shaper

Note that product of noise rate and shaping time $f_n \tau$ determines the required threshold-to-noise ratio, i.e. for a given threshold-to-noise ratio the noise rate is higher at short shaping times

$\Rightarrow$ The noise rate for a given threshold-to-noise ratio is proportional to bandwidth.

$\Rightarrow$ To obtain the same noise rate, a fast system requires a larger threshold-to-noise ratio than a slow system with the same noise level.

Example of noise occupancy (open circles) and efficiency (solid circles) vs. threshold in a practical detector module:



Note that an extended overlap region of high efficiency and low noise occupancy is desired.

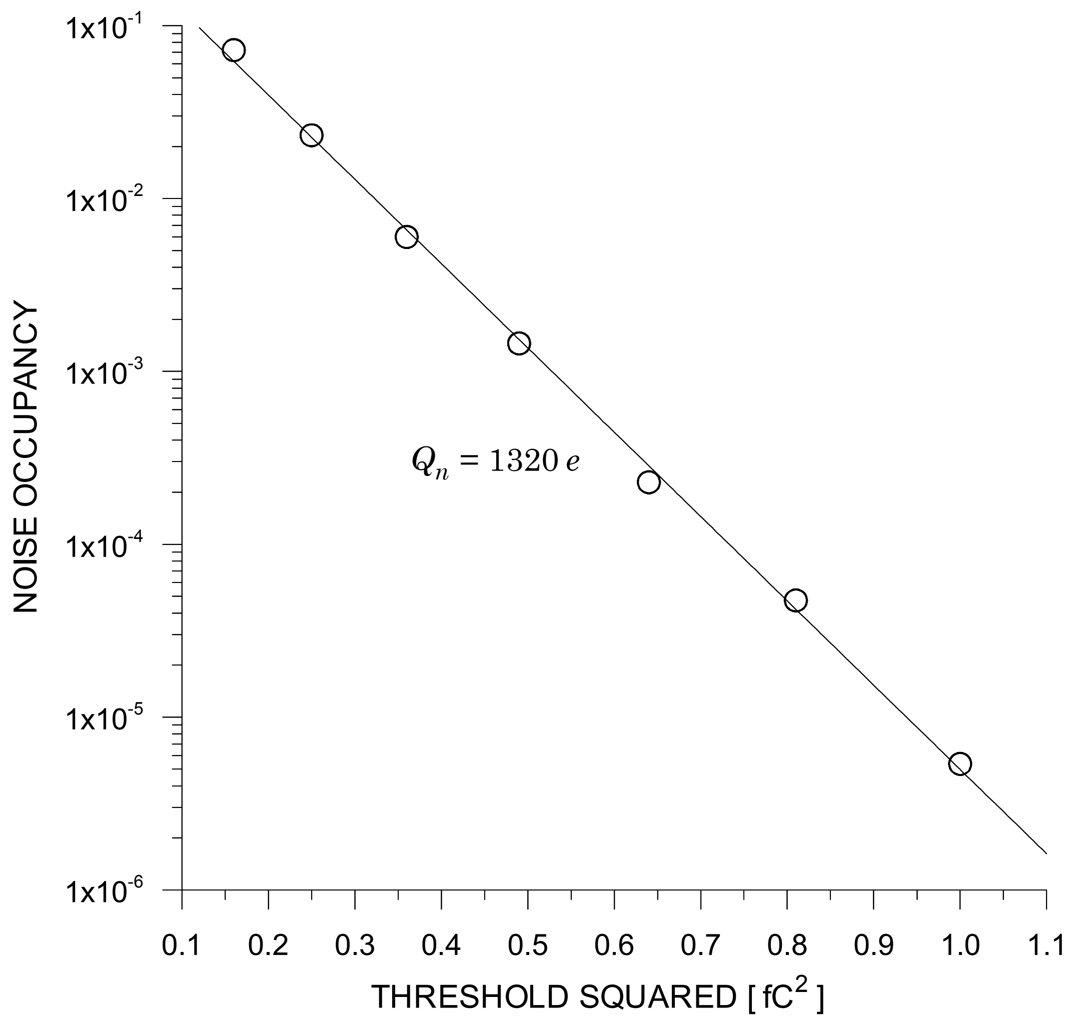
The dependence of occupancy on threshold can be used to measure the noise level.

$$\log P_n = \log\left(\frac{\Delta t}{2\sqrt{3}\ \tau}\right) - \frac{1}{2}\left(\frac{Q_T}{Q_n}\right)^2,$$

so the *slope* of log P_n vs. Q_T^2 yields the noise level.

This analysis is *independent of the details of the shaper*, which affect only the offset.

Measured online during a beam test at KEK in Japan



5. Timing Measurements

Pulse height measurements discussed up to now emphasize accurate measurement of signal charge.

- Timing measurements optimize determination of time of occurrence.
- For timing, the figure of merit is not signal-to-noise, but slope-to-noise ratio.

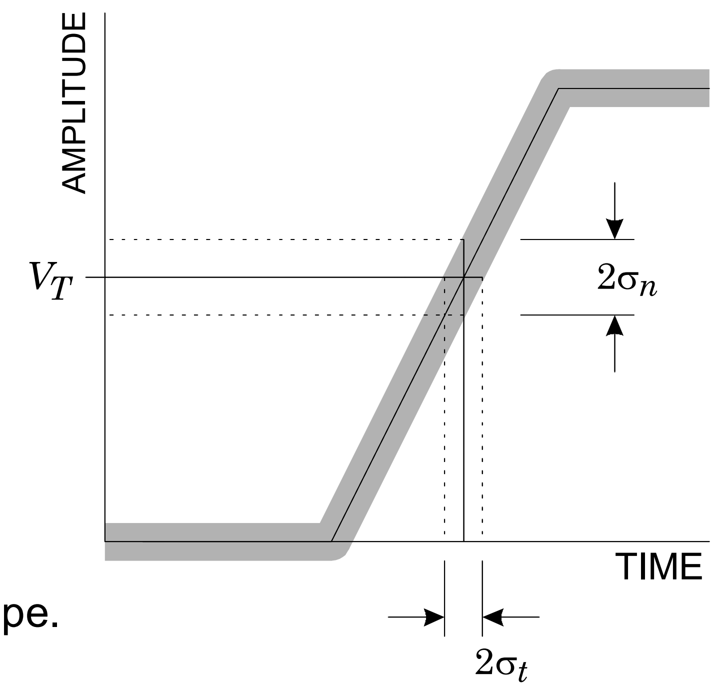
Consider the leading edge of a pulse fed into a threshold discriminator (comparator).

The instantaneous signal level is modulated by noise.

⇒ time of threshold crossing fluctuates

$$\sigma_t = \frac{\sigma_n}{\left.\frac{dV}{dt}\right|_{V_T}} \approx \frac{t_r}{S/N}$$

t_r = rise time



Typically, the leading edge is not linear, so the optimum trigger level is the point of maximum slope.

Pulse Shaping

Consider a system whose bandwidth is determined by a single RC integrator.

The time constant of the RC low-pass filter determines the

- rise time (and hence dV/dt)
- amplifier bandwidth (and hence the noise)

Time dependence: $$V_o(t) = V_0(1 - e^{-t/\tau})$$

The rise time is commonly expressed as the interval between the points of 10% and 90% amplitude

$$t_r = 2.2\ \tau$$

In terms of bandwidth

$$t_r = 2.2\ \tau = \frac{2.2}{2\pi\ f_u} = \frac{0.35}{f_u}$$

Example: An oscilloscope with 100 MHz bandwidth has 3.5 ns rise time.

For a cascade of amplifiers: $$t_r \approx \sqrt{t_{r1}^2 + t_{r2}^2 + \ \ldots\ + t_{rn}^2}$$

Choice of Rise Time in a Timing System

Assume a detector pulse with peak amplitude V_0 and a rise time t_c passing through an amplifier chain with a rise time t_{ra}.

1. Amplifier rise time $\gg$ Signal rise time:

$$\text{Noise} \propto \sqrt{f_u} \propto \sqrt{\frac{1}{t_{ra}}}$$

$$\frac{dV}{dt} \propto \frac{1}{t_{ra}} \propto f_u$$

increase in bandwidth $\Rightarrow$ improvement in dV/dt outweighs increase in noise.

2. Amplifier rise time $\ll$ Signal rise time

increase in noise without increase in dV/dt

Optimum: The amplifier rise time should be chosen to match the signal rise time.

Differentiation time constant: choose greater than rise time constant

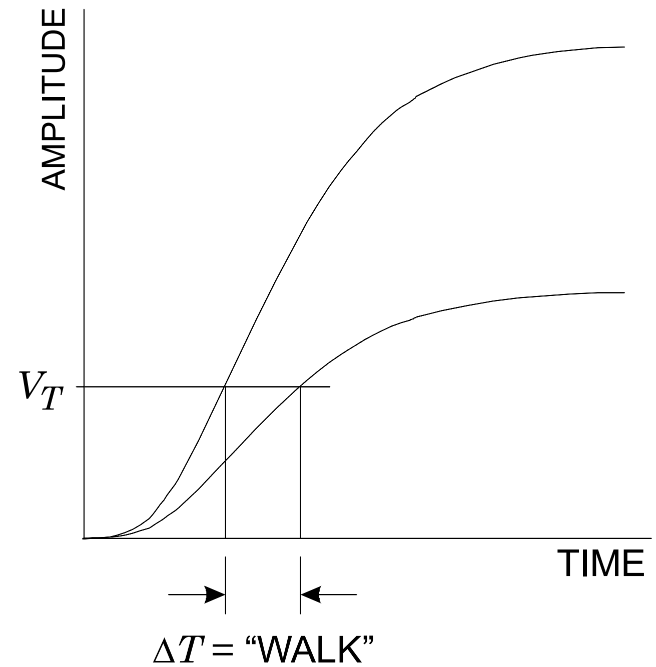
($\tau_{diff} = 10\tau_{\text{int}}$ incurs 20% loss in pulse height)

Time Walk

For a fixed trigger level the time of threshold crossing
depends on pulse amplitude.

⇒ Accuracy of timing measurement limited by

- jitter (due to noise)
- time walk (due to amplitude variations)



If the rise time is known, "time walk" can be compensated in software event-by-event by measuring the pulse height and correcting the time measurement.

This technique fails if both amplitude and rise time vary, as is common.

In hardware, time walk can be reduced by setting the threshold to the lowest practical level, or by using amplitude compensation circuitry, e.g. constant fraction triggering.

Lowest Practical Threshold

Single RC integrator has maximum slope at t= 0: $\dfrac{d}{dt}(1-e^{-t/\tau}) = \dfrac{1}{\tau}\; e^{-t/\tau}$

However, the rise time of practically all fast timing systems is determined by multiple time constants.

For small t the slope at the output of a single RC integrator is linear, so initially the pulse can be approximated by a ramp αt.

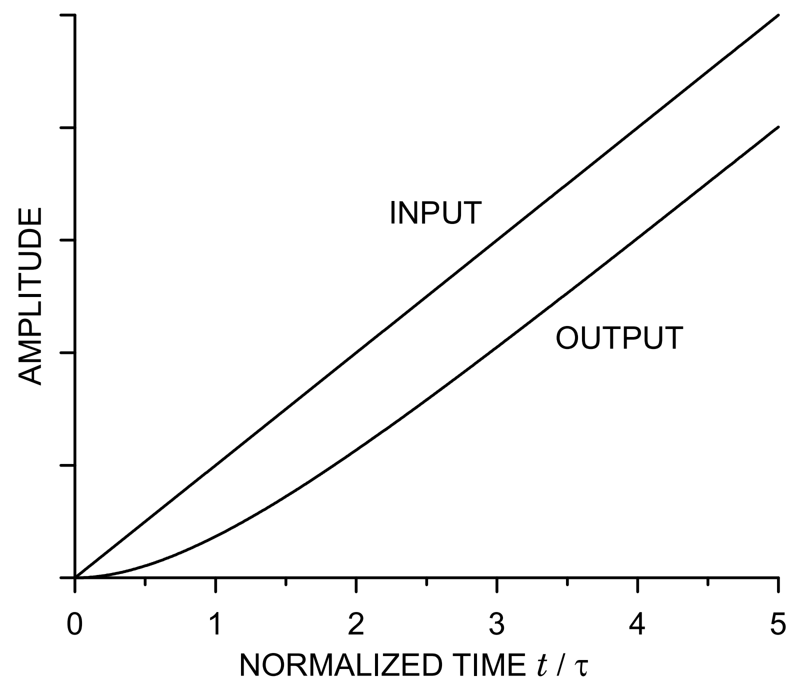
Response of the following integrator

$$V_i = \alpha\; t \quad \rightarrow \quad V_o = \alpha(t-\tau) - \alpha\; \tau\; e^{-t/\tau}$$

$\Rightarrow$ The output is delayed by τ
and curvature is introduced at small t.

Output attains 90% of input slope after $t = 2.3\tau$.

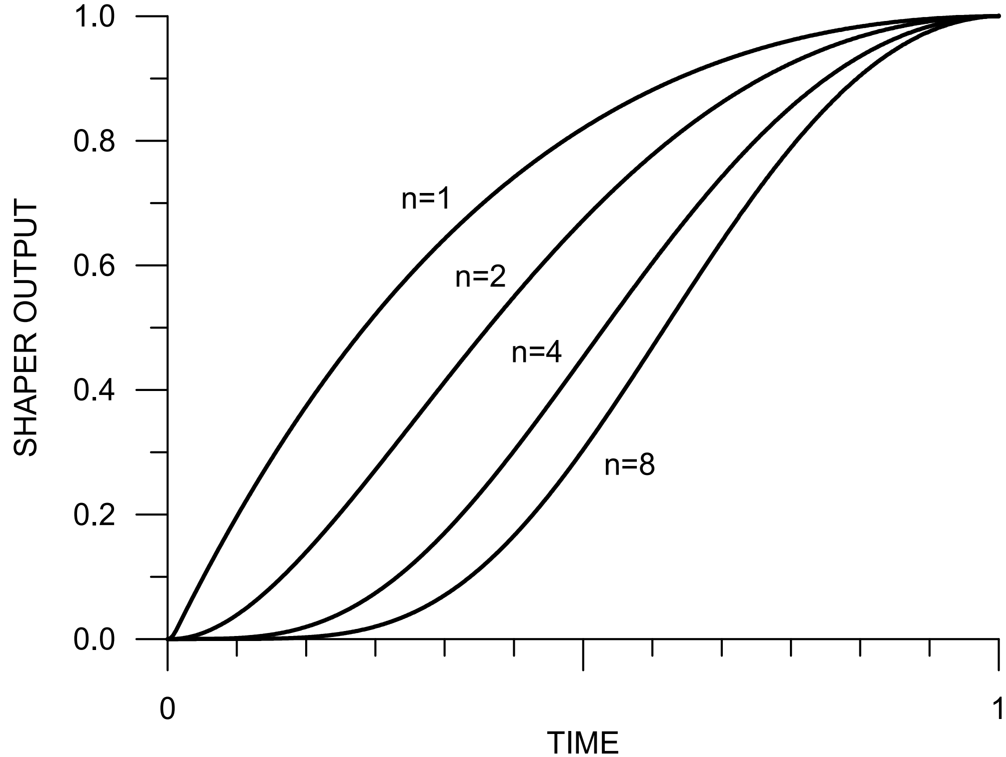
Delay for n integrators = $n\tau$



Additional RC integrators introduce more curvature at the beginning of the pulse.

Output pulse shape for multiple RC integrators

(normalized to preserve the peaking time, $\tau_n = \tau_{n-1} / n$)



Increased curvature at beginning of pulse limits the minimum threshold for good timing.

⇒ One dominant time constant best for timing measurements

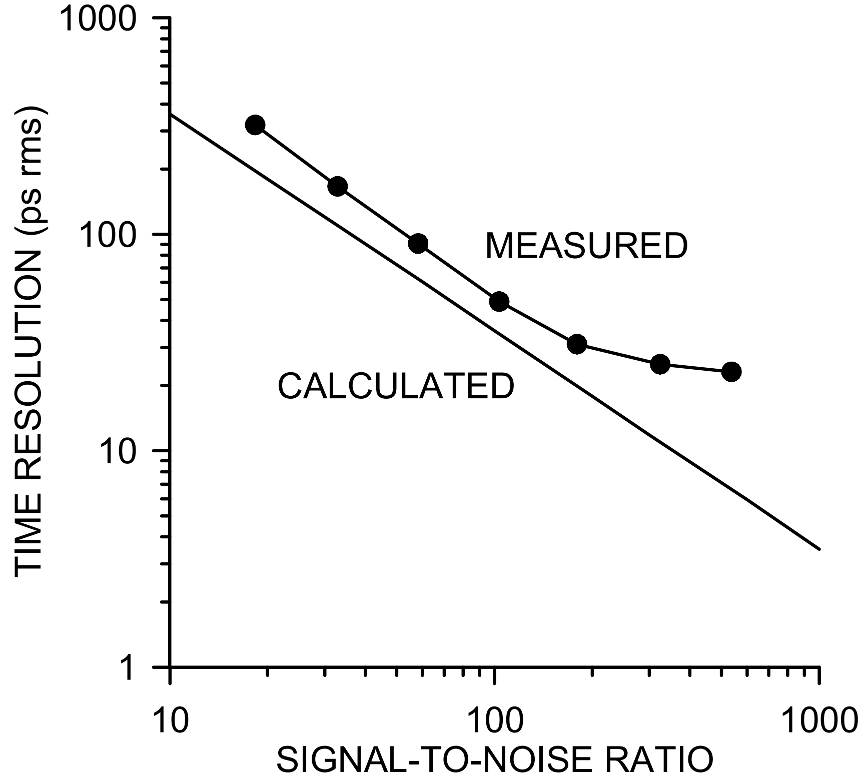
Unlike amplitude measurements, where multiple integrators are desirable to improve pulse symmetry and count rate performance.

Fast Timing: Comparison between theory and experiment

Time resolution $\propto 1/(S/N)$

At S/N <100 the measured curve lies above the calculation because the timing discriminator limited the rise time.

At high S/N the residual jitter of the time digitizer limits the resolution.



For more details on fast timing with semiconductor detectors, see

H. Spieler, IEEE Trans. Nucl. Sci. **NS-29/3** (1982) 1142.

6. Digitization of Pulse Height – Analog to Digital Conversion

For data storage and subsequent analysis the analog signal at the shaper output must be digitized.

Important parameters for ADCs used in detector systems:

1. Resolution
   The "granularity" of the digitized output
2. Differential Non-Linearity
   How uniform are the digitization increments?
3. Integral Non-Linearity
   Is the digital output proportional to the analog input?
4. Conversion Time
   How much time is required to convert an analog signal to a digital output?
5. Count-Rate Performance
   How quickly can a new conversion commence after completion of a prior one without introducing deleterious artifacts?
6. Stability
   Do the conversion parameters change with time?

Instrumentation ADCs used in industrial data acquisition and control systems share most of these requirements. However, detector systems place greater emphasis on differential non-linearity and count-rate performance. The latter is important, as detector signals often occur randomly, in contrast to measurement systems where signals are sampled at regular intervals.

6.1 Resolution

Digitization incurs approximation, as a continuous signal distribution is transformed into a discrete set of values. To reduce the additional errors (noise) introduced by digitization, the discrete digital steps must correspond to a sufficiently small analog increment.

Simplistic assumption:

Resolution is defined by the number of output bits, e.g. 13 bits $\rightarrow \dfrac{\Delta V}{V} = \dfrac{1}{8192} = 1.2 \cdot 10^{-4}$

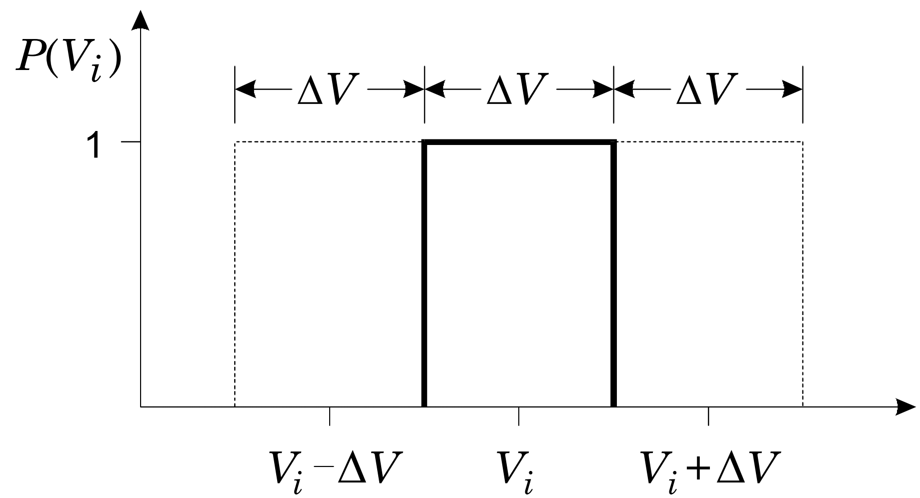
True Measure: Channel Profile

Plot probability vs. pulse amplitude that a pulse height corresponding to a specific output bin is actually converted to that address.

Ideal ADC:

Measurement accuracy:

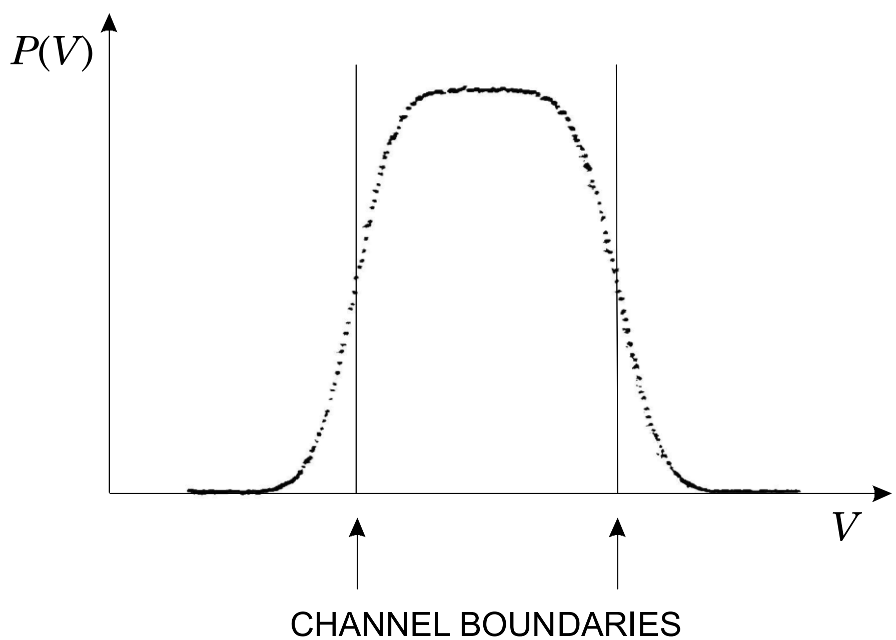
- If all counts of a peak fall in one bin, the resolution is ΔV.
- If the counts are distributed over several bins, peak fitting can yield a resolution of 10^{-1} – 10^{-2} ΔV, *if the distribution is known and reproducible* (not necessarily a valid assumption for an ADC).



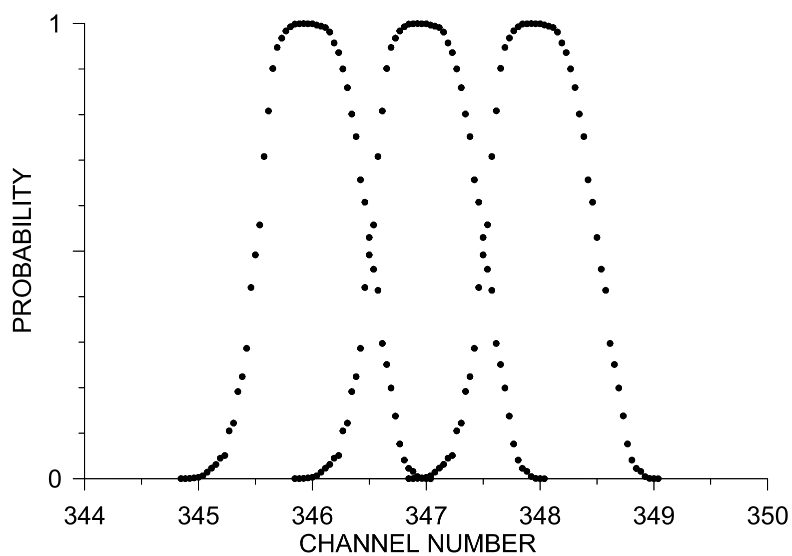
In reality, the channel profile is not rectangular as sketched above.

Electronic noise in the threshold discrimination process that determines the channel boundaries "smears" the transition from one bin to the next.

Measured channel profile (13 bit ADC)



The profiles of adjacent channels overlap.

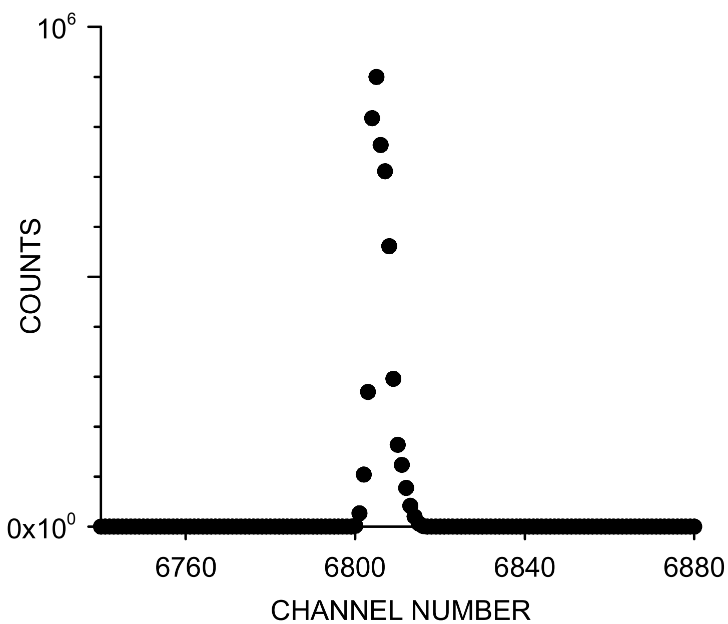


Channel profile can be checked quickly by applying the output of a precision pulser to the ADC.

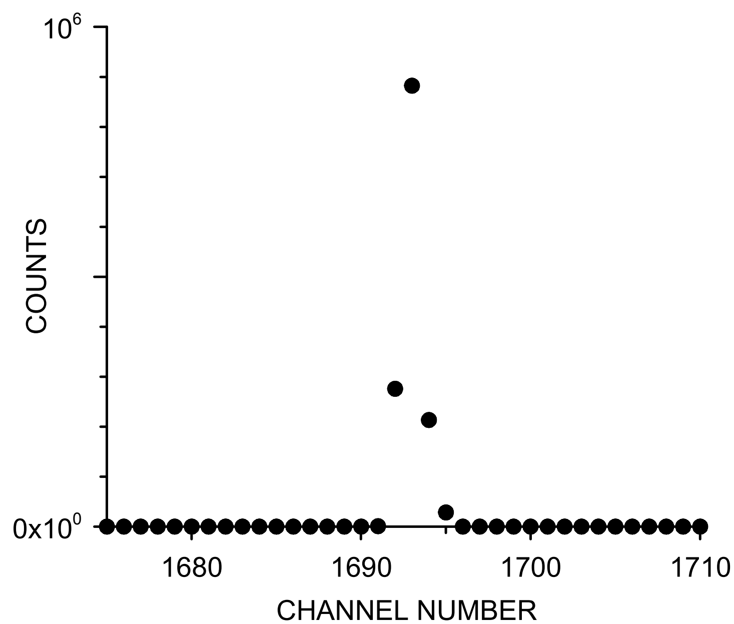
If the pulser output has very low noise, i.e. the amplitude jitter is much smaller than the voltage increment corresponding to one ADC channel or bin, all pulses will be converted to a single channel, with only a small fraction appearing in the neighbor channels.

Example of an ADC whose digital resolution is greater than its analog resolution:

8192 ch conversion range (13 bits)



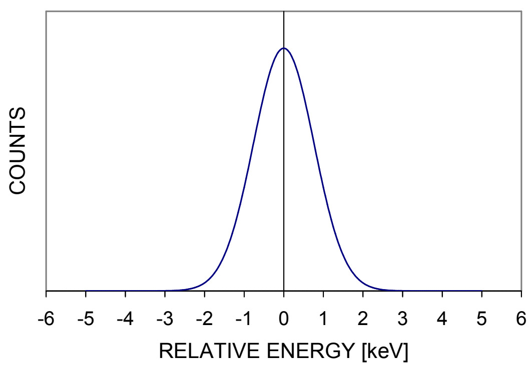
2048 ch conversion range (11 bits)



Here the 2K range provides maximum resolution – higher ranges superfluous.

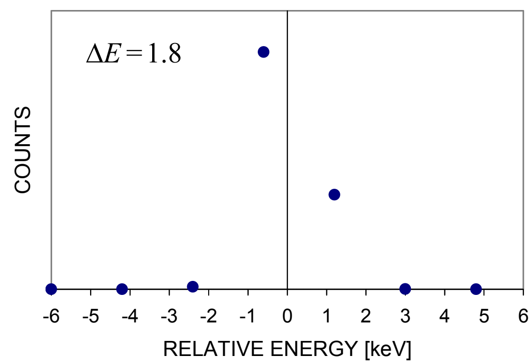
How much ADC Resolution is Required?

Example: Detector resolution ΔE
1.8 keV FWHM

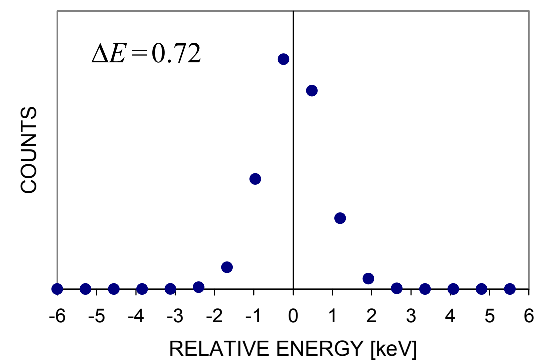


Digitized spectra for various ADC resolutions (bin widths)

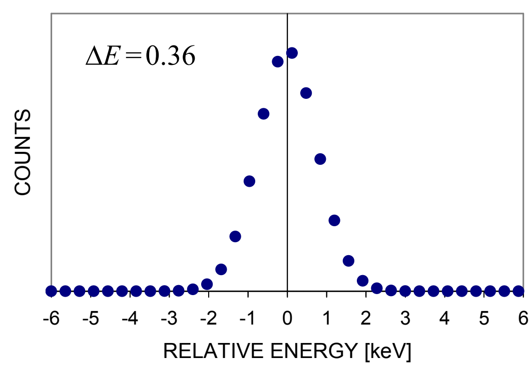
ΔE= 1.8 keV = 1 x FWHM



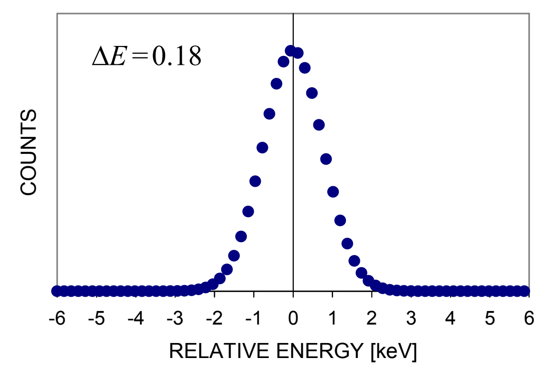
ΔE= 0.72 keV = 0.4 x FWHM



ΔE= 0.36 keV = 0.2 x FWHM



ΔE= 0.18 keV = 0.1 x FWHM



Fitting can determine centroid position to fraction of bin width even with coarse digitization,
if only a single peak is present and the line shape is known.

6.2 Differential Non-Linearity

Differential non-linearity is a measure of the non-uniformity of channel profiles over the range of the ADC.

Depending on the nature of the distribution, either a peak or an rms specification may be appropriate.

$$DNL = \max \left\{ \frac{\Delta V(i)}{\langle \Delta V \rangle} - 1 \right\} \quad \text{or} \quad DNL = \text{r.m.s.} \left\{ \frac{\Delta V(i)}{\langle \Delta V \rangle} - 1 \right\}$$

where $\langle \Delta V \rangle$ is the average channel width and

$\Delta V(i)$ is the width of an individual channel.

Differential non-linearity of < ±1% max. is typical in nuclear science ADCs,

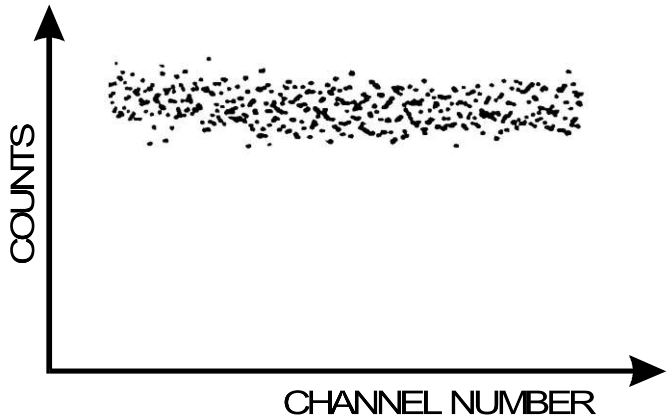
but state-of-the-art ADCs can achieve 10^{-3} rms,

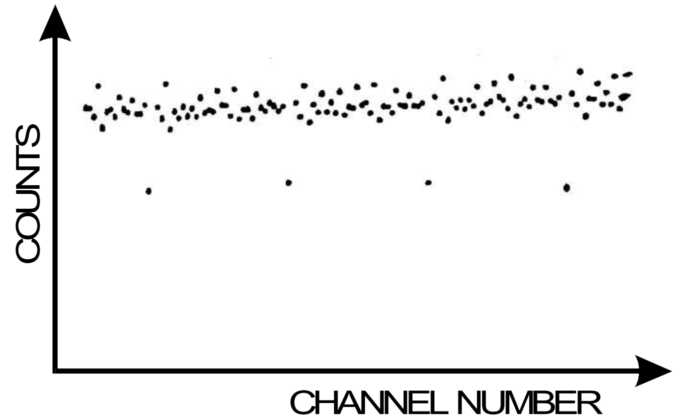
i.e. the variation is comparable to the statistical fluctuation for 10^{6} random counts.

Typical differential non-linearity patterns

“white” input spectrum, suppressed zero   Vertical scale 10x larger than left

↓





An ideal ADC would show an equal number of counts in each bin.

The spectrum to the left shows a random pattern, but
note the multiple periodicities visible in the right hand spectrum.

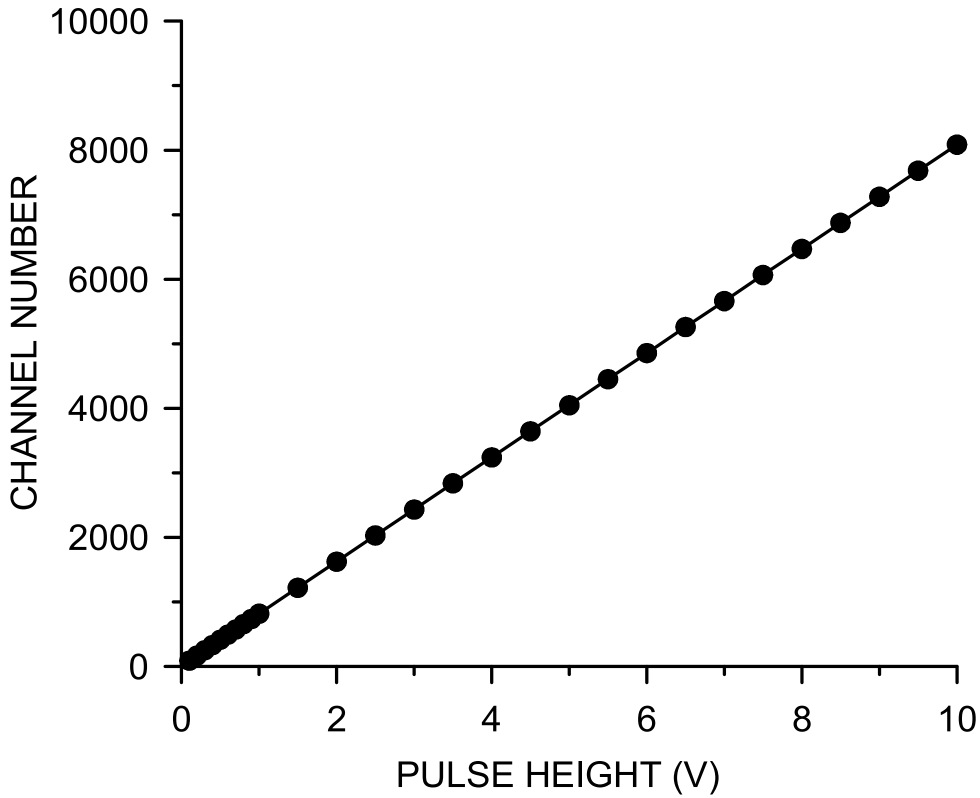
Note: Instrumentation ADCs are often specified with an accuracy of ±0.5 LSB (least significant bit) or more, so

1. the differential non-linearity may be 50% or more,
2. the response may be non-monotonic

$\Rightarrow$ **output may decrease when input rises.**

6.3 Integral Non-Linearity

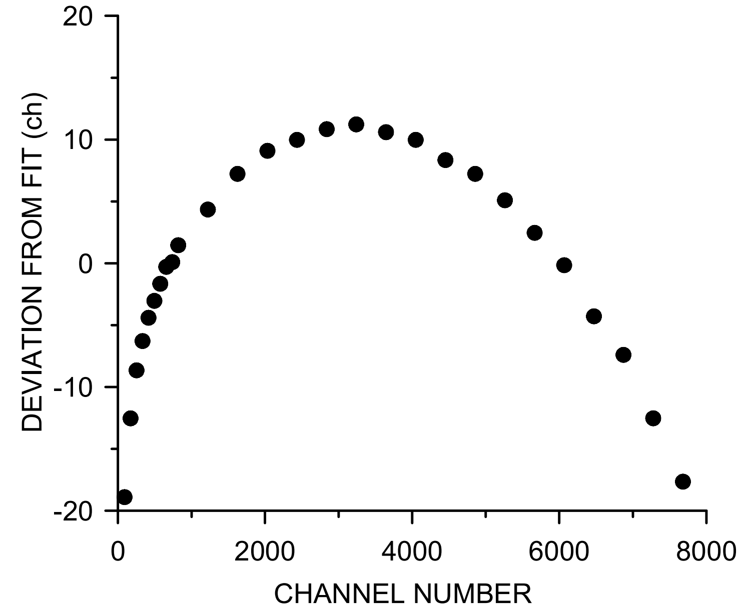
Integral non-linearity measures the deviation from proportionality of the measured amplitude to the input signal level.



The dots are measured values and the line is a fit to the data.

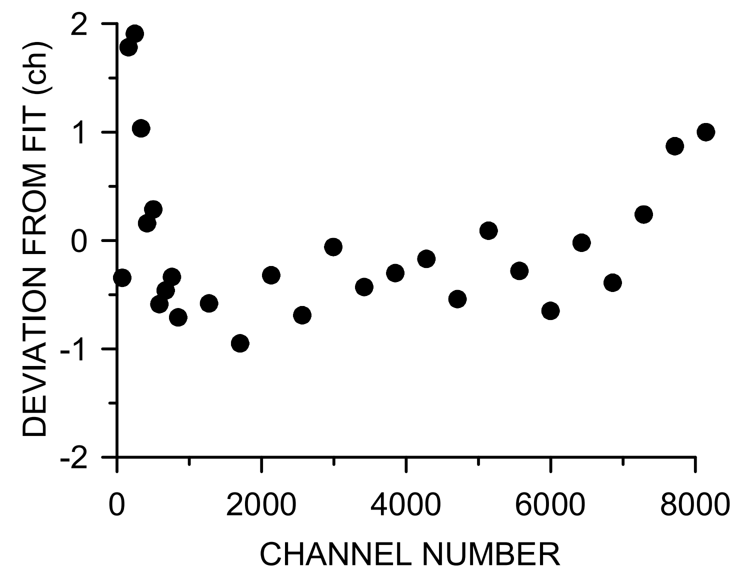
This plot is not very useful if the deviations from linearity are small.

Plotting the deviations of the measured points from the fit yields a more useful result.



Integral non-linearity measured with a 400 ns wide input pulse

The linearity of an ADC can depend on the input pulse shape and duration,
due to bandwidth limitations in the circuitry.



Increasing the pulse width to 3 μs improved the linearity significantly:

6.4 Conversion Time

During the digitization of a signal a subsequent signal is lost ("dead time").

Dead Time =

signal acquisition time	$\rightarrow$	time-to-peak + const.
+ conversion time	$\rightarrow$	can depend on pulse height
+ readout time to memory	$\rightarrow$	depends on speed of data transmission and buffer memory access

Dead time affects measurements of yields or reaction cross-sections. Unless the event rate << 1/(dead time), it is necessary to measure the dead time, e.g. with a reference pulser fed simultaneously into the spectrum.

The total number of reference pulses issued during the measurement is determined by a scaler and compared with the number of pulses recorded in the spectrum.

Does a pulse-height dependent dead time mean that the correction is a function of pulse height?

Usually not. If events in different part of the spectrum are not correlated in time, i.e. random, they are all subject to the same average dead time (although this average will depend on the spectral distribution).

- Caution with correlated events!
  Example: Decay chains, where lifetime is < dead time.
  The daughter decay will be lost systematically.

6.5 Count Rate Effects

Problems are usually due to internal baseline shifts with event rate or undershoots following a pulse.

If signals occur at constant intervals, the effect of an undershoot will always be the same.

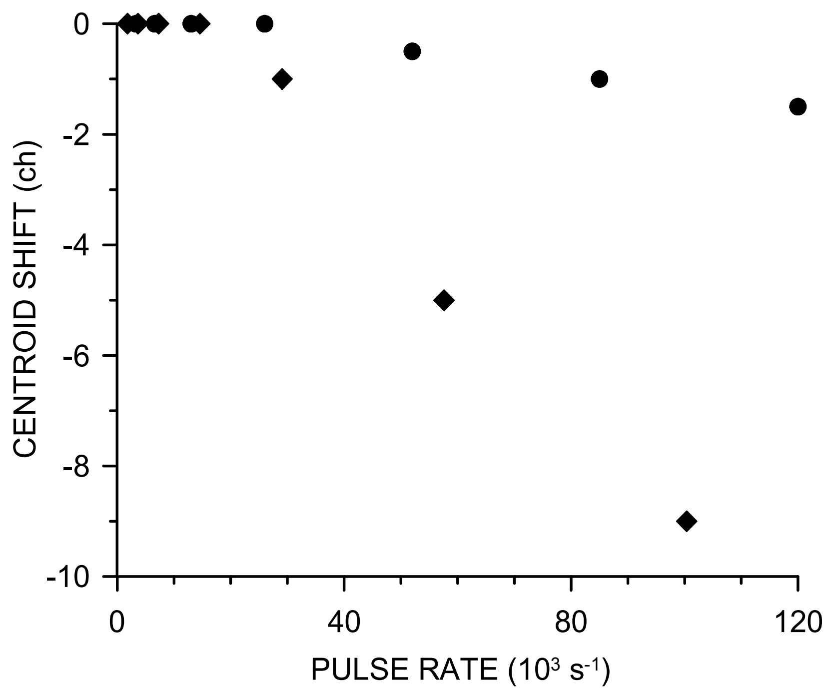
However, in a random sequence of pulses, the effect will vary from pulse to pulse.

⇒ spectral broadening

Baseline shifts tend to manifest themselves as a systematic shift in centroid position with event rate.

Centroid shifts for two 13 bit ADCs vs. random rate:

Not all ADCs work as expected.



6.6 Stability

Stability vs. temperature is usually adequate with modern electronics in a laboratory environment.

- Note that temperature changes within a module are typically much smaller than ambient.

However: Highly precise or long-term measurements require spectrum stabilization to compensate for changes in gain and baseline of the overall system.

Technique: Using precision pulsers place a reference peak at both the low and high end of the spectrum.

(Pk. Pos. 2) – (Pk. Pos. 1) $\rightarrow$ Gain, …

then

(Pk. Pos. 1) or (Pk. Pos. 2) $\rightarrow$ Offset

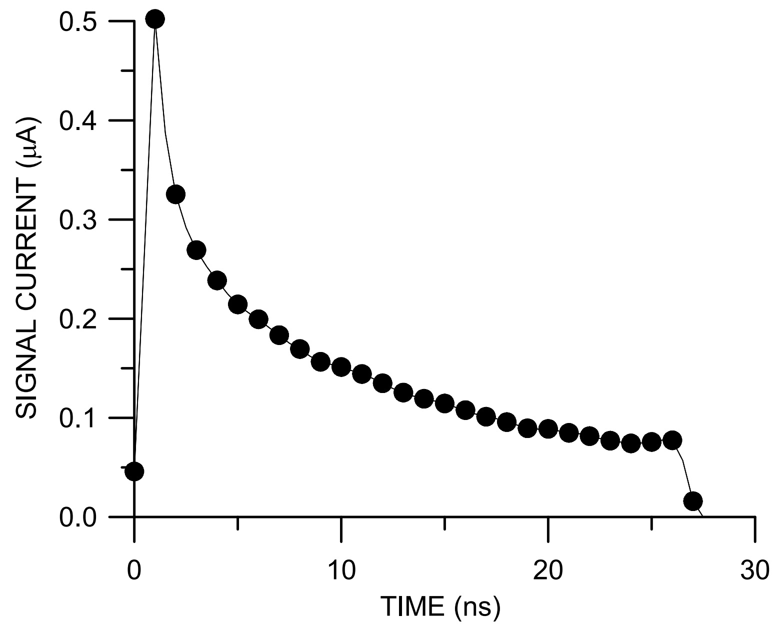
Traditional Implementation: Hardware, spectrum stabilizer module

Today, it is more convenient to determine the corrections in software.

These can be applied to calibration corrections or used to derive an electrical signal that is applied to the hardware (simplest and best in the ADC).

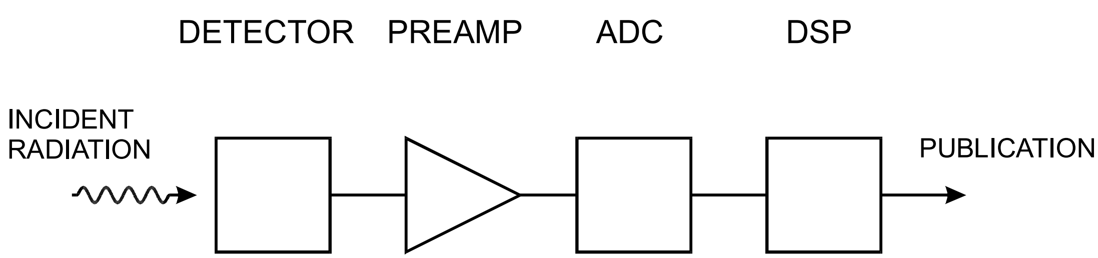
7. Digital Signal Processing

Sample detector signal with fast digitizer to reconstruct pulse:



Then use digital signal processor to perform mathematical operations for desired pulse shaping.

Block Diagram

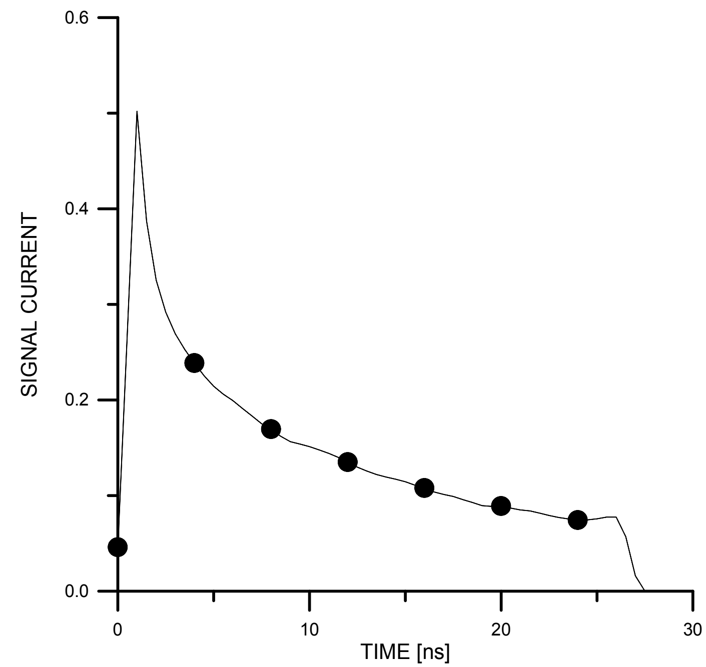


DSP allows great flexibility in implementing filtering functions

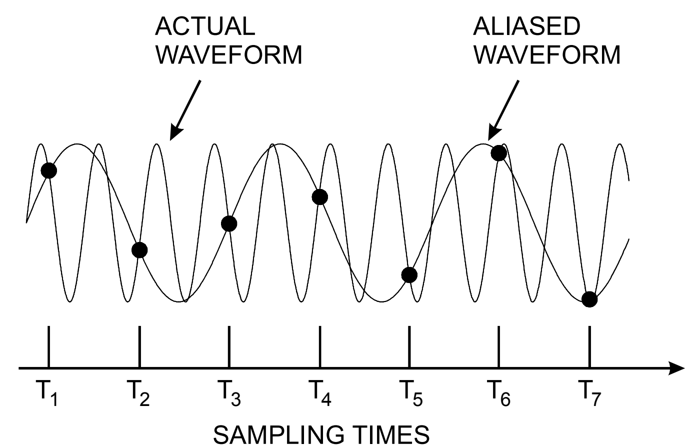
However: increased circuit complexity
increased demands on ADC, compared to traditional shaping.

Important to choose sample interval sufficiently small to capture pulse structure.

Sampling interval of 4 ns misses initial peak.



With too low a sampling rate high frequency components
will be "aliased" to lower frequencies:

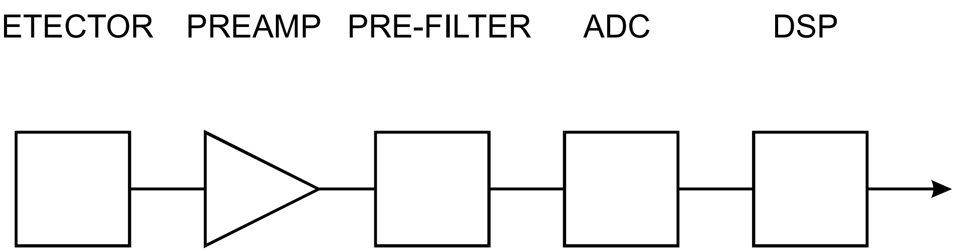


Applies to any form of sampling
(time waveform, image, ...)

Nyquist condition:

Sampling frequency > 2x highest signal frequency

⇒ Fast ADC required + Pre-Filter to limit signal bandwidth



- Dynamic range requirements for ADC may be more severe than in analog filtered system (depending on pulse shape and pre-filter).
- Digitization introduces additional noise ("quantization noise")

  If one bit corresponds to an amplitude interval Δ, the quantization noise

$$\sigma_v^2 = \int_{-\Delta/2}^{\Delta/2} \frac{v^2}{\Delta}\, dv = \frac{\Delta^2}{12} .$$

  (differential non-linearity introduces quasi-random noise)

- Electronics preceding ADC and front-end of ADC must exhibit same precision as analog system, i.e. baseline and other pulse-to-pulse amplitude fluctuations less than order Q_n /10, i.e. typically 10^{-4} in high-resolution systems. For 10 V FS at the ADC input this corresponds to < 1 mV.

⇒ ADC must provide high performance at short conversion times. Today this is technically feasible for some applications, e.g. detectors with moderate to long collection times (γ and x-ray detectors).

Digital Filtering

Filtering is performed by convolution: $$S_o(n) = \sum_{k=0}^{N-1} W(k) \cdot S_i(n-k)$$

$W(k)$ is a set of coefficients that describes the weighting function yielding the desired pulse shape.

A filter performing this function is called a Finite Impulse Response (FIR) filter.

This is analogous to filtering in the frequency domain:

In the frequency domain the result of filtering is determined by multiplying the responses of the individual stages:

$$G(f) = G_1(f) \cdot G_2(f)$$

where $G_1(f)$ and $G_1(f)$ are complex numbers.

The theory of Fourier transforms states that the equivalent result in the time domain is formed by convolution of the individual time responses:

$$g(t) = g_1(t) * g_2(t) \equiv \int_{-\infty}^{+\infty} g_1(\tau) \cdot g_2(t-\tau) d\tau \,,$$

analogously to the discrete sum shown above.

ADCs often have excessive noise.

Increasing gain to increase the signal level into the ADC will reduce the ADC noise contribution, but reduce the dynamic range.

For a preamplifier input noise v_{n1} fed to the ADC input noise v_{n2} with a gain G, the overall noise

$$v_n = \frac{\sqrt{(v_{n1}G)^2 + v_{n2}^2}}{G} = \sqrt{v_{n1}^2 + \left(\frac{v_{n2}}{G}\right)^2}$$

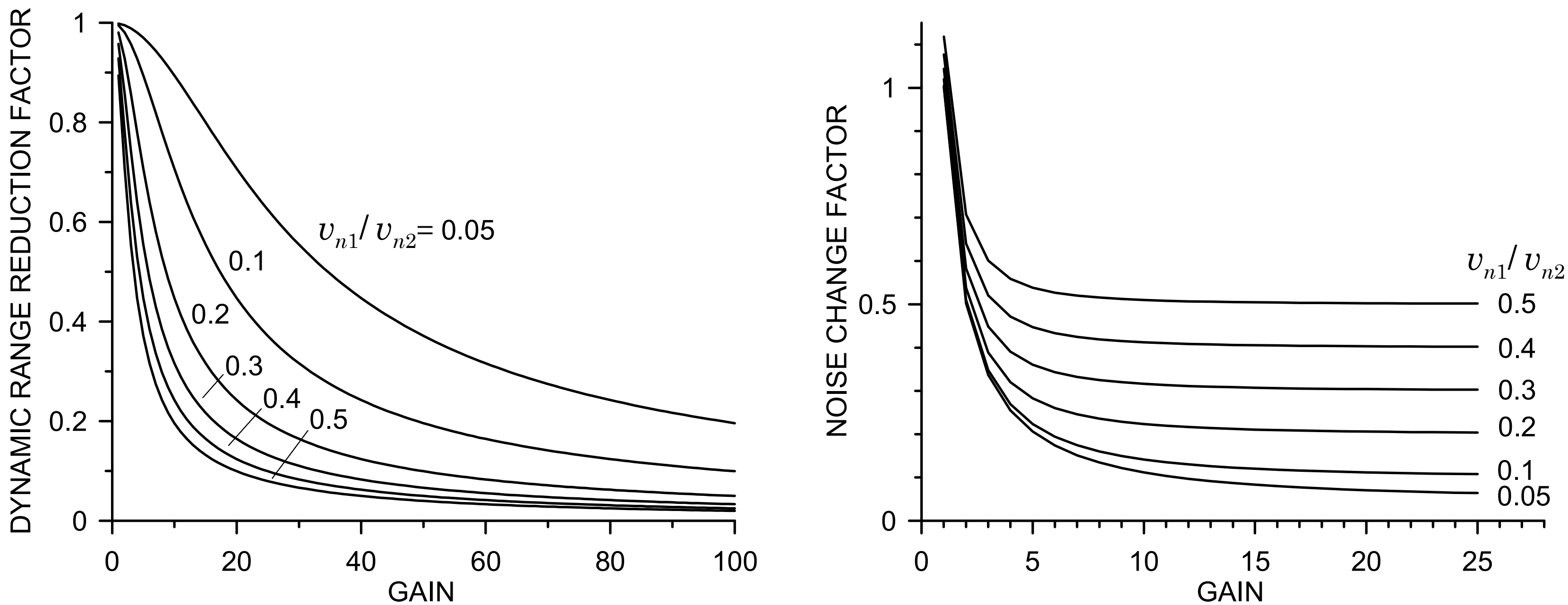
The ratio of the overall noise to the ADC noise $$\frac{v_n}{v_{n2}} = \sqrt{\left(\frac{v_{n1}}{v_{n2}}\right)^2 + \left(\frac{1}{G}\right)^2}$$

The resulting dynamic range given by the ratio of the maximum ADC input signal level to the combined noise is

$$\frac{V_i^{max}}{v_n} = \frac{V_{i0}^{max}/G}{\sqrt{v_{n1}^2 + \left(v_{n2}/G\right)^2}} = \frac{V_{i0}^{max}}{v_{n2}} \frac{1}{\sqrt{\left(\frac{v_{n1}}{v_{n2}}G\right)^2 + 1}}$$

The second factor is the reduction of dynamic range.

Change in dynamic range and noise vs. pre-ADC gain for various ratios of the pre-circuit noise v_{n1} to the ADC noise v_{n2}



Given a sufficient sampling rate, digital signal processing will provide the same results as an analog system.

The equivalent noise charge expression yielding the results of current noise, voltage, and $1/f$ voltage noise

$$Q_n^2 = i_n^2 F_i T_S + e_n^2 F_v \frac{C^2}{T_S} + F_{vf} A_f C^2$$

applies to both analog and digital signal processing, provided the additional noise contributions in digital processing are negligible.

Potential Problems as noted above are insufficient sampling rate and excessive ADC noise, in addition to digital crosstalk.

Digital vs. Analog Processing

Benefits of digital signal processing:

- Flexibility in implementing filter functions
- Filters possible that are impractical in hardware
- Simple to change filter parameters
- Tail cancellation and pile-up rejection easily incorporated
- Adaptive filtering can be used to compensate for pulse shape variations.

Where is digital signal processing appropriate?

- Systems highly optimized for Resolution
  High counting rates
- Variable detector pulse shapes

Where is analog signal processing best (most efficient)?

- Fast shaping
- Systems not sensitive to pulse shape (fixed shaper constants)
- High density systems that require small circuit area or low power

Both types of systems require careful analog design.

Progress in fast ADCs (precision, reduced power) will expand range of DSP applications.

VI. Why Things Don't Work
– Why *S/N* Theory Often Seems to be Irrelevant

Throughout the previous lectures it was assumed that the only sources of noise were

- random
- known
- in the detector, preamplifier, or associated components

In practice, the detector system will pick up spurious signals that are

- not random,
- but not correlated with the signal,

so with reference to the signal they are quasi-random.

⇒ Baseline fluctuations superimposed on the desired signal

⇒ Increased detection threshold, degradation of resolution

Important to distinguish between

- pickup of spurious signals, either from local or remote sources (clocks, digital circuitry, readout lines),
- self-oscillation
  (circuit provides feedback path that causes sustained oscillation due to a portion of the output reaching the input)

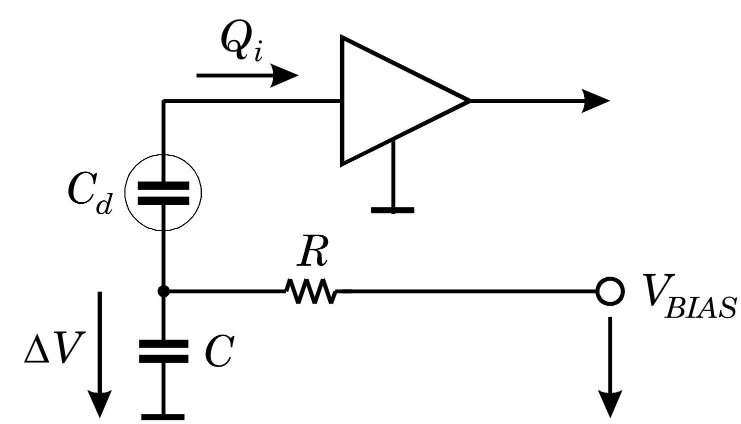
External Pickup is often the cause, but many problems are due to poor work practices or inappropriate equipment

1. Example: Noisy Detector Bias Supplies

The detector is the most sensitive node in the system.

Any disturbance ΔV on the detector bias line will induce charge in the input circuit.

$$\Delta Q = C_d \Delta V$$



$\Delta V = 10$ μV and 10 pF detector capacitance yield
$\Delta Q \approx$ 0.1 fC – about 600 el or 2 keV (Si).

Especially when the detector bias is low (<100V),
it is tempting to use a general laboratory power supply.

Frequently, power supplies are very noisy – especially old units.
The RC circuits in the bias line provide some filtering, but usually not enough for a typical power supply

Beware of switching power supplies. Well-designed switching regulators can be very clean, but most switchers are very noisy.

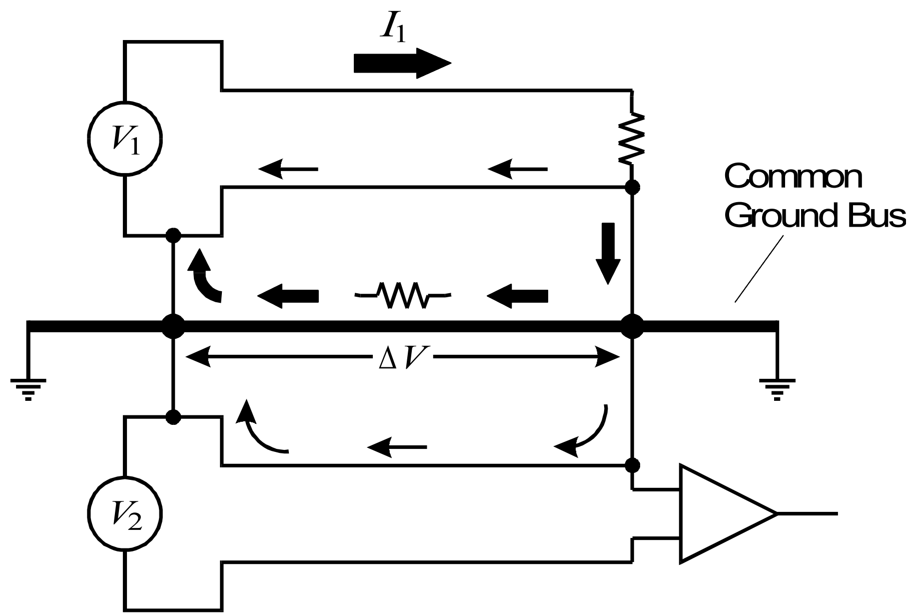
Spikes on the output can be quite large, but short, so that the rms noise specification may appear adequate.

2. Shared Current Paths – Grounding and the Power of Myth

Although capacitive or inductive coupling cannot be ignored, the most prevalent mechanism of undesired signal transfer is the existence of shared signal paths.

Mechanism:

A large alternating current I_1 is coupled into the common ground bus.



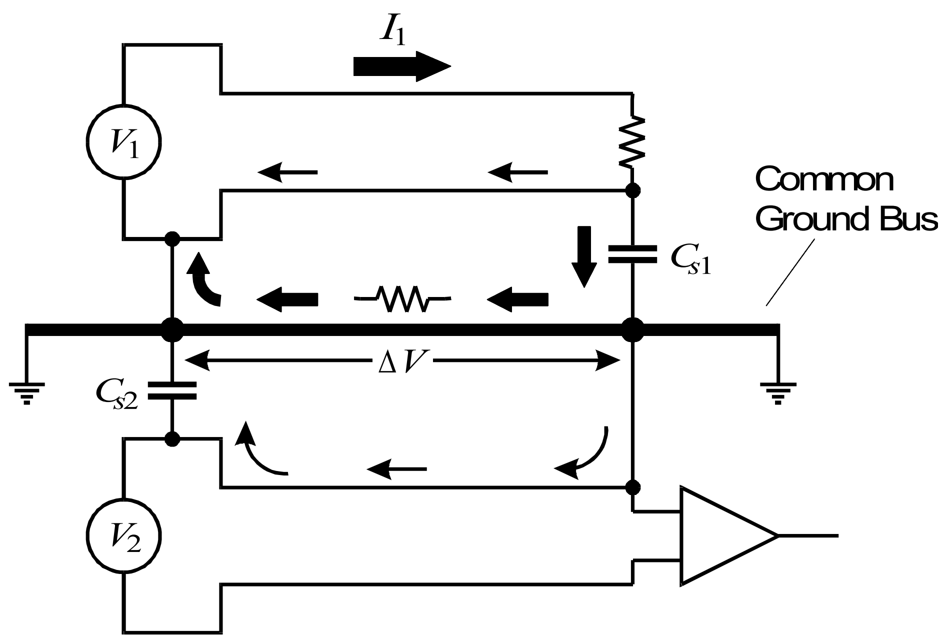
Although the circuit associated with generator V_1 has a dedicated current return, the current seeks the past of least resistance, which is the massive ground bus.

The lower circuit is a sensitive signal transmission path. Following the common lore, it is connected to ground at both the source and receiver.

The large current flowing through the ground bus causes a voltage drop ΔV, which is superimposed on the low-level signal loop associated with V_2 and appears as an additional signal component.

- Cross-coupling has *nothing to do with grounding per se*, but is due to the common return path. However, the common ground caused the problem by establishing the shared path.

In systems that respond to transients (i.e. time-varying signals) rather than DC signals, secondary loops can be closed by capacitance. A DC path is not necessary.



The loops in this figure are the same as shown before, but the loops are closed by the capacitances C_{s1} and C_{s2}.

Frequently, these capacitances are not formed explicitly by capacitors, but are the stray capacitance formed by a power supply to ground, a detector to its support structure (as represented by C_{s2}), etc.

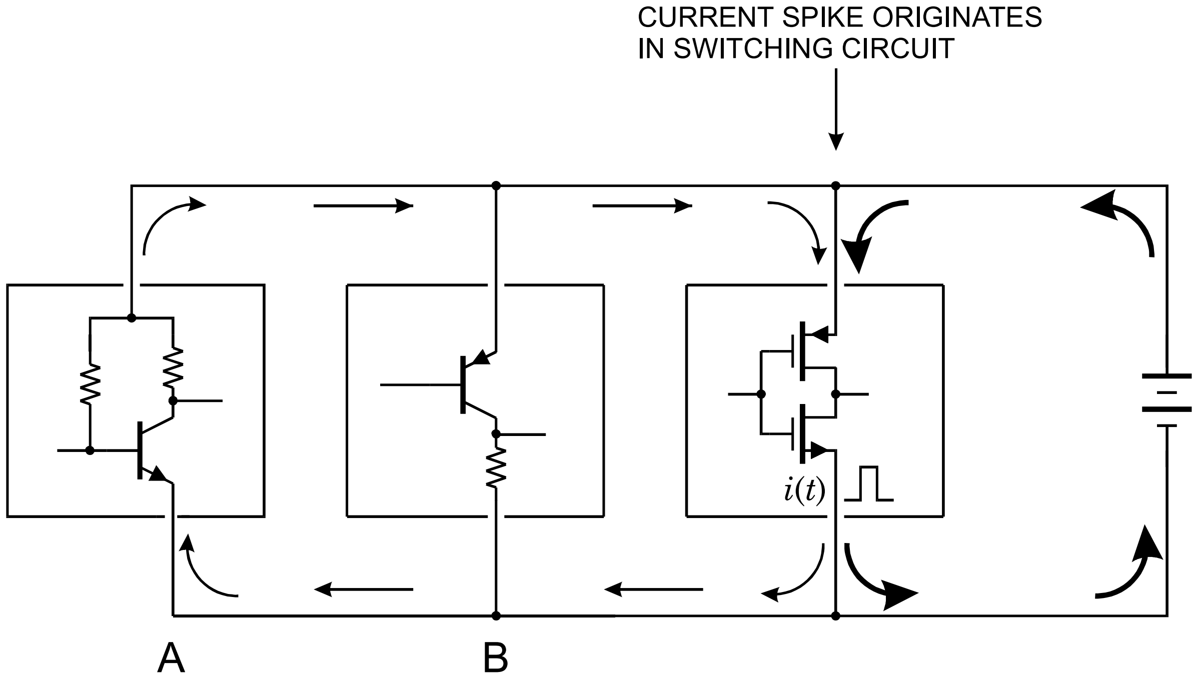
For AC signals the inductance of the common current path can increase the impedance substantially beyond the DC resistance, especially at high frequencies.

This mode of interference occurs whenever spurious voltages are introduced into the signal path and superimpose on the desired signal.

Interference does not cross-couple by voltage alone, but also via current injection.

Current spikes originating in logic circuitry, for example, propagate through the bussing system as on a transmission line.

Individual connection points will absorb some fraction of the current signal, depending on the relative impedance of the node.



Current also flows into low impedance node A (common base stage), which closes the secondary loop,

but not into high impedance node B

Remedial Techniques

Reduce impedance of the common path

$\Rightarrow$ Copper Braid Syndrome

Colloquially called “improving ground”.

(sometimes fortuitously introduces an out-of-phase component of the original interference, leading to cancellation)

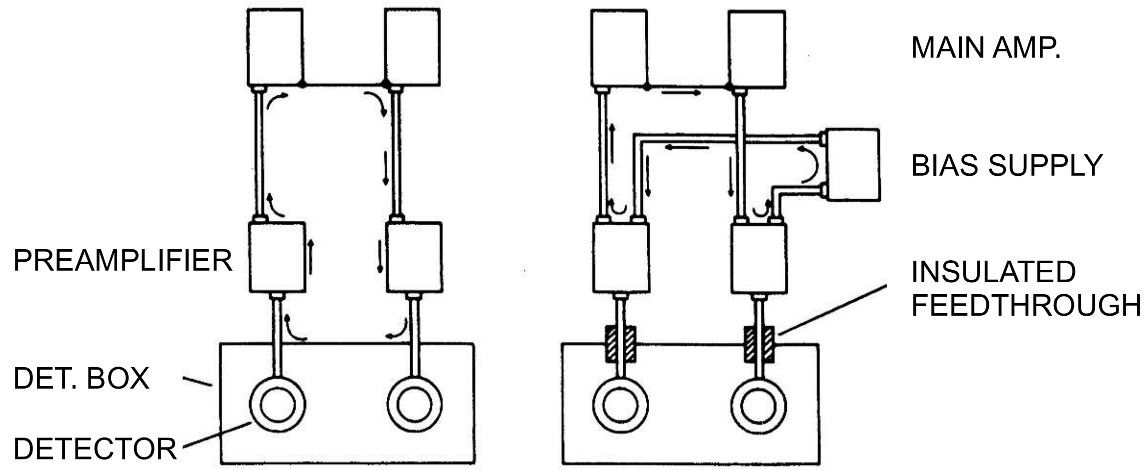
Rather haphazard, poorly controlled $\Rightarrow$ continual surprises

Avoid Grounds and break parasitic signal paths

Example:

The configuration at the left has a loop that includes the most sensitive part of the system – the detector and preamplifier input.

By introducing insulated feed-throughs, the input loop is broken.



Note that a new loop is shown, introduced by the common detector bias supply. This loop is restricted to the output circuit of the preamplifier, where the signal has been amplified, so it is less sensitive to interference.

- Note that the problem is not caused by loops *per se*, i.e. enclosed areas, but by the multiple connections that provide entry paths for interference.

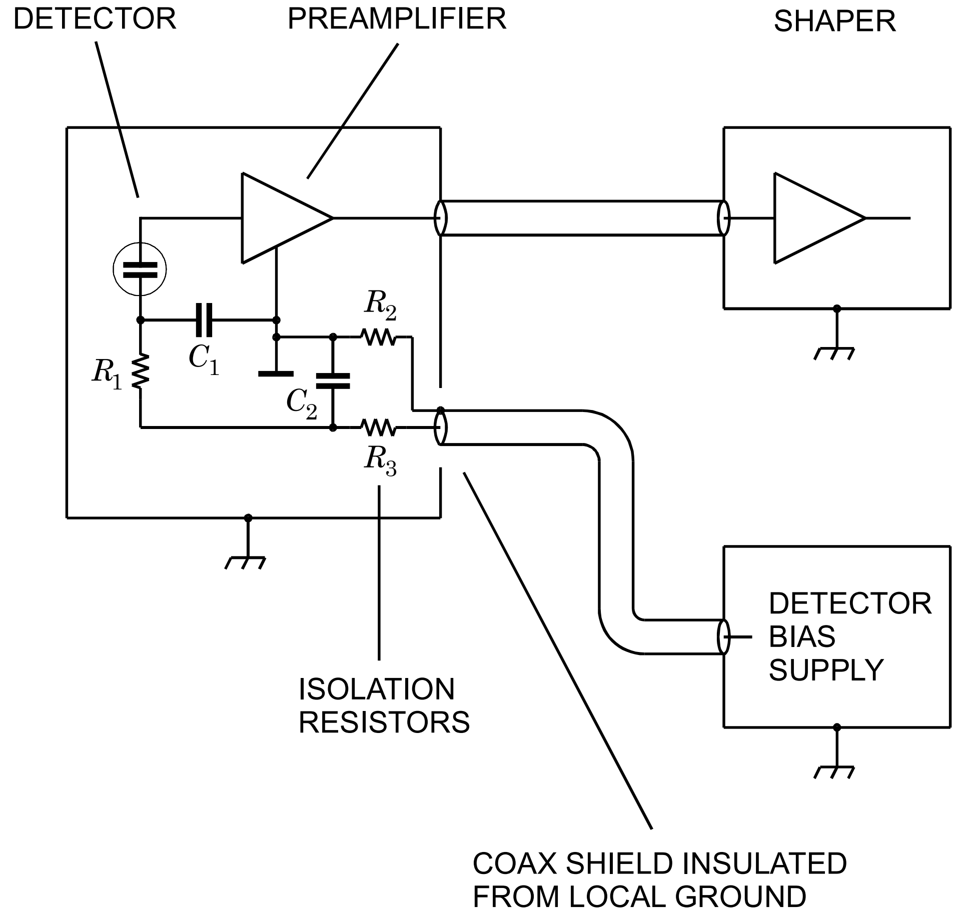
- Although not shown in the schematic illustrations above, both the “detector box” (e.g. a scattering chamber) and the main amplifiers (e.g. in a NIM bin or VME crate) are connected to potential interference sources, so currents can flow through parts of the input signal path.

Series resistors isolate parasitic ground connections.

Example: Detector bias voltage connection

Isolation resistors can also be mounted in an external box that is looped into the bias cable. Either use an insulated box or be sure to isolate the shells of the input and output connectors from another.

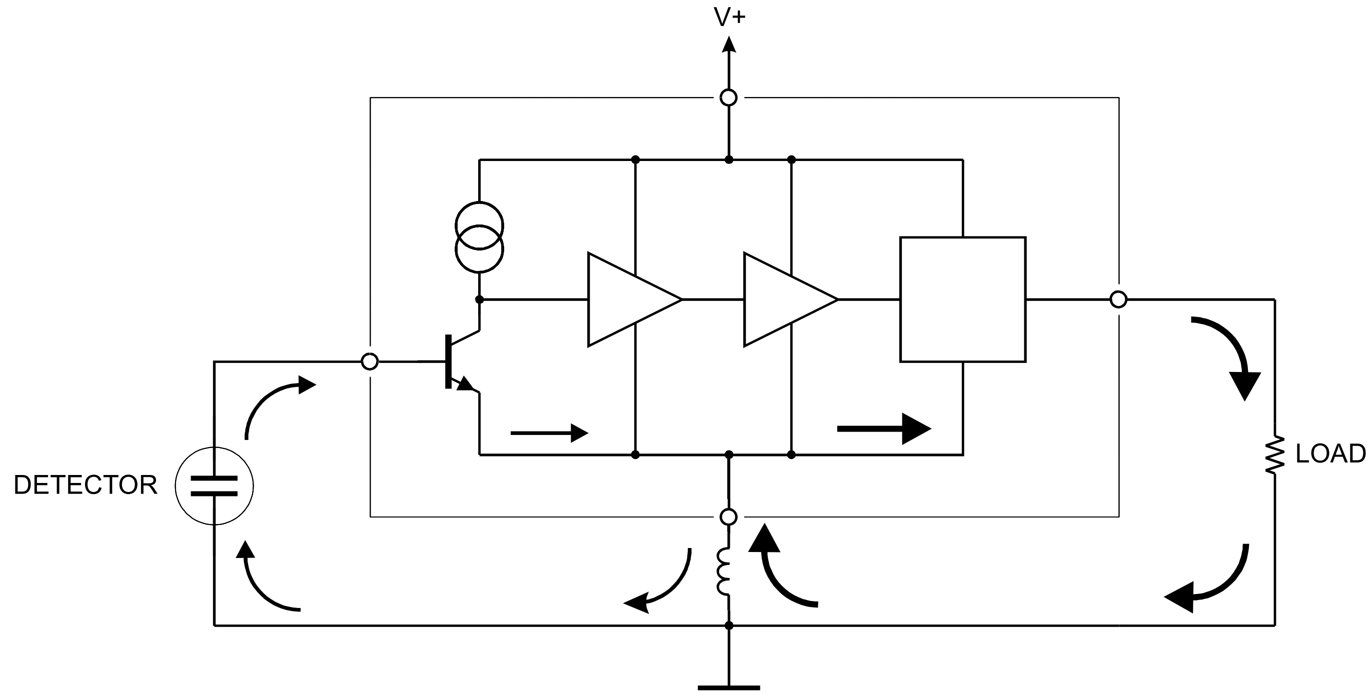
A simple check for noise introduced through the detector bias connection is to use a battery.



"Ground loops" are often formed by the third wire in the AC power connection. Avoid voltage differences in the "ground" connection by connecting all power cords associated with low-level circuitry into the same outlet strip.

“Ground” Connections in Multi-Stage Circuits

IC combining a preamplifier, gain stages and an output driver:

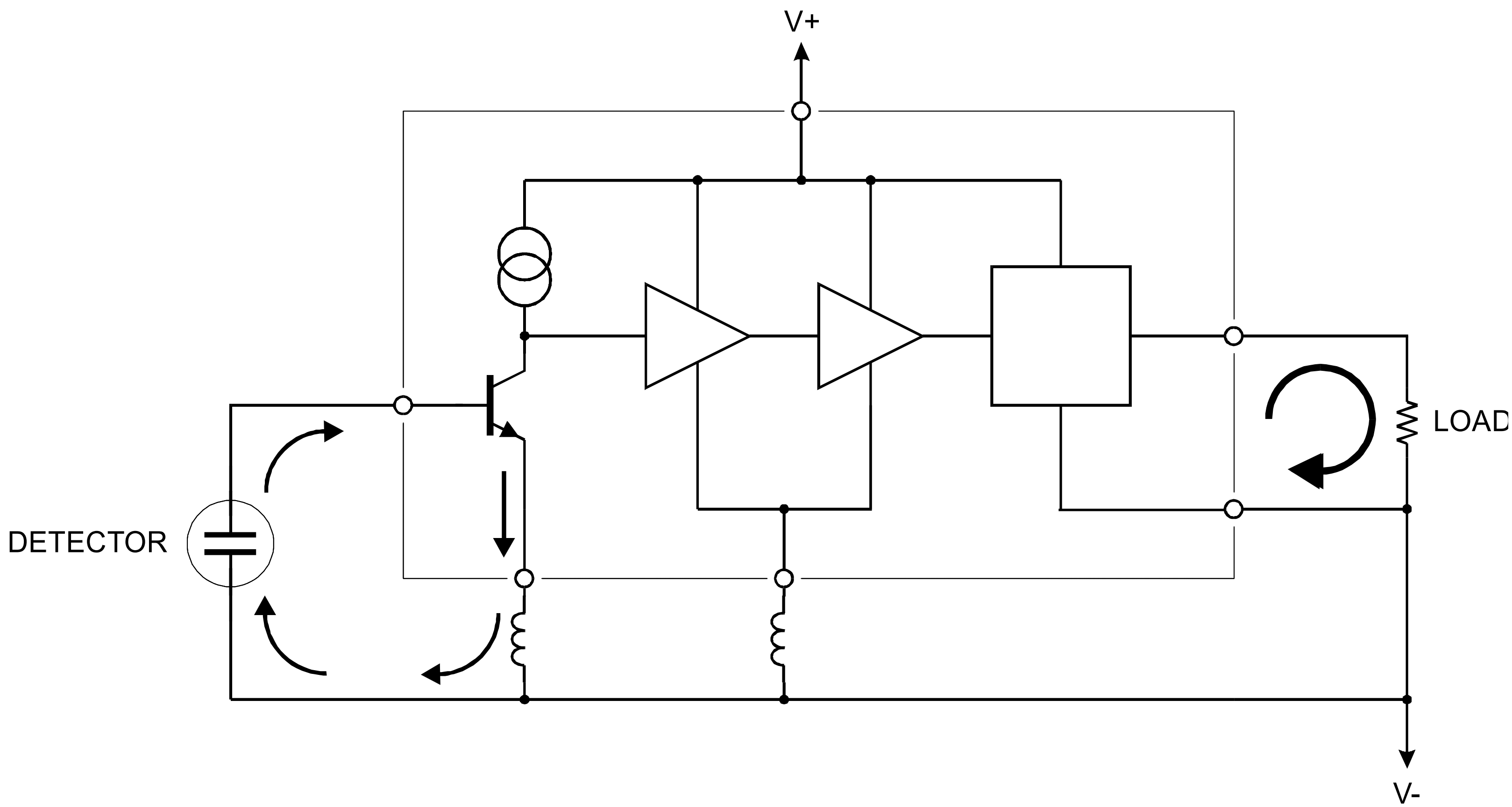


The output current is typically orders of magnitude greater than the input current (due to amplifier gain, load impedance).

Combining all ground returns in one bond pad creates a shared impedance (inductance of bond wire).

This also illustrates the use of a popular technique – the “star” ground – and its pitfalls.

Separating the “ground” connections by current return paths routes currents away from the common impedance and constrains the extent of the output loop, which tends to carry the highest current.

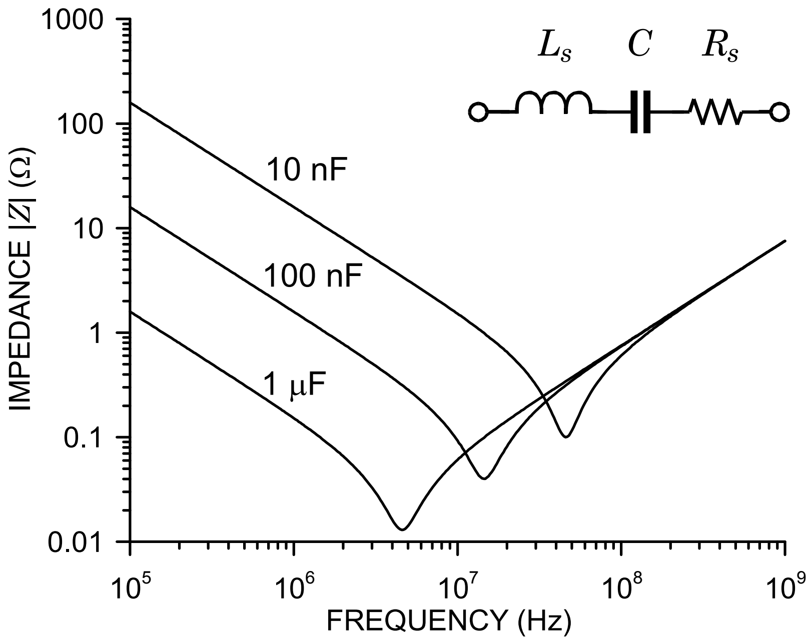


3. Choice of Capacitors

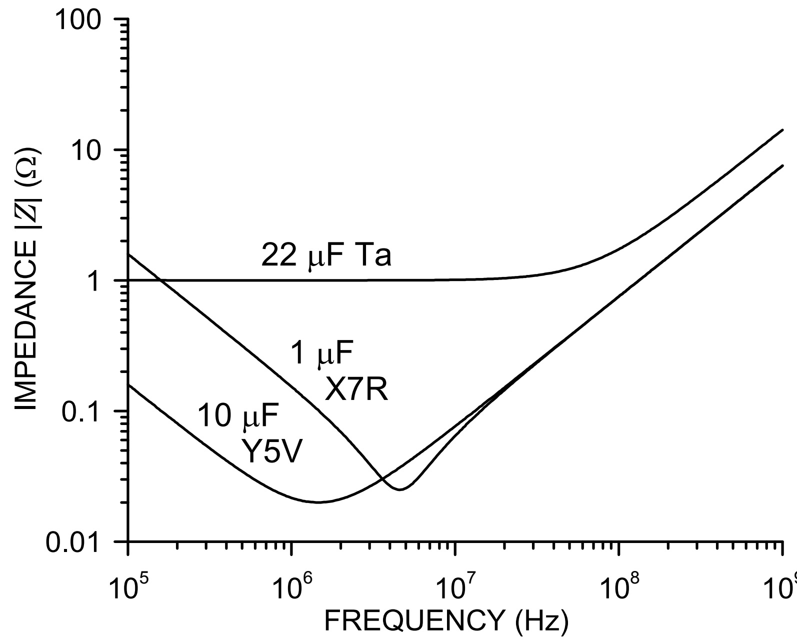
Capacitors are often in signal return paths. As bypass capacitors they must buffer large current pulses.

All capacitors form a series resonant circuit. Above the resonance frequency they act inductive. The loss resistance (equivalent series resistance, ESR) limits the impedance at resonance.

Impedance of X7R ceramic chip capacitors



Comparison between X7R, Y5V ceramic and Tantalum chip capacitors.



Y5V has large voltage dependence, typ. ~20% of nominal capacitance at rated voltage. X7R better.

Summary

- Detectors involve a wide range of interacting functions – often conflicting.

  Requires understanding the physics of the

  experiment, detector, and readout,

  rather then merely following recipes.

- Physics requirements must be translated to engineering parameters.
- Many details interact, even in conceptually simple designs.

  View in different aspects, e.g. analysis in time and frequency domain

- Single-channel recipes tend to be incomplete
  – Overall interactions must be considered.
- Don't blindly accept the results of simulations. Do cross checks!
- Novel detectors often build on a range of different concepts.
  Appropriate compromises often enable systems that were called impractical.

The broad range of physics in novel detector development brings you into more science than run-of-the-mill data analysis.