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Refresher Course

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Silicon Detectors

Many different applications, but built on the same basic physics

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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

I. Introduction

- Traditional Applications
- Multi-Electrode Position Sensing

II. Detector Signals

- Interactions of Charged Particles
- Interactions of Photons
- Detector Sensitivity
- Avalanche Photodiodes
- Silicon Photomultipliers
- Electric Signal Formation
- Semiconductor Materials

III. Front-End Electronics

- Voltage vs. Current Mode
- Charge-Sensitive Amplifier
- Input Impedance

IV. Electronic Noise

- Noise Types
- Amplifier Noise
- Cross-Coupled Noise

V. Signal Processing

- Requirements
- Shaper Examples
- Noise Charge vs. Shaping Time
- Threshold Discriminator Systems
- Timing Measurements

VI. Summary

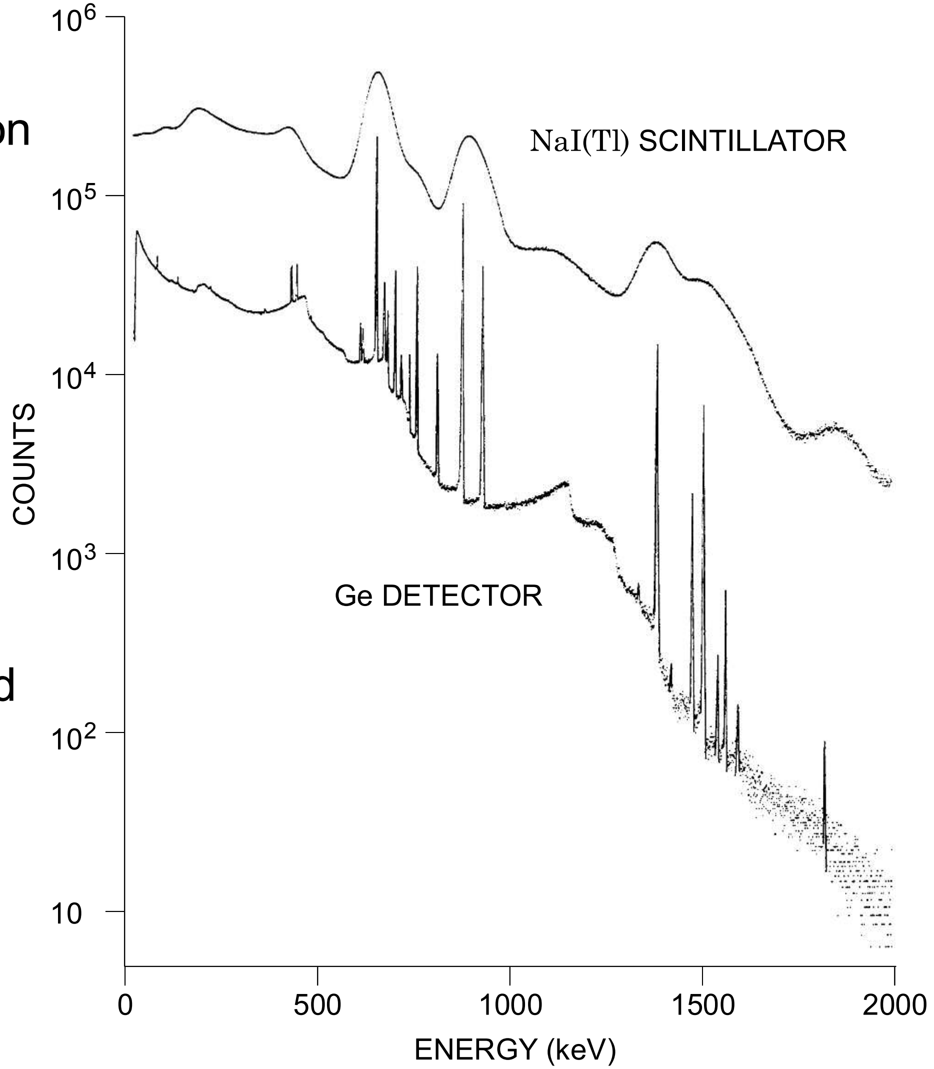
Traditional Applications of Semiconductor Detectors

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 semiconductor 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.

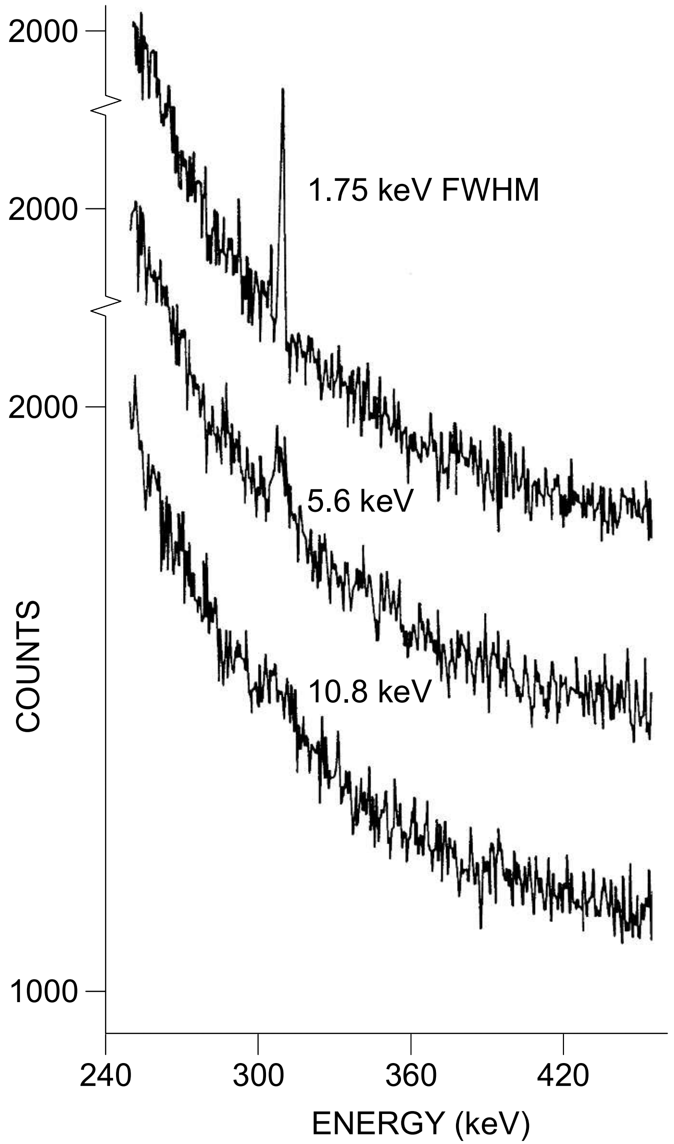
Energy resolution improves sensitivity because

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

X-Ray Fluorescence

When excited by radiation of sufficient energy, atoms emit characteristic x-rays that can be used to detect trace contaminants.

⇒ high sensitivity with small samples.

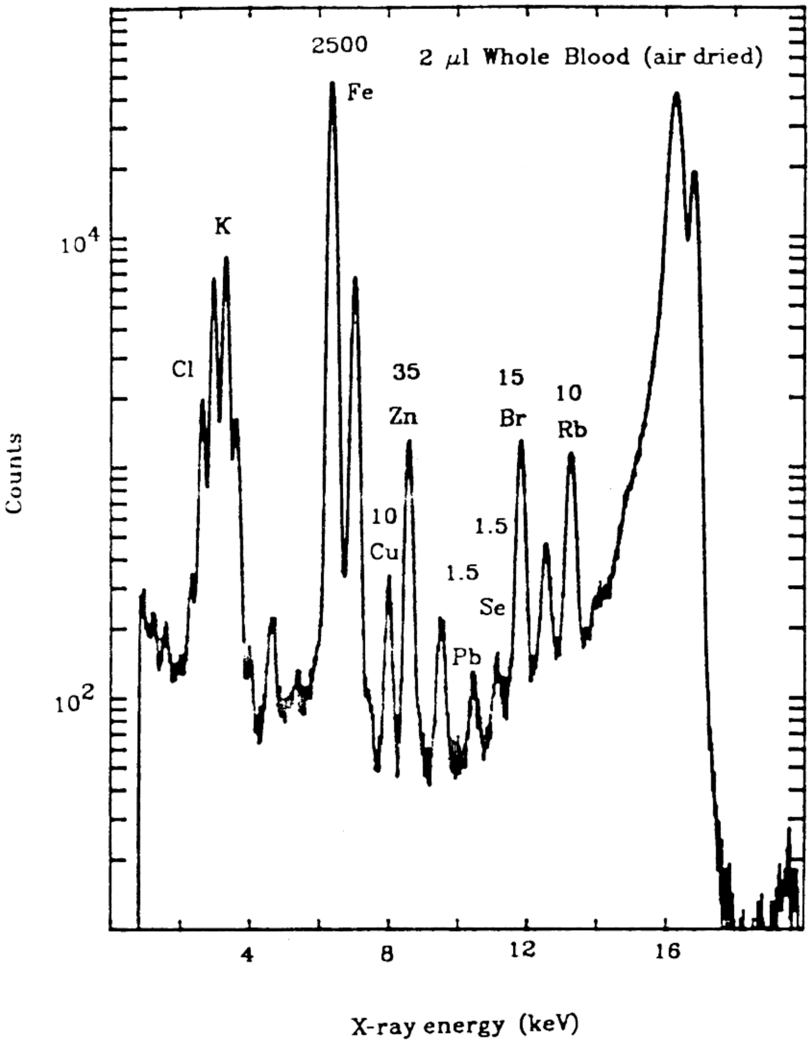
Spectrum taken from 2 μl (2 mm^3) of blood:

Concentrations are given in parts per million

Note the Pb peak

(measurement taken before the introduction of unleaded gasoline).

In many applications weak signals are to be recognized next to strong signals.



(Joe Jaklevic, Engineering Div. LBNL)

Nuclei Z and A Identification

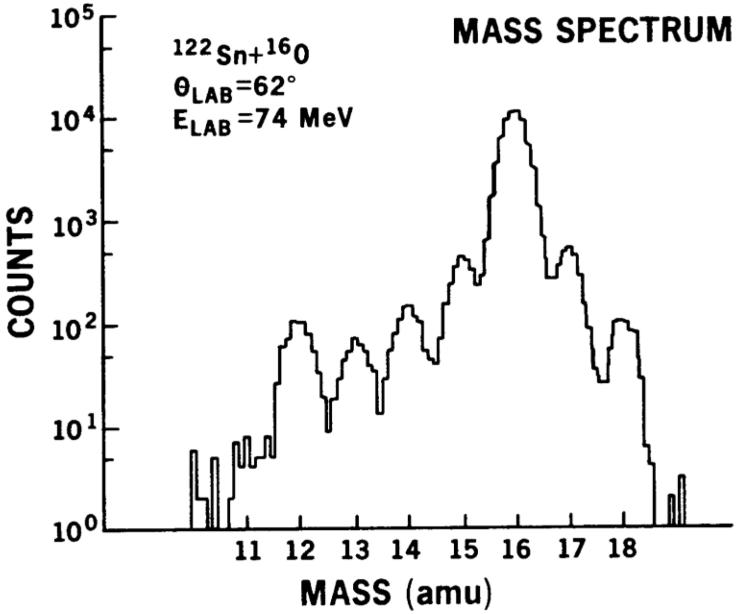
Thin Si detectors have ns collection times and provide ps time resolution.

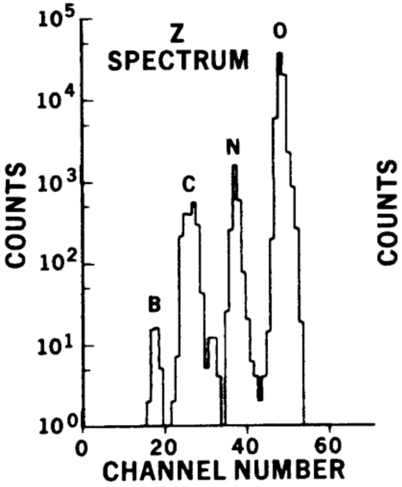
Heavy ion interactions yield a wide range of nuclei.

A detector system that determines partial energy loss ΔE and total energy E yields the nuclear charge Z.

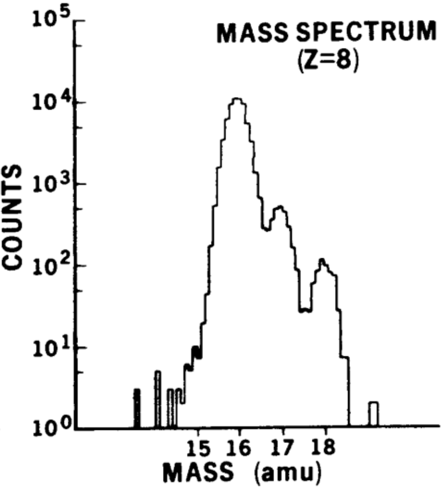
Measuring the time of flight yields the mass.

a)





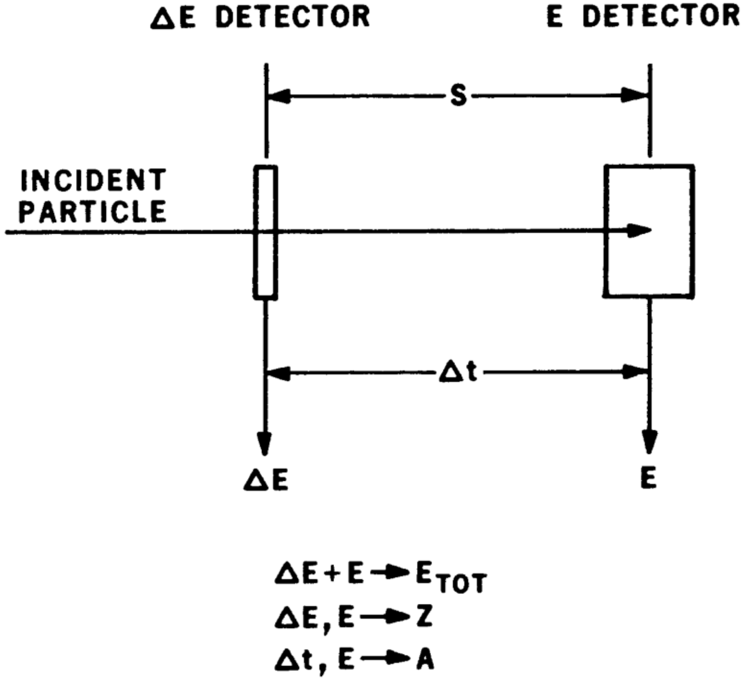
b)



c)

Spieler et al., Z. Physik **A278** (1976) 241

Nuclear Time-of-Flight System



Combination of energy and timing

Thin Si detectors (~100 μm) have ns collection times and can provide ps time resolution.

Example

ΔE-detector: 27 μm thick, A= 100 mm^2
$\langle E_{bias} \rangle$ =1.1 ·10^4 V/cm

E-detector: 142 μm thick, A= 100 mm^2
$\langle E_{bias} \rangle$ =2 ·10^4 V/cm

For 230 MeV ^{28}Si:

ΔE = 50 MeV ⇒ V_s= 5.6 mV
E = 180 MeV ⇒ V_s= 106 mV

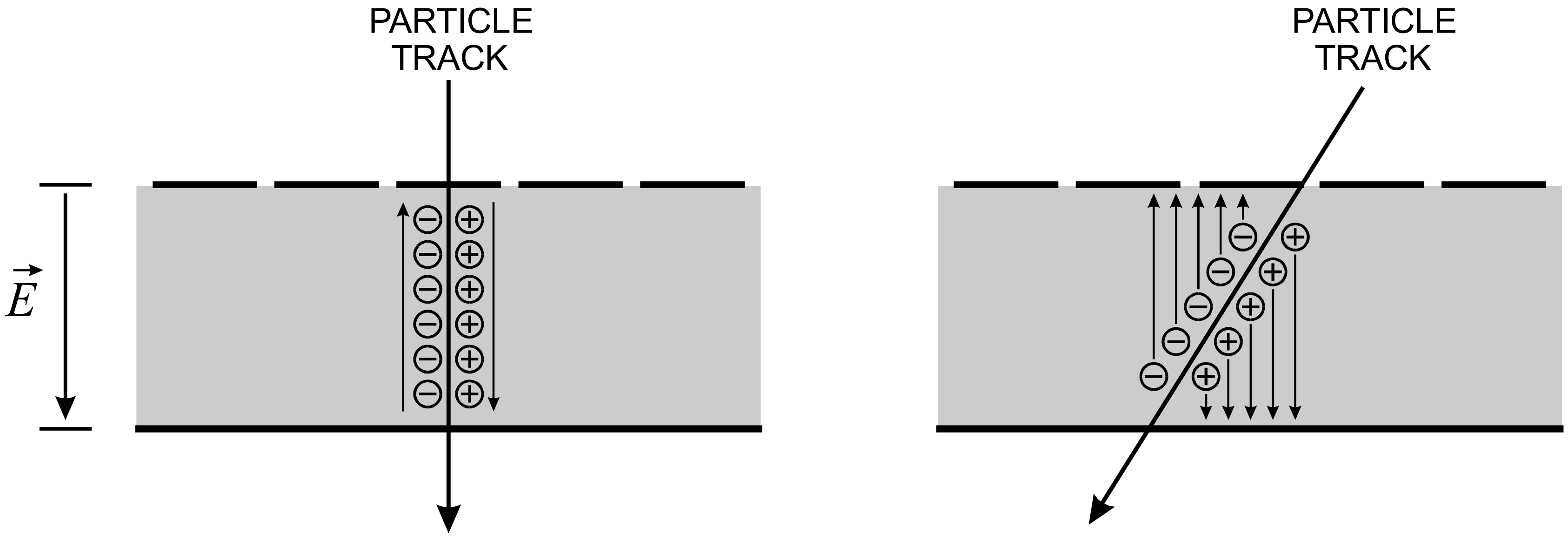
⇒ Δt = 32 ps FWHM

σ_t = 14 ps (both detectors combined) – < 10 ps for E detector

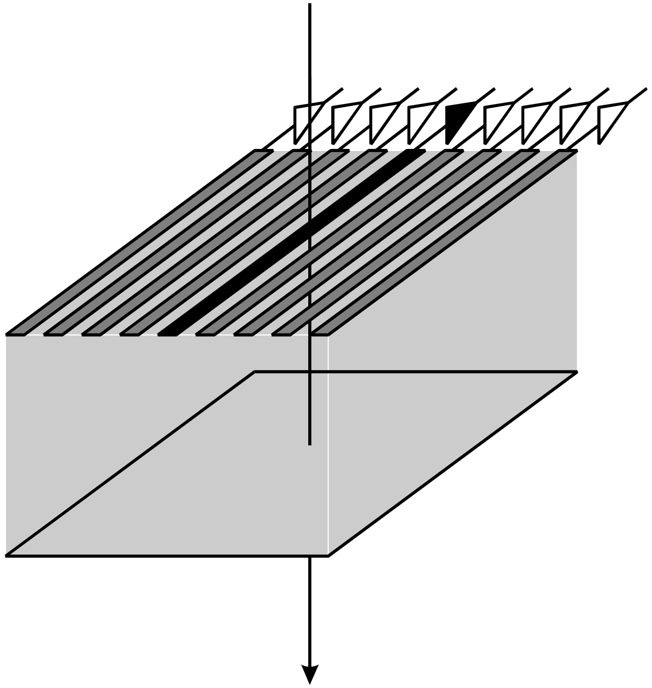
Silicon Detectors Now Yield Position Sensitivity

The electrodes of the sensor can be segmented to provide position information.

Now the magnitude of the signal measured on a given electrode depends on its position relative to the sites of charge formation:



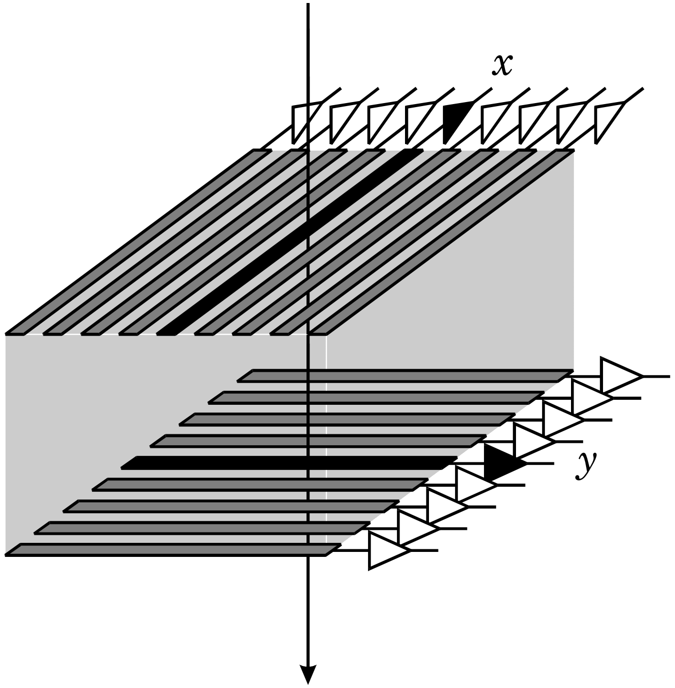
Segmenting one electrode into strips provides position information in one dimension.



Angled tracks will deposit charge on two or more strips.

Evaluating the ratio of charge deposition allows interpolation to provide position resolution better than expected from the electrode pitch alone.

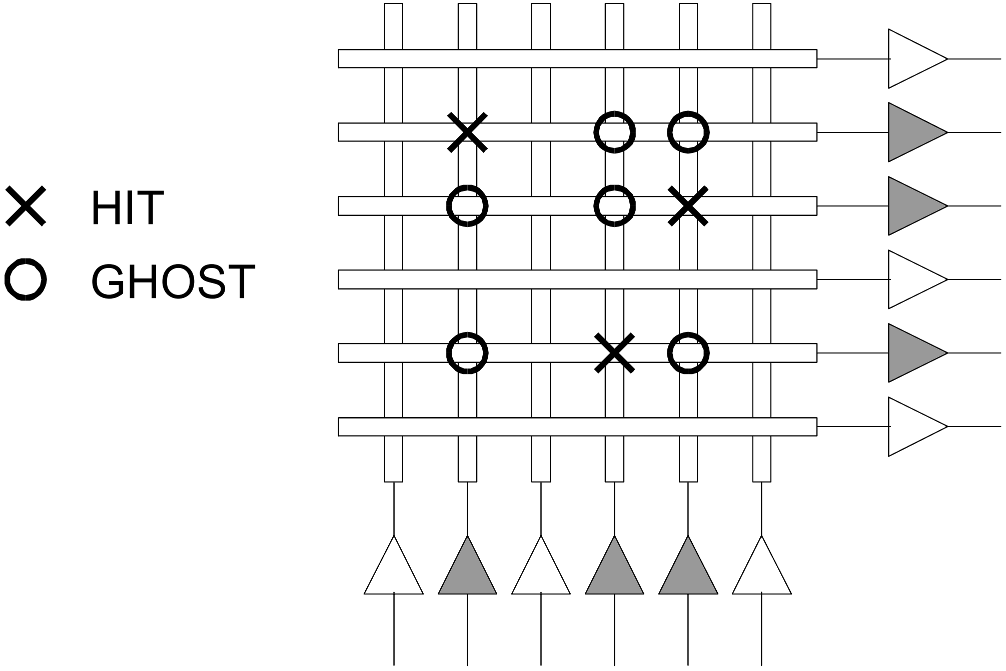
A second orthogonal set of strips on the opposite face gives two-dimensional position readout.



In a colliding-beam experiment the strip pitch (center-to-center distance) is typically 25 – 100 μm and lengths range from centimeters to tens of centimeters, usually aligned parallel to the beam axis to provide $r\varphi$ coordinates.

The maximum strip length per sensor is limited by wafer size, so multiple sensors are ganged to form longer electrodes. Practical detectors have used strips as long as 30 or 40 cm, limited by electronic noise and the hit rate per strip.

Problem: Ambiguities with multiple simultaneous hits (“ghosting”)



n hits in acceptance field $\Rightarrow$ n x-coordinates
n y-coordinates

$\Rightarrow$ n^2 combinations

of which $n^2 - n$ are “ghosts”

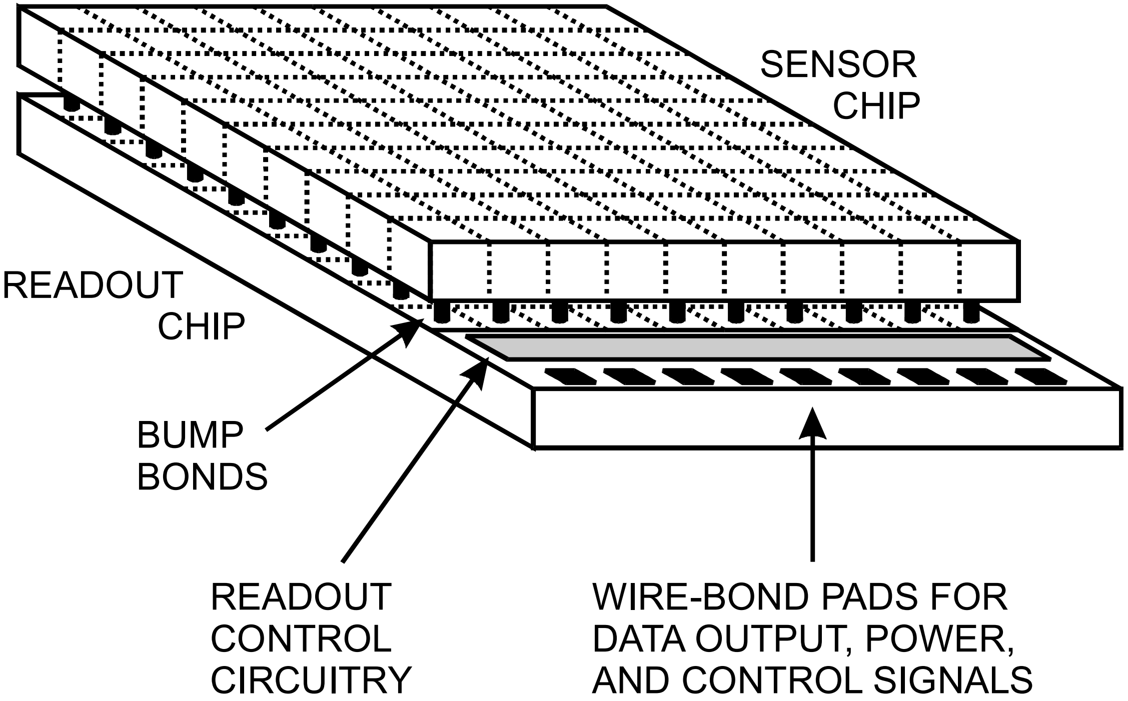
Pixel Devices

To obtain unambiguous two-dimensional information the sensor must provide fine segmentation in both dimensions.

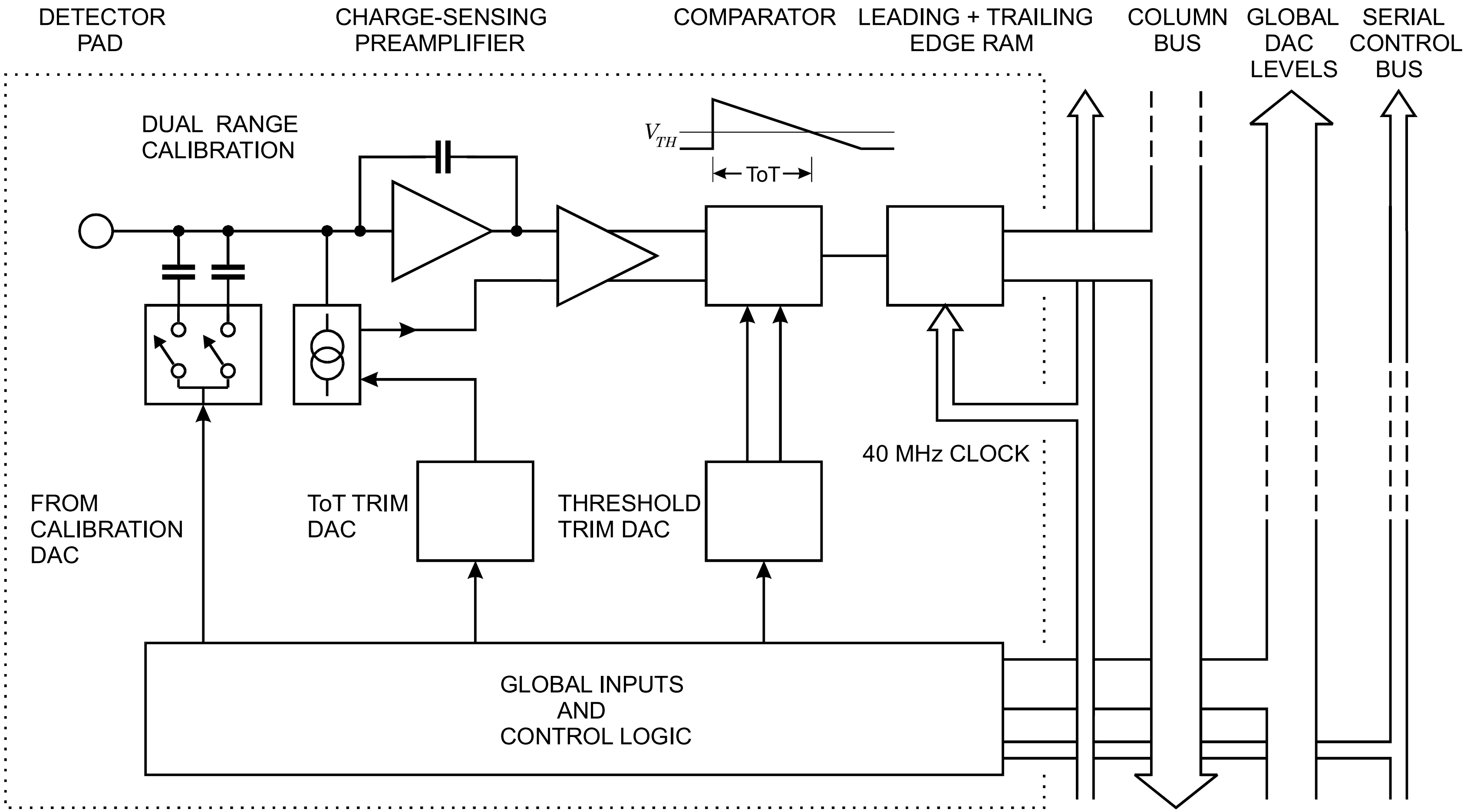
Charge coupled devices (CCDs), random access pixel devices, and silicon drift chambers represent different approaches to obtaining nonprojective two-dimensional information.

The most flexible is the hybrid pixel device:

The sensor electrodes are patterned as a checkerboard and a matching two-dimensional array of readout electronics is connected via a two-dimensional array of contacts, for example solder bumps.



Electronics per Pixel Cell can be quite complex (e.g. ATLAS Pixel Detector)



- After introduction in high-energy physics (LHC), hybrid pixel devices with complex electronic readouts are now used in a variety applications, e.g. high-rate x-ray detection and medical imaging (e.g. Medipix).

- In this scheme the pixel size is limited by the area required by each electronic readout cell.

- Pixel sizes of 30 – 100 μm for multi-component electronics per pixel cell are practical today, depending on the complexity of the circuitry required in each pixel.

- The hybrid figure also shows that the readout IC requires more area than the pixel array to accommodate the readout control and driver circuitry and additional bond pads for the external connections.

- Since multiple readout ICs are needed to cover more than several cm^2, this additional area constrains designs that require full coverage.

- Pixel readouts can yield energy measurement, count rates (threshold), and timing. However, not all can be optimized, but acceptable functions may be practical.

- Implementing this structure monolithically would be a great simplification and some work has proceeded in this direction.

However, all applications build on the basics of semiconductor detectors.

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
>
> … and don’t just follow recipes.

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.

Measurement Contributions

1. Energy

Full charge deposited in Detector
Integration of the detector current pulse
Pulse shaping to reduce electronic noise and pile-up
Resolution determined by
    signal magnitude and
    variations due to detector and electronic noise

2. Timing

Sufficient charge for signal
Preamplifier that accepts required pulse speed
Sufficiently fast electronics
Resolution determined by
    time variations in the detector
    pulse rise-time
    signal magnitude
    electronic noise

3. Position

Detector configuration
Sufficient signal-to-noise ratio

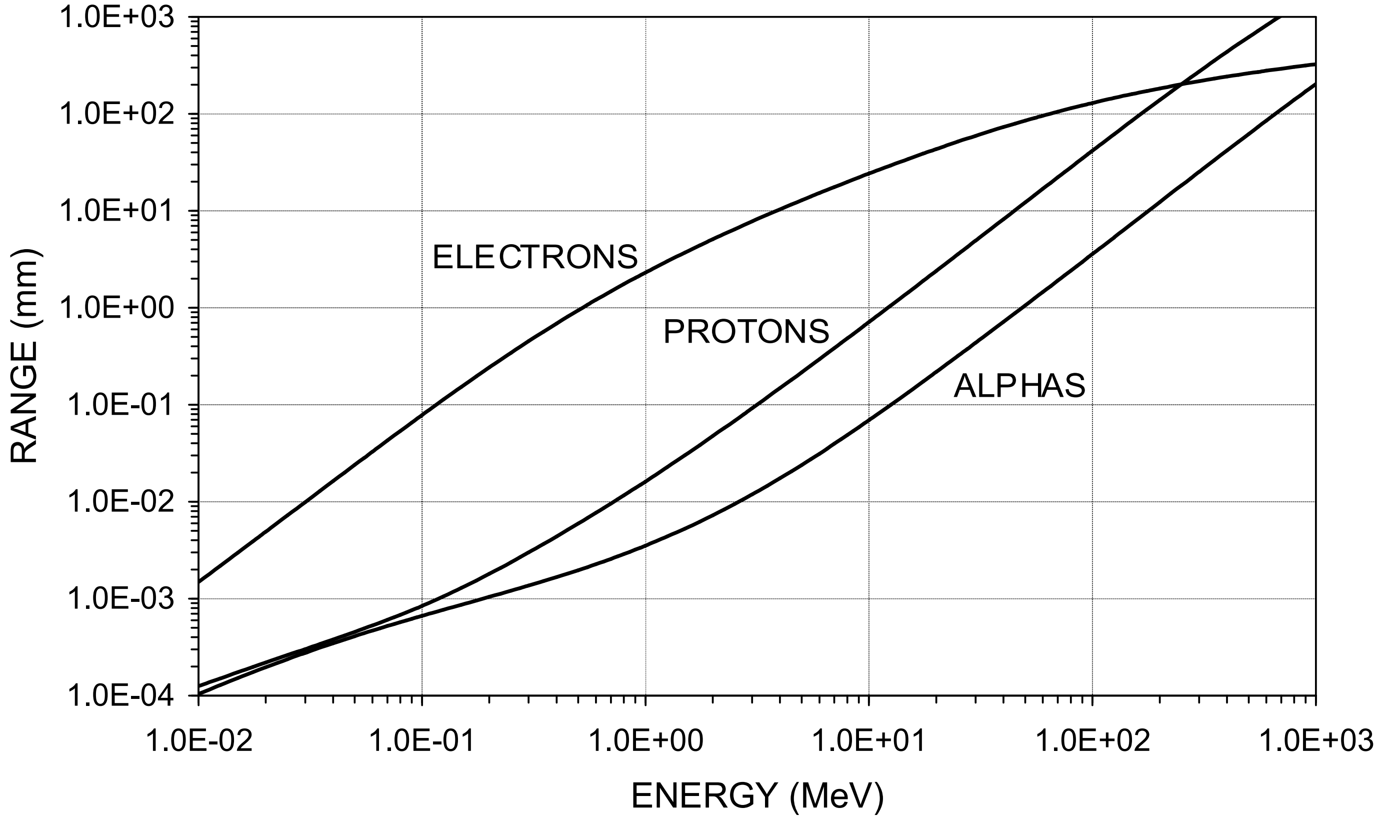
4. Threshold Counts

Sufficient signal-to-noise ratio to maintain negligible noise counts

Detector Signals

1. Charged Particles Deposit Energy Along the Track

Range of Charged Particles in Silicon



1 MeV electrons have a range of about 2 mm (substantial fluctuations due to straggling)
1 MeV alphas have a range of 3.3 μm.

At low energies the range of particles decreases drastically with increasing projectile charge.

For E = 5 MeV in Si:

p	R = 220 μm
α	R = 25 μm
^{16}O	R = 4.3 μm
^{40}Ca	R = 3.0 μm
^{132}Xe	R = 2.0 μm
^{197}Au	R = 1.4 μm

2. Interactions of Gamma Rays

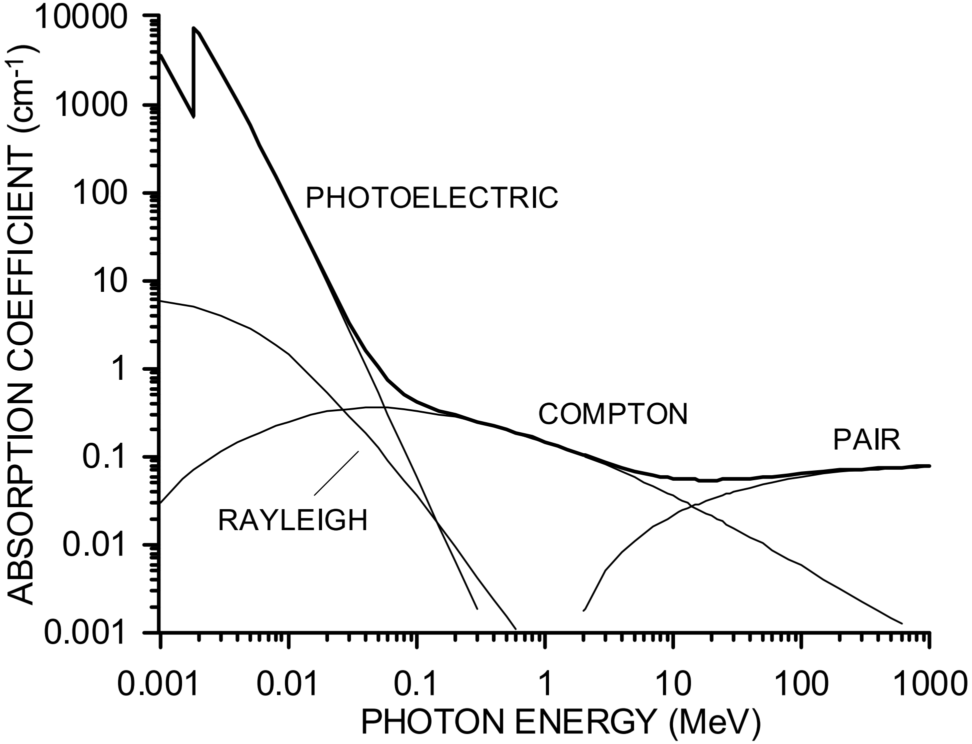
In contrast to charged particles, which deposit energy continuously along their track, photon interactions are localized. **Photons must free mobile electrons for an electric signal.**

In passing through a medium, photons will traverse a certain distance unaffected, until depositing energy either by

a) Photoelectric absorption

b) Compton scattering

c) Pair production



The probability of undergoing an interaction is an exponential function of distance. The fraction of photons that suffered any interaction after traversing a distance x is

$$f = 1 - \exp(-\mu x)$$

where μ is the total absorption coefficient, expressed in cm^{-1} (sum of the individual coefficients)

The inverse value $1/\mu$ gives a direct estimate of the range.

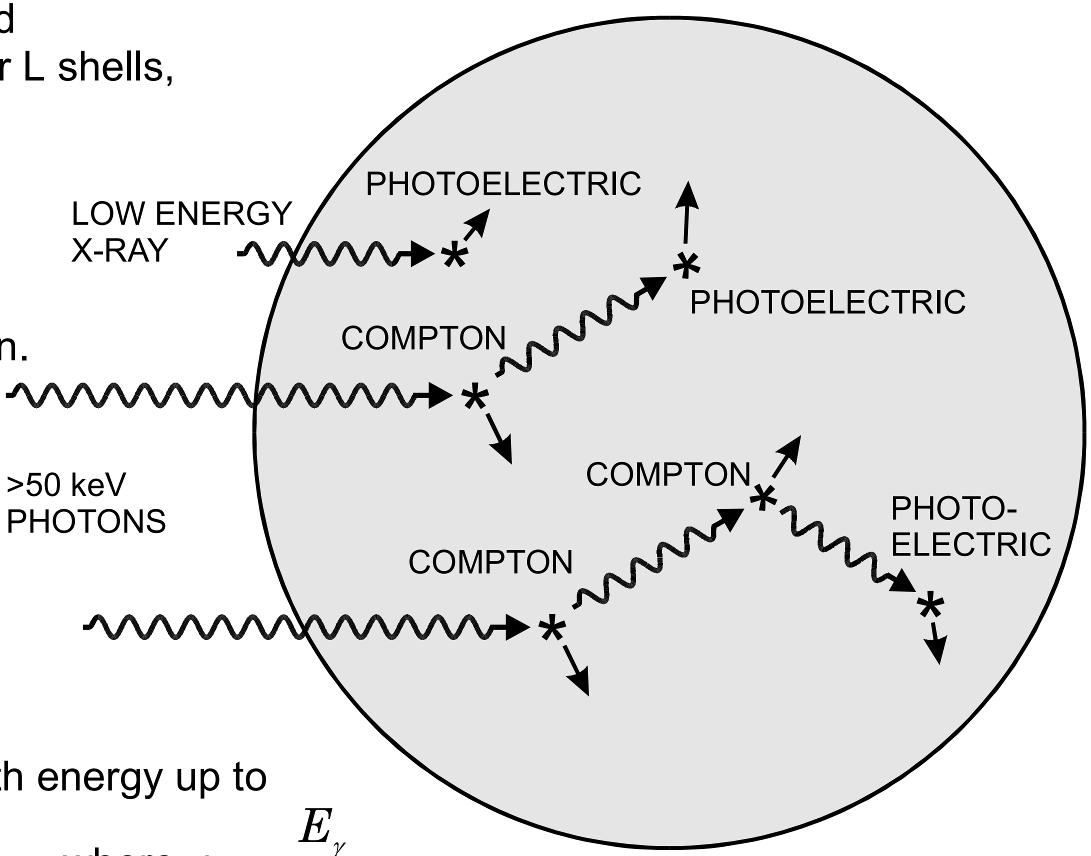
Photon Ranges in Si: 20 keV ~ 5 μm
100 keV ~ 80 μm.

Interactions

Photoelectric Absorption

An electron is emitted with the energy $E_{pe} = E_\gamma - E_b$ E_b is the binding energy of the photo electron.

The cross section for 20 keV and 100 keV is maximum for the K or L shells, for which the binding energies E_b = 1.8 keV and 0.1 keV in Si.



Compton Scattering

Photon scatters off of an electron.

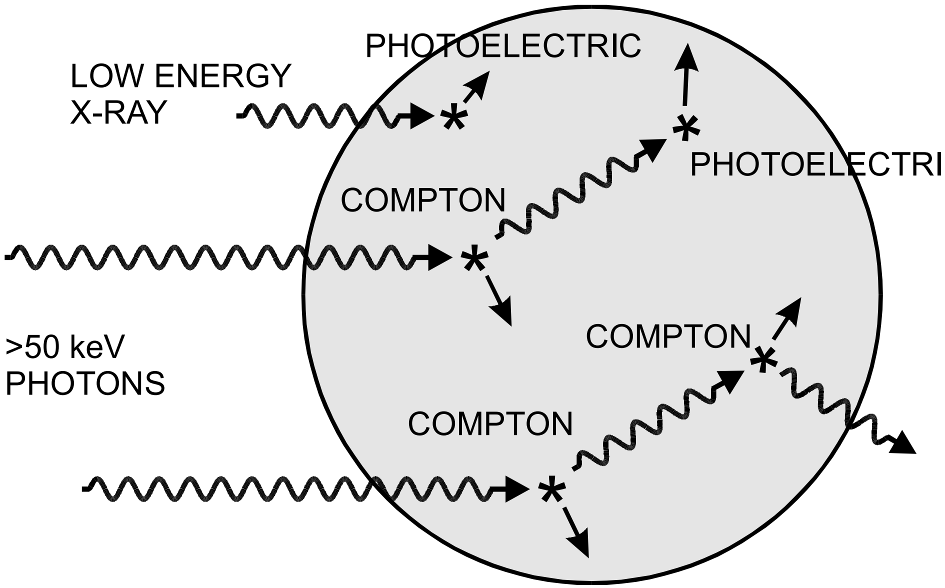
⇒ photon deflected with decreased energy

$$E_\gamma^{Comp} = \frac{E_\gamma}{1 + \frac{E_\gamma}{m_0 c^2}(1 - \cos\Theta)}$$

The recoil electron is emitted with energy up to

$$E_e^{max} = \frac{E_\gamma}{1 + (1/2\alpha)} \qquad \text{where } \alpha \equiv \frac{E_\gamma}{m_0 c^2}$$

However, if the detector volume is too small, the secondary photon from the second Compton scatter will leave the detector.

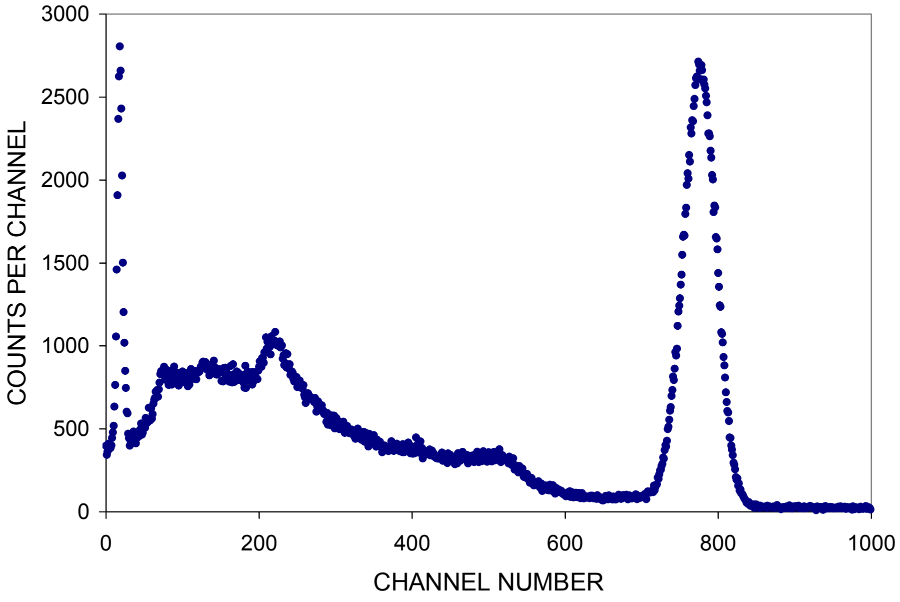


⇒ The detector will only absorb a part of the incident energy.

The resulting signal will be in the Compton continuum.

A typical spectrum from an atomic decay Atom 1 → Atom 2

- the incident photon peak
- the spread-out Compton continuum
- the x-ray from the Atom 2



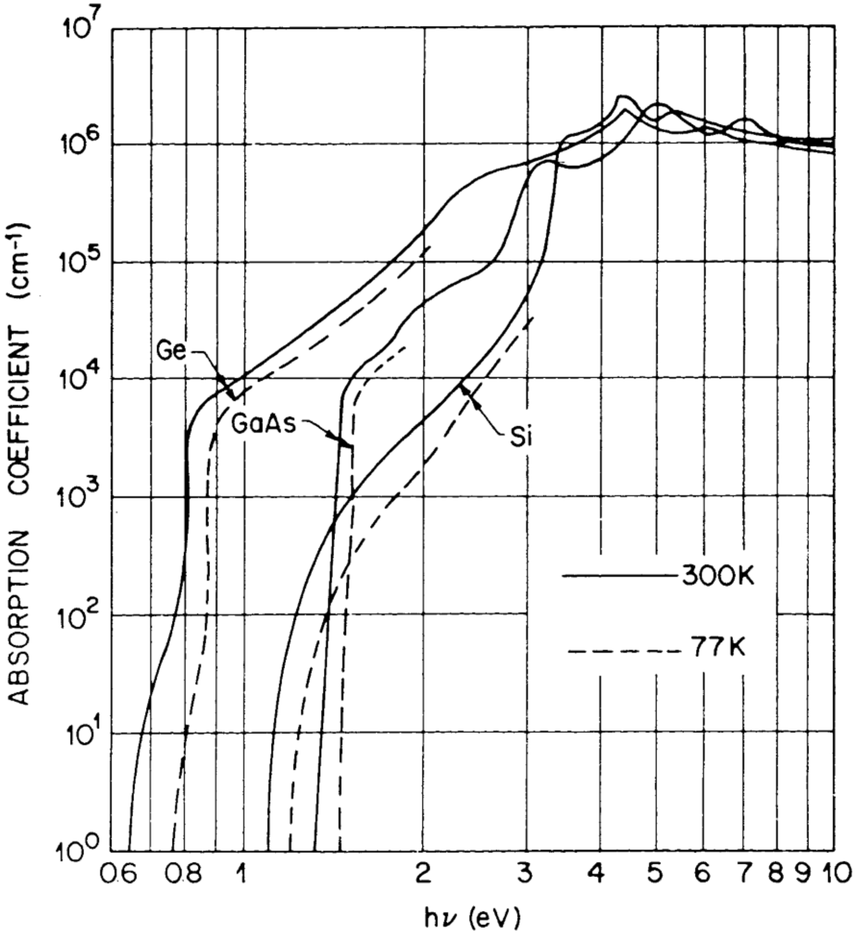
Sensitivity – Conversion of Energy to Signal Charge

a) Visible light
(energies near band gap)

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

Measurements on silicon photodiodes show that for photon energies below 4 eV one electron-hole (*e*-*h*) pair is formed per incident photon.

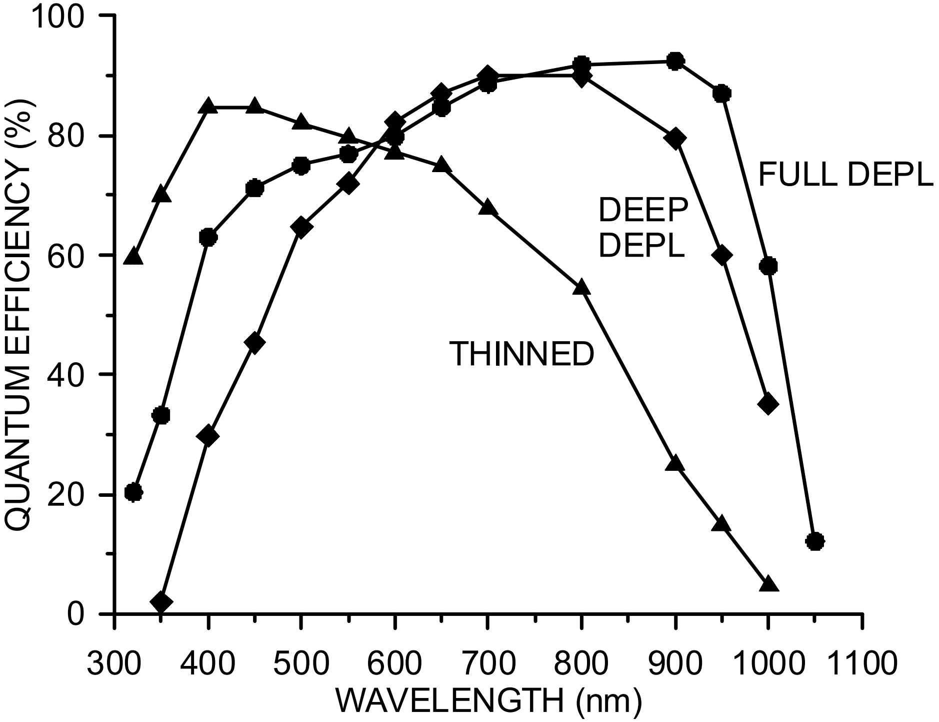
The mean energy E_i required to produce an *e*-*h* pair peaks at 4.4 eV for a photon energy around 6 eV.

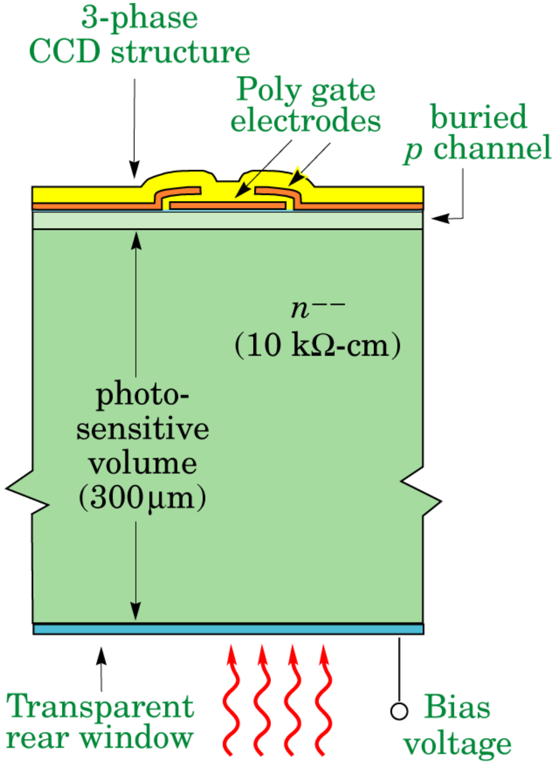


(From Sze 1981, ©Wiley and Sons)

Example: Fully Depleted CCDs
Developed at LBNL

- high resistivity n-type substrate, fully depleted
- backside illumination
- thin backside dead layer
- 300 µm substrate thickness
  - ⇒ 300 µm active thickness
  - ⇒ good QE up to λ= 1 µm





Developed up to
650 µm depletion depth

⇒ ~50% efficiency at 20 keV

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.

Why?

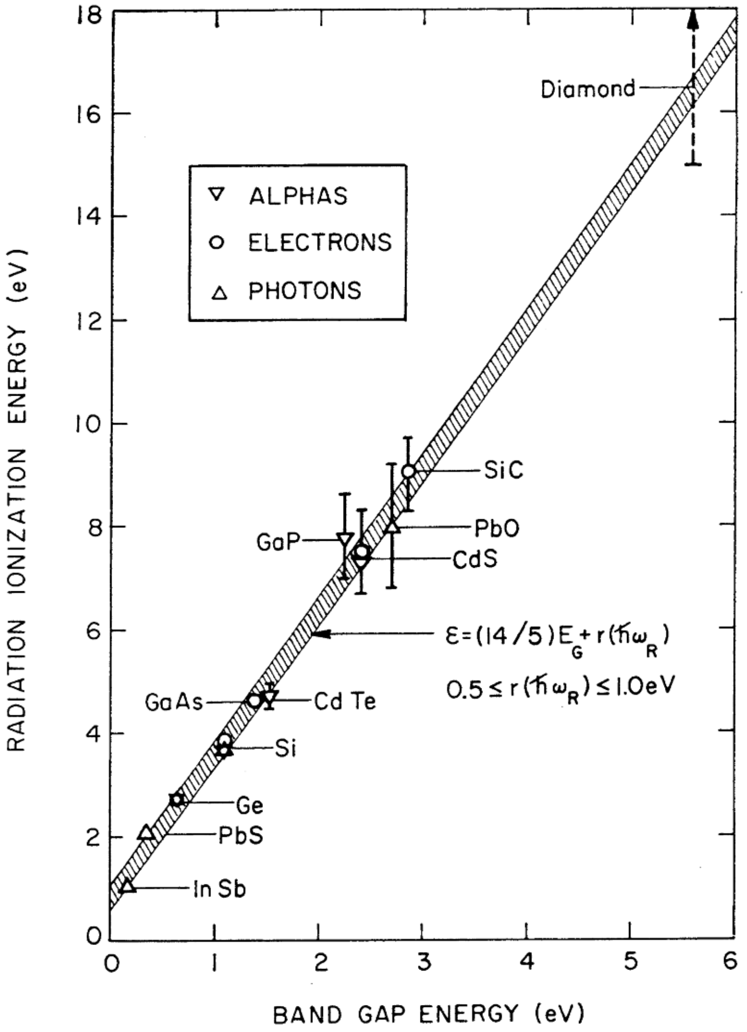
When a particle deposits energy one must conserve both

energy and momentum

momentum conservation is not fulfilled by transition across the gap

⇒ excite phonons

~60% of the deposited energy goes into phonon excitation.



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

To increase sensitivity: Instead of detecting electron-hole pairs, detect heat or phonons

Energy scale: 10 meV ⇒ lower energy threshold

Signal Fluctuations: Intrinsic Resolution of Semiconductor Detectors

The number of charge-pairs: $N_Q = \dfrac{E}{E_i}$

E = particle energy
E_i = energy per electron-hole

The corresponding energy fluctuation: $\Delta E = E_i \sqrt{FN_Q} = E_i \sqrt{F \dfrac{E}{E_i}} = \sqrt{FEE_i}$

F is the Fano factor (Chapter 2, pp 52-55).

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

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

Since the total energy must be conserved,

a) the fluctuation cannot exceed the absorbed energy

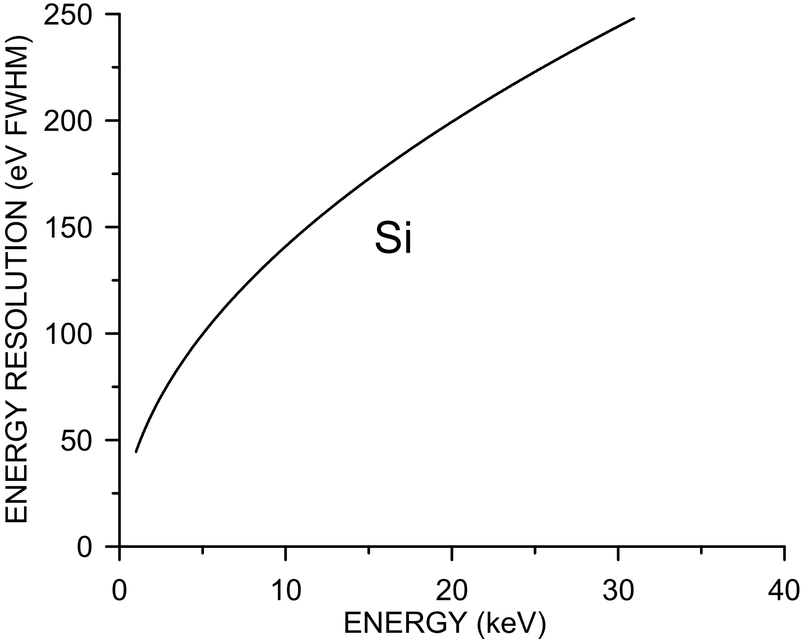
b) any fluctuation in the number of signal charges must be balanced by the fluctuation in the number of phonons. As the number of phonons is much greater, its relative variance is small and this reduces the overall fluctuations.

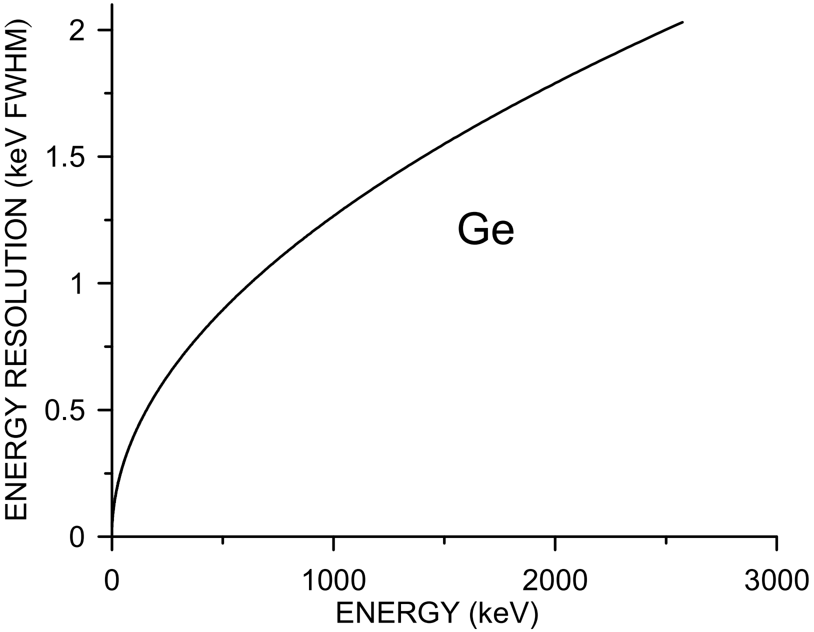
The magnitude of the Fano factor depends on the energy paths that lead to the signal quanta. It often is >1:

In Xe gas $F = 0.15$, but in liquid Xe $F \approx 20$.

Many applicants view Fano as a universal resolution factor from all contributions – wrong!

Inherent Detector Energy Resolution





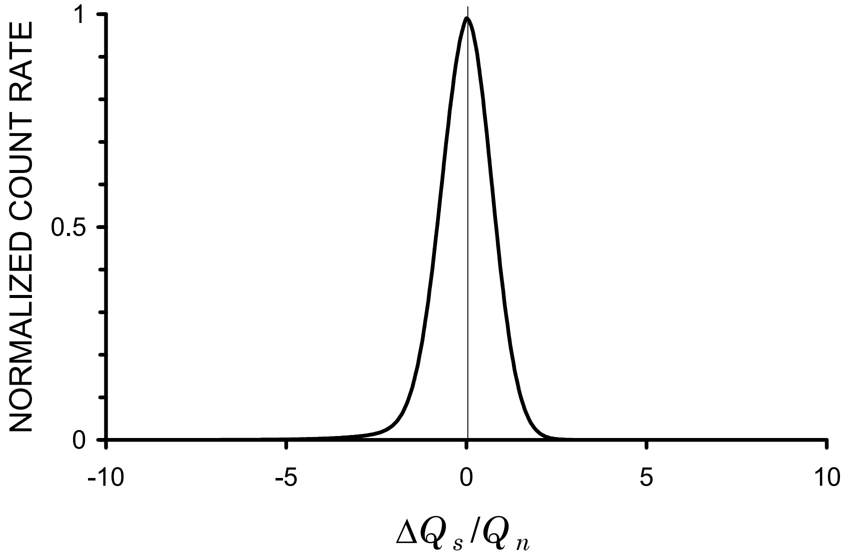
Detectors with good efficiency in the 10s of keV range can have sufficiently small capacitance to allow electronic noise of ~100 eV FWHM, so the variance of the detector signal is a significant contribution.

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

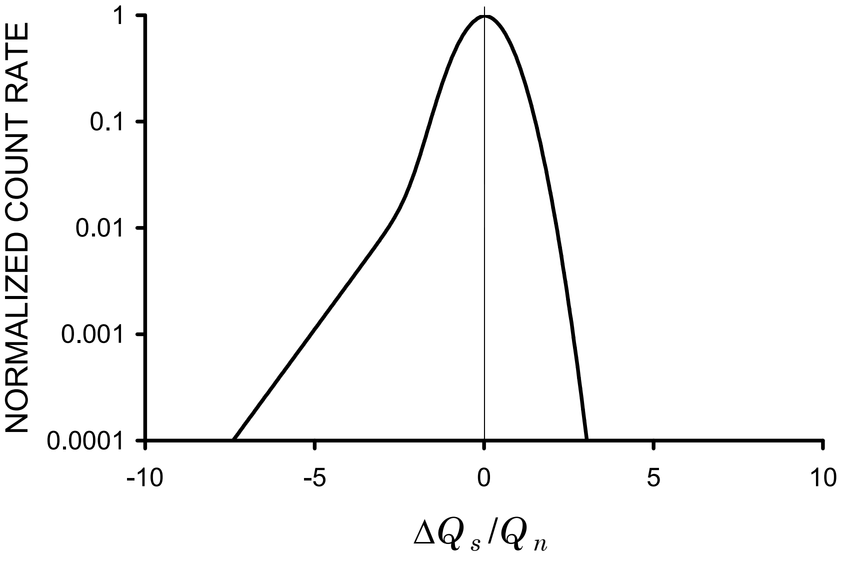
Is this the full resolution?

If not all of the energy is converted to charge, e.g. because of minor traps, a low-energy tail will form.

Linear Plot



Logarithmic Plot

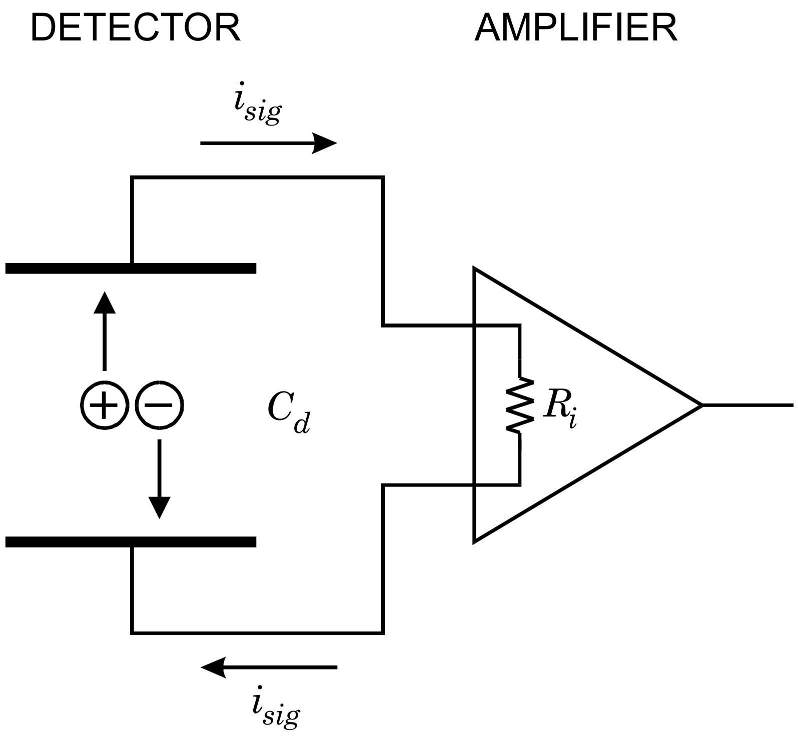


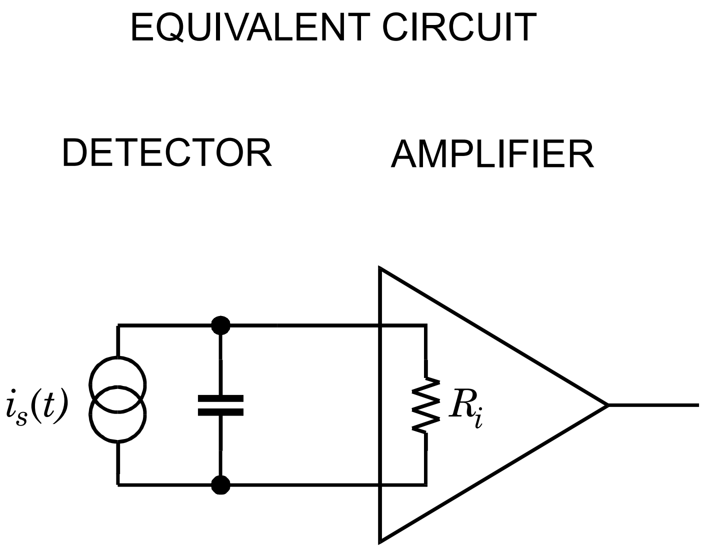
Check a logarithmic plot to determine the full resolution!

Signal Formation

Deposited energy yields freely moved electrons (-Q) and holes (+Q)

Applying an electric field to the detector makes them move in opposite directions to yield a current of the same polarity.

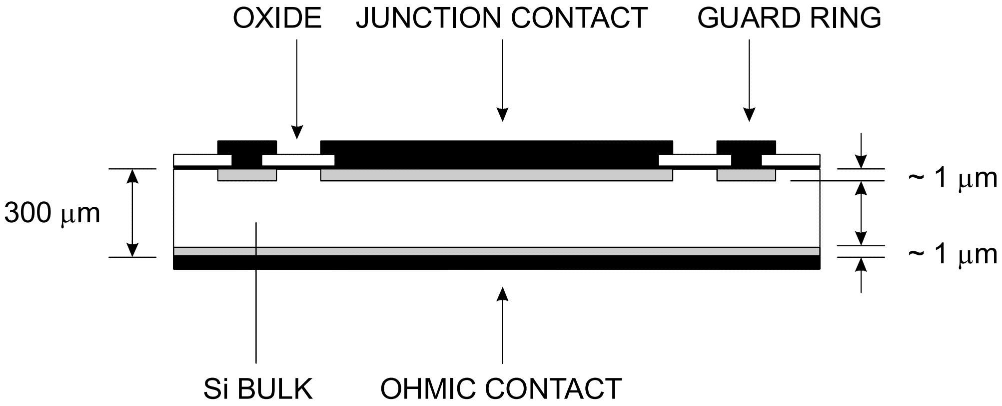




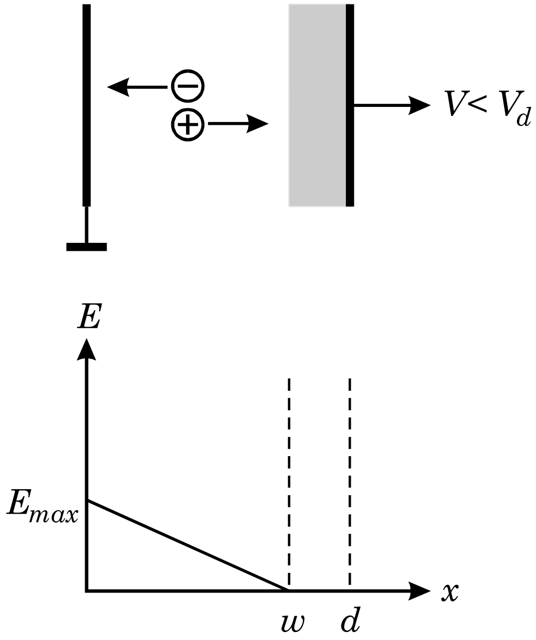
A Typical Silicon Detector Configuration

The p-n junction is asymmetric, i.e. one side is much more highly doped than the other.

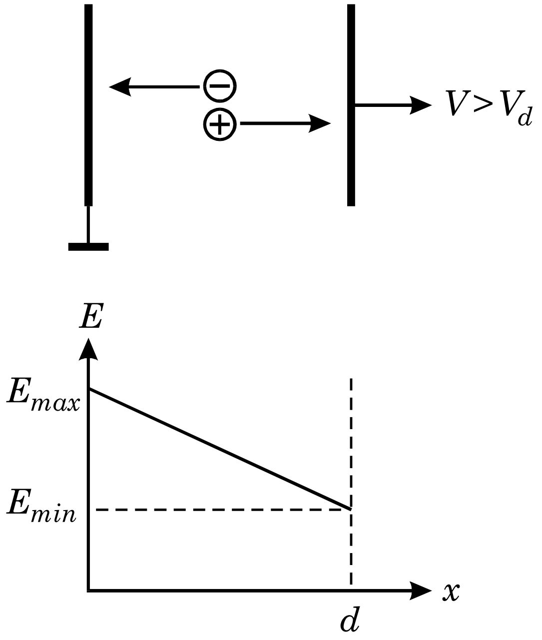
The detector volume then extends into the lightly doped portion.



Partial Depletion ⇒ zero field at ohmic side



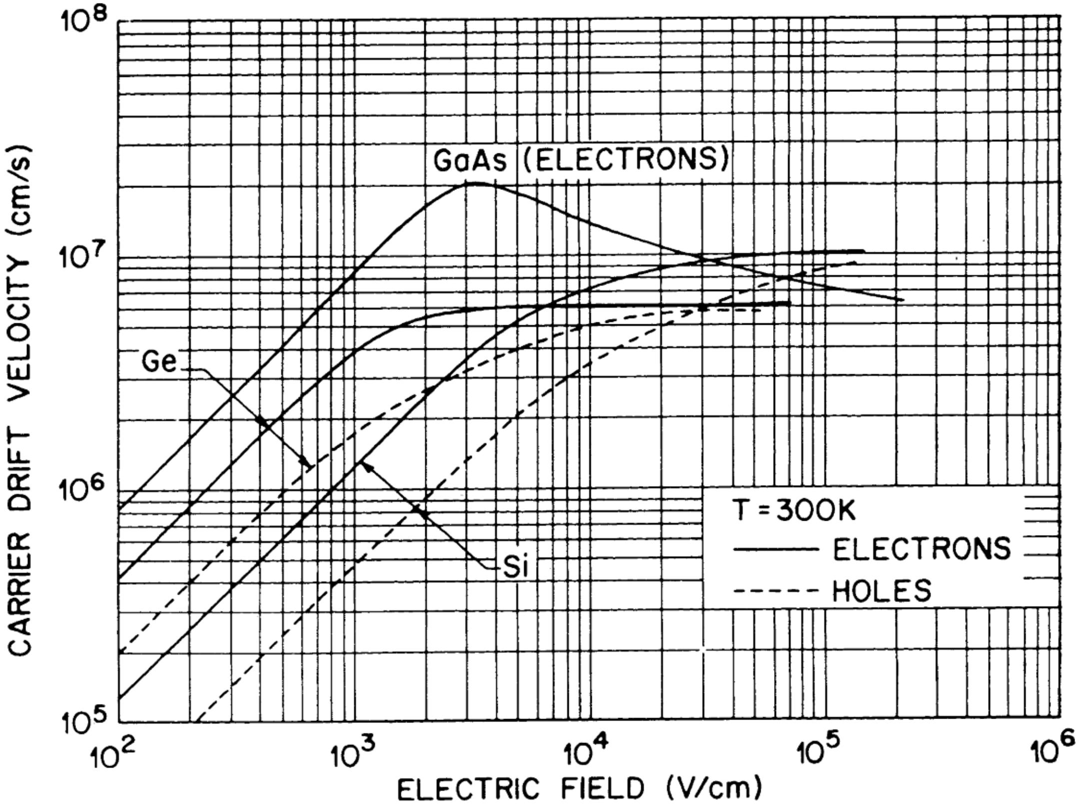
Over-Depletion ⇒ high field throughout



The velocity of the charges determines the instantaneous current level.

In Si at 300K the mobility μ at low fields is

- 1350 cm^2/ Vs for electrons
- 480 cm^2/ Vs for holes.



(From Sze 1981, ©Wiley and Sons, reproduced with permission)

The mobility is constant up to about 10^4 V/cm, but then increased phonon emission reduces the energy going into electron motion, so the mobility decreases.

At high fields $E > 10^5$ V/cm the mobility $\mu \propto \frac{1}{E}$ and carriers attain a constant drift velocity of 10^7 cm/s.

At fields $E > 10^5$ V/cm electrons can gain enough energy to excite additional electron-hole pairs.

This can lead to breakdown, but when well-chosen, it can also yield inherent gain.

However, for visible light photodiodes yield only one electron-hole pair per incident photon, so signals are small.

Photomultiplier tubes provide high gain without introducing significant electronic noise, whereas photodiode systems depend critically on low noise electronics.

Unlike PMT systems, photodiode readouts must be very carefully optimized.

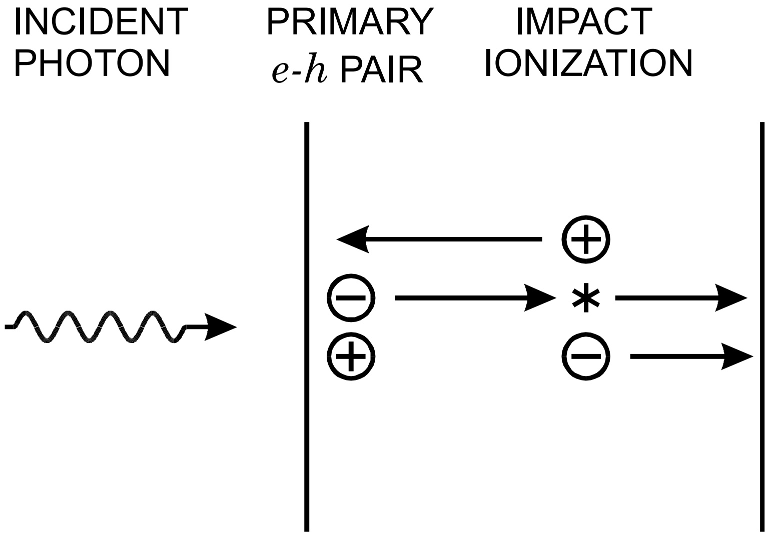
⇒ Reduce demands on electronics by developing photodiodes with internal gain,

- Avalanche Photodiodes (APDs).

Principle of an Avalanche Photodiode

An electron-hole pair is created at the left-most electrode by incident light.

Under the influence of the electric field the electron drifts towards the right, gaining sufficient energy for ionization, i.e. formation of an additional electron-hole pair.



The gain of this process

$$G_n = e^{\alpha_n d}$$

where the electron ionization coefficient

$$\alpha_n = \alpha_{n0} \exp(-E_n / E)$$

is a function of the electric field E. The parameters α_{n0} and E_n are material constants.

The ionization coefficient is also strongly temperature dependent.

With increasing fields the probability of the slower holes getting enough energy to add additional gain leads to breakdown

This leads to the following limits of gain and detector thickness vs. electric field

$E = 2\cdot 10^{5}$ V/cm $\quad G_n = 2.2\cdot 10^{3}$ $\quad d = 520\ \mu$m $\quad V_b = 10$ kV

$E = 3\cdot 10^{5}$ V/cm $\quad G_n = 50$ $\quad d = 5\ \mu$m $\quad V_b = 150$ V

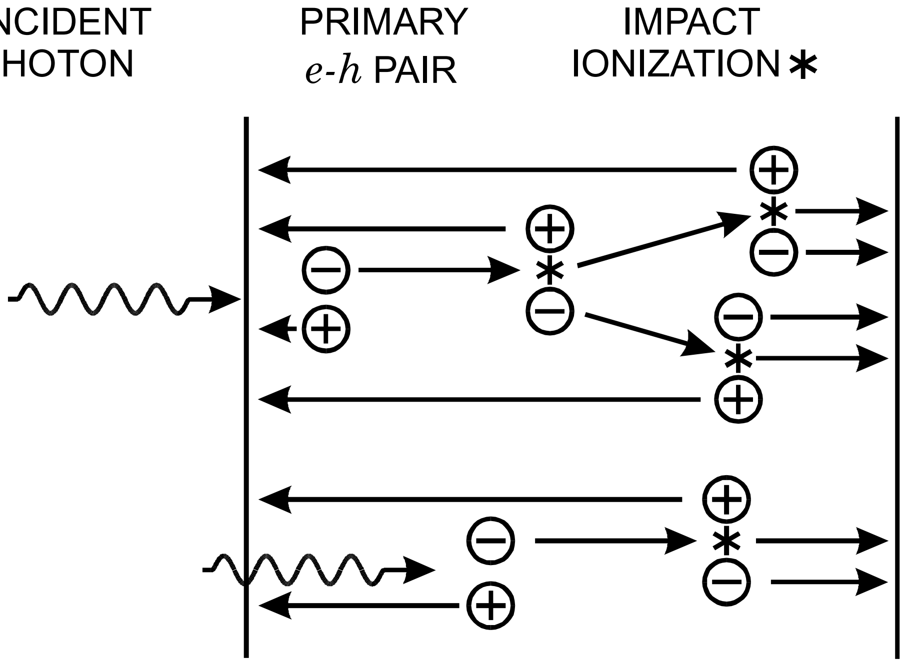
$E = 4\cdot 10^{5}$ V/cm $\quad G_n = 6.5$ $\quad d = 0.5\ \mu$m $\quad V_b = 20$ V

$E = 5\cdot 10^{5}$ V/cm $\quad G_n = 2.8$ $\quad d = 0.1\ \mu$m $\quad V_b = 5$ V

To achieve gains in the range 100 – 1000 requires

- a depletion region of several hundred microns thick
- bias voltages in the range 500 – 1000 V
- excellent control of the field distribution
  provide stable operation without local breakdown
  reduce avalanche noise

In the single field volume configuration the gain depends on where the initial electron-hole pair is formed.

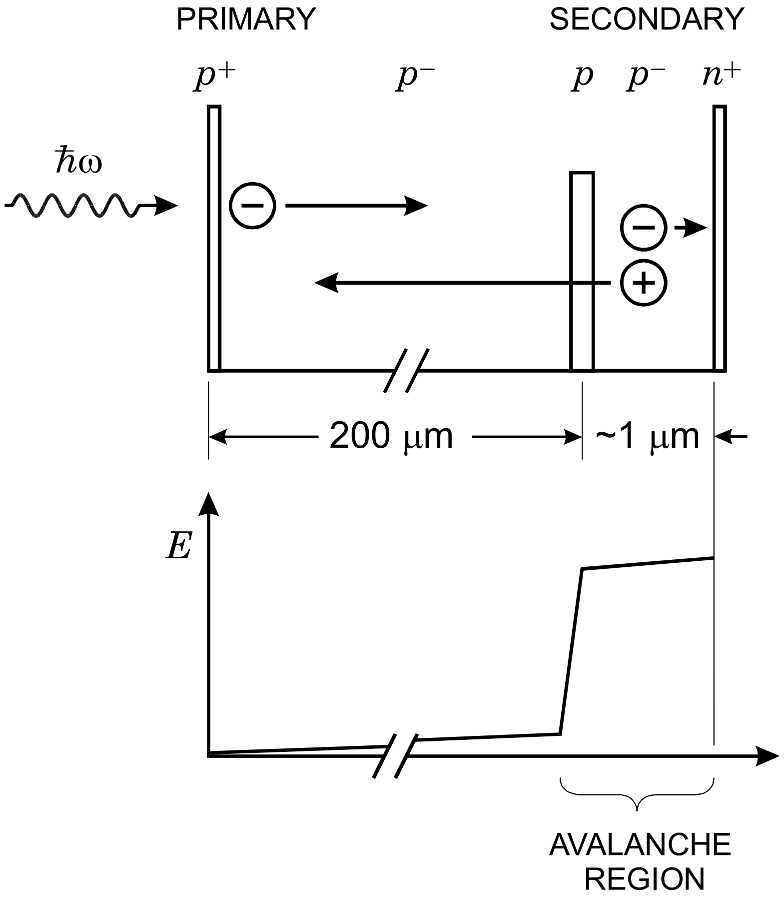


In addition to the statistical variation of the gain, this can form significant changes in the signal for a given photon energy.

However, for timing this is not so critical, as the rise of the pulse is only varied by the light speed of the incident photons.

To optimize energy resolution, a different configuration is advantageous,
the "reach-through" APD.

"Reach-Through" APD.



Lightly doped p-type material is used for the bulk.

A local high-field region is created by introducing an intermediate p-layer through deep diffusion.

When a depletion voltage is applied, the diode depletes from the right-hand side. Initially the depletion region progresses with voltage until the intermediate p-layer is reached. Since this layer is more highly doped, the voltage required to deplete the intermediate layer is rather high.
As a result, a high field is set up in the region between the junction and the p-layer.

Depletion beyond the p-layer requires less voltage, due to low doping.

Photons impinge on the left surface.
Electrons drift towards the high field region, where they avalanche.

Secondary holes drift through the low-field region, contributing most of the induced signal.

The advantage of this structure is that the primary holes remain in the low-field region. Secondary holes drift into the low-field region, thus reducing the hole partial gain and the risk of breakdown.

Silicon Photomultipliers (SiPM)

At high gains APDs go into a sustained avalanche mode.

This can be triggered by an incident photon. Typical gain ~10^6.

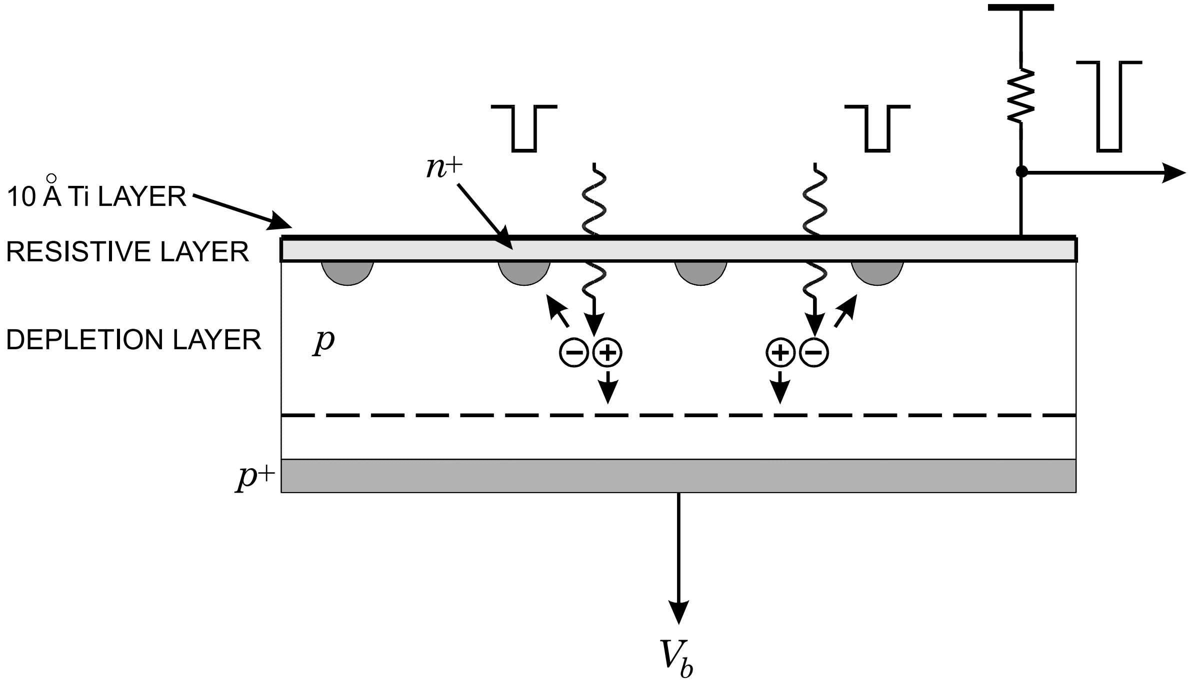
If the current and time duration of the sustained avalanche are limited, the diode does not suffer damage.

Inserting a sufficiently resistance into the bias “quenches” the avalanche, as the momentary high current increases the voltage drop and reduces the diode bias to a stable level (analogous to Geiger mode).

This yields a short current pulse of uniform magnitude for each incident photon.

However, in a single sensor all intensity information of the incident scintillation light is lost.

The silicon photomultiplier subdivides the APD into many small pixels (~50 μm), so that individual pixels are struck by only one scintillation photon.



Summing the current pulses from all pixels ⇒ signal proportional to the number of photons.

Advantage:	single photon sensitivity, fast response (~100 ns)
Downside:	electrons due to diode reverse bias current (thermal excitation) initiate avalanches, so dark current rates are ~10^5 s^{-1}.

In experiments with external triggers or coincidence conditions the dark counts can be suppressed.

Detector Charge Collection – 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.

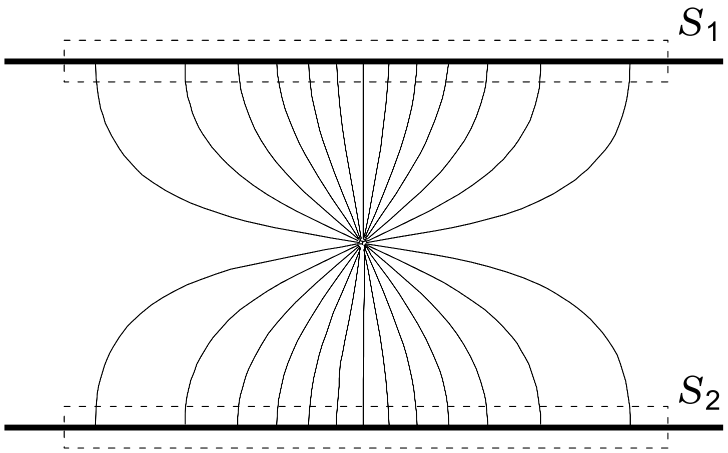
The following discussion applies to ALL types of structures that register the effect of charges moving in an ensemble of electrodes, i.e. not just semiconductor or gas-filled ionization chambers, but also resistors, capacitors, photoconductors, vacuum tubes, etc.

The effect of the amplifier on the signal pulse will be discussed in the following section.

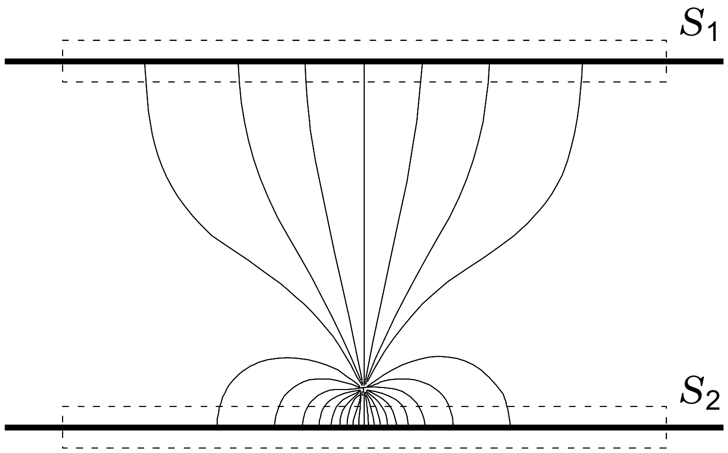
Induced Charge

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).



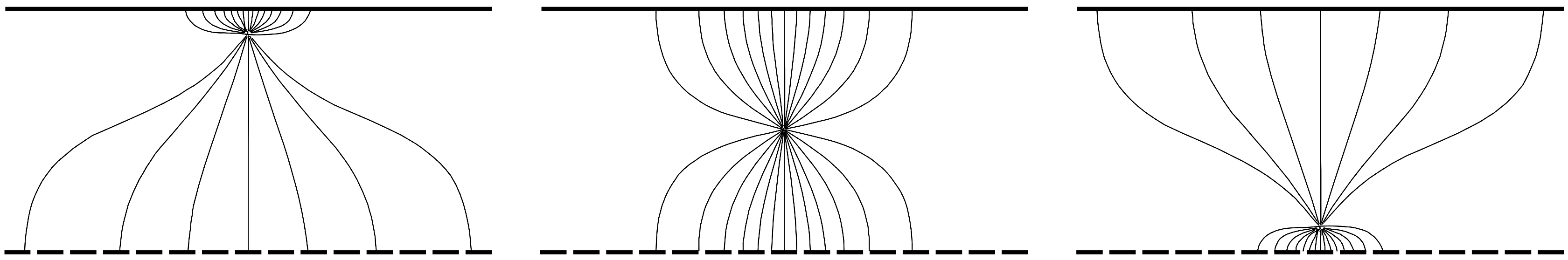
When the charge is close to one plate, most of the field lines terminate on that plate and the induced charge is much greater.



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

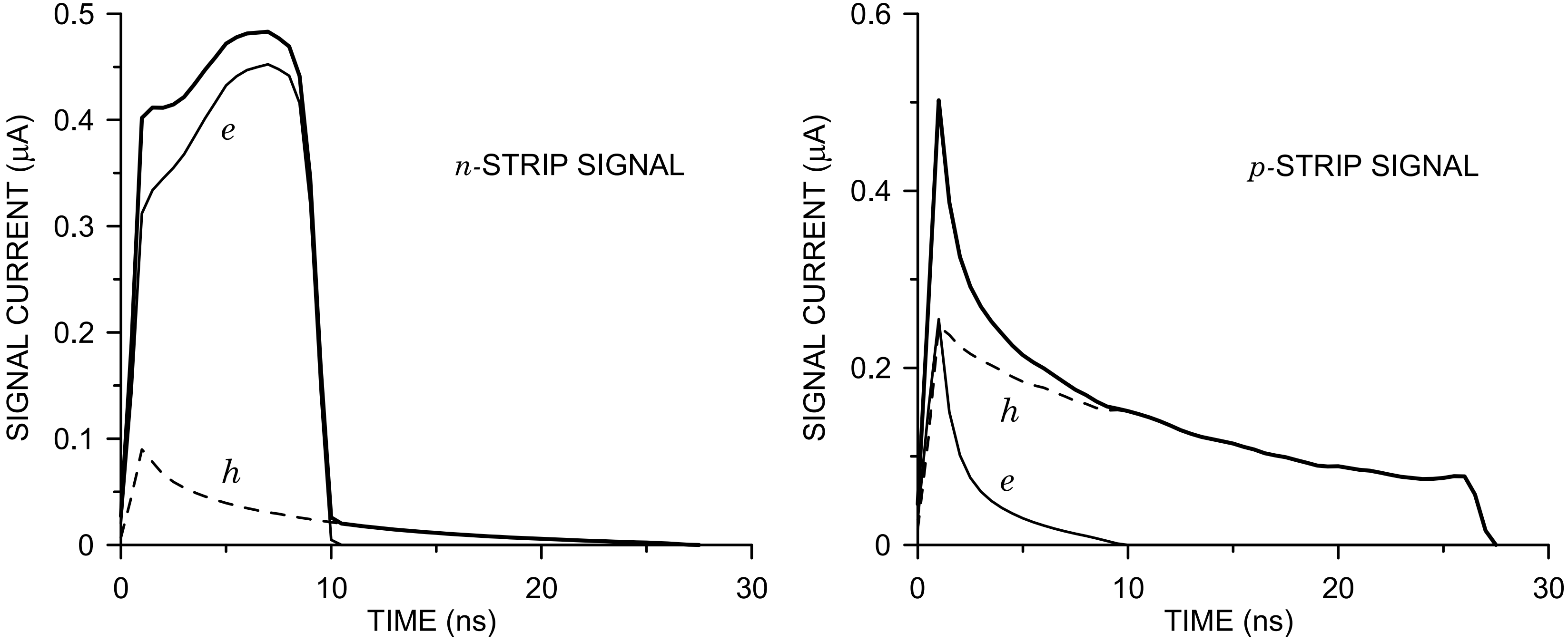
The magnitude of the induced current due to the moving charge depends on the coupling between the charge and the individual electrodes.

Mathematically this can be analyzed conveniently by applying Ramo’s theorem.

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

A common fallacy bases induced charge on energy balance, where it is claimed that the energy gained by the particle in traversing the sensor equals the change in potential on the capacitor plates. Energy distribution is more complex.

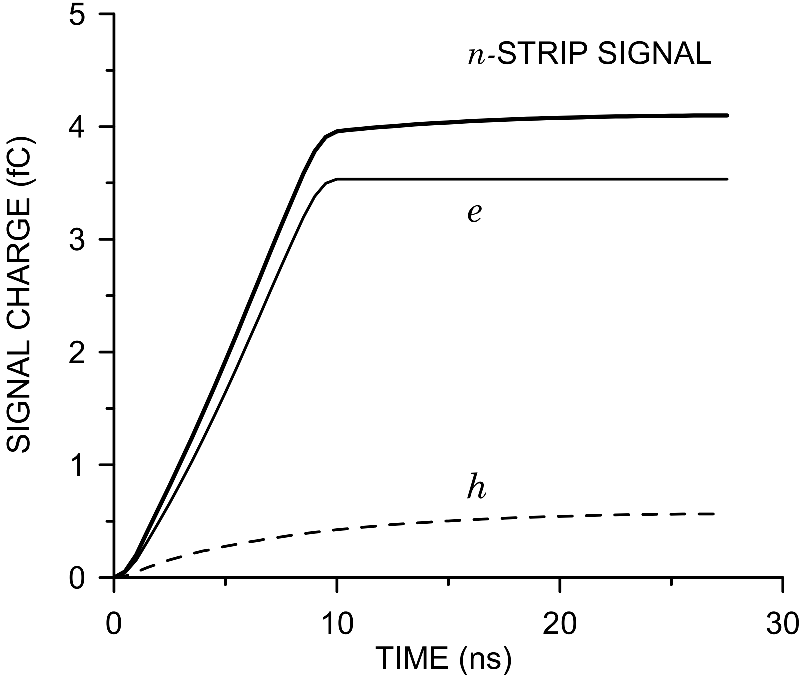
Current pulses in strip detectors (track traversing the detector)

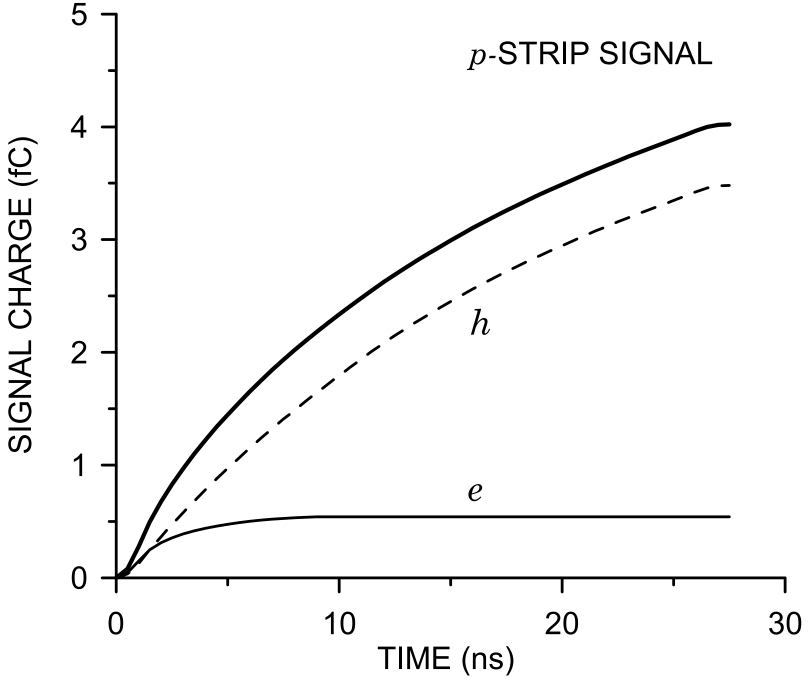


Depletion voltage= 60V   Bias voltage= 90V

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



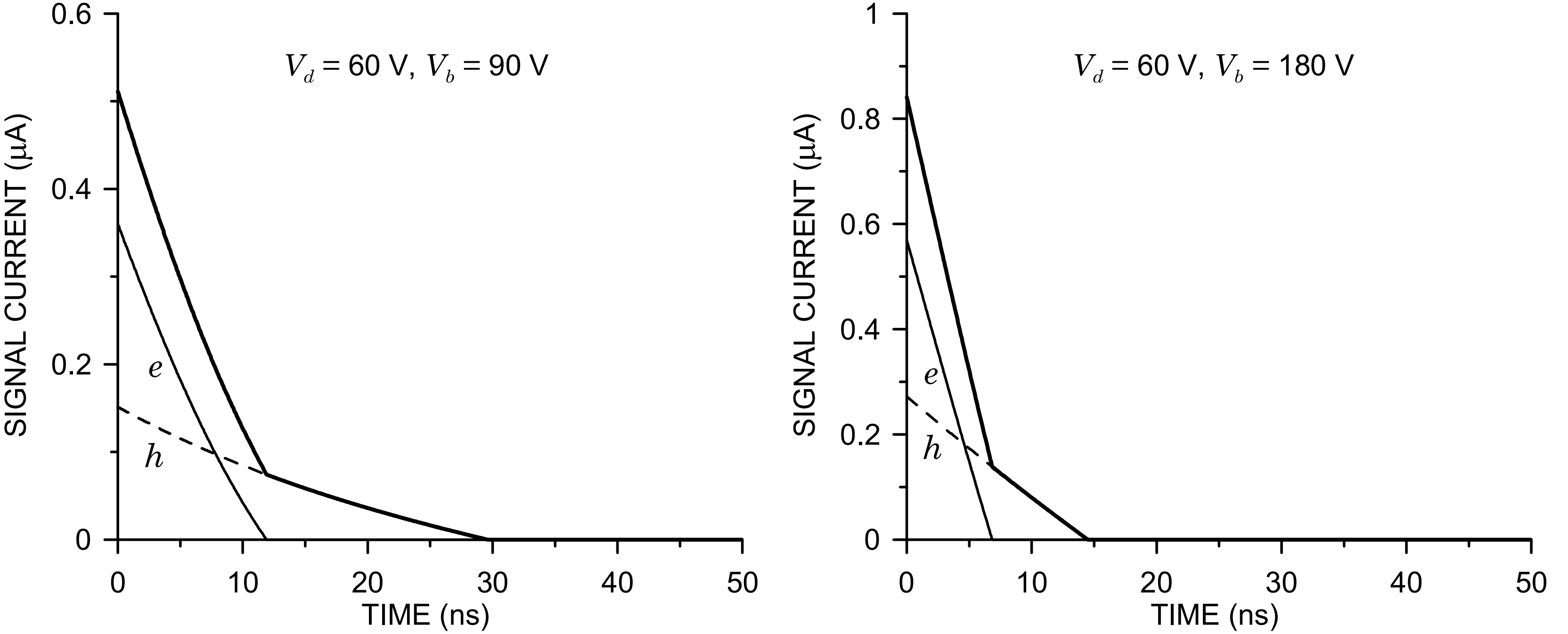


In both electrodes the induced current must be integrated over the full collection time to optimize energy resolution.

For comparison:

Current pulses in pad detectors (track traversing the detector)

Only one electrode on each side:
Same current shape and magnitude on both electrodes – just opposite polarity

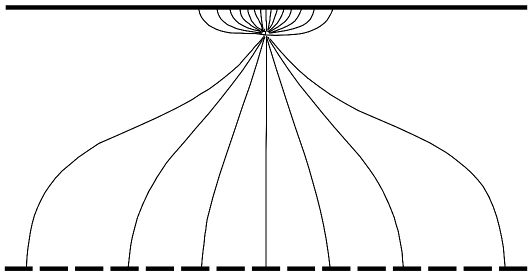


For the same depletion and bias voltages the pulse durations are the same as in strip detectors, although the shapes are very different.

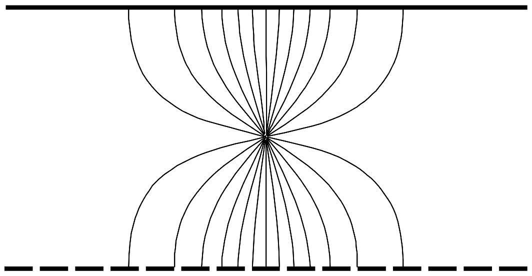
Overbias decreases the collection time.

Varying time delays in signal amplitude can affect timing measurements.

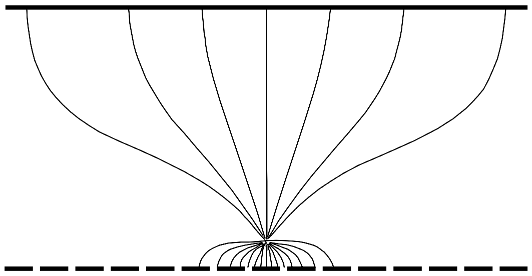
The coupling of the charge increases greatly when the charge comes close to the electrode in strip or pixel detectors.



Initially, charge is induced over many strips.

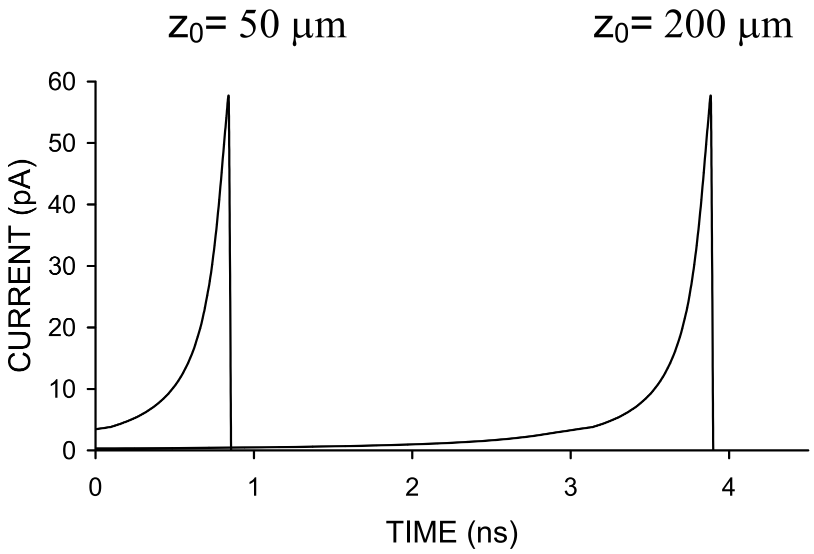


As the charge approaches the strips, the signal distributes over fewer strips.



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.



Semiconductor Materials

Material	E_g (eV)	E_i (eV)	ε	μ_e	μ_h	$(\mu\tau)_e$	$(\mu\tau)_h$	ρ	$\langle Z \rangle$
Si	1.12	3.6	11.7	1350	450	>1	>1	2.33	14
Ge	0.67	2.96	16	3900	1900	>1	>1	5.33	32
GaAs	1.43	4.2	12.8	8000	400	$8\cdot10^{-5}$	$4\cdot10^{-6}$	5.32	31.5
Diamond	5.5	13	5.7	1800	1200	*	*	3.52	6
4H-SiC	3.26	8	9.7	1000	115	$4\cdot10^{-4}$	$8\cdot10^{-5}$	3.21	10
GaN	3.39	8 – 10		1000	30			6.15	19
InP	1.35	4.2	12.4	4600	150	$5\cdot10^{-6}$	$<10^{-5}$	4.78	32
CdTe	1.44	4.43	10.9	1100	100	$3\cdot10^{-3}$	$2\cdot10^{-4}$	5.85	50
$Cd_{0.9}Zn_{0.1}Te$	1.572	4.64	10	1000	120	$4\cdot10^{-3}$	$1.2\cdot10^{-4}$	5.78	49.1
HgI2	2.15	4.2	8.8	100	4	$3\cdot10^{-4}$	$4\cdot10^{-5}$	6.4	62
TlBr	2.68	6.5	30	30	4	$5\cdot10^{-4}$	$2\cdot10^{-5}$	7.56	58
a-Si	1.9	6	12	1 – 4	0.05	$2\cdot10^{-7}$	$3\cdot10^{-8}$	2.3	14

* In diamond the maximum drift length is typically specified. Typically grown by thin-film deposition, material quality depends on the growth rate, with 200 μm drift length obtained for optimal growth.

Higher Z materials would provide higher absorption, but typically suffer from limited carrier lifetime. ⇒ This often leads to incomplete charge collection, typically for holes

Si and Ge provide the best overall properties for precision spectroscopy.

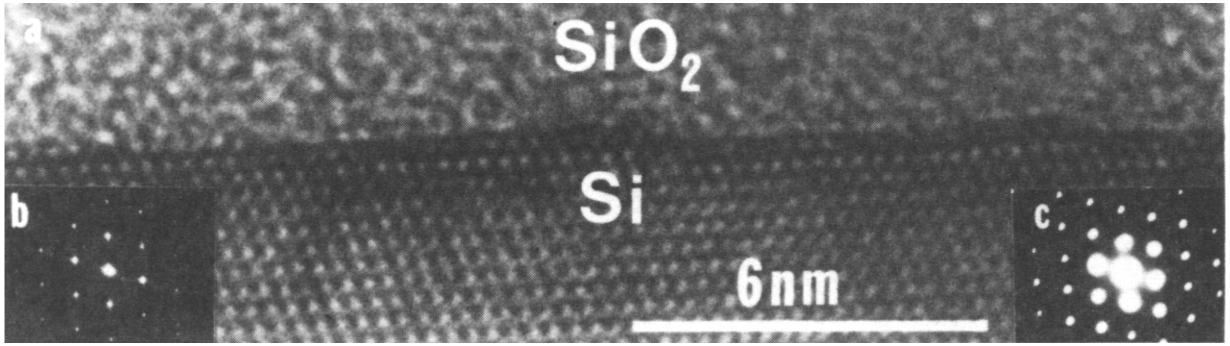
In strip and pixel detectors the individual electrodes must be resistively isolated from one another.

The silicon surface must be covered by a layer that established a well-controlled termination for the “dangling bonds” where the crystal lattice is truncated.

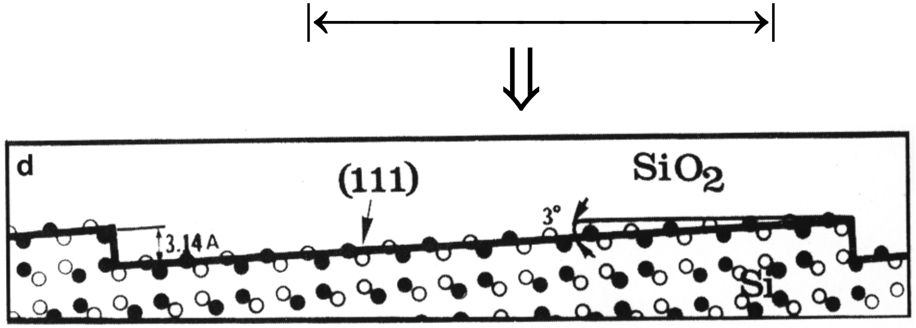
If the interelectrode-oxide does not connect to the silicon with a matching lattice formation, localized carriers can increase resistive coupling.

Furthermore, the thermal coefficient of expansion must be well matched to silicon
⇒ Thermally grown silicon dioxide has proven to be ideal for this purpose.

Atomic resolution electron microscope image of SiO_2-Si interface



Si is unique in this respect.

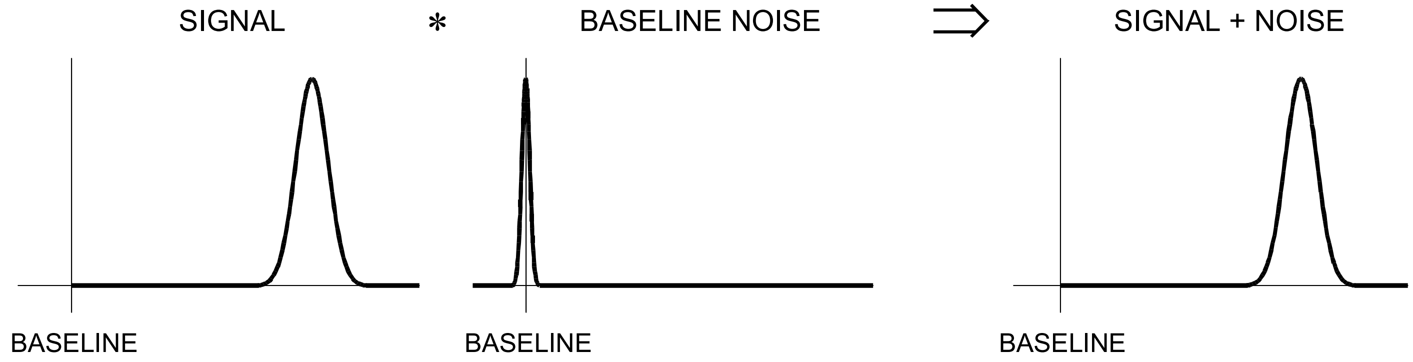


Currently, for many applications Si is still the material of choice.

(Gronsky et al., LBNL National Center for Electron Spectroscopy)

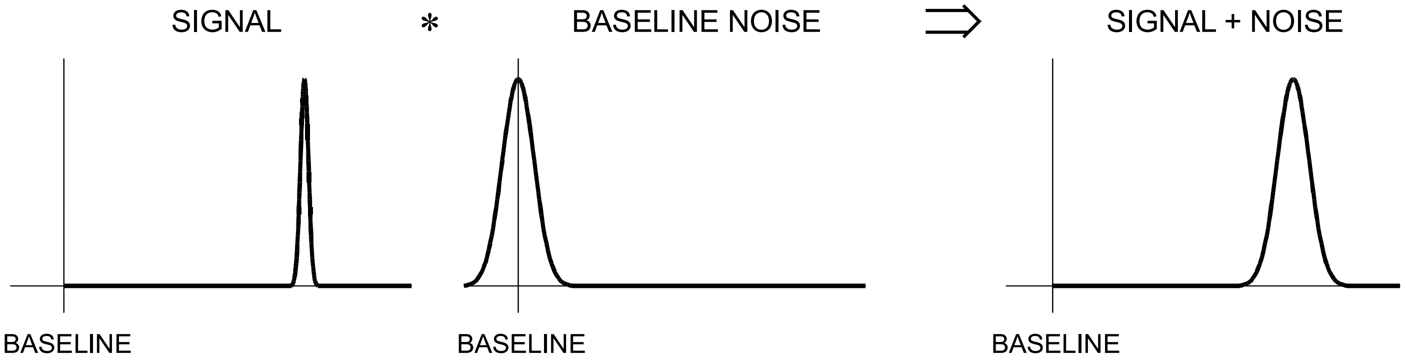
Overall Sensitivity or Resolution

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



⇒ Electronic (baseline) noise not important

2. Signal Variance << Baseline Variance



⇒ Electronic (baseline) noise is key!

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

⇒ Power distributed over wide frequency range

⇒ Contribution to energy fluctuations depends on signal processing

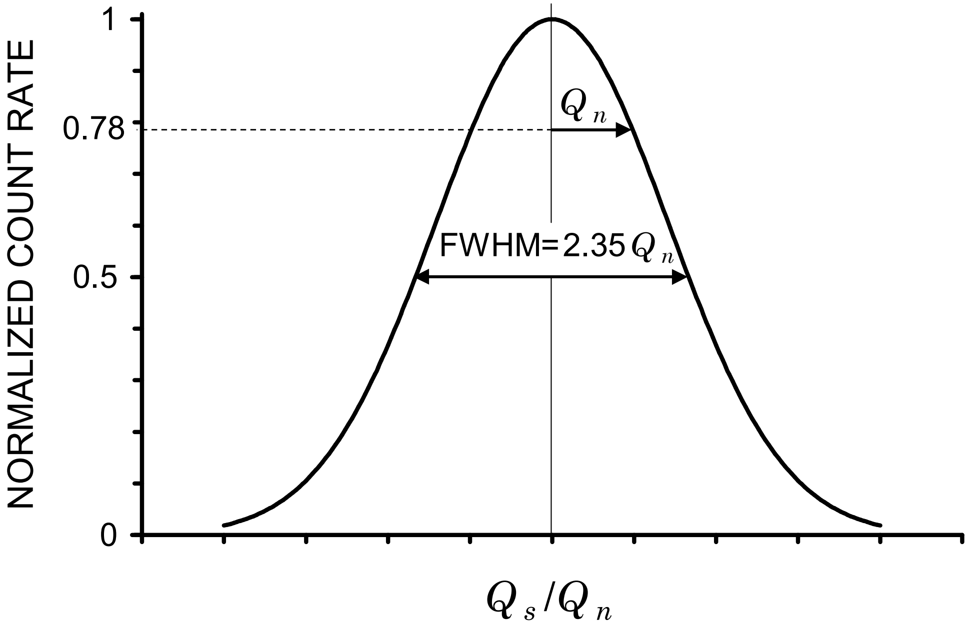
Electronic noise is purely random.

⇒ amplitude distribution is Gaussian

⇒ noise modulates baseline

⇒ baseline fluctuations superimposed on signal

⇒ output signal has Gaussian distribution



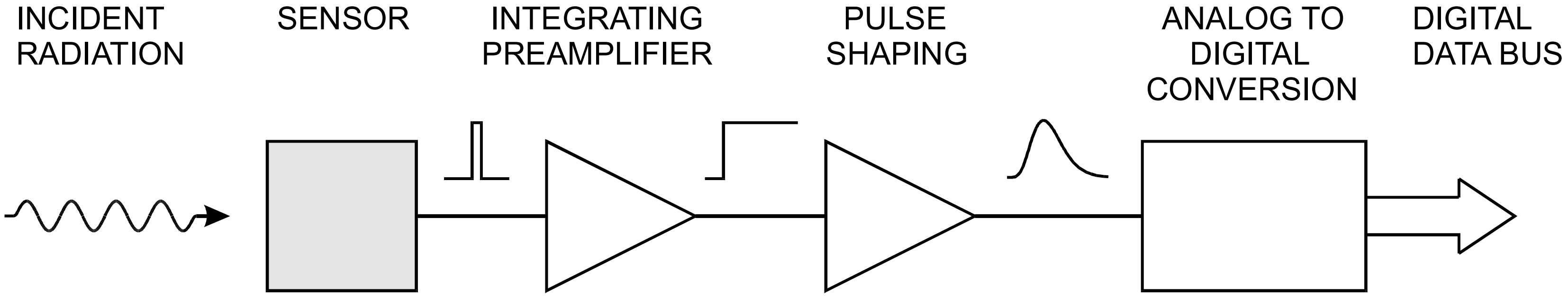
Measuring Resolution

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

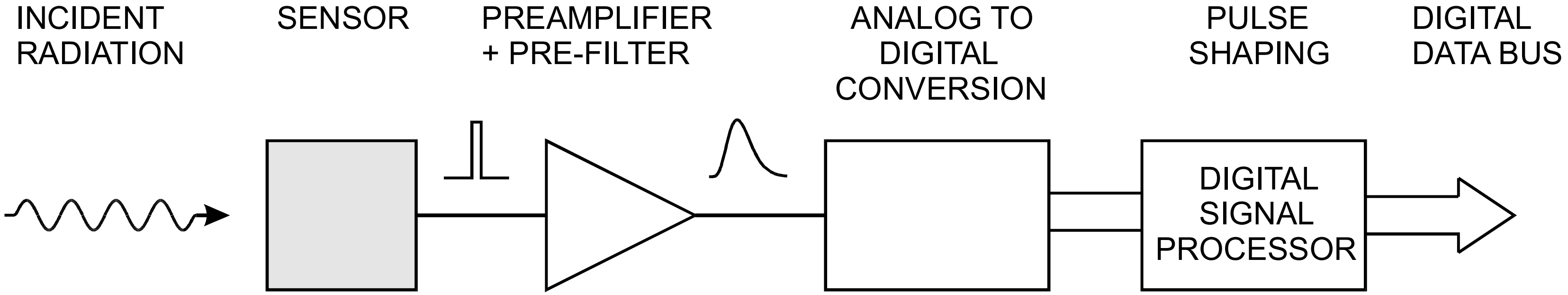
Measure the pulse height spectrum.

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

Basic Functions of Front-End Electronics



Pulse shaping can also be performed with digital circuitry:

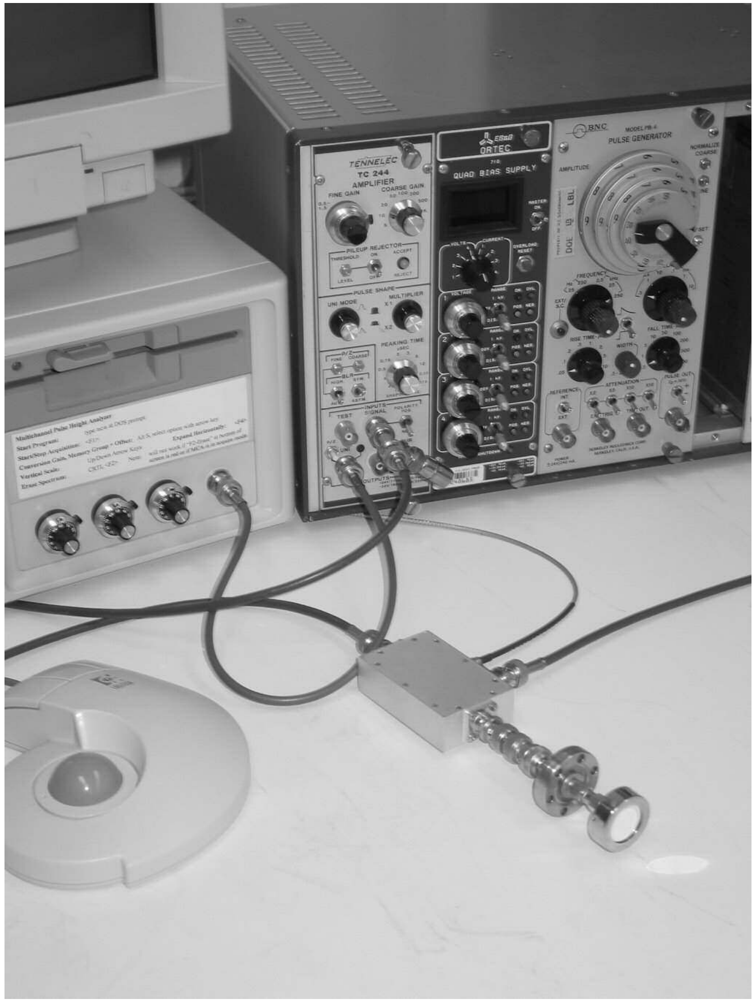


Digital pulse shaping provides great flexibility, but analog pulse shaping is more practical in high-density detectors such as strip and pixel systems.

- Analog shapers must not be complicated – *Every amplifier is a pulse shaper!*

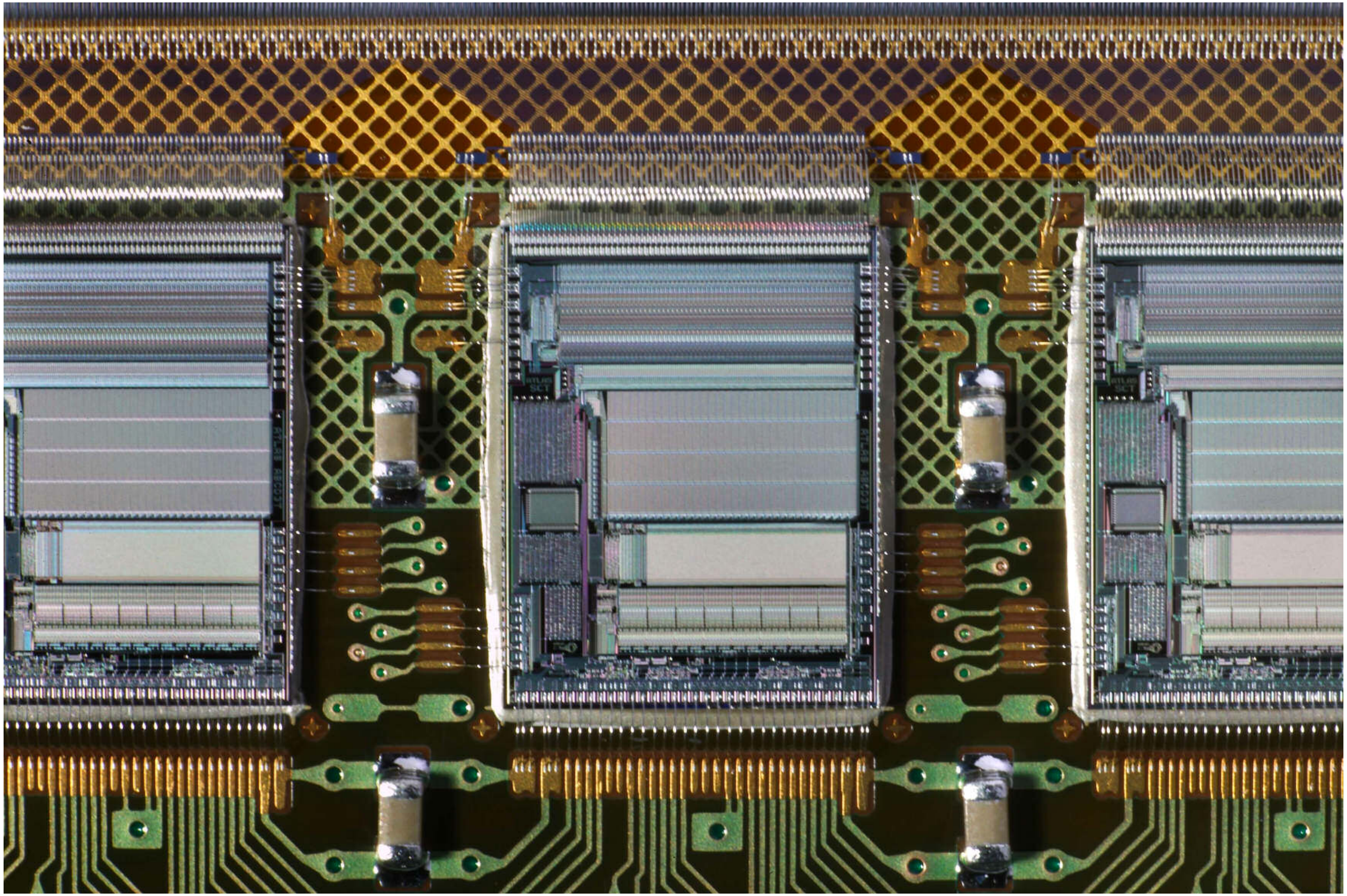
Many Different Implementations

“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

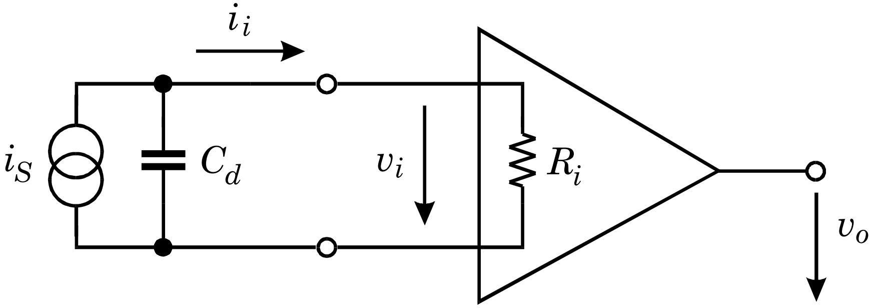
Technology choices

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”!

Signal Acquisition

A given amplifier can operate in either Voltage or Current Mode depending on the detector capacitance.



Output voltage:

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

The detector capacitance discharges through the amplifier input resistance R_i.

Operating mode depends on charge collection time t_c and the input time constant R_iC_d :

a) $R_iC_d \ll t_c$

detector capacitance discharges rapidly

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

current sensitive amplifier

b) $R_iC_d \gg t_c$

detector capacitance discharges slowly

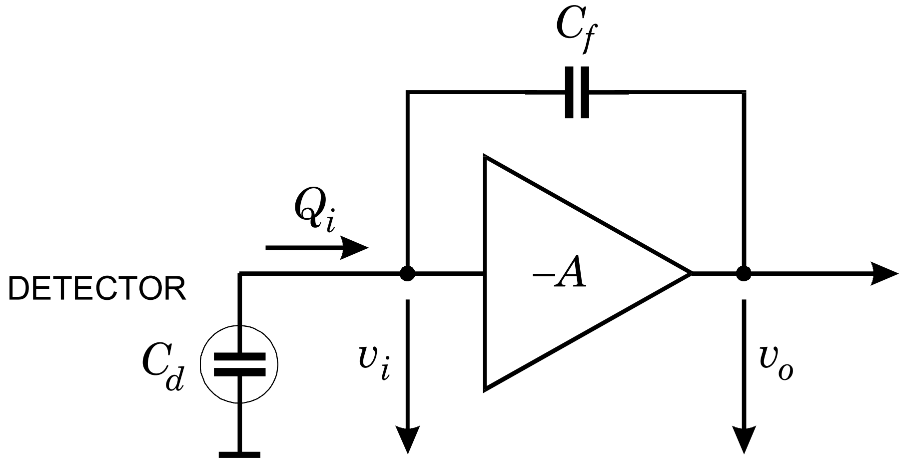
$\Rightarrow \quad v_o \propto \int i_s(t)dt$

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.

If the goal is to measure signal charge, 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: $v_f = (A+1)v_i$

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

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

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

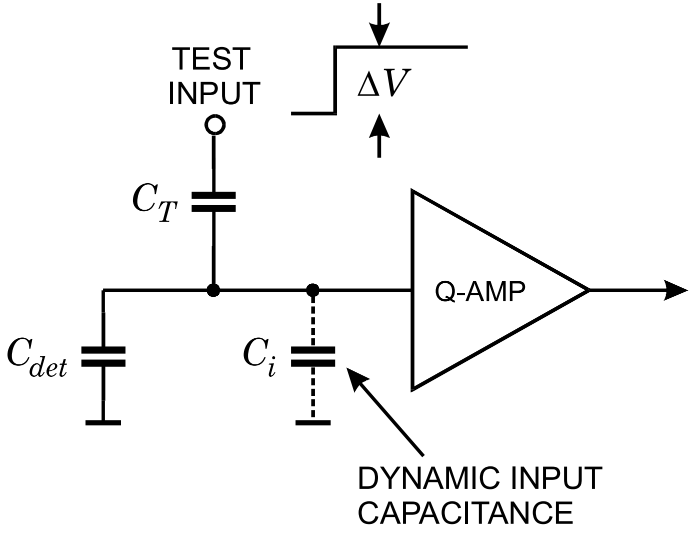
Gain $$A_Q = \frac{dV_o}{dQ_i} = \frac{A \cdot v_i}{C_i \cdot v_i} = \frac{A}{C_i} = \frac{A}{A+1} \cdot \frac{1}{C_f} \approx \frac{1}{C_f} \quad (A >> 1)$$

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

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.

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

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

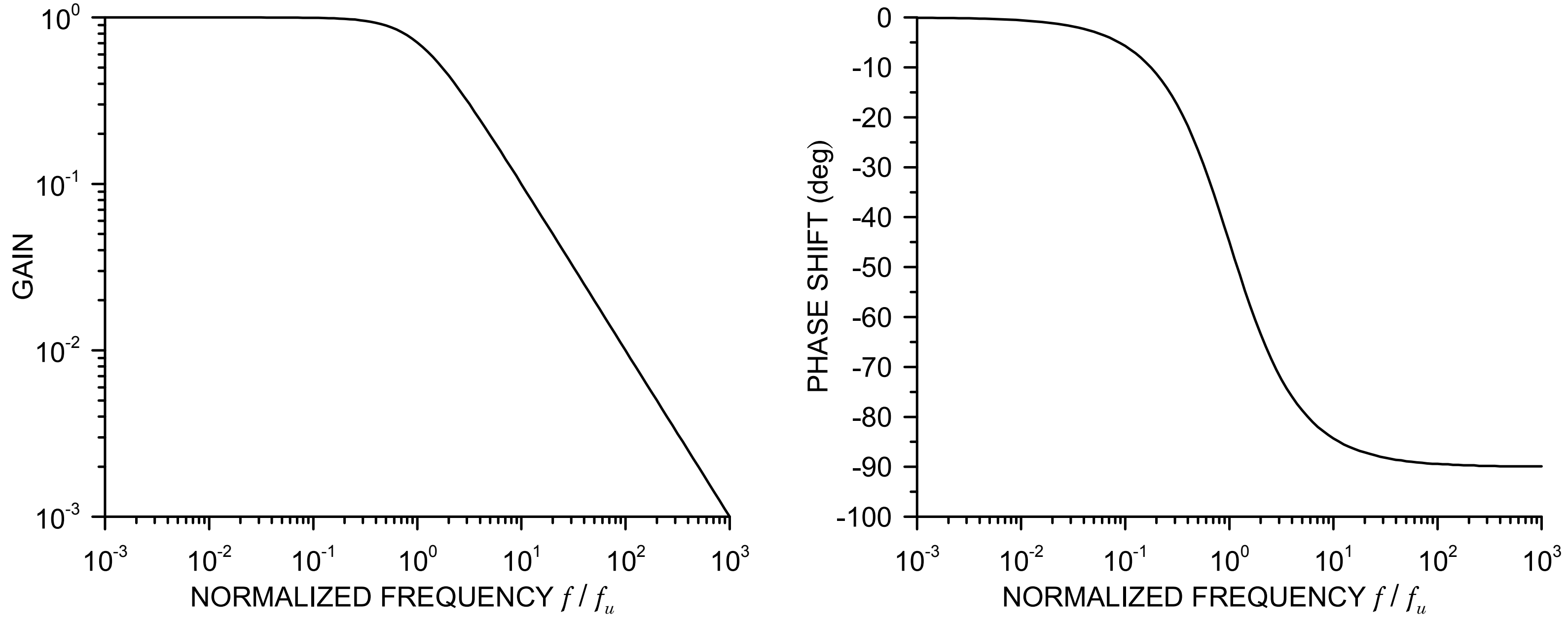
Typically: $C_T / C_i = 10^{-3} - 10^{-4}$

Realistic Charge-Sensitive Preamplifiers

The preceding discussion assumed idealized amplifiers with infinite speed.

Practical amplifiers have a limited bandwidth, which increases the pulse rise time.

Without the feedback capacitor C_f a practical frequency response is shown below



Beyond the cutoff frequency where the gain begins to drop, the phase shift of the output signal Phase shows change from low-frequency response.

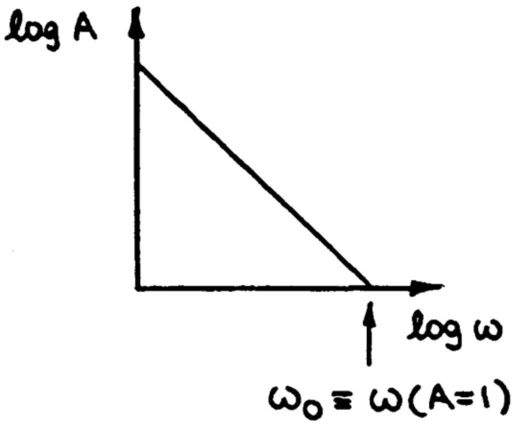
For an inverting amplifier, as used in the charge sensitive preamp, add 180°.

Input Impedance of a Charge-Sensitive Amplifier

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

Amplifier gain vs. frequency beyond the upper cutoff frequency

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



Gain-Bandwidth Product

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

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

$\mathbf{i}$ *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.

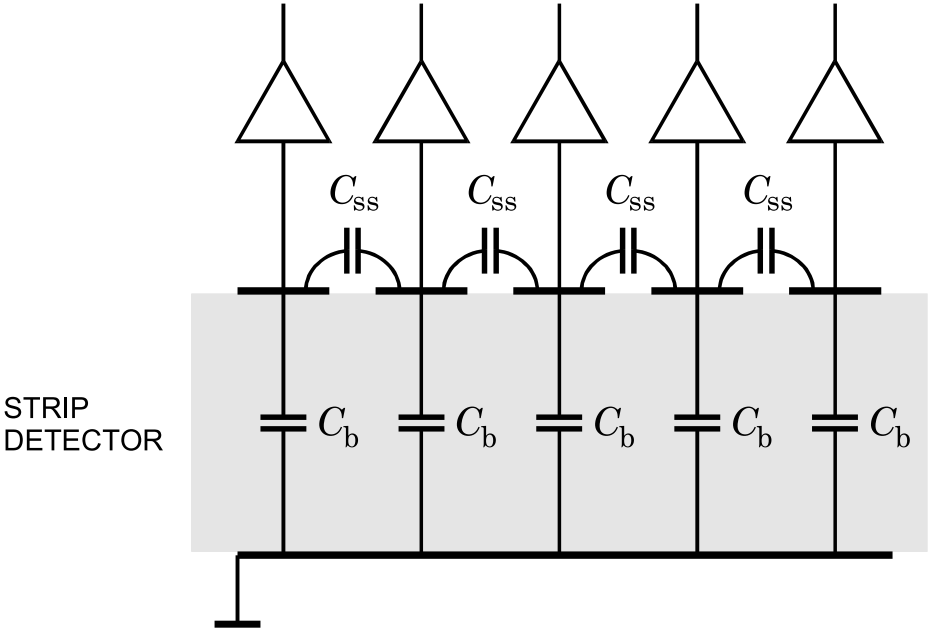
⇒ Resistive input

⇒ **Together with the detector capacitance it yields the rise time** $\tau_i = R_i C_D$

Rise time increases with detector capacitance: $\tau_i = \frac{1}{\omega_0 C_f} \cdot C_D$

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 or pixels.



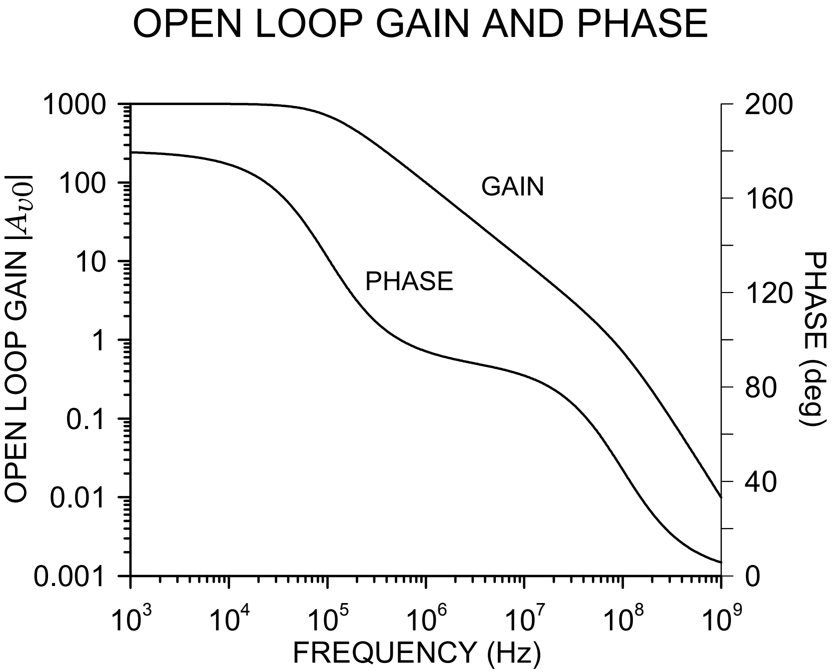
For strip pitches that are smaller than the bulk thickness, the capacitance is dominated by the fringing capacitance to the neighboring strips C_{SS}.

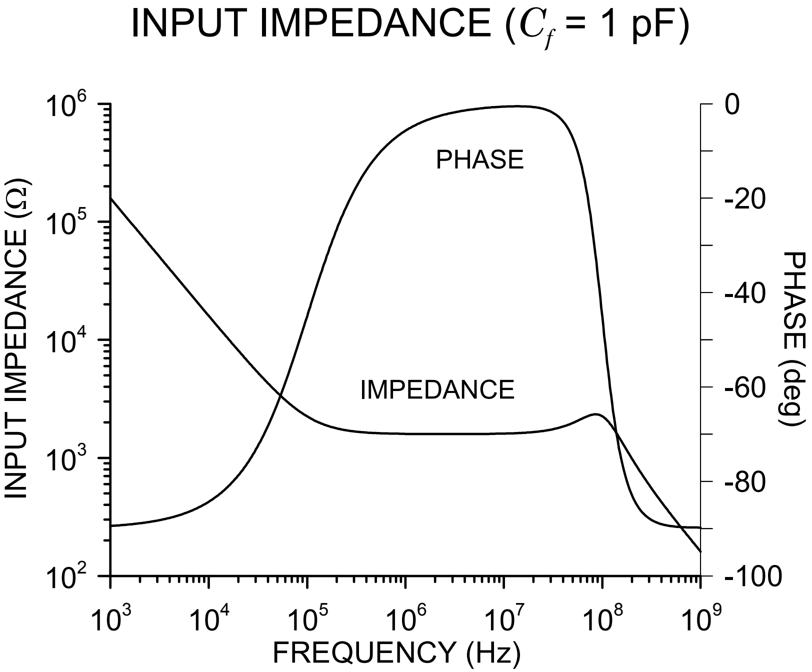
Typically: 1 – 2 pF/cm for strip pitches of 25 – 100 μm on Si.

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

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

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



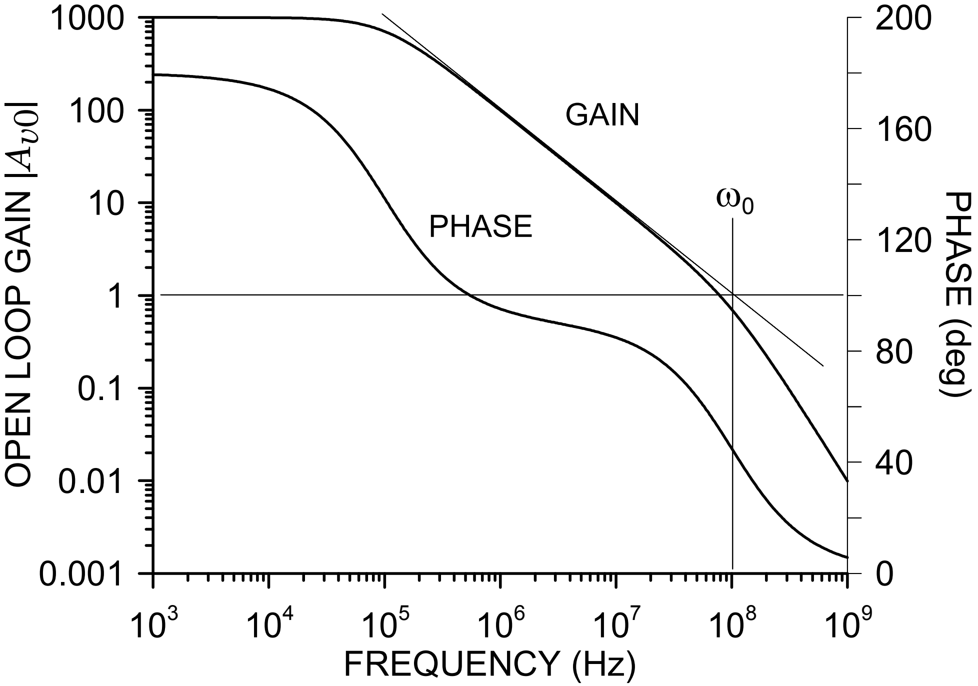


In the capacitive regime the input impedance drops with frequency, but then levels off in the resistive regime.

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.



At 10 MHz ($\triangleq$ ~20 ns peaking time)
$Z_i \approx$ 1.6 kΩ, corresponding to 10 pF

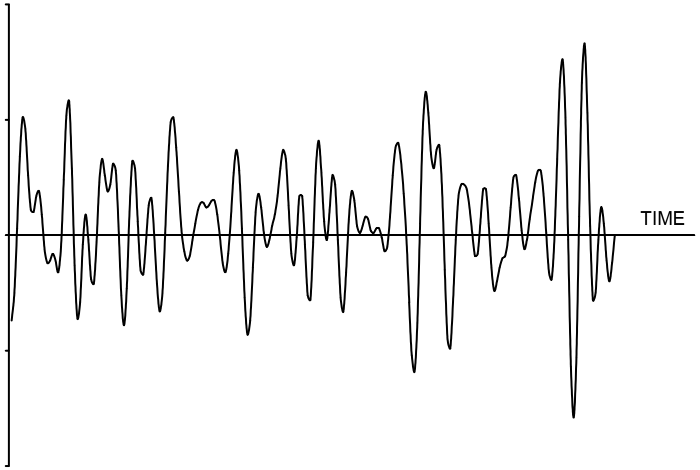
⇒ with 6 cm long strips about half of the signal current will go to the neighbors.

It is essential to confirm that the input impedance is low enough to reduce cross-talk through the inter-electrode capacitance to acceptable levels.

Electronic Noise

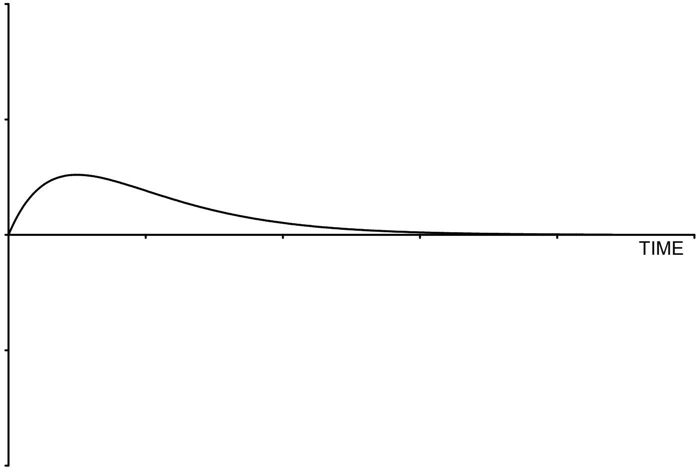
Choose a time when no signal is present.

Amplifier’s quiescent output level (baseline):

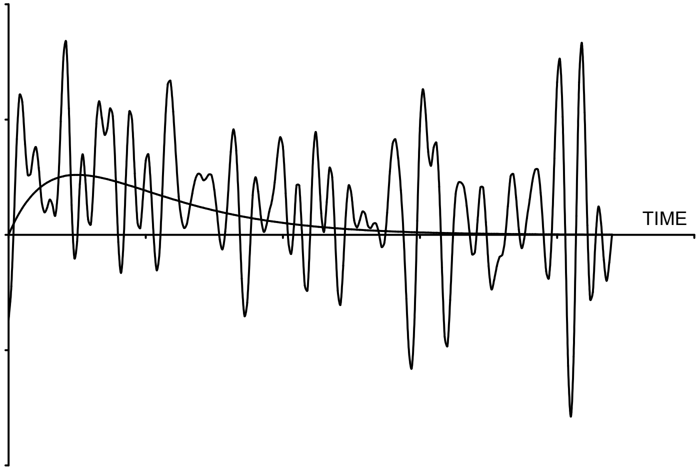


In the presence of a signal, noise + signal add.

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:

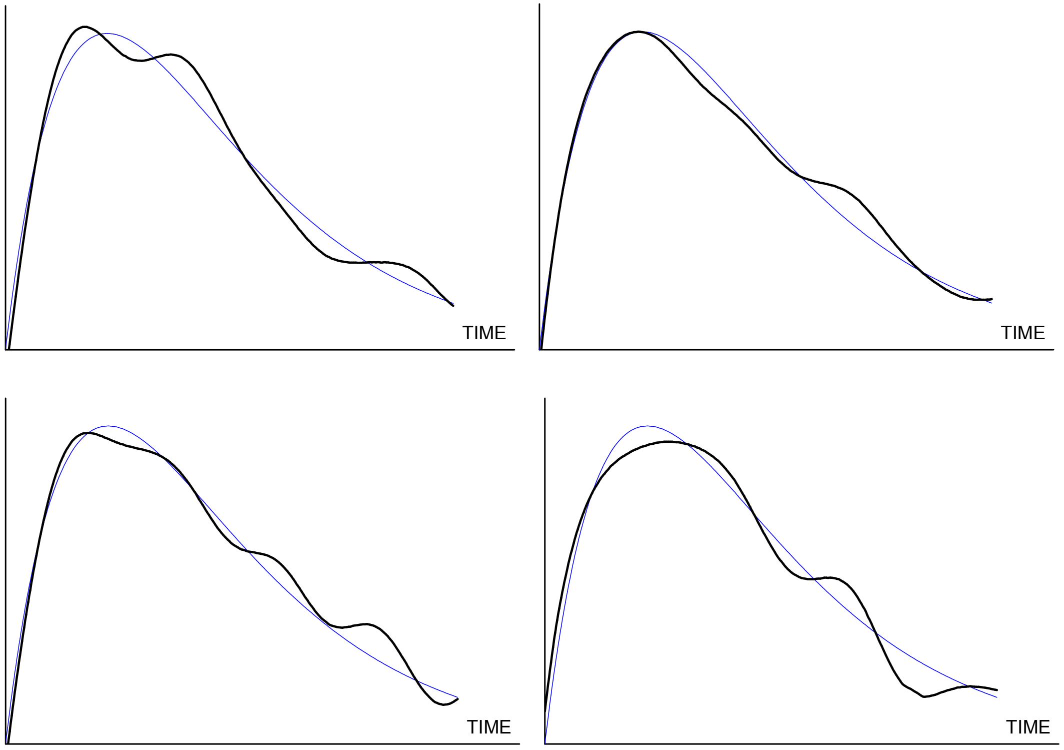
noiseless signal superimposed for comparison

S/N = 20

Noise affects

Peak signal

Time distribution



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:

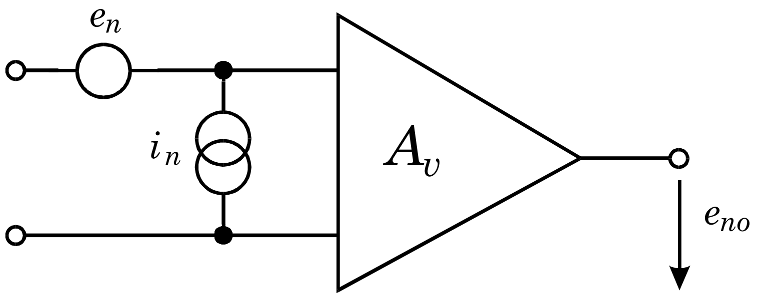
- velocity fluctuations, *e.g.* thermal noise ⇒ voltage noise
- number fluctuations, *e.g.* shot noise ⇒ current noise

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

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

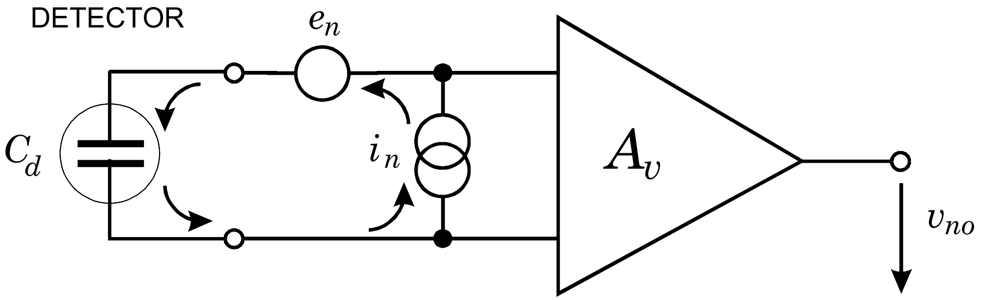
Additional noise at low frequencies often occurs because of charge trapping and release in the preamplifier input transistor ⇒ “1/*f* ” noise

Amplifier Noise Components

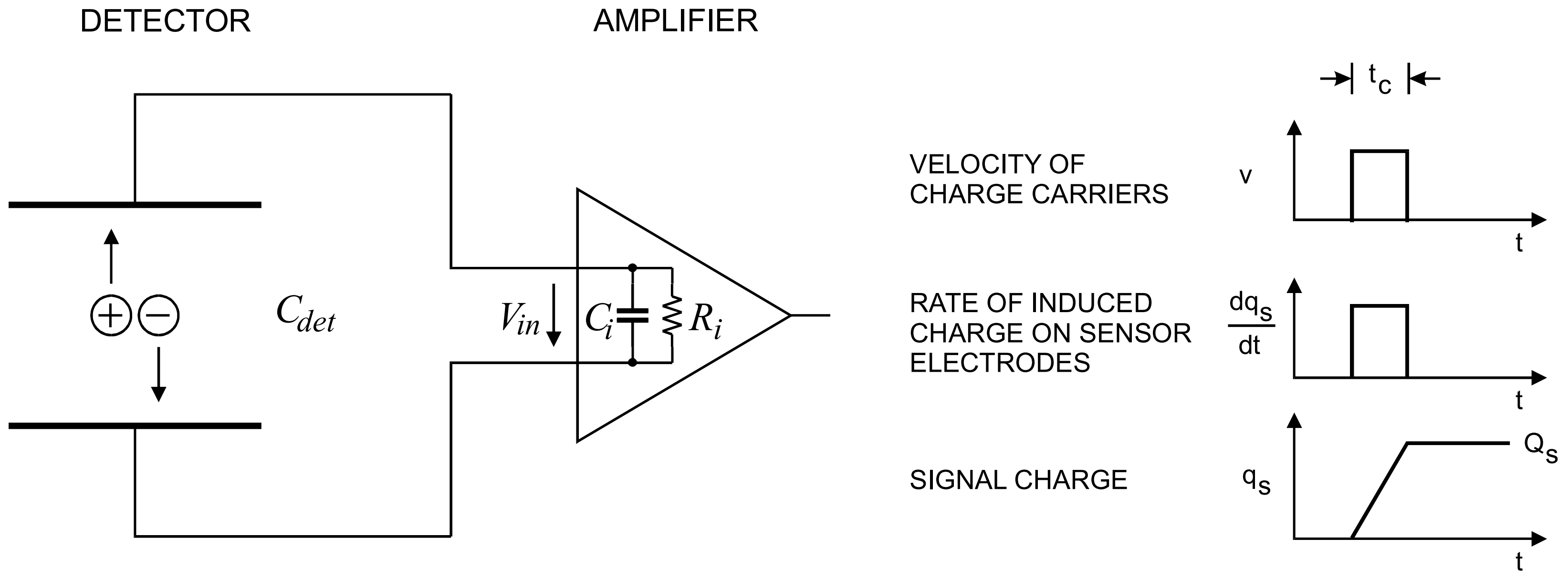


The noise voltage mainly originates in the inner of the amplifier. The noise voltage e_{no} at the output is divided by the amplifier voltage gain to yield the input related noise voltage e_n .

With a detector at the input the noise current flows through the capacitance and forms a noise voltage that increases with decreasing frequency, i.e. longer shaping times.



Signal-to-Noise Ratio vs. Detector Capacitance



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

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

↑

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

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

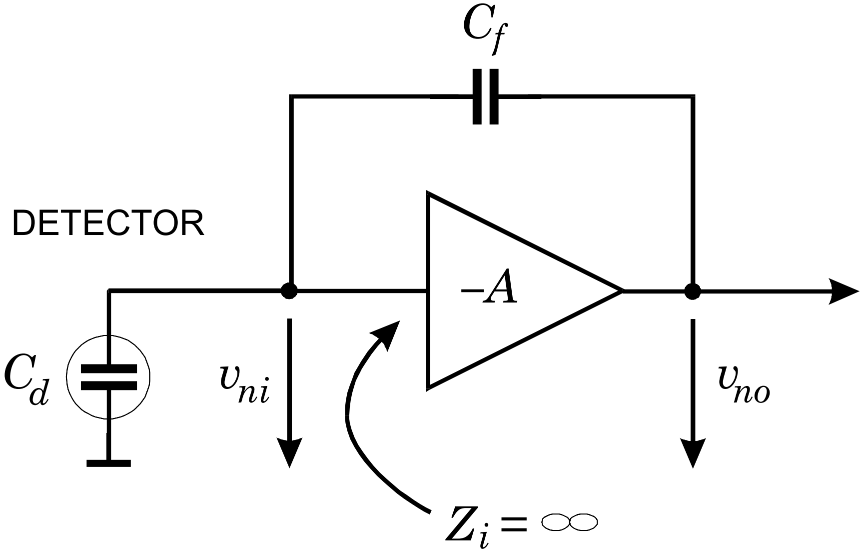
- However, S/N does not become infinite as $C \to 0$
  (then front-end operates in current mode)
- The result that $S/N \propto 1/C$ generally applies to systems that measure signal charge.
- Feedback amplifiers cannot increase S/N. They can add noise.

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 $C_f - C_d$.

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

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



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.

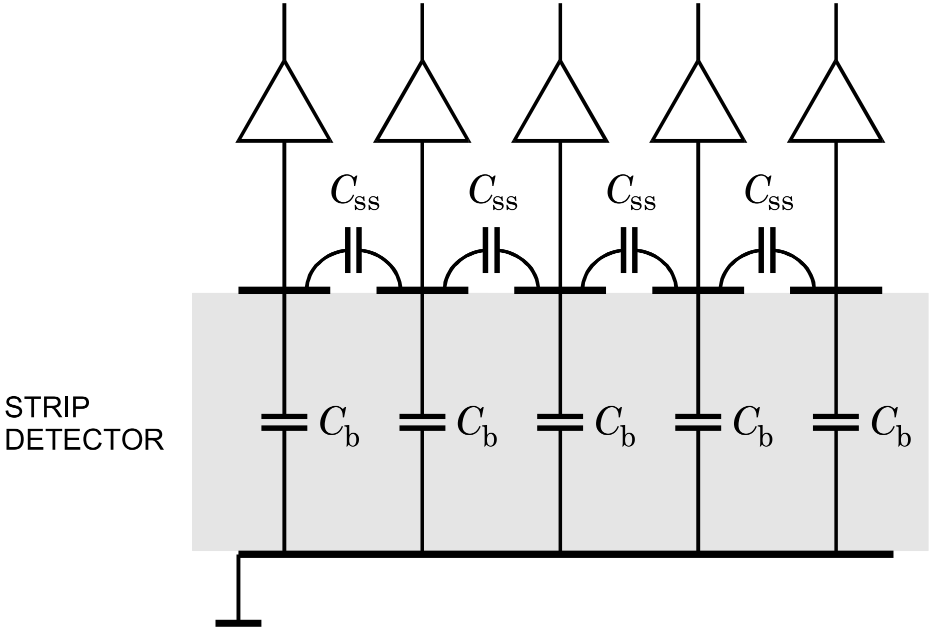
Conclusion

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.

Cross-Coupled Noise in Strip and Pixel Sensors



Noise at the input of an amplifier is cross-coupled to its neighbors through the inter-electrode capacitance

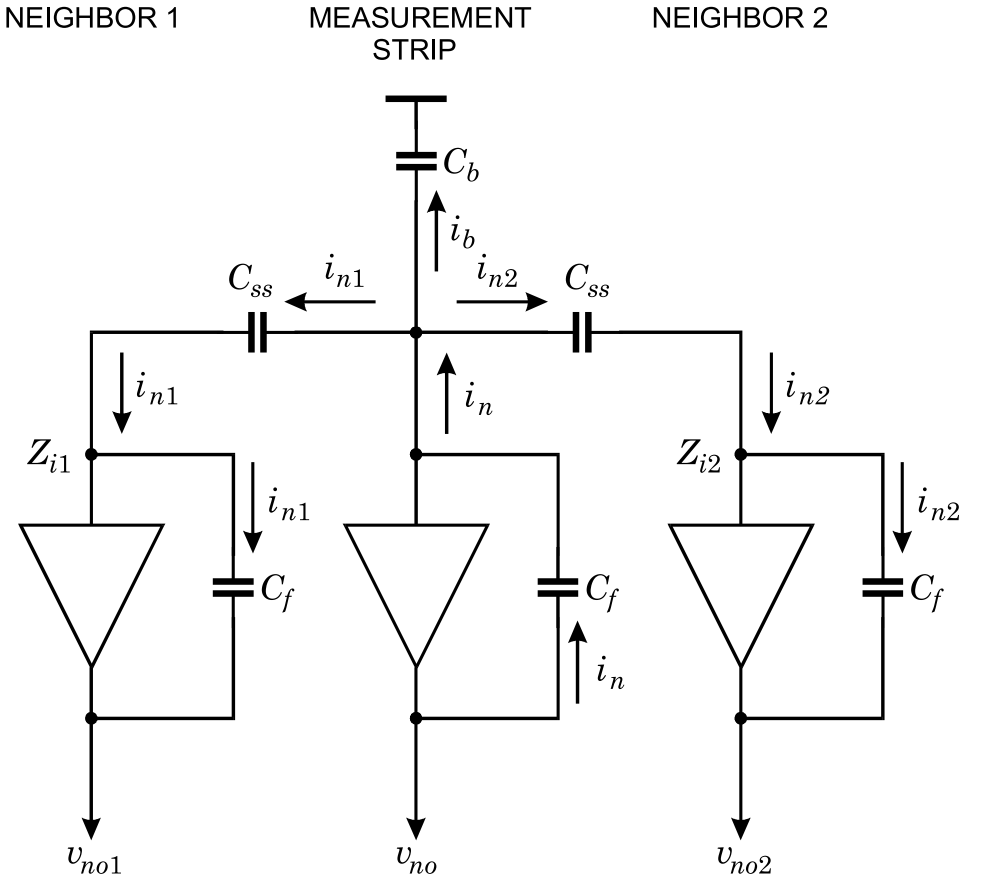
Noise Cross-Coupling Function in Strip and Pixel Detectors

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).

Signal Processing

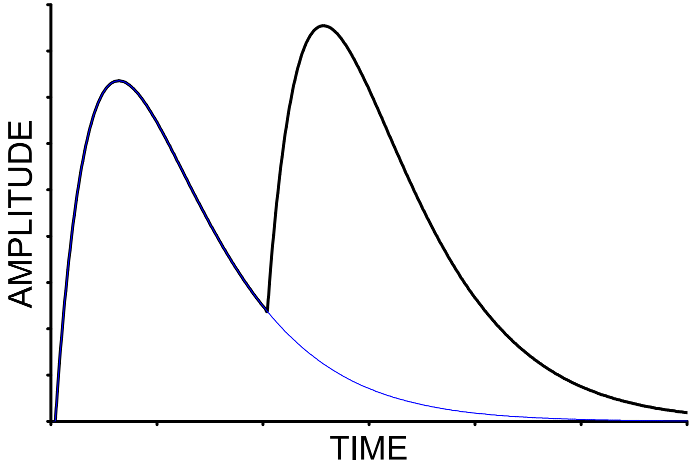
Requirements – Two conflicting objectives:

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

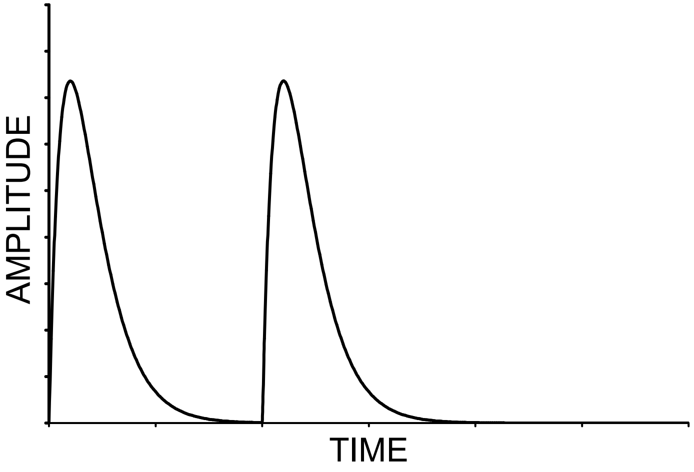
   Restrict bandwidth to match measurement time ⇒ Increase pulse width

2. Improve Pulse Pair Resolution ⇒ Decrease pulse width

Pulse pile-up distorts amplitude measurement.



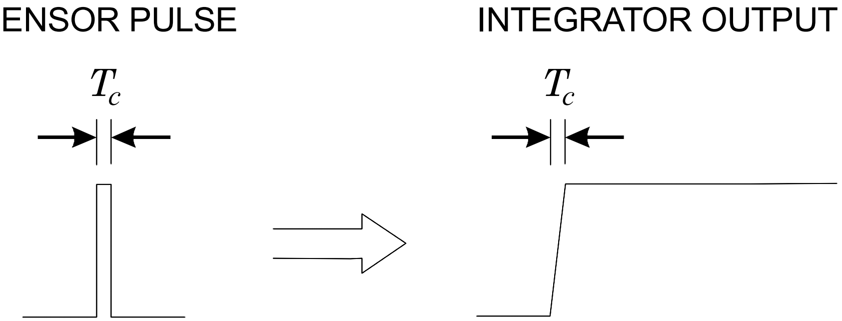
Reducing pulse shaping time avoids pile-up.



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

Goal: Improve energy resolution

Procedure: Integrate detector signal current ⇒ Step impulse



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

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

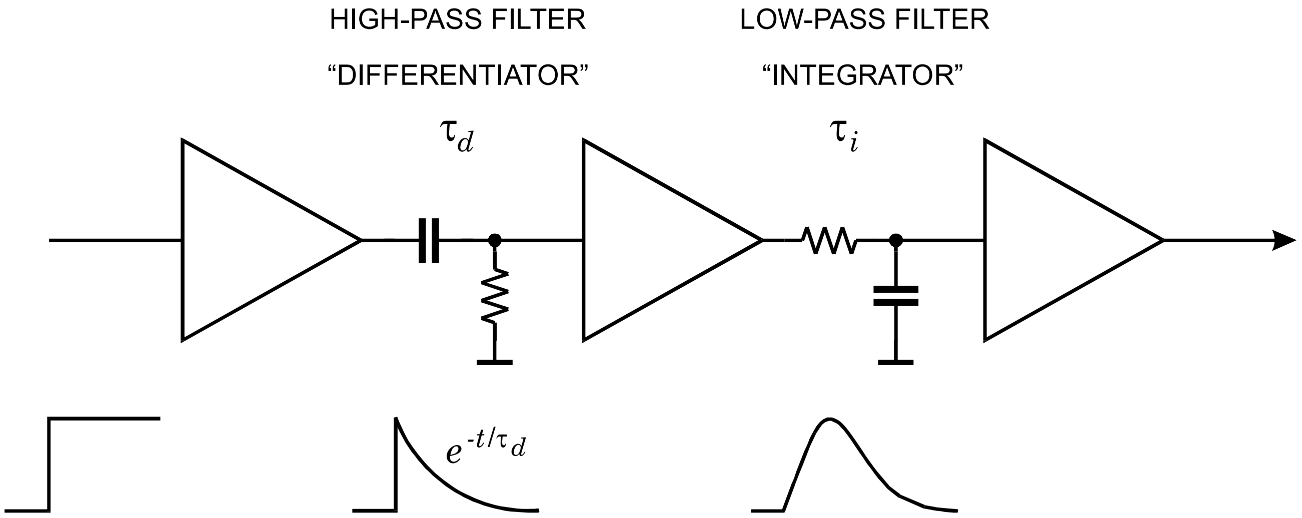
⇒ Allows reduced bandwidth and great flexibility in selecting shaper response.

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.

⇒ Useful for estimates, since simple to evaluate

Key elements:
- lower frequency bound ($\hat{=}$ pulse duration)
- upper frequency bound ($\hat{=}$ 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 $S/N = 1$

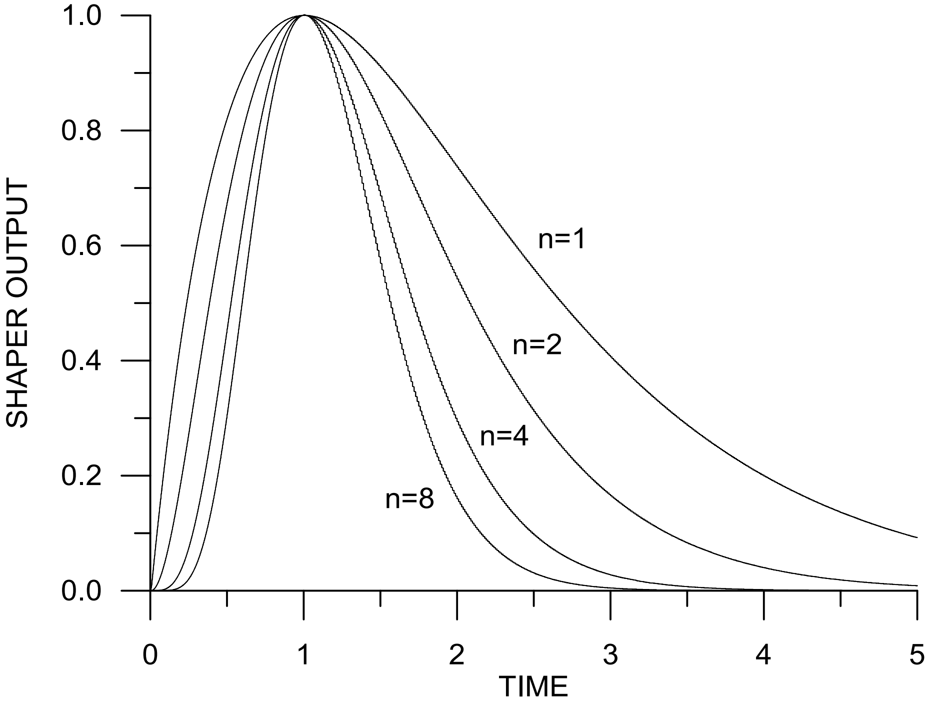
Shapers with Multiple Integrators

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

Reduce the integration time constant with the number of integrators to maintain the peaking time.

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

⇒ improved rate capability at the same peaking time



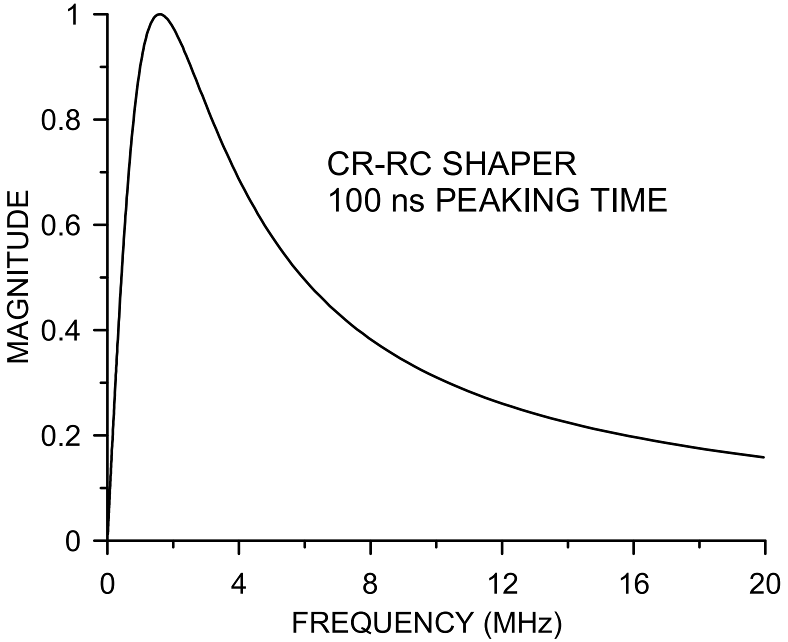
Multiple integrators often do not require additional circuitry.

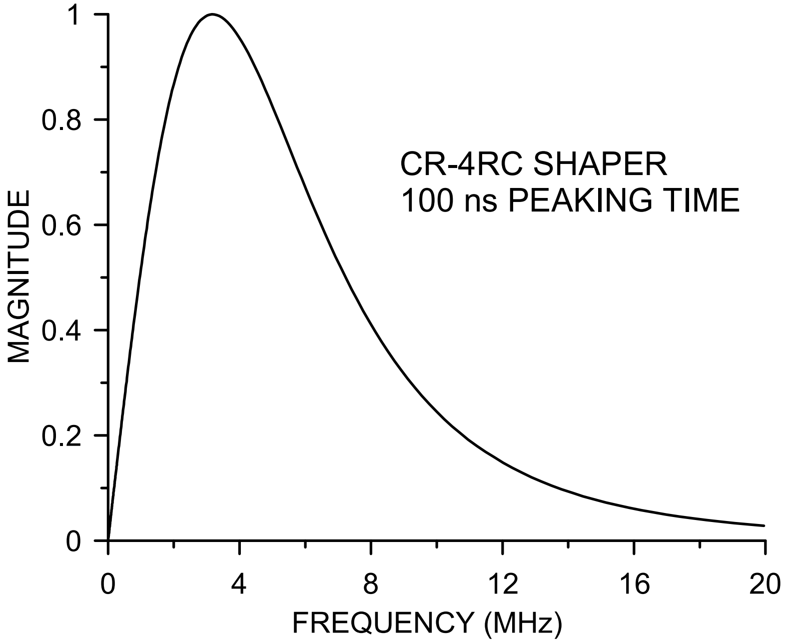
Several gain stages are typically necessary to bring the signal to the level required for a threshold discriminator or analog-to-digital converter.

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.

Frequency Response of a $CR\text{-}RC$ and $CR\text{-}4RC$ shaper





Both have a 100 ns peaking time.

The peaking frequencies are 1.6MHz for the $CR\text{-}RC$ shaper and 3.2 MHz for the $CR\text{-}4RC$.

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

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

Noise Charge vs. Pulse Shaping

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

Equivalent Noise Charge:

$$Q_n^2 = \underset{\substack{\uparrow \\ \text{current noise} \\ \propto \tau \\ \text{independent of } C_d}}{i_n^2 \, T_s \, F_i} + \underset{\substack{\uparrow \\ \text{voltage noise} \\ \propto 1/\tau \\ \propto C_d^2}}{C^2 \, e_n^2 \, \frac{F_v}{T_s}}$$

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 + …)

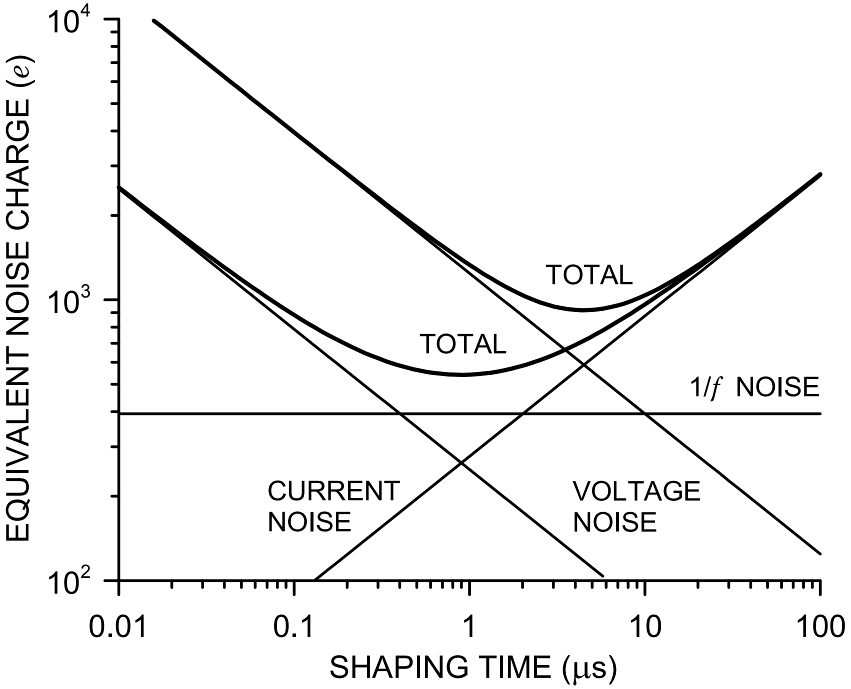
Typical values of F_i , F_v

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 ($T_{opt} \approx$ 130 K)

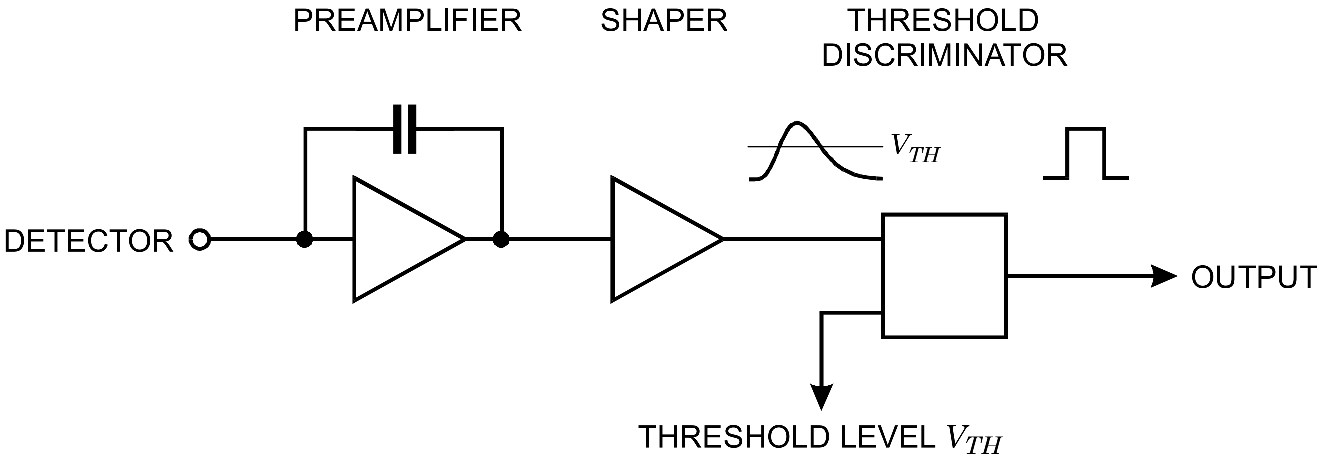
Bipolar transistors advantageous at short shaping times (<100 ns).
When collector current is optimized, bipolar transistor equivalent noise charge is independent of shaping time (see Chapter 6).

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.



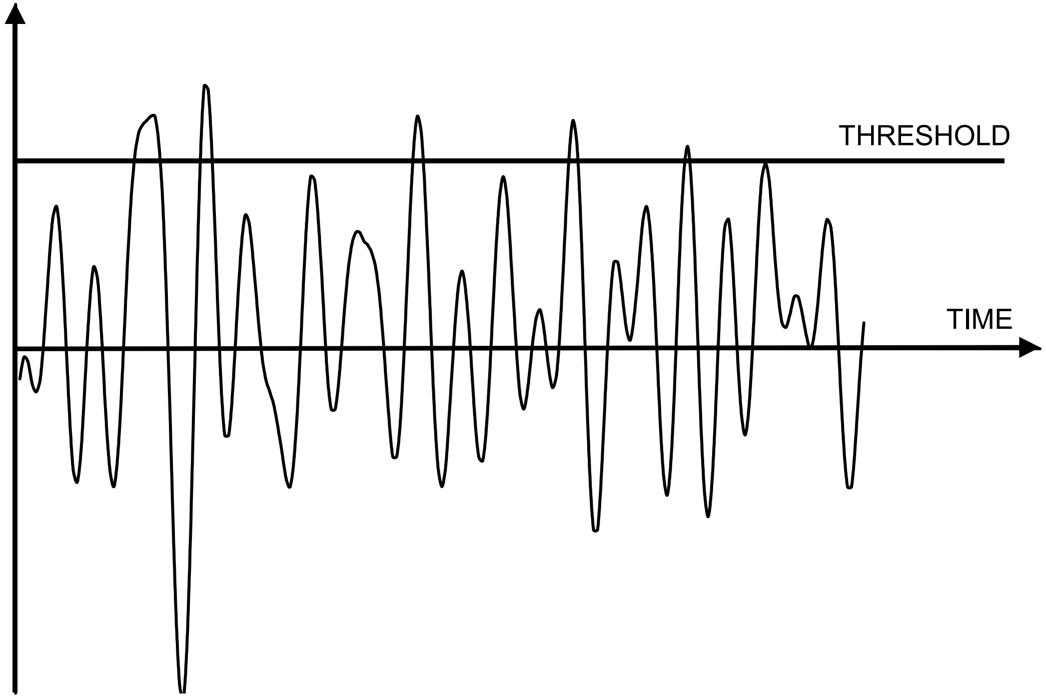
How small a detector pulse can still be detected reliably?

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.

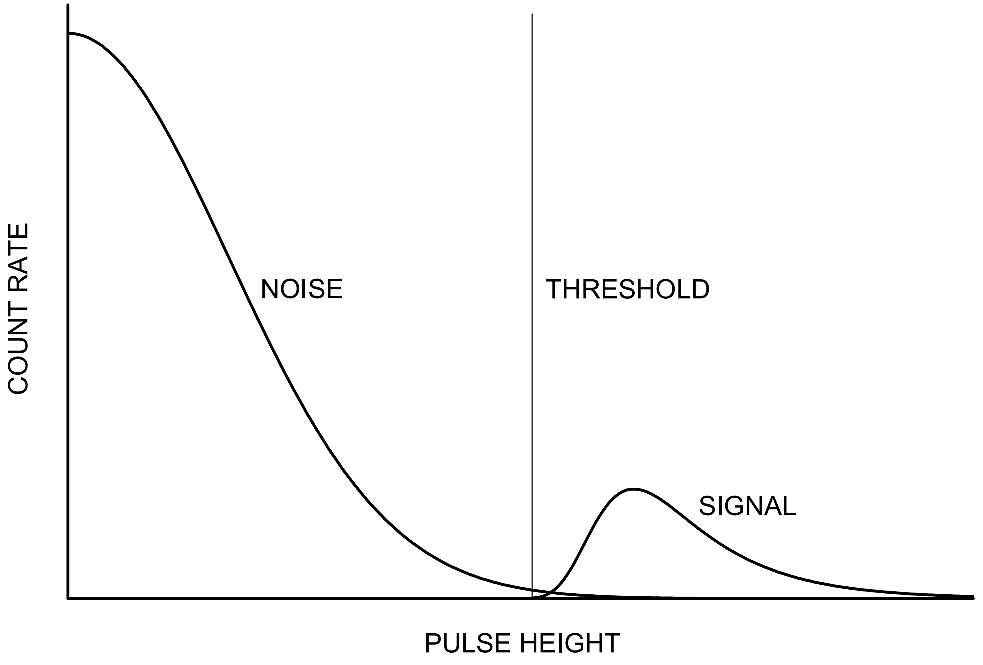
This is always true since the amplitude spectrum of Gaussian noise extends to infinity



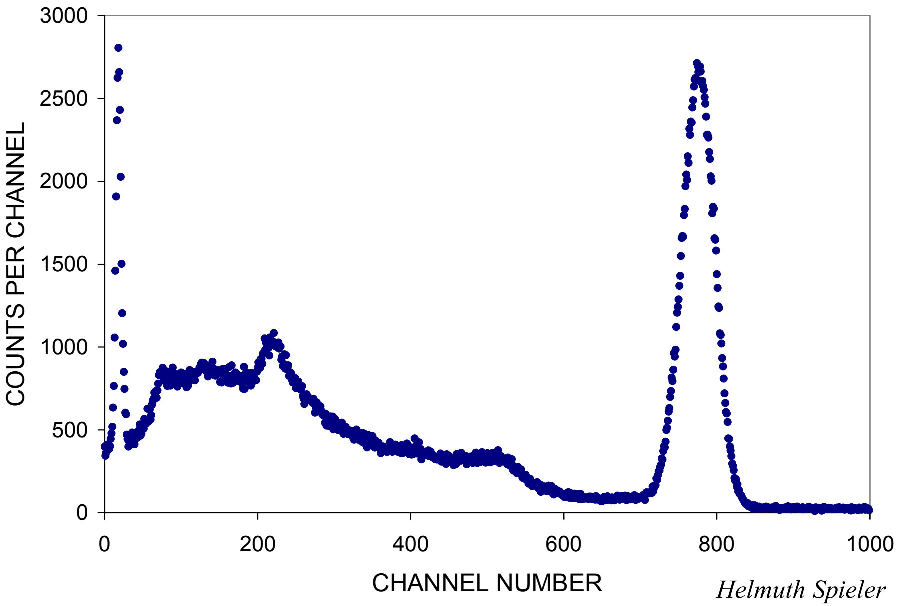
The threshold must be set

1. high enough to suppress noise hits

2. low enough to capture the signal



In a typical photon spectrum with Compton scattering, the threshold must be set much lower to record all interactions of the upper-energy photons.



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).

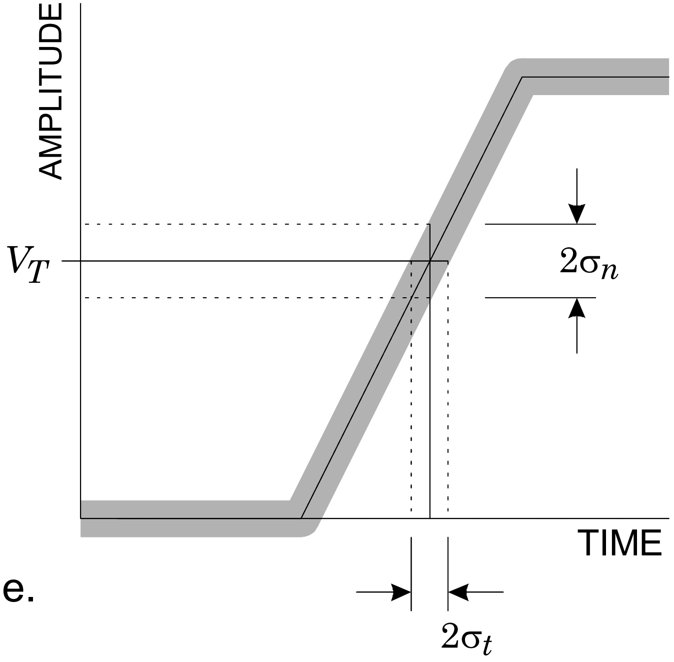
The instantaneous signal level is modulated by noise.

⇒ time of threshold crossing fluctuates

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

t_r = rise time

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



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}.

Rise times add in quadrature $t_r \approx \sqrt{t_c^2 + t_{ra}^2}$

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 S/N: The amplifier rise time should be chosen to match the signal rise time.

Differentiation time constant: choose greater than rise time constant

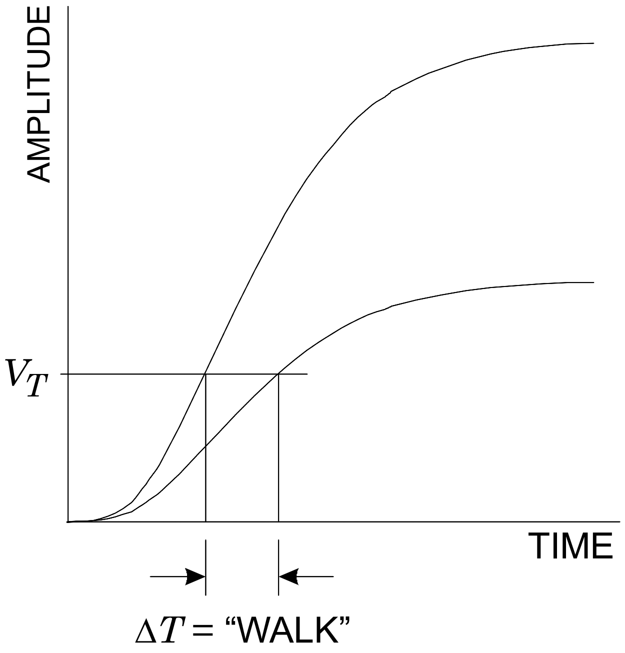
($\tau_{diff} = 10\tau_{\text{int}}$ incurs 20% loss in pulse height)

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.

For more details on fast timing with semiconductor detectors, see

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

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.