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Excellence in Detectors and Instrumentation Technologies

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Silicon Readout – Where The Bugs Can Hide

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

or simply Google "spieler detectors"

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

Course Contents

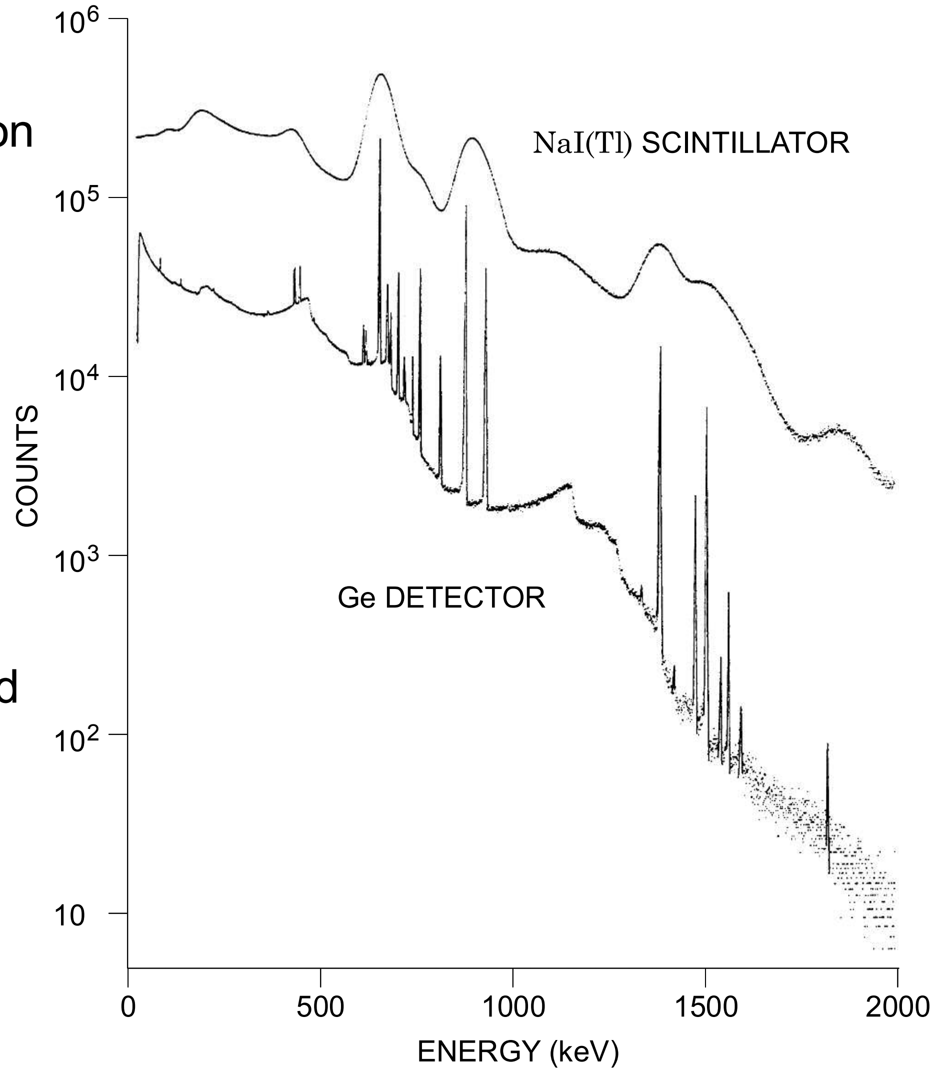
Why understand front-end electronics?

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.

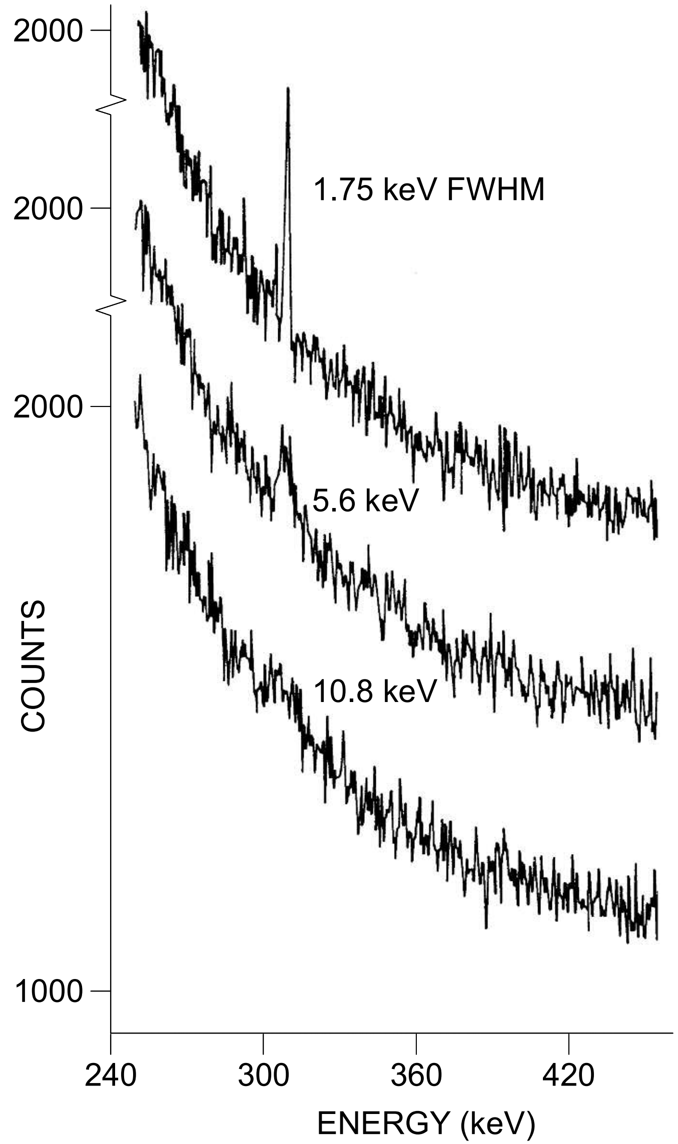
Energy resolution improves sensitivity because

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,

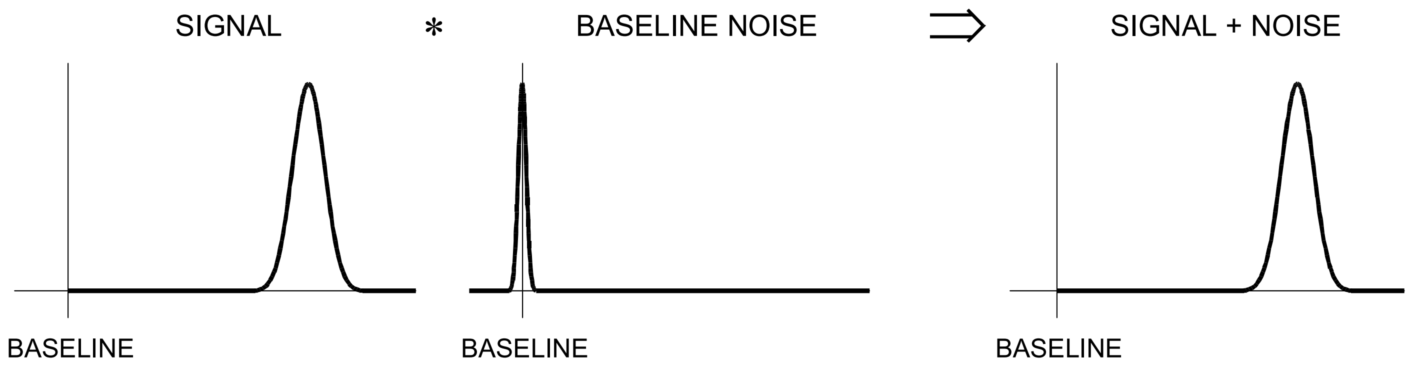
> i.e. recognize which aspects of physics apply in practical situations.
>
> … nope, real life doesn't tell you which chapter to follow!

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 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

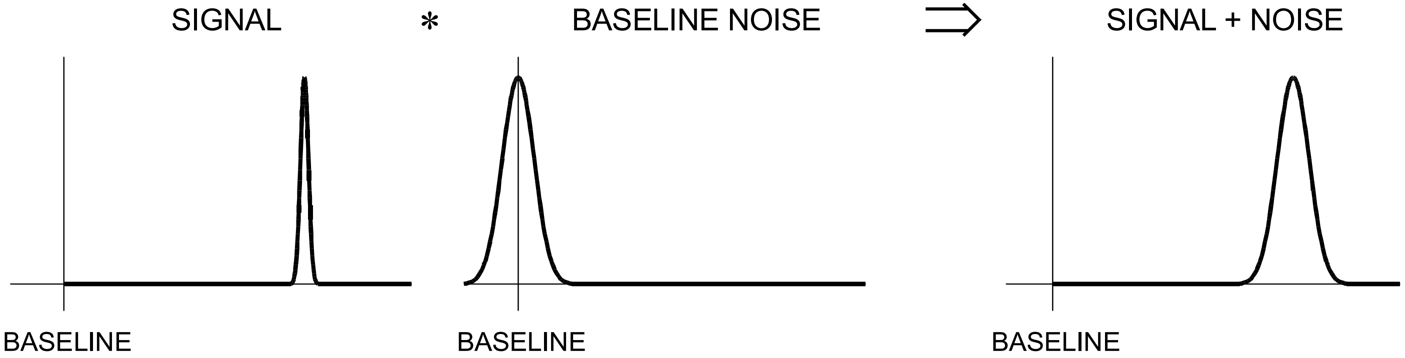
Number of photoelectrons: $N_{pe} \approx 8 \cdot 10^4$ [MeV^{-1}] $\times E_\gamma \times QE \approx 2.4 \cdot 10^4$

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 and
$\sigma_{sig} = G_{PMT}\, \sigma_{pe} \approx 1.2 \cdot 10^7$ 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 $N_{ep} = \dfrac{E_{dep}}{3.6\text{ eV}}$

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.

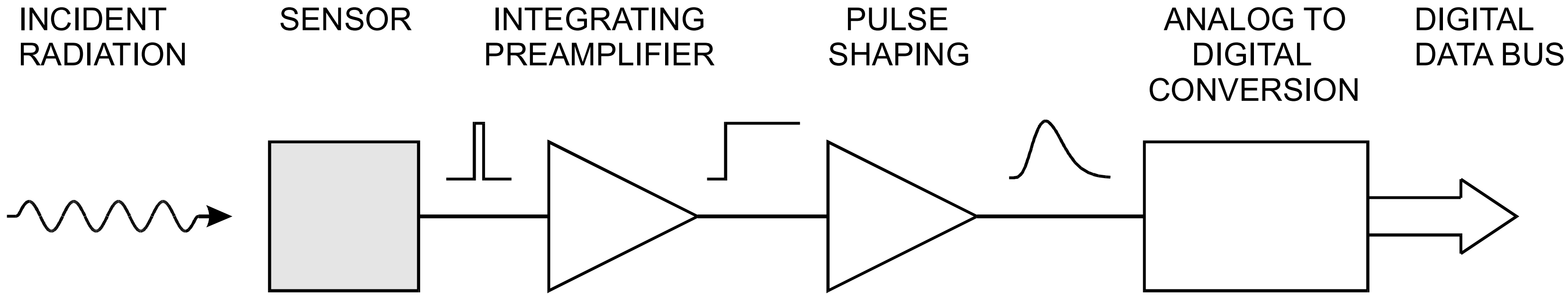
Although detectors appear to be very different, basic principles of the readout apply to all.

- 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.

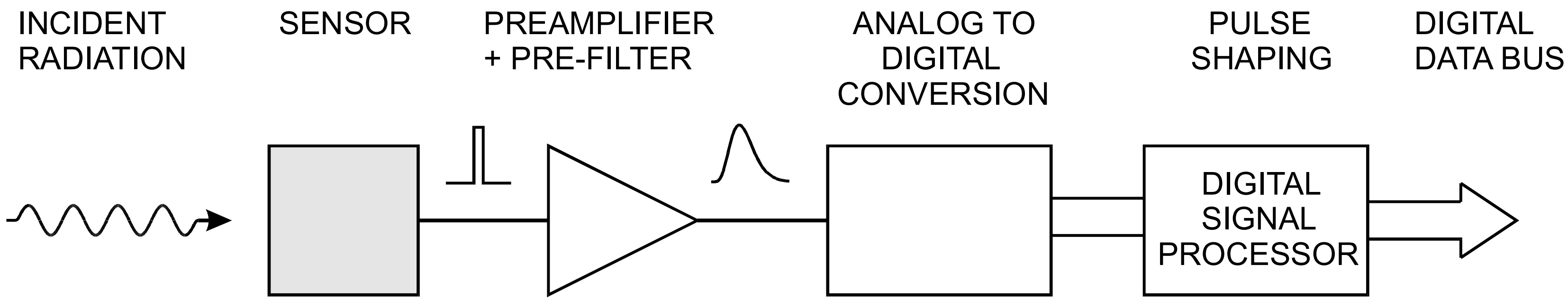
Readout systems include the following functions:

- Signal acquisition
- Pulse shaping
- Digitization
- Data Readout

1. Basic Functions of Front-End Electronics

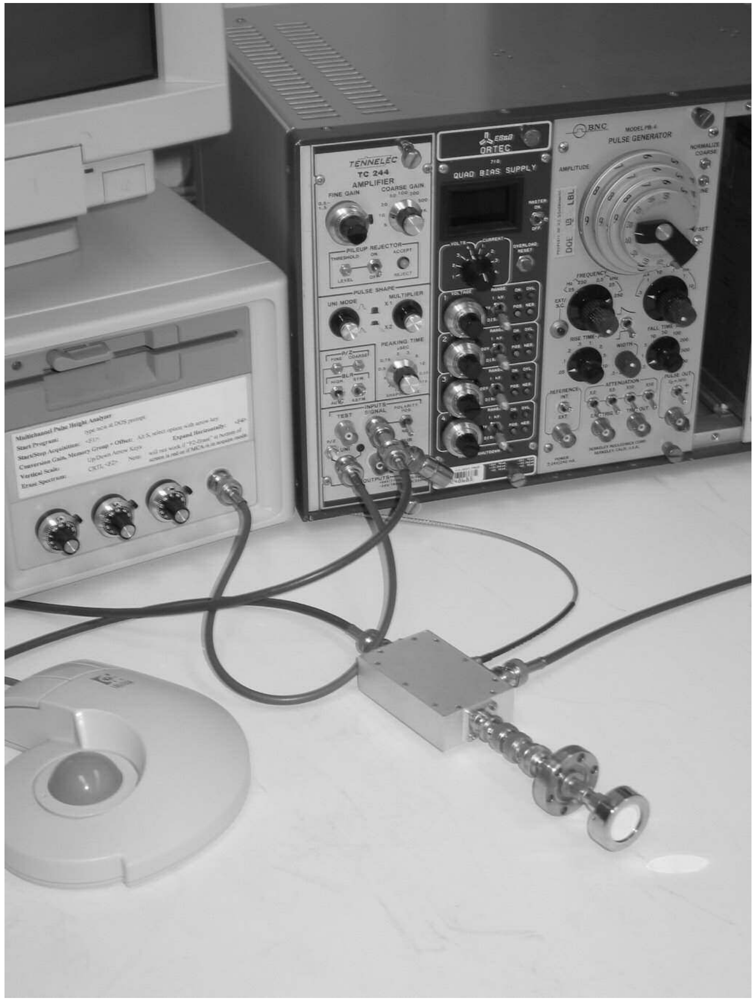


Pulse shaping can also be performed with digital circuitry:



Many Different Implementations

"Traditional" Si detector system for charged particle measurements



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

Spectroscopy systems highly optimized!

By the late 1970s improvements were measured in single %.

Separate system components:

1. detector
2. preamplifier
3. amplifier
    - adjustable gain
    - adjustable shaping
        - (unipolar + bipolar)
    - adjustable pole-zero cancellation
    - baseline restorer

Beam times typ. few days with changing configurations, so equipment must be modular and adaptable.

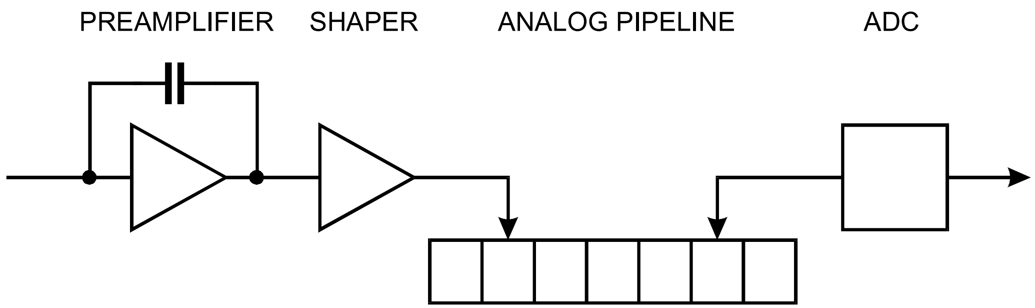
Today, systems with many channels are required in many fields.

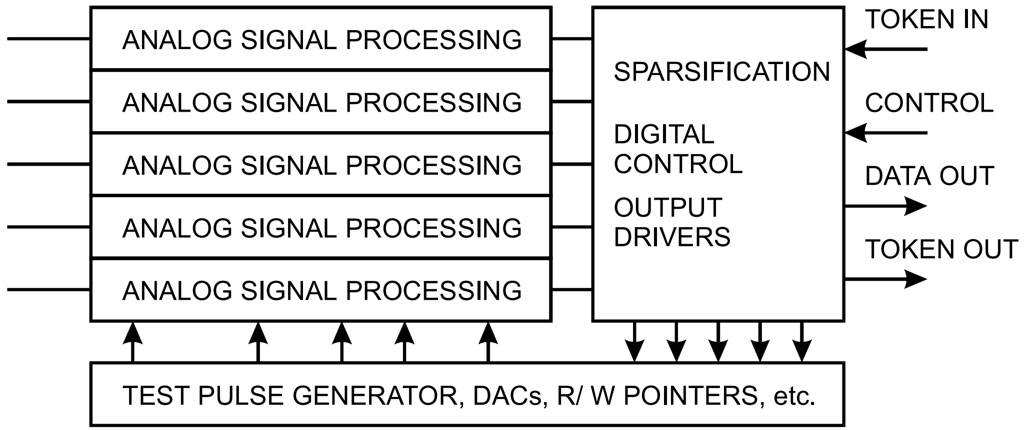
In large systems power dissipation and size are critical, so systems are not necessarily designed for optimum noise, but *adequate* noise, and circuitry is tailored to specific detector requirements.

Large-Scale Readout Systems

Example: Si strip detector

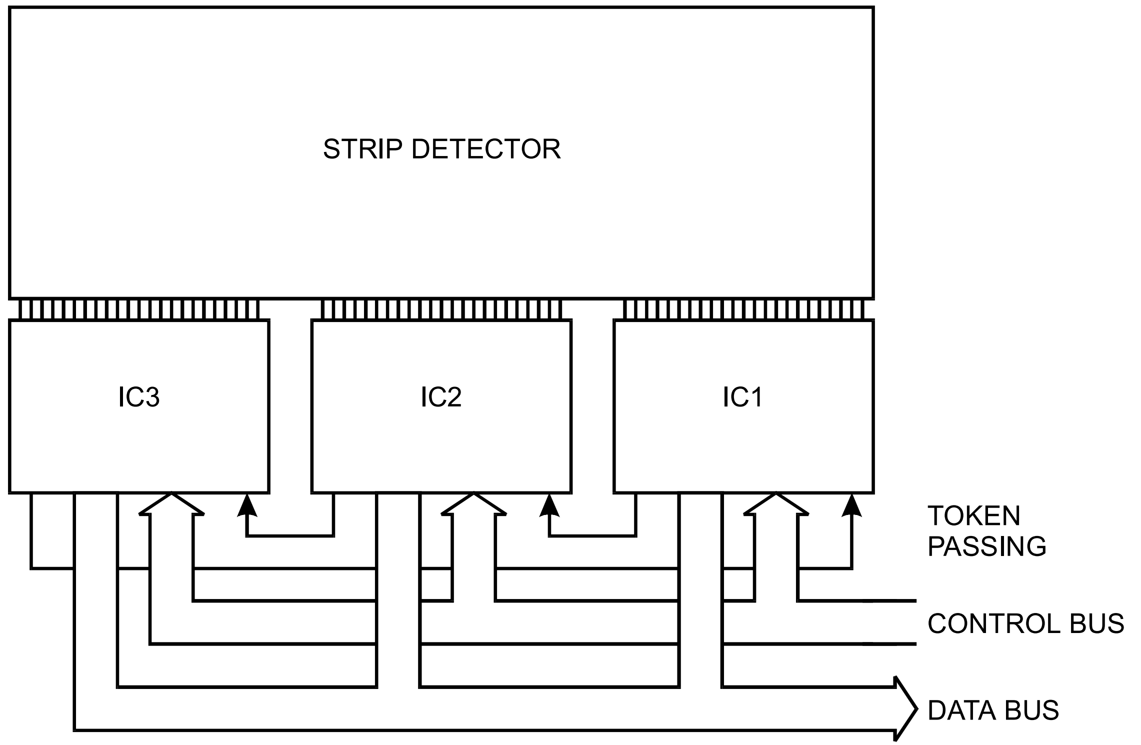
On-chip Circuits





Inside a typical readout IC:
- 128 parallel channels of analog front-end electronics
- Logic circuitry to decode control signals, load DACs, etc.
- Digital circuitry for zero-suppression, readout

Readout of Multiple ICs



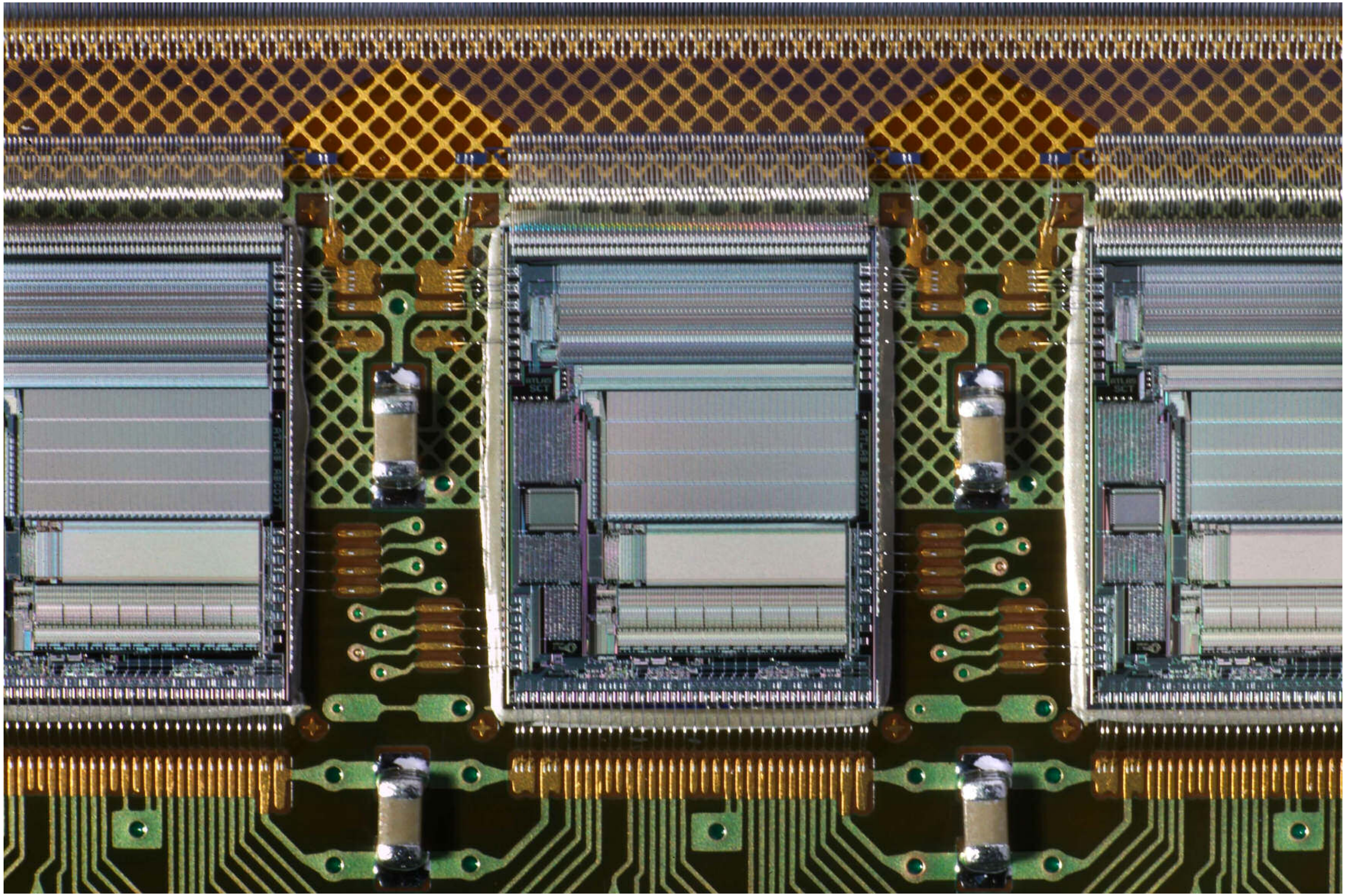
IC1 is designated as master.

Readout is initiated by a trigger signal selecting appropriate time stamp to IC1.

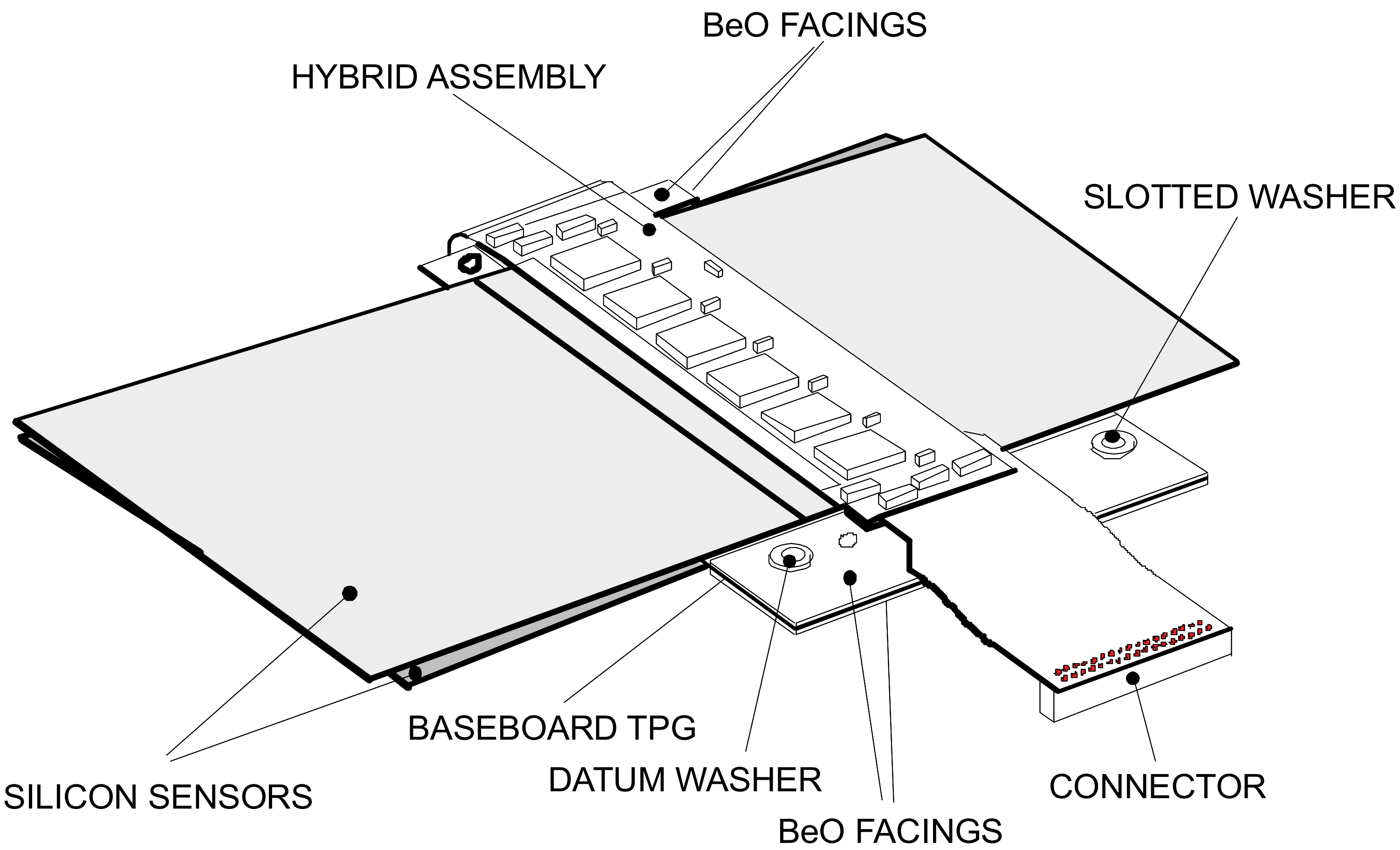
When all data from IC1 have been transferred, a token is passed to IC2.

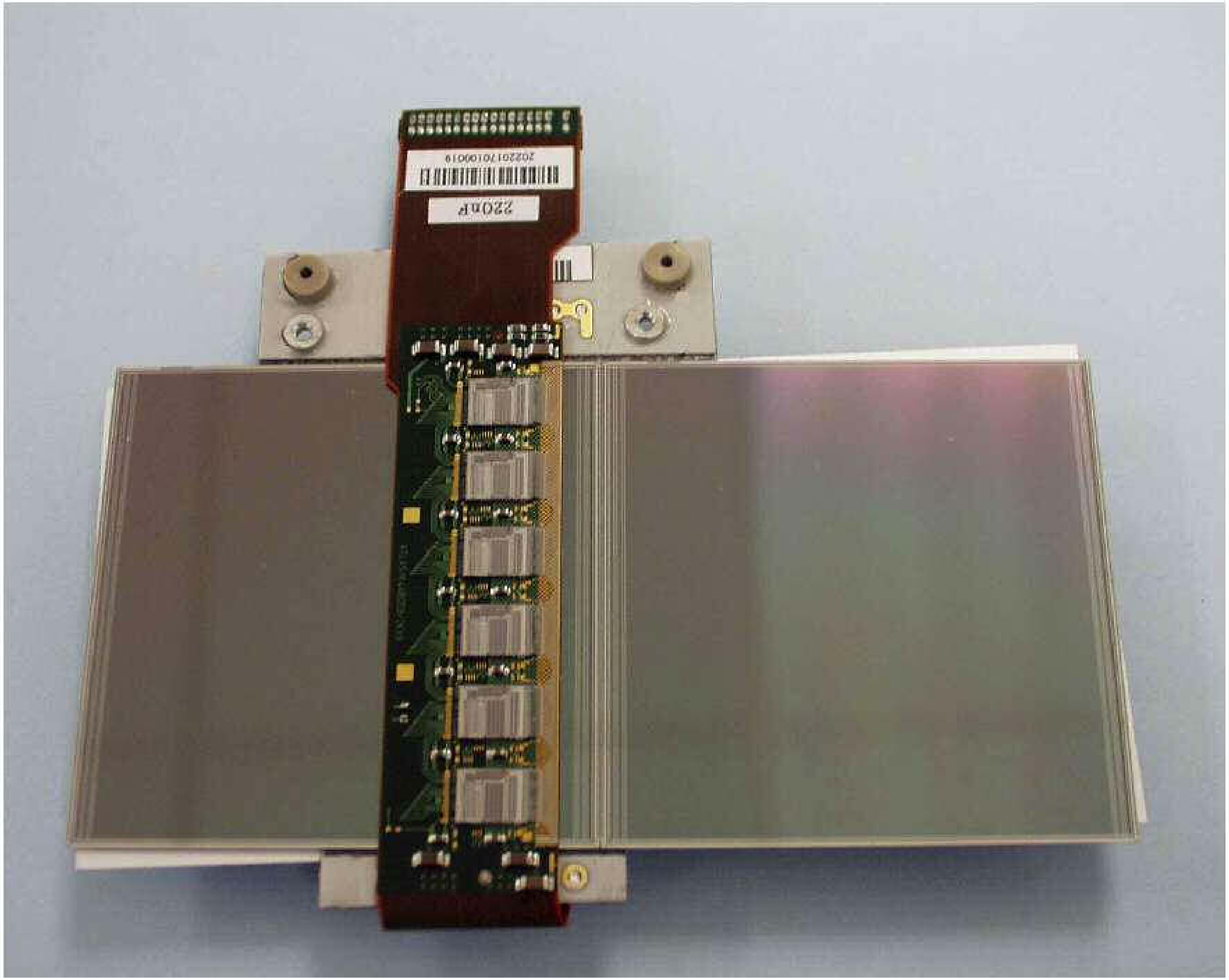
When IC3 has finished, the token is passed back to IC1, which can begin a new cycle.

ATLAS Silicon Strip system (SCT): ABCD chips mounted on hybrid

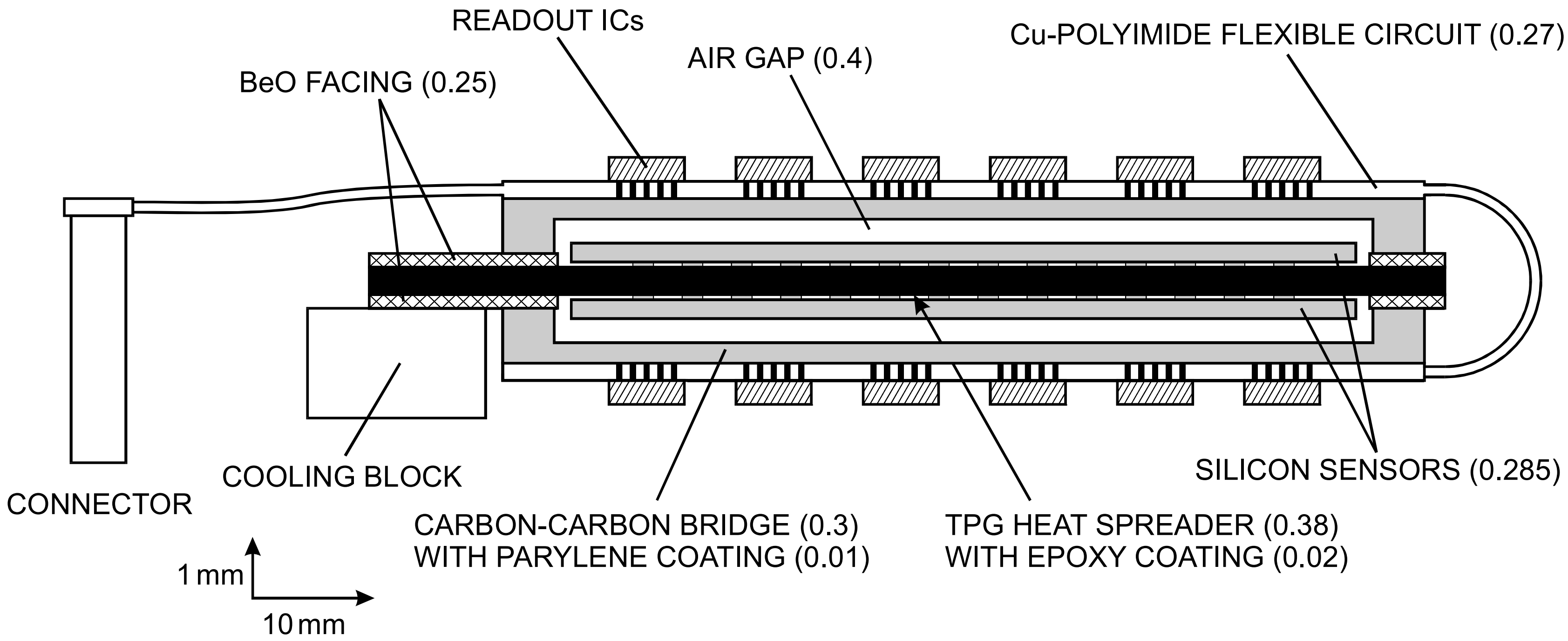


ATLAS SCT Detector Module





Cross Section of Module



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. Signal Formation

    Induced Charge

    Energy Balance Calculations
    and Failures

2. Signal Magnitude and Fluctuations

3. Signal Acquisition

    Amplifier Types

    Active Integrator –
    Charge-Sensitive Amplifiers

    Calibration

    Realistic Charge-Sensitive Amplifier

    Pulse Response

    Input Impedance of a
    Charge-Sensitive Amplifier

    Time Response of a
    Charge-Sensitive Amplifier

    Input Impedance in Strip and
    Pixel Detectors

II. Signal Formation and Acquisition

We consider detectors that provide electrical signal outputs.

To extract the amplitude or timing information the electrical signal is coupled to an amplifier, sent through gain and filtering stages, and finally digitized to allow data storage and analysis.

Optimal signal processing depends on the primary signal.

In general, the signal can be

1. a continuously varying current or voltage
2. a sequence of pulses, occurring
   - periodically
   - at known times
   - randomly

All of these affect the choice of signal processing techniques.

First steps in signal processing:

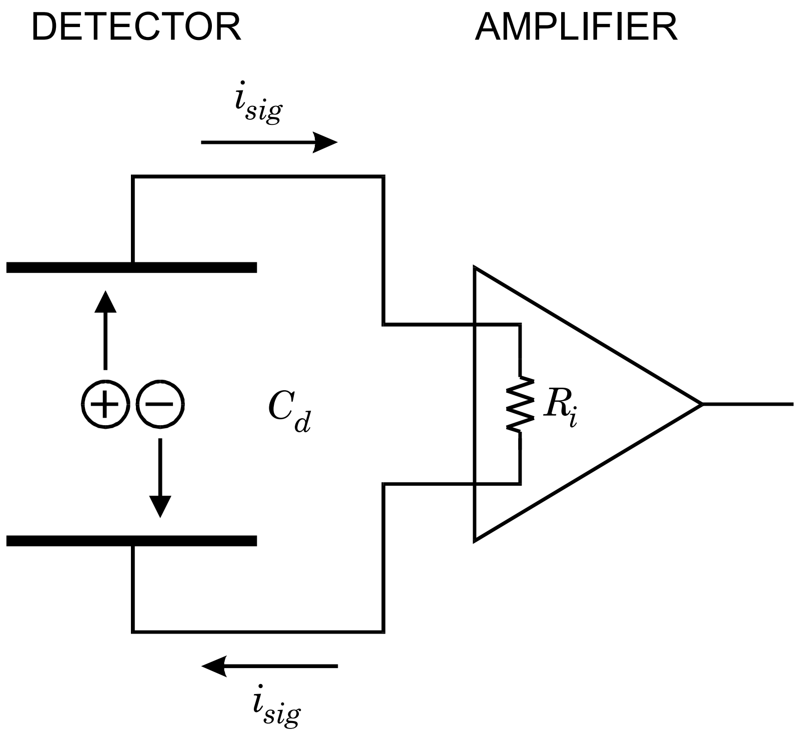
- Formation of the signal in the detector
- Coupling the sensor to the amplifier

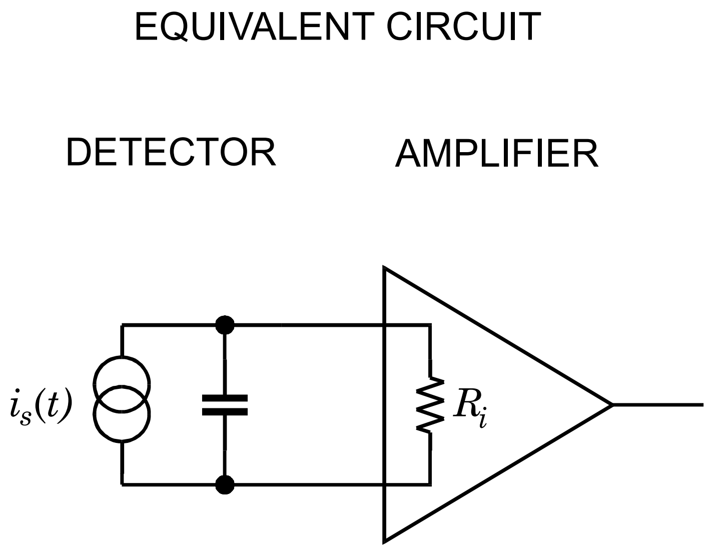
Radiation detectors use either

- direct detection (e.g. ionization chambers)

  or
- indirect detection (e.g. scintillators)

1. Signal Formation





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 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.

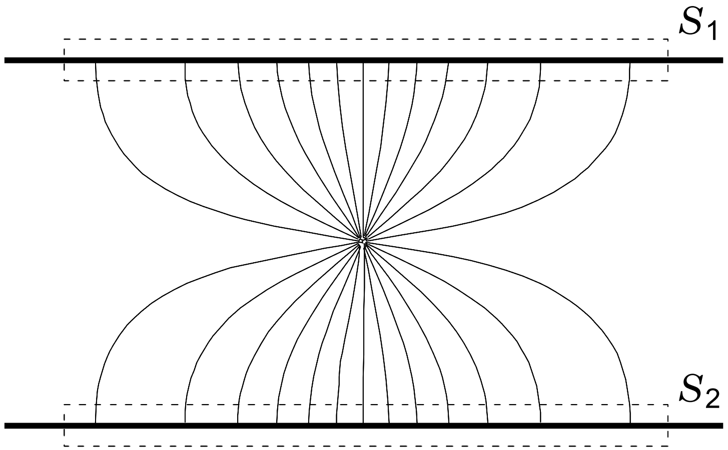
The effect of the amplifier on the signal pulse will be discussed in the Electronics part.

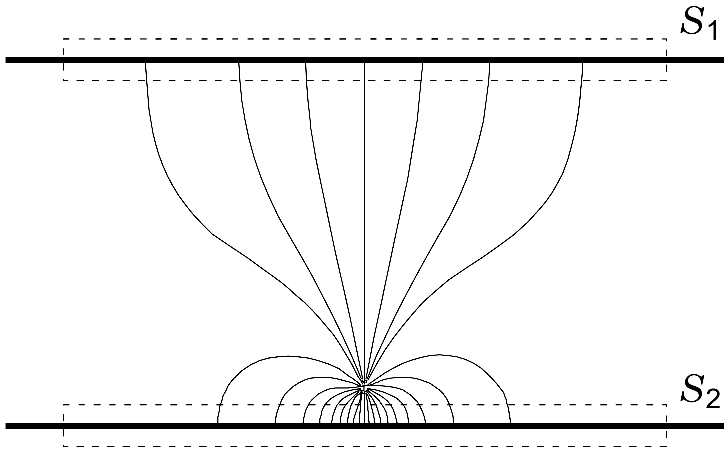
Induced Charge

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.

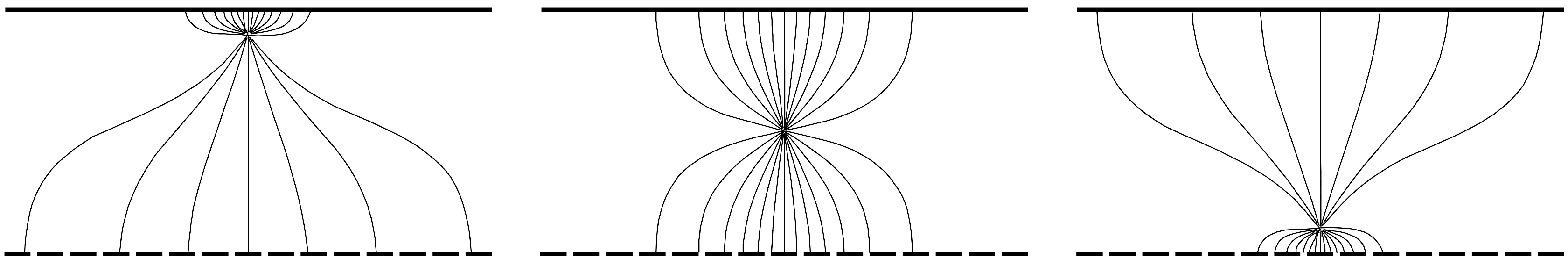




As a charge traverses the space between the two plates the induced charge changes continuously, so current flows in the external circuit as soon as the charges begin to move.

Induced Signal Currents in a Strip Detector

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



Initially, charge is induced over many strips.

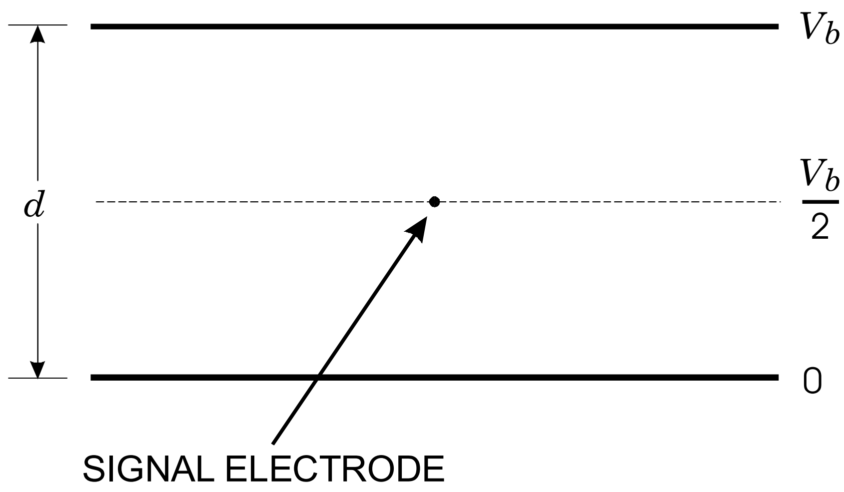
As the charge approaches the strips, the signal distributes over fewer strips.

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

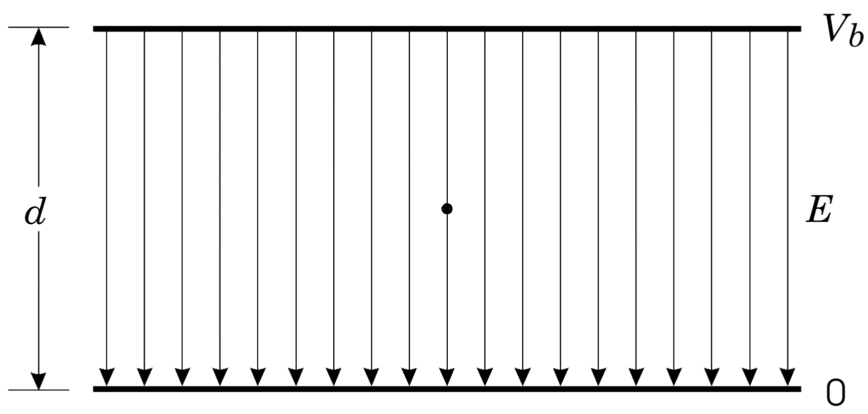
The magnitude of the induced current due to the moving charge depends on the coupling between the charge and the individual electrodes.

Quantifying Induced Current

Assume a parallel plate detector with a small diameter signal electrode in the middle



The signal electrode is biased so that the electric field is uniform throughout the active volume.



A mobile charge will move at a constant velocity

$$v = \mu E = \mu \frac{V_b}{d}$$

at any position within the active volume.

The induced current depends on

- the velocity of the moving charge
- the coupling of the moving charge to the signal electrode

The coupling to the signal electrode is determined by applying a unit charge to the signal electrode and determining the field F_S.

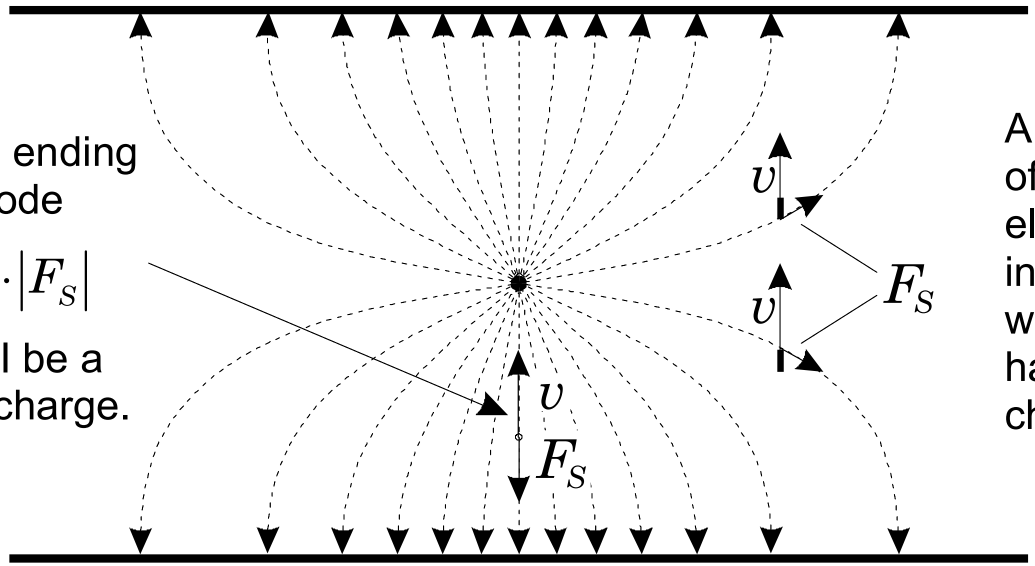
The induced current for a moving charge q is $\quad i_S = q \cdot \underbrace{\vec{v} \cdot \vec{F}_S}$

The magnitude of the dot product sets the current.

For a charge ending on the electrode

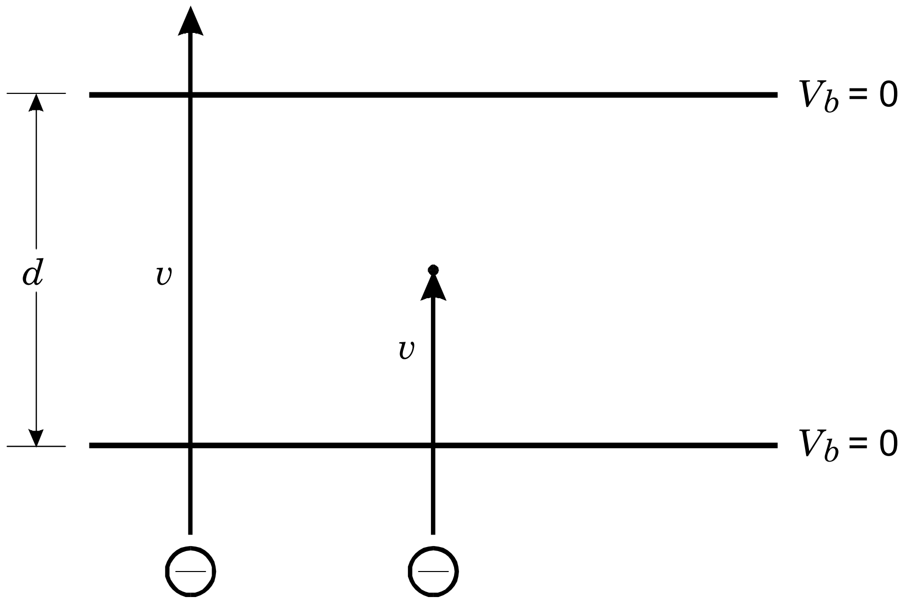
$\vec{v} \cdot \vec{F}_S = -|v| \cdot |F_S|$

and there will be a net induced charge.



A charge moving along off to the side of the electrode will initially induce a current, which will invert in the upper half. The total induced charge will be zero.

Note: The bias voltage is not a key component in signal formation. The signal derived above will be the same if electrons are injected from the outside.



The key parameters are the carrier's velocity + path and the electrode geometry.

The more detailed derivation resulting in a simple application technique was published by Ramo (Proc. IRE 27 (1939) 584-585). Also see Spieler, Chapter 2, pp 71-82.

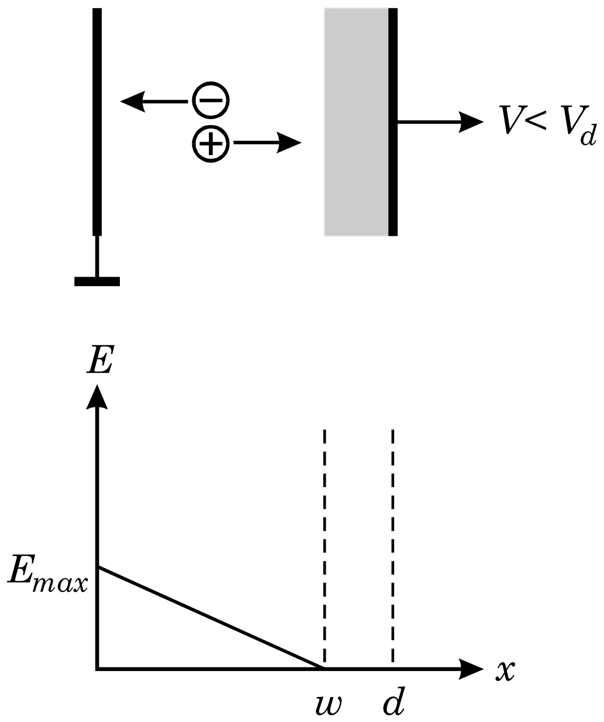
"Ramo's theorem" is a direct derivation from Maxwell's equations. Calling it a theorem does not make it a speculative recipe, as is the case for some theories.

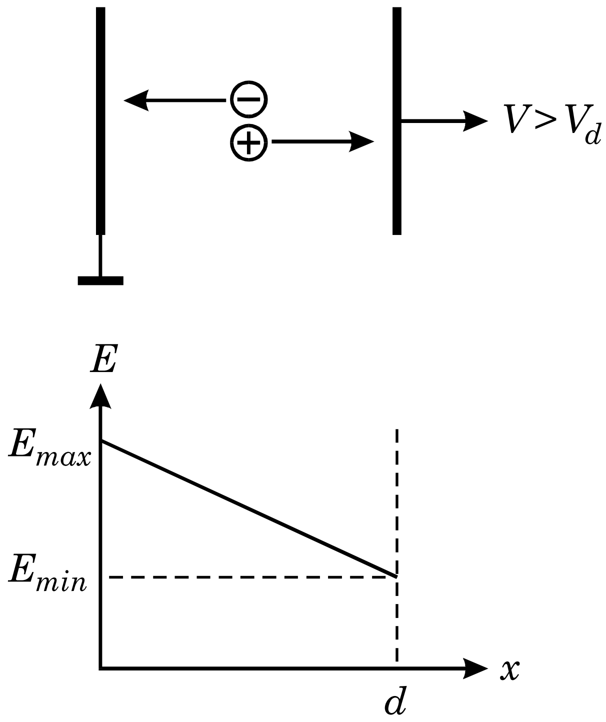
Signals from a silicon pad detector

Electric Field

Partial Depletion

Overdepletion



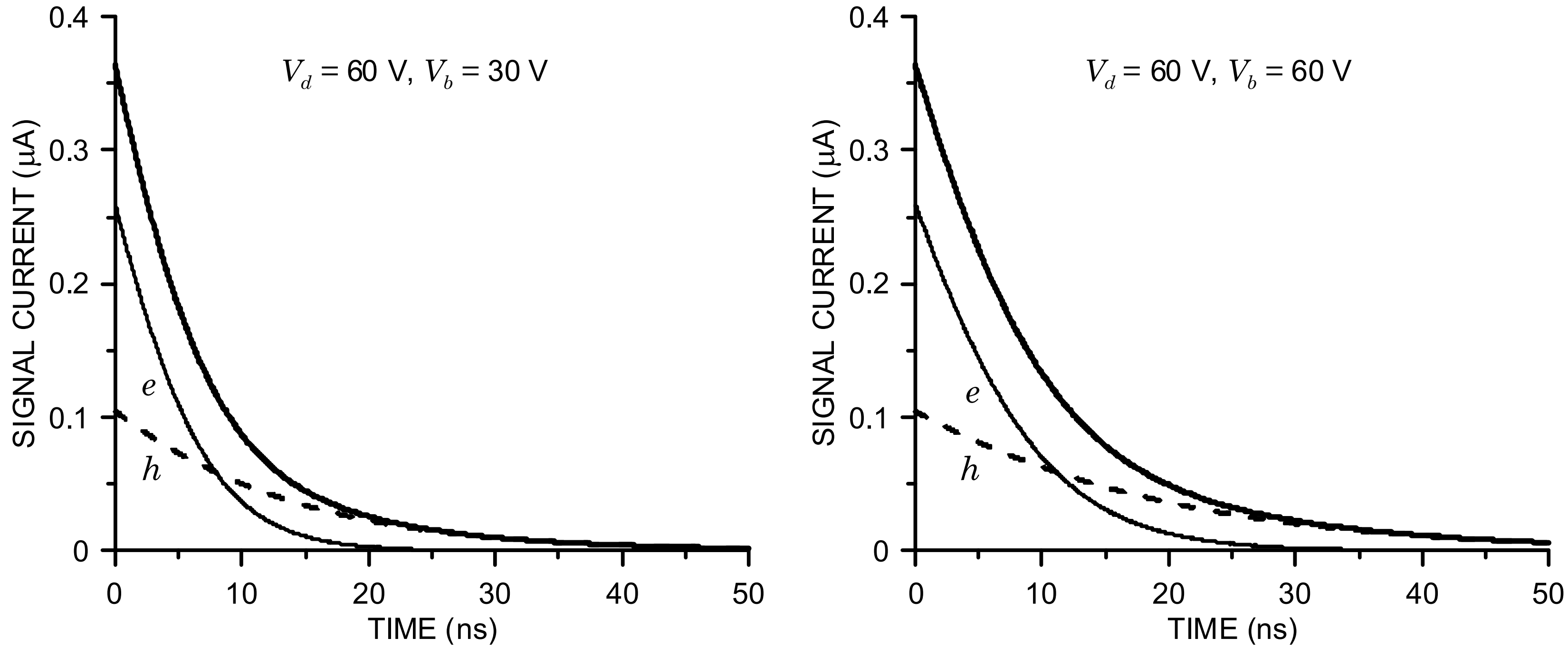


In a parallel plate configuration the weighting field is uniform throughout the active region:

$F_S = \dfrac{1}{w}$ in partial depletion and $F_S = \dfrac{1}{d}$ in full or overdepletion.

Current pulses in pad detectors (track traversing the detector)

Pulses in partial and full depletion

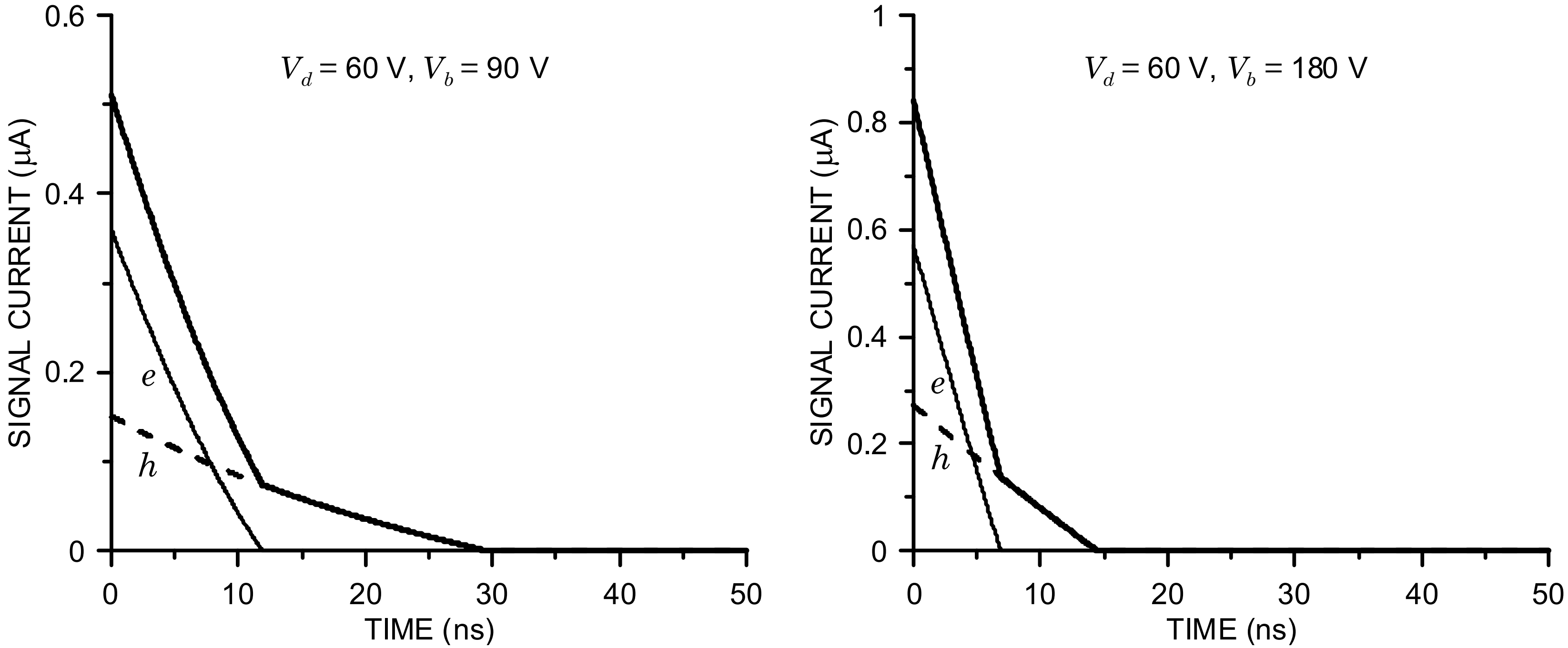


In both partial and full depletion the pulse shapes and durations are nearly the same.

As the depletion width increases, the electric field increases correspondingly.

Current pulses in pad detectors (track traversing the detector)

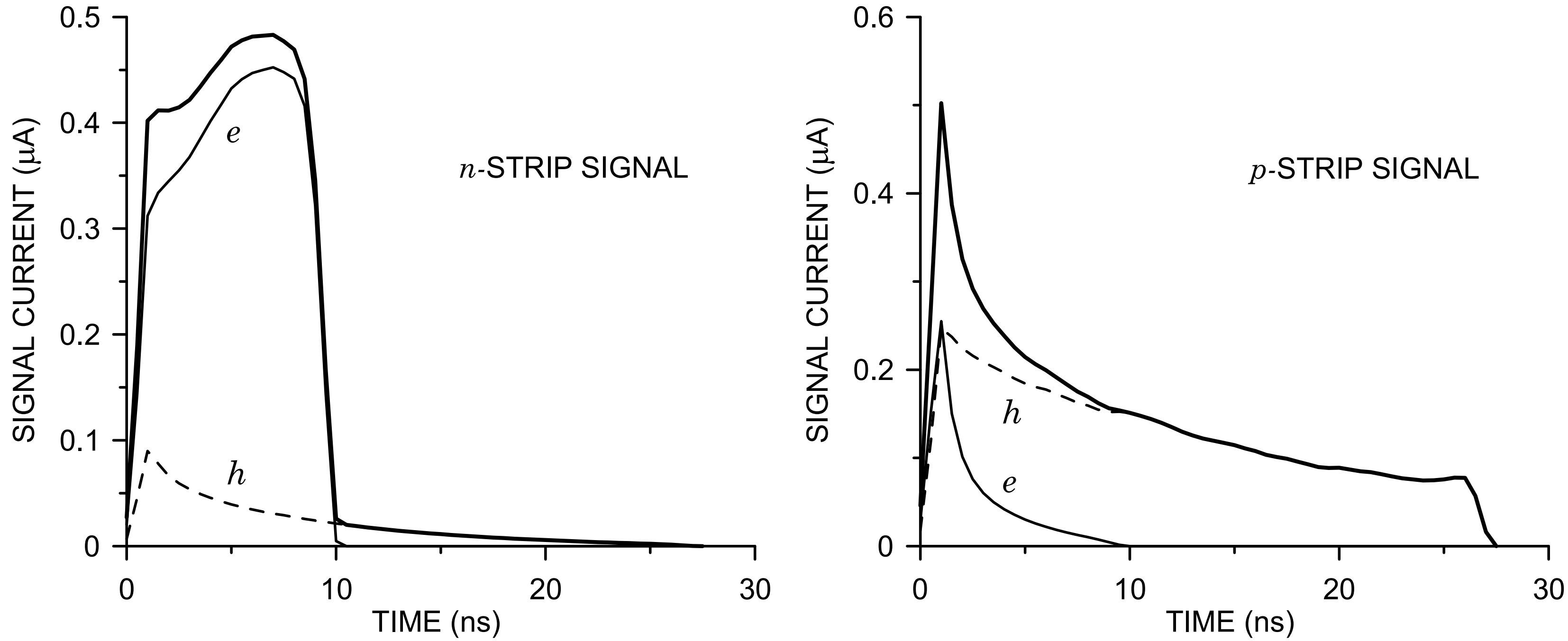
Pulses in overdepletion



Overdepletion increases the electric field in the whole active volume, so increasing the overbias significantly reduces the collection time.

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

Current pulses in strip detectors (track traversing the detector)
(V_d = 60V, V_b = 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.

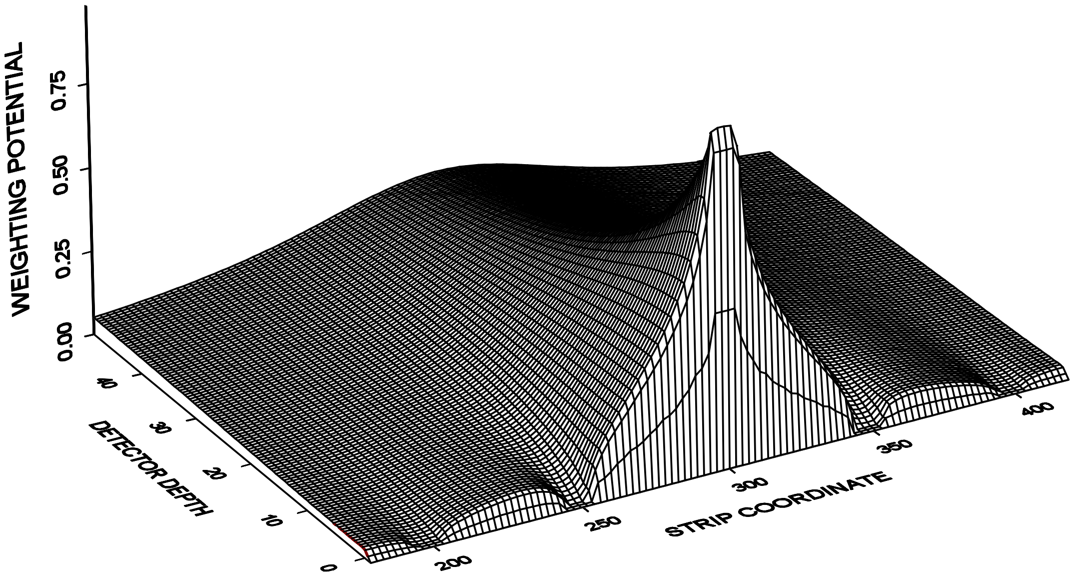
This is because in segmented detectors the weighting field is very different.

Weighting field in a strip detector

The strip pitch is assumed to be small compared to the thickness.

The electric field is similar to a parallel-plate geometry, except in the immediate vicinity of the strips.

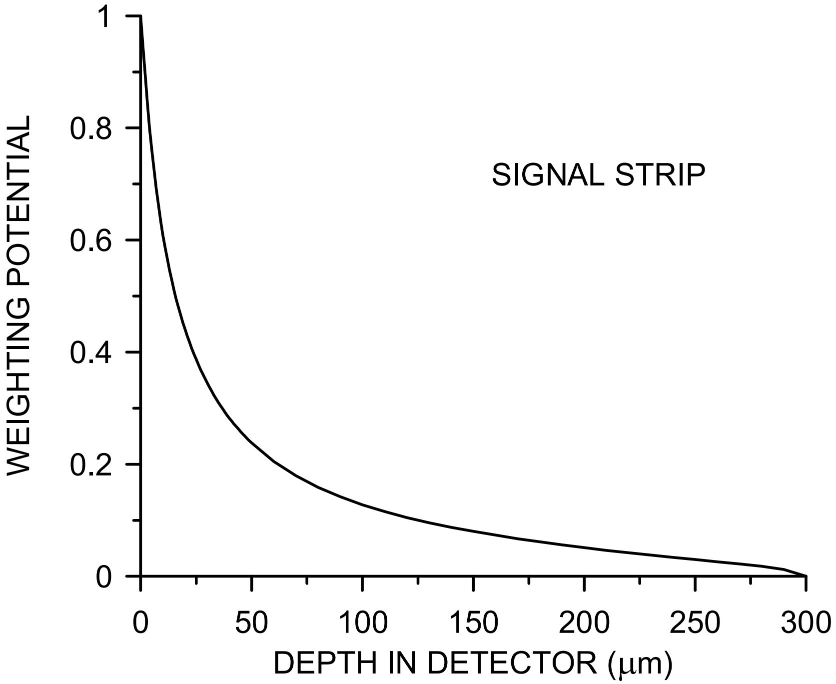
The signal weighting potential, i.e. the integral of the weighting field, however is very different.

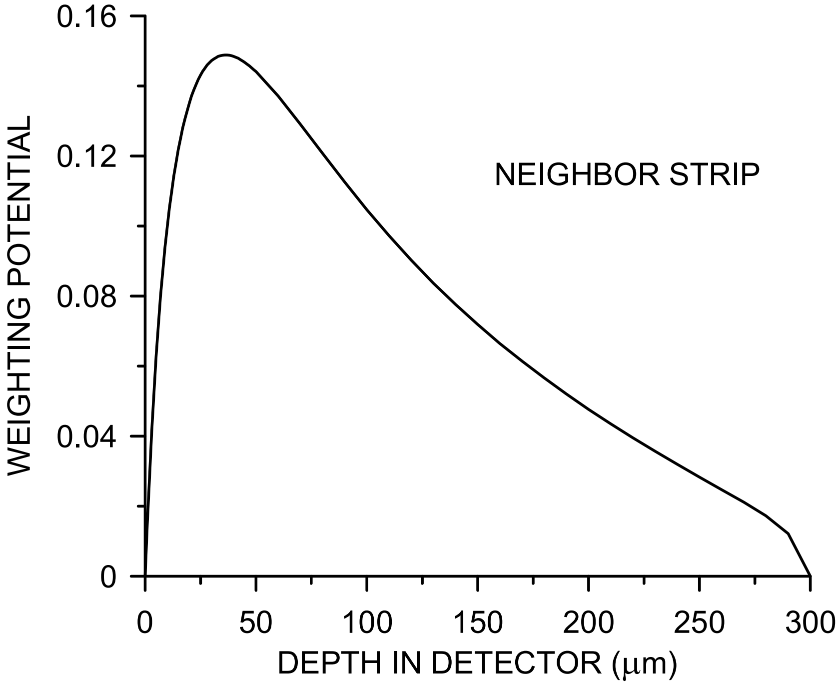


Weighting potential for a 300 μm thick strip detector with strips on a pitch of 50 μm. Only 50 μm of depth are shown. Most of the induced charge occurs near the strip electrodes.

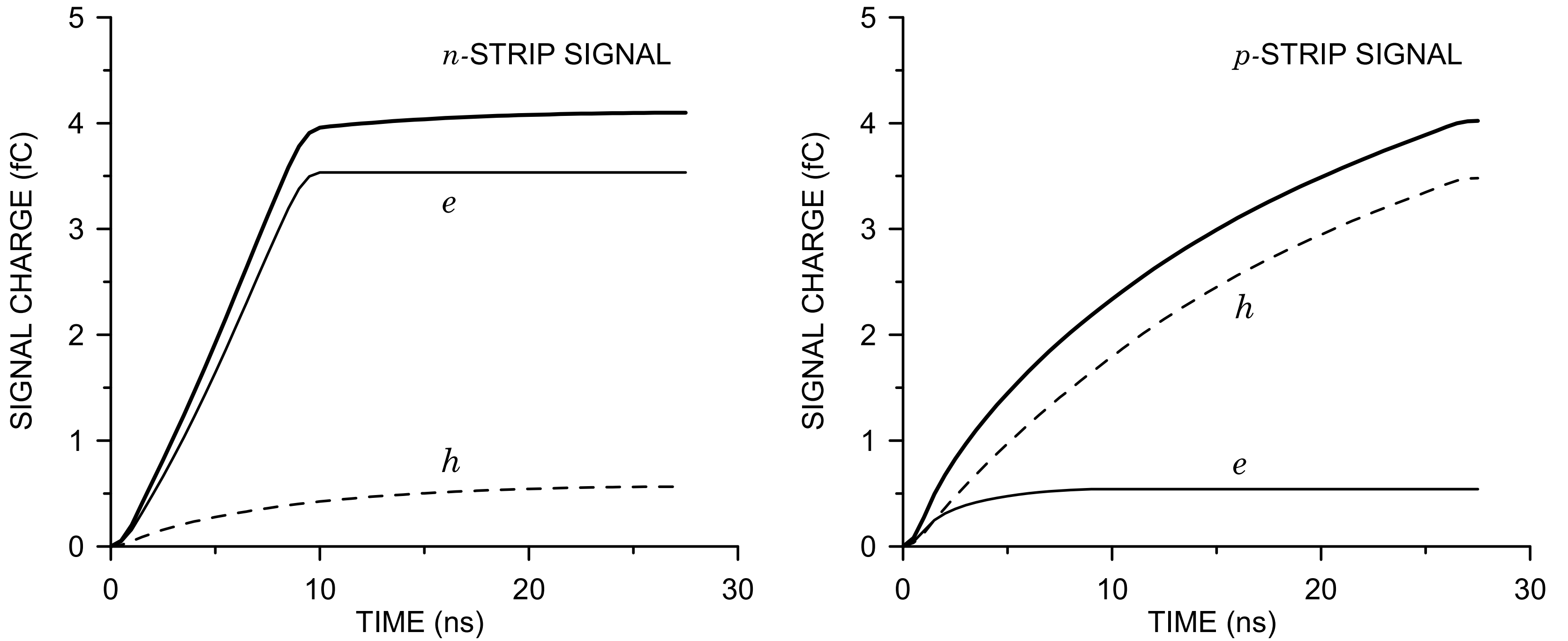
Cuts through the weighting potential

The relative induced charge for a signal charge moving from one depth to another is determined by the difference in weighting potentials at the corresponding coordinates.





Strip Detector Signal Charge



Energy Balance Calculation

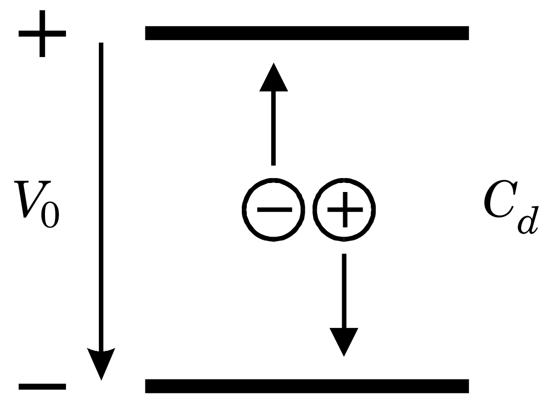
A popular technique for calculating the signal charge applies energy conservation.

Some references:

Dan Green, *The Physics of Particle Detectors*, Cambridge University Press, 2000
Konrad Kleinknecht, *Detectors for Particle Radiation*, Cambridge University Press, 1998
Glenn F. Knoll, *Radiation Detection and Measurement*, Wiley, 2000

Assume a detector that is completely disconnected, so it is a charged capacitor where the total available energy is stored in the electric field.

$$U = \frac{1}{2} C V^2$$



A charge driven by the electric field will extract the required energy from the total stored energy.

Assume that an additional signal charge dQ is induced. This will change the voltage by dQ/C and the energy

$$U = \frac{1}{2}C\left(V_0 + \frac{dQ}{C}\right)^2 = \frac{1}{2}C\left(\frac{Q_0}{C} + \frac{dQ}{C}\right)^2 = \frac{1}{2C}\left(Q_0 + dQ\right)^2 \approx \frac{1}{2C}\left(Q_0^2 + 2Q_0 dQ\right).$$

Thus the change in energy stored in the electric field for a change in signal charge

$$dU = Q_0 \frac{dQ(t)}{C}$$

The field $E(x)$ imparts a force $F(x)$ on the mobile signal charge, so it will gain the energy $F(x)dx$ and change the energy stored in the field by

$$dU = Fdx = qE(x)dx = qE(x)v(x)dt\,,$$

and change the electrode charge by

$$dQ = \frac{C}{Q_0} qE(x)v(x)dt = \frac{q}{V_0} E(x)v(x)dt\,.$$

Hence, the instantaneous signal current

$$i(x) = \frac{dQ}{dt} = \frac{q}{V_0} E(x) v(x) .$$

The signal charge

$$Q_s = \int i(x) dt = \frac{q}{V_0} \int E(x) v(x) dt .$$

At constant velocity $dx = v dt$, so

$$Q_s = \frac{q}{V_0} \int E(x) v dt = \frac{q}{V_0} \int E(x) dx = q \frac{\Delta V}{V_0} .$$

Assume a constant field V_0 / d and a charge traversing the detector thickness d at constant velocity. Then $\Delta V = V_0$ and $Q_s = q$.

For a charge traversing a fraction of the active width x / d, then for a constant field $\Delta V / V_0 = x / d$, so

$$Q_s = q \frac{x}{d} .$$

This agrees with the induced charge technique, but required a constant field V_0 / d and constant carrier velocity. **This is not a general result!**

Silicon Pad Detector

Depletion voltage= 60V
Bias voltage= 90V

$\Rightarrow$ sloping field

Point deposition: $x_0 = d/2$

Induced Charge Calculation

Electron contribution =
Hole contribution

Independent of field profile!

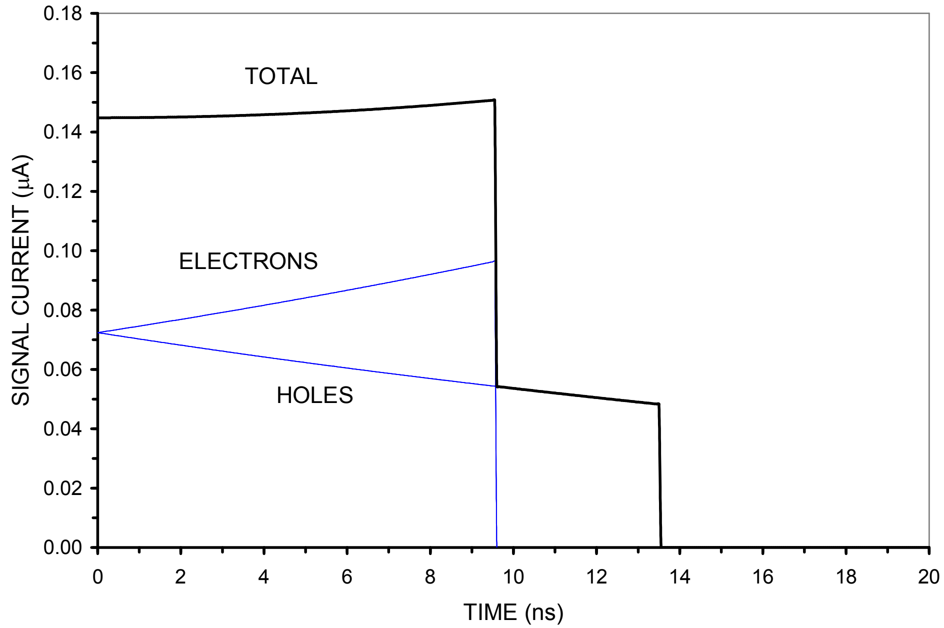
Energy Balance Calculation

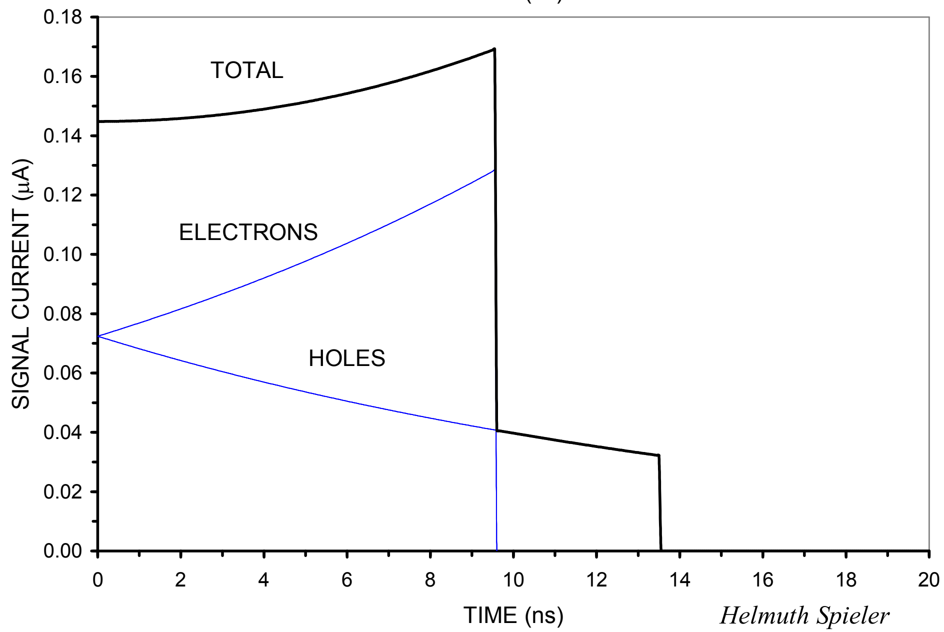
Electron contribution greater than from holes (58% vs. 42%)

At 120V bias: 63% vs. 37%.

In reality the induced charge ratio is independent of bias.

The induced charge calculation gives the correct result.



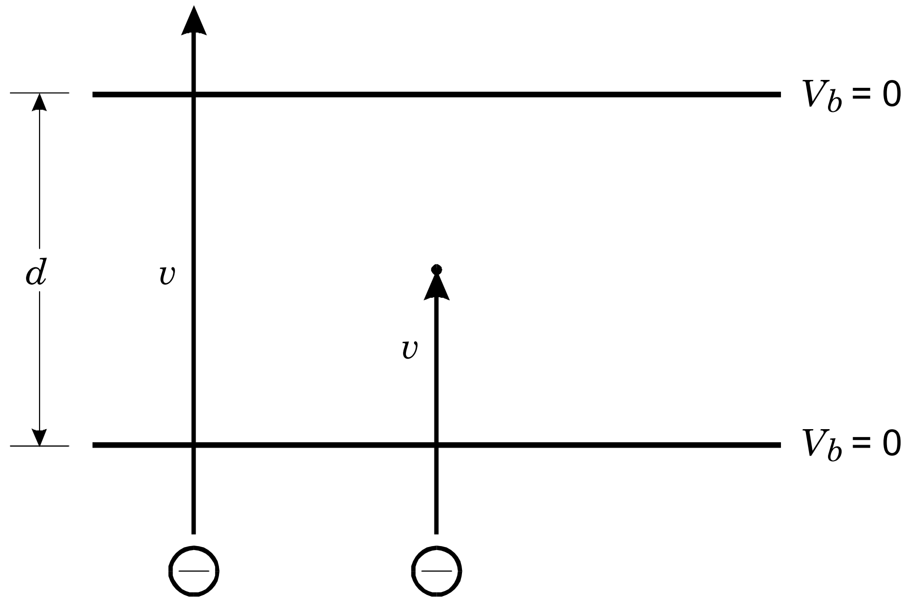


Consider the previous example with zero field:

According to the energy balance derivation

$$i(x) = \frac{q}{V_0} E(x) v(x),$$

so with zero field there is no signal.



Take it a step further and assume operating in a vacuum.

Then the left hand electron will not deposit any energy, so no energy conservation scheme will yield a signal.

The right hand electron ending on the electrode will create a signal because of the deposited charge, but this has nothing to do with energy conservation.

Obviously, considering “energy balance” to be a generally applicable technique is overly optimistic.

What’s wrong?

The energy balance calculation yielded the result

$$Q_s = q\frac{x}{d}\,,$$

which appears to agree with the induced charge result. However, energy balance assumes special conditions, a constant field V_0/d and constant carrier velocity.

This is hardly ever the case and not a general result!

Furthermore, detector electronics commonly maintain constant potential because they rapidly extract the signal current, so charge doesn’t build up on the electrodes.

In calculating energy conservation one has to consider kinetic energy. The mobility limited velocity $v = \mu E$ is caused by collisions, so only a fraction of the total energy goes into the net motion. The instantaneous velocity is much greater and since the kinetic energy $E_{kin} \propto v^2$, its fluctuations do not scale the same as the velocity.

It is not clear how the energy balance approach can be applied to multi-electrode detectors. At the charge collection electrode of a strip detector it provides the same result as for a pad detector, which is totally wrong.

Claiming that a technique is generally correct because in a few specific cases it appears to provide the correct result is rather naïve.

Quote from

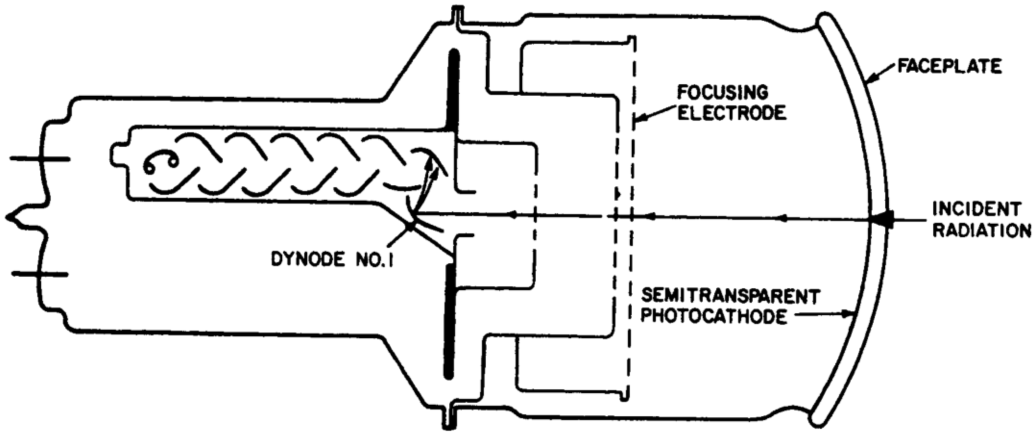
Dan Green, *The Physics of Particle Detectors*, Cambridge University Press, 2000

to support the energy conservation derivation:

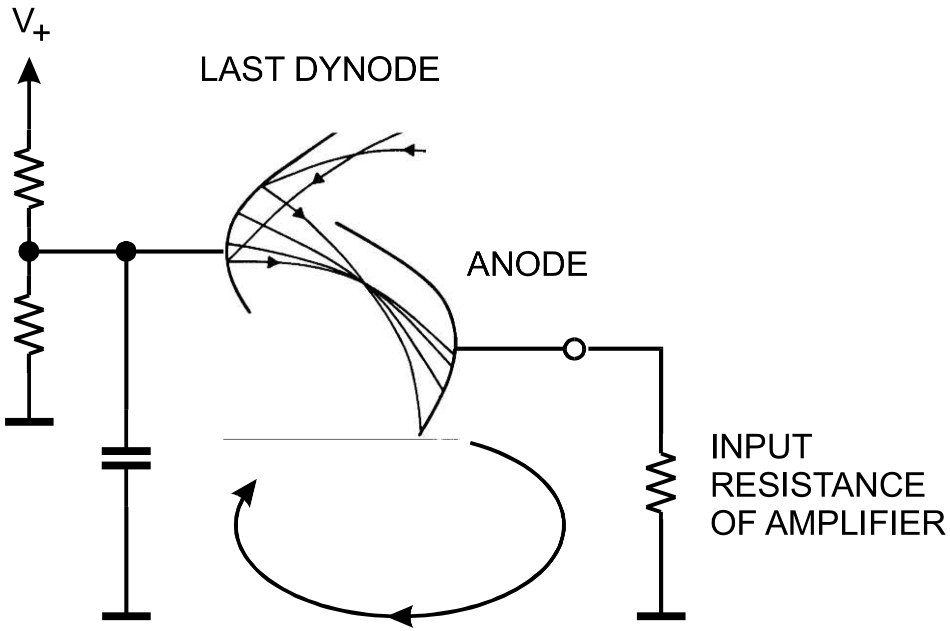
"Note that this treatment is quite general, having used only energy conservation. Thus it can be used … in the discussion of wire chambers and silicon detectors."

- Indeed it can – and provide the wrong results.
- Applying energy conservation requires an understanding of all physical processes that are contributing.
- In general, one should begin by identifying the basic physics interactions, i.e. the interaction of a moving charge by its electric field.
- The detailed calculations may be quite complex, but attempting to circumvent full physics understanding by applying some overall rule does not always solve the problem.
- The title of a book or paper does not ensure what is actually done.
- Claiming "physics" does not guarantee science!

Induced current is a common phenomenon, e.g. in a photomultiplier tube

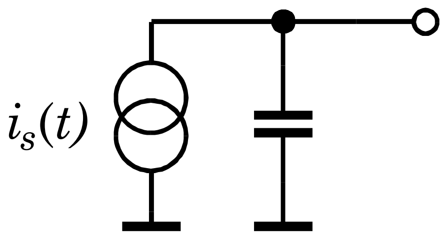


Detail of output circuit



Output Current Loop

Model



The closed current path from the last dynode to the anode must be well configured.

2. Signal Magnitude and Fluctuations

Any form of elementary excitation can be used to detect the radiation signal.

An electrical signal can be formed directly by ionization.

Incident radiation quanta impart sufficient energy to individual atomic electrons to form electron-ion pairs (in gases) or electron-hole pairs (in semiconductors and metals).

Other detection mechanisms are

Excitation of optical states (scintillators)

Excitation of lattice vibrations (phonons)

Breakup of Cooper pairs in superconductors

Formation of superheated droplets in superfluid He

The number of Signal Quanta for a signal energy E: $N_{SQ} = \frac{E}{E_{SQ}}$

Typical excitation energies E_{SQ}

Ionization in gases	~30 eV
Ionization in semiconductors	1 – 5 eV
Scintillation	~10 – 1000 eV
Phonons	meV
Breakup of Cooper Pairs	meV

Detector Sensitivity

Example:

Ionization signal in semiconductor detectors ($E_{SQ} = E_i$)

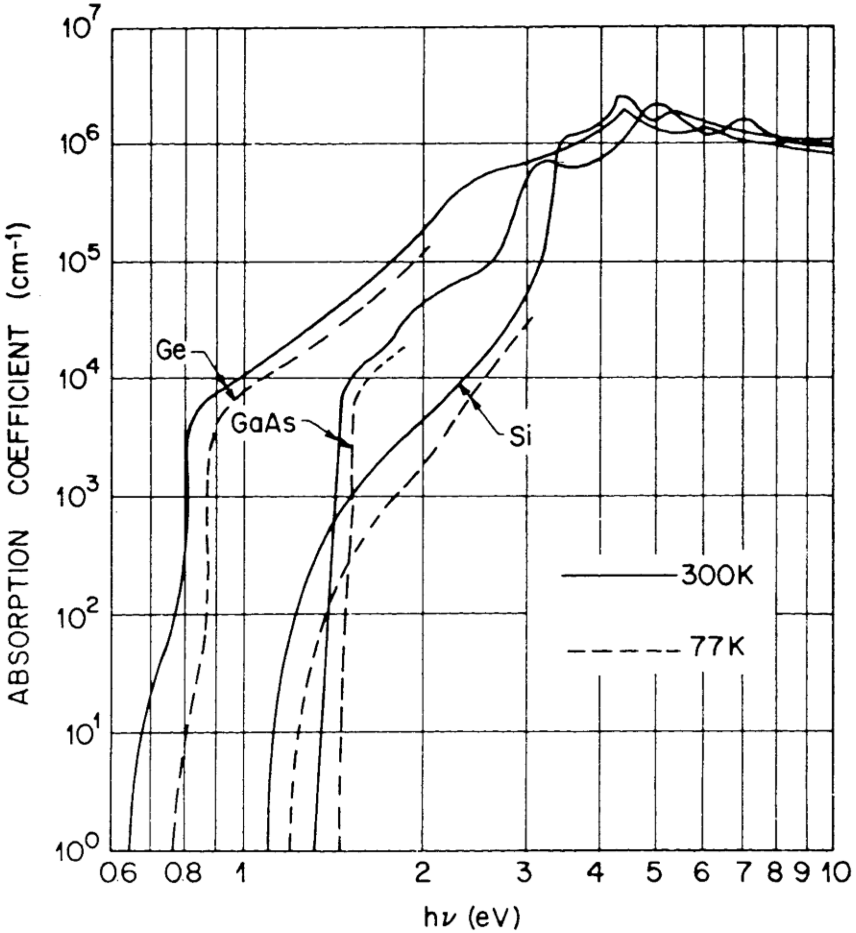
a) Visible light (energies near band gap)

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

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

For photon energies below
4 eV one electron–hole pair is formed per incident photon in Si.

The ionization energy E_i attains a peak maximum of 4.4 eV at photon energies around 6 eV and assumes a constant value above 1.5keV.



(from Sze)

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 E_g.

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

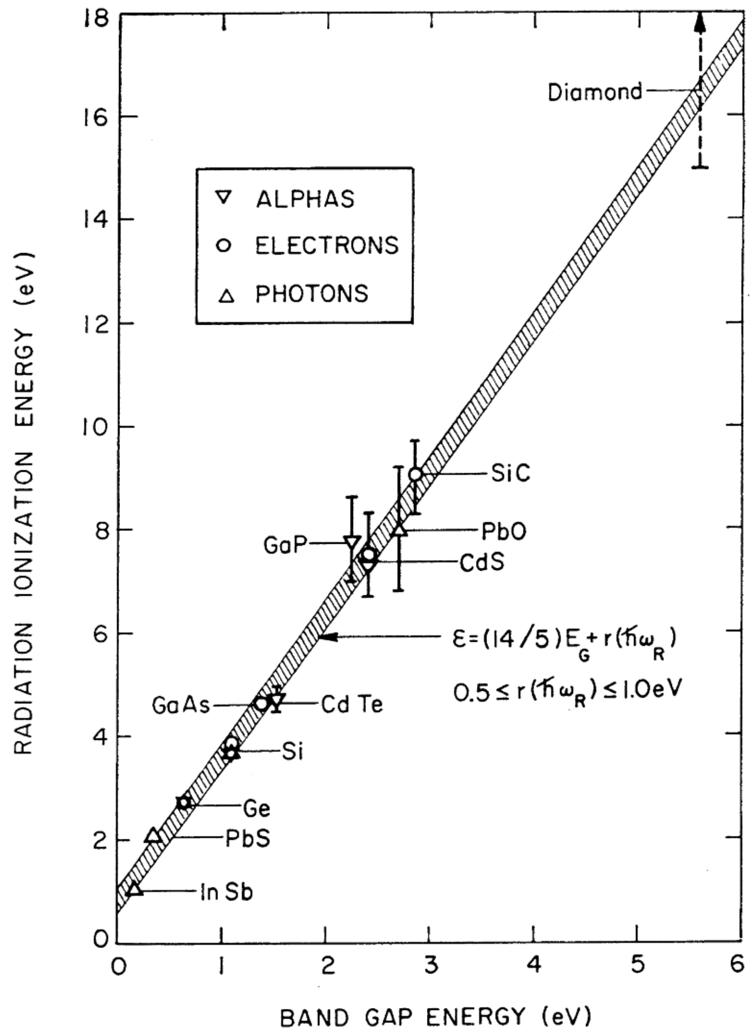
Why?

When a 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

The ratio of ionization to bandgap energy is nearly the same in all semiconductors.

Signal Fluctuations: Intrinsic Resolution of Semiconductor Detectors

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

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

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,

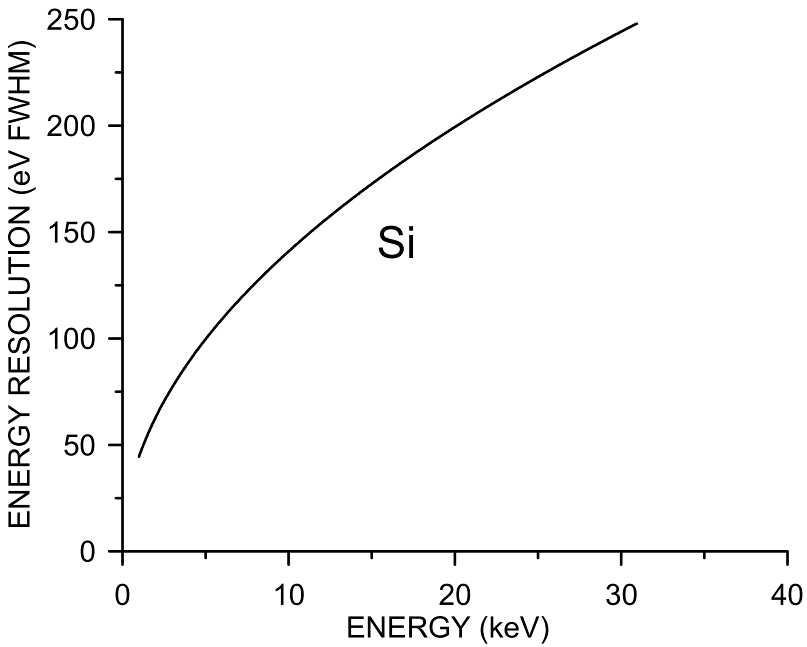
a) the fluctuation cannot exceed the absorbed energy

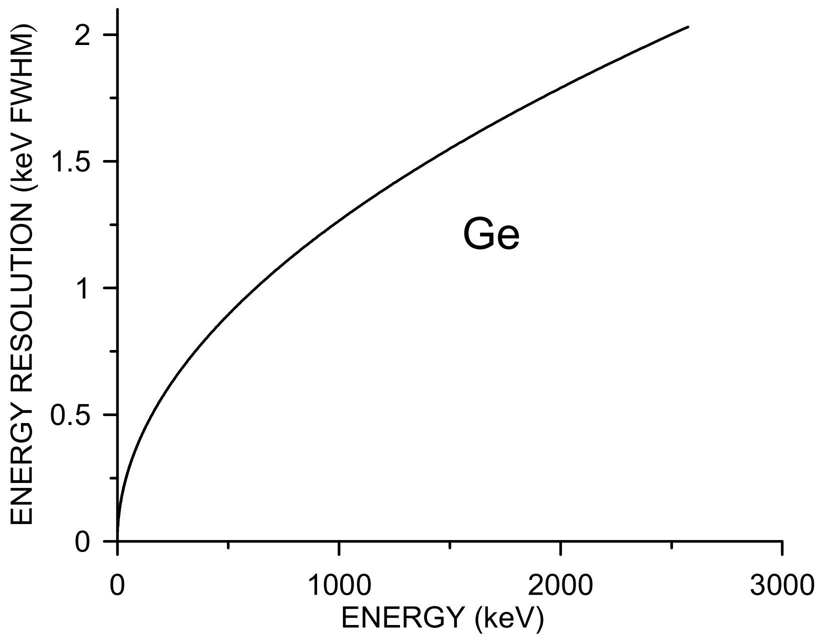
b) any fluctuation in the number of signal charges must be balanced by the fluctuation in the number of phonons. As the number of phonons is much greater, its relative variance is small and this reduces the overall fluctuations.

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 $F \approx 20$.

Inherent Detector Energy Resolution





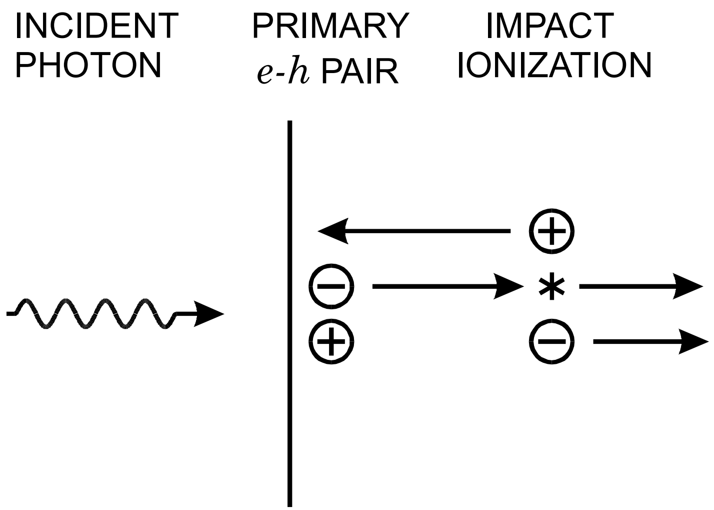
Detectors with good efficiency in the 10s of keV range can have sufficiently small capacitance to allow electronic noise of ~100 eV FWHM, so the variance of the detector signal is a significant contribution.

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

Internal Gain

For low-energy signals or detector geometries that in combination with the electronics incur a small signal-to-noise ratio, increasing the inherent detector signal can be very helpful.

By accelerating signal charges to a sufficient energy, they can collide with atoms in the base material and release secondary charges.



By increasing the width of the high-field region, multiple secondary electrons are released.

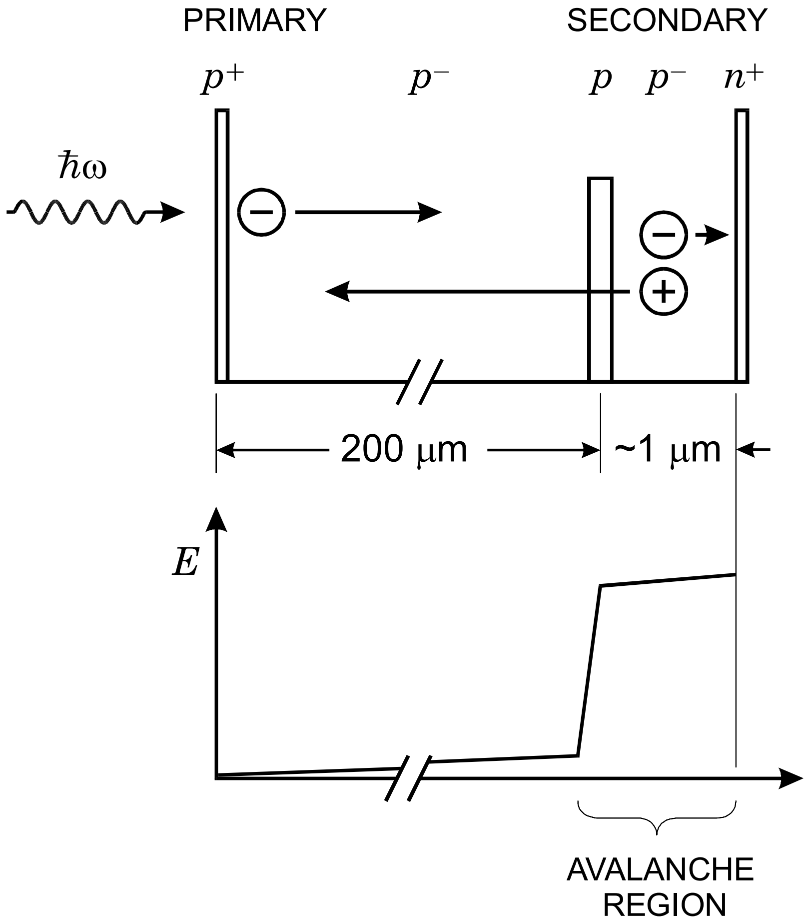
It is important that the ionization coefficient of holes is much smaller than of electrons.

If both generate secondary particles, voltage breakdown will occur at low fields.

Many conventional avalanche techniques introduce large statistical gain fluctuations. This can be controlled by choosing certain field profiles and limiting the gain.

Moderate gain of 5 or 10 can lead to significant improvements in energy resolution.

Example of a low-fluctuation avalanche gain detector



The desired size and parameters of the avalanche region must be compatible with practical fabrication techniques.

Dopant distributions

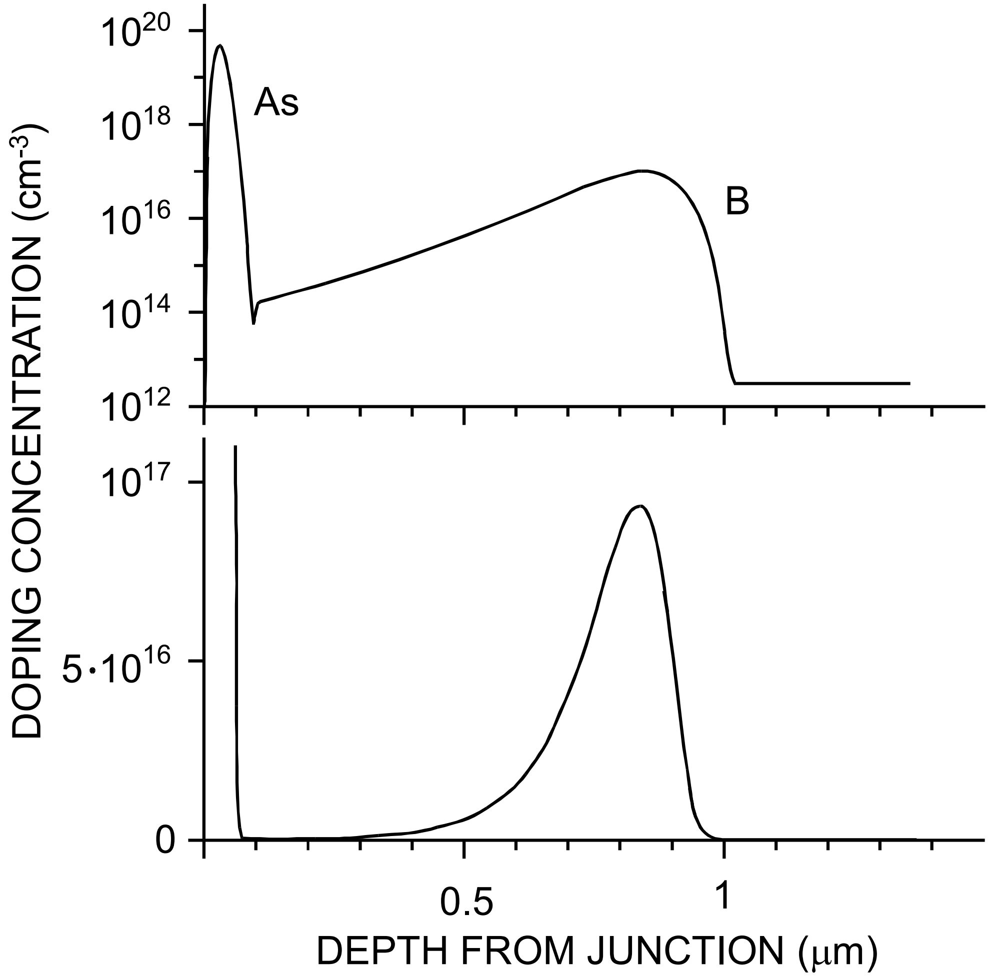
Note that the orientation is reversed with respect to the previous figure.

The boron dopant distribution expands with annealing.

Arsenic is chosen to maintain a small width.

See Spieler pp 86-91 and H.G. Spieler and E.E. Haller, IEEE Trans. Nucl. Sci. **NS-32** (1985) 419

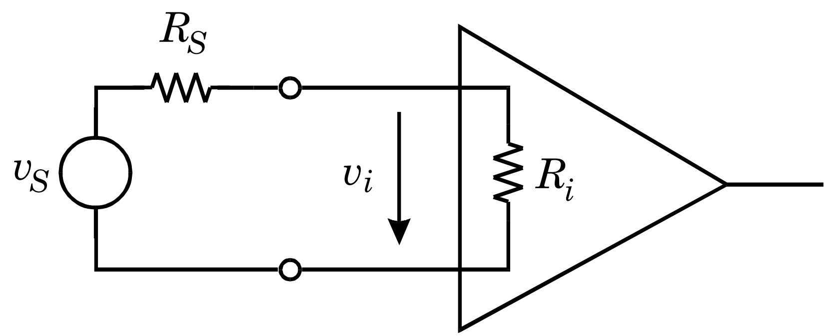
In gaseous detectors gain fluctuations can be kept low by utilizing electroluminescent gain.



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

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

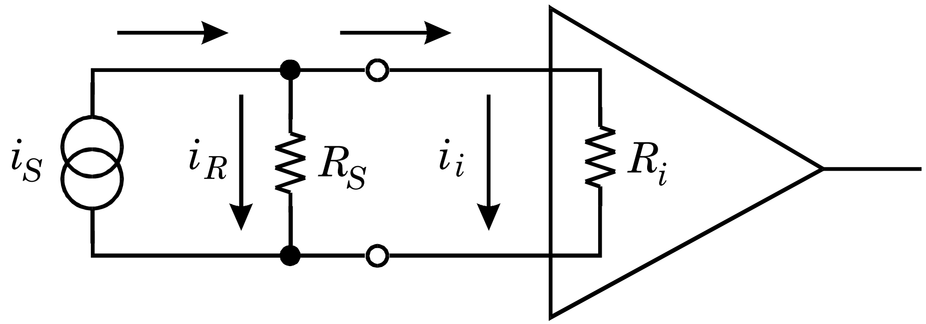
To operate in the voltage-sensitive mode, the amplifier's input resistance (or impedance) must be large compared to the source resistance (impedance).

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).

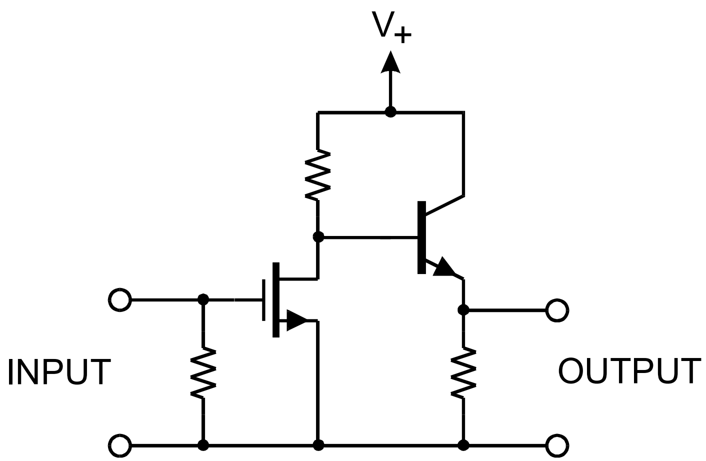
One can also model a current source as a voltage source with a series resistance. For the signal current to be unaffected by the amplifier input resistance, the input resistance must be small compared to the source resistance, as derived above.

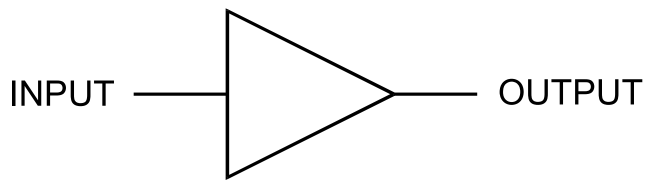
At the output, to provide current drive the output resistance should be high, i.e. large compared to the input resistance of the next stage.

- 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:

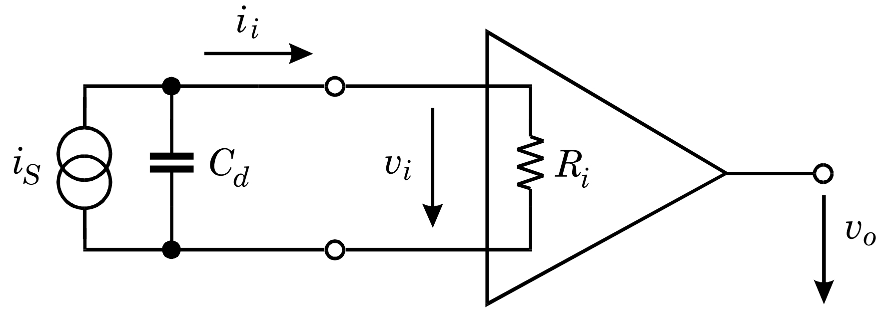




c) Voltage and Current Mode with Capacitive Sources

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

$\Rightarrow \quad v_o = A_v \cdot (Q_S / C) \propto \int i_s(t) dt$

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 (can vary with bias voltage or strip length).

Active Integrator ("charge-sensitive amplifier")

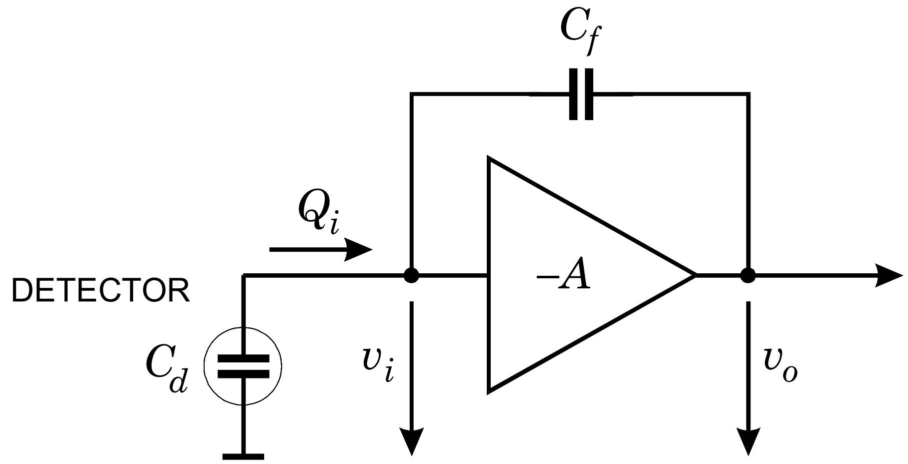
Start with an ideal inverting voltage amplifier

Voltage gain $dv_o / dv_i = -A$

$\Rightarrow \quad v_o = -A v_i$

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

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

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

What fraction of the signal charge is measured?

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

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

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

$\uparrow$

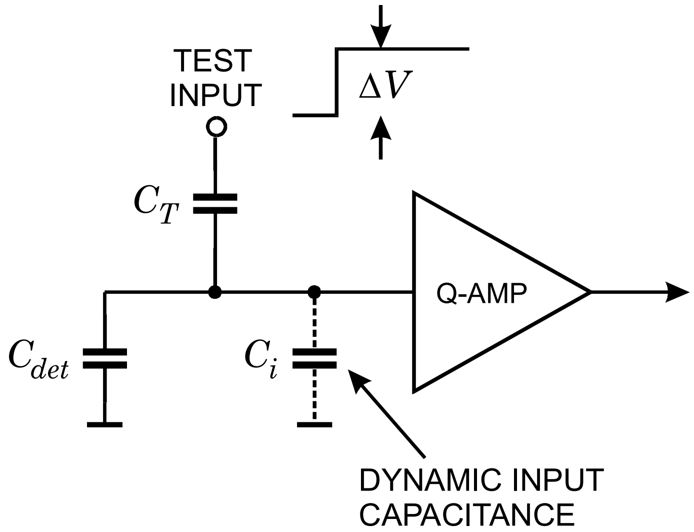
Si Det.: 50 μm thick, 250 mm^2 area

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.

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

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

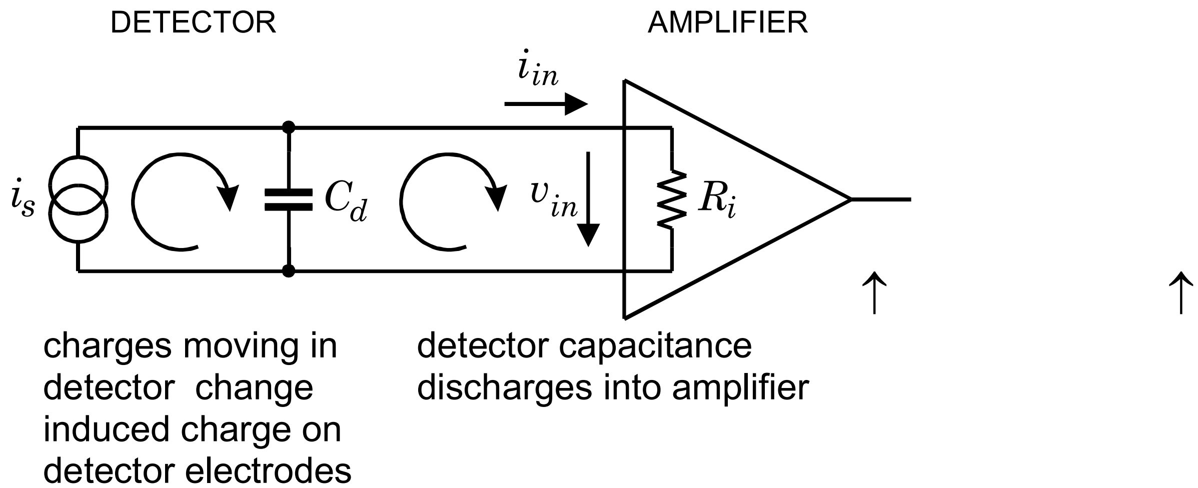
Realistic Charge-Sensitive Preamplifiers

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:

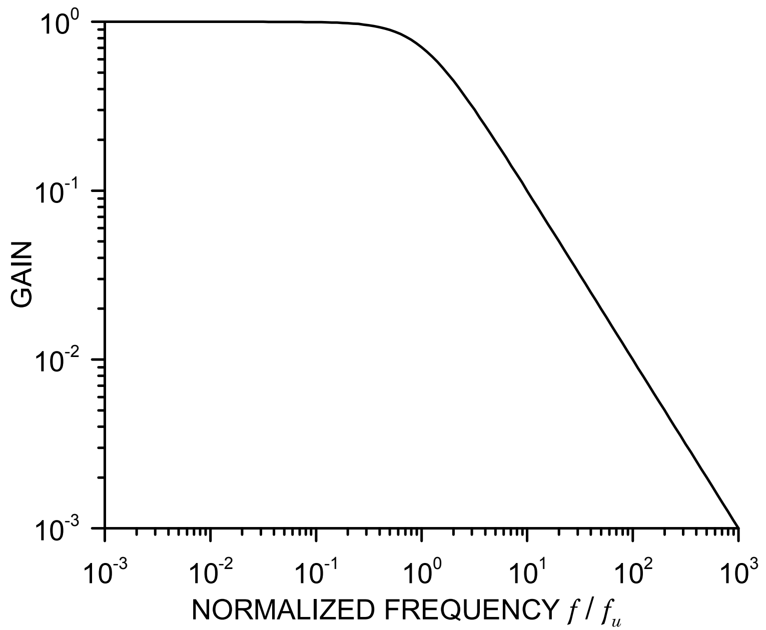


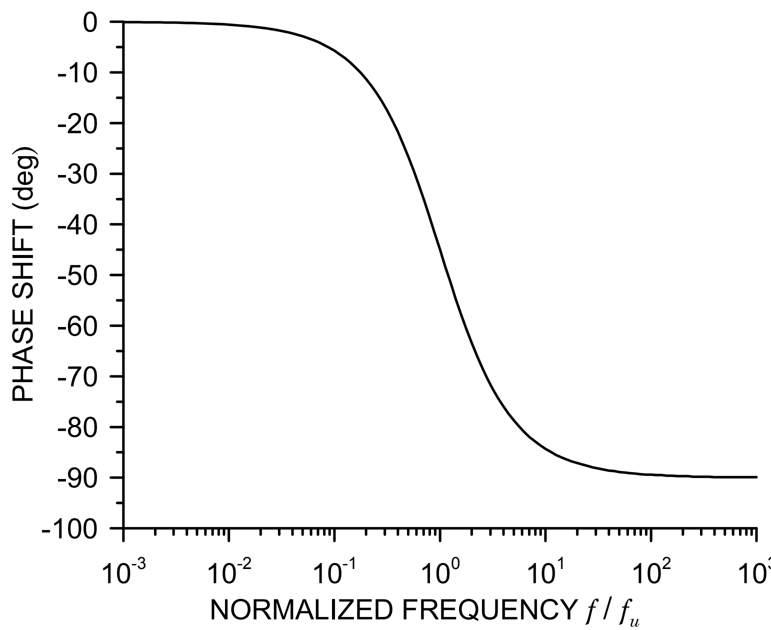
Signal is preserved even if the amplifier responds much more slowly than the detector signal.

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?

Frequency and phase response of a simple amplifier:





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

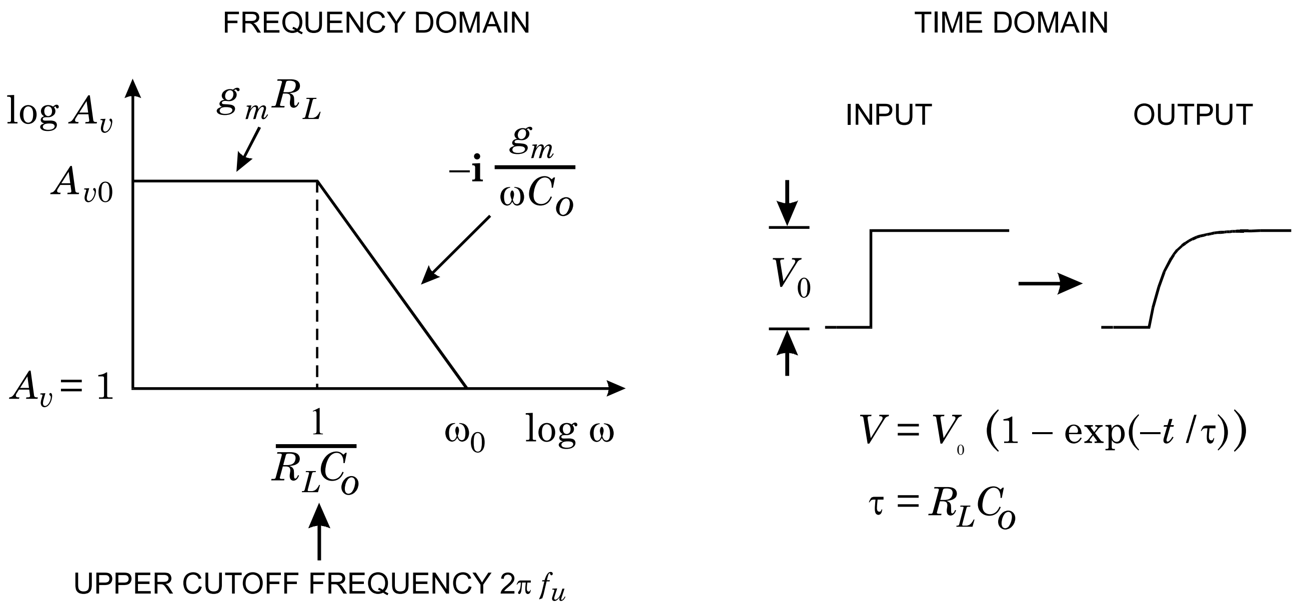
The corner (cutoff) frequency is often called a “pole”.

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 $\tau = R_L C_O$ (R_L is the load resistance)



The imaginary term above the cutoff frequency indicates the 90° phase shift.

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

Input Impedance of a Charge-Sensitive Amplifier

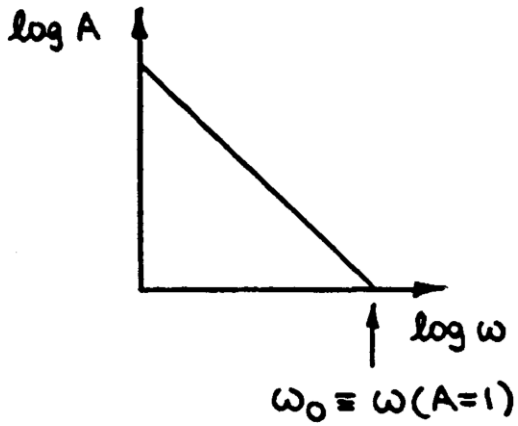
Input impedance $Z_i = \frac{Z_f}{A+1} \approx \frac{Z_f}{A} \quad (A >> 1)$

Amplifier gain vs. frequency beyond the upper cutoff frequency

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

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

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



Gain-Bandwidth Product

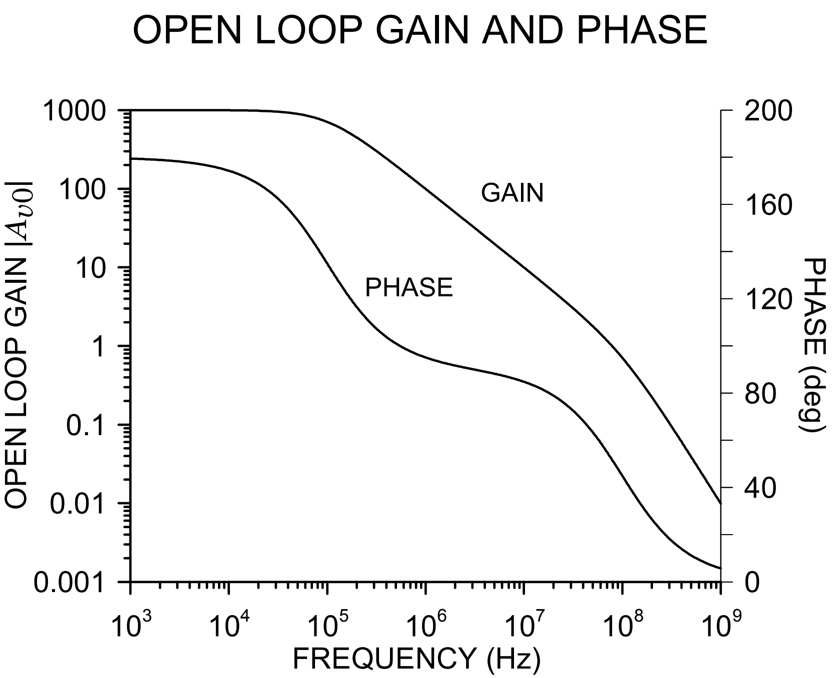
Imaginary component vanishes $\Rightarrow$ *Resistance:* $Z_i \rightarrow R_i$

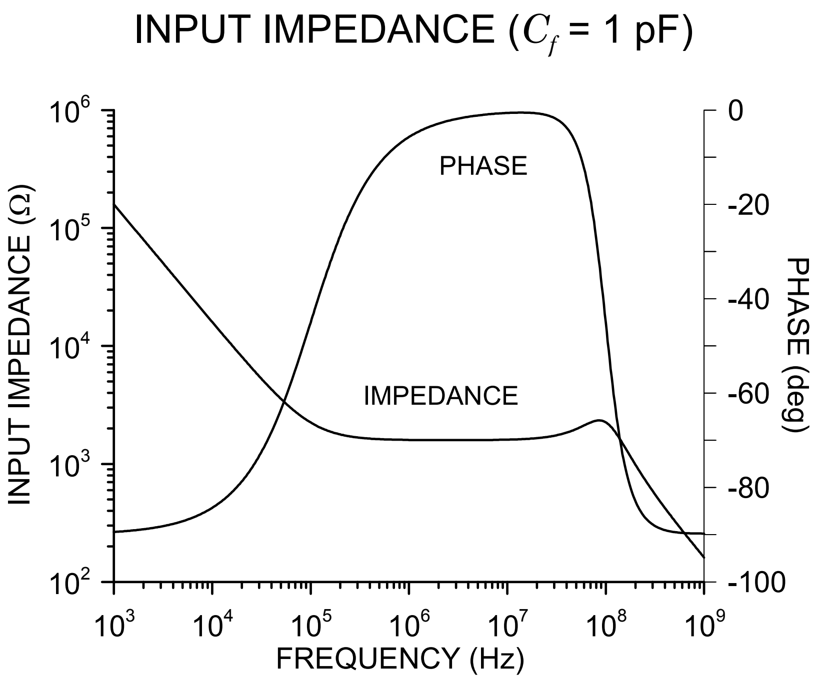
$\Rightarrow$ low frequencies ($f < f_u$): capacitive input
high frequencies ($f > f_u$): resistive input

Practically all charge-sensitive amplifiers operate in the 90° phase shift regime.
$\Rightarrow$ 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



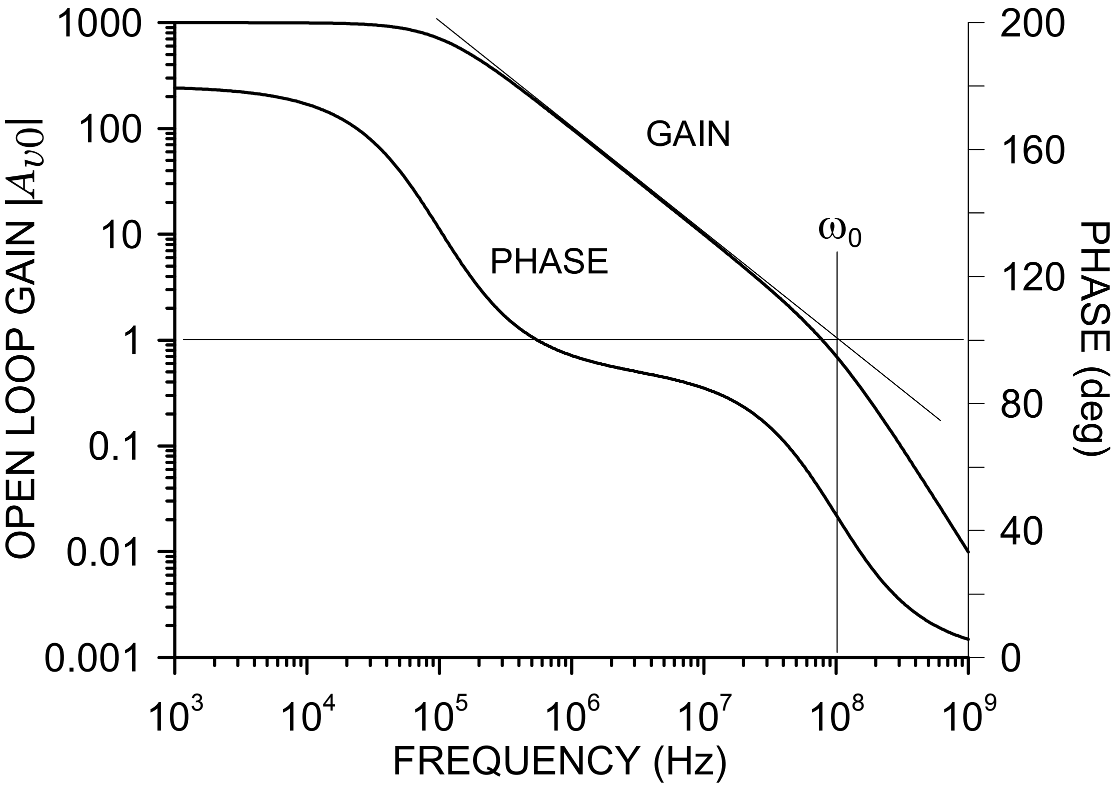


The relevant frequency range is determined by the frequency passband of the pulse shaper. This is 5 – 15 MHz for a typical 20 ns shaper, so in this example the ohmic input is effective at much longer shaping times.

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.



Low-power amplifiers with a gain-bandwidth product much greater than in this example are quite practical, so smaller feedback capacitances are also possible.

Time Response of a Charge-Sensitive Amplifier

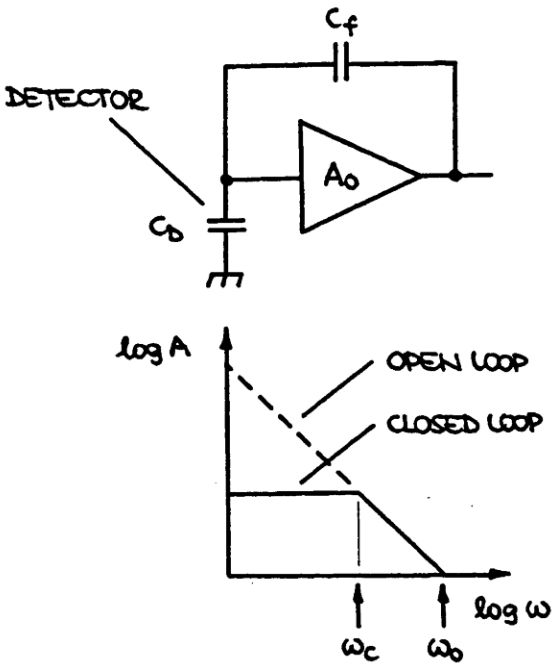
Input resistance and detector capacitance form RC time constant:

$$\tau_i = R_i C_D$$

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

⇒ Rise time increases with detector capacitance.

Or apply feedback theory:



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

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

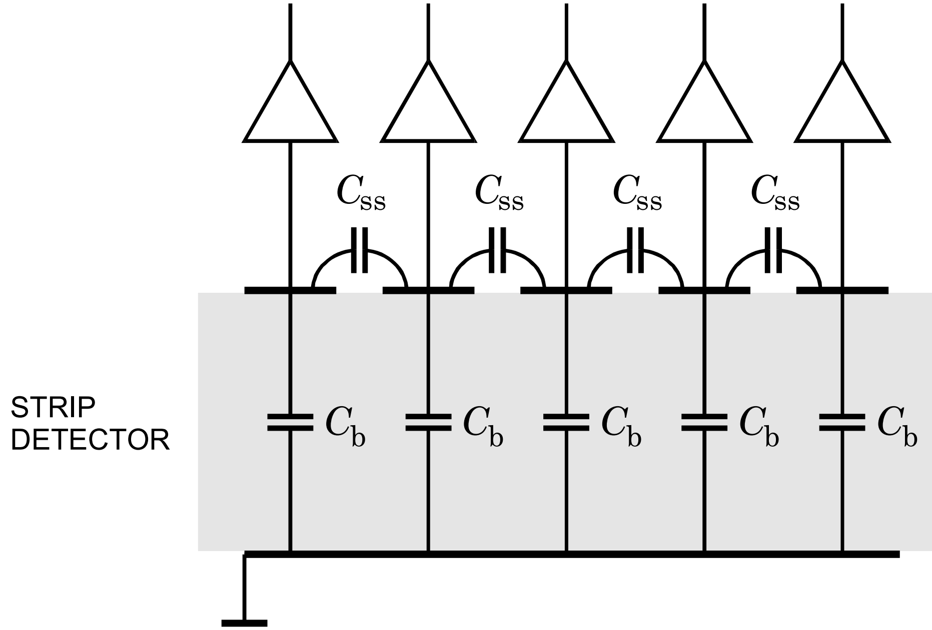
Closed Loop Bandwidth $\omega_C A_f = \omega_0$

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

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
10 MHz ($\triangleq$ ~20 ns peaking time)
$Z_i \approx 1.6$ kΩ, corresponding to 12 pF

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

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 $T_P > (2 \ldots 3) \times R_i C_D$ and if $C_i \gg C_D$.

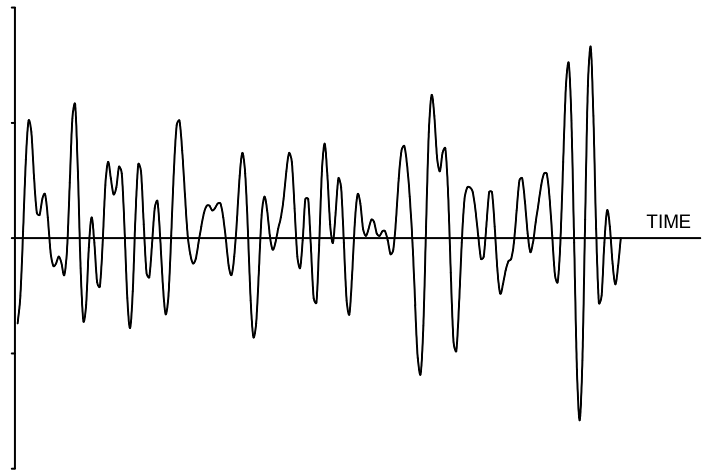
III. Electronic Noise

1. Signals and Noise
2. Basic Noise Mechanisms
    - Thermal Noise
    - Shot Noise
    - Low Frequency ("$1/f$") Noise
3. Signal-to-Noise Ratio vs. Detector Capacitance
    - Voltage Amplifier
    - Charge-Sensitive Amplifier
4. Complex Sensors
    - Cross-coupled noise
    - Strip Detector Model
5. Quantum Noise Limits in Amplifiers

1. Signals and Noise

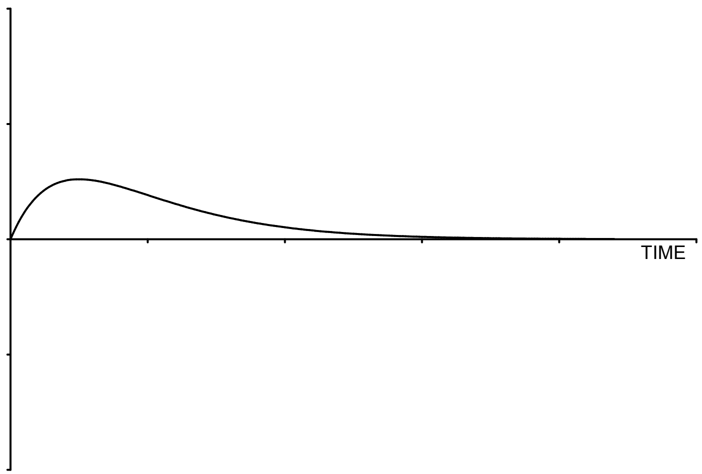
Choose a time when no signal is present.

Amplifier's quiescent output level (baseline):

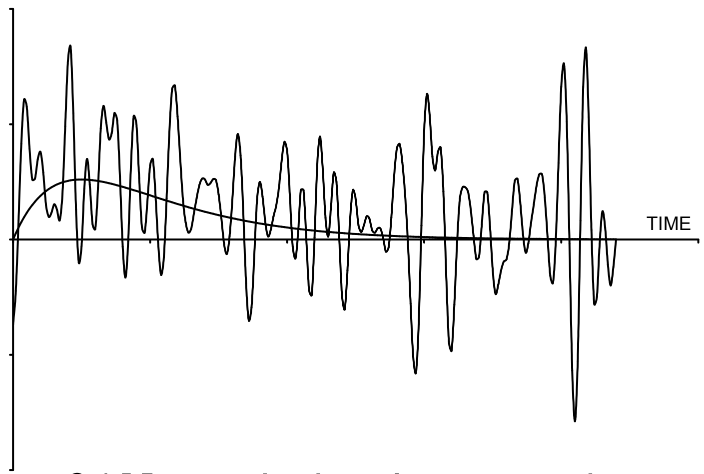


In the presence of a signal, noise + signal add.

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.

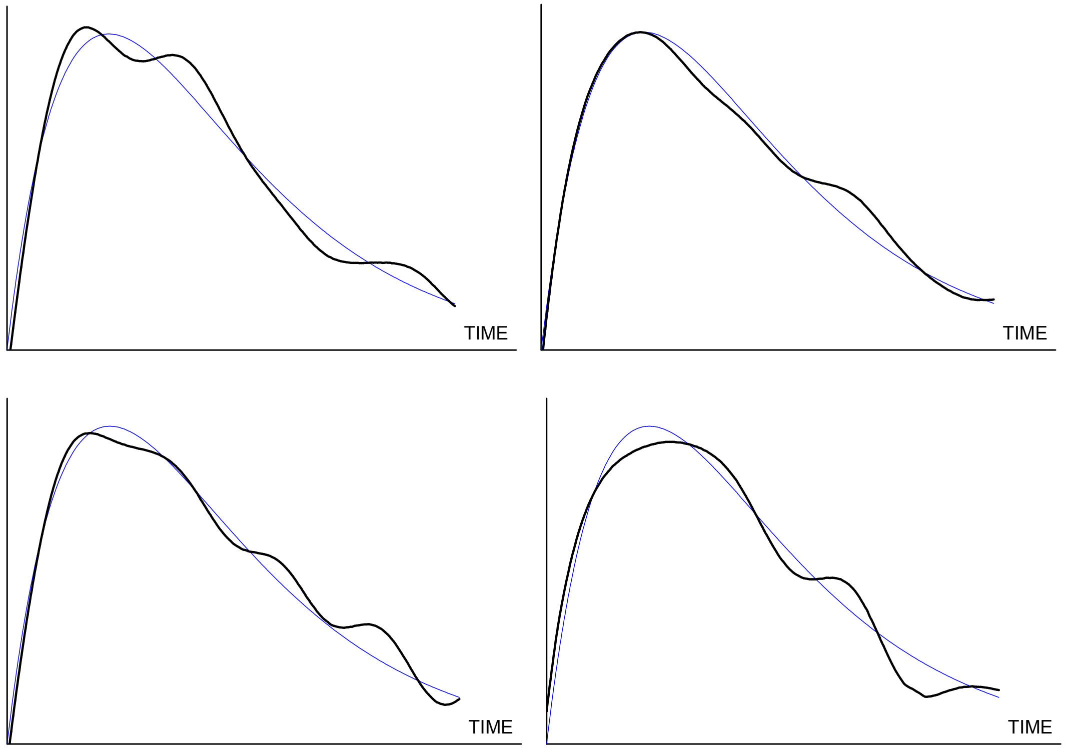
In an optimized system, the time scale of the fluctuation is comparable to the signal peaking time.

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

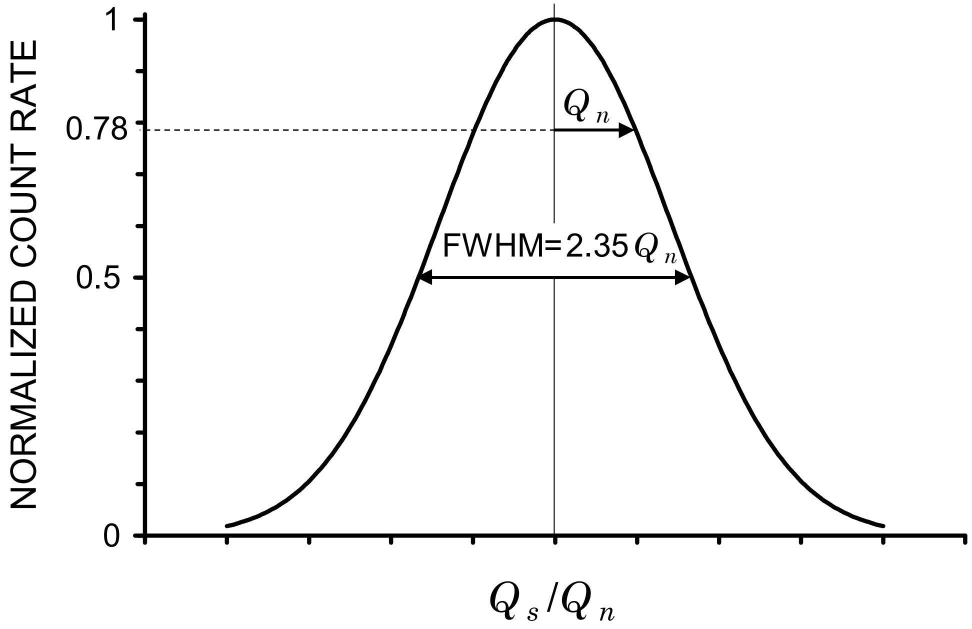
Electronic noise is purely random.

⇒ 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)

2. 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: $\frac{dP_{noise}}{df} = const.$

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 $\frac{dV_{noise}^2}{df} \equiv e_n^2 = 4kTR$

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

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).

Spectral Density of Thermal Noise (Johnson Noise)

Two approaches can be used to derive the spectral distribution of thermal noise.

1. The thermal velocity distribution of the charge carriers is used to calculate the time dependence of the induced current, which is then transformed into the frequency domain.
2. Application of Planck's theory of black body radiation.

The first approach clearly shows the underlying physics, whereas the second "hides" the physics by applying a general result of statistical mechanics. However, the first requires some advanced concepts that go well beyond the standard curriculum, so the "black body" approach will be used.

In Planck's theory of black body radiation the energy per mode

$$\overline{E} = \frac{h\nu}{e^{h\nu / kT} - 1}$$

and the spectral density of the radiated power

$$\frac{dP}{d\nu} = \frac{h\nu}{e^{h\nu / kT} - 1}.$$

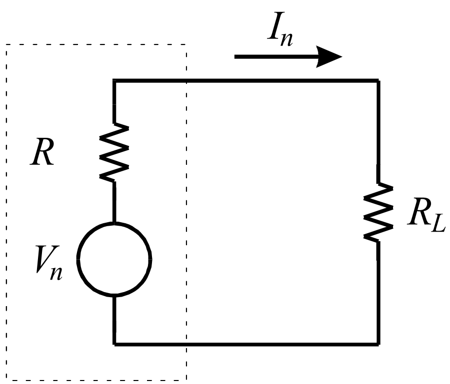
This is the power that can be extracted in equilibrium.

At low frequencies $h\nu \ll kT$:
$$\frac{dP}{d\nu} \approx \frac{h\nu}{\left(1+\frac{h\nu}{kT}\right)-1} = kT \ ,$$

so at low frequencies the spectral density is independent of frequency and for a total bandwidth B the noise power that can be transferred to an external device

$$P_n = kTB \, .$$

To apply this result to the noise of a resistor, consider a resistor R whose thermal noise gives rise to a noise voltage V_n. To determine the power transferred to an external device consider the circuit



The dotted box encloses the equivalent circuit of the resistive noise source.

The power dissipated in the load resistor R_L

$$\frac{V_{nL}^2}{R_L} = I_n^2 R_L = \frac{V_n^2 R_L}{(R+R_L)^2}$$

The maximum power transfer occurs when the load resistance equals the source resistance $R_L = R$, so

$$V_{nL}^2 = \frac{V_n^2}{4} \; .$$

Since the maximum power that can be transferred to R_L is kTB ,

$$\frac{V_{nL}^2}{R} = \frac{V_n^2}{4R} = kTB$$

$$P_n = \frac{V_n^2}{R} = 4kTB$$

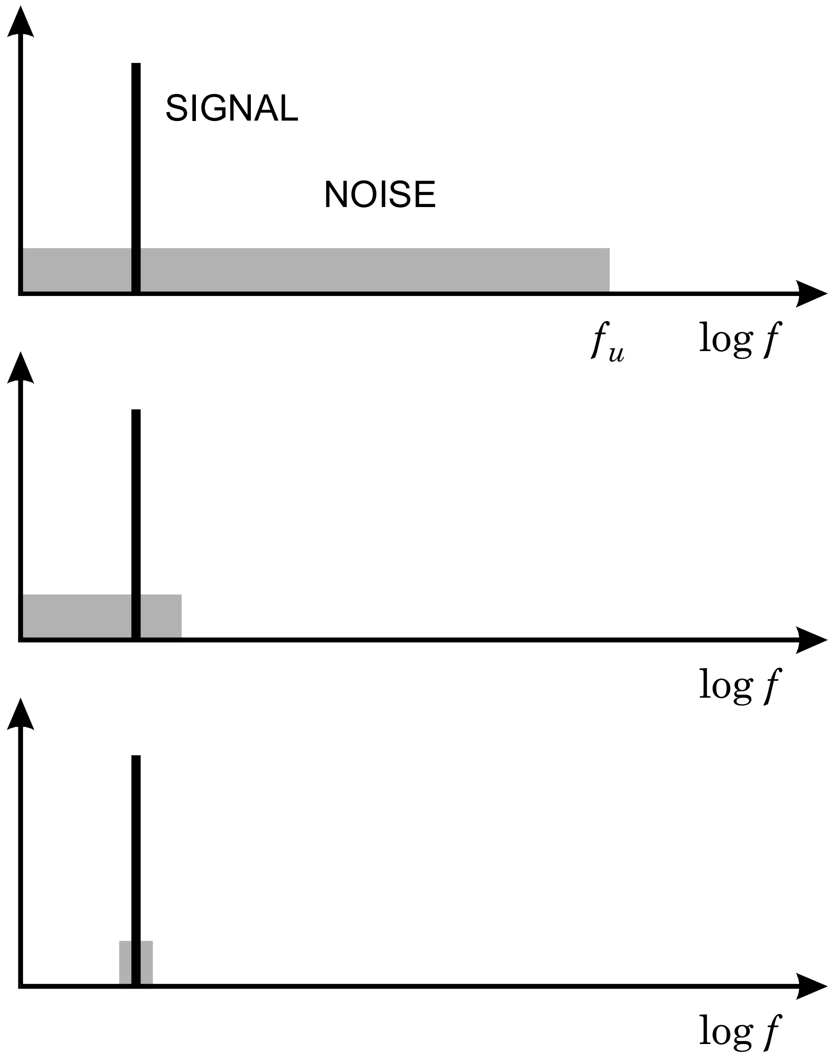
and the spectral density of the noise power in the resistor

$$\frac{dP_n}{d\nu} = 4kT \; .$$

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.



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

A more intuitive interpretation of this expression will be given in Part IV.

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.

Low Frequency (“$1/f$”) Noise

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

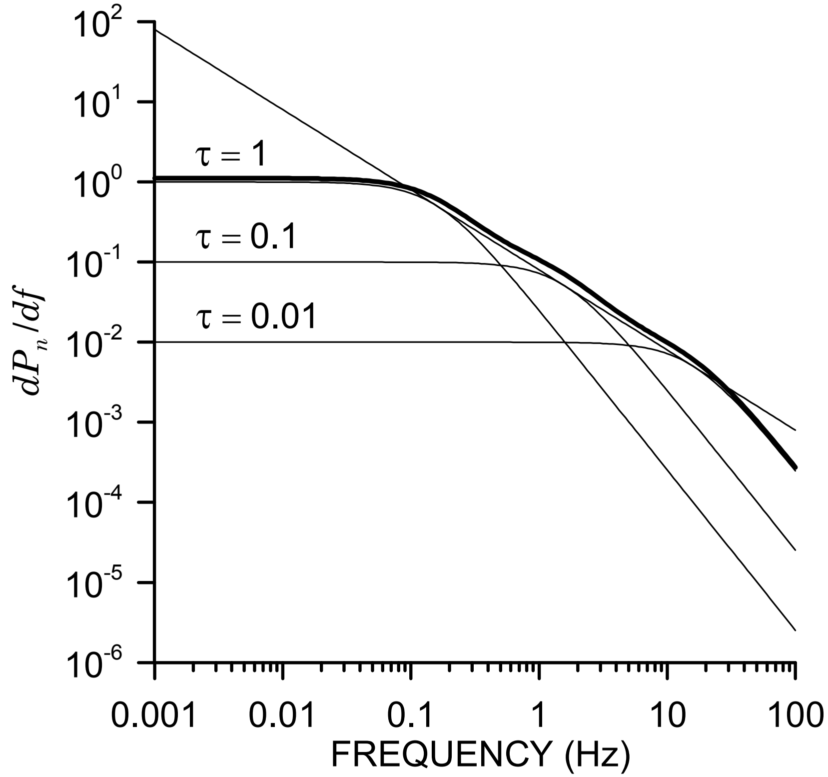
The spectral density for a single lifetime

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

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

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

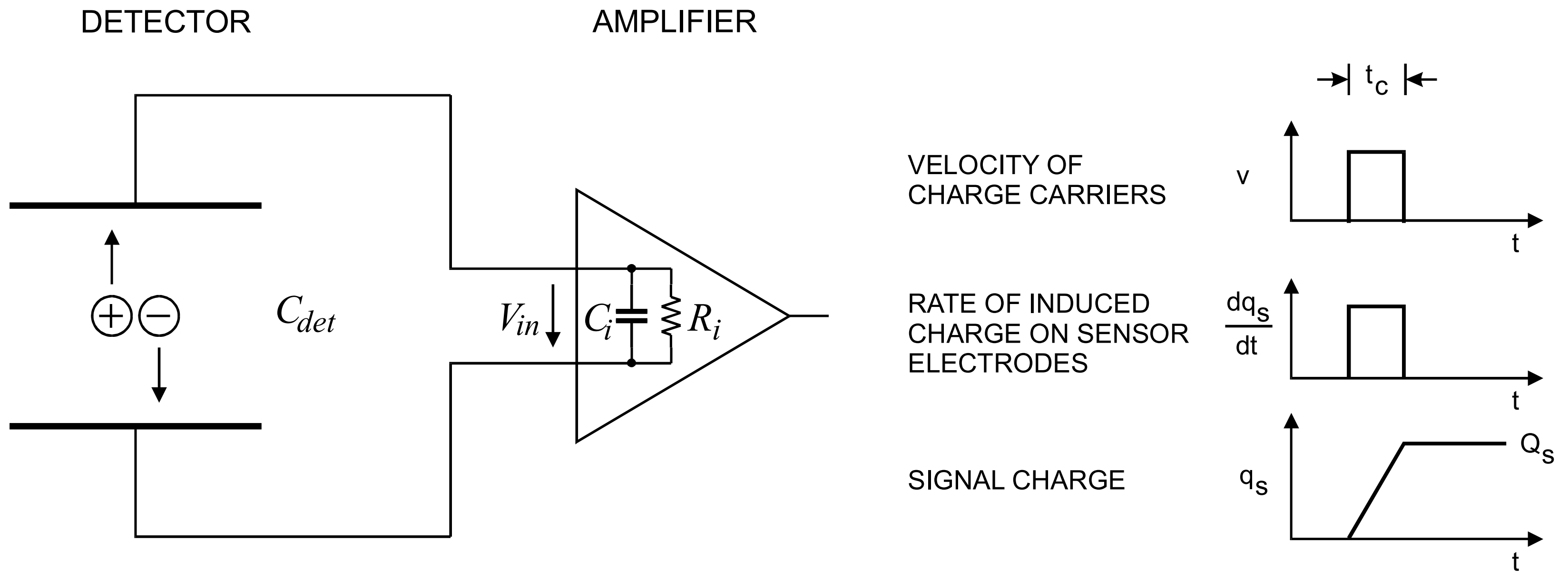
Traps with three time constants of 0.01, 0.1 and 1 s yield a $1/f$ distribution over two decades in frequency.



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

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

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



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

the 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.

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.

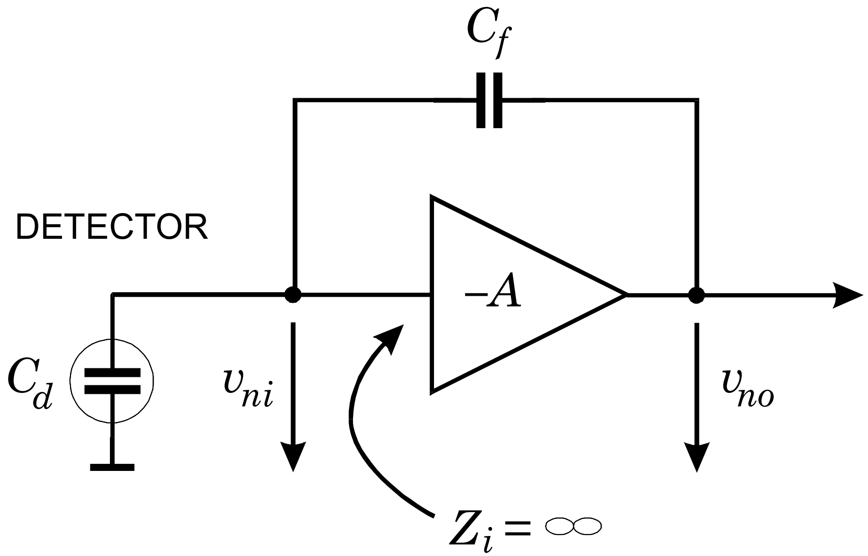
$\Rightarrow$ ***S/N* at the amplifier output depends on feedback.**

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.

In an ideal current-sensitive amplifier the signal-to-noise ratio can be independent of capacitance, but the effect of external noise sources often increases with detector capacitance.

Conclusion

In general

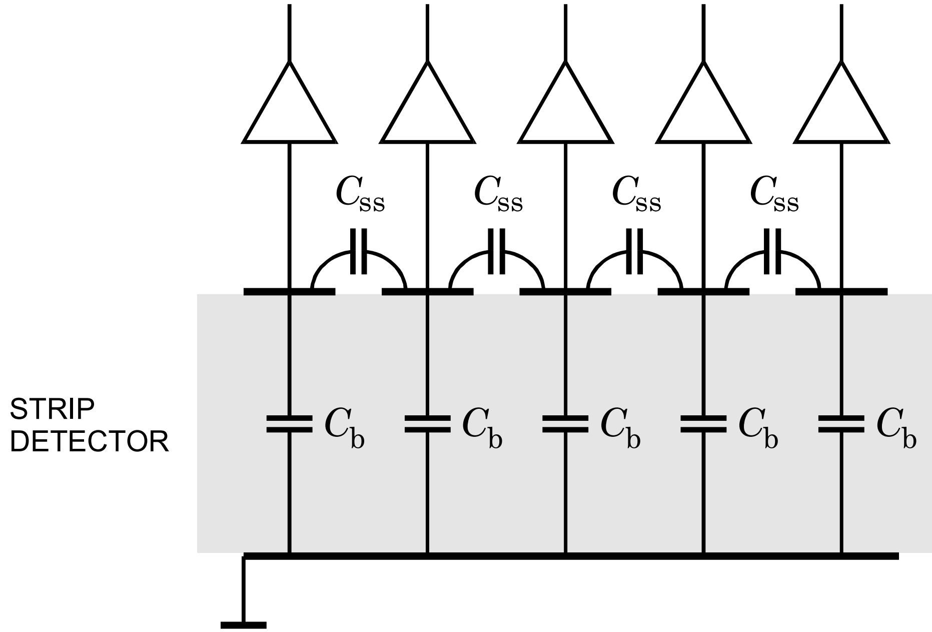
- optimum S/N is independent of whether the voltage, current, or charge signal is sensed.

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

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

4. Complex Sensors

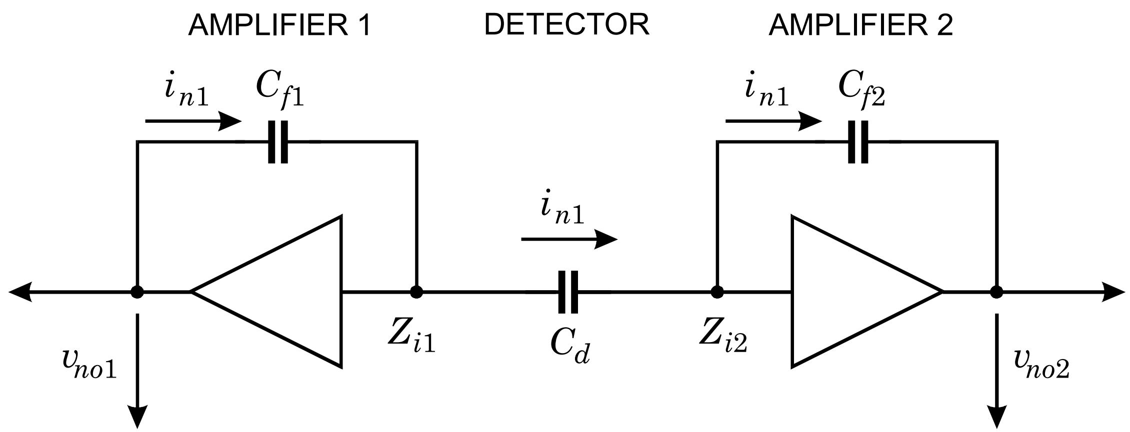
Cross-coupled noise



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

Principle of Noise Cross-Coupling

Consider a capacitance connecting two amplifier inputs, e.g. amplifiers at opposite electrodes of a detector with capacitance C_d:



First, assume that amplifier 2 is noiseless.

The noise voltage v_{no1} causes a current flow i_{n1} to flow through the feedback capacitance C_{f1} and the detector capacitance C_d into the input of amplifier 2.

Note that for a signal originating at the output of amplifier 1, its input impedance Z_{i1} is high (∞ for an idealized amplifier), so all of current i_{n1} flows into amplifier 2.

Amplifier 2 presents a low impedance to the noise current i_{n1}, so its magnitude

$$i_{n1} = \frac{v_{no1}}{X_{C_{f1}} + X_{C_d}} = \frac{v_{no1}}{\dfrac{1}{\omega C_{f1}} + \dfrac{1}{\omega C_d}}.$$

The voltage at the output of amplifier is the product of the input current times the feedback impedance,

$$v_{no12} = \frac{i_{n1}}{\omega C_{f2}} = \frac{v_{no1}}{\dfrac{1}{\omega C_{f1}} + \dfrac{1}{\omega C_d}} \cdot \frac{1}{\omega C_{f2}} = \frac{v_{no1}}{\dfrac{C_{f2}}{C_{f1}} + \dfrac{C_{f2}}{C_d}}.$$

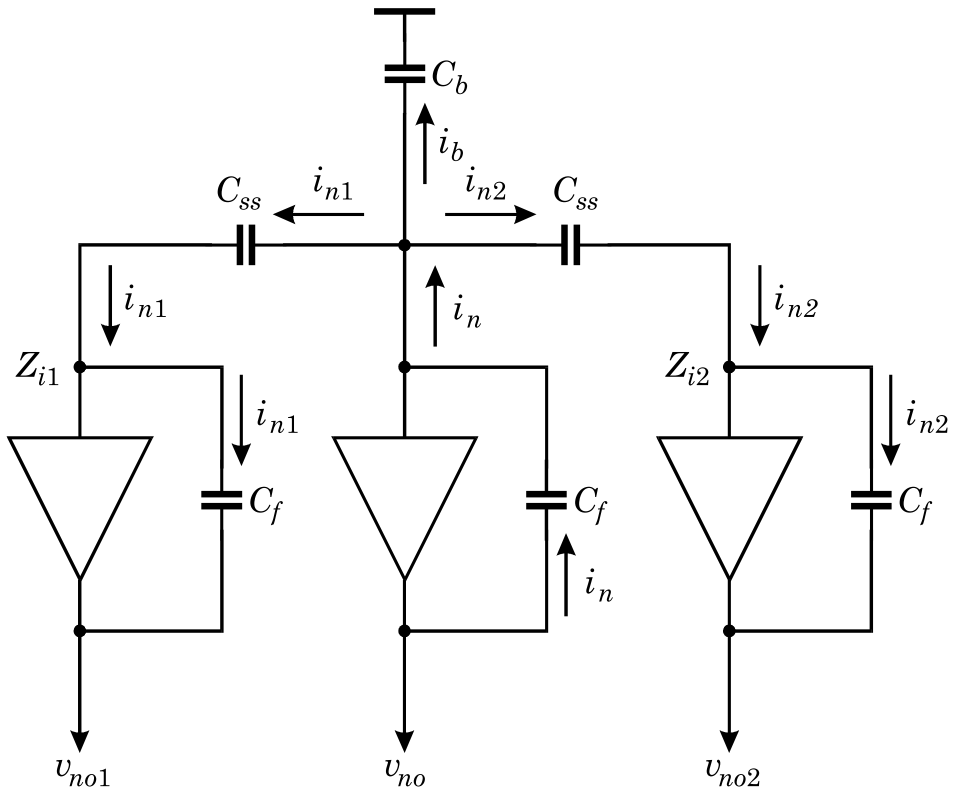
For identical amplifiers $C_{f1} = C_{f2}$. Furthermore, $C_{f2} \ll C_d$, so the additional noise from amplifier 1 at the output of amplifier 2 is

$$v_{no12} = v_{no1}.$$

This adds in quadrature to the noise of amplifier 2. Since both amplifiers are same, $v_{no1} = v_{no2}$, so cross-coupling increases the noise by a factor $\sqrt{2}$.

Cross-Coupling in Strip Detectors

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 $C_b = 0$, $v_{no1} = v_{no2} = v_{no} / 2$ and the total noise increases by a factor

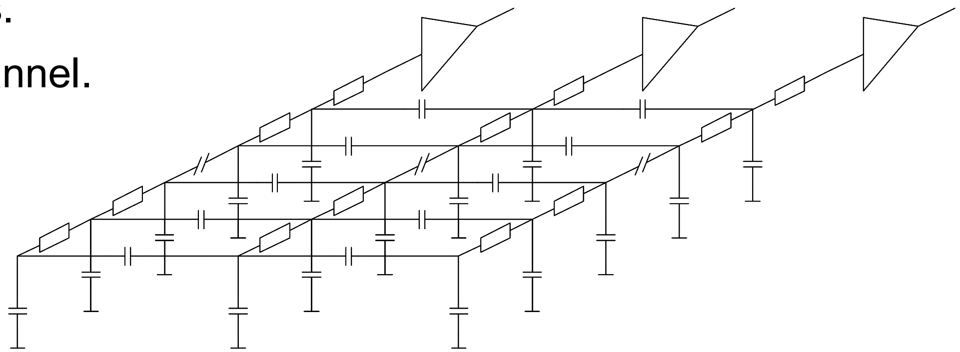
$$\sqrt{1 + 0.5^2 + 0.5^2} = 1.22$$

For a backplane capacitance $C_b = C_{ss} / 10$ the noise increases by 16%.

Strip Detector Model for Noise Simulations

Noise coupled from neighbor channels.

Analyze signal and noise in center channel.



Includes:
a) Noise contributions from
   neighbor channels
b) Signal transfer to
   neighbor channels
c) Noise from distributed strip resistance
d) Full SPICE model of preamplifier

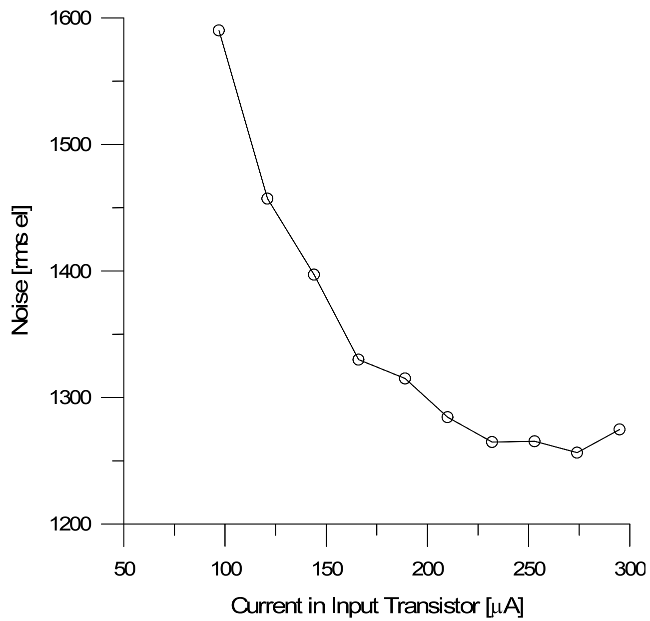
Measured Noise of Module:

p-strips on n-bulk, BJT input transistor

Simulation Results: 1460 el (150 μA)

1230 el (300 μA)

⇒ Noise can be predicted with good accuracy.



5. Quantum Noise Limits in Amplifiers

What is the lower limit to electronic noise?

Can it be eliminated altogether, for example by using superconductors and eliminating devices that carry shot noise?

Starting point is the uncertainty relationship

$$\Delta E \Delta t \geq \frac{\hbar}{2}$$

Consider a narrow frequency band at frequency ω. The energy uncertainty can be given in terms of the uncertainty in the number of signal quanta

$$\Delta E = \hbar \omega \Delta n$$

and the time uncertainty in terms of phase

$$\Delta t = \frac{\Delta \varphi}{\omega},$$

so that

$$\Delta \varphi \Delta n \geq \frac{1}{2}$$

We assume that the distributions in number and phase are Gaussian, so that the equality holds.

Assume a noiseless amplifier with gain G, so that n_1 quanta at the input yield

$$n_2 = Gn_1$$

quanta at the output.

Furthermore, the phase at the output φ_2 is shifted by a constant relative to the input.

Then the output must also obey the relationship $\Delta\varphi_2 \Delta n_2 = \frac{1}{2}$

However, since $\Delta n_2 = G\Delta n_1$ and $\Delta\varphi_2 = \Delta\varphi_1$:

$$\Delta\varphi_1 \Delta n_1 = \frac{1}{2G} \;,$$

which is smaller than allowed by the uncertainty principle.

This contradiction can only be avoided by assuming that the amplifier introduces noise per unit bandwidth of

$$\frac{dP_{no}}{d\omega} = (G-1)\hbar\omega \ ,$$

which, referred to the input, is

$$\frac{dP_{ni}}{d\omega} = \left(1 - \frac{1}{G}\right)\hbar\omega$$

If the noise from the following gain stages is to be small, the gain of the first stage must be large, and then the minimum noise of the amplifier

$$\frac{dP_{ni}}{d\omega} = \hbar\omega$$

At 2 mm wavelength the minimum noise corresponds to about 7K.

This minimum noise limit applies to phase-coherent systems. In systems where the phase information is lost, e.g. bolometers, this limit does not apply.

For a detailed discussion see C.M. Caves, Phys. Rev. D **26** (1982) 1817-1839
H.A. Haus and J.A. Mullen, Phys. Rev. 128 (1962) 2407-2413

IV. Signal Processing

1. The Problem

   "Optimum" Filtering

   Pulse Shaping Objectives

2. Pulse Shaping and Signal-to-Noise Ratio

   Equivalent Noise Charge

   Ballistic Deficit

   Noise vs. Shaping Time

   Analytical Analysis of a Detector Front-End

   Other Types of Shapers

   Examples

   Detector Noise Summary

3. Noise vs. Power Dissipation

1. The Problem

Radiation impinges on a sensor and creates an electrical signal.

The signal level is low and must be amplified to allow digitization and storage.

Both the sensor and amplifiers introduce signal fluctuations – noise.

1. Fluctuations in signal introduced by sensor
2. Noise from electronics superimposed on signal

The detection limit and measurement accuracy are determined by the signal-to-noise ratio.

Electronic noise affects all measurements:

1. Detect presence of hit: Noise level determines minimum threshold.
   If threshold too low, output dominated by noise hits.
2. Energy measurement: Noise "smears" signal amplitude.
3. Time measurement: Noise alters time dependence of signal pulse.

How to optimize the signal-to-noise ratio?

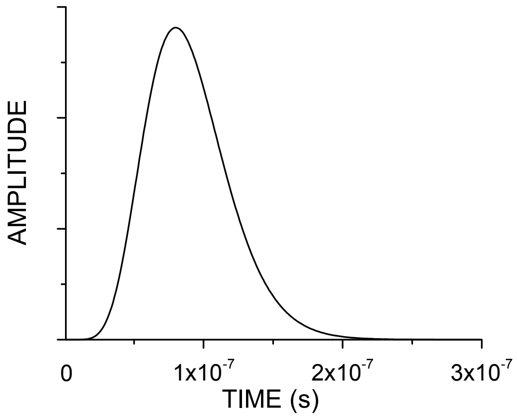
1. Increase signal and reduce noise
2. For a given sensor and signal: reduce electronic noise

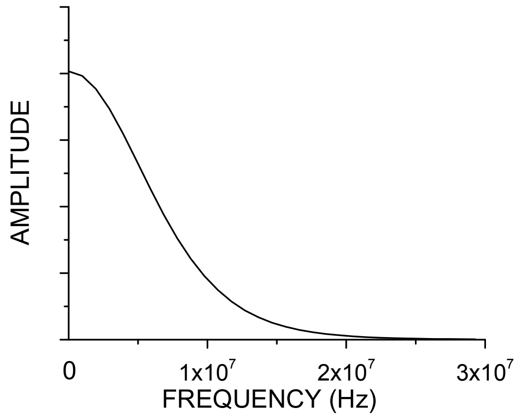
Assume that the signal is a pulse.

The time distribution of the signal corresponds to a frequency spectrum (Fourier transform).

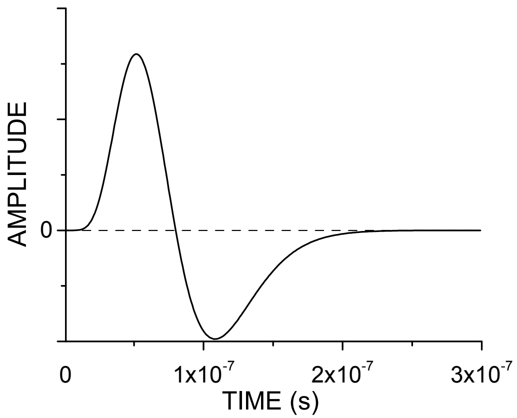
Examples: Time Domain Frequency Domain

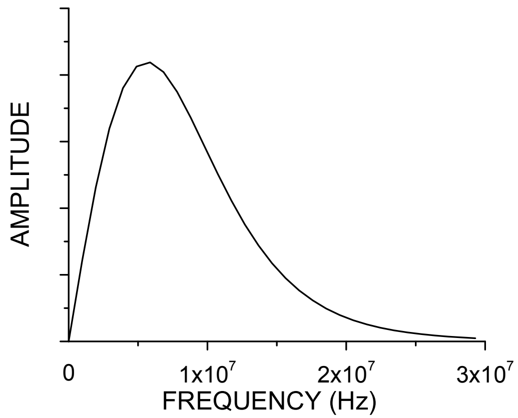
1. The pulse is unipolar, so it has a DC component and the frequency spectrum extends down to zero.





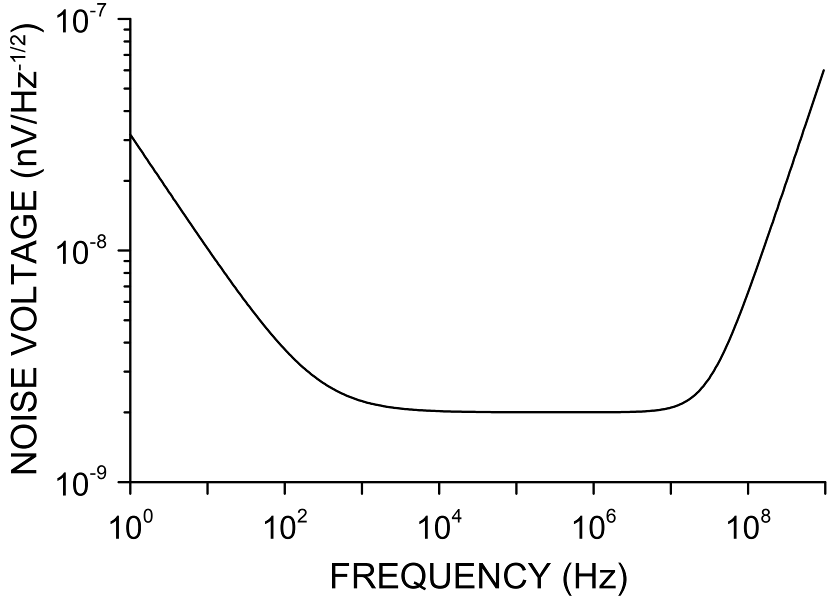
2. This bipolar pulse carries no net charge, so the frequency spectrum falls to zero at low frequencies, but extends to higher frequencies because of the faster slope.





The noise spectrum is generally not the same as the signal spectrum.

Typical Noise Spectrum:



⇒ Tailor frequency response of measurement system to optimize signal-to-noise ratio.

Frequency response of the measurement system affects both

- signal amplitude and
- noise.

Signal Processing Objectives

Two conflicting objectives:

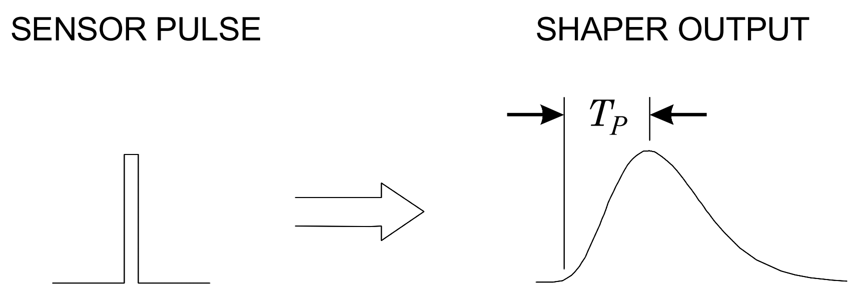
1. Improve Signal-to-Noise Ratio S/N

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),

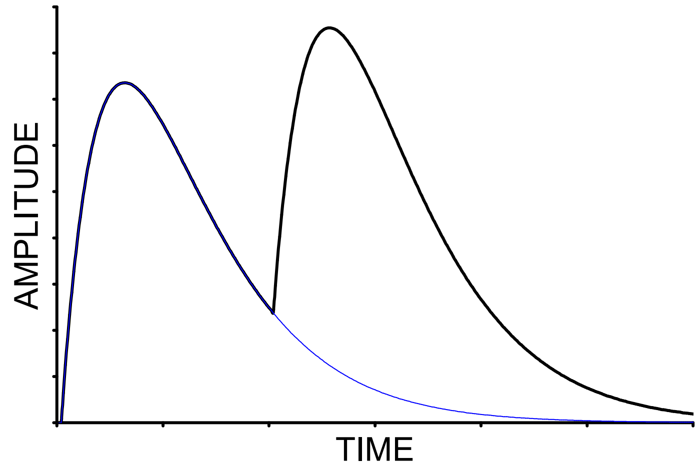
with a gradually rounded maximum at the peaking time T_P
(to facilitate measurement of the peak amplitude)



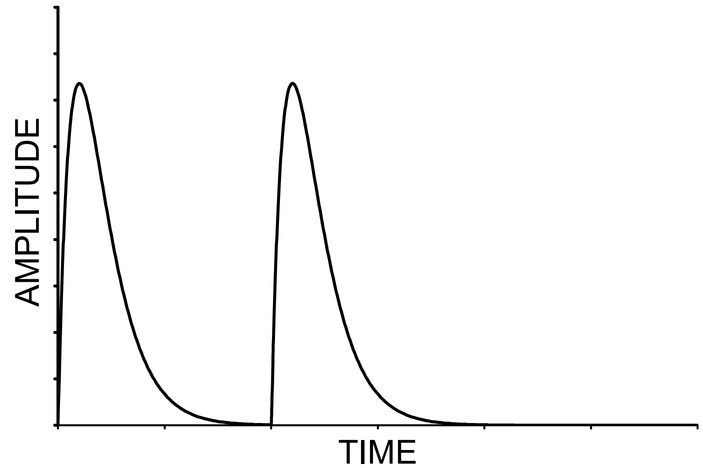
If the shape of the pulse does not change with signal level, the peak amplitude is also a measure of the energy, so one often speaks of pulse-height measurements or pulse height analysis. The pulse height spectrum is the energy spectrum.

2. Improve Pulse Pair Resolution ⇒ Decrease pulse width

Pulse pile-up distorts amplitude measurement.



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



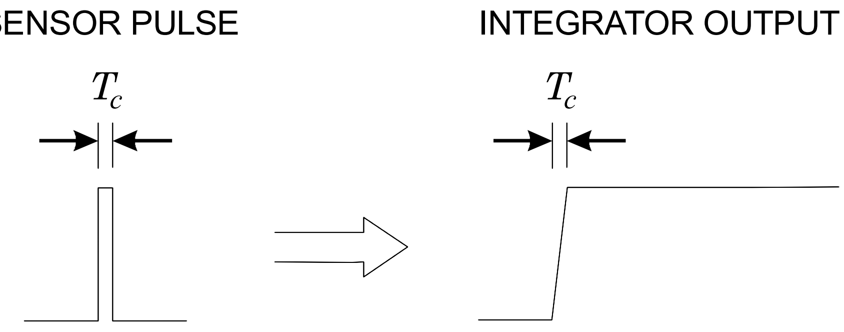
Necessary to find balance between these conflicting requirements. Sometimes minimum noise is crucial, sometimes rate capability is paramount.

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).

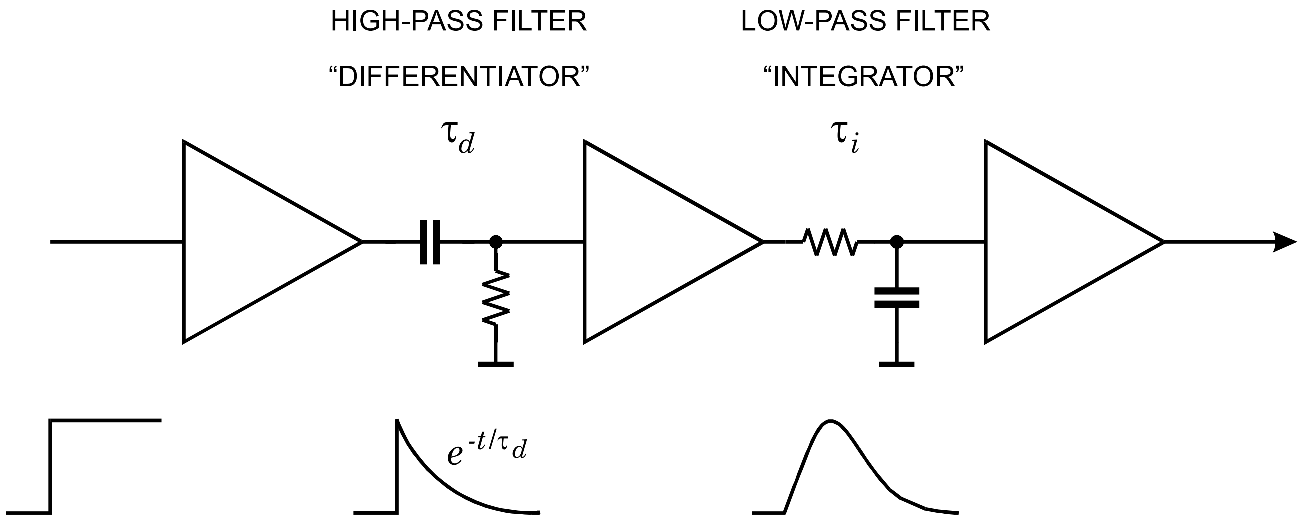
Long "flat top" allows measurements at times well beyond the collection time T_C.

$\Rightarrow$ 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).

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

are common to all shapers.

2. 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 $S/N = 1$

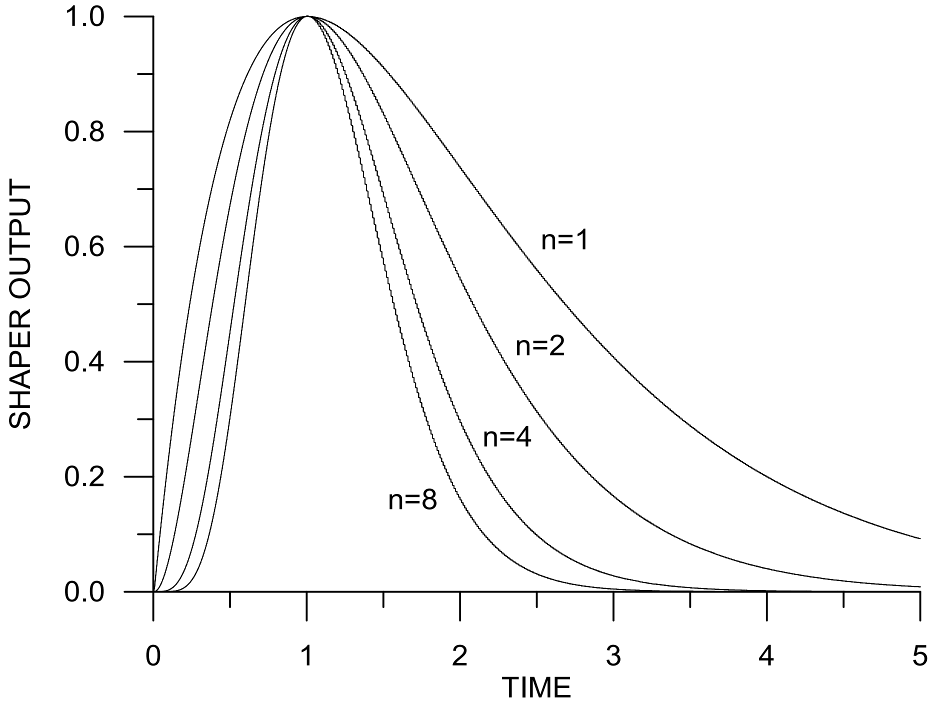
Shapers with Multiple Integrators

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.

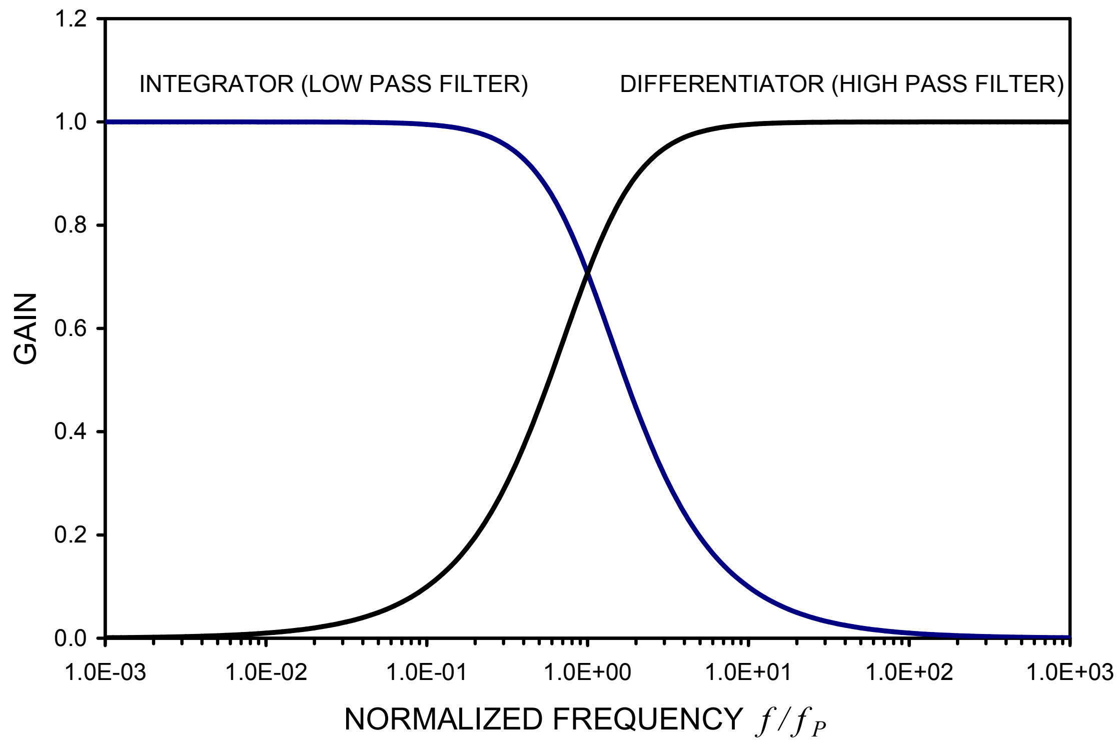
In γ-spectroscopy systems shapers with the equivalent of 8 RC integrators are common. Usually, this is achieved with active filters.

Noise Charge vs. Shaping Time

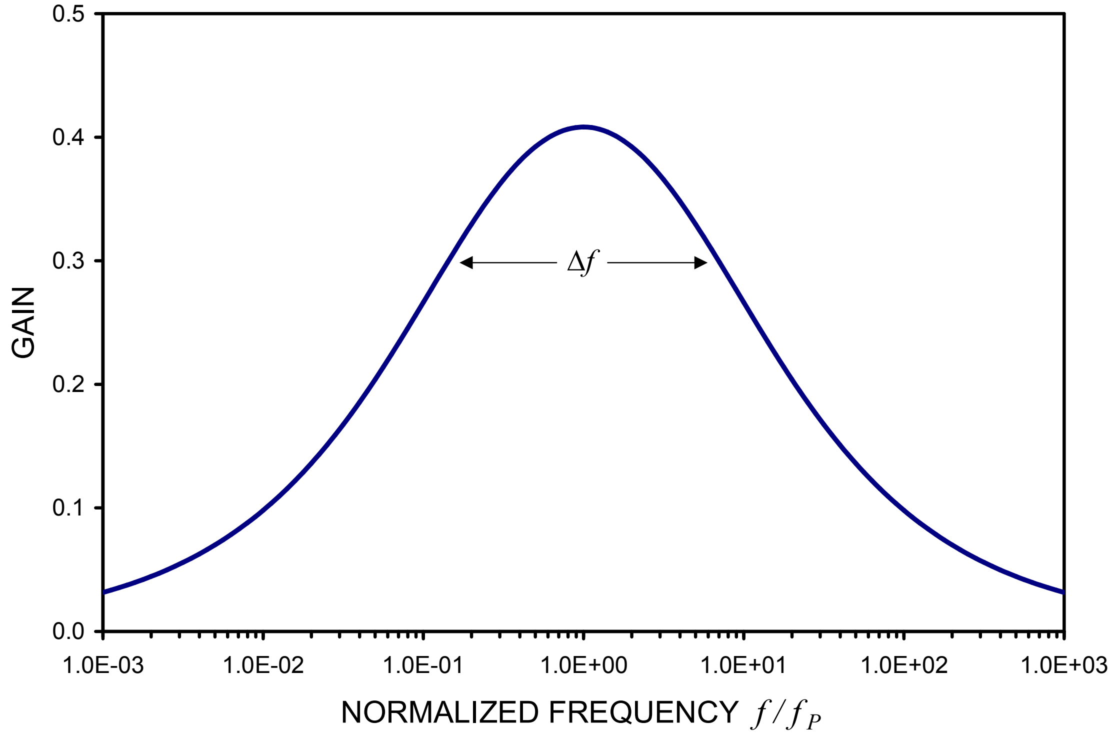
Assume that differentiator and integrator time constants are equal $\tau_i = \tau_d \equiv \tau$.

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

Frequency response of individual pulse shaping stages



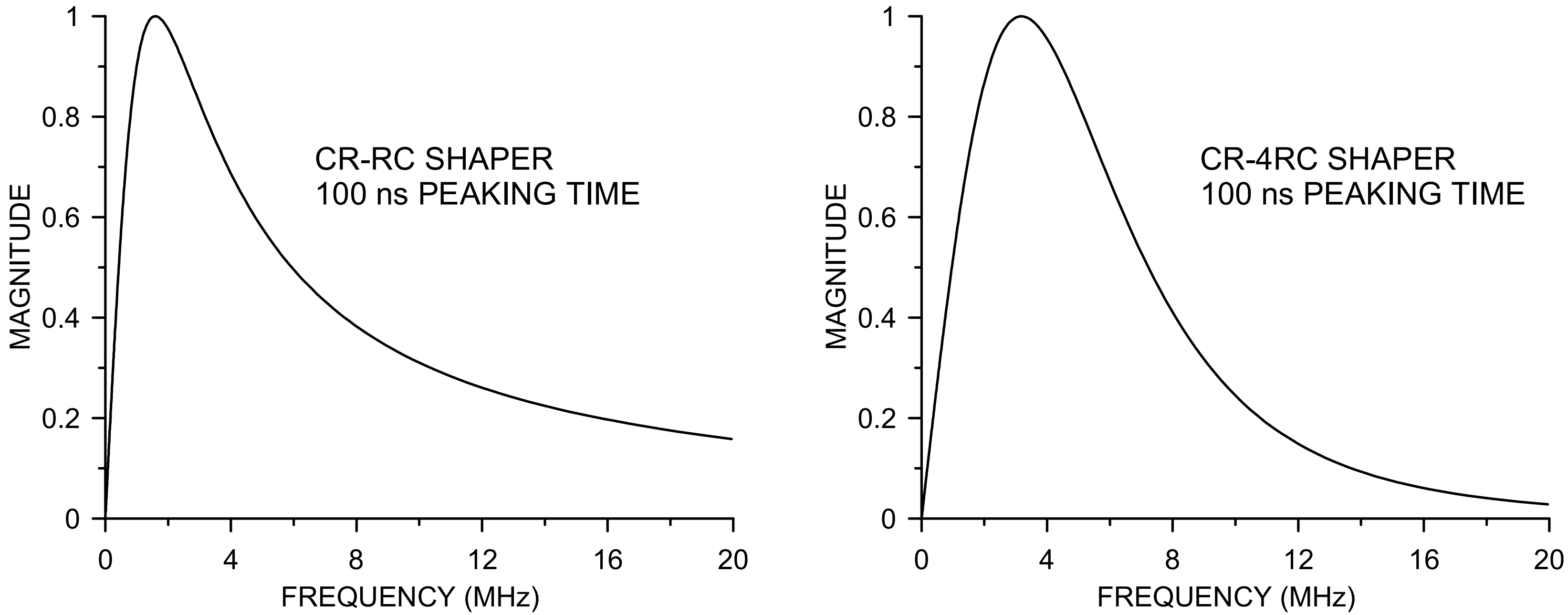
Combined frequency response



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

Bandwidth Δf decreases with increasing time constant τ.

Comparison with CR-$4RC$ shaper



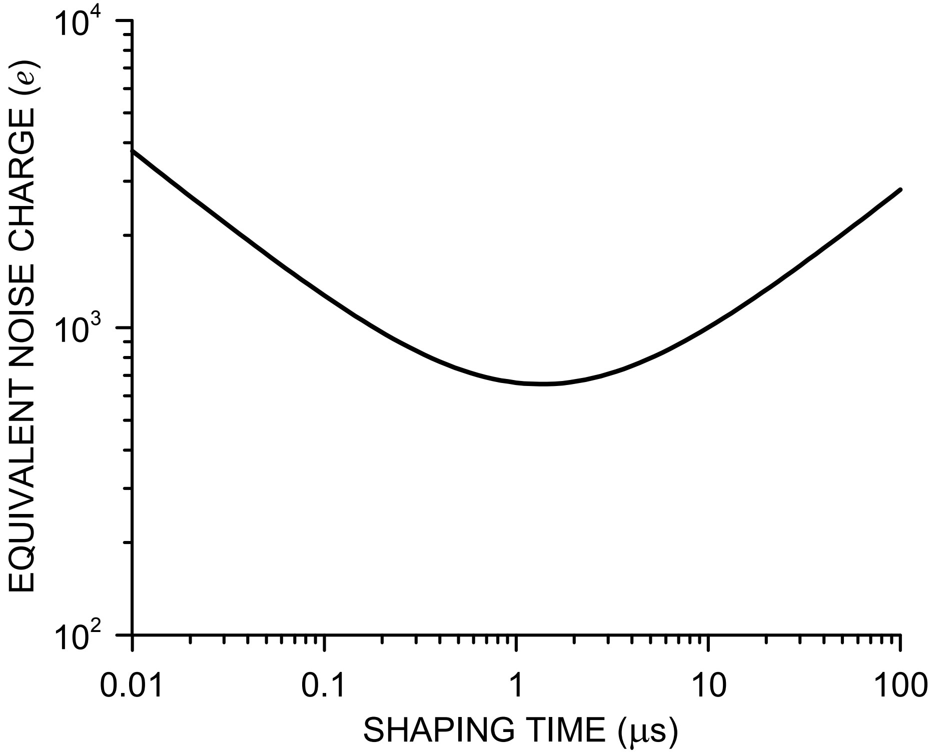
Both have a 100 ns peaking time.

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.

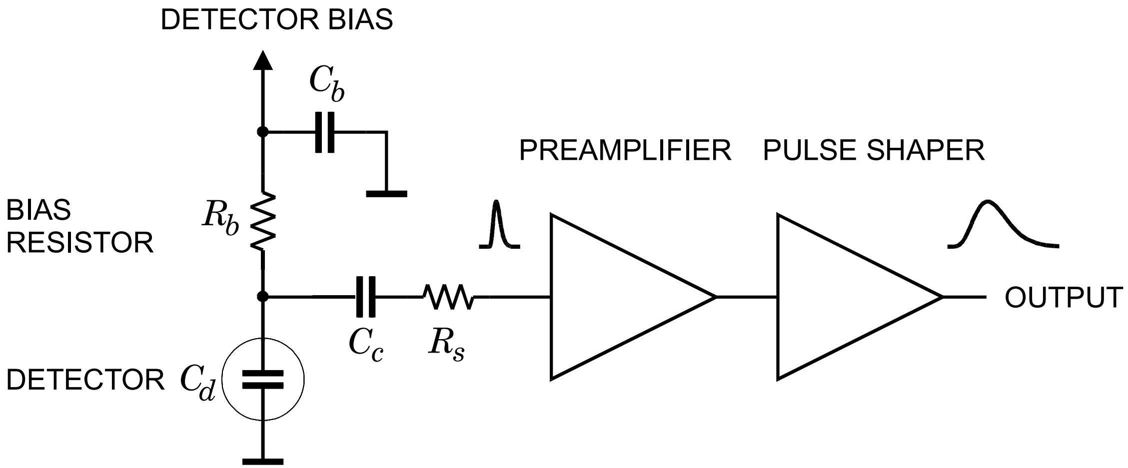
Result of typical noise measurement vs. shaping time



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



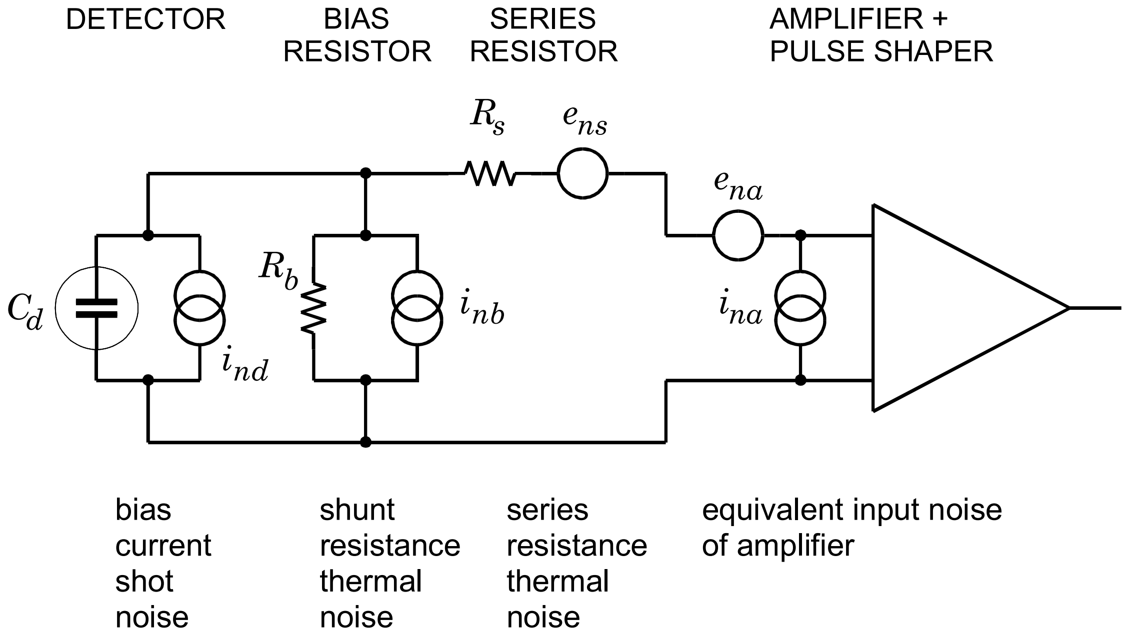
Detector bias voltage is applied through the resistor R_B. The bypass capacitor C_B serves to shunt any external interference coming through the bias supply line to ground. For AC signals this capacitor connects the “far end” of the bias resistor to ground, so that R_B appears to be in parallel with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

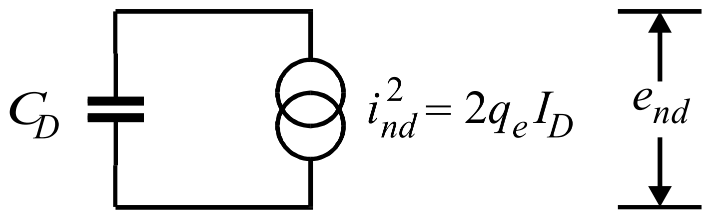
1. Determine the frequency distribution of the noise voltage presented to the amplifier input from all individual noise sources
2. Integrate over the frequency response of a CR-RC shaper to determine the total noise output.
3. Determine the output signal for a known signal charge and calculate equivalent 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.

Changes in Noise Spectrum – Detector Bias Current



This model results from two assumptions:

1. The input impedance of the amplifier is infinite
2. The shunt resistance R_P is much larger than the capacitive reactance of the detector in the frequency range of the pulse shaper.

Does this assumption make sense?

If R_P is too small, the signal charge on the detector capacitance will discharge before the shaper output peaks. To avoid this

$$R_P C_D \gg T_P \approx \frac{1}{\omega_P},$$

where ω_P is the midband frequency of the shaper.

Therefore, $R_P \gg \dfrac{1}{\omega_P C_D}$ as postulated.

Under these conditions the noise current will flow through the detector capacitance, yielding the voltage

$$e_{nd}^2 = i_{nd}^2 \frac{1}{(\omega\, C_D)^2} = 2q_e I_D \frac{1}{(\omega\, C_D)^2}$$

⇒ The noise contribution decreases with increasing frequency (shorter shaping time)

Note: Although shot noise is “white”, the resulting noise spectrum is strongly frequency dependent.

Shot Noise Viewed in the Time Domain

In the time domain this result is more intuitive. Since every shaper also acts as an integrator, one can view the total shot noise as the result of “counting electrons”.

Assume an ideal integrator that records all charge uniformly within a time T. The number of electron charges measured is

$$N_e = \frac{I_D T}{q_e}$$

The associated noise is the fluctuation in the number of electron charges recorded

$$\sigma_n = \sqrt{N_e} \propto \sqrt{T}$$

Does this also apply to an AC-coupled system, where no DC current flows, so no electrons are “counted”?

Since shot noise is a fluctuation, the current undergoes both positive and negative excursions. Although the DC component is not passed through an AC coupled system, the excursions are. Since, on the average, each fluctuation requires a positive and a negative zero crossing, the process of “counting electrons” is actually the counting of zero crossings, which in a detailed analysis yields the same result.

Equivalent Noise Charge

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

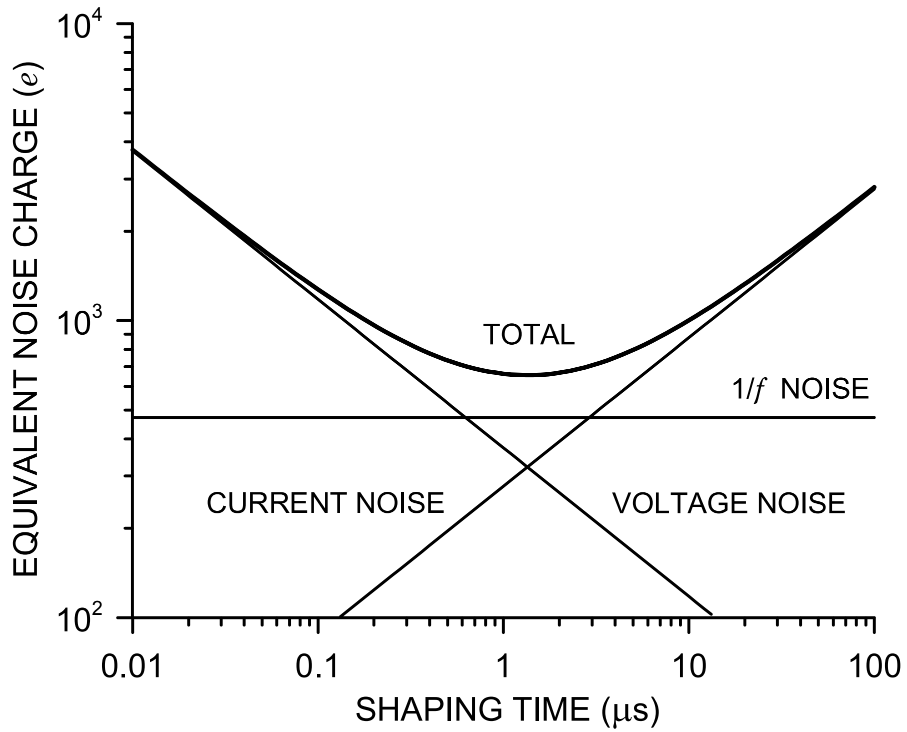
	↑	↑	↑
$\mathbf{e} = \exp(1)$	current noise	voltage noise	$1/f$ noise
	$\propto \tau$	$\propto 1/\tau$	independent of τ
	independent of C_D	$\propto C_D{}^2$	$\propto C_D{}^2$

- Current noise is independent of detector capacitance, consistent with the notion of "counting electrons".
- Voltage noise increases with detector capacitance (reduced signal voltage)
- $1/f$ 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.

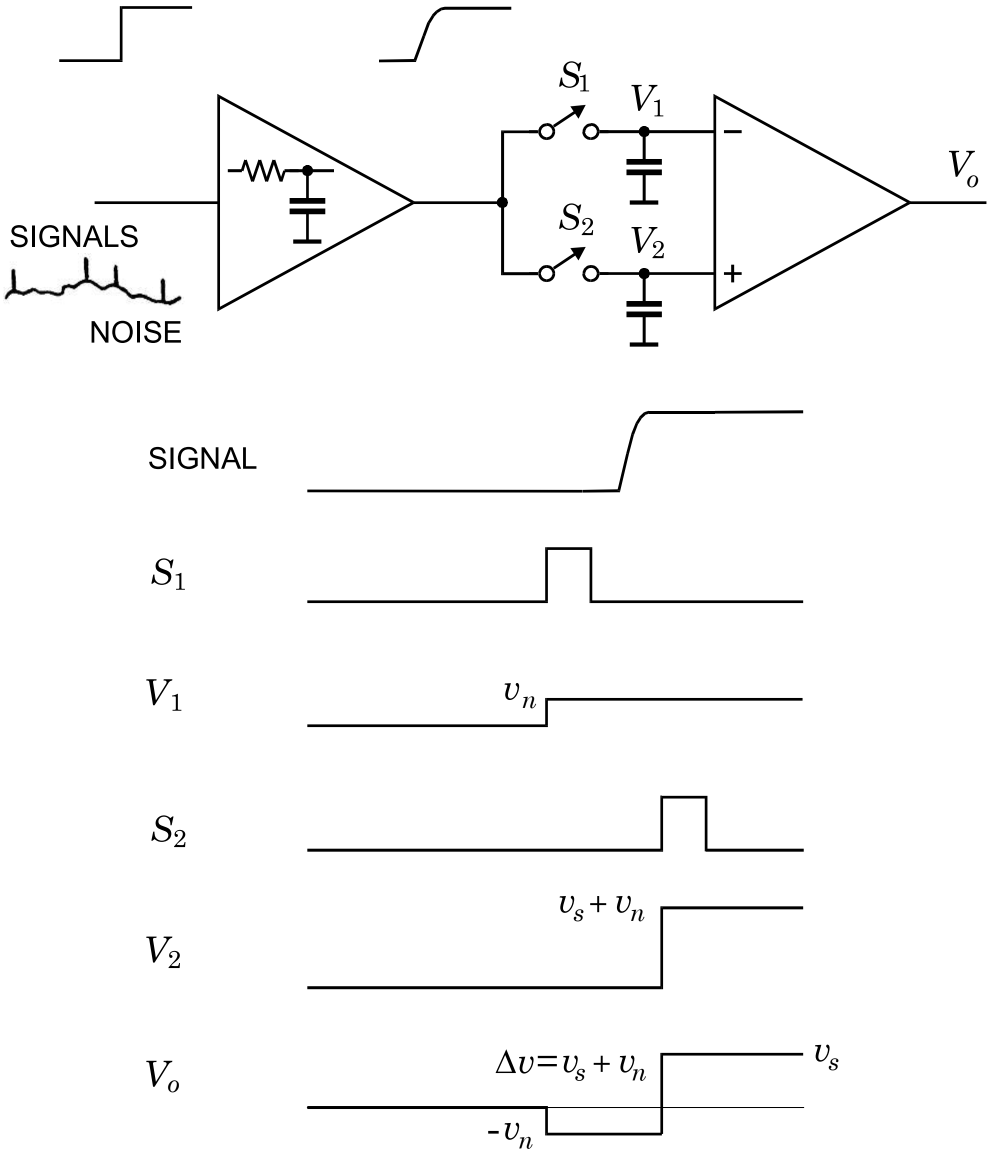
A commonly used time-variant filter is the correlated double-sampler.

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.

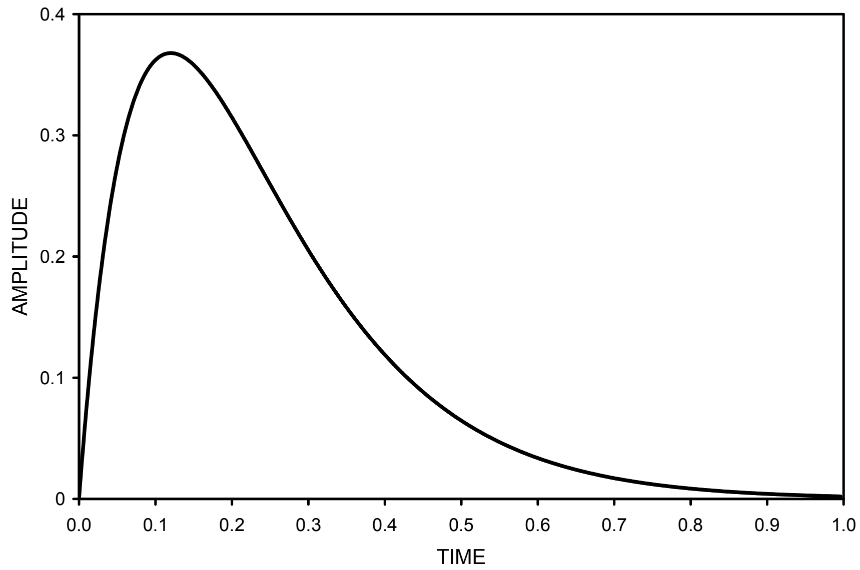
The rest of the world uses equivalent noise voltage and current.

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

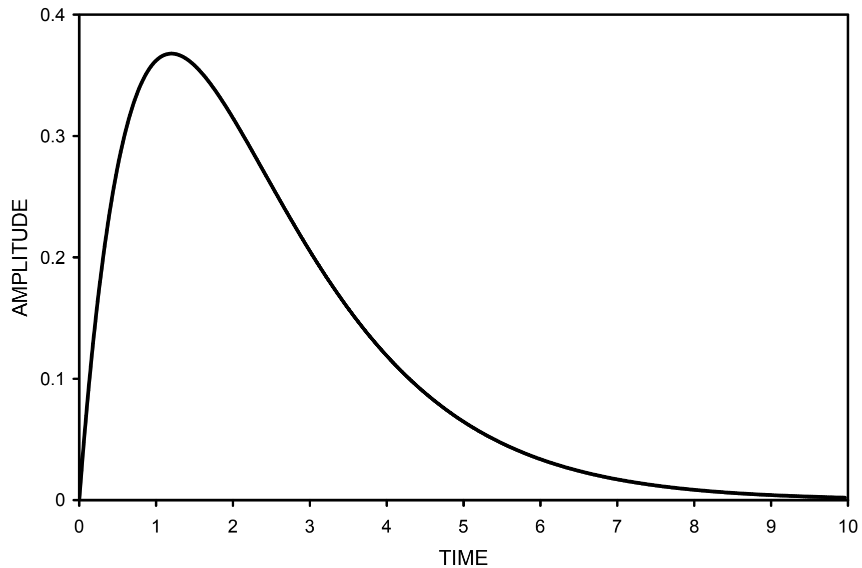
Scaling of Filter Noise Parameters

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

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

$\uparrow$ current noise, $\propto \tau$, independent of C_D

$\uparrow$ voltage noise, $\propto 1/\tau$, $\propto C_D{}^2$

$\uparrow$ $1/f$ noise, independent of τ, $\propto C_D{}^2$

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

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.

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.

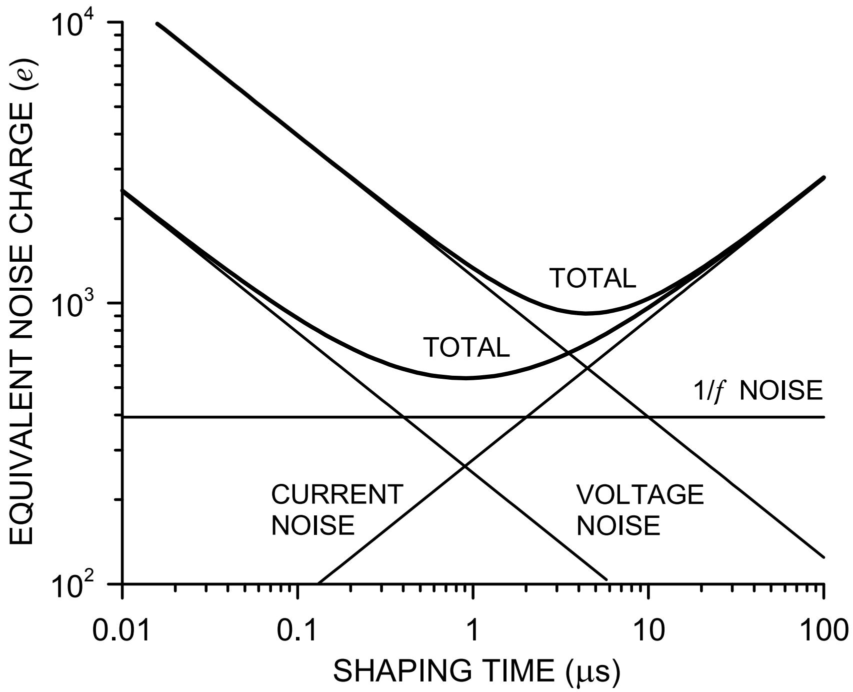
Typical values of F_i , F_v

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 ($T_{opt} \approx$ 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).

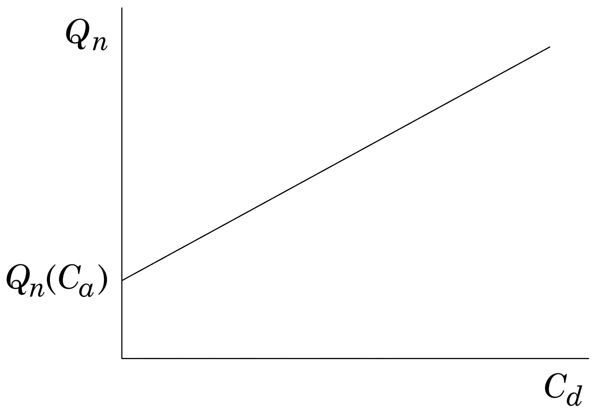
Equivalent Noise Charge vs. Detector Capacitance ($C = C_d + C_a$)

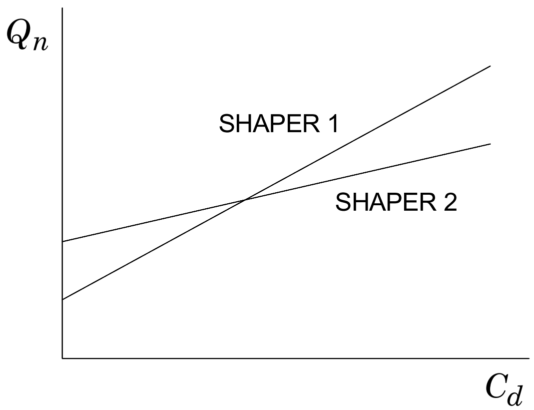
$$Q_n = \sqrt{i_n^2 F_i T + (C_d + C_a)^2 e_n^2 F_v \frac{1}{T}}$$

$$\frac{dQ_n}{dC_d} = \frac{2C_d e_n^2 F_v \frac{1}{T}}{\sqrt{i_n^2 F_i T + (C_d + C_a)^2 e_n^2 F_v \frac{1}{T}}}$$

If current noise $i_n^2 F_i T$ is negligible, i.e. **voltage noise dominates:** $\frac{dQ_n}{dC_d} \approx \underbrace{2e_n}_{\text{input stage}} \cdot \underbrace{\sqrt{\frac{F_v}{T}}}_{\text{shaper}}$

Zero intercept: $Q_n\big|_{C_d=0} = C_a e_n \sqrt{F_v / T}$





Noise vs. Power Dissipation

Under optimum scaling to maintain signal-to-noise ratio,

input transistor power (≈ preamp power) scales with $(S/N)^2$.

Power Reduction

1. Segmentation reduces detector capacitance

   ⇒ lower noise for given power

2. Segmentation reduces the hit rate per channel

   ⇒ longer shaping time, reduce voltage noise

3. Segmentation reduces the leakage current per channel
   (smaller detector volume)

   ⇒ reduced shot noise, increased radiation resistance

Segmentation is a key concept in large-scale detector systems.
(also to increase radiation resistance)

Example: Optimization of Si detector strip length

Assume reduced signal charge S_{rad}/S_0 due to trapping (radiation damage):

Under optimum scaling to maintain signal-to-noise ratio,

input transistor power (≈ preamp power) scales with $(S_0/S_{rad})^2$.

see Spieler, *Semiconductor Detector Systems*, Ch. 6

Alternative: reduce sensor capacitance

Best to scale strip length by S_{rad}/S_0.

Increases number of readout ICs by S_0/S_{rad}, so

increases power by S_0/S_{rad}

- Digital readout power per channel independent of strip length
- Front-end power dominated by input transistor – scales with $\propto C_{strip}^2 \propto L_{strip}^2$

Total power: $$P_{tot} = N_{strip}\left(P'_{analog}L^2 + P_{digital}\right)$$

Number of strips: $$N_{strip} = \frac{A}{p \cdot L}$$ where $A =$ Area and $p =$ strip pitch

$\Rightarrow$ Power per unit area $$\frac{P_{tot}}{A} = \frac{1}{p}\left(P'_{analog}L + \frac{P_{digital}}{L}\right)$$

Assume analog power for 10 cm strip length: 0.2 mW
(SiGe design by E. Spencer, UCSC)

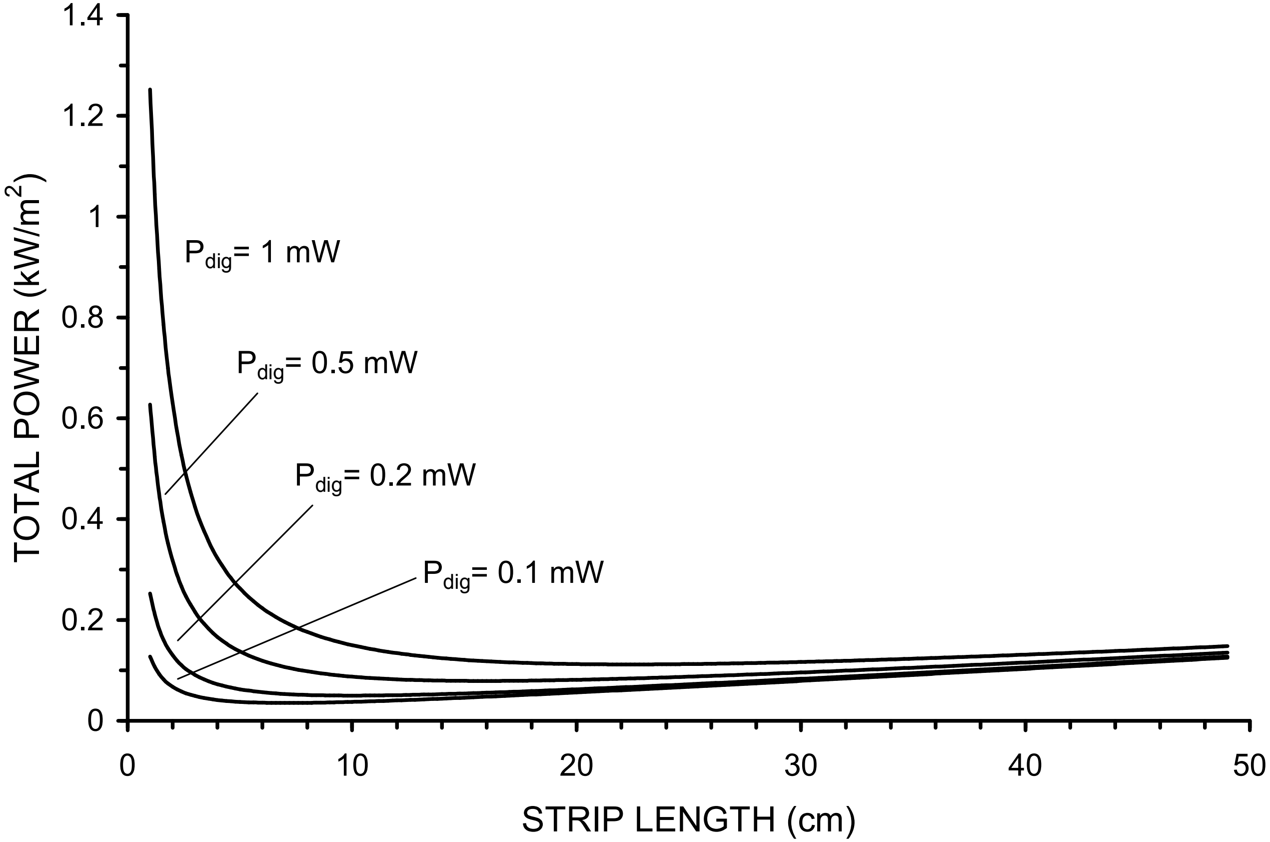
For comparison
ABCD chip digital power: 1.1 mW/ch at 40 MHz clock frequency, $V_{DD} =$ 4V

Digital power scales $\propto$ clock frequency and $\propto 1/$(supply voltage)^2

Note: max strip length also constrained by occupancy

Total Power (kW) per Square Meter vs. Strip Length and Digital Power P_{dig}

(strip pitch = 80 μm, analog power 0.2 mW for 10 cm strip length)



- Power increases rapidly at strip lengths below about 3 cm.
  (Dominated by digital circuitry)
- Important to streamline digital circuitry to reduce its contribution.
  e.g. analyze contributions of individual circuit blocks and assess usefulness.

Digital Power Dissipation

CMOS logic circuits requires little power, but power is absorbed during switching.

Energy dissipated in wiring resistance R:

$$E = \int i^2(t) R \, dt$$

$$i(t) = \frac{V}{R} \exp\left(-\frac{t}{RC}\right)$$

$$E = \frac{V^2}{R} \int_0^\infty \exp(-2t / RC) dt = \frac{1}{2} C V^2$$

If pulses (rising + falling edge transitions) occur at frequency f,

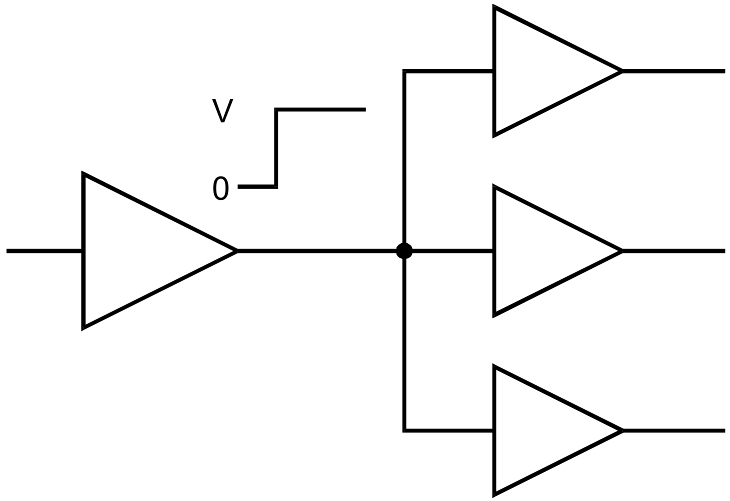
$$P = fCV^2$$

Power dissipation increases with clock frequency and (logic swing)^2.

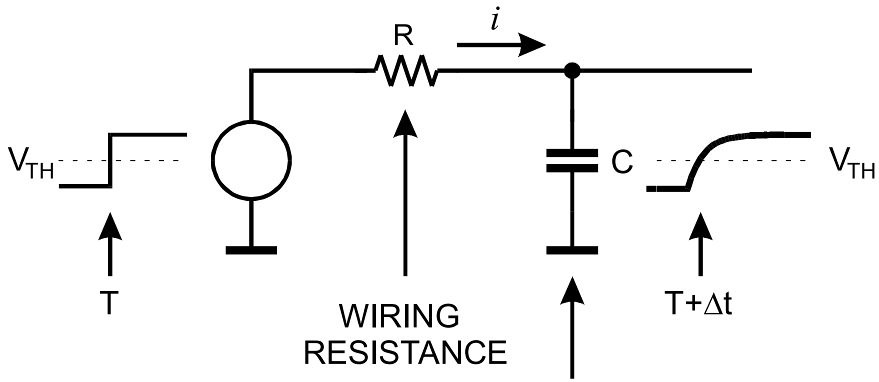
The RC time constant also introduces a time delay.

This depends on the number of driven inputs and wiring lengths.

CASCADED CMOS STAGES



EQUIVALENT CIRCUIT



The above result is often derived from the energy stored in a capacitor.

$$E = \frac{1}{2} C V^2$$

However, this energy will be returned when the capacitor is discharged, so after the leading and trailing edges of a pulse the net energy is zero.

In reality, the power is dissipated by the charge and discharge current flow in the circuit's series resistance.

This is one of many examples where the wrong physics yields a correct result

– until one digs deeper.

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

(Clearly just a partial discussion. For more see Spieler, Chapter 9)

There are many reasons why things don't work ...

1. Idiocy

2. Incompetence

3. Meetings
   (where decisions are made,
   but many participants are incompetent)

However, often there are technical issues.

External Noise Sources

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.

$\Rightarrow$ Baseline fluctuations superimposed on the desired signal

$\Rightarrow$ 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)

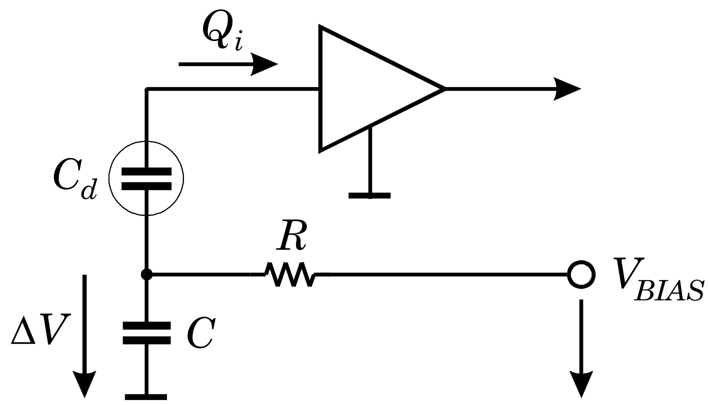
External Pickup is often the cause, but many problems are due to poor work practices or inappropriate equipment

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.

Don't take power supplies for granted ! $\Rightarrow$ Use very low noise power supplies.

Light Pick-Up

Every semiconductor detector is also a photodiode

Sources

- Room lighting (Light Leaks)
- Vacuum gauges

Interference is correlated with the power line frequency
(60 Hz in U.S., 50 Hz in Europe, Japan)

Pickup from incandescent lamps has twice the line frequency
(light intensity $\propto$ voltage squared)

Diagnostics:

a) Inspect signal output with oscilloscope set to trigger mode “line”. Look for stationary structure on baseline. Analog oscilloscope better than digital.

b) Switch off light

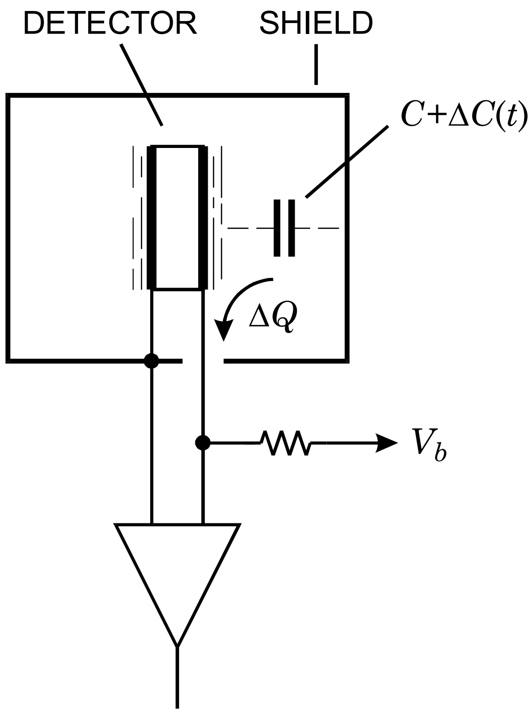
c) Cover system with black cloth (preferably felt, or very densely woven – check if you can see through)

Microphonics

If the electrode at potential V_B vibrates with respect to the enclosure, the stray capacitance C is modulated by $\Delta C(t)$, inducing a charge

$$\Delta Q(t) = V_b \Delta C(t)$$

in the detector signal circuit.



Typically, vibrations are excited by motors
(vacuum pumps, blowers),
so the interference tends to be correlated with the line frequency.

Check with a) oscilloscope on line trigger

b) hand to feel vibrations

This type of pickup only occurs between conductors at different potentials, so it can be reduced by shielding the relevant electrode.

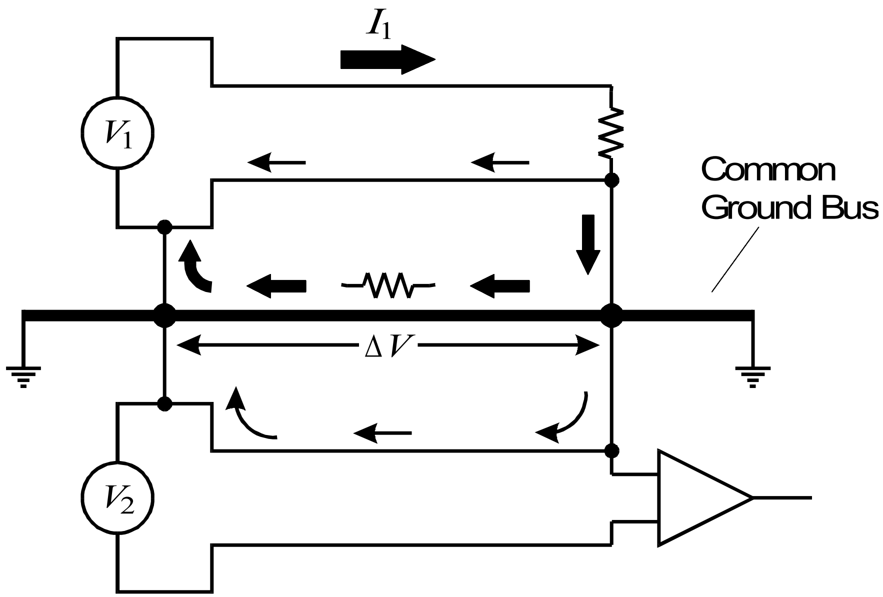
The shield should be at the same potential as the sensitive node.

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.

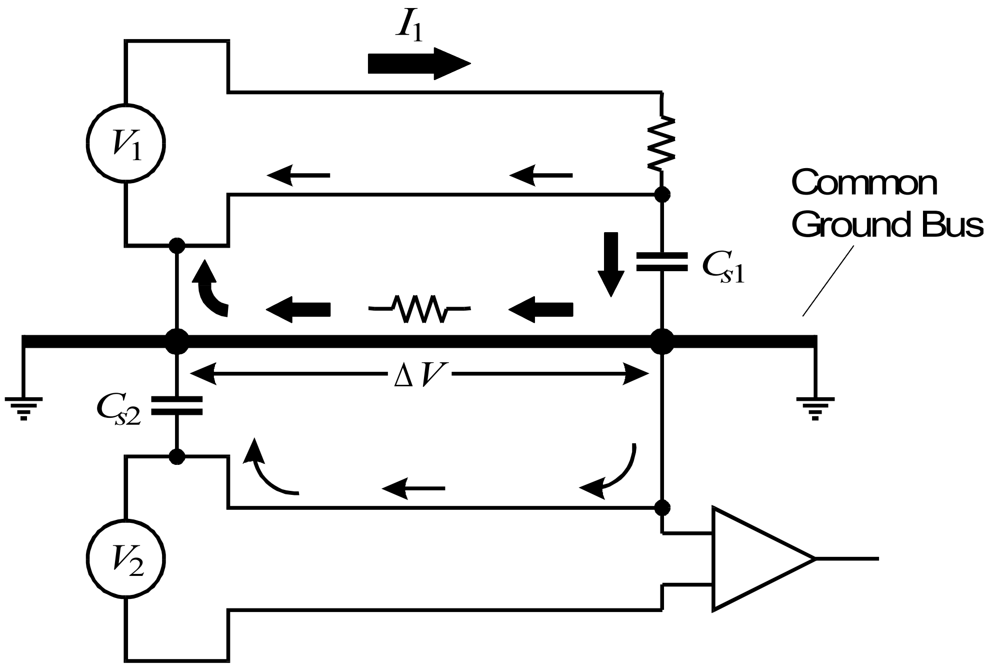
The lower circuit is a sensitive signal transmission path. Following the common lore, it is connected to ground at both the source and receiver.

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.

Remedial Techniques

1. Reduce impedance of the common path

$\Rightarrow$ Copper Braid Syndrome

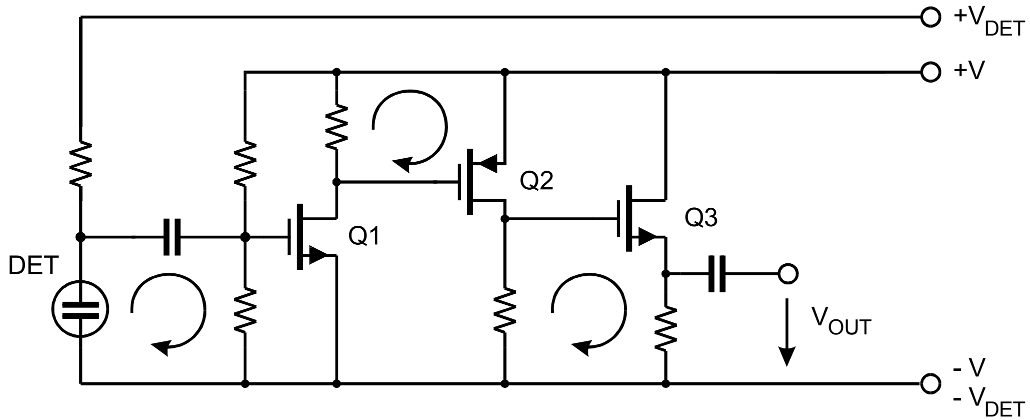
Colloquially called “improving ground”.

Sometimes fortuitously introduces an out-of-phase component of the original interference, leading to cancellation.

Rather haphazard, poorly controlled $\Rightarrow$ continual surprises

2. Avoid Grounds

Circuits rely on current return paths, not a ground connection!



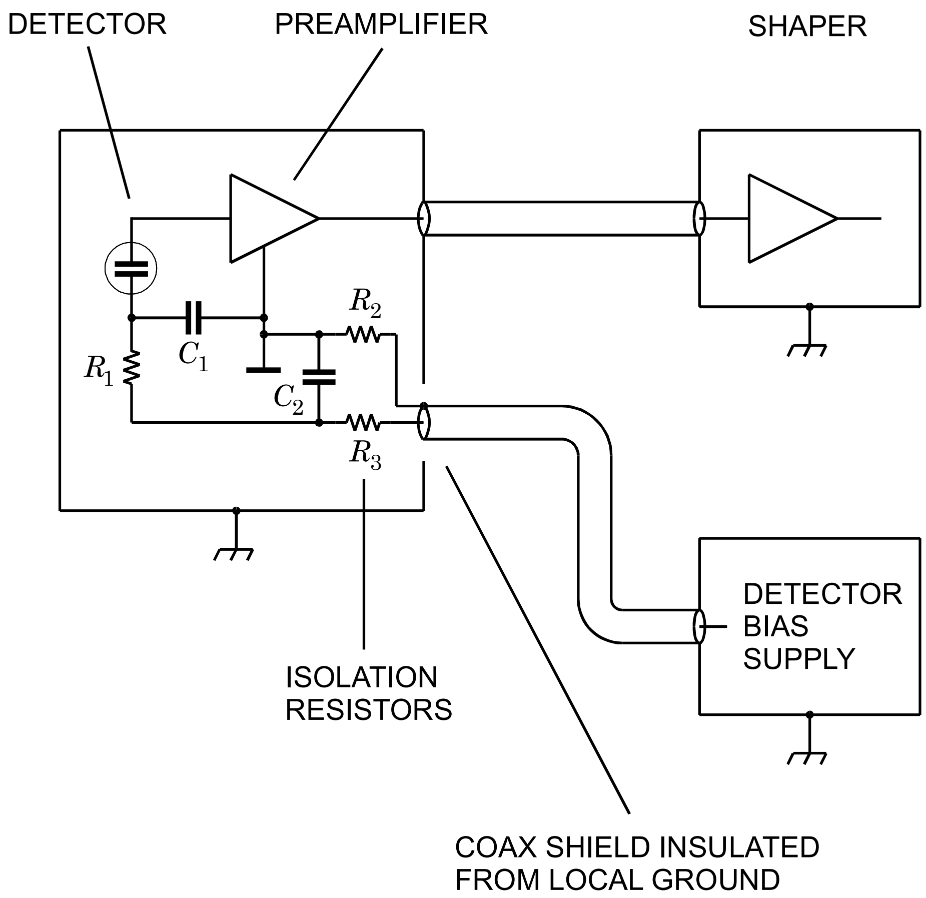
In transferring from stage to stage the signal current flows through local return loops.

1. At the input the detector signal is applied between the gate and source of Q1
2. At the output of Q1 the signal is developed across the load resistor in the drain of Q1 and applied between the gate and source of Q2.
3. The output of Q2 is developed across the load resistor in its drain and applied across the gate and source resistor and load.

Note that – disregarding the input voltage divider that biases Q1 – varying either +V or –V does not affect the local signals.

Series resistors isolate parasitic ground connections.

Example:
detector bias voltage connection



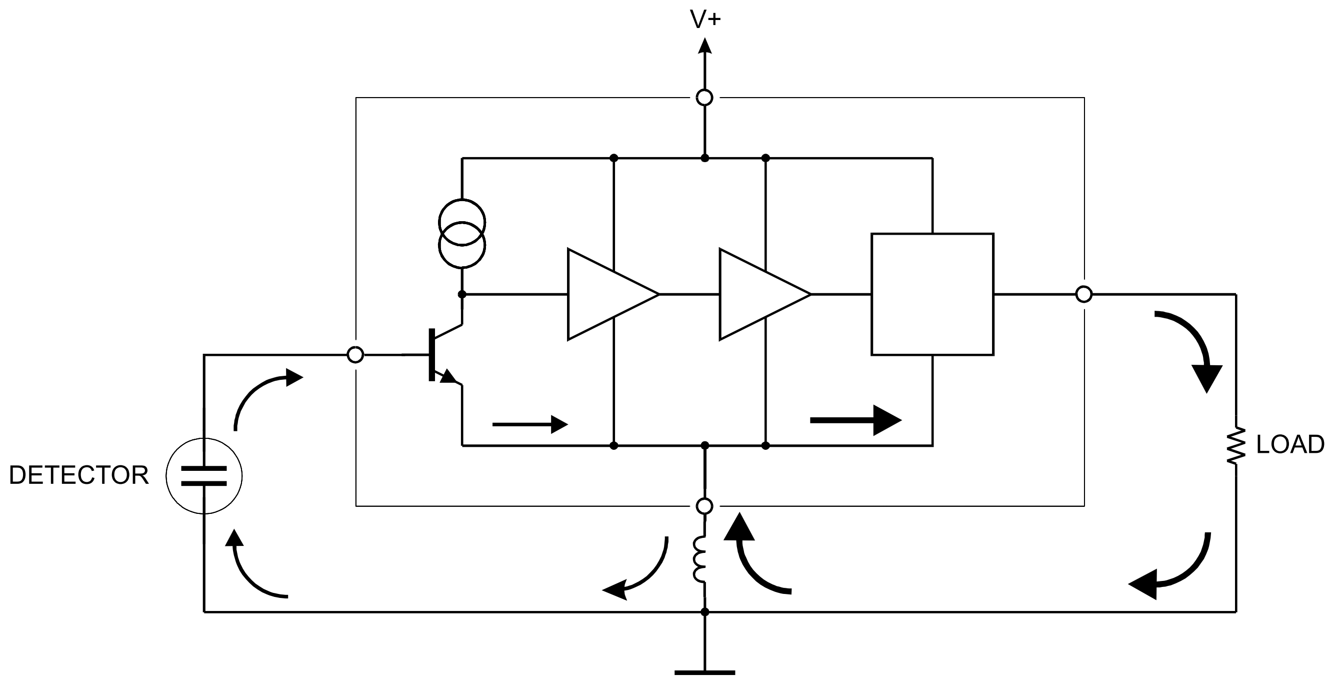
Isolation resistors can also be mounted in an external box that is looped into the bias cable. Either use an insulated box or be sure to isolate the shells of the input and output connectors from another.

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.

A common recipe is to use just one ground connection.

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

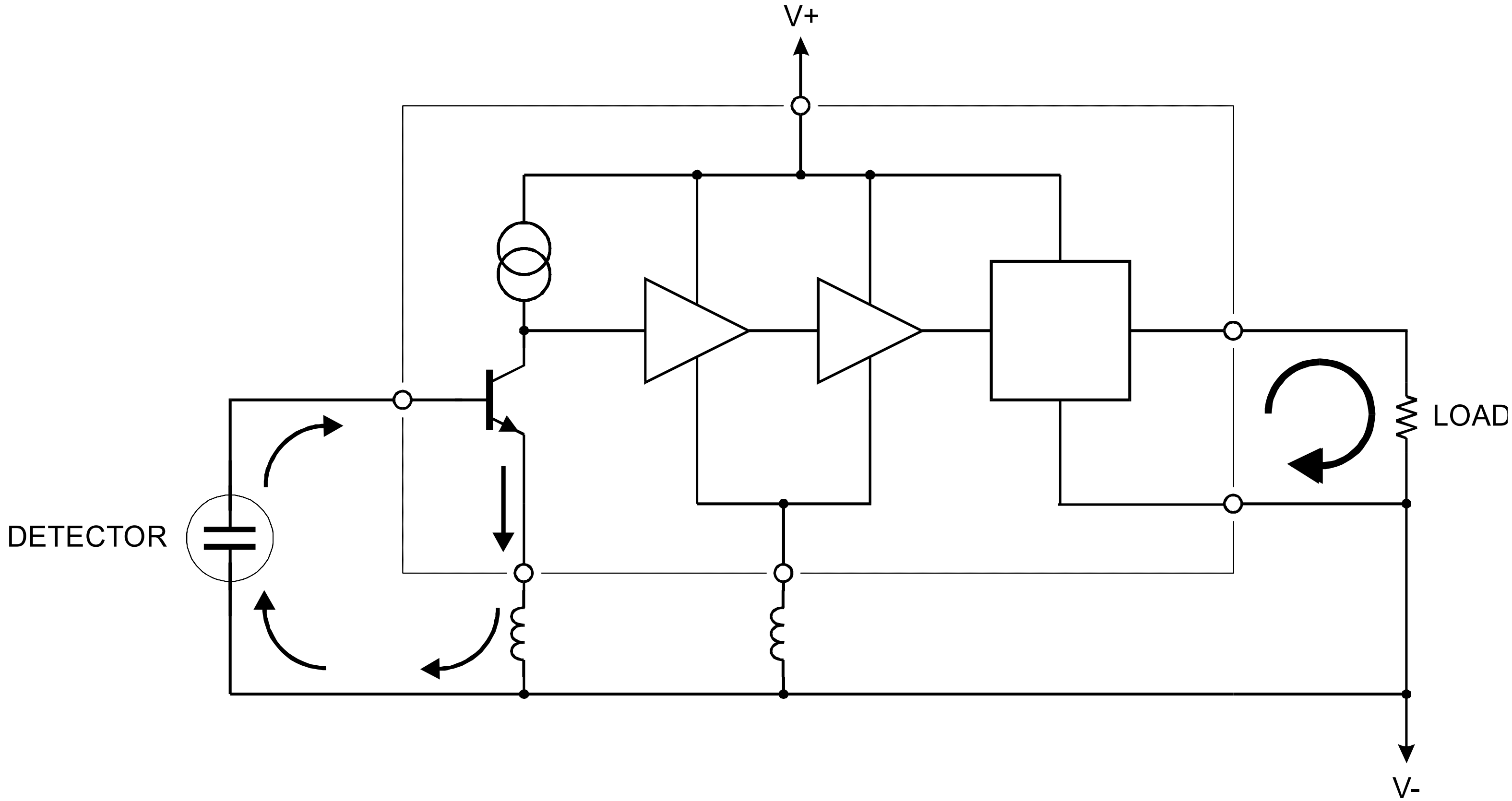


The output current to the load is typically orders of magnitude greater than the input current (due to amplifier gain, load impedance).

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

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.

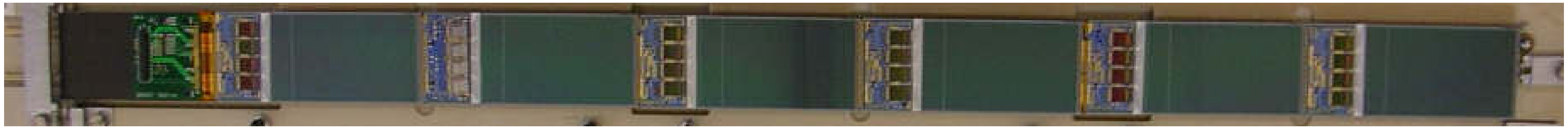
The input current path should be well-controlled and isolated from others.



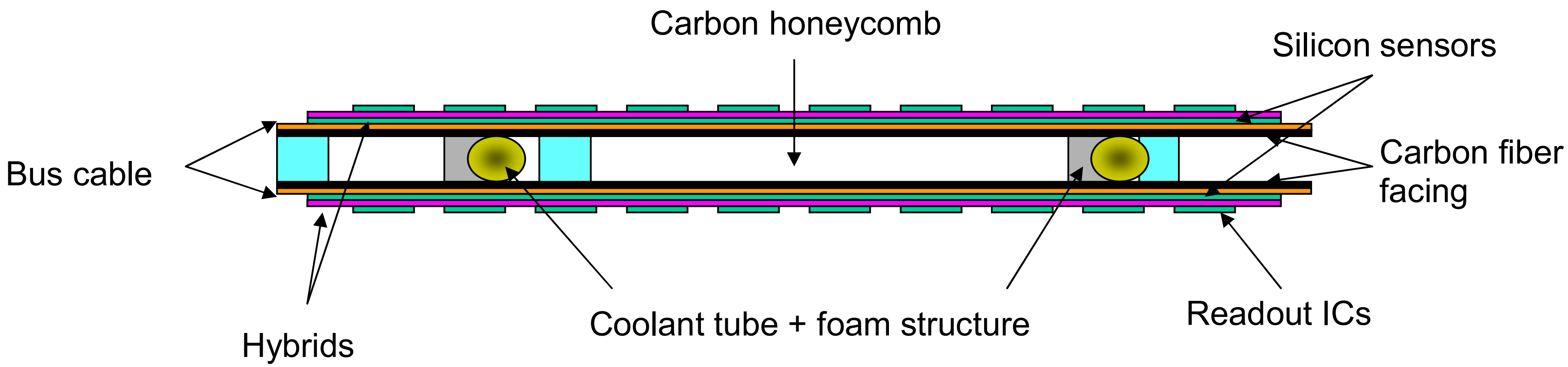
Module Integration

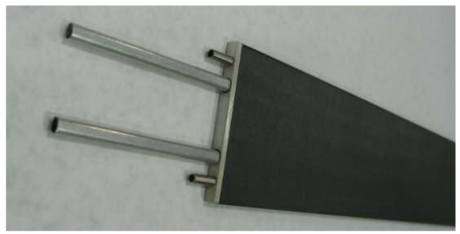
(ATLAS sLHC: figures courtesy of Carl Haber)

Combining mechanical supports with cooling and wiring can reduce material and simplify assembly.



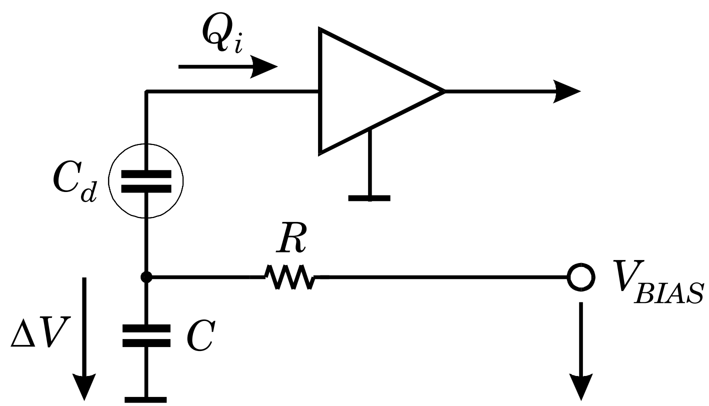
Cross Section:





Pickup

As noted above, the detector bias line is a common source of external noise.



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

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

⇒ Crucial to control pickup on the detector bias line.

Pickup cont'd

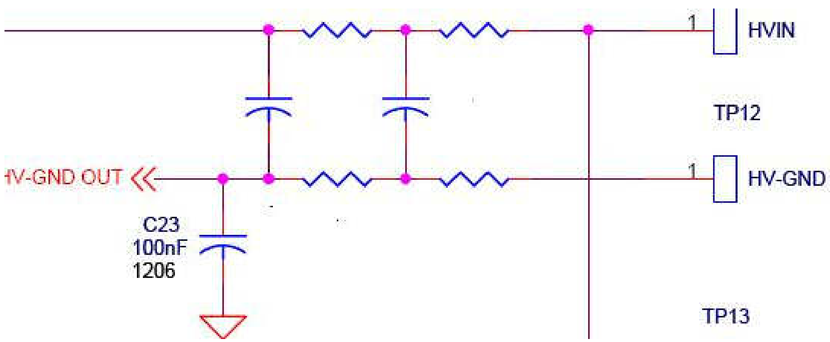
Key design aspects: Hybrid shield layer

Local "grounds"

Symmetric HV filter

To minimize material use "self-shielding" techniques:

Both signal and power lines must be treated as balanced feeds, so any pickup becomes a common-mode effect.



Resistors in "ground" lead must have the same values as those in the "HV" line.
$\Rightarrow$ 10 modules operated in parallel with good performance.

2. Local Referencing

Capacitive coupling is critical when fluctuating voltages are near a sensitive node.

Noise currents on the cooling or support structures can couple to the detector input node.

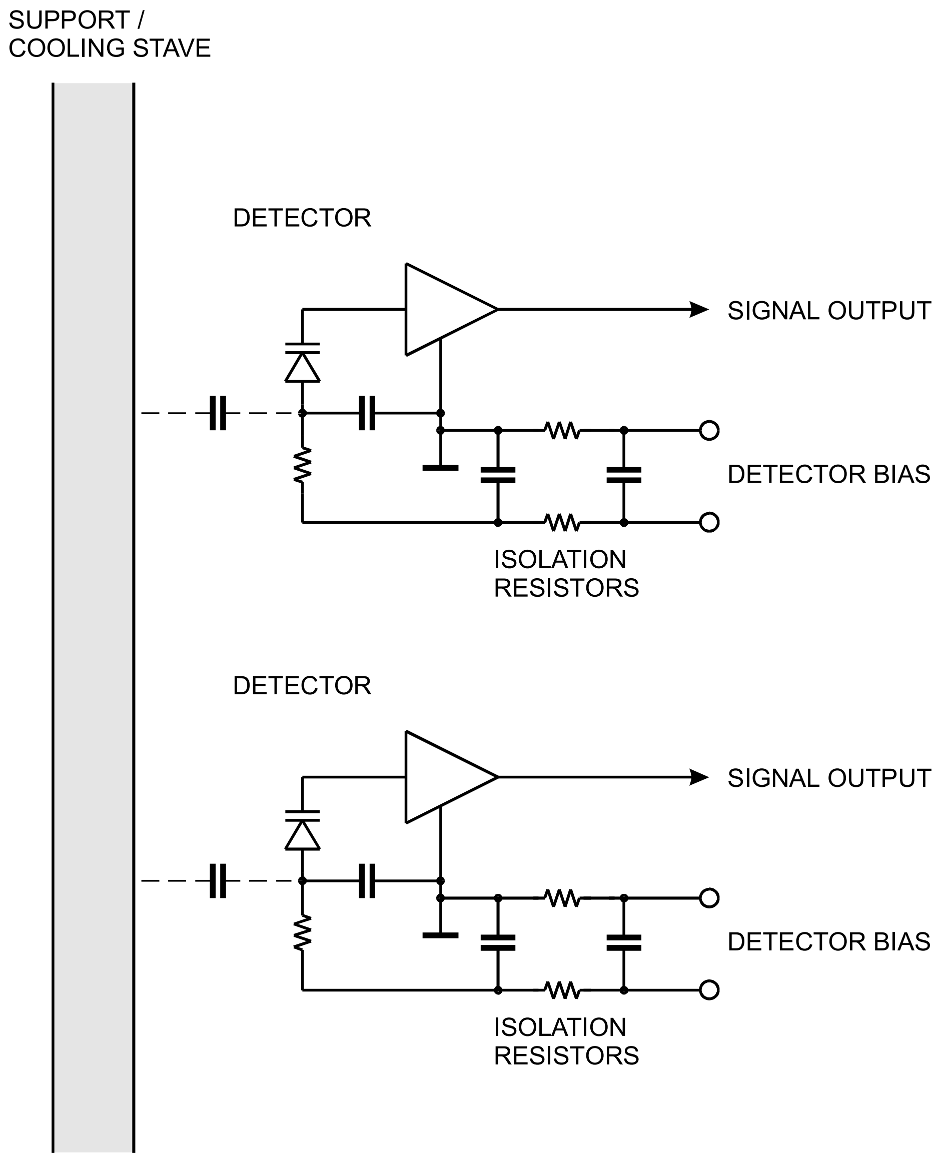
⇒ Keep stave and module at same high-frequency potential

Keep mounting capacitance small,

Control spurious signals on mount

Easier said than done.

Often the rationale for “grounding”, but think of the physical mechanisms before blindly following recipes.



Closing Comments

1. Don't just follow recipes – think physics!
2. Statements may sound good, but what counts is whether they are correct.
3. The key aspects of detectors and electronics can be recognized if you really understand physics.
4. Just talking physics is not good enough – implementing ideas to test their correctness is essential
5. Don't blindly accept the results of simulations. Do cross checks!
6. Bugs are not just technical, but also intellectual.

Following Confucius, yins at the extreme turn into yangs.

Science projects driven by recipes turn into engineering
and engineering projects driven by physics can turn into science.

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