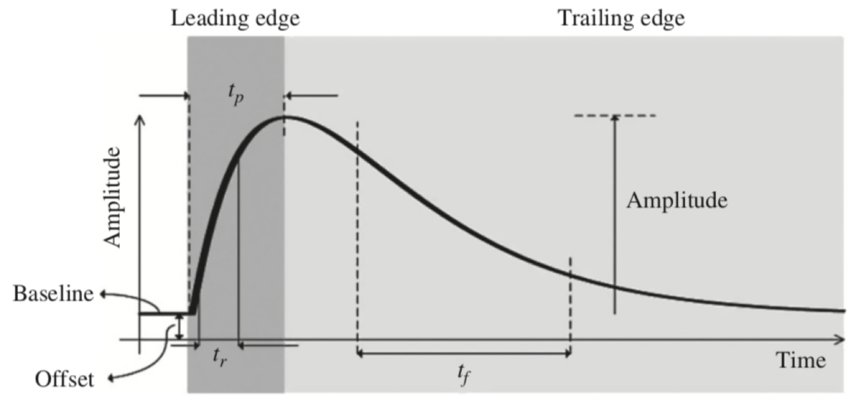
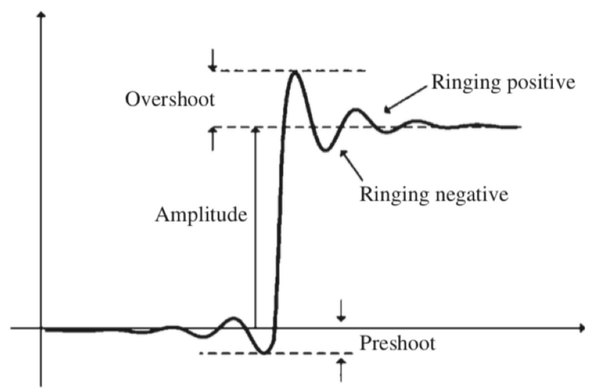
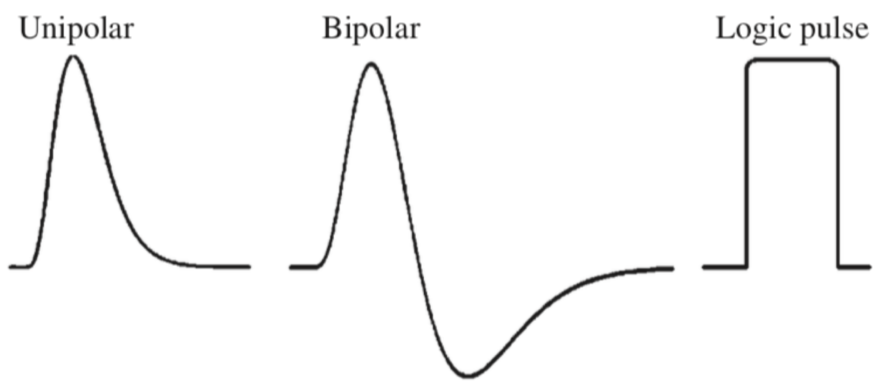
chap.7 Electronics for Pulse Processing

Pulse signal terminology

1) Baseline
2) Pulse Height or Amplitude
3) Signal Width
4) Leading Edge
5) Falling Edge
6) Rise Time
7) Fall Time
8) Unipolar and Bipolar







Fast and Slow signals

- Fast pulses are very important for timing applications and high count rates; in these applications it is very important to preserve their rapid rise time throughout the electronics system.
  ex. Time of flight, coincidence or anti-coincidence...

- Slow pulses, on the other hand, are generally less susceptible to noise and offer better pulse height information for spectroscopy work.
  ex. Amplitude, Energy..

- Fast signals must be treated differently from slow signals, because of their much greater susceptibility to distortion from small, stray capacitances, inductances and resistances in the circuits and interconnections.

- For this reason, essentially two standardized electronics(NIM etc.) have arisen: one designed for fast ns pulses and the other for slower pulses.

Fourier transform

Suppose f(x) is absolutely integrable in $(-\infty, +\infty)$, we can describe the f(x) in terms of an infinite sum of sines and cosines

$$f(x) = \frac{a_0}{2} + \sum_{m=1}^{\infty}(a_m \cos mx + b_m \sin mx),$$

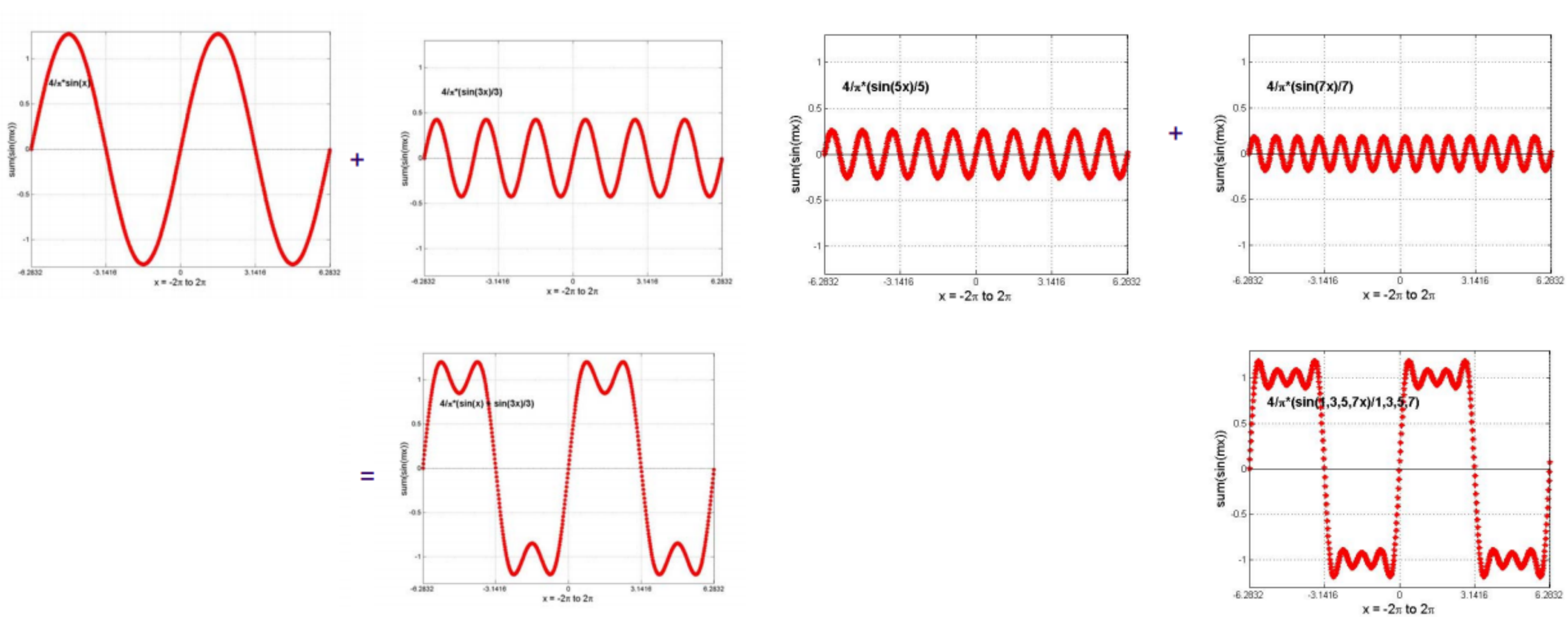
$$a_m = \frac{1}{\pi}\int_{-\pi}^{\pi} f(x)\cos mx dx$$

$$b_m = \frac{1}{\pi}\int_{-\pi}^{\pi} f(x)\sin mx dx$$

$$a_0 = \frac{1}{\pi}\int_{-\pi}^{\pi} f(x) dx$$

Period function $$f(x) = \begin{cases} 1 & 0 < x < \pi \\ -1 & -\pi < x < 0 \end{cases}$$

$$f(x) = \frac{4}{\pi}\left(\frac{\sin x}{1} + \frac{\sin 3x}{3} + \frac{\sin 5x}{5} + \ldots\right)$$



In general, for any non-periodic function $f(t)$

$$f(t) = \frac{1}{2\pi}\int_{-\infty}^{\infty} F(\omega)e^{i\omega t}\,d\omega \qquad F(\omega) = \int_{-\infty}^{\infty} f(t)e^{-i\omega t}\,dt$$

Time domain ↔ Frequency domain

■ Example

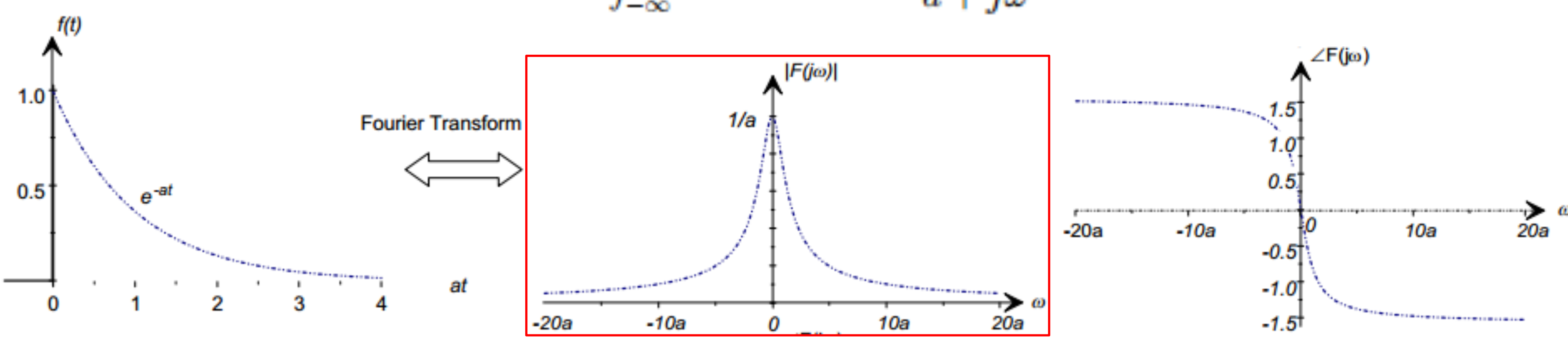
Find the Fourier transform of the one-sided real exponential function

$$f(t) = \begin{cases} 0 & t < 0 \\ e^{-at} & t \geq 0. \end{cases}$$

(for $a > 0$) as shown in Figure 17.

Solution: From the definition of the forward Fourier transform

$$F(j\omega) \;=\; \int_{-\infty}^{\infty} f(t)e^{-j\omega t}dt \;=\; \frac{1}{a + j\omega}$$

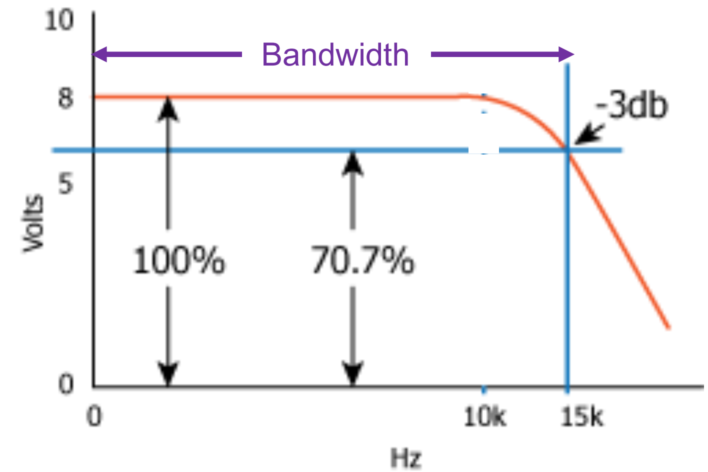


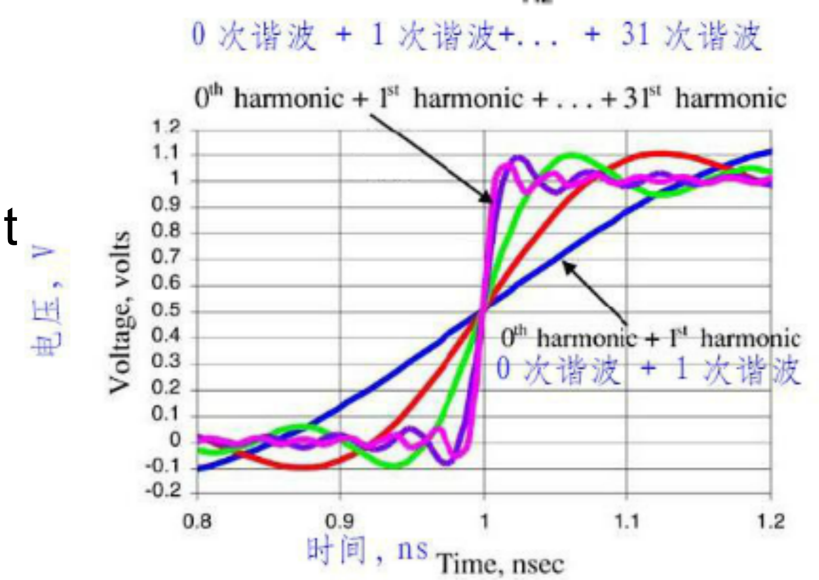
$$|F(j\omega)| \;=\; \frac{1}{\sqrt{a^2 + \omega^2}} \qquad \angle F(j\omega) \;=\; \tan^{-1}\left(\frac{-\omega}{a}\right)$$

Bandwidth

- In any real circuit, the frequency response is limited to a finite range in ω.
- Bandwidth: the range of frequencies delimted by the points at which the response falls by 3dB, represents the range of acceped frequencies.
- high frequency components allow the signal to rise sharply, while the lower frequencies account for the flat parts.
- For fast nuclear electronics, must be capable of accepting frequencies up t 500MHz.

$$dB = 20\, log_{10}(V_2/V_1)$$





Effect of Bandwidth on Rise Time

Bandwidth x Rise Time = 0.35 For square pulse

Impedance

- For an inductor, capacitor, resistor, or combination, a sinusoidal current produces a sinusoidal voltage of the same frequency (the magnitude and phase may change)

$$i(t) = I_m \cos(\omega t + \phi_i) \qquad \mathbf{I} = I_m e^{j\phi_i}$$

$$v(t) = V_m \cos(\omega t + \phi_v) \qquad \mathbf{V} = V_m e^{j\phi_v}$$

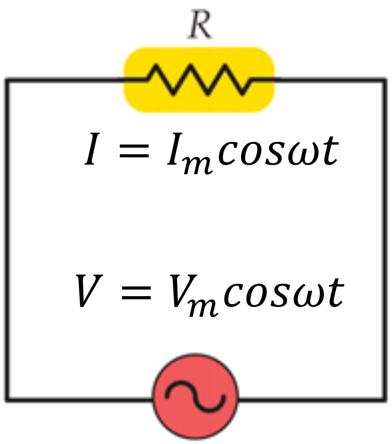
$\mathbf{V} = \mathbf{IZ}$ Ohm's law

$\mathbf{Z} = Z_m e^{j\phi_z}$, which we call the **impedance**.

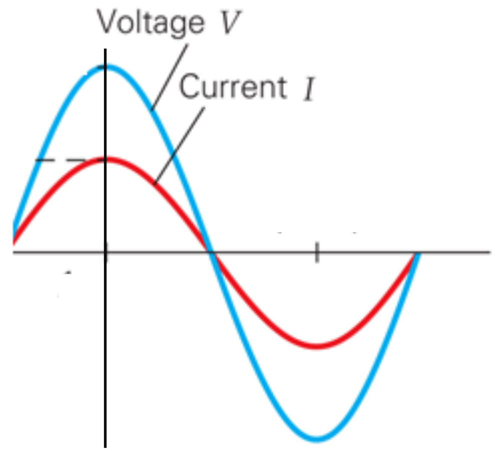
- impedance can be determined by determining the relationship between voltage and current magnitudes and phases, for a particular element (or combination of elements).

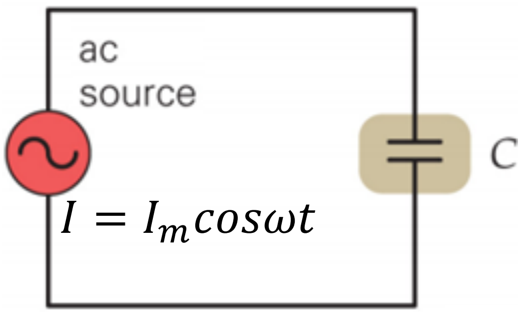
$$\mathbf{Z} = Z_m e^{j\phi_z} = \frac{\mathbf{V}}{\mathbf{I}} = \frac{V_m e^{j\phi_v}}{I_m e^{j\phi_i}} = \frac{V_m}{I_m} e^{j(\phi_v - \phi_i)}$$

$$Z_m = \frac{V_m}{I_m} \qquad \phi_z = \phi_v - \phi_i$$



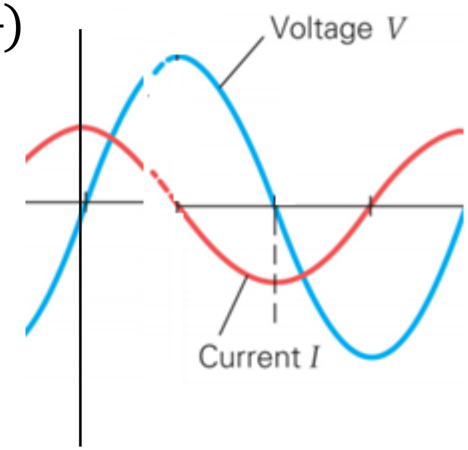
Z=R=V/I





$$v_c = \frac{Q}{C} = \frac{\int Idt}{C} = \frac{1}{\omega C} I_m \cos(\omega t - \frac{\pi}{2})$$

$$Z_c = \frac{1}{\omega C} e^{j(\varphi_v - \varphi_i)} = \frac{1}{j\omega C}$$



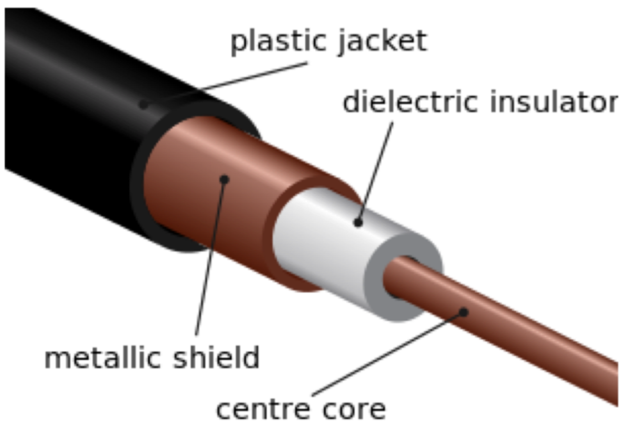
$$j \equiv \cos\left(\frac{\pi}{2}\right) + j\sin\left(\frac{\pi}{2}\right) \equiv e^{j\frac{\pi}{2}}$$

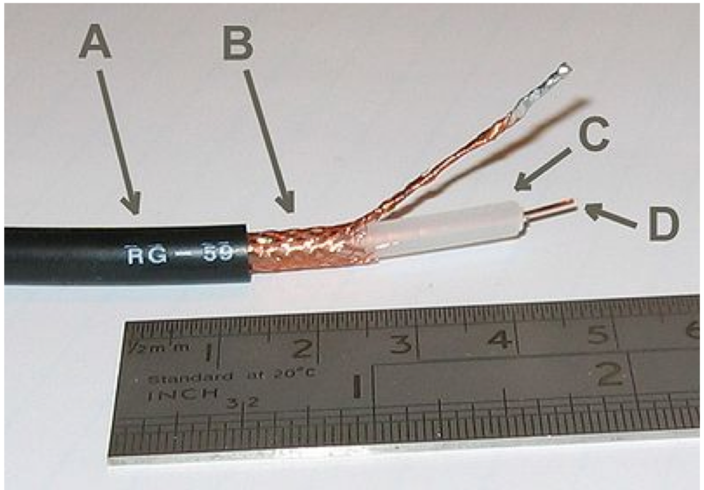
$$\frac{1}{j} \equiv -j \equiv \cos\left(-\frac{\pi}{2}\right) + j\sin\left(-\frac{\pi}{2}\right) \equiv e^{j(-\frac{\pi}{2})}$$

$$Z_{\text{capacitor}} = \frac{1}{j\omega C} \qquad Z_{\text{inductor}} = j\omega L$$

Signal Transmission

- In nuclear electronics, the standard transmission line is the coaxial cable, which minimize losses and noise pick-up.



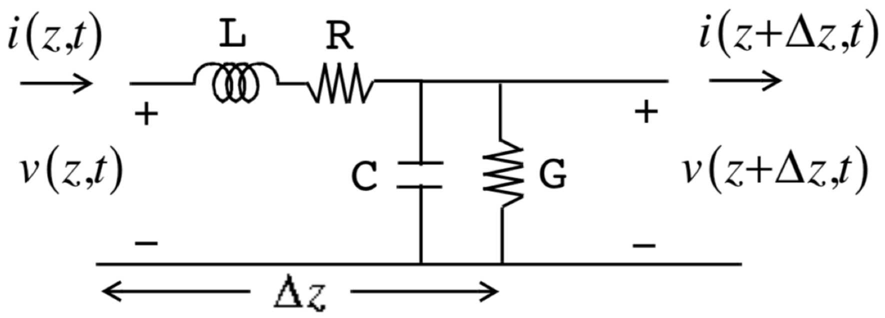


- The entire cable is protected by a plastic outer covering. One advantage of this type of construction is that outer cylindrical conductor, besides serving as the ground return, also shields the central wire from stray electromagnetic fields.

- Frequencies down to ~ 100 kHz are effectively attenuated in most standard cables.

The General Wave Equation for a Coaxial Line

consider a length Δz of a transmission line at location z .



R, L, C, G : per unit length of line.

G: Leakage conductance of the dielectric per unite length.

http://web.mst.edu/~kosbar/ee3430/ff/transmissionlines/leakage_conductance/index.html

A straight forward application of Kirchoff's Loop Law gives

$$v(z+\Delta z,t) - v(z,t) = -\mathrm{L}\,\Delta z \frac{\partial i(z,t)}{\partial t} - \mathrm{R}\,\Delta z\, i(z,t)\,,$$

and Kirchoff's Current Law at the upper node gives

$$i(z+\Delta z,t) - i(z,t) = -\mathrm{G}\,\Delta z\, v(z+\Delta z,t) - \mathrm{C}\,\Delta z \frac{\partial v(z+\Delta z,t)}{\partial t}\,.$$

Dividing through by Δz and taking the limit $\Delta z \to 0$,

$$\frac{\partial v}{\partial z} = -\mathrm{L}\frac{\partial i}{\partial t} - \mathrm{R}\,i$$

$$\frac{\partial i}{\partial z} = -\mathrm{C}\frac{\partial v}{\partial t} - \mathrm{G}\,v$$

$$\Longrightarrow \frac{\partial^2 v}{\partial z^2} = \mathrm{LC}\frac{\partial^2 v}{\partial t^2} + (LG + RC)\frac{\partial v}{\partial t} + RGV$$

For the ideal lossless cable:

R=0 (perfect conductor) G=0 (lossless dielectric)

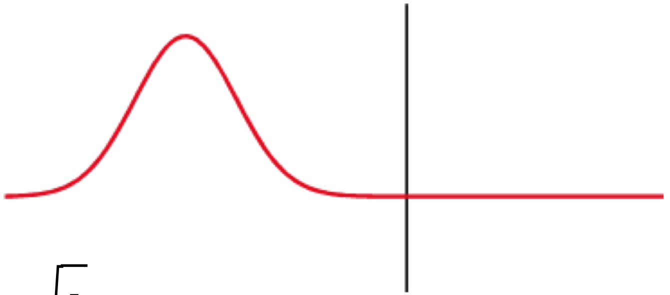
$$\frac{\partial^2 v}{\partial z^2} = \mathrm{LC}\frac{\partial^2 v}{\partial t^2}$$

$suppose\ v(z,t) = v(z)\exp(j\omega t)$,

$$\frac{\partial^2 v}{\partial z^2} = -\omega^2 \mathrm{LC} v = -k^2 v$$

The solution is a traveling wave:

$$v(z,t) = V_+ e^{j(\omega t - kz)} + V_- e^{j(\omega t + kz)}$$



Similarly,

$$\mathrm{i}(z,t) = \frac{V_+}{z_0} e^{j(\omega t - kz)} - \frac{V_-}{z_0} e^{j(\omega t + kz)} \qquad z_0 = \sqrt{\frac{L}{C}}$$

The velocity of propagation is $\mathrm{v} = \frac{\omega}{k} = \frac{1}{\sqrt{LC}}$

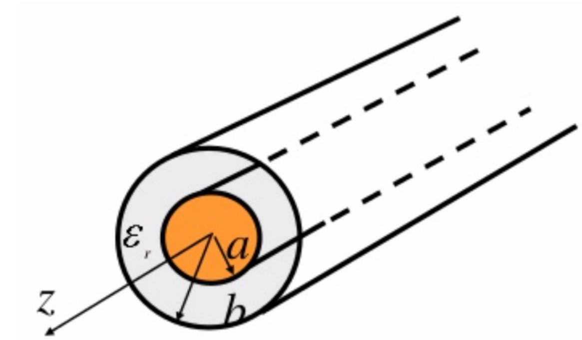
$$\mathrm{T} = v^{-1} = \sqrt{LC}$$

T is known as the delay of the cable, and is typically on the order of ~ 5ns for standard 50Ω cables currently found in the lab.

- Z_o (Characteristic impedance) is the ratio of voltage to current of **one** wave direction.

$$Z_o = \frac{\mathbf{v}_+}{\mathbf{i}_+} = \frac{|\mathbf{v}_-|}{|\mathbf{i}_-|}$$

For this lossless line, Z_0 of a coax. :

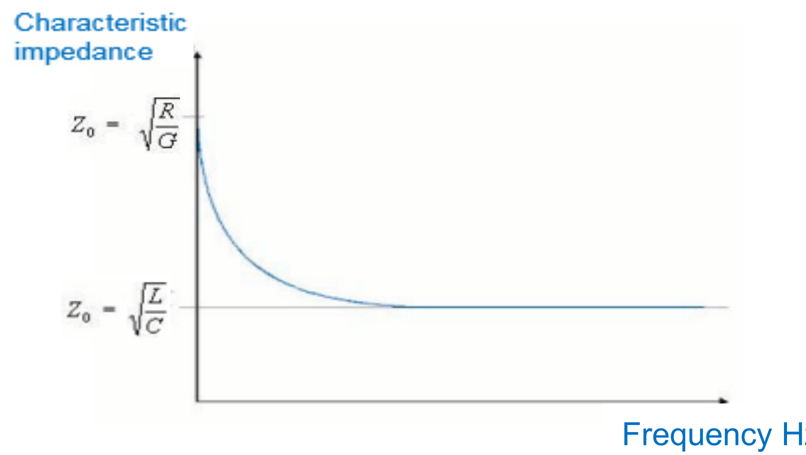


$$C = \frac{2\pi\varepsilon_0\varepsilon_r}{\ln(\frac{b}{a})} \quad F/m \qquad L = \frac{\mu_0}{2\pi}\ln\left(\frac{b}{a}\right) \quad H/m$$

$$z_0 = \sqrt{\frac{L}{C}} = \frac{60}{\sqrt{\varepsilon_r}}\ln\left(\frac{b}{a}\right)$$

For general transmission line,

$$z_0 = \sqrt{\frac{R + j\omega L}{G + j\omega C}}$$



At $DC(0kHz)$: $z_0 = \sqrt{R/G}$

At voice frquencies ($eg.\ 1kHz$): $z_0 = \sqrt{R/\omega C}$

At high frquencies ($> 100kHz$): $z_0 = \sqrt{L/C}$

- Z_o is totally independent of the cable length and only dependent on the cross sectional geometry and the materials used.

Table 13.1. Some common coaxial cable types and their characteristics (data from *LeCroy catalog* [13.1])

Type [RG]	Delay [ns/m]	Diameter [cm]	Capacitance [pF/m]	Max. operating voltage [kV]	Remarks
50 Ω, single braided cables: very widely used					
58U	5.14	0.307	93.5	1.9	Standard cable for fast NIM electronics
58A/U	5.14	0.305	96.8	1.9	
58C/U	5.06	0.295	93.5	1.9	

- Nowadays, almost all high-speed instruments, interconnects, circuits are 50Ω-based

Reflections

- The signal in a coaxial cable is, in general, the sum of the original signal and a reflected signal traveling in the opposite direction. For an arbitrary signal form f

$$V = f(x - vt) + g(x + vt)$$

$$I = \frac{1}{Z_0}(f(x - vt) - g(x + vt))$$

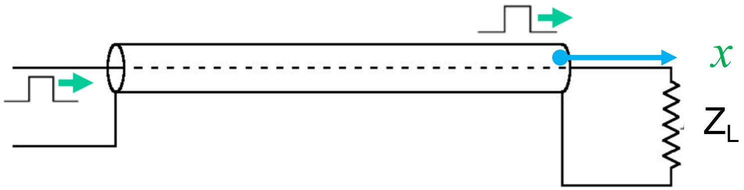
- Cable of characteristic impedance Z_0 terminated by an impedance Z_L

$$V = IZ_L$$

$$f + g = (f - g)\frac{Z_L}{Z_0}$$

$$f(1 - \frac{Z_L}{Z_0}) = -g(1 + \frac{Z_L}{Z_0})$$

$$g = f\frac{Z_L - Z_0}{Z_L + Z_0}$$

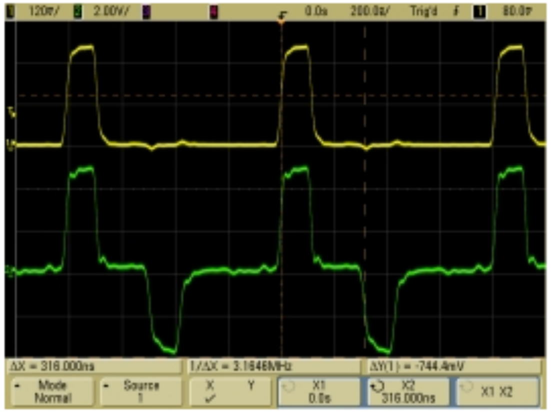


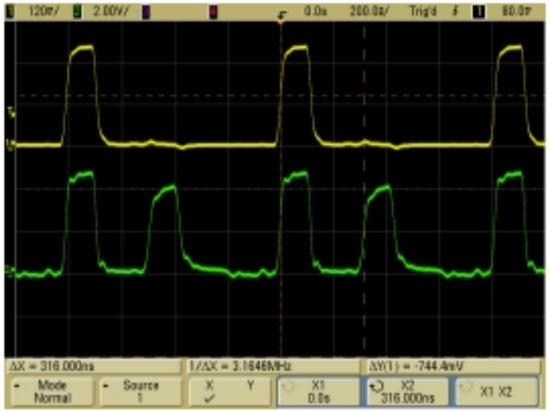
$$\rho = g/f = \frac{Z_L - Z_0}{Z_L + Z_0}$$

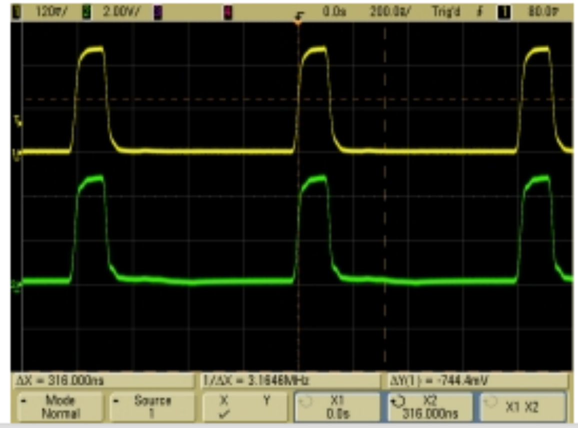
Voltage reflection coefficient

$$\rho = g/f = \frac{Z_L - Z_0}{Z_L + Z_0}$$

- Open circuit termination: $Z_L \to \infty$, $\rho = +1$
- Short circuit termination: $Z_L = 0$, $\rho = -1$
- Impedance matching: $Z_L = Z_0$, $\rho = 0$

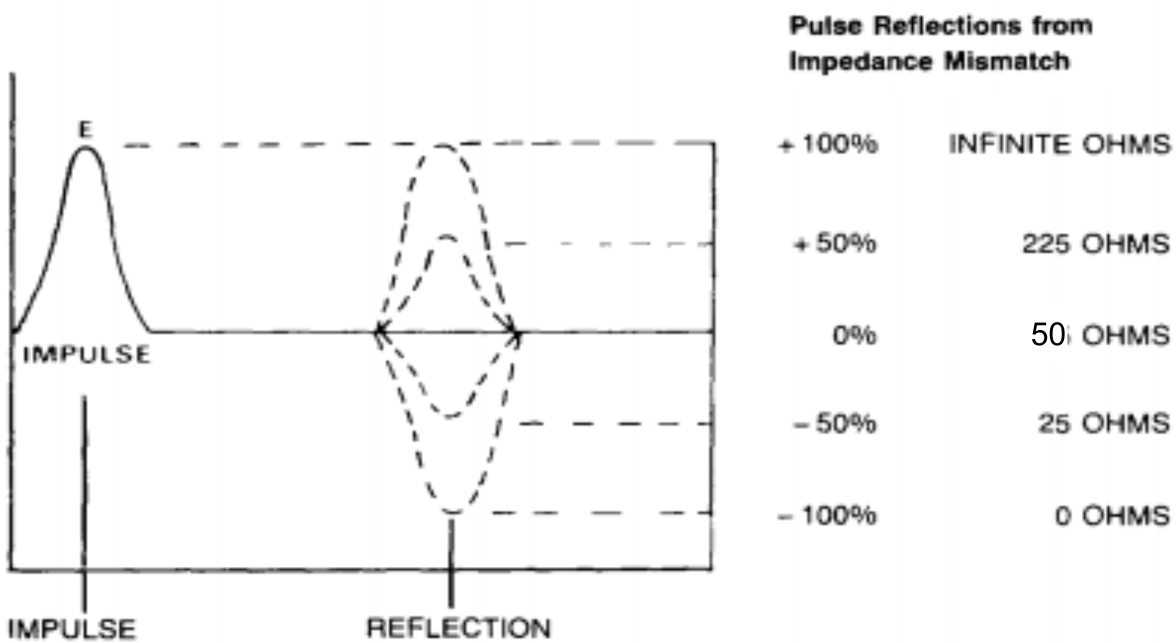






Voltage reflection coefficient: $\rho = \frac{Z_L - Z_0}{Z_L + Z_0}$

The polarity and amplitude of the reflected signal are dependent on the relative values of two impedances. ρ vanishes when $Z_L = Z_0$ and there is no reflection.

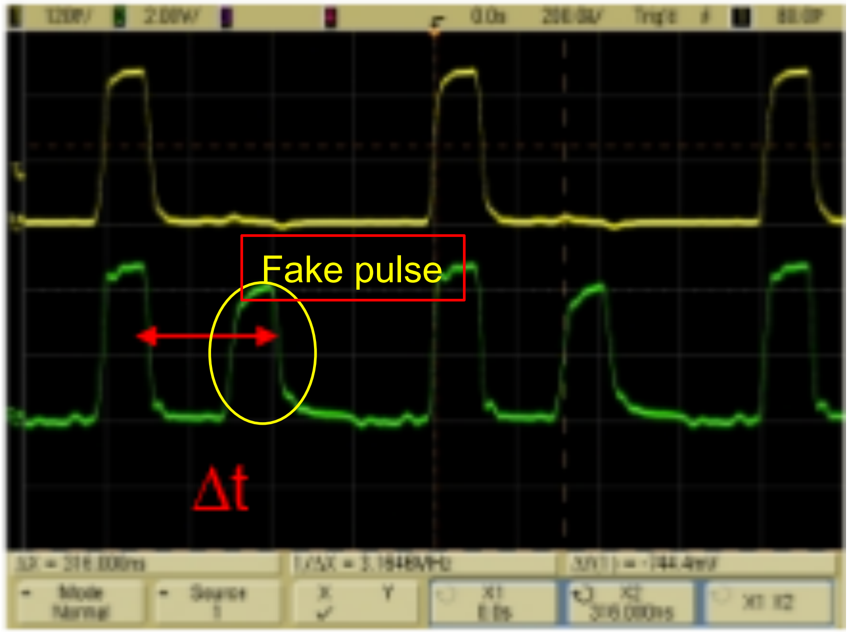


- Reflection can be used for checking cable connection, cable length and cable quality.

Ex: check cable connection inside/outside chamber

When $\Delta t > \tau_r$, Distortion of the signal by reflection is not negligible.

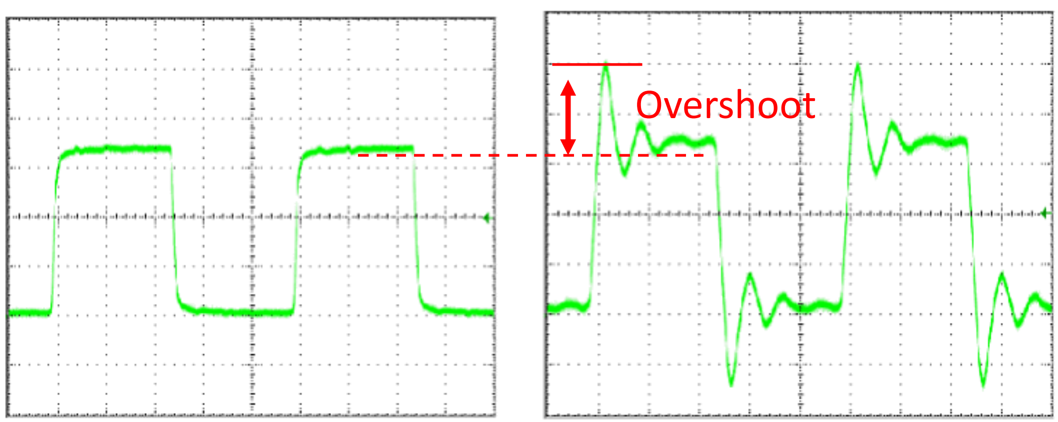
$\Delta t > \tau_r$



Critical for fast signal

- Fake pulse
- Wrong counting

$\Delta t < \tau_r$



Pulse distortion

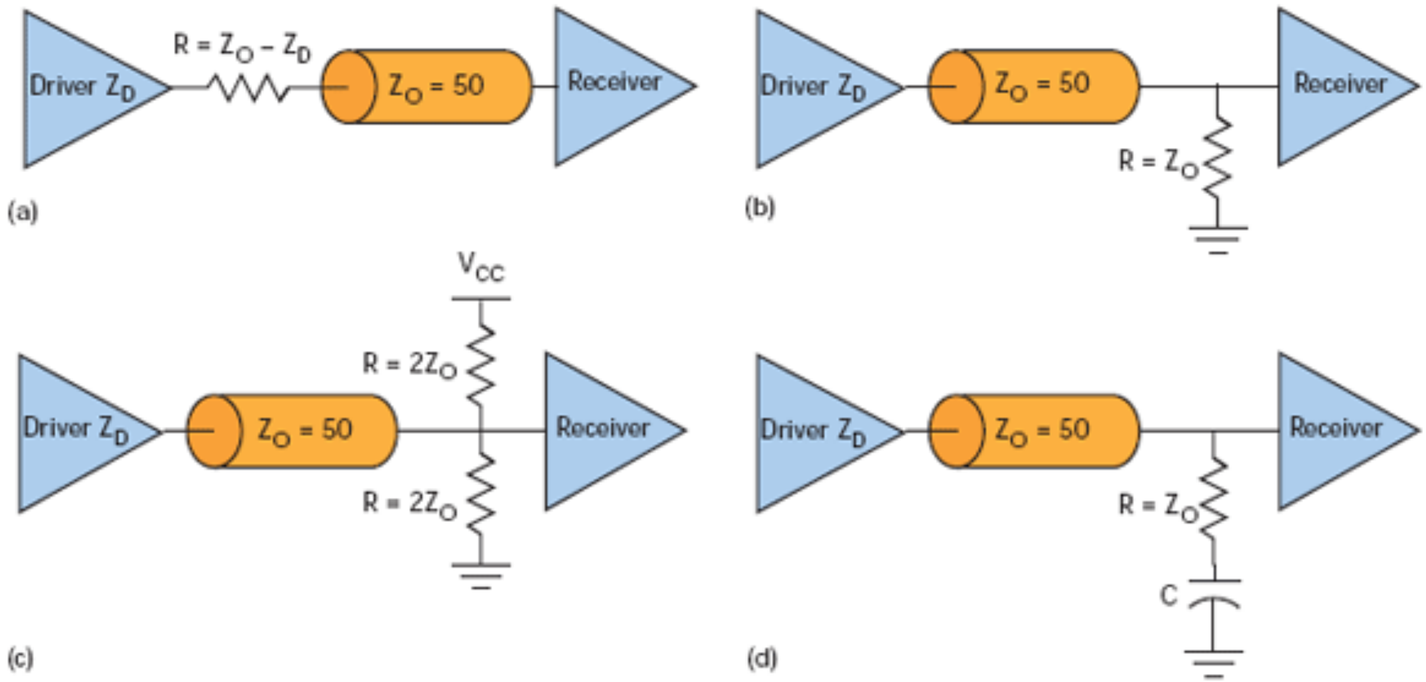
For slow signal(Δt_w>10μs) reflection is negligible.

- The distinction between fast and slow pulses depends on a comparison of the fastest component of the pulse (usually the rise or fall time) with the transit time of the pulse through the cable.

- Pulses with rise times that are slow compared to the transit time are considered slow, while those with rise times comparable with the transit time (or faster) are considered fast pulses. Most pulses derived from `timing or fast logic electronic modules' (e.g leading-edge or constant- fraction discriminators) are fast pulses, while those from linear devices (e.g. amplifiers) are slow.

Cable Termination. Impedance Matching

Generally, termination can be done in two ways:
Either by adding an impedance in series with the load or in parallel.



2. Common termination techniques for clock distribution include series termination (a), parallel termination (b), Thevenin termination (c), and ac termination (d). Series termination is achieved at the driver side, while the other three eliminate reflection at the receiver end.

In most cases, a simple termination at the receiving end is usually sufficient.

Impedance Matching: the other ways

Transformer

$N = N_s/N_p$ = turns ratio

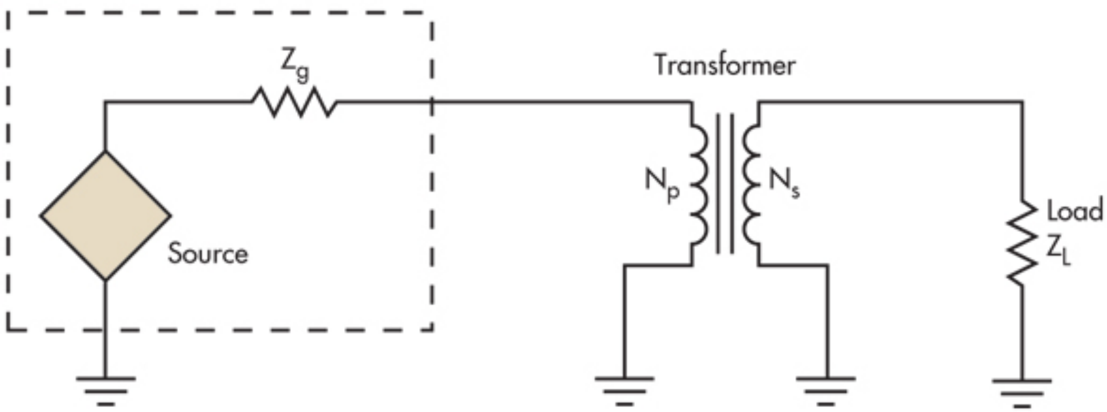


Fig 9. A transformer offers a near ideal method for making one impedance look like another.

Signal transition by optical fiber

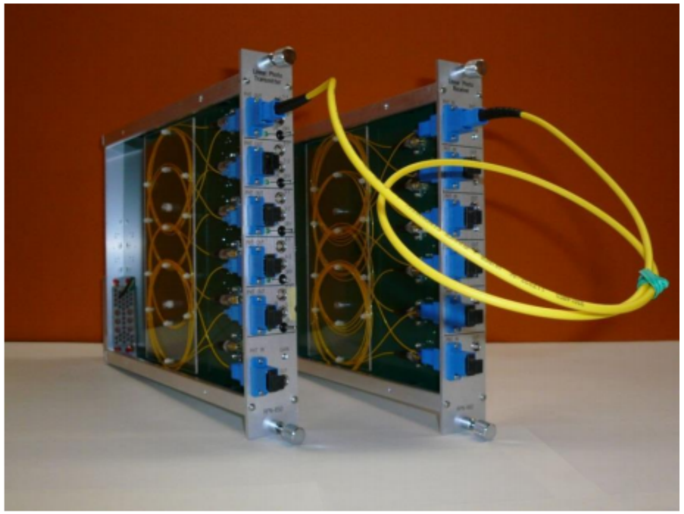
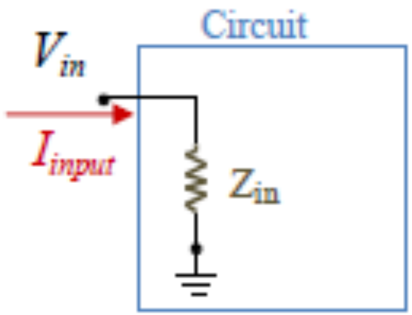
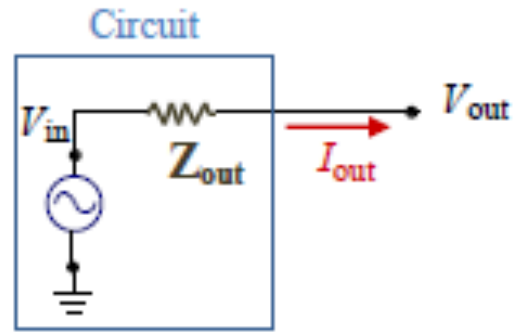


Fig. 22. Photograph of the optical transmitter and receiver NIM modules.

Input impedance & output impedance





Z_{in} = input voltage / (input current)

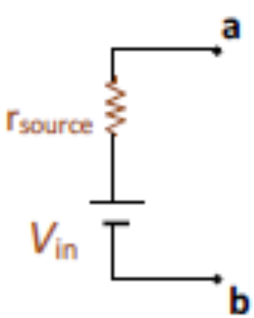
Z_{out} = [V_{in} -V_{out}] / (output current)

Circuit loading The undesirable reduction of the open-circuit voltage V_{ab} by the load.

Voltage source: $\mathbf{v}_{in}$; Internal resistance of the source : $\mathbf{r}_{source}$

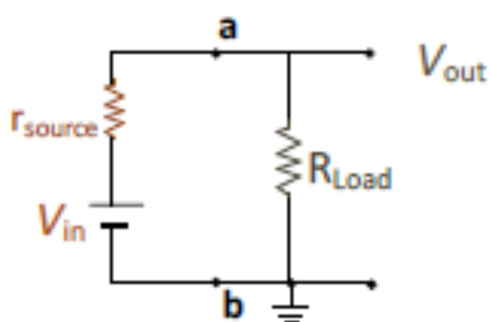
External load resistance: $\mathbf{R}_{load}$

$$V_{out} = V_{in}\frac{1}{1+(r_{source}/R_{Load})}$$

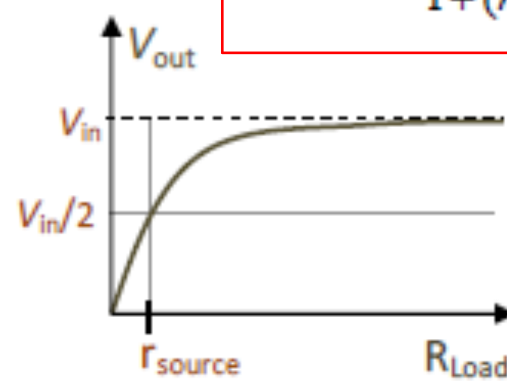


Open circuit

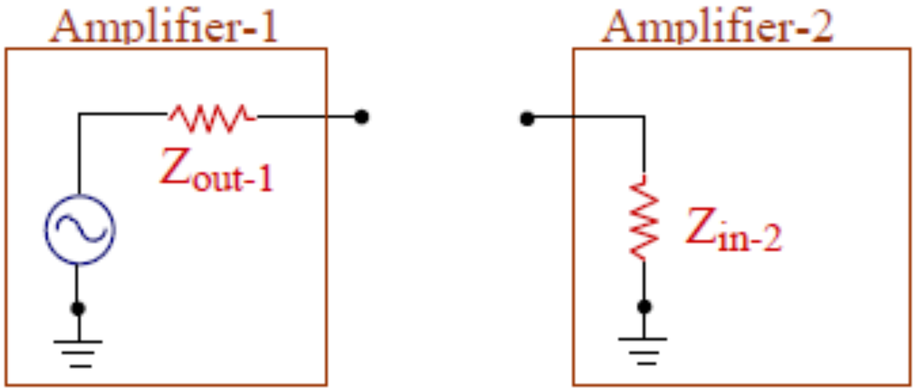
$V_{ab} = V_{in}$



$V_{ab} < V_{in}$



$$V_{out} = V_{in} \frac{1}{1+(r_{source}/R_{Load})}$$



Solution to avoid "loading" the circuit: Use $R_{Load} >> r_{source}$

(Rule of thumb: To use $R_{Load} > 10\ r_{source}$)

Connecting circuits one after another

In electronic circuits, stages are connected one after another.

i) it is always better to have a "stiff source" ($Z_{out} << Z_{in}$), so that signal levels do not change when a load is connected. For energy measurement (slow)

ii) there are situations in which it is rather required to have $Z_{out} = Z_{in}$. That is the case in radiofrequency circuits to avoid signal reflections. For time measurement(fast)

So, be aware to respond accordingly depending on the situation.

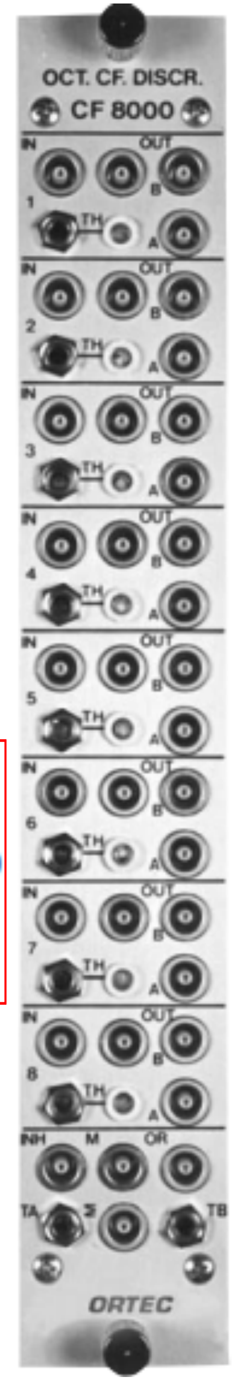
Shaping amplifier
Energy measurement

INPUT BNC front- and rear-panel connectors accept either positive or negative pulses with rise time of 10 to 650 ns and decay times of 40 μs to ∞; Z_{in} = 1000 Ω dc-coupled; linear maximum 10 V; absolute maximum 20 V.

OUTPUTS

UNI Front-panel BNC connector with Z_o <1 Ω and rear-panel connector with Z_o = 93 Ω, short-circuit proof; with full scale linear range of +10 V; active filter shaped; dc-restored, dc level adjustable to ±100 mV.

Output ->ADC (Z_{in}=1K ohm)



Discriminator
time measurement

INPUTS

INPUTS Front-panel LEMO connector for each channel.

INPUT RANGE 0 to −5 V.

PROTECTED TO −100 V for pulse duty cycles <0.05%.

IMPEDANCE 50 Ω, dc-coupled.

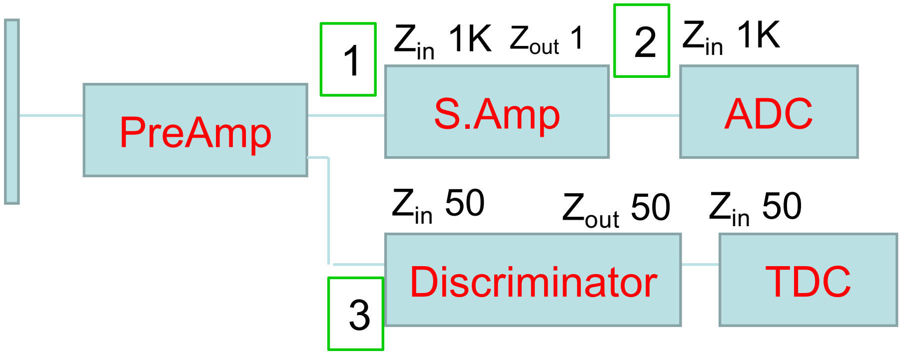
OUTPUTS

Output Impedance 50-Ω, ac-coupled.

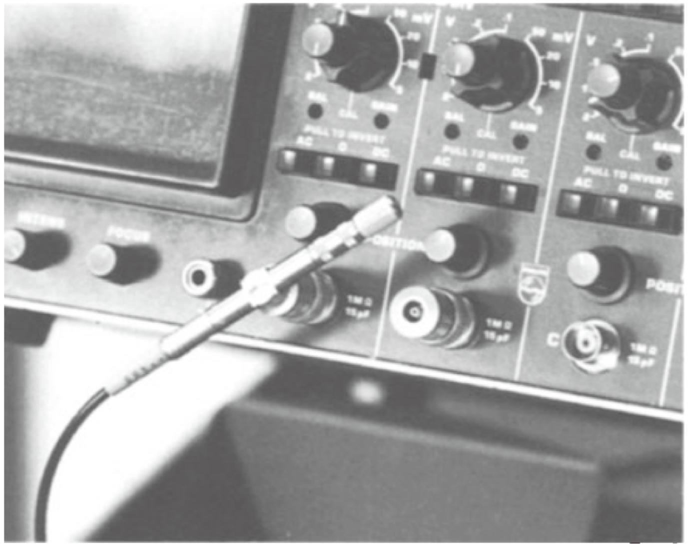
Output -> time circuit(Z_{in}=50 ohm)
-> TDC (Z_{in}=50 ohm)

用示波器观察探测器或设备的输出幅度：

必须选择合适的示波器输入阻抗观察，即选择示波器Z_{in}接近于上一级设备的z_{out}

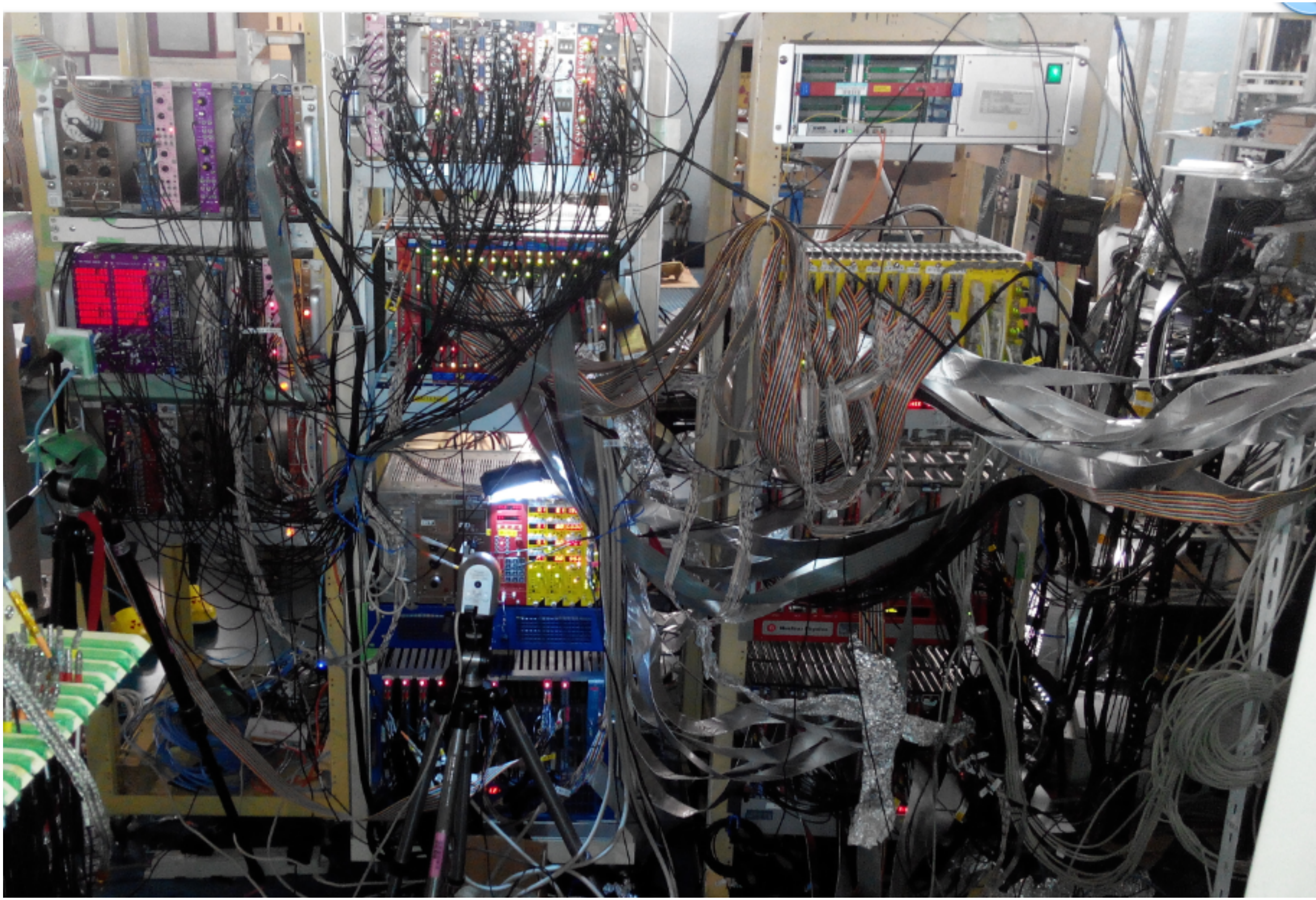


50Ω termination in an Oscilloscope



For an oscilloscope only offers a 1MΩ input impedance, add a T-connector with the signal cable into one end and a 50Ω terminator into the other end.

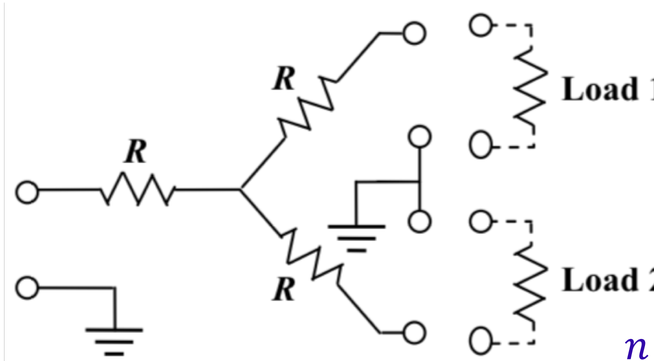
50ohm terminator



Bad habit: hang cables on the modules without proper termination

Pulse splitter

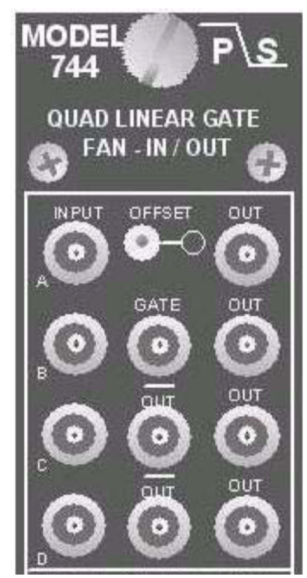
A pulse applied to any terminal will be supplied to the other two, while maintaining a constant impedance. For a 50 Ω splitter, the resistance values should each be 16.6 Ω. In this case, the impedance `looking' into any terminal will be 50 Ω, if the other two terminals are connected to 50 Ω loads. The signal delivered to each load will be only half what it would be if the load were directly applied to the source.



$$R = Z_0 \frac{n-1}{n+1}$$

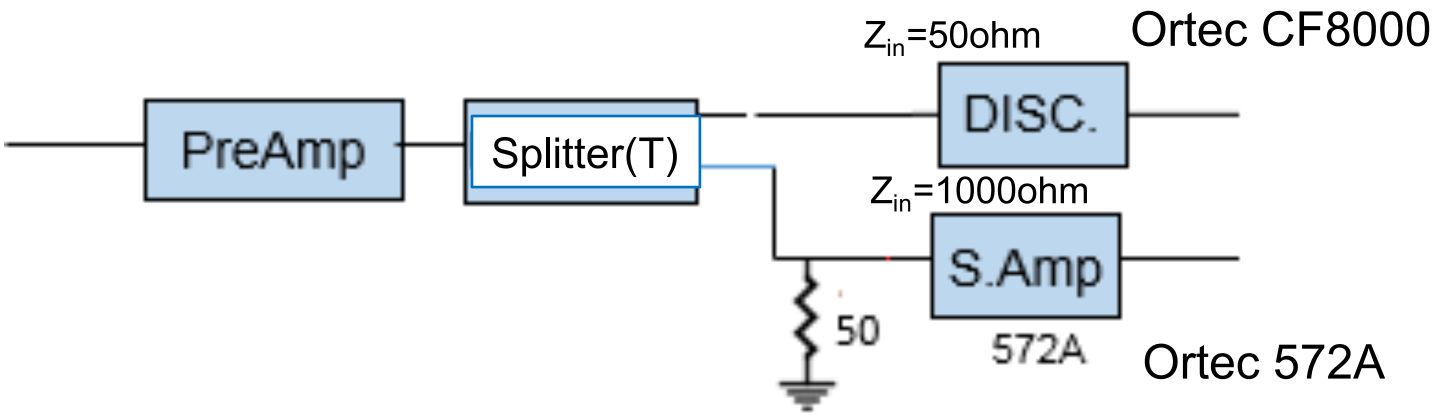
Figure 1.5 A symmetric pulse-splitter circuit. This can be used to supply two loads while maintaining matched impedance levels.

T-connector :
no impedance matching, not good for fast signal.



Linear Fan In / Fan out

Example: signal splitting & Impedance marching



Losses in Coaxial cables

$$\frac{\partial^2 v}{\partial z^2} = \mathrm{LC}\frac{\partial^2 v}{\partial t^2} + (LG + RC)\frac{\partial v}{\partial t} + RGV$$

The solution is

$$v(z,t) = V_+ \boxed{e^{-\alpha z}} e^{j(\omega t - kz)} + V_- e^{\alpha z} e^{j(\omega t + kz)}$$

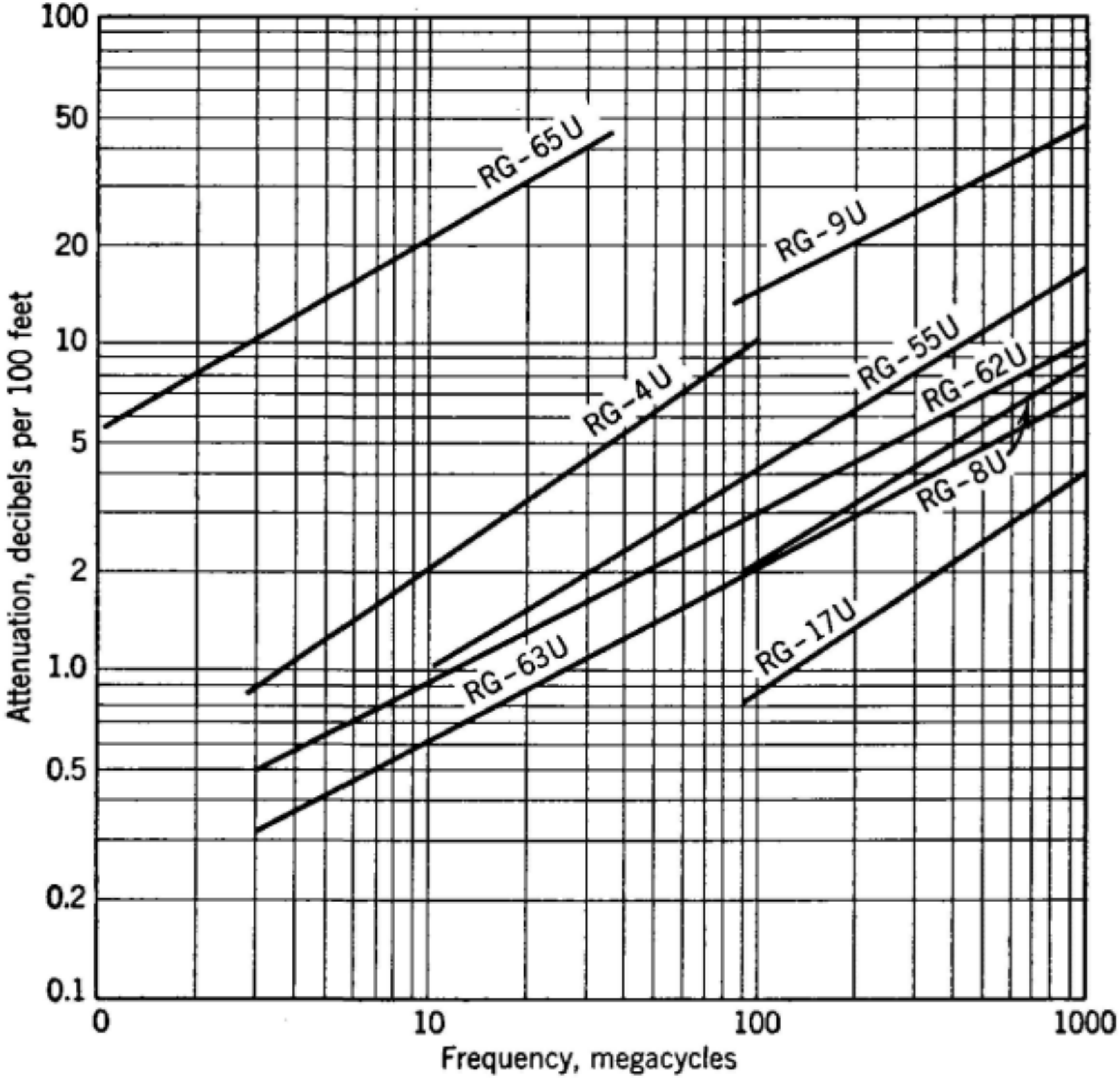
$$\alpha + jk = \sqrt{(R + j\omega L)(G + j\omega C)}$$

The introduction of R and G, leads to an exponential attenuation of the signal with distance at a rate given by α

The dependence of α and $v = \frac{dk}{d\omega}$ on ω. This implies a differential attenuation of the frequency components which leads to a dispersion of the pulse packet.

the high-frequency dependence of α can then be written as

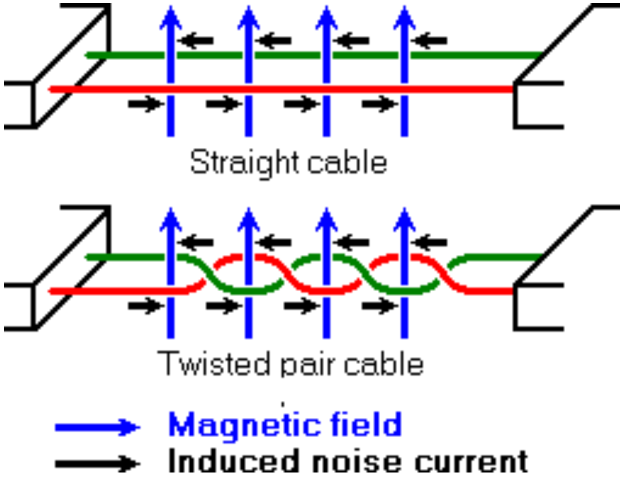
$\alpha(f) = a\sqrt{f} + bf,$ where $f = 2\pi\omega$ and *a* and *b* are constants.



100
50
20
10
5
2
1.0
0.5
0.2
0.1
Attenuation, decibels per 100 feet
0
10
100
1000
Frequency, megacycles
RG-65U
RG-9U
RG-4U
RG-55U
RG-62U
RG-8U
RG-63U
RG-17U

Other types of transmission line:

Twisted pair wire – often with $Z_0 \approx 120$ ohm



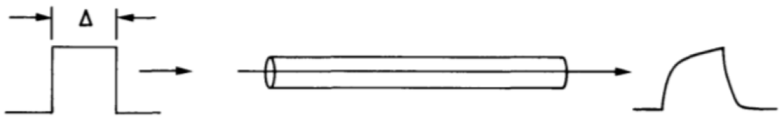
Pulse distortion

$$\tau_0 = \frac{(x\alpha)^2}{\pi f}$$

x: length of cable

a: attenuation constant;

f :frequency at which *a* is evaluated.



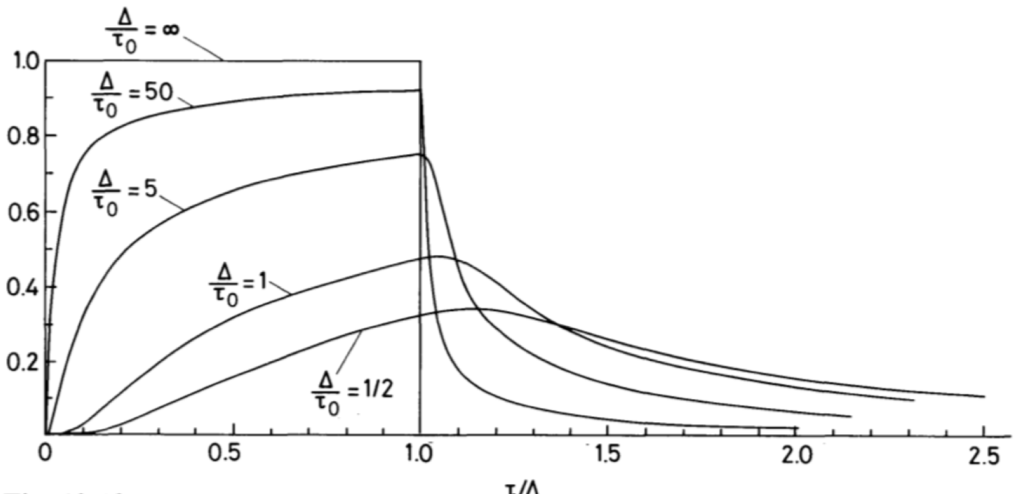


Fig. 13.10

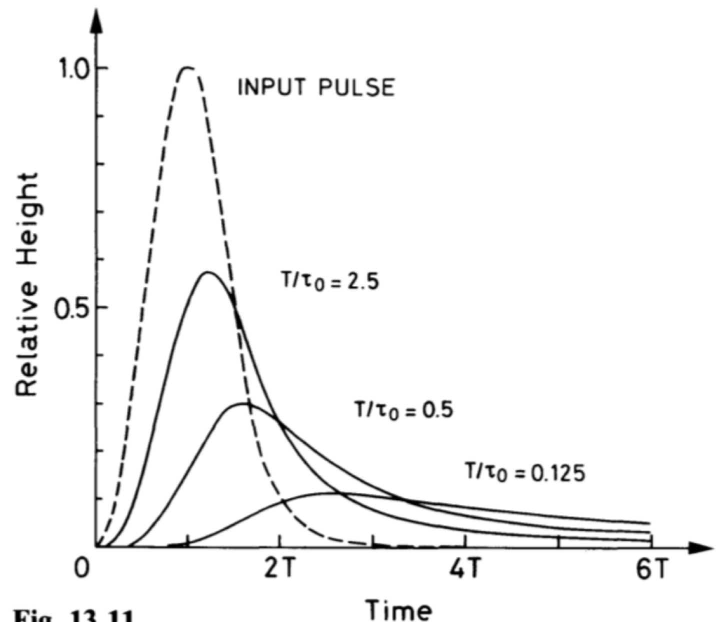


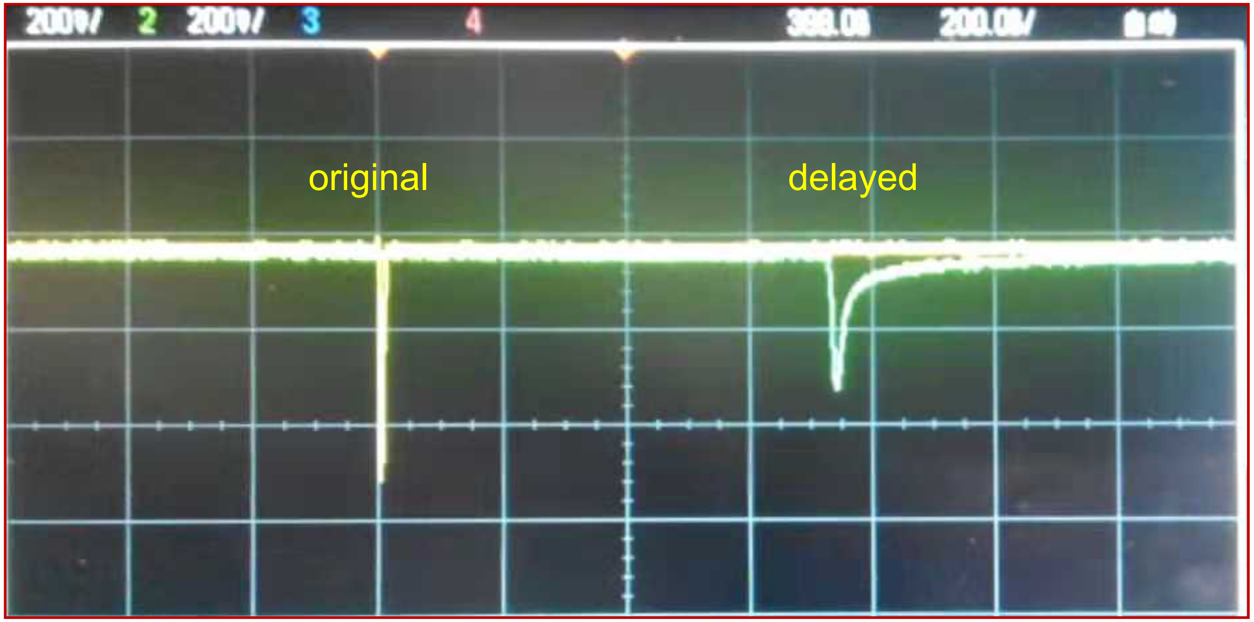
Fig. 13.11

Distortion of a rectangular pulse

Distortion of a photomultiplier

Signal losses in very long cables, critical for fast signal (high frequency)

Distortion of a fast signal after long cable delay (~780ns)

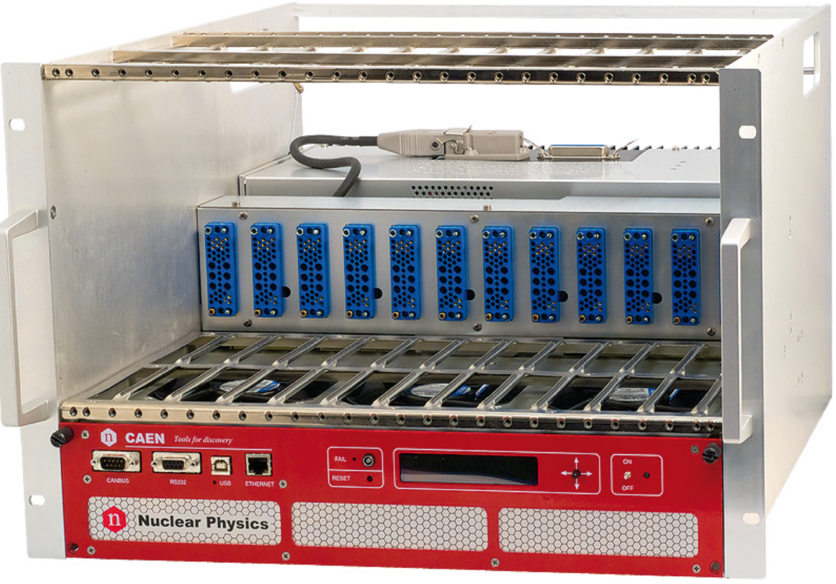


signal	dE/E(%)	
Original	2	Resolution is completely destroyed due to noise pickup through a long cable
Delayed	40	
Amplified(x10) and delayed	5	Improve the signal/noise

The NIM standard

In NIM(Nuclear Instrument Module) standard, the basic electonics apparatus are constructed in the form of modules according to standard mechanical and electrical specifications.

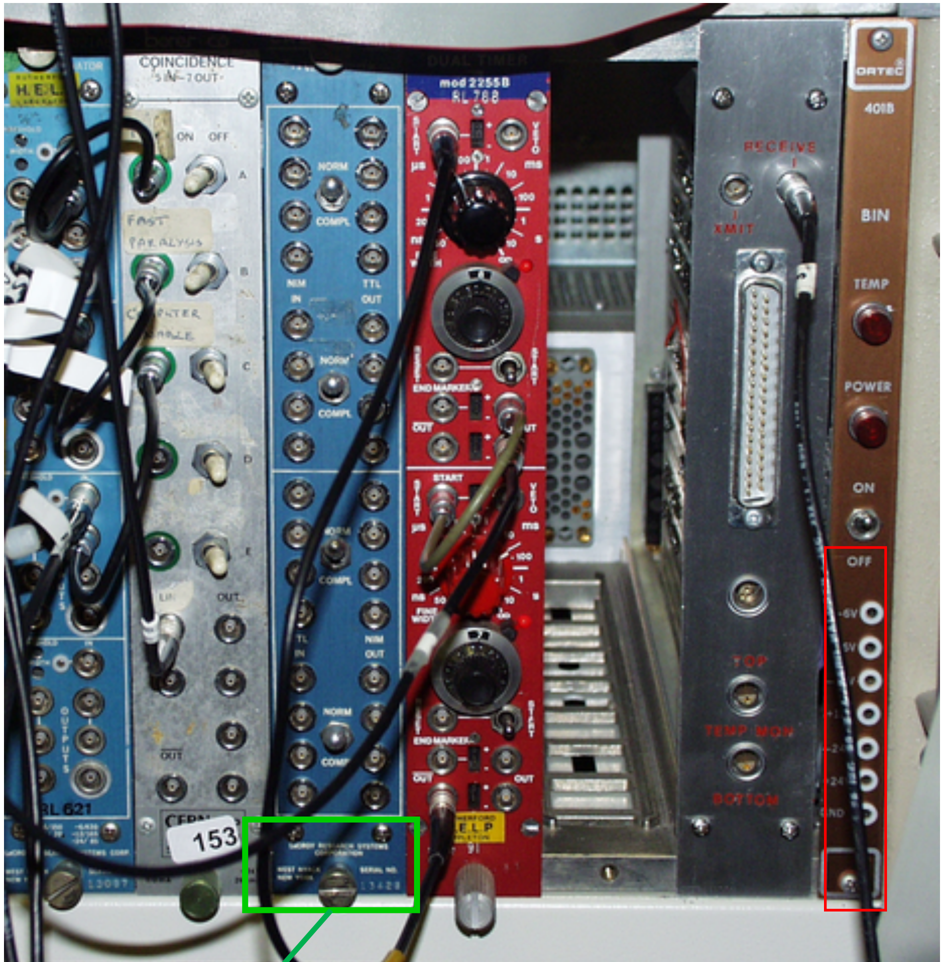
Mechanically, NIM modules must have a minimum standard width of 1.35 inches(3.43 cm) and a height of 8.75 in(22.225 cm). They can, however, also be built in multiples of this standard, that is, double-width, triple-width, etc.



NIM crate

The rear power connectors must provide, at the very least, four standard dc voltages, -12V, +12V, -24V and +24V, as designated by the NIM convention. However, many bins also provide -6V and +6V.

Power



NIM Crate Specification:

Output power

300W:
17A @ ±6V
3.4A @ ±12V
3.4A @ ±24V

600 W:
45A @ ±6V
8A @ ±12V
8A @ ±24V

20A @ ±6V
15A @ ±12V
8A @ ±24V

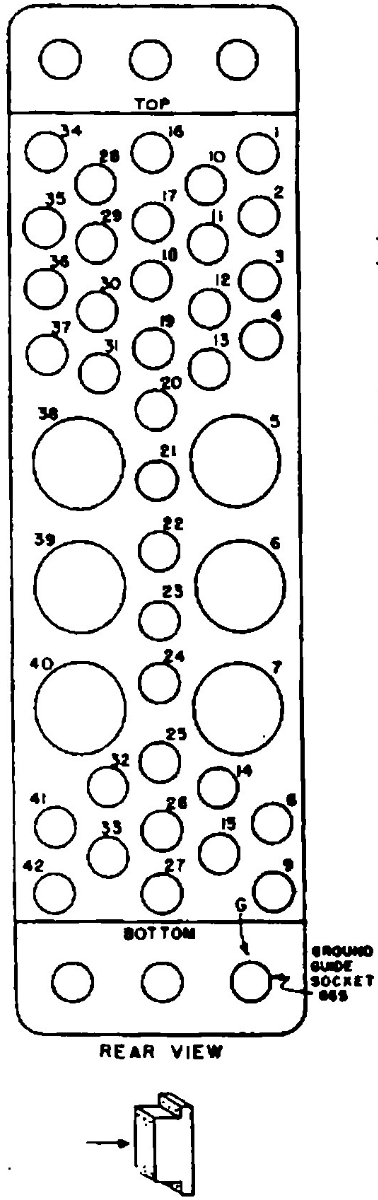
45A @ ±6V
18A @ ±12V
±24V not present

Voltage monitor: +/-6,+/-12V, +/-24V , GND

POWER REQUIRED +12 V, 85 mA; –12 V, 50 mA; +24 V, 100 mA; –24 V, 105 mA.

Voltage must be checked before use(No modules in)
during use(All modules in) to avoid over current

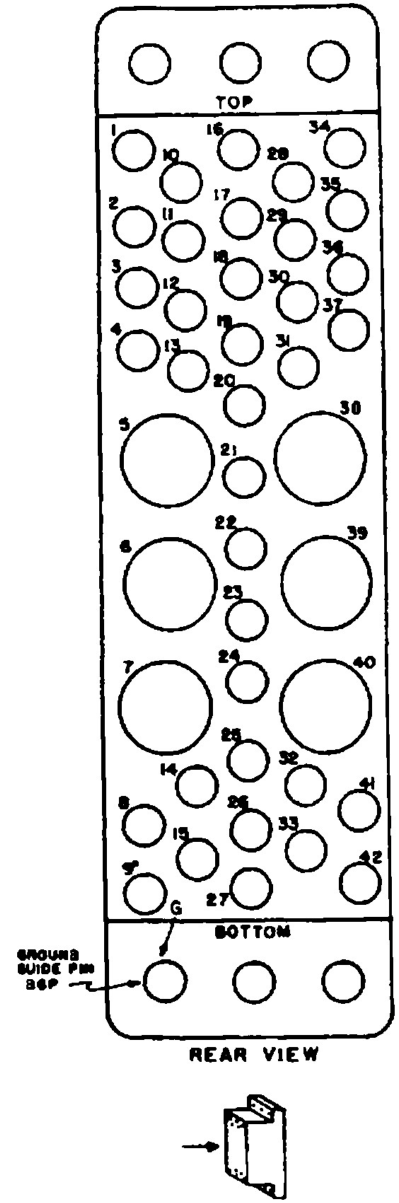
BIN CONNECTOR



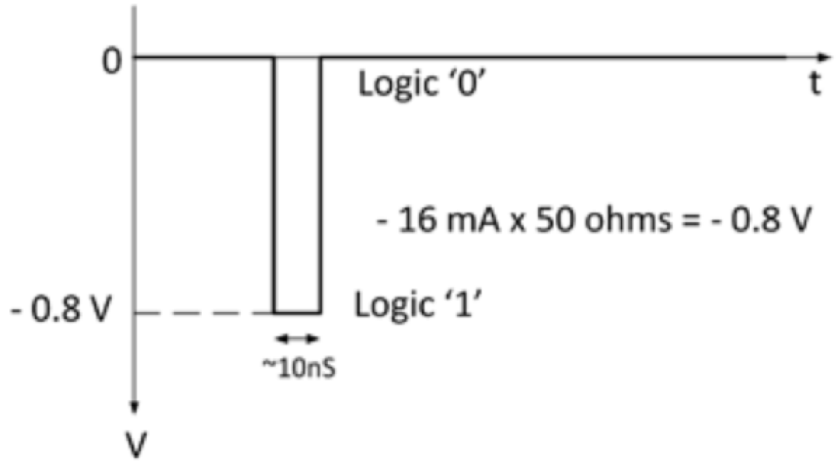
PIN	FUNCTION
1	+3 VOLTS
2	-3 VOLTS
3	SPARE
4	RESERVED
5	COAXIAL
6	COAXIAL
7	COAXIAL
8	+200 VOLTS D.C.
9	SPARE
♦ 10	+6 VOLTS
♦ 11	-6 VOLTS
12	RESERVED
13	CARRY NO. 1
14	SPARE
15	RESERVED
♦ 16	+12 VOLTS
♦ 17	-12 VOLTS
18	SPARE
19	RESERVED
20	SPARE
21	SPARE
22	RESERVED
23	RESERVED
24	RESERVED
25	RESERVED
26	SPARE
27	SPARE
♦ 28	+24 VOLTS
♦ 29	-24 VOLTS
30	SPARE
31	CARRY NO. 2.
32	SPARE
♦ 33	117 VOLTS A.C. (HOT)
♦ 34	POWER RETURN GND
35	RESET
36	GATE
37	SPARE
38	COAXIAL
39	COAXIAL
40	COAXIAL
♦ 41	117 VOLTS A.C.(NEUTRAL)
♦ 42	HIGH QUALITY GND
G	GROUND GUIDE PIN

♦ MUST BE BUSSED TO ALL BIN CONNECTORS PG10 THROUGH PG128

MODULE CONNECTOR



•**Fast negative logic pulses**, nS timescale, now commonly called NIM signals. These make up most of our Timing System signals.

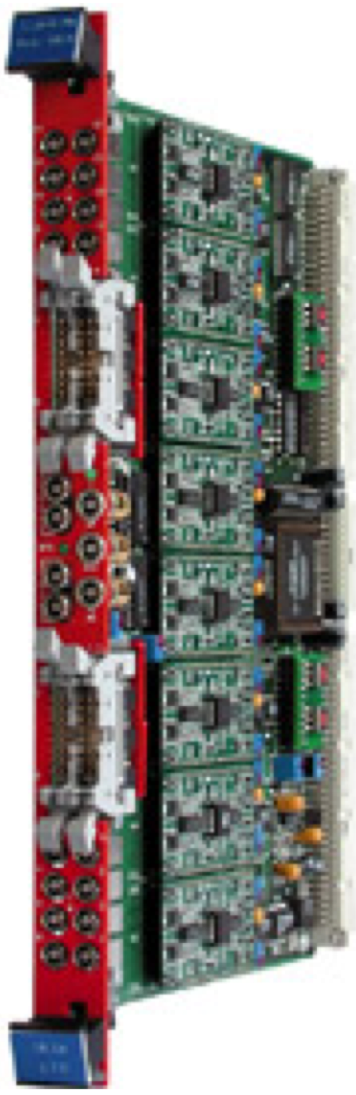


Typical Negative Logic NIM Signal as seen on a scope terminated with 50 ohms

Only for logic signals!

Table 12.3. TTL and ECL signal levels

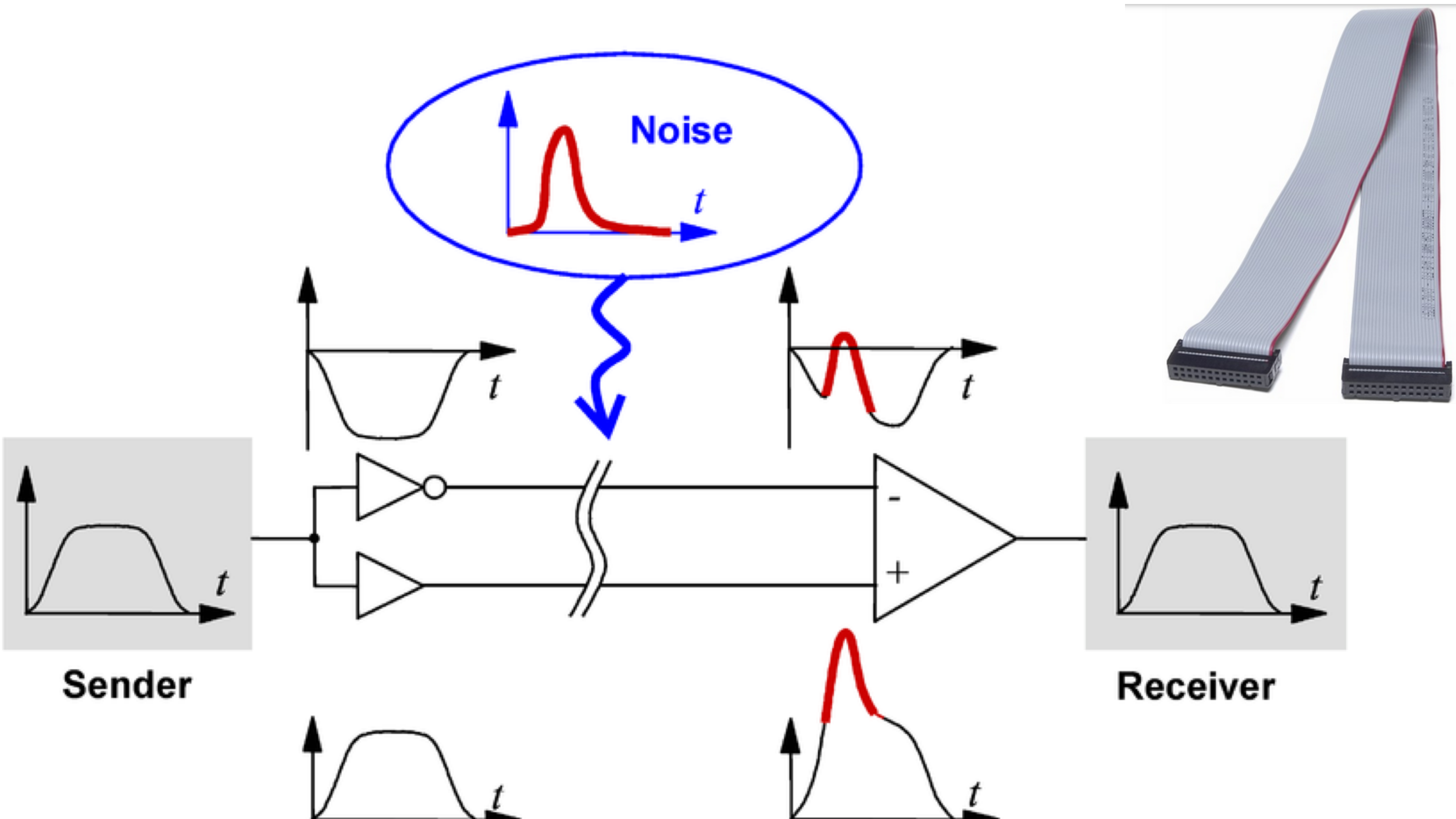
	TTL	ECL
Logic 1	2 – 5V	– 1.75 V
Logic 0	0 – 0.8 V	– 0.90 V



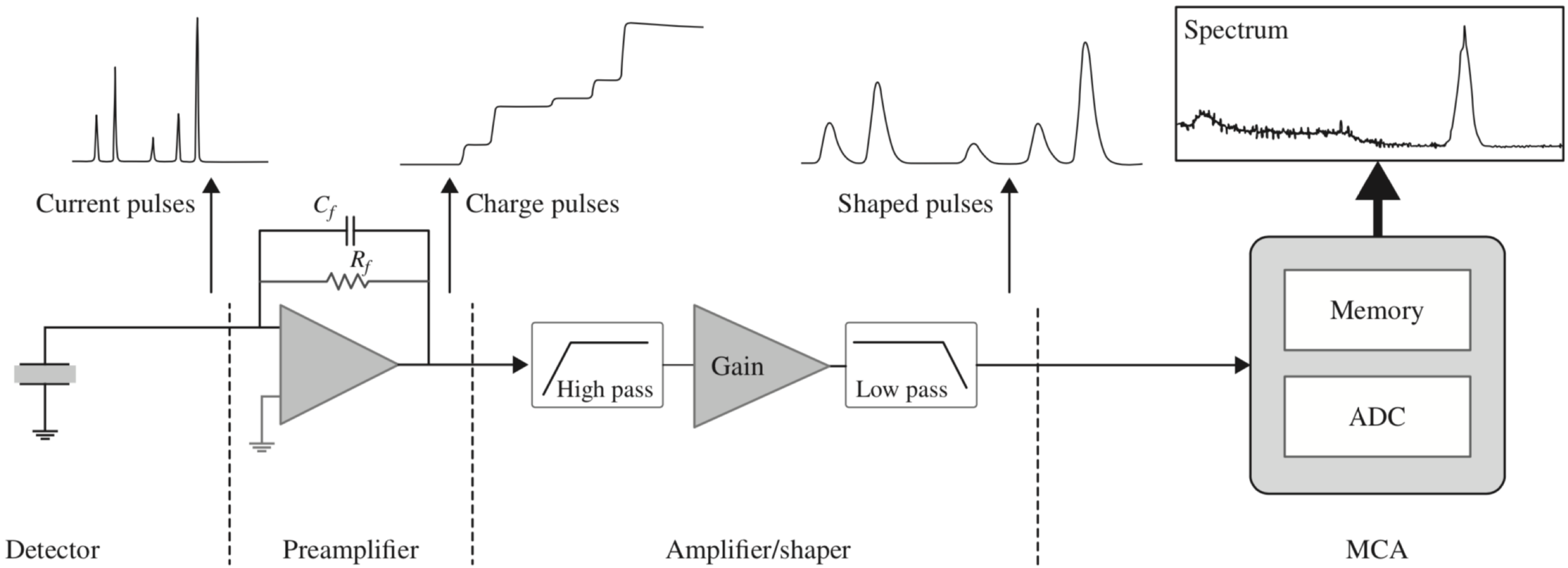
16 ch. discriminator
ECL output

Usually, differential receivers are used for ECL inputs to a module to take dvantage of the complementary outputs from the ECL output driver. This has the benefit of avoiding ground loops between modules, and minimizing common mode noise interference. ECL signals have rise times less than 2 ns.

Connected with a ribbon cable or a cable containing 100-Ω twisted pairs



Energy spectroscopy system



Electronics for pulse signal processing

Typical Signal Output and Pulse Duration of Various Radiation Detectors

Detector	Signal (V)	Pulse Duration (µsec)
Sodium iodide scintillator with photomultiplier tube	10^{-1}-1	0.23*
Lutetium oxyorthosilicate scintillator with photomultiplier tube	10^{-1}-1	0.04*
Liquid scintillator with photomultiplier tube	10^{-2}-10^{-1}	10^{-2}*
Lutetium oxyorthosilicate scintillator with avalanche photodiode	10^{-5}-10^{-4}	0.04*
Direct semiconductor detector	10^{-4}-10^{-3}	10^{-1}-1
Gas proportional counter	10^{-3}-10^{-2}	10^{-1}-1
Geiger-Müller counter	1-10	50-300

Most of them produce pulse signals of relatively small amplitude.

In addition, most of the detectors listed have relatively high output impedance, that is, a high internal resistance to the flow of electrical current.

Preamplifier

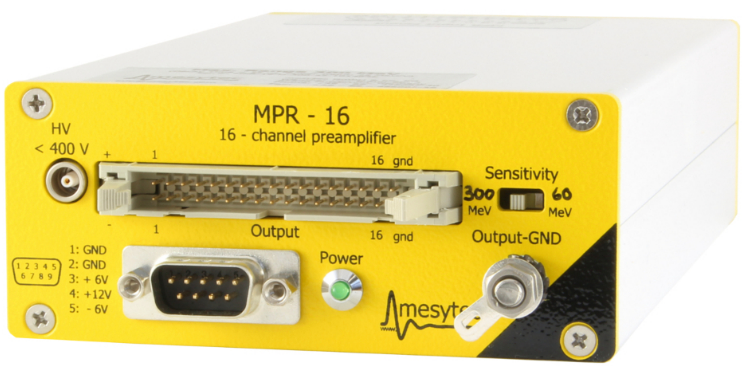
The purposes of a ***preamplifier*** (or preamp):

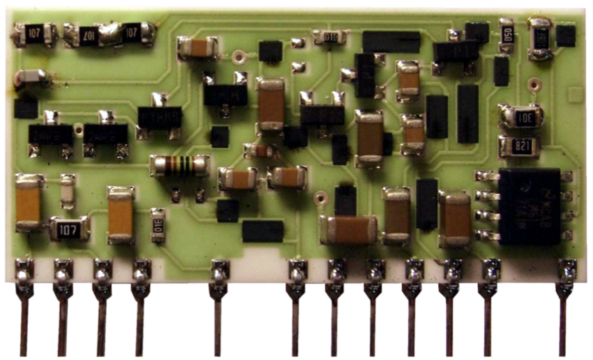
(1) To amplify, if necessary, the relatively small signals produced by the radiation detector for transmission over long distances

(2) To match impedance levels between the detector and subsequent components in the system.

(3) To shape the signal pulse for optimal signal processing by the subsequent components.

Since the input signal at the preamp. is generally weak, preamp, are normally mounted as close as possible to the detector so as to minimize the cable length.

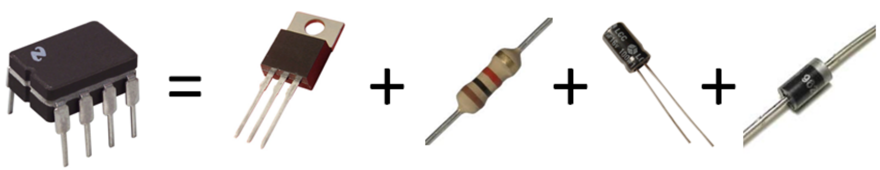




Operational Amplifier(Op-Amp)

Op-Amps are one of the basic building blocks of analog electronic circuits.

- An “Op-Amp” is an integrated circuit that sets an output voltage based on the input voltages provided.
- In a circuit, it is used to perform an *operation* and an *amplification* where the *operation* may be add, subtract, filter, integrate, differentiate, etc.
- Op-Amps are composed of transistors, resistors, capacitors, and diodes.



V_{S+} : positive power supply

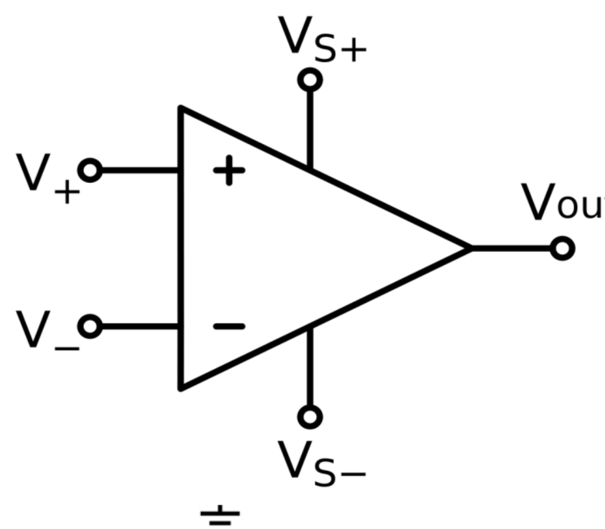
V_{S-} : negative power supply

V_{+}: non-inverting input terminal

V_{-}: inverting input terminal

V_{out} : output terminal

V_{+}, V_{-} , V_{out} are all referenced to ground



Parameter Name	Symbol	Value (ideal)	Value (real)
Input impedance	Z_{in}	∞	10^9~$10^{12}\Omega$
Output impedance	Z_{out}	0	$20\sim100\Omega$
Open-loop gain	A	∞	$10^4\sim10^7$
Bandwidth	B	∞	$10^6\sim10^9$Hz

- The maximum output voltage value is the supply voltage (saturation):
  $V_{S-} \leq V_{out} \leq V_{S+}$

$$V_{out} = G(V_+ - V_-) = G \cdot V_{in}$$

- Current flow into the op-amp from either input terminal is zero.
  $I_- = I_+ = 0$

- Differential voltage between the two input terminals is zero for negative feed back.
  $V_+ - V_- = 0$

- The maximum output voltage value is the supply voltage (saturation):
  $V_{S-} \leq V_{out} \leq V_{S+}$

Inverting Op-Amp

to amplify the input voltage to output voltage with a negative gain.

- $V_+ = 0\ V$
- $V_{in} = R_{in} \cdot I$
- $V_{out} = R_f \cdot (-I)$
- $\frac{V_{out}}{V_{in}} = \frac{-I \cdot R_f}{I \cdot R_{in}}$
- $V_{out} = -\frac{R_f}{R_{in}} \cdot V_{in}$

- Current flow into the op-amp from either input terminal is zero. I

  $I_- = I_+ = 0$

- Differential voltage between the two input terminals is zero for negative feed back.

  $V_+ - V_- = 0$

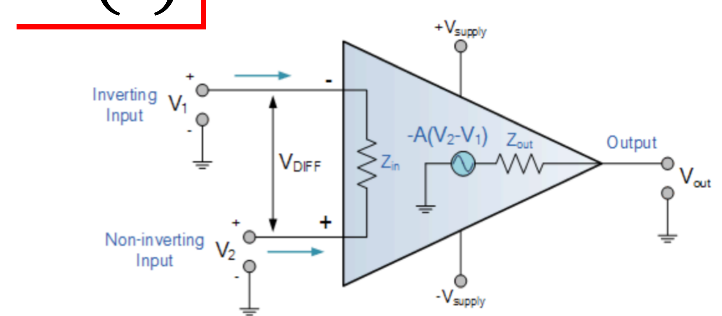
Integrating Op-Amp

takes the summation of input voltages over time and provides that as the output signal

- $V_+ = 0\ V$
- $I(t) = \frac{V_{in}(t)}{R}$
- $V_{out} = -\frac{1}{C} \cdot \int_0^t I(\tau) d\tau$
- $V_{out} = -\frac{1}{RC} \cdot \int_0^t V_{in}(\tau) d\tau$

Equivalent Circuit of an Ideal Operational Amplifier

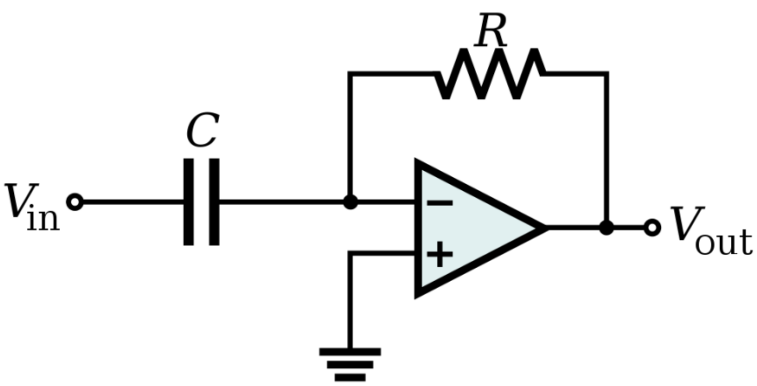
$I(t)$



Derivative Op-Amp

takes the rate of change of the inverted input voltage signal and provides that as the output signal

- $V_+ = 0\ V$
- $V_{in}(t) = \frac{1}{C} \cdot \int I(t)dt$
- $I(t) = C \cdot \frac{dV_{in}(t)}{dt}$
- $V_{out} = -R \cdot I(t)$
- $V_{out} = -RC \cdot \frac{dV_{in}(t)}{dt}$

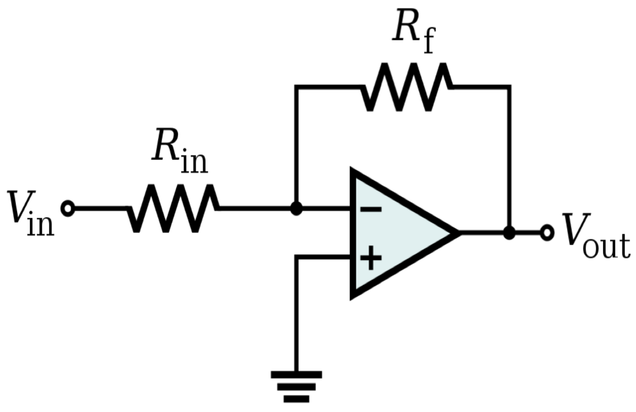


Differential Op-Amp

takes the difference between two signals and provides that as the output

$$V_{out} = \left(\frac{R_1 + R_f}{R_1}\right)\left(\frac{R_g}{R_g + R_2}\right)V_2 - \frac{R_f}{R_1}V_1$$

- If $\frac{R_f}{R_1} = \frac{R_g}{R_2}$: $\quad V_{out} = \frac{R_f}{R_1}(V_2 - V_1)$
- Moreover, if $R_f = R_1$: $\quad V_{out} = V_2 - V_1$

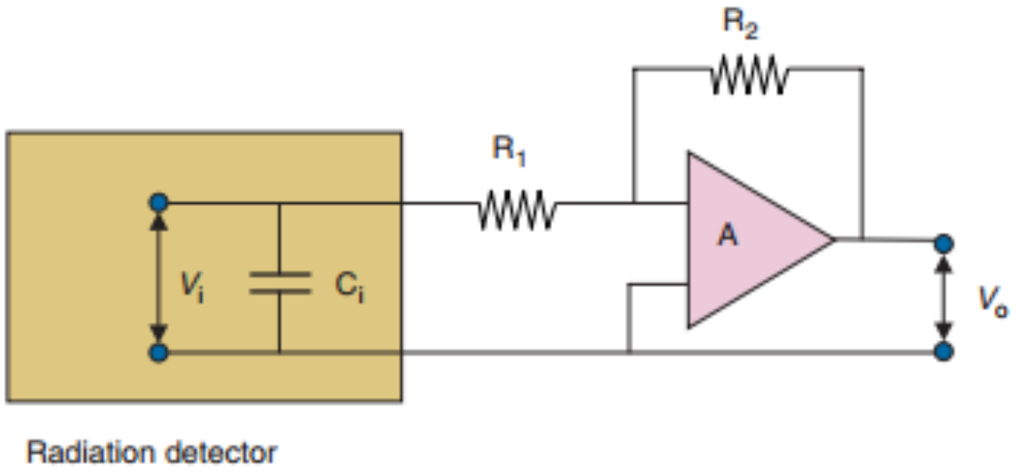


Three basic types of preamplifier exist:

1) voltage sensitive (Conventional)
2) charge sensitive (for semiconductor, most common)
3) current sensitive

Voltage sensitive preamplifier

The voltage-sensitive preamp amplifies any voltage that appears at its input.



Intrinsic capacitance of the detector and other components in the input circuit, c_i; $V_i = Q/C_i$

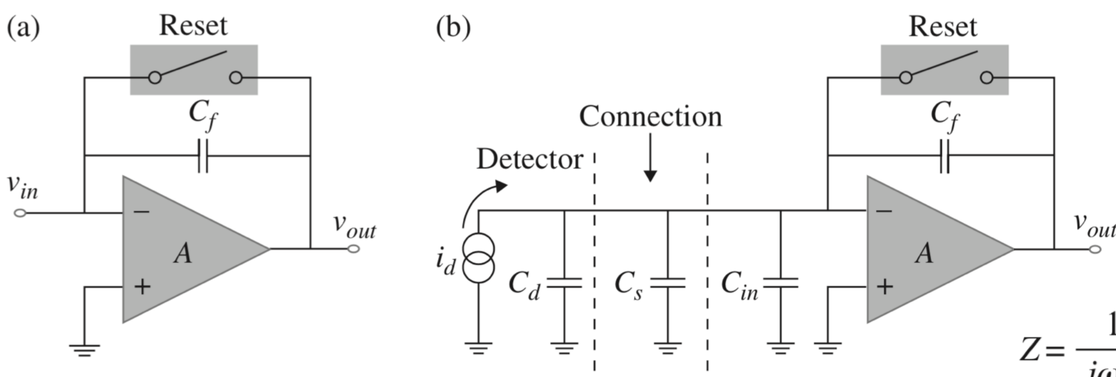
$$V_0 = -\frac{R_2}{R_1} V_i = -\frac{R_2}{C_i R_1} Q$$

The magnitude of the signal is dependent on the sensor capacitance.

Such a preamplifier design works fine for detectors whose capacitance does not changes, such as PMT's, proportional counters and Geiger-Muller tubes.

Charge-Sensitive Preamplifiers

In a system with varying sensor capacitances, a partially depleted semiconductor sensor, where the capacitance varies with the applied bias voltage, or intrinsic capacitance changing with temperature, one would have to deal with additional calibrations. This can be achieved rather simply with a charge-sensitive amplifier.

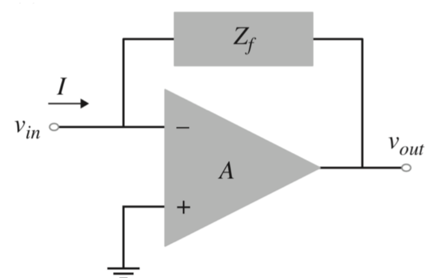


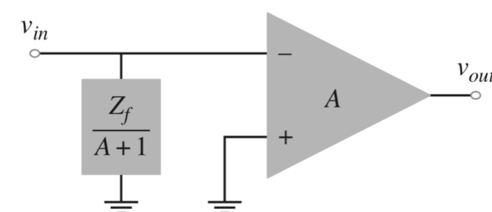
$$C = C_d + C_{in} + C_s$$

C_{in}: input capacitance of the op-amp

C_s: stray capacitances

$$Z = \frac{1}{j\omega C}$$





$$v_{out} = -Av_{in}$$

$$I = \frac{v_{in} - v_{out}}{z_f} = \frac{(A+1)v_{in}}{z_f} = \frac{v_{in}}{z_{eff}}$$

$C_{eff} = (1 + A)C_f$ effective input capacitance

all the capacitances appearing at the input of the preamplifier,

$$C_{tot} = C + (1 + A)C_f$$

the voltage at the input of the preamplifier can be written as

$$v_{in} = \frac{Q_\circ}{C_{tot}} = \frac{Q_\circ}{C + (1 + A)C_f}$$

$$v_{out} = -\frac{AQ_\circ}{C + (A + 1)C_f} = -\frac{Q_\circ}{\frac{A+1}{A}C_f + \frac{C}{A}}$$

$$\boxed{v_{out} \approx -\frac{Q_\circ}{C_f}} \qquad (A \gg 1)$$

Sensitivity: the output of the energy deposited in the detector in mV/MeV.

$$Q_\circ = \frac{Ee(10^6)}{w} \qquad v_{out} = \frac{E(10^6)(1.6 \times 10^{-19})}{wC_f}$$

$$\frac{V_{out}}{E} = \frac{160}{wC_f}$$ [mV/MeV] , as ω in eV , C_f in pF

88 mV/MeV: C_f = 0.5 pF for a silicon detector (w = 3.62)

12 mV/MeV: C_f = 0.5 pF for an argon-filled ionization chamber (w = 26.5 eV)

in practice, the charge produced in the detector (Q) is shared between the C_f and sum of other capacitances at the input of the preamplifier (C), and therefore, some charge is always lost without contribution to the output pulse.

$$\frac{Q_f}{Q_\circ} = \frac{Q_f}{Q_l + Q_f} = \frac{(A+1)C_f}{C + (A+1)C_f} = \frac{1}{1 + \dfrac{C}{(A+1)C_f}}$$

to achieve high charge transfer efficiency, a low capacitance detector is desirable, and the stray capacitance between the detector and preamplifier should be minimized.

C_s : ~ 95pF/mm for lemo-cable

This can be achieved by placing the preamplifier as close as possible to the detector and by using proper wiring.

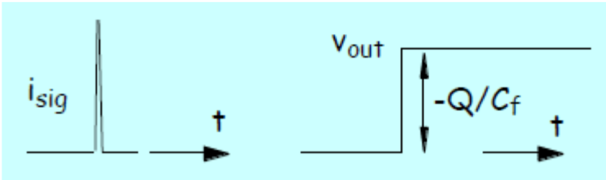
ORTEC 142A, B, and C Preamplifiers

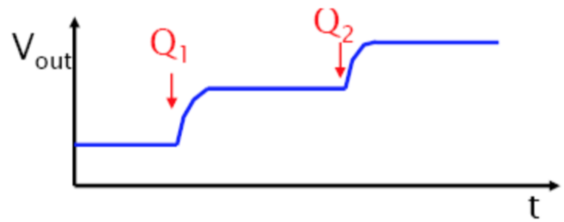
OPEN LOOP GAIN
142A >40,000.
142B >80,000.
142C >80,000.

CHARGE SENSITIVITY (Si equivalent)
142A Nominally 20 mV/MeV.
142B Nominally 10 mV/MeV.
142C Nominally 10 mV/MeV.

ENERGY RANGE
142A 0–200 MeV.
142B 0–400 MeV.
142C 0–400 MeV.

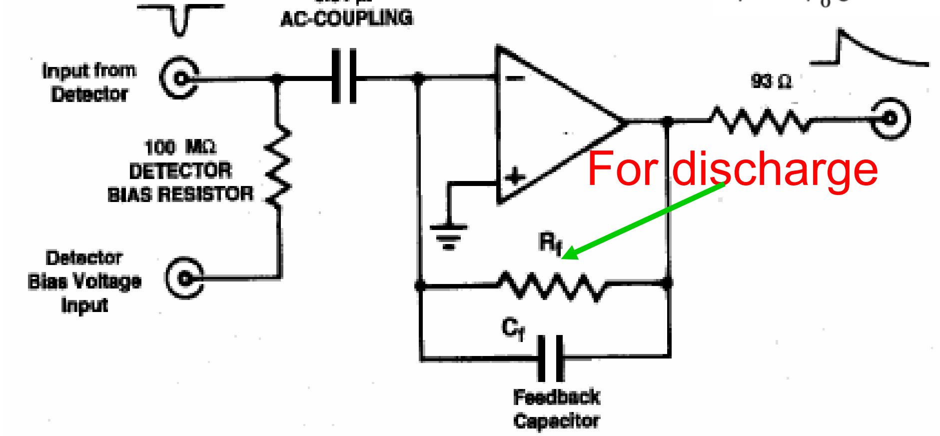
Without the reset circuit

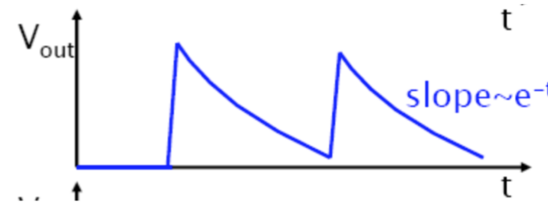




Resistive Feedback Preamplifiers

$$V = V_o e^{-t/R_f C_f}$$

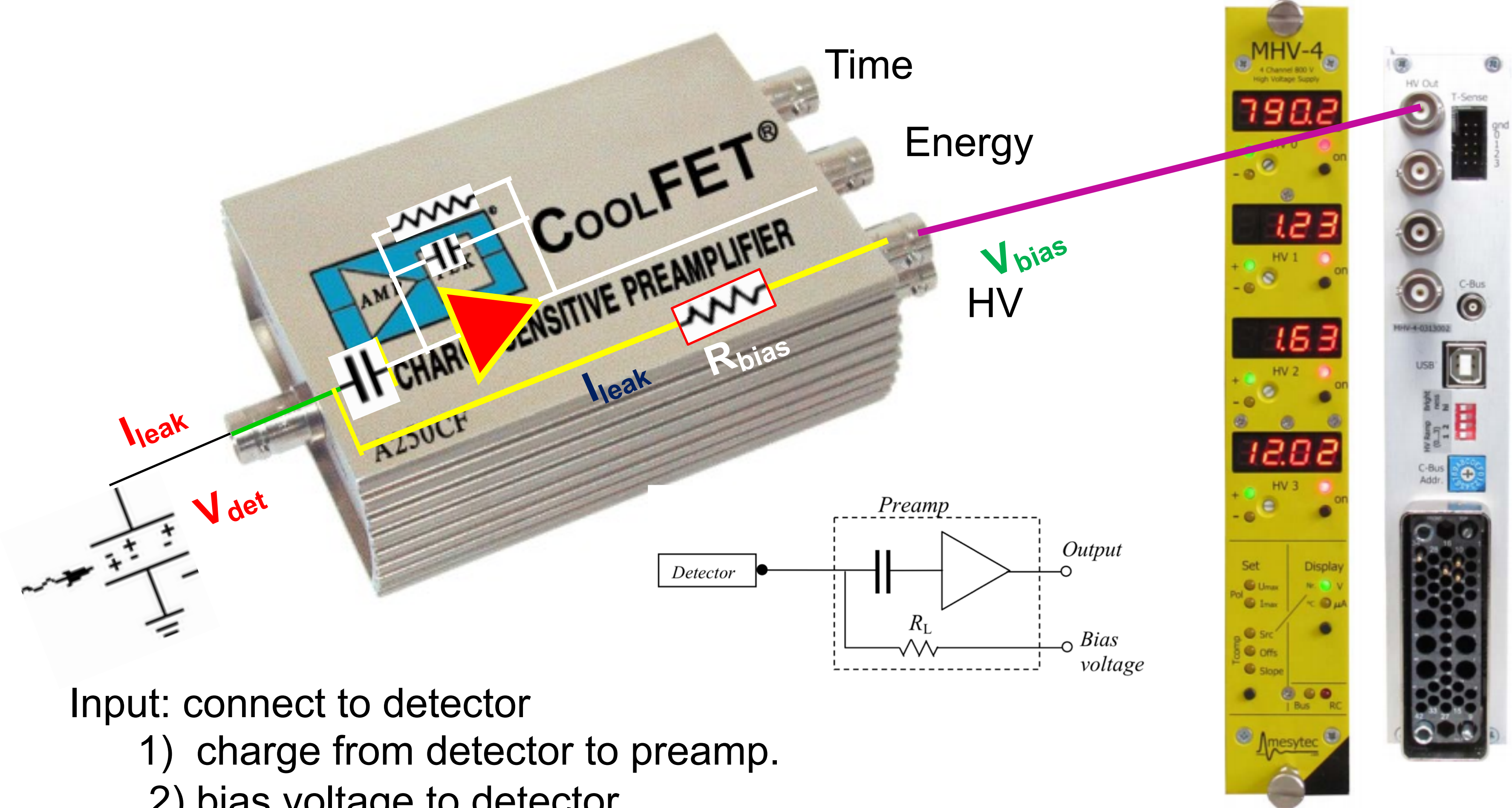




Typical value:
C_f : few pF
R_f: 10M-100MΩ
RC: 20-200μs

R_f is a noise source and in direct-coupled system, is made as large as possible consistent with the signal energy-rate product and the detector leakage current.

Although originally developed for use with semiconductor diode detectors, this charge-sensitive configuration has proved its superiority in a number of other applications, so that preamp. used with other detectors in which the capacitance does not necessarily change are also often of the charge-sensitive design.



Input: connect to detector

1) charge from detector to preamp.
2) bias voltage to detector

$V_{det}=V_{bias}-I_{leak} \times R_{bais}$

Be aware: $V_{det} \neq V_{bias}$

Typically,

V_{det}= 100V

R_{bias}= 100Mohm

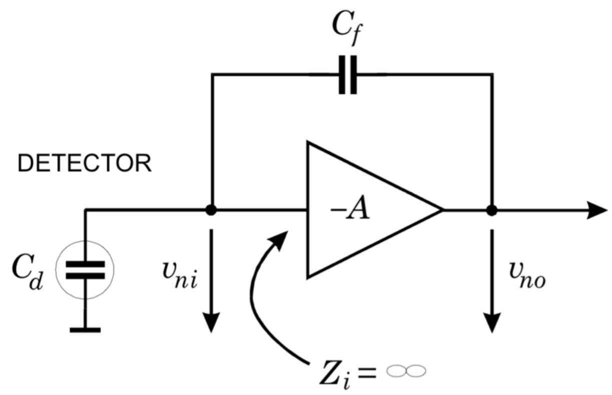
I_{leak}= 0.3μA

Voltage drop to the bias resistance = 30V

V_{bias} > 130V !

Signal-to-Noise Ratio vs. Detector Capacitance

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



$$v_{no} = v_{ni} \frac{Z_{C_f} + Z_{C_D}}{Z_{C_D}} = v_{ni} \frac{\frac{1}{\omega C_f} + \frac{1}{\omega C_D}}{\frac{1}{\omega C_D}}$$

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

Equivalent input noise charge

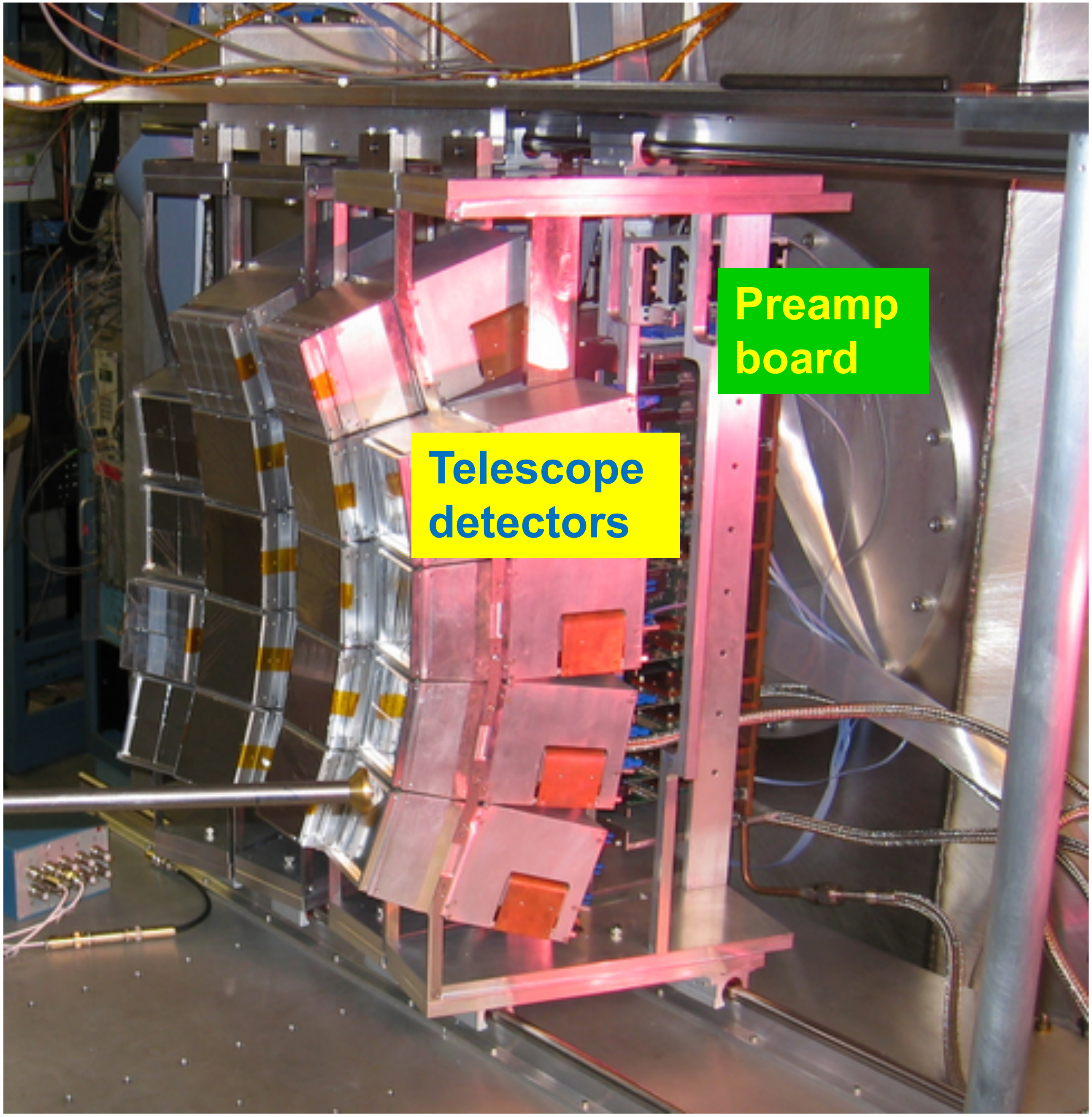
$$Q_{ni} = \frac{v_{no}}{A_Q} = v_{no} C_f = v_{ni}\left(C_D + C_f\right)$$

Signal-to-noise ratio

$$\frac{Q_s}{Q_{ni}} = \frac{Q_s}{v_{ni}(C_D + C_f)} = \frac{1}{C} \frac{Q_s}{v_{ni}}$$

the signal-to-noise ratio for a given signal charge is inversely proportional to the **total capacitance at the input node**.

The preamplifier package is kept small to permit mounting it as close as practical to the detector, thus reducing input capacitance caused by cabling and decreasing micro phonic noise, ground loops, and radio frequency pickup, all of which are sources of noise for the charge-sensitive preamplifier.


Preamp
board
Telescope
detectors

Current sensitive preamplifier:

Several detector types, such as photomultiplier tubes and microchannel plates, generate a moderately large and fast-rising output signal through a high output impedance. Pulse processing for timing or counting with these detectors can be rather simple. A properly terminated 50-Ω coaxial cable is attached to the detector output, so that the current pulse from the detector develops the desired voltage pulse across the 50-Ω load presented by the cable.

the amplitude of the voltage pulse at the preamplifier output will be $V_{out} = 50\, I_{in}\, A$

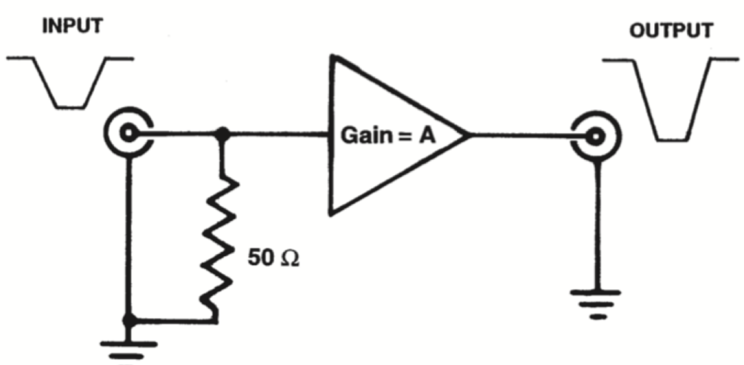
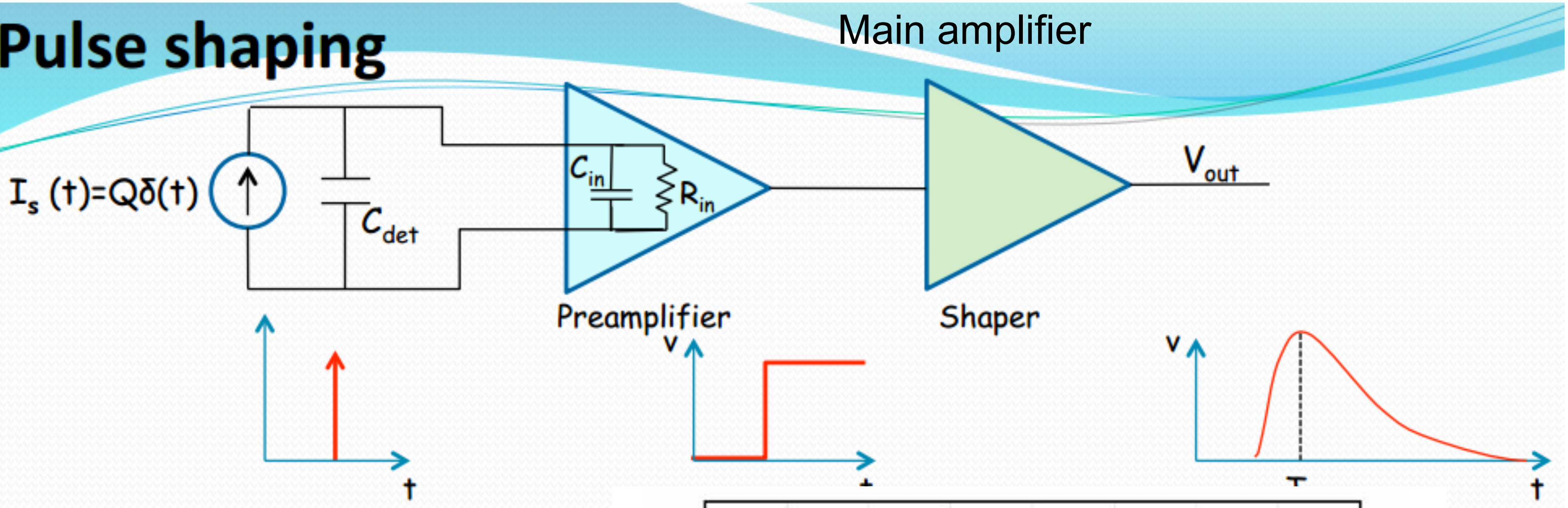


Fig. 1. A Simplified Schematic of the Current-Sensitive Preamplifier.

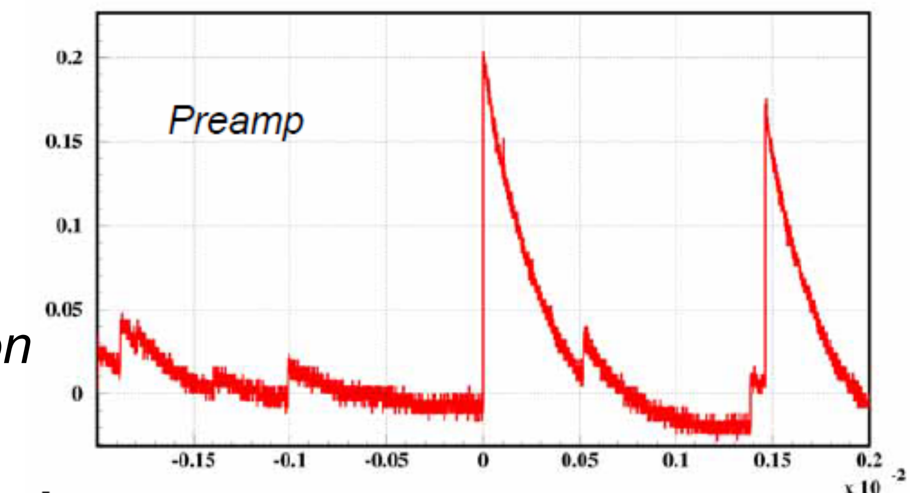
Pulse shaping

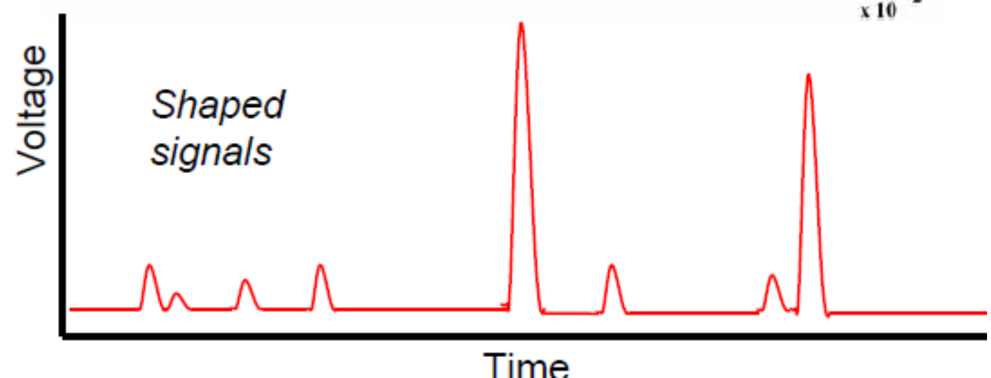


Spectroscopy amp.
Shape pulses to:

- *Keep signal height information*
- *Improve signal to noise*
- *Reduce pileup effects*

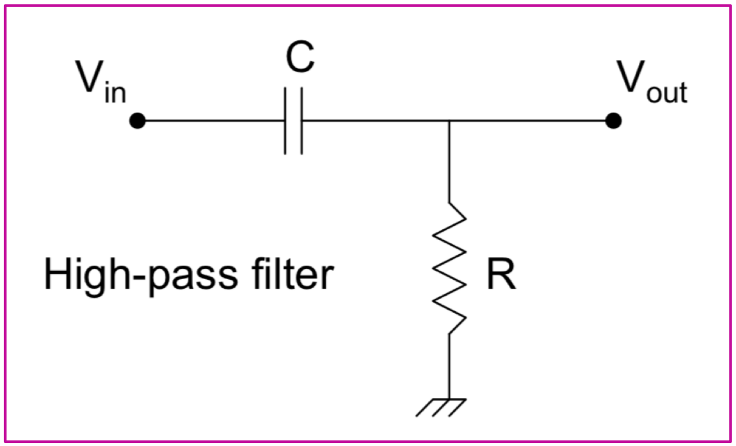
Shortcoming:
Lose fast time information
(eg rise time)

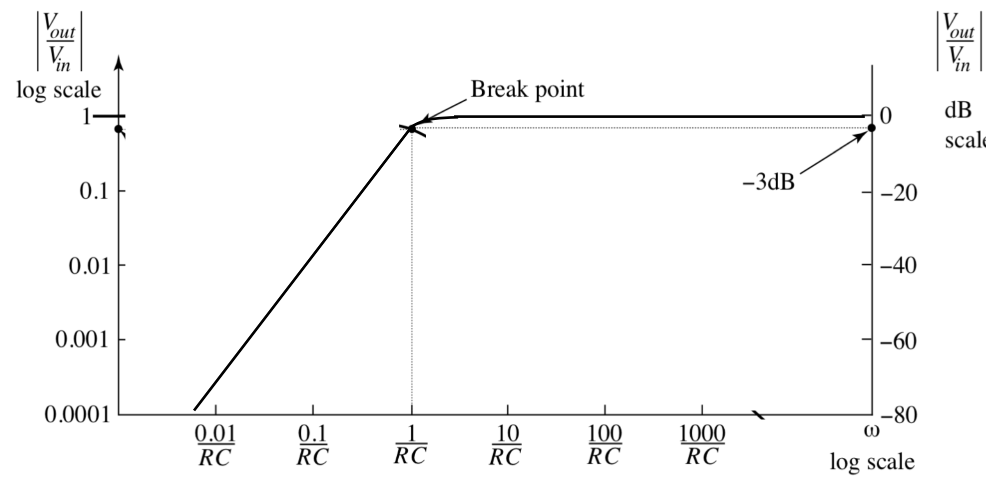




High-Pass Filter or CR Differentiating Circuit

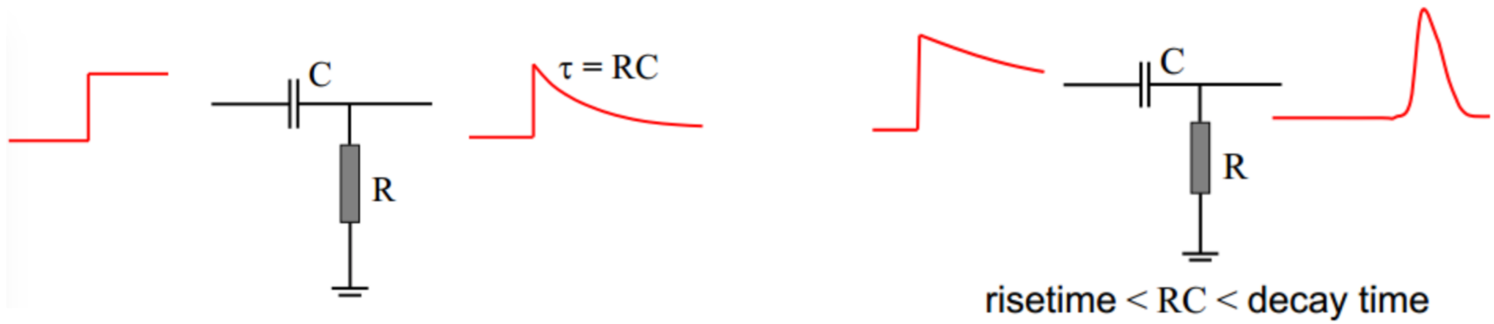
The CR high-pass filter, consisting of a single capacitor and resistor, acts as a low frequency filter, attenuating frequency below $f \leq \dfrac{1}{2\pi RC}$



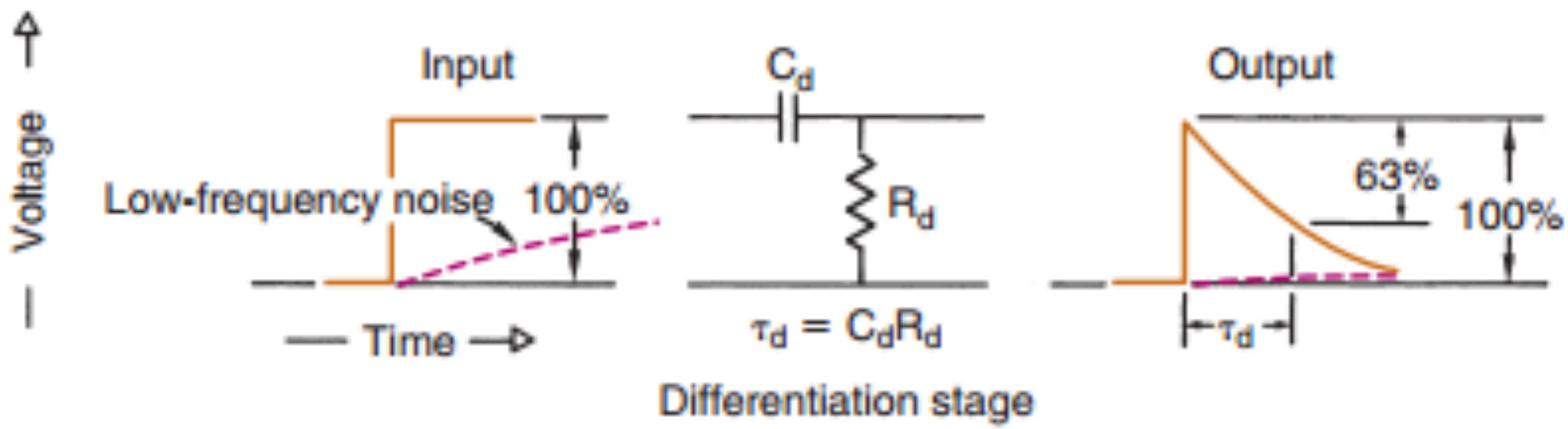


$$\frac{V_{out}}{V_{in}} = \frac{R}{R + 1/j\omega C}$$

$$\left|\frac{V_{out}}{V_{in}}\right| = \frac{\omega RC}{\sqrt{1 + (\omega RC)^2}}$$

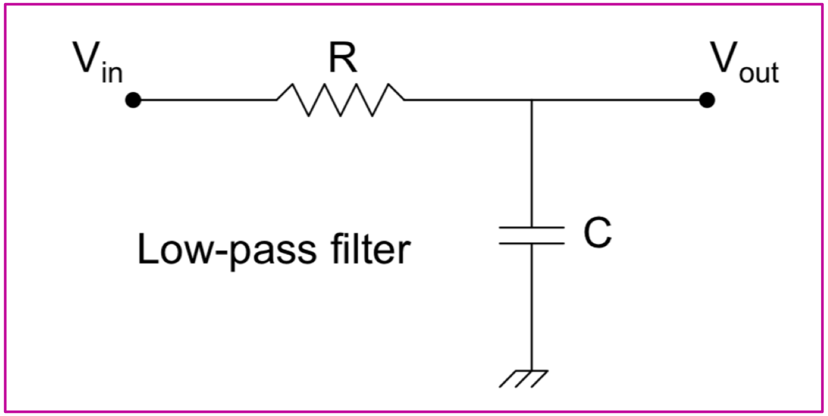


Its effect on a step function pulse. The flat part of the pulse is degraded and made to decay to the baseline, thereby shortening the pulse. In contrast, the fast rising part of the pulse, which depends on the higher frequencies, is not effected.



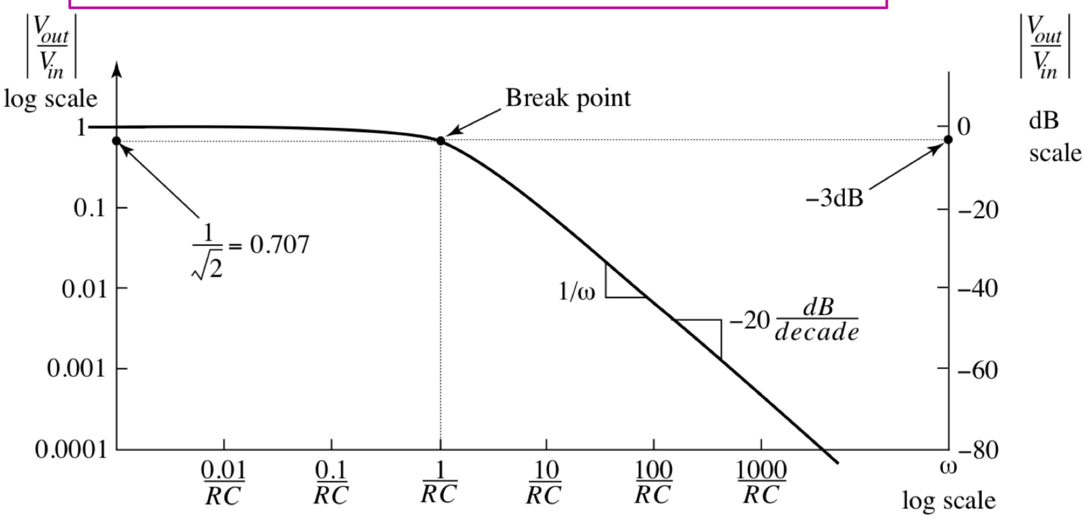
discriminates against low-frequency noise.

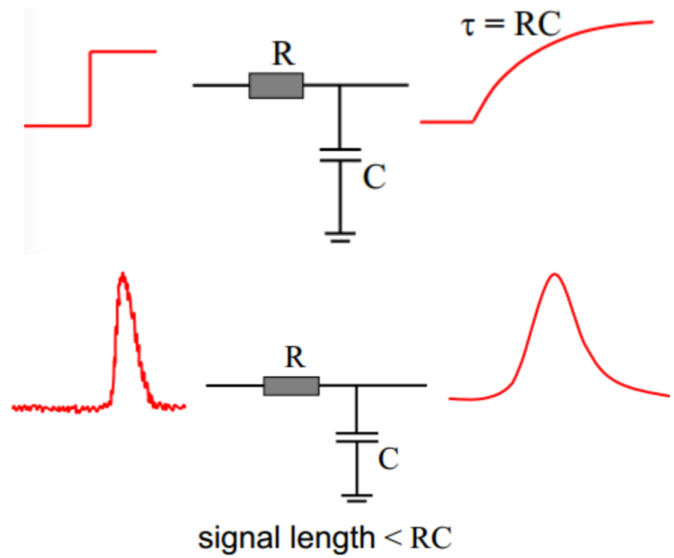
Low-Pass Filter or Integrating Circuit



$$\frac{V_{out}}{V_{in}} = \frac{1/j\omega C}{R + 1/j\omega C} = \frac{1}{1 + j\omega RC}$$

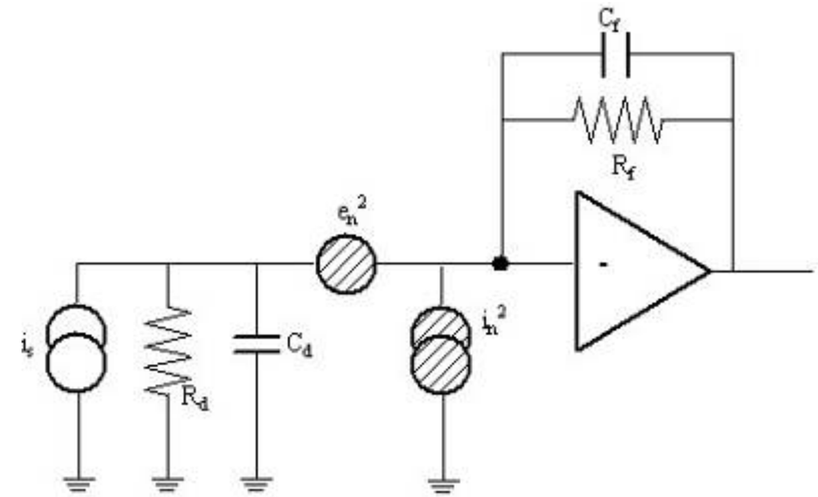
$$\left|\frac{V_{out}}{V_{in}}\right| = \frac{1}{\sqrt{1 + (\omega RC)^2}}$$



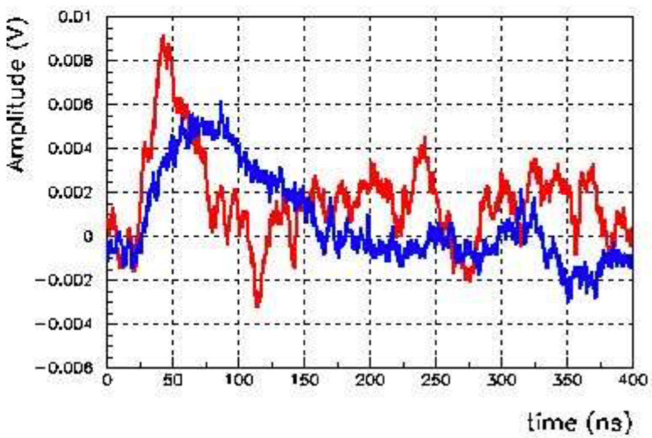


The principle application of the integrator is to smooth out fluctuations in a noisy signal.

Noise in charge sensitive preamplifiers



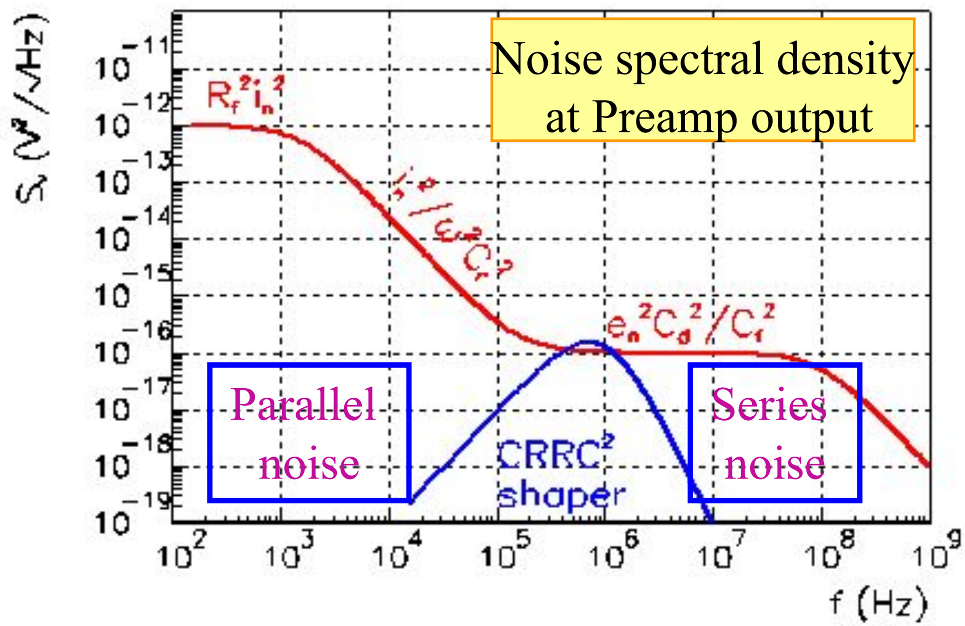
- 2 noise generators at the input
  - Parallel noise : ($\frac{dV_n^2}{df} \equiv i_n^2$) (leakage currents)
  - Series noise : ($\frac{dI_n^2}{df} \equiv e_n^2$) (preamp)

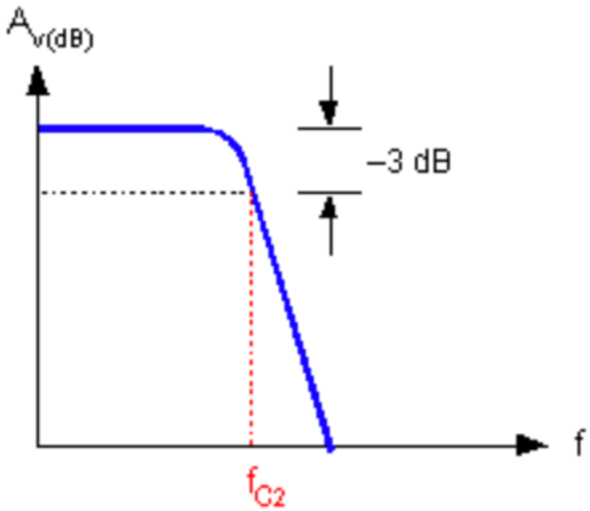


- Output noise spectral density :

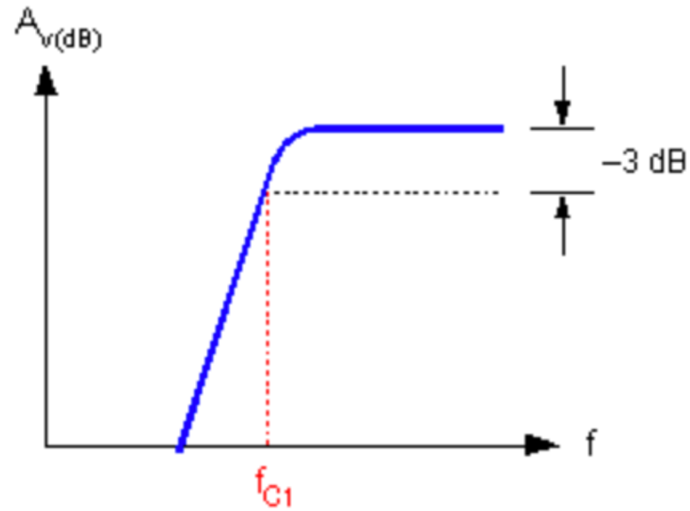
$$S_v(\omega) = \frac{i_n^2 + \frac{e_n^2}{|Z_d|^2}}{\omega^2 C_f^2} = \frac{i_n^2}{\omega^2 C_f^2} + \frac{e_n^2 C_d^2}{C_f^2}$$

  - Parallel noise in $1/\omega^2$
  - Series noise is flat, with a « noise gain » of C_d/C_f

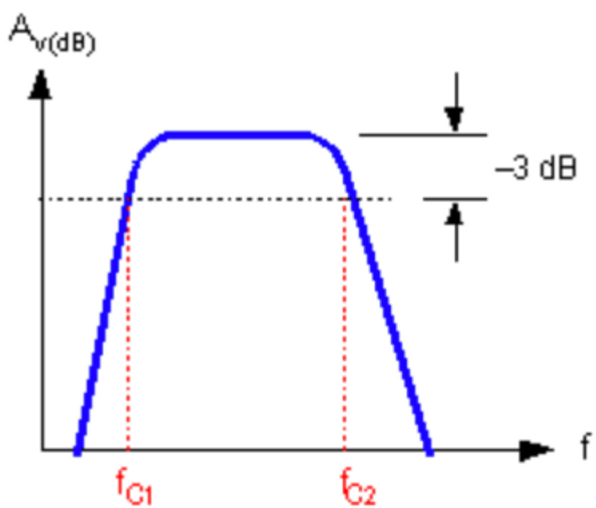




Low-pass filter



High-pass filter

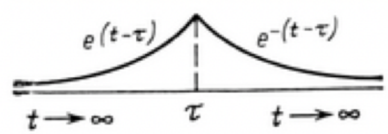
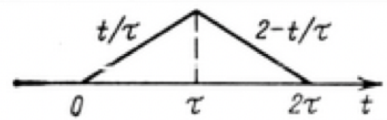
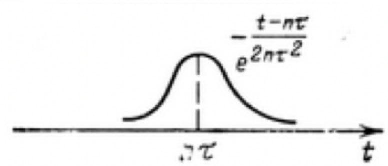
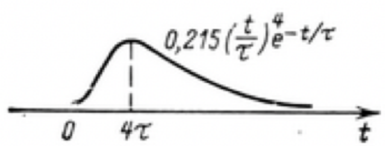
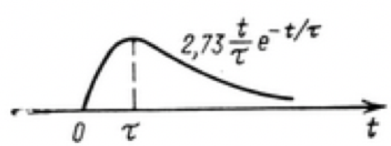


Band-pass filter

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

Relative noise contribution of different pulse shapes

For a given noise spectrum, there usually exists an optimum pulse shape in which the signal is least disturbed by noise.

	Pulse shape	Relative 'noise'
Cusp		1,00
Triangle		1,08
Gaussian (CR+RC^n)		1,12
Semi-Gaussian (CR+RC^5)		1,16
Unipolar CR-RC		1,36

Simple shaper: CR-RC

The simplest Pseudo-Gaussian filter is the CR-RC shaper because :

1. The high-pass filter is made with CR network
2. The low-pass filter is made with RC network

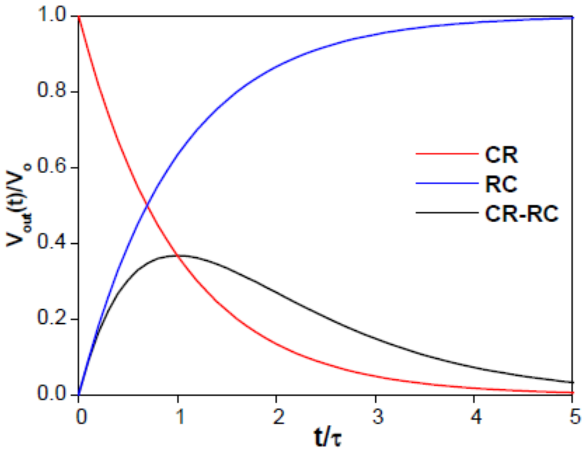
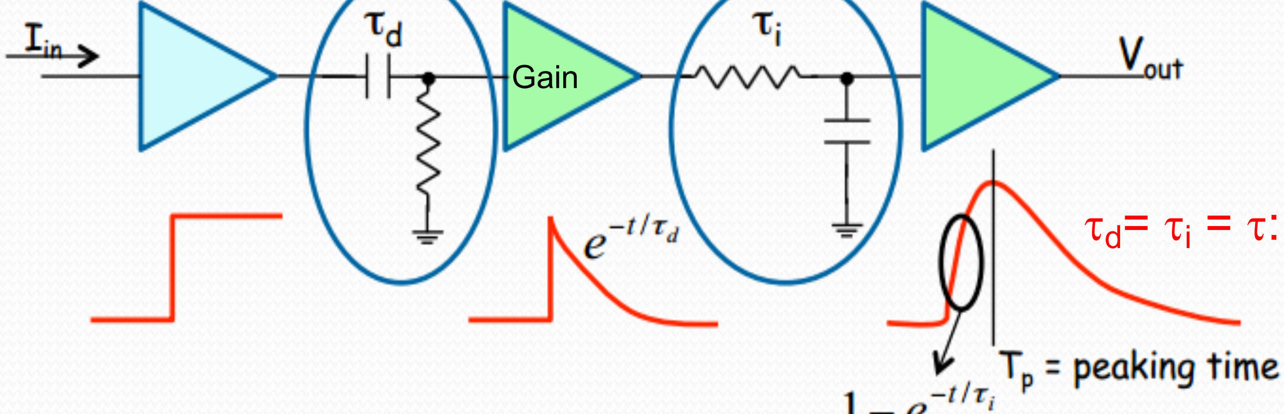


Fig. 6.9. CR, RC and CR-RC time domain responses for a step function input.



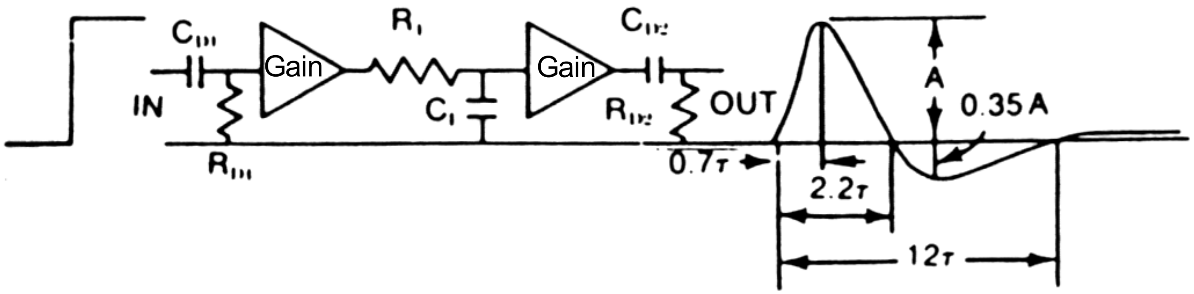
$\tau_d = \tau_i = \tau$: shaping time

- This shaper is called CR-RC because the high-pass filter is made with CR network, while the low-pass filter with a RC network

- The noise is 36% worse than "optimum filter" with the same time constants

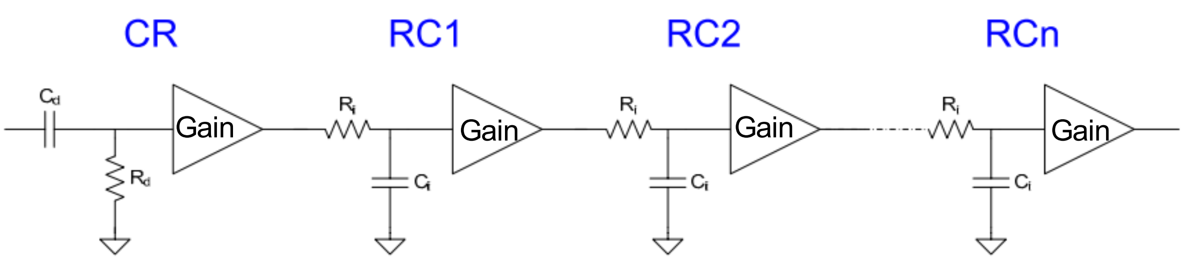
it has been shown that the best signal- to-noise ratio is achieved when the *CR* and *RC* time constants are equal

Doubly-Differentiated CR-RC-CR Shaping



$$\tau = R_{D1}C_{D1} = R_I C_I = R_{D2}C_{D2}$$

The shapers are often more complicated, with <u>multiple (n) integrators</u> → CR-RC^n



Same peaking time if $\tau_n = \tau_{(n=1)}/n$

With same peaking time

1. More symmetrical
2. Faster return to baseline
3. Improved rate capability

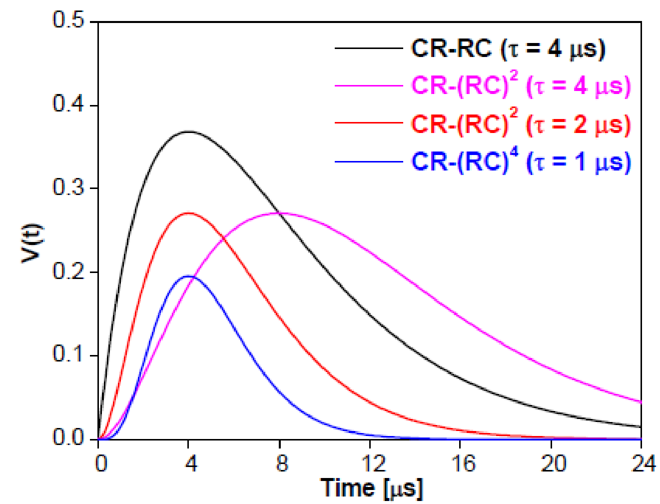


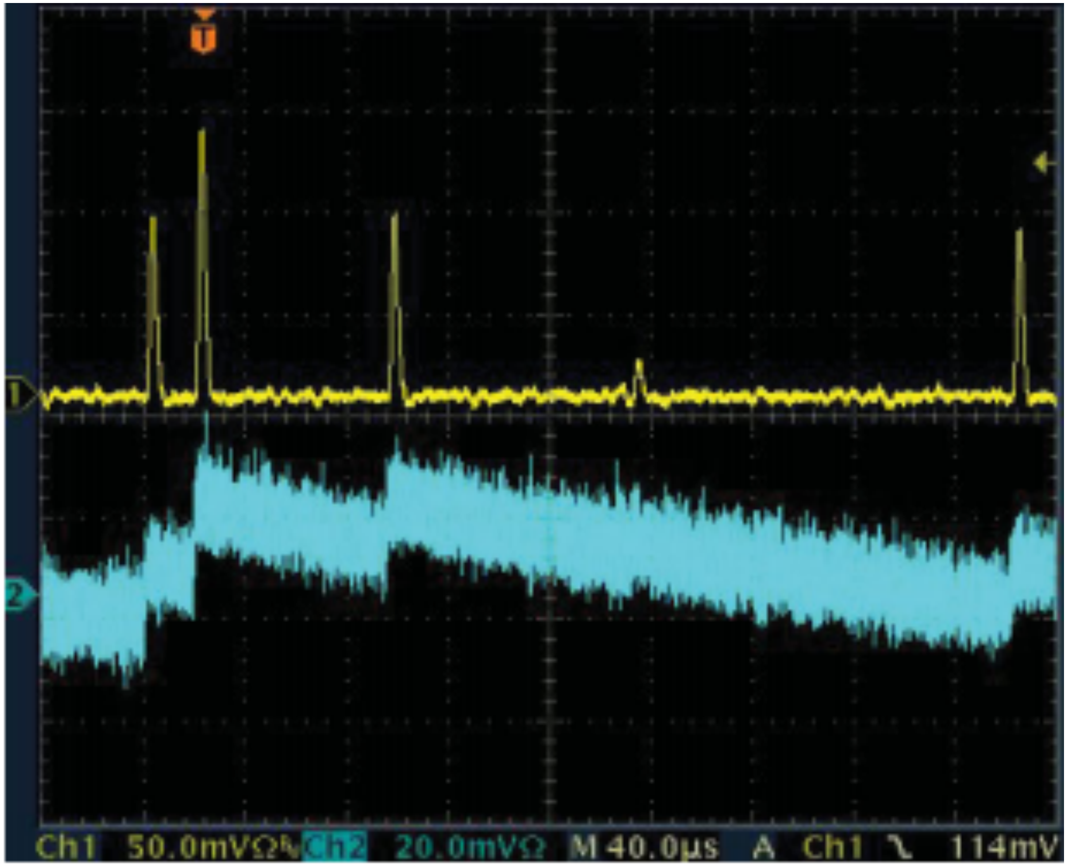
Fig. 6.13. Pulse shape of the CR-(RC)^2 circuit.

www.**ortec**-online.com/download/**amplifier**-introduction.pdf

www.**ortec**-online.com/download/**572a**.pdf

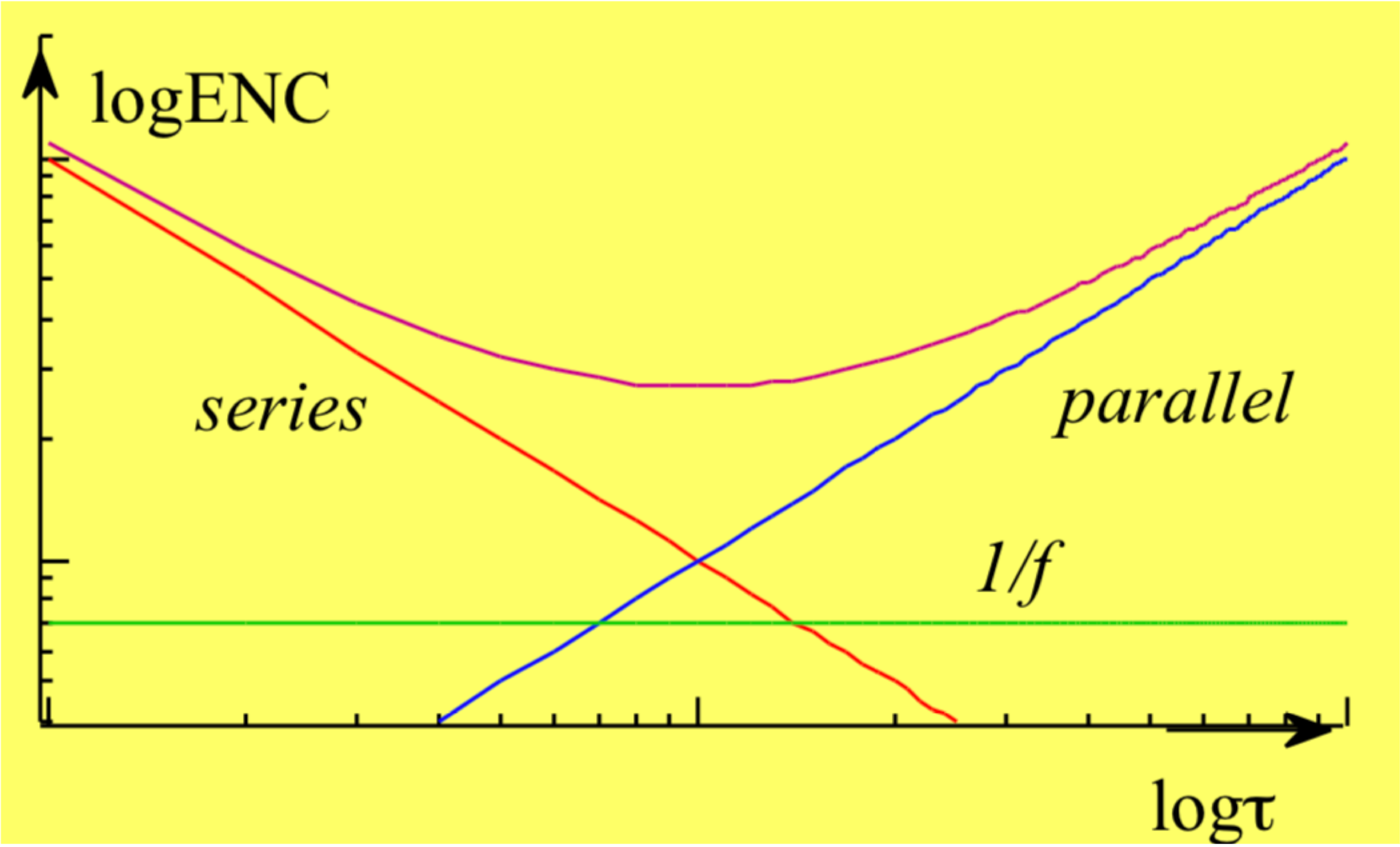
Semi-Gaussian shaping

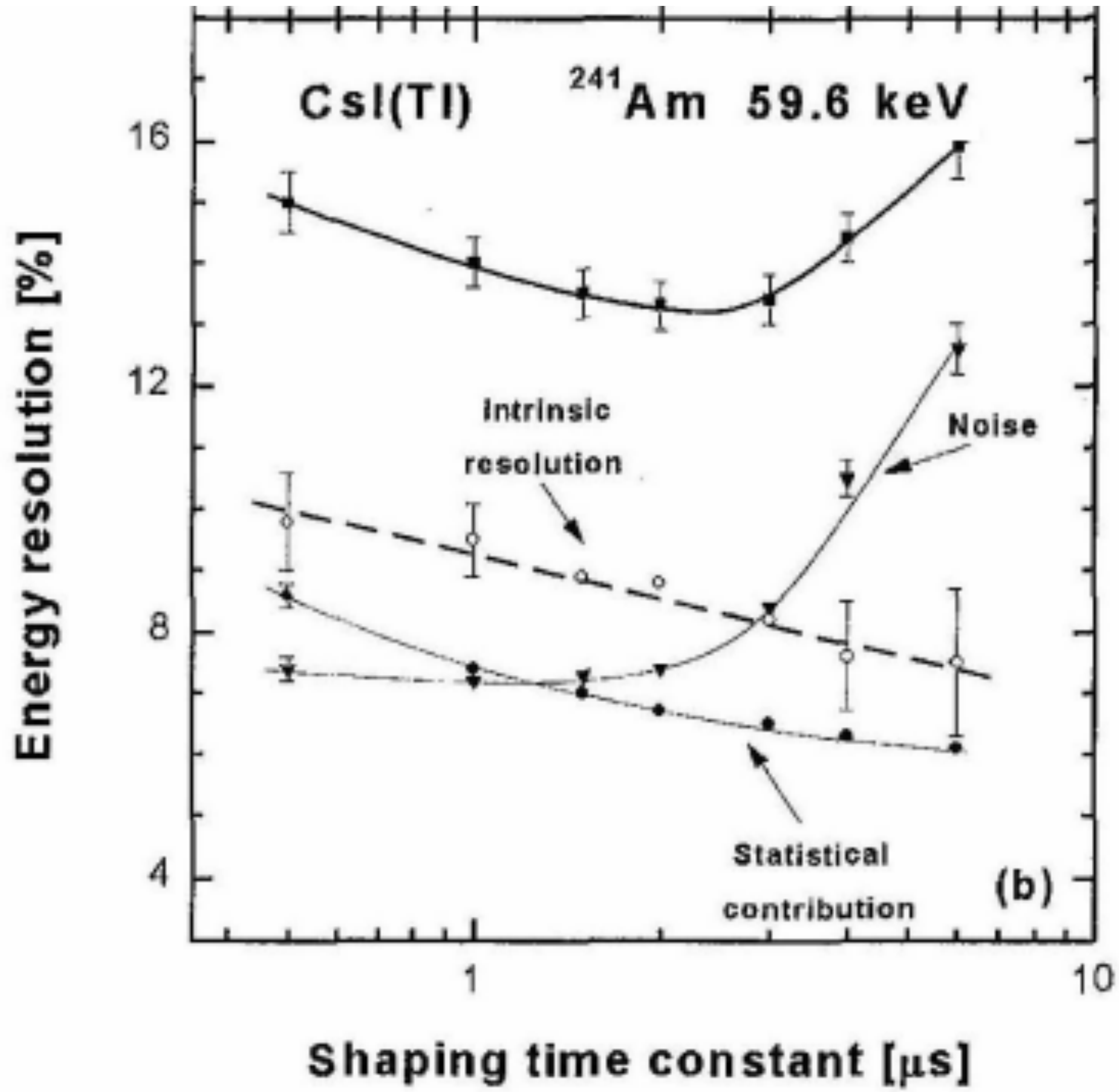
four or five RC stages are sufficient to produce a form having close to the theoretical signal-to-noise improvement.



The optimum pulse shaping time constant

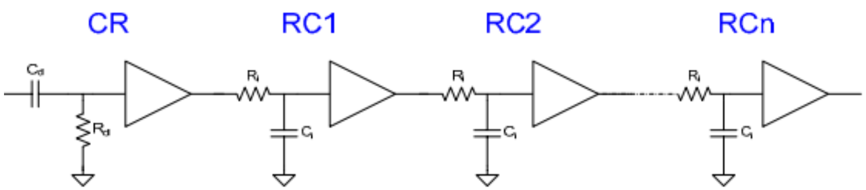
For semiconductor detectors, the electronic noise at the preamplifier input makes a noticeable contribution to the energy resolution. This noise contribution can be minimized by choosing the appropriate amplifier shaping time constant.





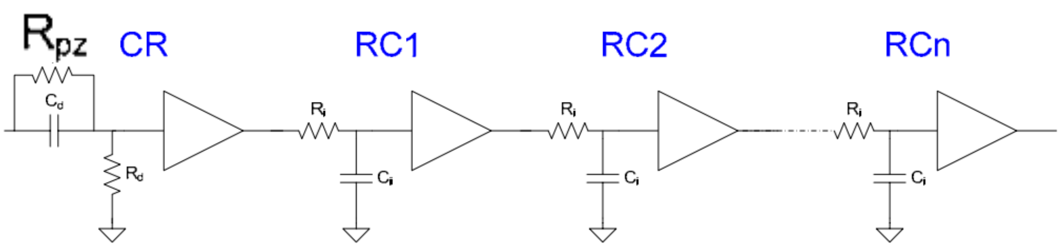
CsI(Tl) ^{241}Am 59.6 keV
16
12
8
4
Energy resolution [%]
Intrinsic resolution
Noise
Statistical contribution
(b)
1
10
Shaping time constant [μs]

Pole-zero Cancellation



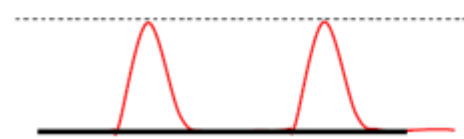
Undershoot !

Undershoot leading to degraded energy resolution

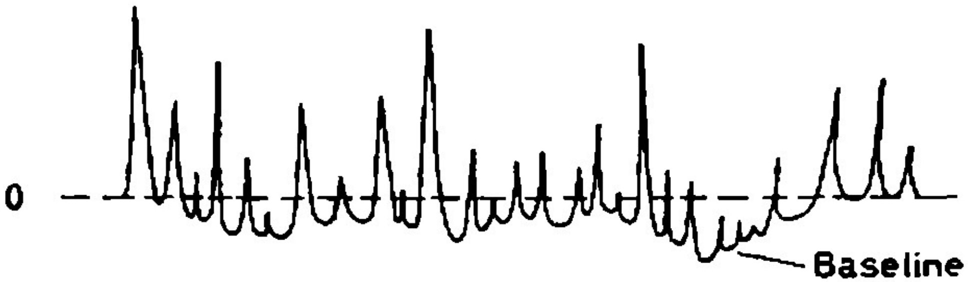


Undershoot leading to degraded energy resolution

- Zero: $R_{pz}C$ = pulse decay time $C_f R_f$



Baseline Restoration



Baseline shift at high count rates

It has to cut-off the low-frequency noise and disturbances and the drift of the DC level.

This function cannot be performed by a simple high-pass filter since it would alter also the signal pulse shape-> **Time-variant differentiator filter**

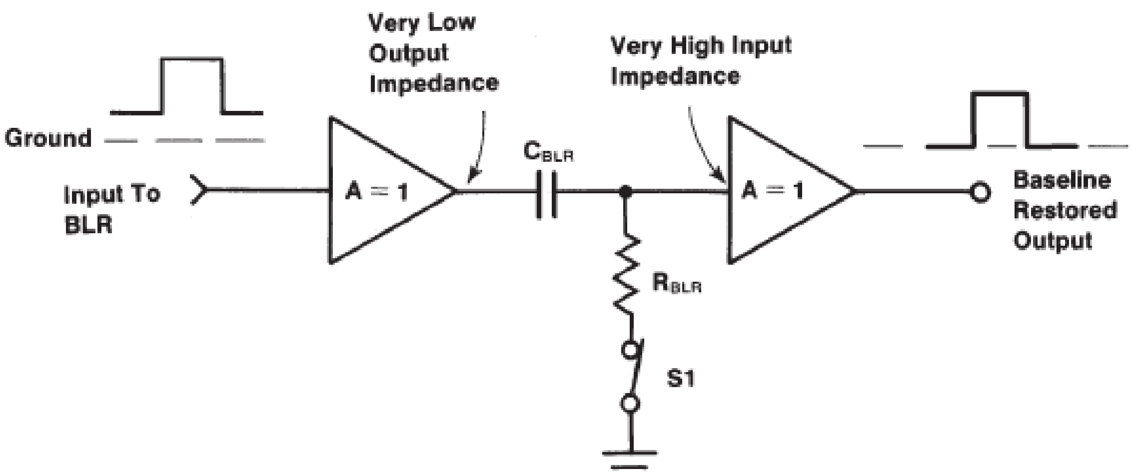


Fig. 20. A Simplified Diagram of a Baseline Restorer.

Pile-Up Rejection

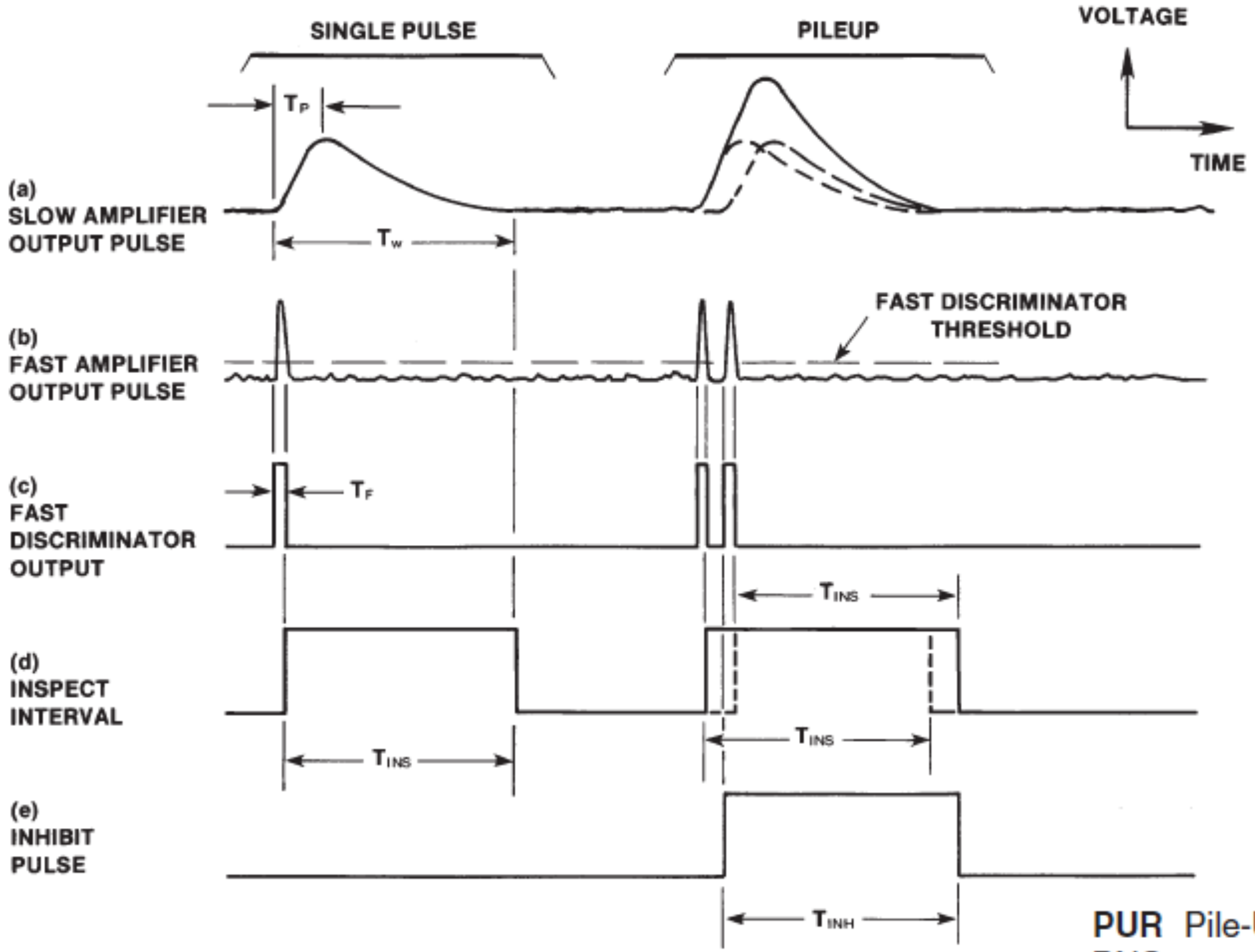
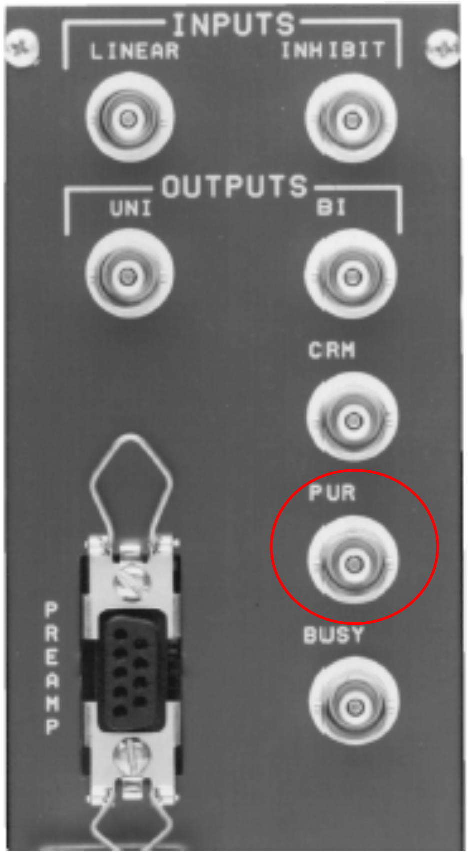


Fig. 22. Basic Waveforms in the Pile-Up Rejector.

Rear panel



PUR Pile-Up Reject output is a rear-panel, BNC connector. Provides a +5-V NIM standard logic pulse when pulse pile-up is detected. Output also present for a pulsed reset

www.**ortec**-online.com/download/**672**.pdf

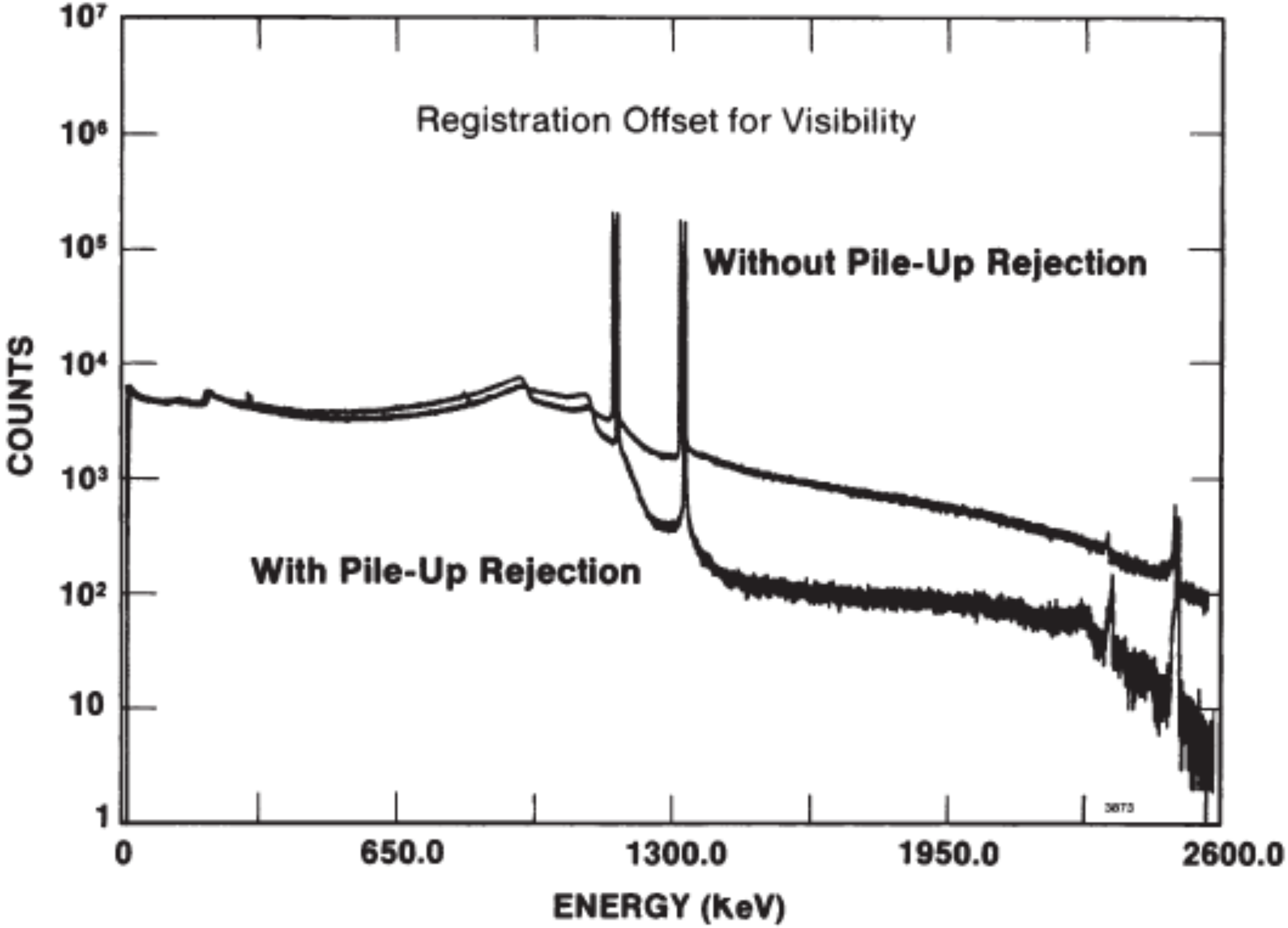


Fig. 23. Demonstration of the Effectiveness of the Pile-Up Rejector in Suppressing the Pile-Up Spectrum with a Germanium Detector and a ^{60}Co Spectrum at 50,000 Counts/s.

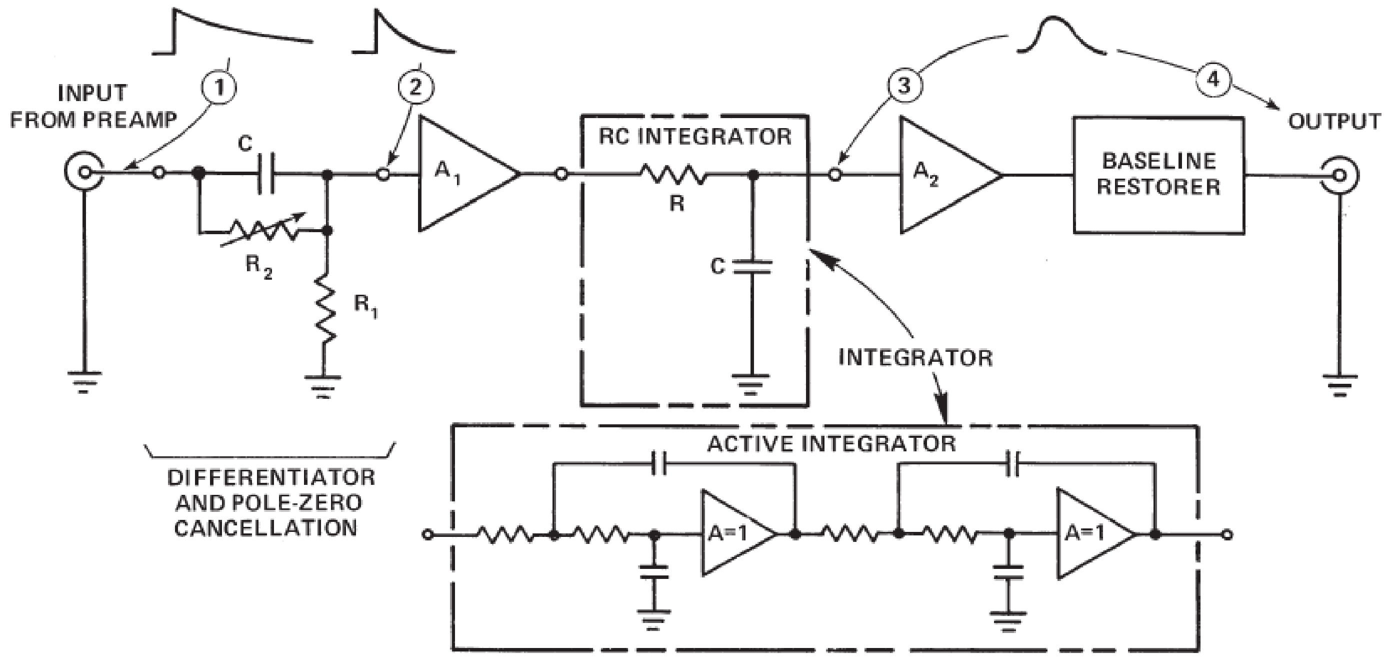
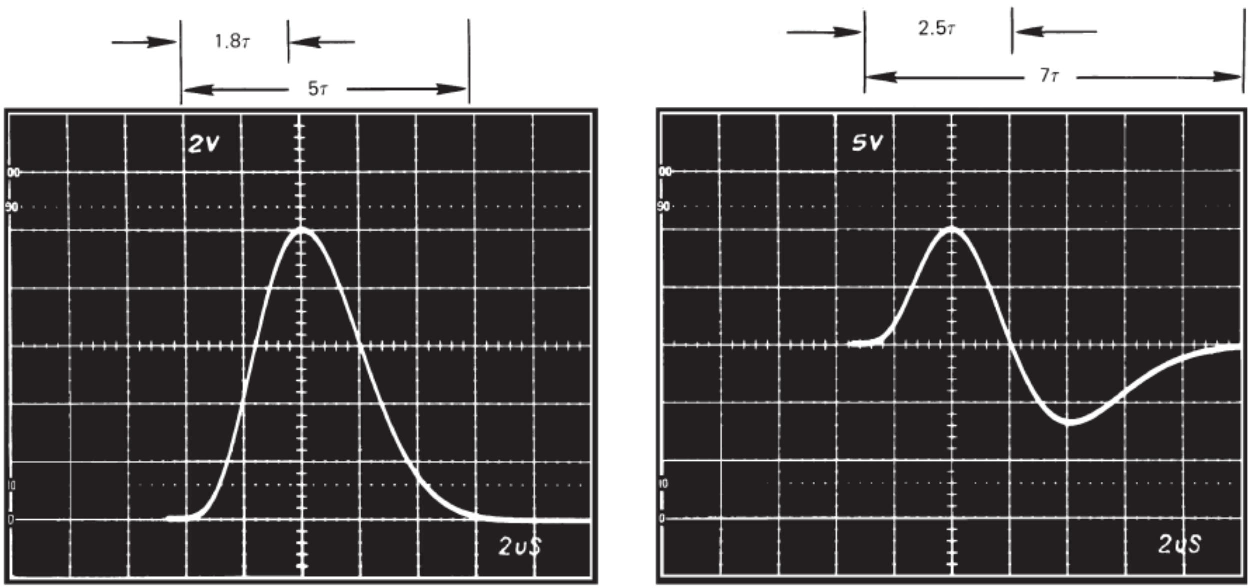


Fig. 13. Pulse Shaping in the Semi-Gaussian Shaping Amplifier.

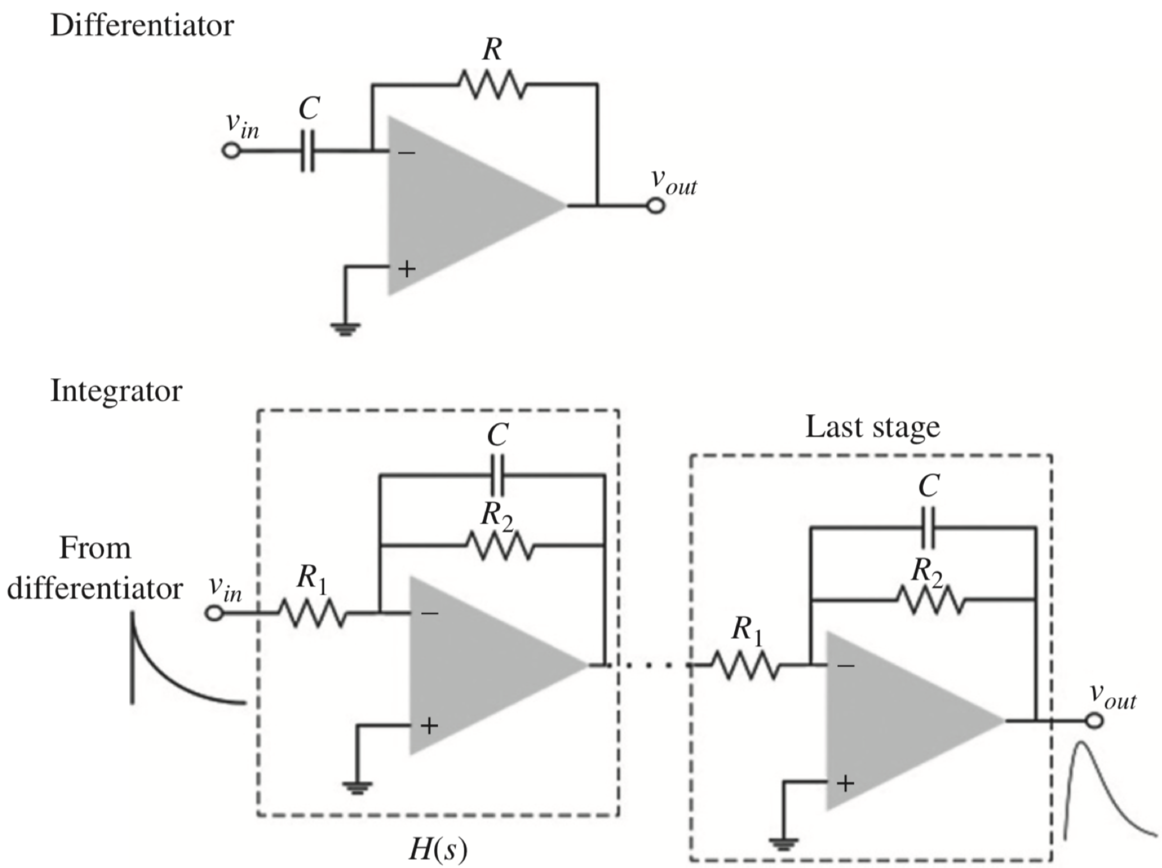


a. Unipolar Output.

b. Bipolar Outputs.

Fig. 14. Typical (a) Unipolar, and (b) Bipolar Output Pulse Shapes from a Semi-Gaussian Shaping Amplifier.

A more practical approach to implement *CR*–$(RC)_n$ filters is to use first-order active differentiator and integrator filters instead of passive *RC* and *CR* filters isolated by buffer amplifiers.



Single-Channel Analyzer

SCA

A single-channel analyzer produces an output logic pulse only if the peak amplitude of its input signal falls within the pulse-height window that is established with two preset threshold levels.

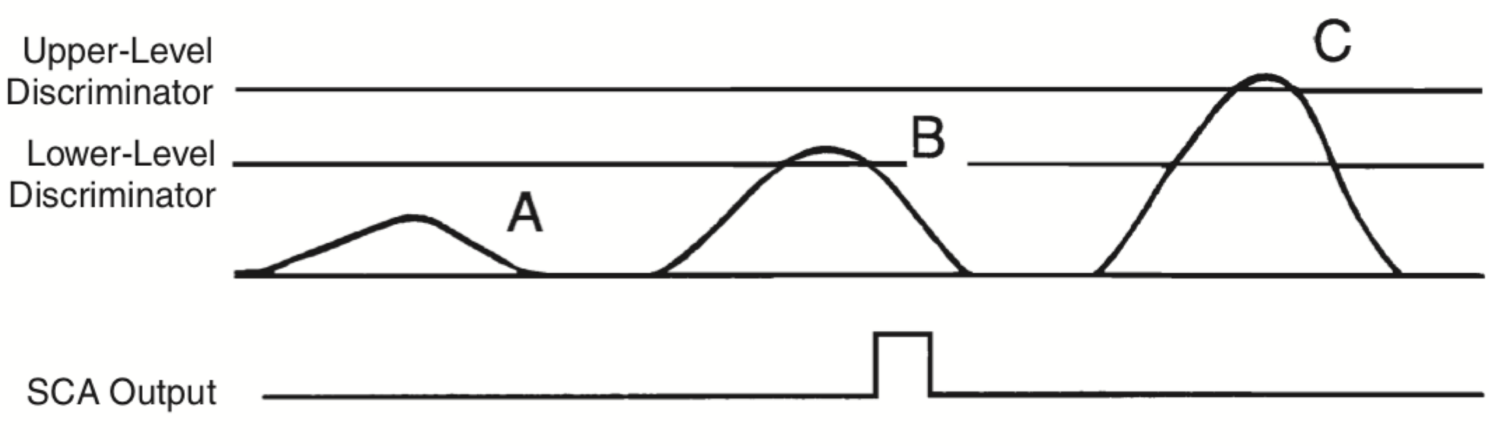


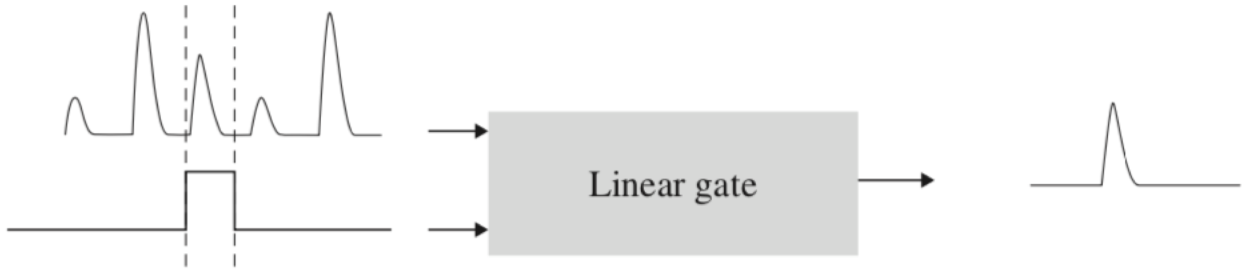
Fig. 2. Single-Channel Analyzer Function.

Analog-to-Digital Converters (ADC)

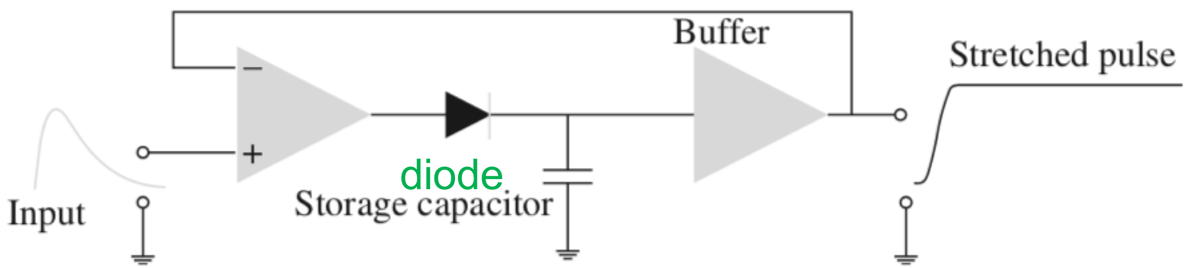
ADC measures the height of an analog pulse and converts that value to a digital number. The digital output is a proportional representation of the analog pulse height at the ADC input.

Linear Gates

the transmission of analog pulses is controlled by applying a logic pulse at a control input.



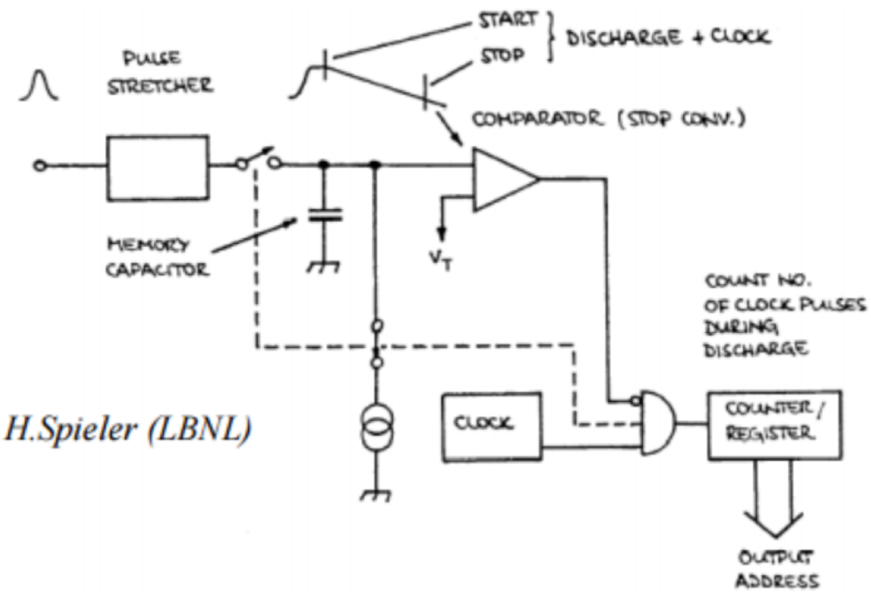
Peak Stretcher



$V_{in}>V_{out}$, the current from op-amp charges the storage capacitor.

$V_{in}<V_{out}$, the hold capacitor cannot be discharged as the diode is in reverse bias, and thus it holds the maximum value of the input.

Wilkinson-Type ADC



H.Spieler (LBNL)

The input signal is stored as a charge on a capacitor that is discharged by a constant current source with a clock (100 up to 400 MHz)

Variable conversion time.
Linearity: one comparitor but stretcher
Input rate limited by clock rate

Highest resolution devices in nuclear physics, used for Ge detectors.

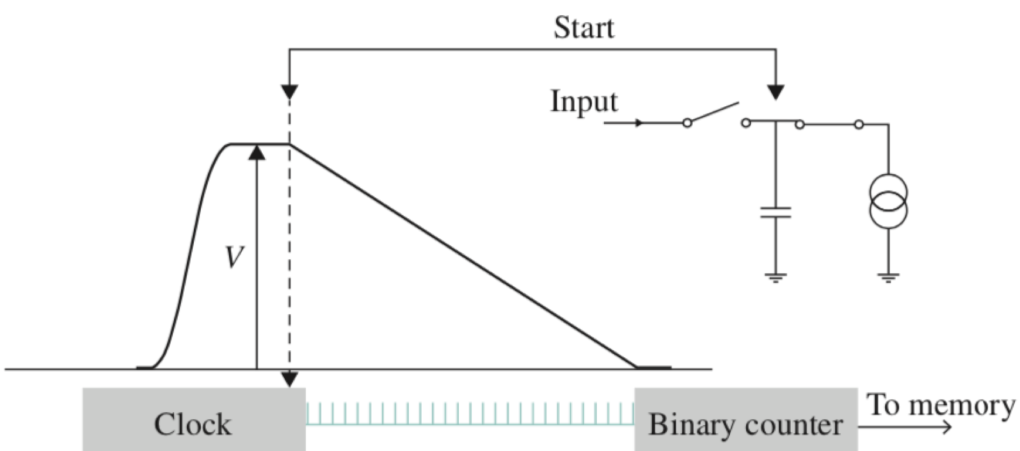


Figure 4.40 Principle of a Wilkinson-type ADC.

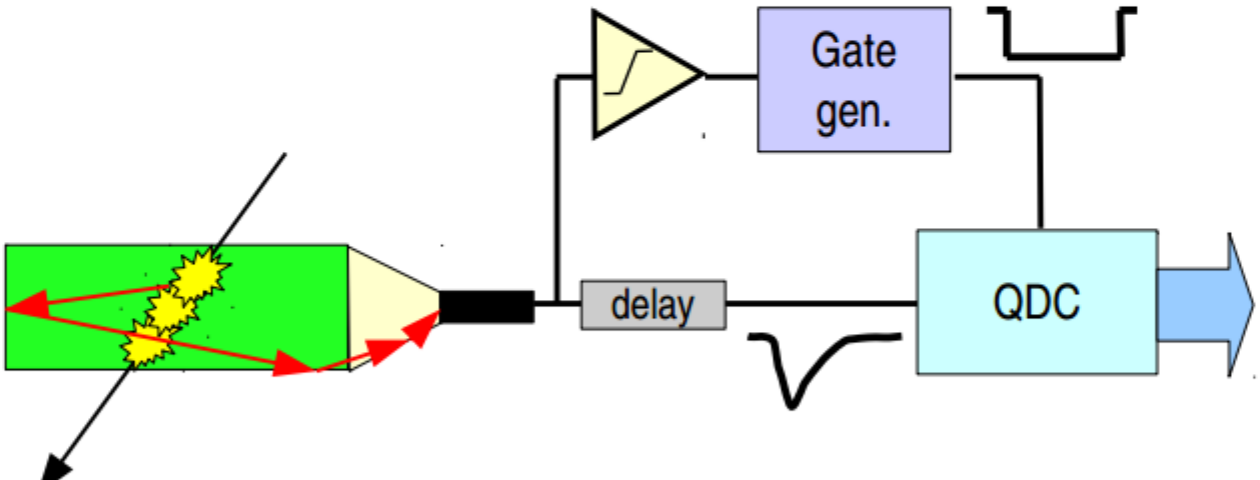
Amplitude to channel

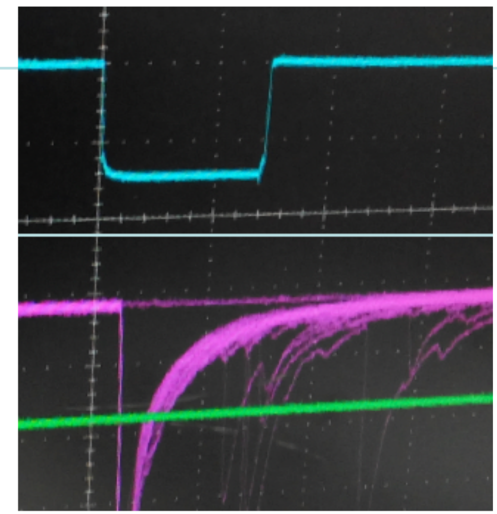
➔ADC converts a voltage into a digital representation.

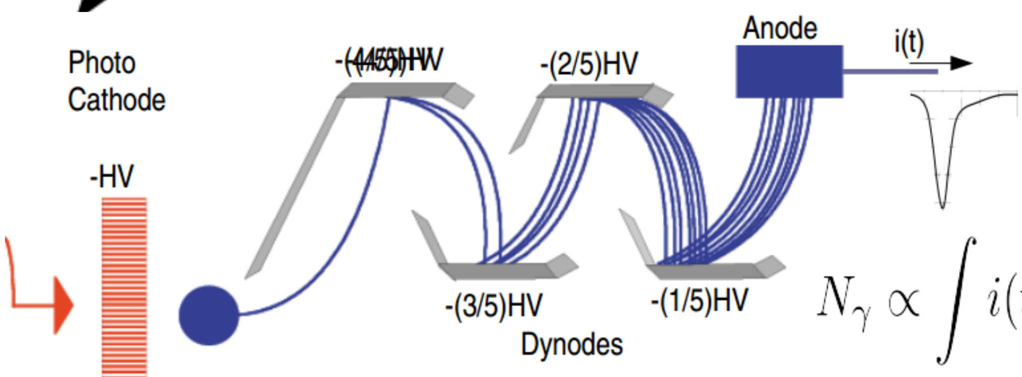
➔QDC → Charge to Digital Converter

- Essentially an integration step followed by an ADC

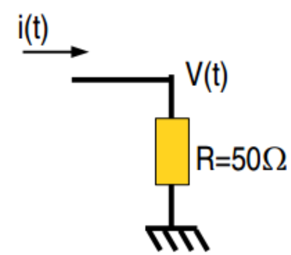
➔Integration requires limits → gate





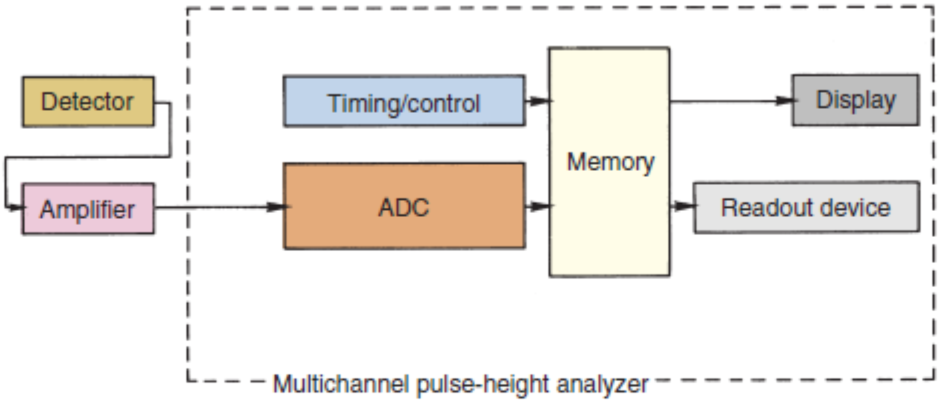


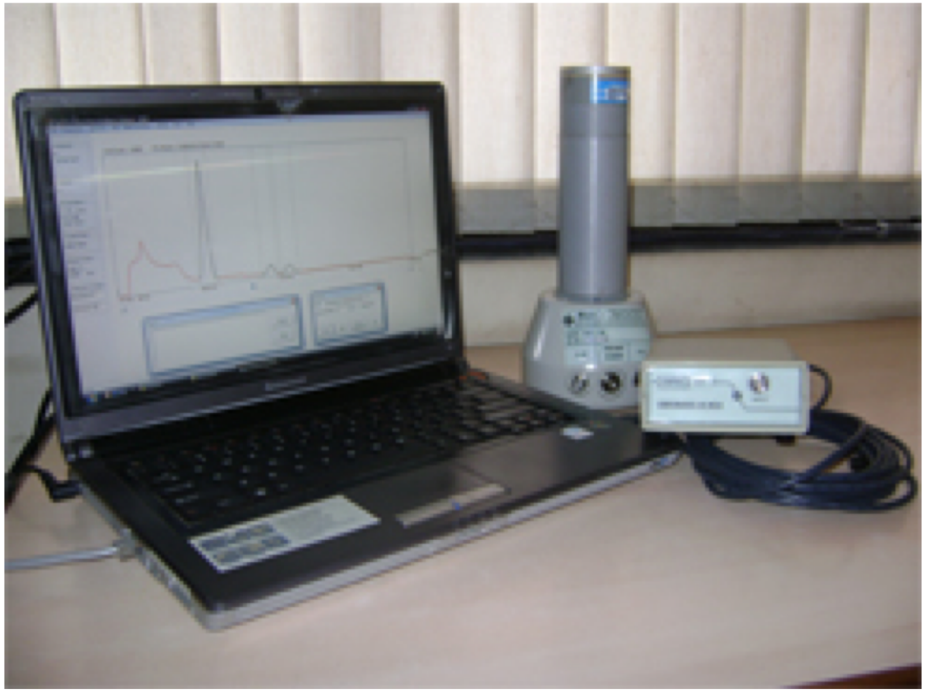
$$N_\gamma \propto \int i(t)\,dt = Q$$



$$Q = \int i(t)\,dt = \frac{1}{R}\int V(t)\,dt$$

Multi-Channel Analyzers(MCA)





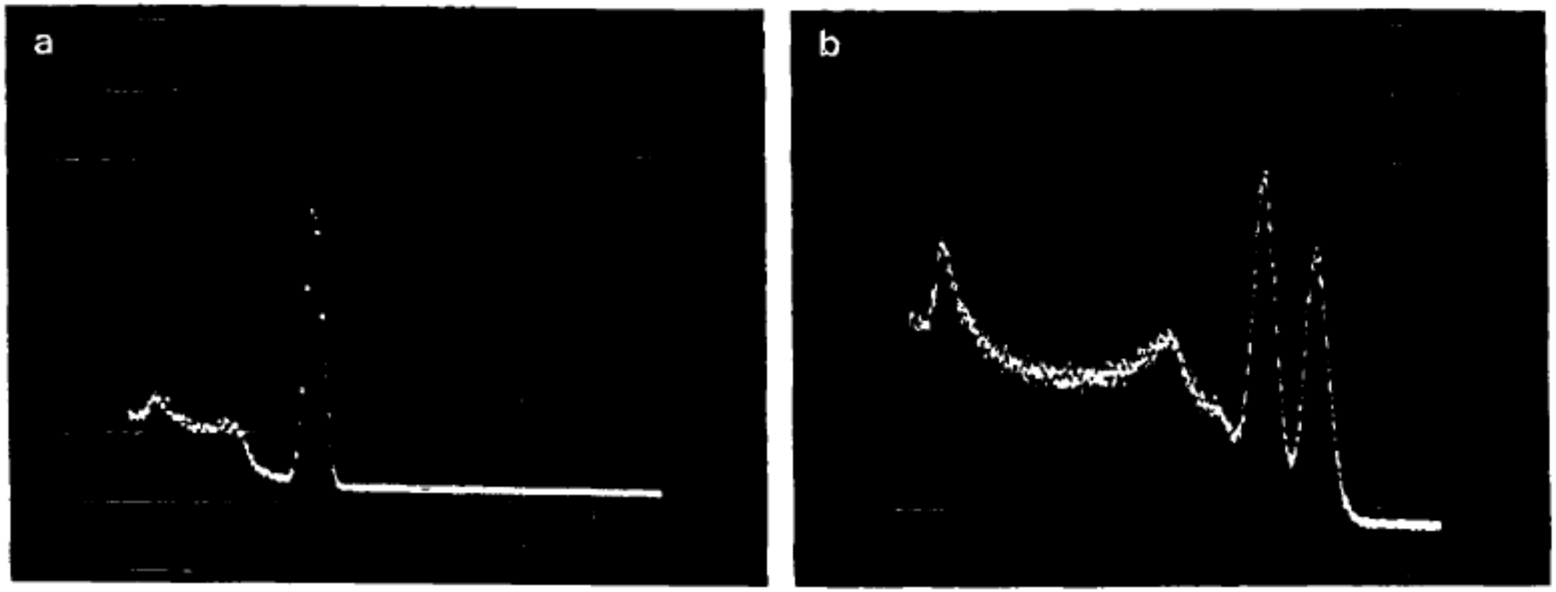
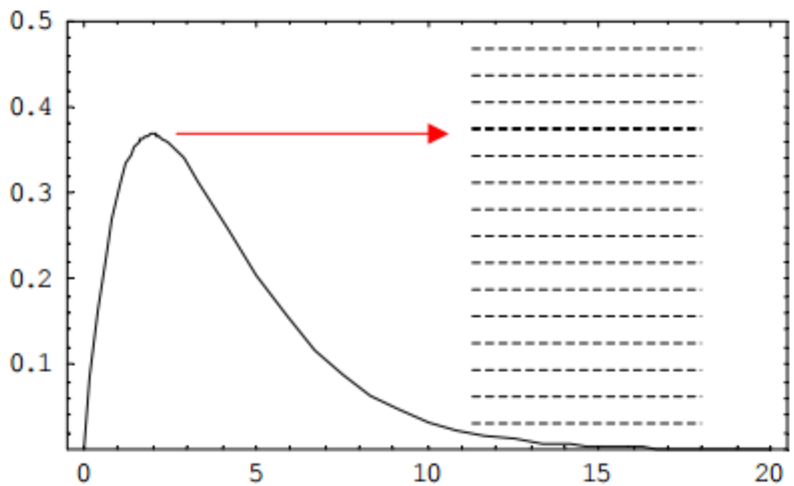


Fig. 15.6. Sample MCA pulse height spectra from a NaI detector: (a) ^{137}Cs, (b) ^{60}Co

Resolution: the resolution of an ADC is specified in terms of both the (voltage) range and the digital range (number of bits).

The voltage associated with the least significant bit (LSB) is $(V_{max} - V_{min}) / 2^N$

Perfect device sorts the data into 2^N bins of equal width = 1 LSB



Example: $\Delta V = 0.5$ V

$N=4$, $2^N=16$ $V_{LSB}= 0.03125$

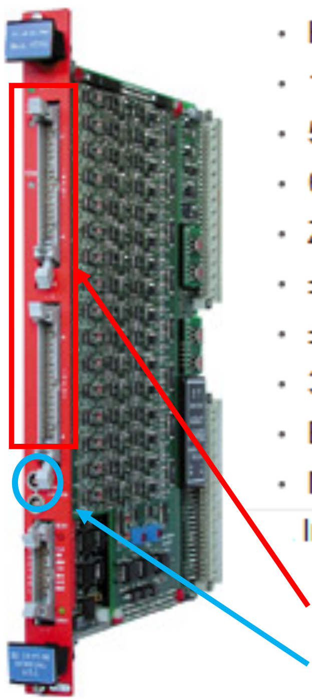
Peak in bin #:

Decimal: 12 binary: 1100, Hex: C

Home / Products / Front End / VME / ADCs (Peak Sensing) / **V785**

V785

32 Channel Multievent Peak Sensing ADC → 4096 channels



+ Image

- High channel density
- 12-bit resolution
- 5.7 µs / 32 ch conversion time
- 600 ns fast clear time
- Zero and overflow suppression for each channel
- ±1.5% differential non linearity
- ±0.1% integral non linearity
- 32 event buffer memory
- BLT32/MBLT64/CBLT32/CBLT64 data transfer
- Multicast commands

Input voltage range is 0 ÷ 4 V.

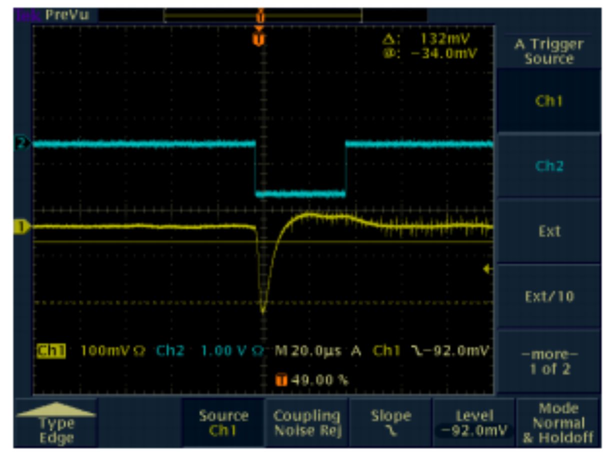
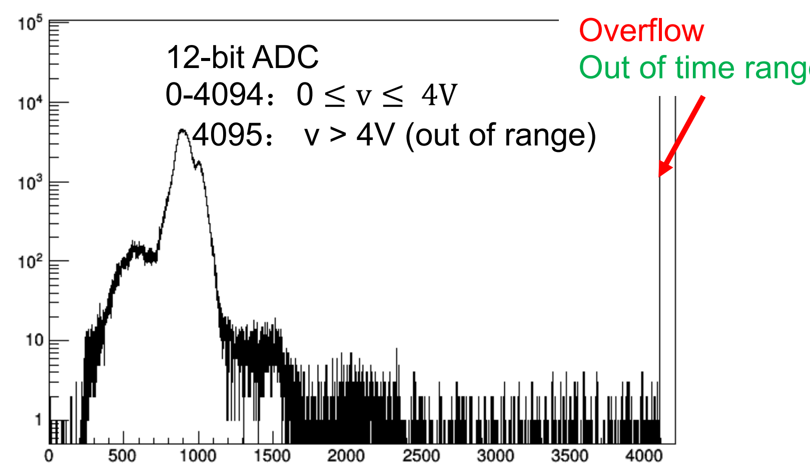
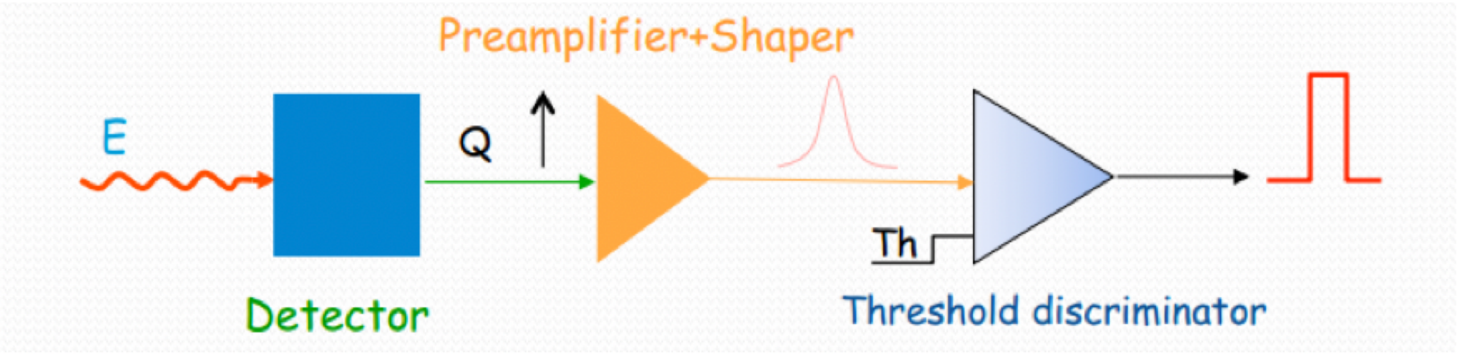


Figure 2.21: ADC gate and signal



Time measurement



In many applications, information on the accurate arrival time of a quantum of radiation in the detector is required. For timing information, detector pulses are usually processed quite differently than the pulse height analysis.

Impedance matching

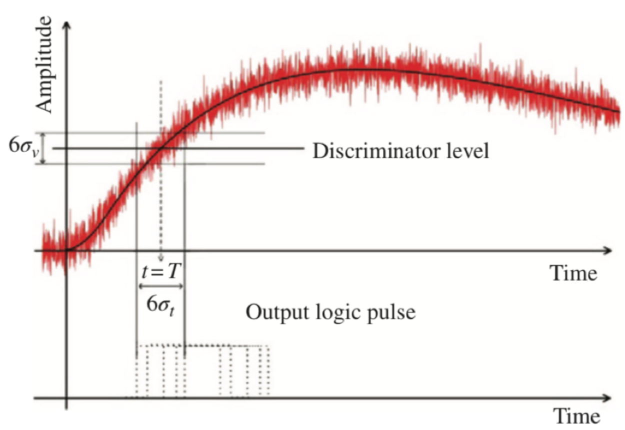
there are situations in which it is rather required to have $Z_{out} = Z_{in}$. That is the case in radiofrequency circuits to avoid signal reflections. For time measurement(fast)

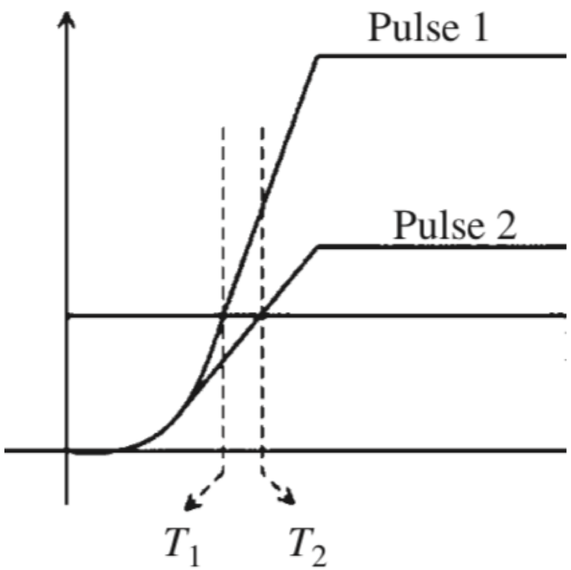
Z=50ohm

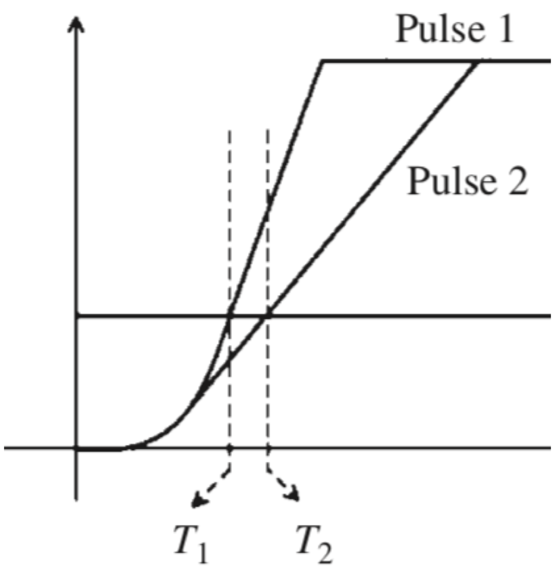
Time Pick-Off Techniques

Time pick-off is the fundamental operation generating a logic pulse to indicate the time of occurrence of an input linear pulse.The leading edge of the logic pulse corresponds to the time of occurrence.

There are two cases of inaccuracy in time pick-off: time jitter and amplitude walk. Time jitter is usually induced by random fluctuations in the signal pulse size and shape. Amplitude walk is the effect induced by the variable amplitudes and risetime of input pulses.

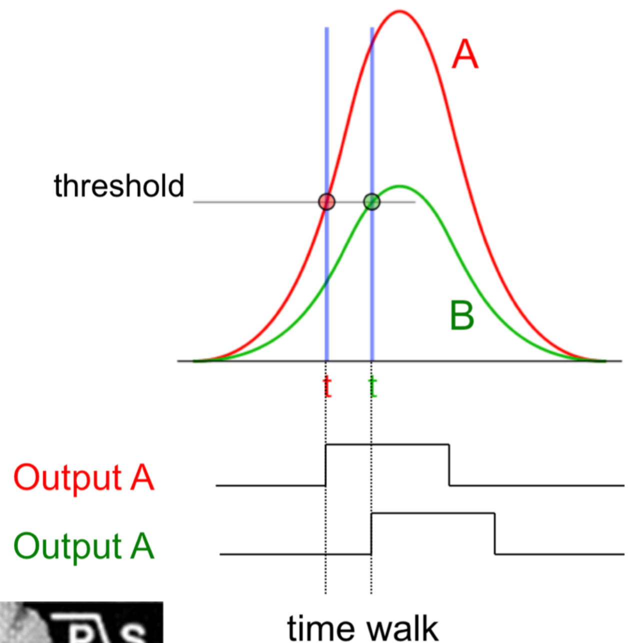


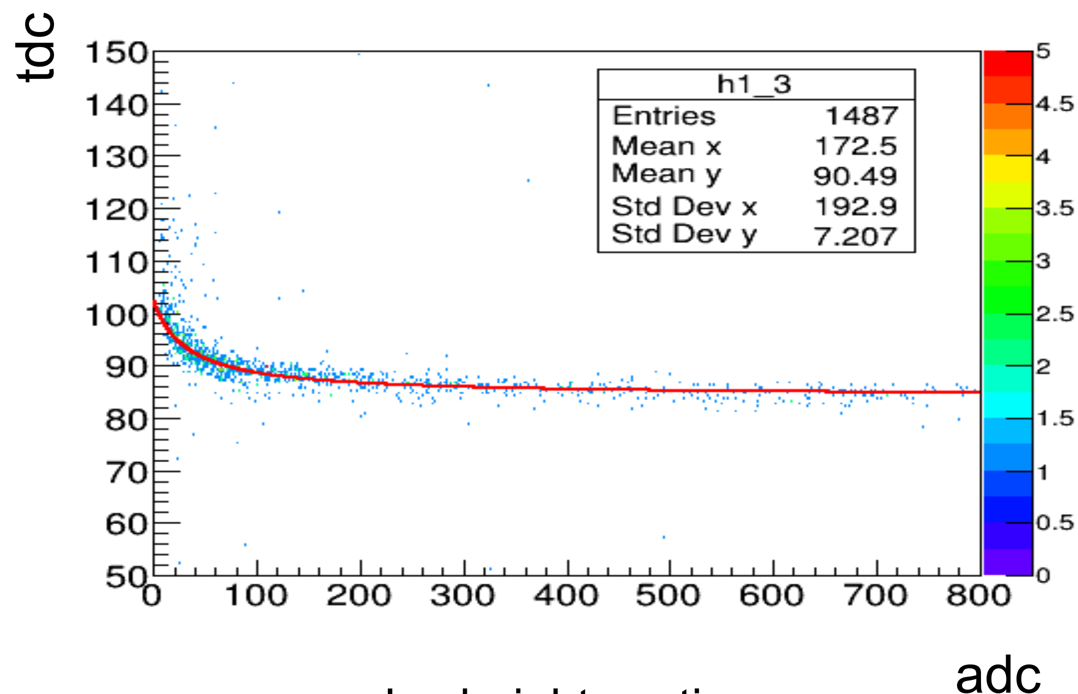




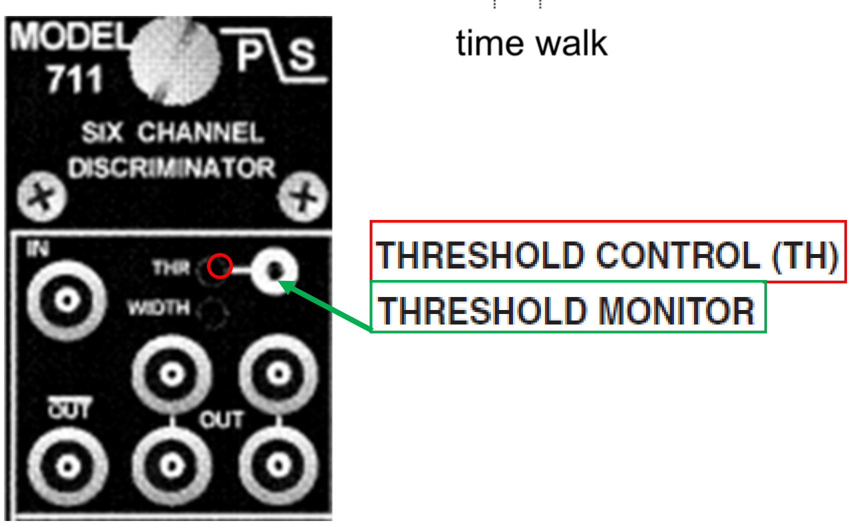
Leading edge timing

Leading edge timing is the simplest time pick-off method and generates the output timing logic pulse when an input pulse crosses a fixed discrimination level. This method is easy to implement and is effective when the dynamic range of the input pulses is not large.

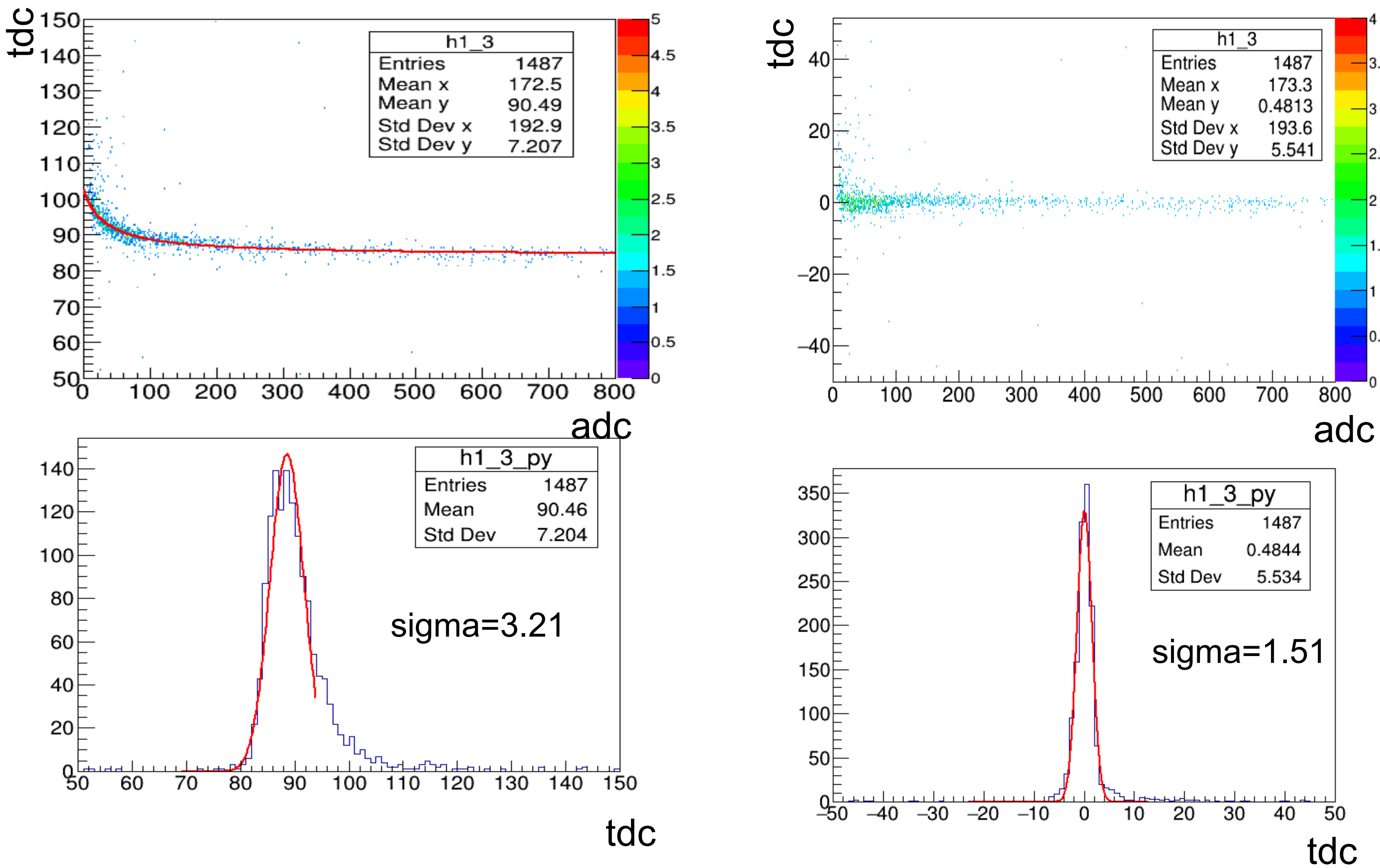




pulse height vs. time

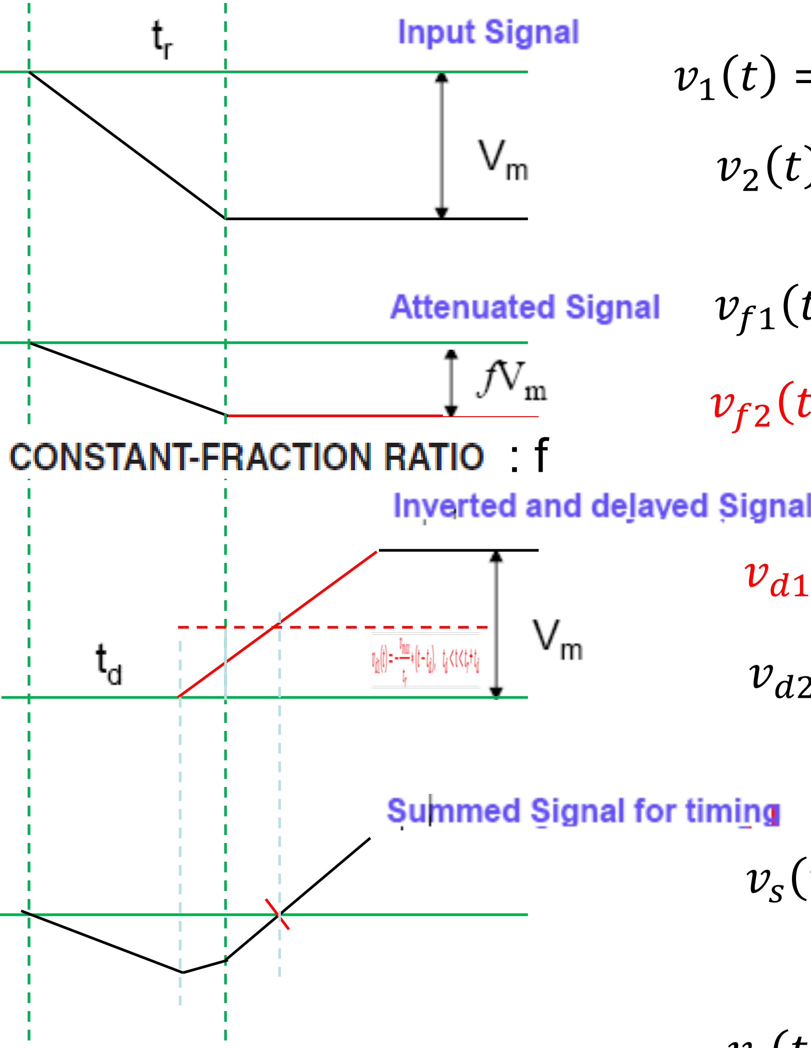


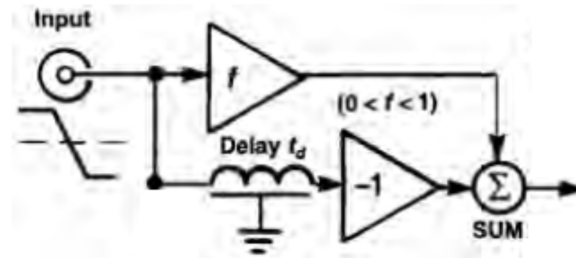
Time walk/slew correction



Constant Franction Discrimination CFD

rise times of all signals are the same





$$v_1(t) = \frac{v_{max}}{t_r} * t, \qquad 0 < t < t_r$$

$$v_2(t) = v_{max}, \qquad t > t_r$$

$$v_{f1}(t) = f\frac{v_{max}}{t_r} * t, \qquad 0 < t < t_r$$

$$v_{f2}(t) = f v_{max}, \qquad t > t_r$$

$$v_{d1}(t) = -\frac{v_{max}}{t_r} * (t - t_d), \qquad t_d < t < t_r + t_d$$

$$v_{d2}(t) = -v_{max}, \qquad t > t_r + t_d$$

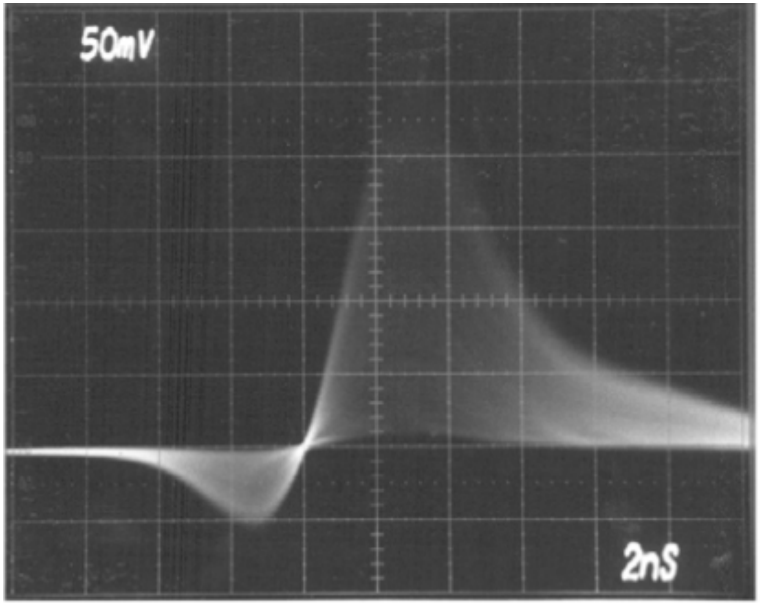
$$v_s(t) = v_{f2}(t) + v_{d1}(t)$$
$$= f v_{max} - \frac{v_{max}}{t_r} * (t - t_d), \qquad t_r < t < t_r + t_d$$

$$v_s(t_{cross}) = 0 \qquad , t_{cross} > t_r$$

The effect is to trigger a timing signal at a constant fraction of the input amplitude, usually chosen to be around 20%

$$t_{cross} = f t_r + t_d$$

No amplitude dependence



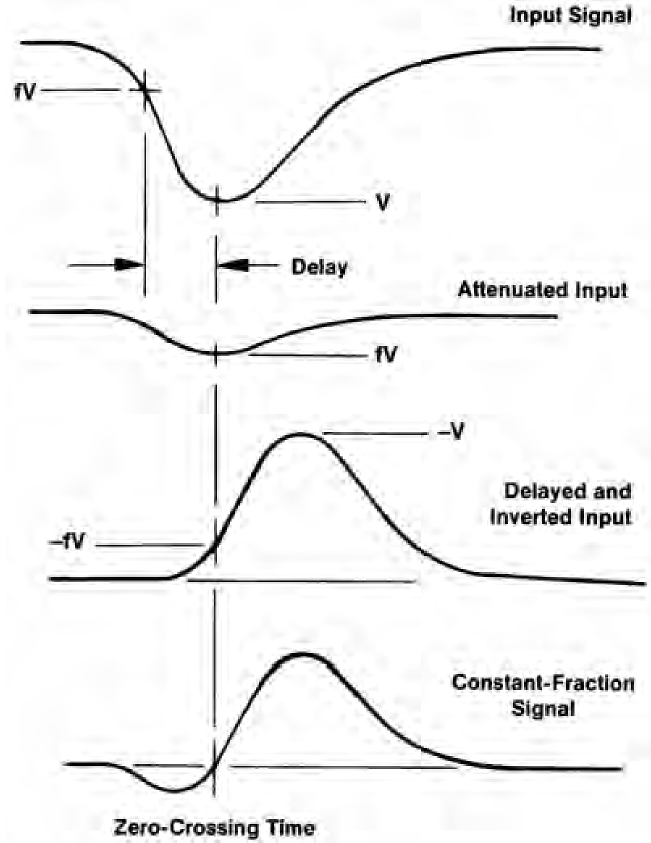


Fig. 3. Formation of the Constant-Fraction Signal.

$t_{cross} = ft_r + t_d$, $t_{cross} > t_r$

$t_d > t_r(1-f)$

CFD timing:

the zero-crossing time occurs after the attenuated input signal has reached its maximum pulse height.

Practical timing experiments involve input signals with finite pulse width, therefore the shaping delay must also be made sufficiently short to force the zero crossing signal to occur during the time that attenuated signal is at its peak.

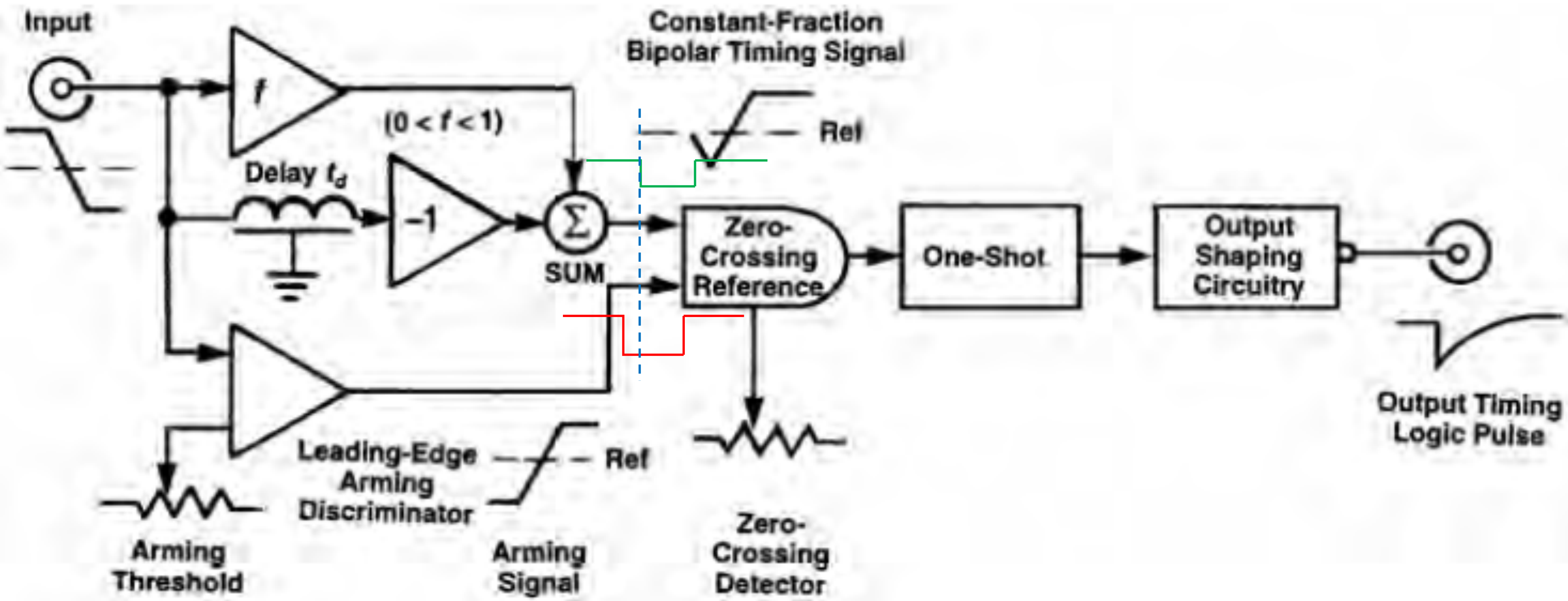


Fig. 4. Functional Representation of a Constant-Fraction Discriminator.

A leading-edge arming discriminator provides energy selection and prevents the sensitive zero-crossing comparator from triggering on any noise inherent in the baseline preceding the pulse.

CONSTANT-FRACTION RATIO 0.4.

THRESHOLD CONTROL (TH) 20-turn front-panel screwdriver adjustment for each discriminator channel; nominally variable from –30 mV to –1 V.

DELAY Internal PCB jumper setting allows the proper shaping delay to be selected. Five possible positions: 2, 4, 6, 8, or 10 ns. Other delays available on order.

PERFORMANCE

WALK <±250 ps from –50 mV to –5 V for a pulse rise time of 1 ns, a pulse width of 10 ns, a 2-ns delay, and the threshold set at minimum.

For plastic scintillator+PMT

Rise time= 4ns

T_{delay} ~ 4x(1-0.4)=2.4 ns

$t_d > t_r(1-f)$

For silicon time signal

Rise time = 60ns

T_{delay}~ 60x(1-0.4)=36 ns

Slew/walk correction
time walk vs. pulse height

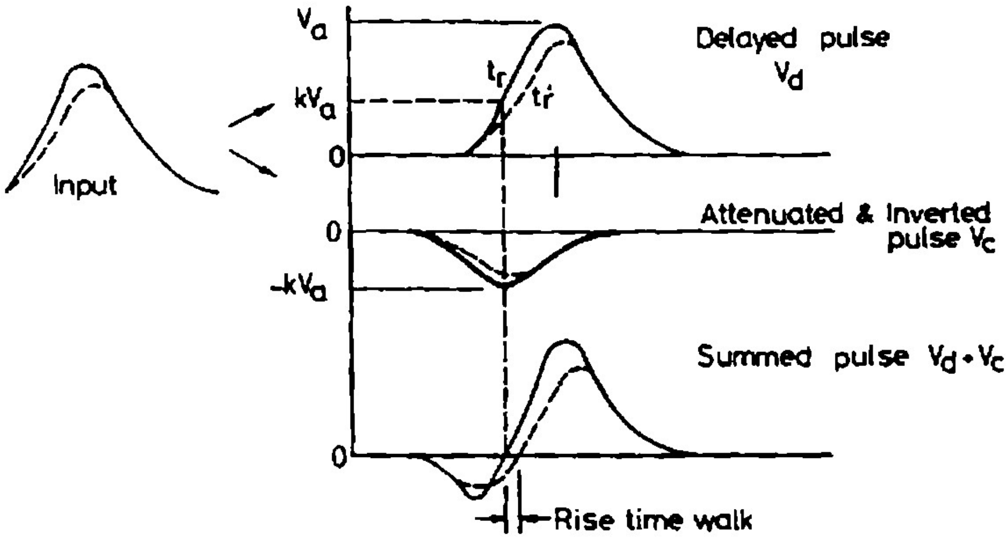
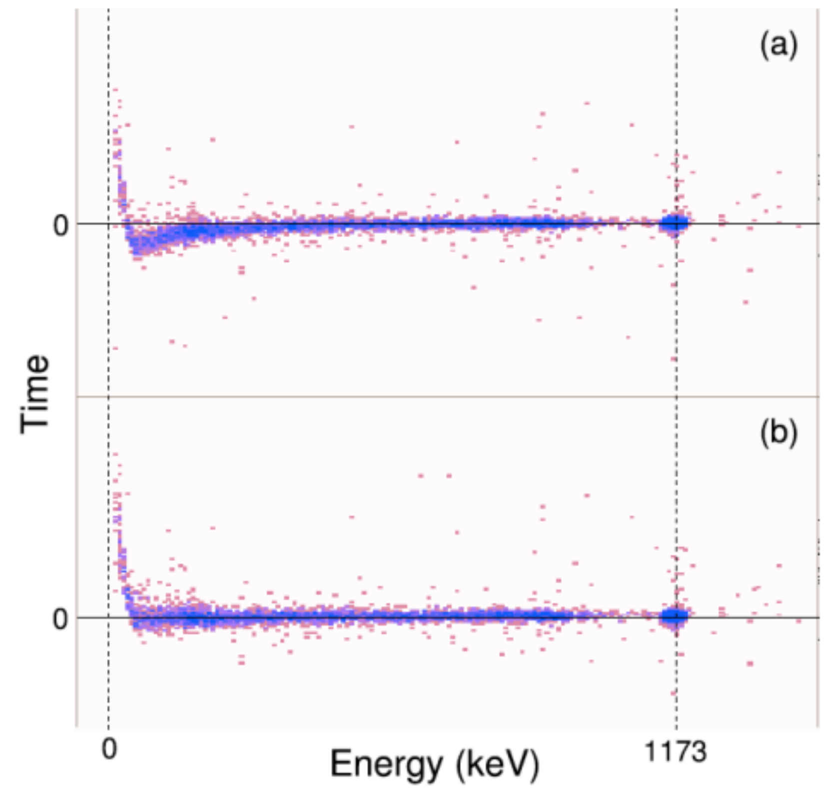


Fig. 17.5. Technique for constant fraction triggering. In order for this technique to work rise times of all signals must be the same. The dotted line shows the result with a different rise time signal



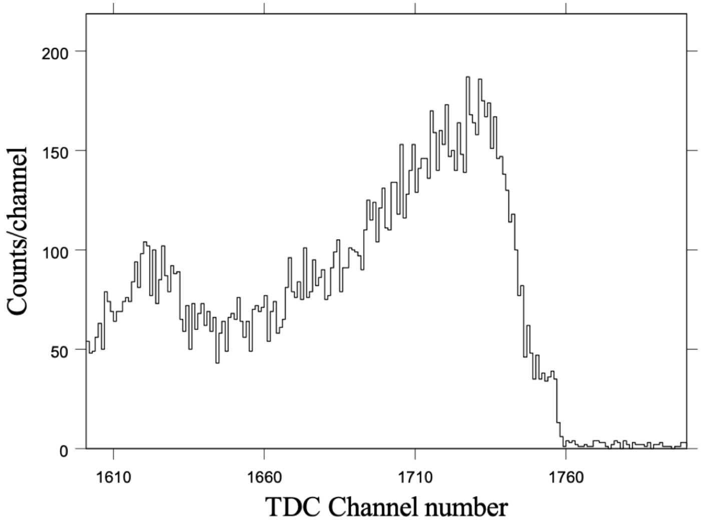


Figure 2. The TDC spectrum for neutron and gamma-ray events in the detector at 20° (compare with Fig. 1). Note the lack of a prompt gamma-ray peak due to poor timing resolution resulting from excessive CFD walk.

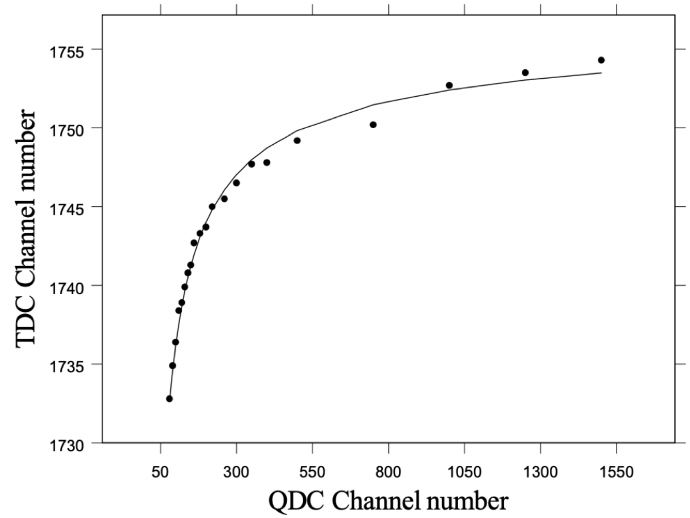


Figure 5. Plot of the data from Table I. The values used for QDC channel number are the midpoints of the QDC ranges reported in Table I. Each TDC channel is equal to 250 ps.

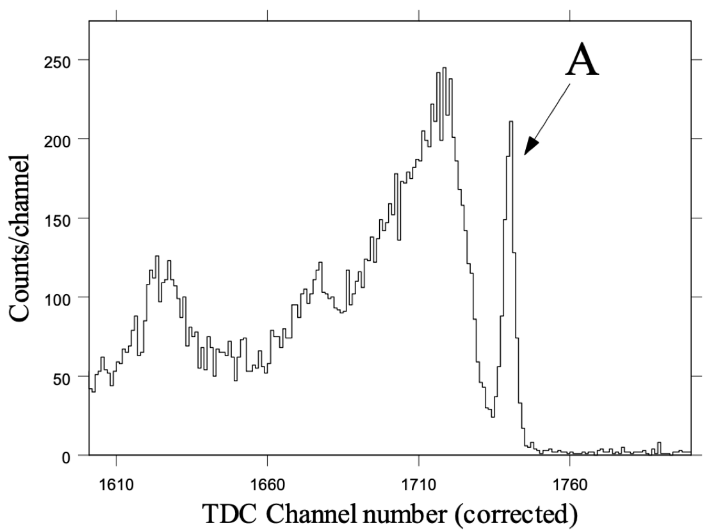
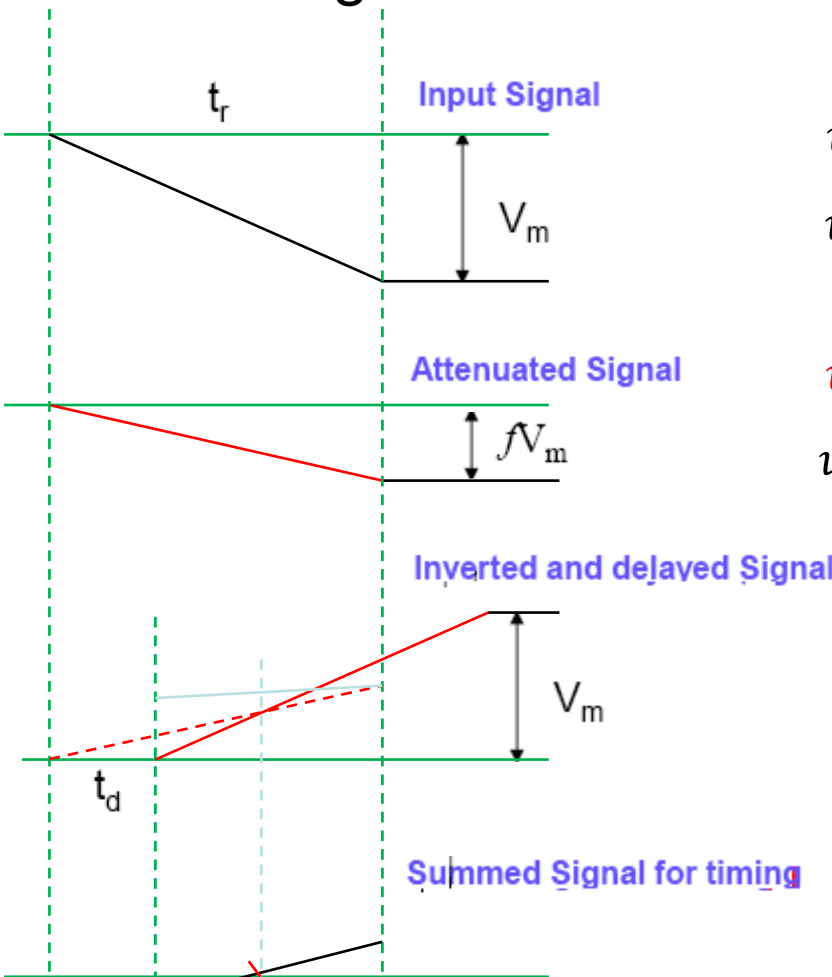


Figure 6. The TDC spectrum for neutron and gamma-ray events in the 20° detector, after applying a correction for CFD walk to the TDC value based on its corresponding QDC value. Figure 2 shows the same data before correction. Note the presence of a prompt gamma-ray peak ("A") not seen in Fig. 2.

http://www.if.pw.edu.pl/~alice/CFD8.pdf

ARC timing

For signals with different rise times



$$v_1(t) = \frac{v_{max}}{t_r} * t, \qquad 0 < \mathrm{t} < t_r$$

$$v_2(t) = v_{max}, \qquad \mathrm{t} > t_r$$

$$v_{f1}(t) = f\frac{v_{max}}{t_r} * t, \qquad 0 < \mathrm{t} < t_r$$

$$v_{f2}(t) = f v_{max}, \qquad \mathrm{t} > t_r$$

$$v_{d1}(t) = -\frac{v_{max}}{t_r} * (t - t_d), \qquad t_d < \mathrm{t} < t_r + t_d$$

$$v_{d2}(t) = -v_{max}, \qquad \mathrm{t} > t_r + t_d$$

$$v_s(t) = v_{f1}(t) + v_{d1}(t)$$
$$= f\frac{v_{max}}{t_r} * t_r - \frac{v_{max}}{t_r} * (t - t_d), \quad t_d < \mathrm{t} < t_r$$

$$v_s(t_{cross}) = 0$$

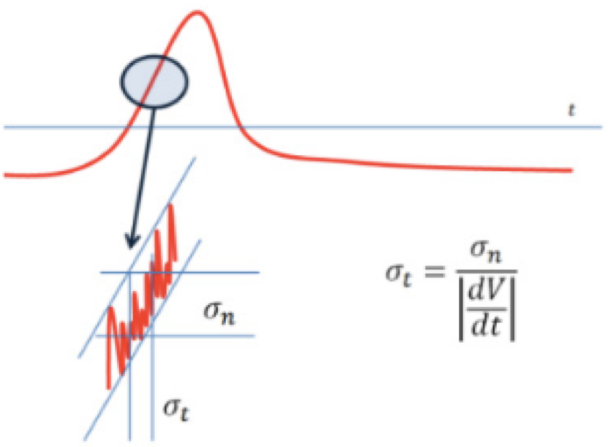
$$t_{cross} = t_d/(1-f), \quad , t_{cross} < t_r$$

$$t_d < t_r/(1-f)$$

No rise time & amplitude dependence

the time of zero crossover occurs *before* the attenuated input signal has reached it's maximum pulse height.

the slope of the ARC timing signal at zeros crossing is almost always less than the slope of the CFD signal at zero crossing. The timing error due to noise- induced jitter is usually worse for ARC timing than it is for CFD timing



- CFD timing is most effective when used with input signals having a wide range of amplitudes but a narrow range of rise times and pulse.

- ARC timing is most effective when used with input signals having a wide range of amplitudes and rise times regardless of pulse width widths.

- For a very narrow dynamic range of input signal amplitudes and rise times, leading-edge timing may provide better timing resolution if the timing jitter is dominated by noise rather than be statistical amplitude variations of the detector signals.

Timing Filter Amplifiers (TFA)

the slope-to-noise ratio of a pulse can be improved by using a proper pulse filter before feeding it into a time pick-off circuit.

The time constants of the integrator and differentiator are adjusted to reduce the high and low frequency noise, respectively.

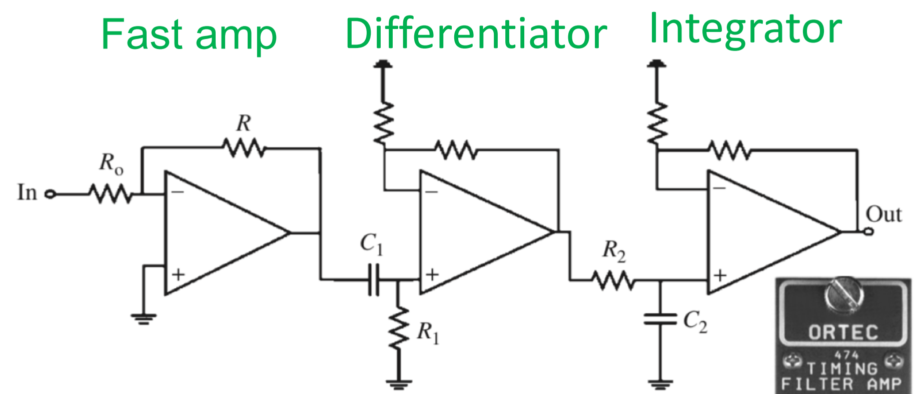


Figure 6.17 Block diagram of a timing filter amplifier.

Timing with Ge(Semiconductor) Detectors

The TFA output is used for time pick-off with a CFD that is operated in ARC timing mode by setting the delay to ~30% of the minimum risetime of TFA output pulses.

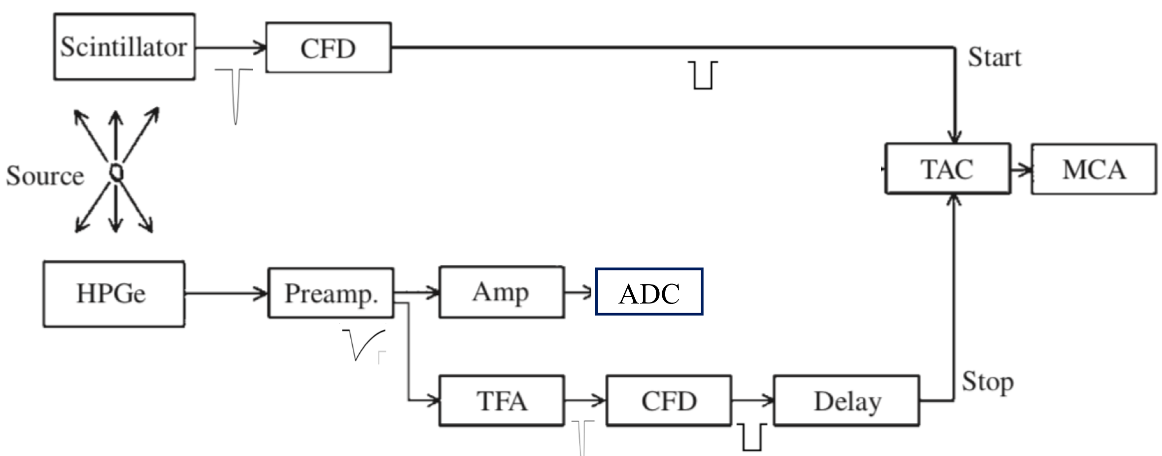
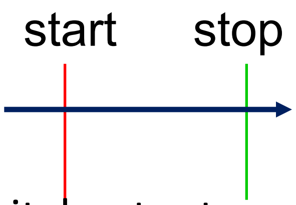


Figure 6.33 A setup for time resolution measurement of a germanium detector.

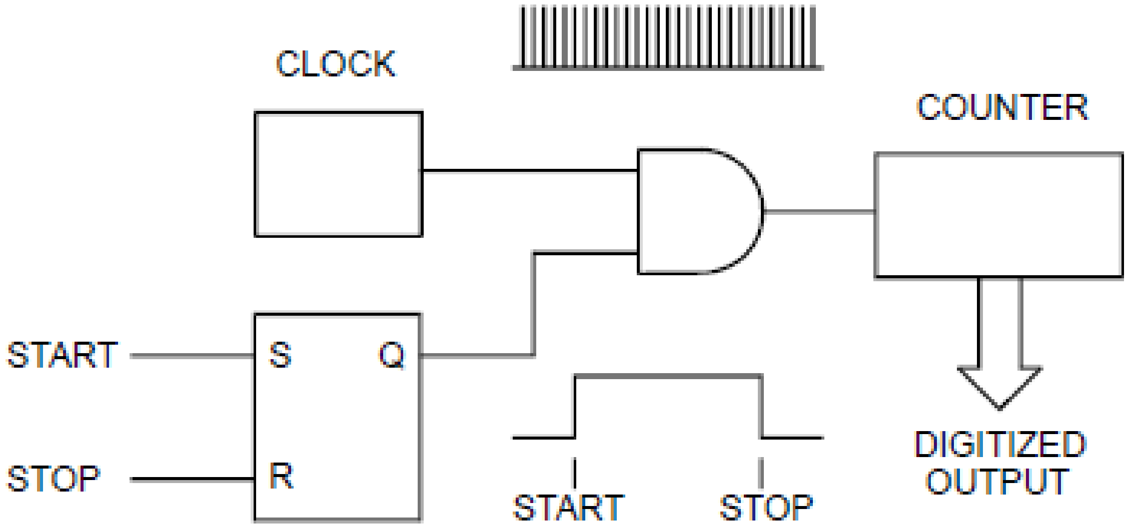
Real pulses from Ge detectors deviate from linear rise with slope changes, ARC timing cannot completely remove the time-walk. the time resolution of large volume HPGe detectors is generally limited to about 8–10 ns.

Time to Digital convertor TDC



Time Digitizers TDC directly convert the time intervals into a digital output.

1. Counter



Simplest arrangement: Count clock pulses between start and stop.

Limitation: Speed of counter

Current technology limits speed of counter system to about 1 GHz

$\Rightarrow \quad \Delta t = 1$ ns

2. Analog Ramp

Commonly used in high-resolution digitizers (Δt = 10 ps)

Principle:
Charge capacitor through switchable current source

Start pulse: turn on current source

Stop pulse: turn off current source

⇒ Voltage on storage capacitor

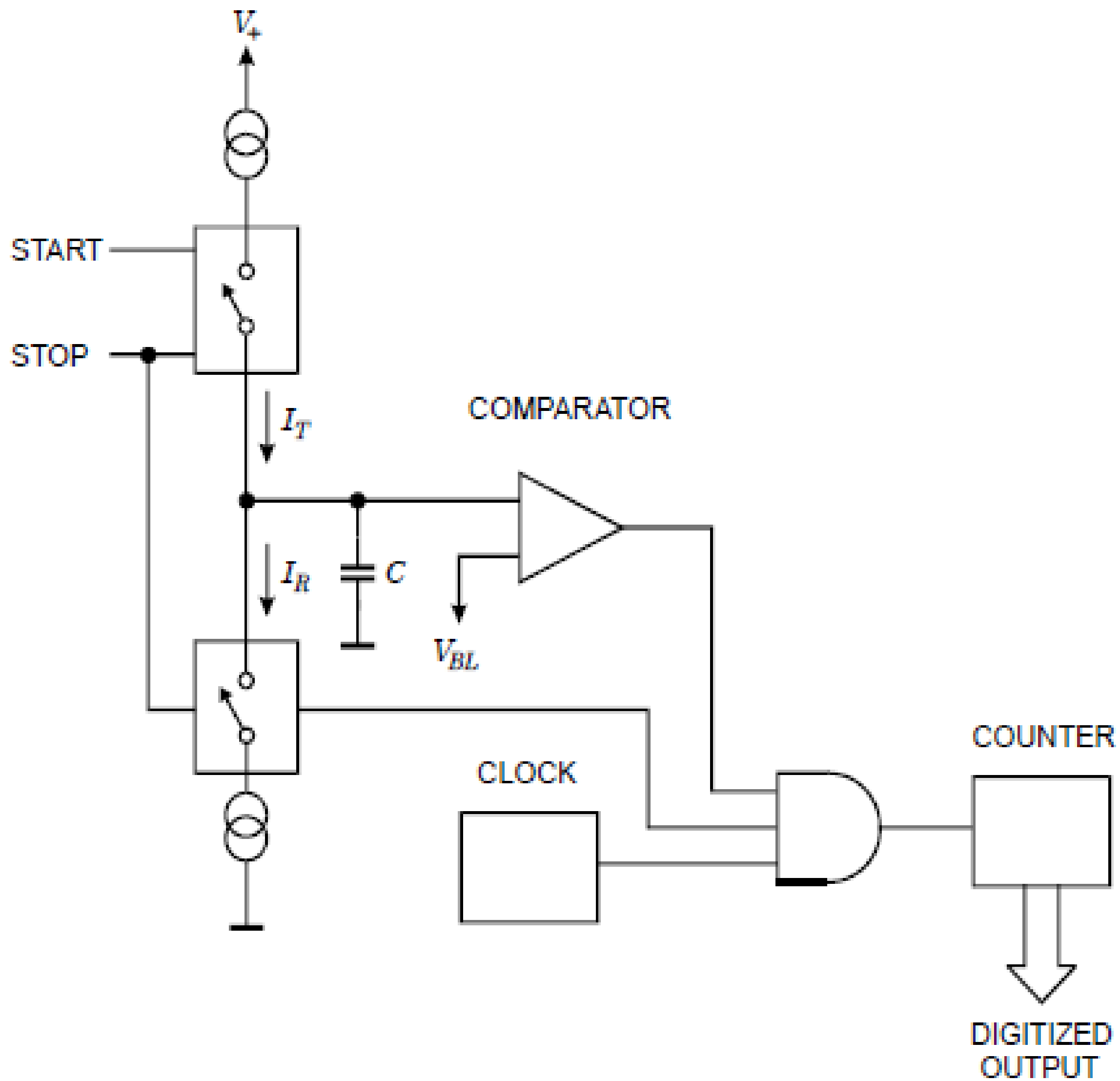
Use Wilkinson ADC with smaller discharge current to digitize voltage.

Drawbacks: No multi-hit capability

Deadtime

Advantages: High resolution (~ps)

Excellent differential linearity



Time to Amplitude Converters (TAC or TPHC)

The TAC is a unite which converts a time period between two logic pulses into an output pulse whose height is proportional to this duration. This pulse may then be analyzed by a multichannel analyzer to give a spectrum as a function of the time interval. An ADC may also be placed after TAC to digitize the output pulse.

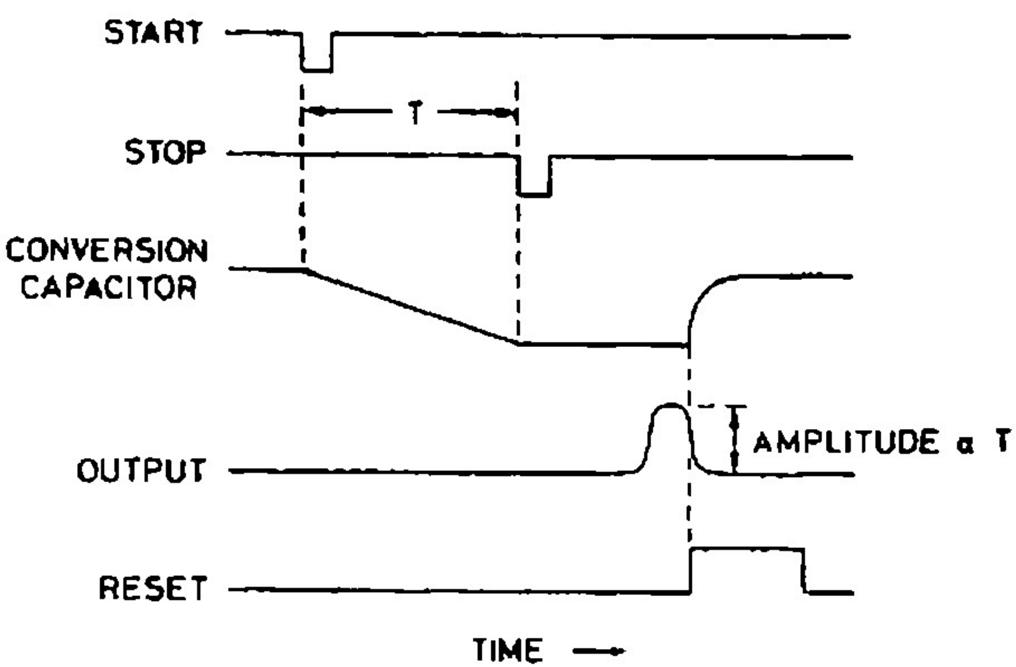
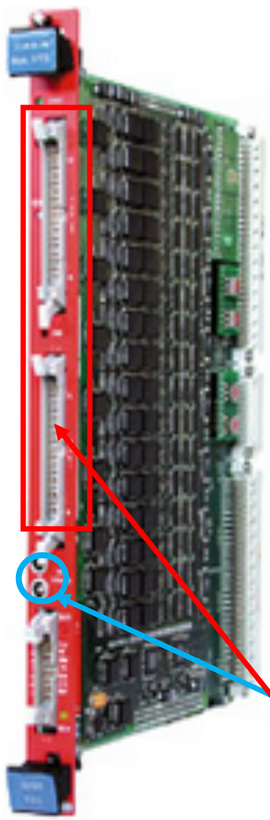


Fig. 14.22. Operation of a time-to-amplitude converter (TAC)

V775

32 Channel Multievent TDC (35÷300 ps)



- High channel density
- 12 bit resolution
- 5.7 μs / 32 ch conversion time
- 600 ns fast clear time
- Zero and overflow suppression for each channel
- ±0.1% integral non linearity
- ±1.5% differential non linearity
- 32 event buffer memory
- BLT32/MBLT64/CBLT32/CBLT64 data transfer
- Multicast commands

Common start

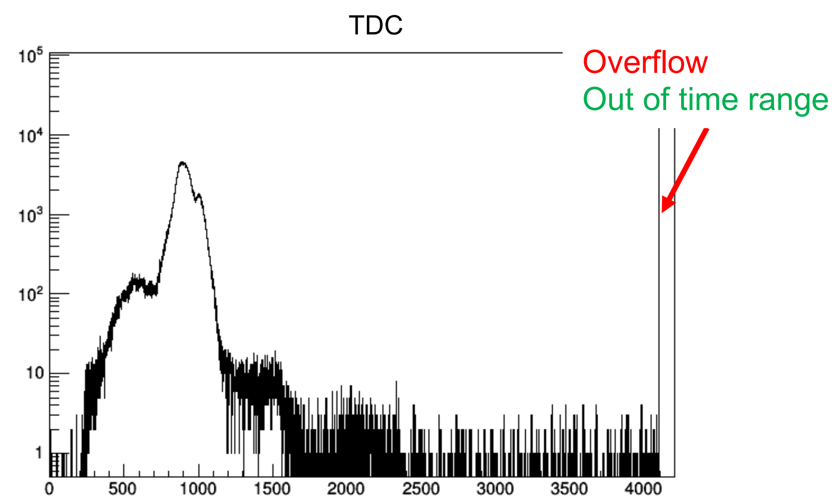
Individual stops

12-bit TDC

0-4094：$0 \leq t \leq$ full range

4095： t > full range (out of range)

Time resolution=full range/4095 ch.



Timing resolution in Time difference measurement

in an actual timing system, there are always some sources of uncertainty or error in the determination of the arrival times of detector pulses, and thus even a prompt coincidence time spectrum shows a distribution rather than a delta function.

If we assume that the independent timing uncertainties in two branches are of Gaussian type, which is usually the case, then the time difference spectrum will display a Gaussian distribution.

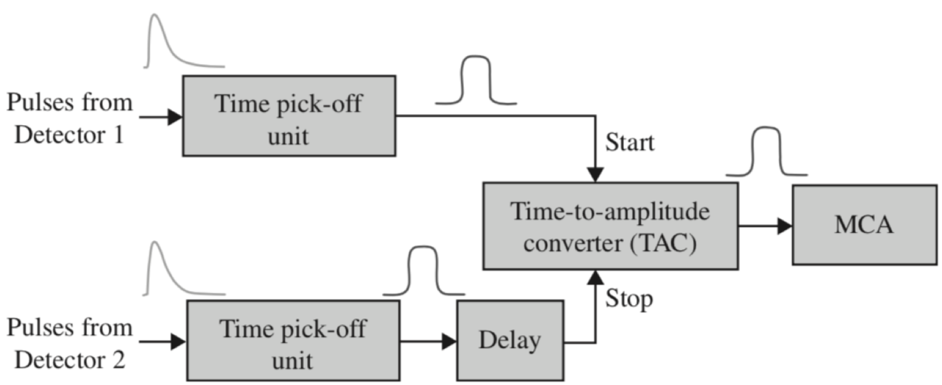
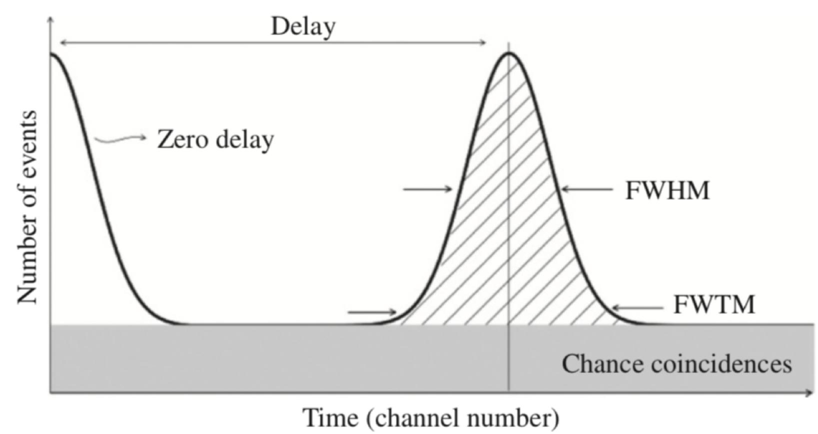


Figure 6.2 A simple block diagram of a typical time spectrometer.



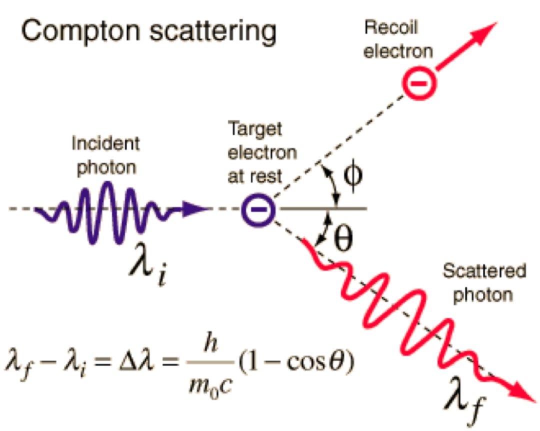
FWHM of the time distribution is called the time resolution of the system.
The FWHM of the time spectrum can be written as $FWHM = FWHM_1^2 + FWHM_2^2$
where $FWHM_1$ and $FWHM_2$ are the contributions of each branch to the time resolution of the system.

Coincidence method

An extremely important technique in nuclear and particle physics. Like pulse height selection, coincidence in time between two or more events serves as a very powerful criterion for distinguishing reactions.

Walther Bothe shared the Nobel Prize for Physics in 1954 "for his discovery of the method of coincidence and the discoveries subsequently made by it".
Bruno Rossi invented the electronic coincidence circuit for implementing the coincidence method.

Waltber W.G. Bothe
1891~1957



The experiment was designed to answer the questions as follows: is it exactly a scatter quantum and a recoil electron that are simultaneously emitted in the elementary process, or is there merely a statistical relationship between the two?

http://www.nobelprize.org/nobel_prizes/physics/laureates/1954/bothe-lecture.html

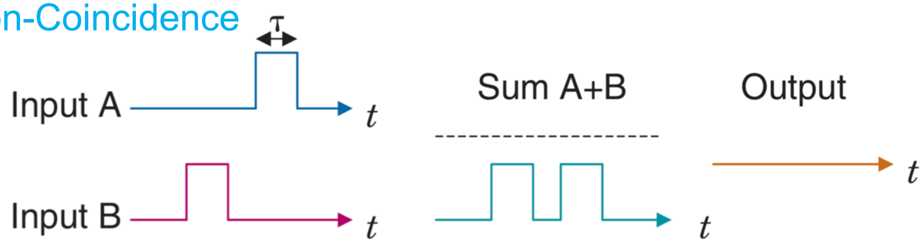
Coincidence circuit (logic AND)

In physics, a coincidence circuit is an electronic device with one output and two (or more) inputs.

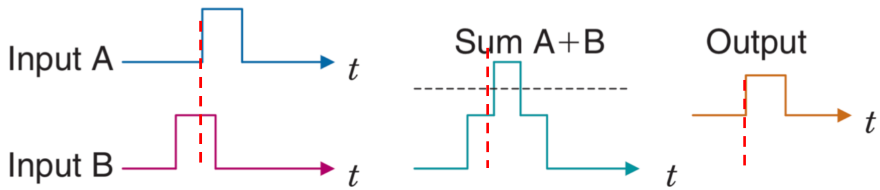
A coincidence output is produced if any part of the two incoming signals overlap. All pulses arriving within a time equal to the sum of their widths are registered as coincident.

Coincidence circuits are widely used in particle physics experiments and in other areas of science and technology.

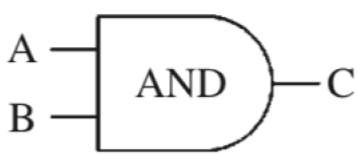
Non-Coincidence



Coincidence

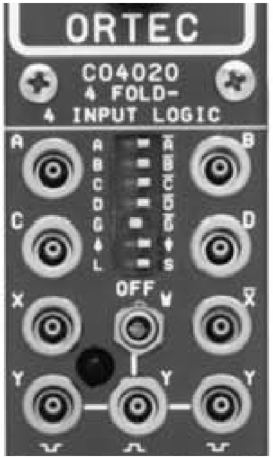


The summing method for determining the coincidence of two signals. The pulse are first summed and then sent through a discriminator set at a level just below twice the logic signal amplitude.

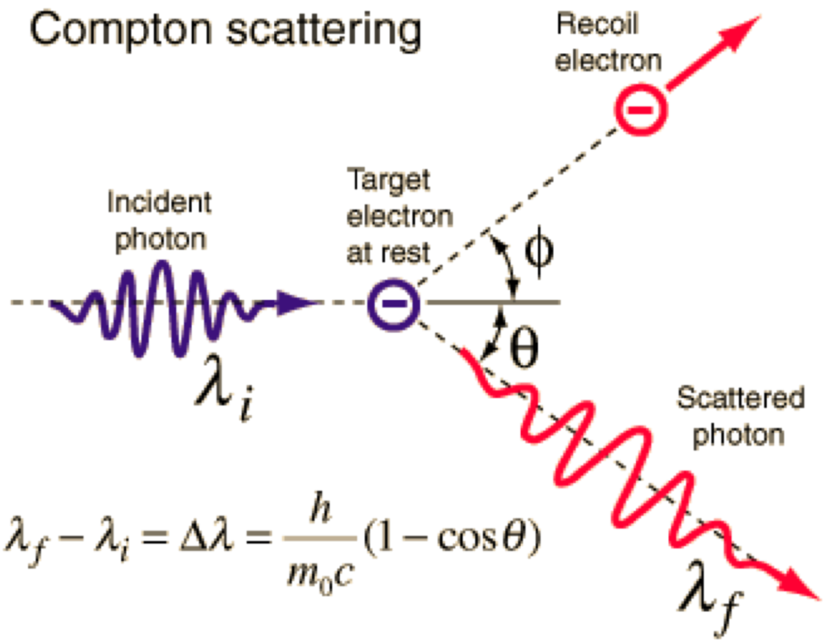


Coincidence=A*B

Anti-coincidence=$\overline{A*B}$



Compton scattering

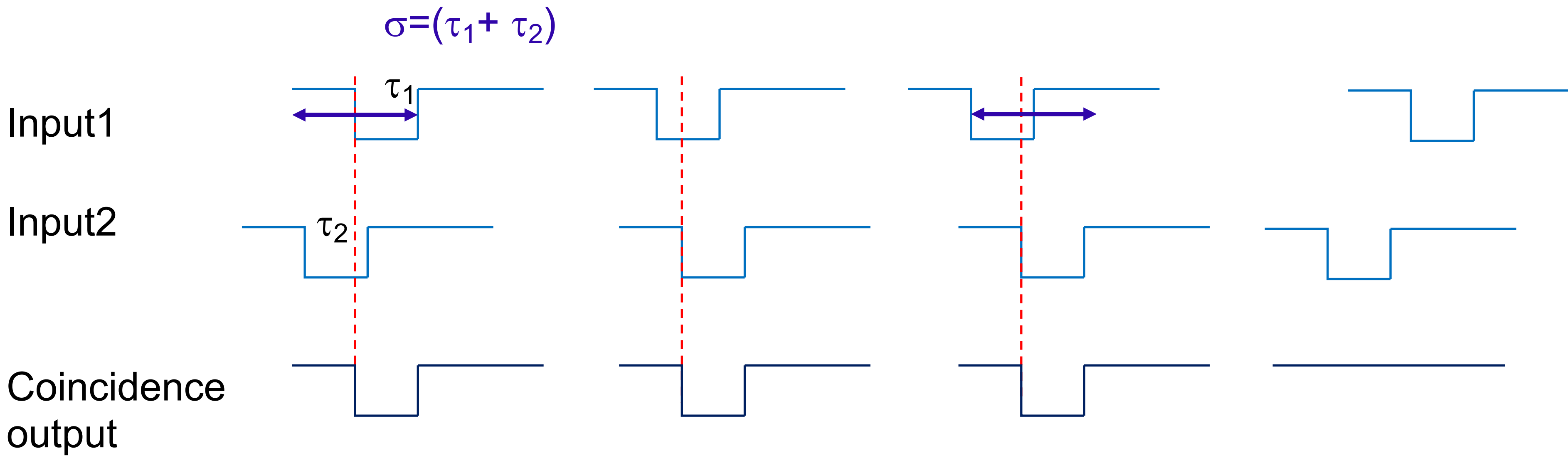


$$\lambda_f - \lambda_i = \Delta\lambda = \frac{h}{m_0 c}(1 - \cos\theta)$$

The final result we obtained was that systematic coincidences do indeed occur with the frequency that could be estimated from the experimental geometry and the response probabilities of the counters on the assumption that, in each elementary Compton process, a scatter quantum and a recoil electron are generated ***simultaneously***.

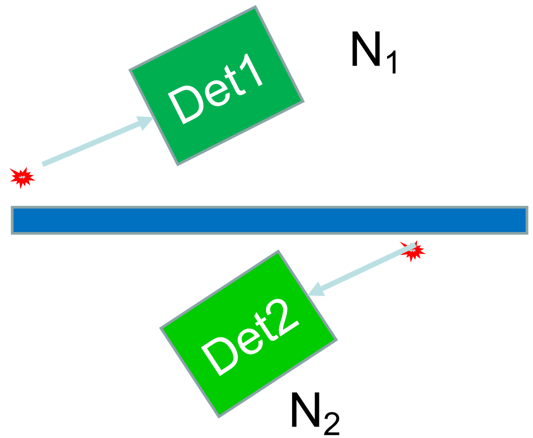
For time spectroscopy, the coincidence time unit performs the equivalent function and selects from all intervals only those for which the time difference between inputs is less than a circuit parameter, known as the resolving time.

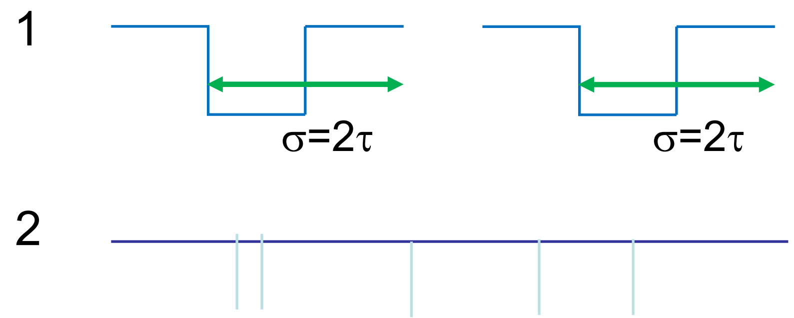
coincidental resolving time σ：符合分辨时间



Accidental coincidences

Uncorrelated two inputs at rates N_1 and N_2

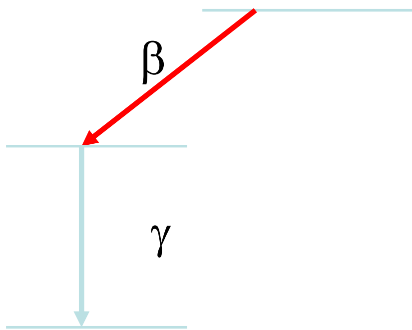


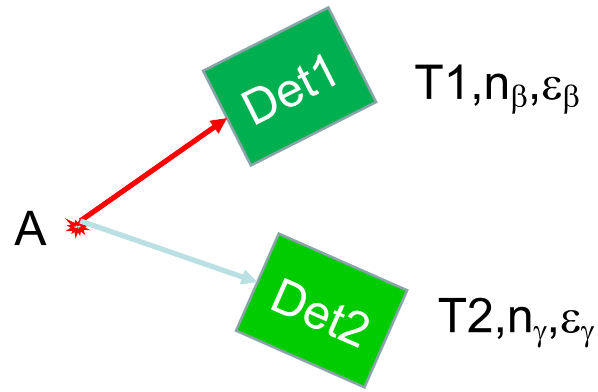


Suppose N_1 and N_2 are the singles counting rates for branches 1 and 2 respectively and σ is the resolution. Since any overlap in these pulses produces a coincidence, this means that the signals need only be within a period σ of each other in order to trigger the module. Assuming a constant singles rate, then, for each signal which arrives from branch 1, there will be $N_2\,\sigma$ pulses from branch 2 which fall into this allowable time period. Since there are N_1 pulses/unit time in branch 1, the total number of accidentals per unit time will be

$$N_r=2\tau N_1 N_2$$

For i uncorrelated inputs: $N_r = i\,\tau^{i-1} N_1 N_2 \,\ldots\ldots N_i$





Activity of the source: A

$$n_\beta = A * \varepsilon_\beta$$

$$n_\gamma = A * \varepsilon_\gamma$$

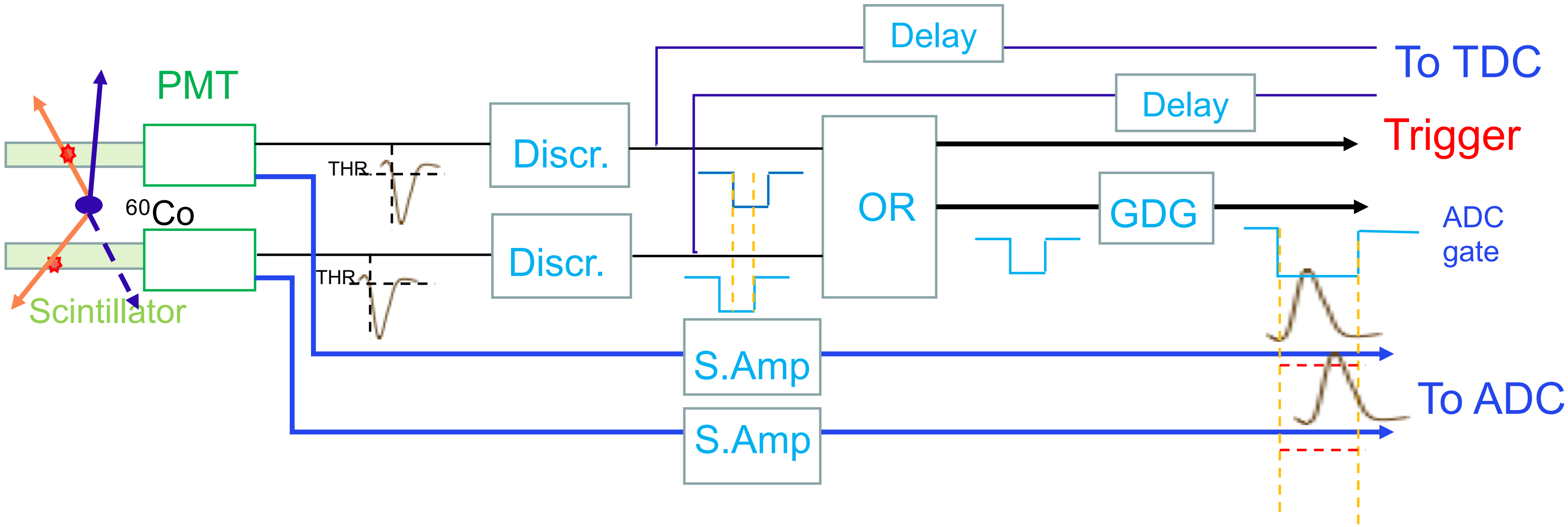
True coincidence: $n_{\beta\gamma} = A * \varepsilon_\beta * \varepsilon_\gamma$

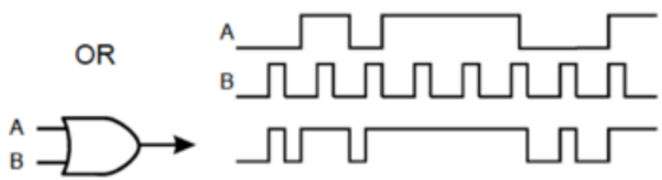
Accidental coincidence: $n'_{\beta\gamma} = 2\tau * n_\beta * n_\gamma = 2\tau * A^2 * \varepsilon_\beta * \varepsilon_\gamma$

true-to-chance coincidence ratio $n_{\beta\gamma}/n'_{\beta\gamma} = \frac{1}{2\tau A}$

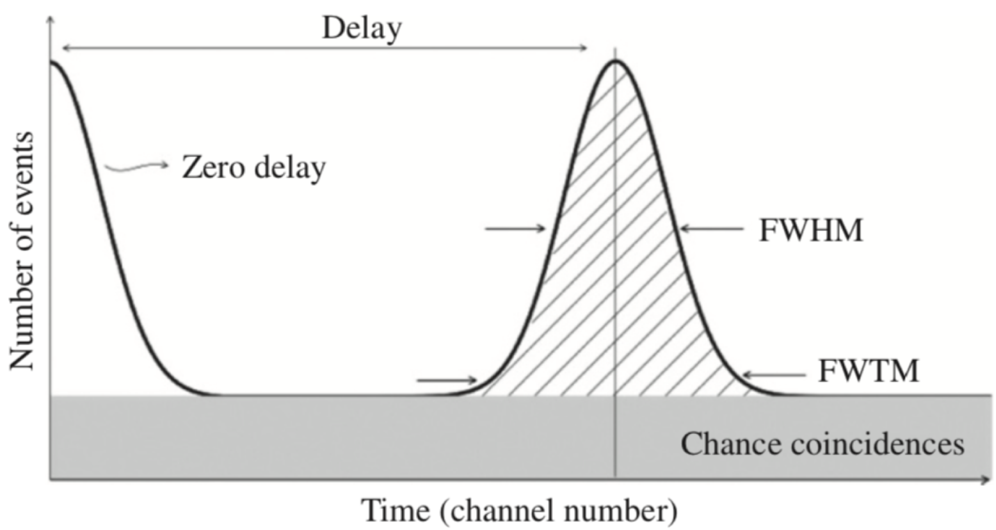
The measured coincidence rate n_{coin} is the sum of the true coincidence rate and the accidental coincidence rate: $n_{coin} = n_{\beta\gamma} + n'_{\beta\gamma}$

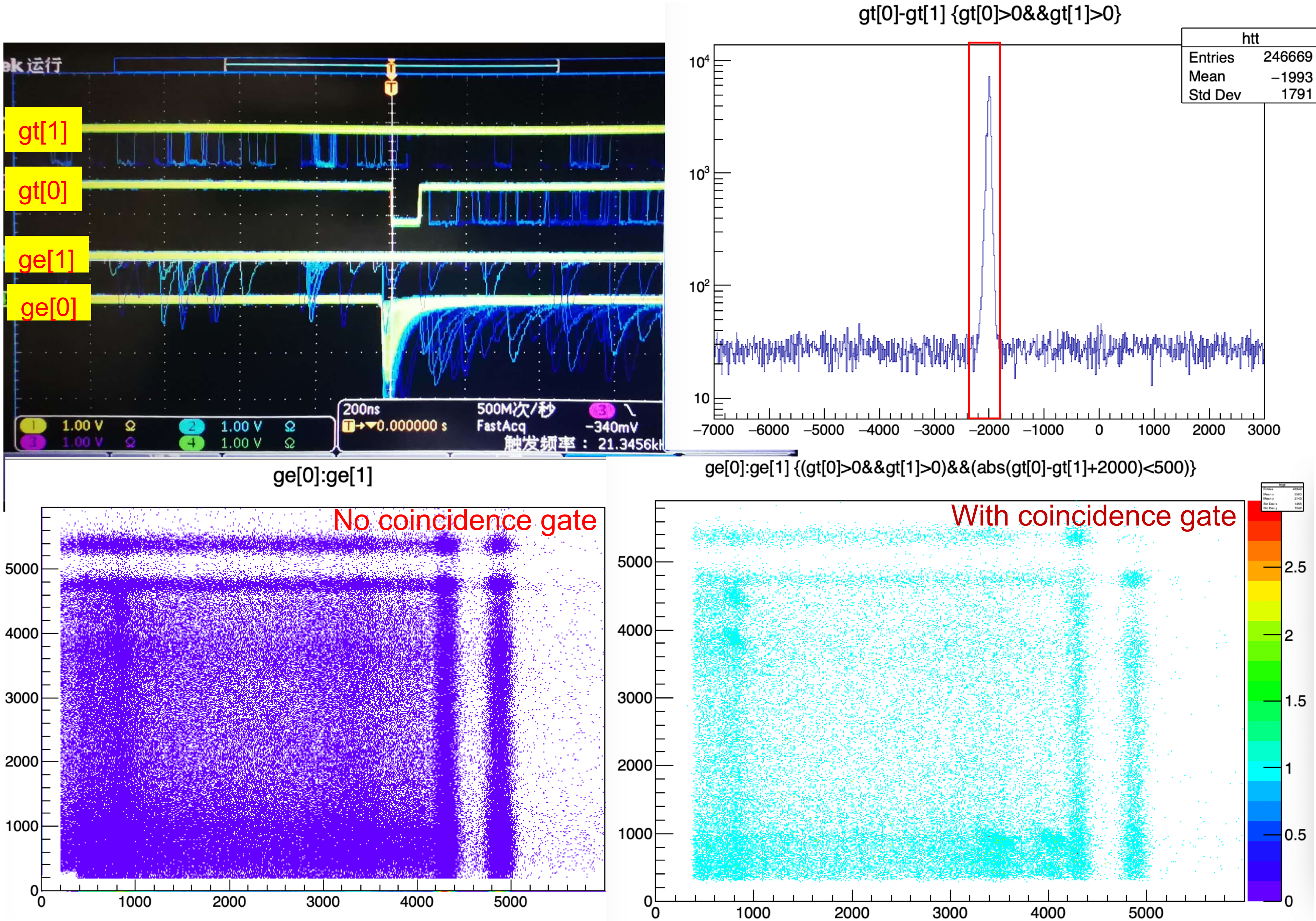
Ture and accidental coincidence



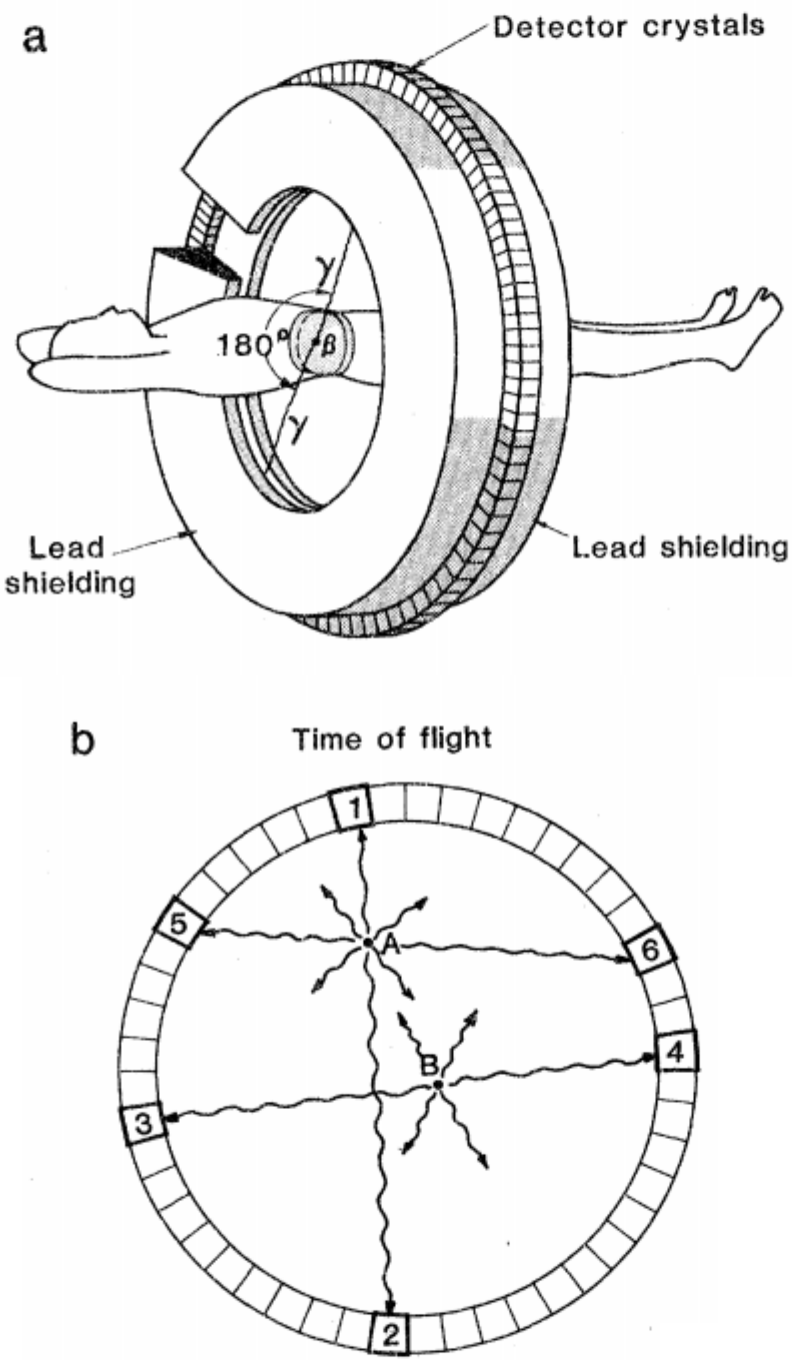


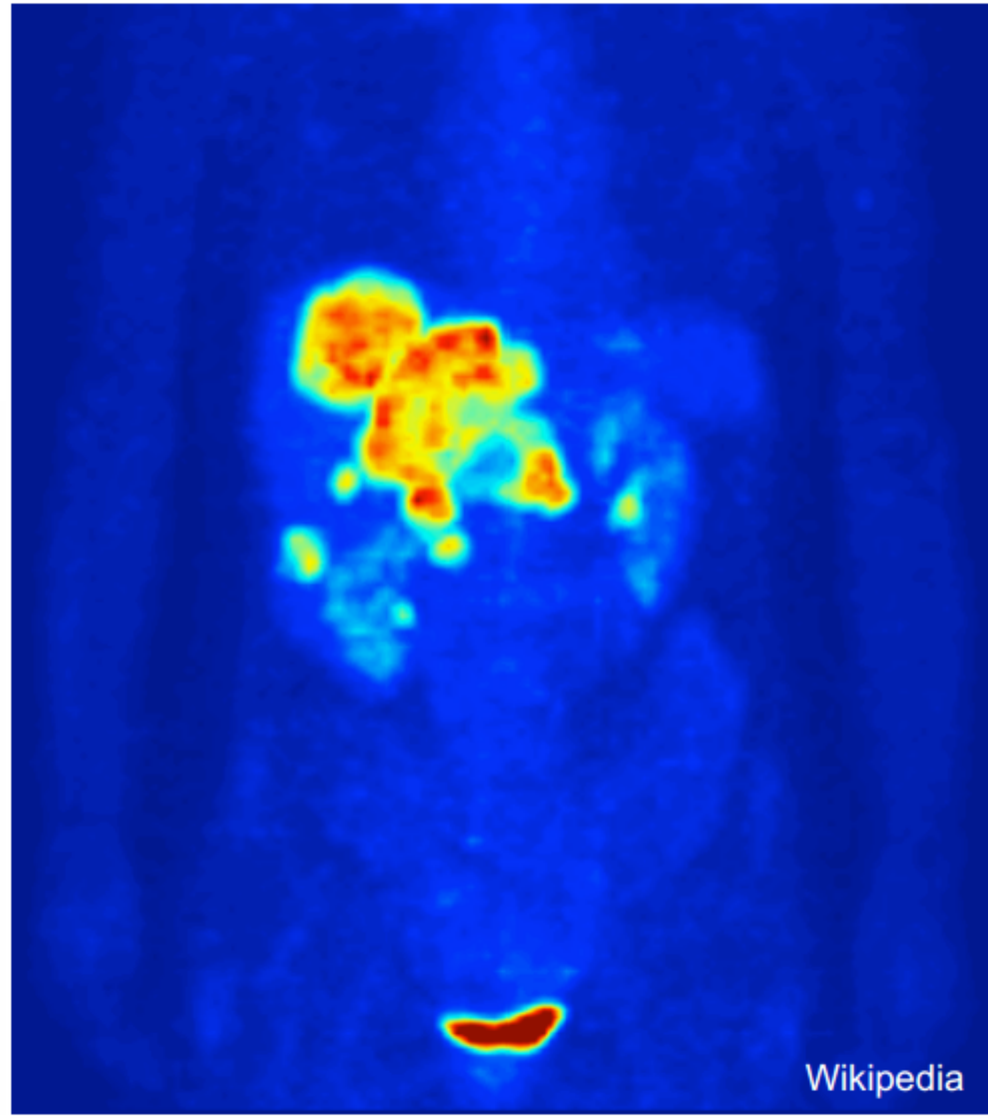
→ OR: Output high, when any input is high.





运行
gt[1]
gt[0]
ge[1]
ge[0]
200ns
500M次/秒
FastAcq
-340mV
1.00 V
1.00 V
1.00 V
1.00 V
gt[0]-gt[1] {gt[0]>0&&gt[1]>0}
htt
Entries 246669
Mean −1993
Std Dev 1791
ge[0]:ge[1]
No coincidence gate
ge[0]:ge[1] {(gt[0]>0&&gt[1]>0)&&(abs(gt[0]-gt[1]+2000)<500)}
With coincidence gate


a
Detector crystals
γ
180°
β
γ
Lead
shielding
Lead shielding
b
Time of flight
1
5
A
6
4
B
3
2


Wikipedia

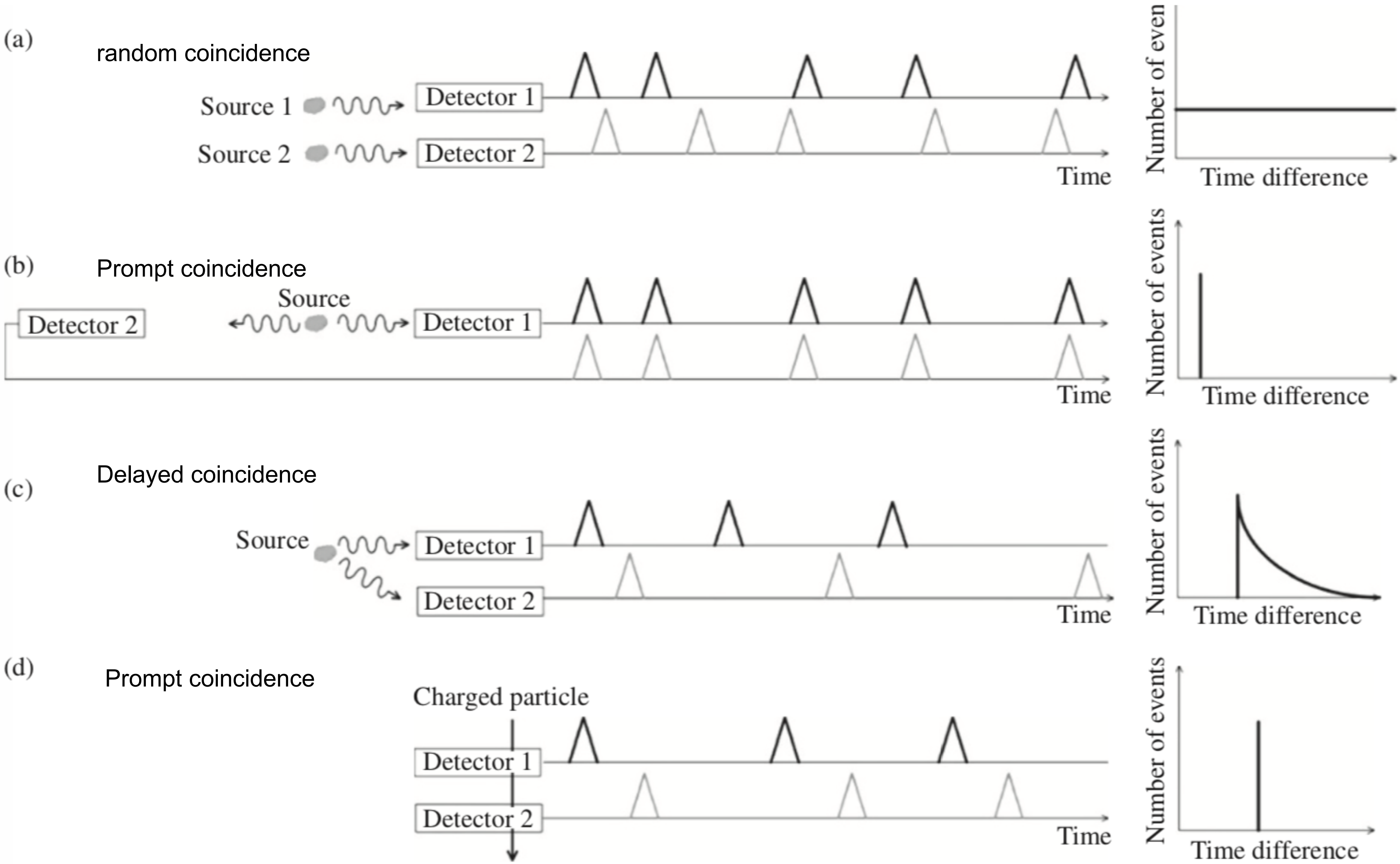


Figure 6.1 Some examples of time-difference measurements between two detectors and the corresponding distributions of time differences under ideal measurement conditions (see text for details).

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- 427A Delay Amplifier
- 460 Delay Line Amplifier
- 474 Timing Filter Amplifier
- 533 Dual Sum and Invert Amplifier
- 570 Amplifier
- 572A Amplifier