Front-End Electronics for Water Cherenkov Detectors

Hyper-K and E61 Case

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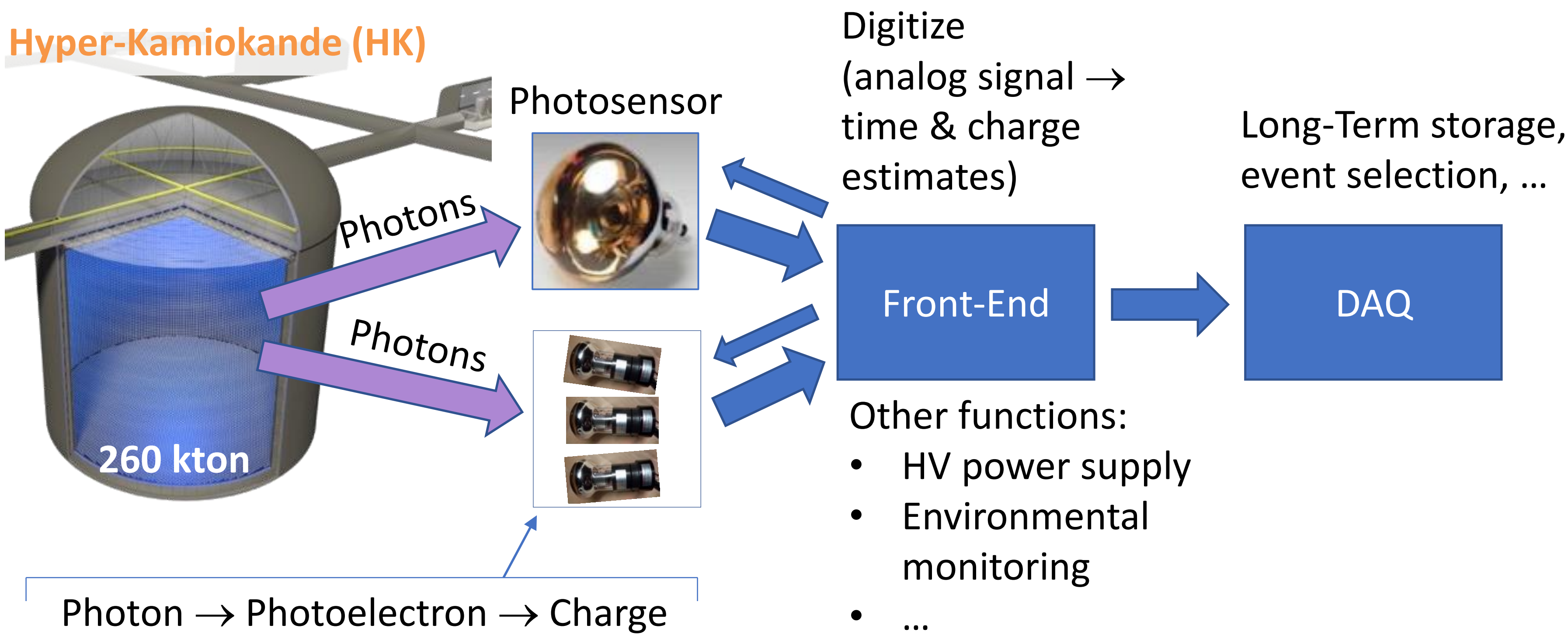
TMEX 2018

European Workshop on Water Cherenkov Precision Detectors
for Neutrino and Nucleon Decay Physics

2018-09-20, Warsaw, Poland

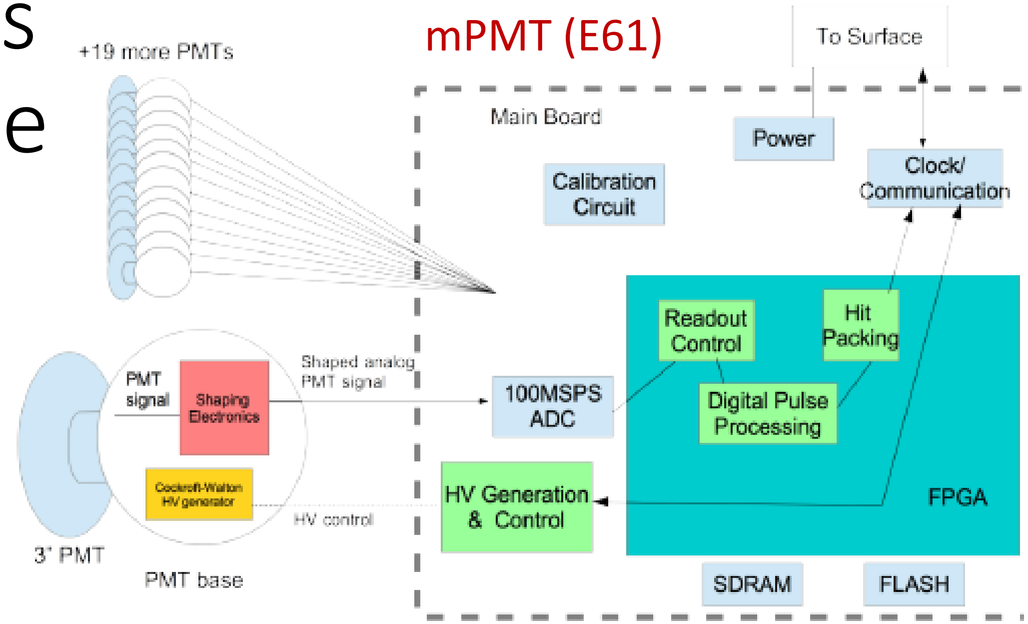
Introduction

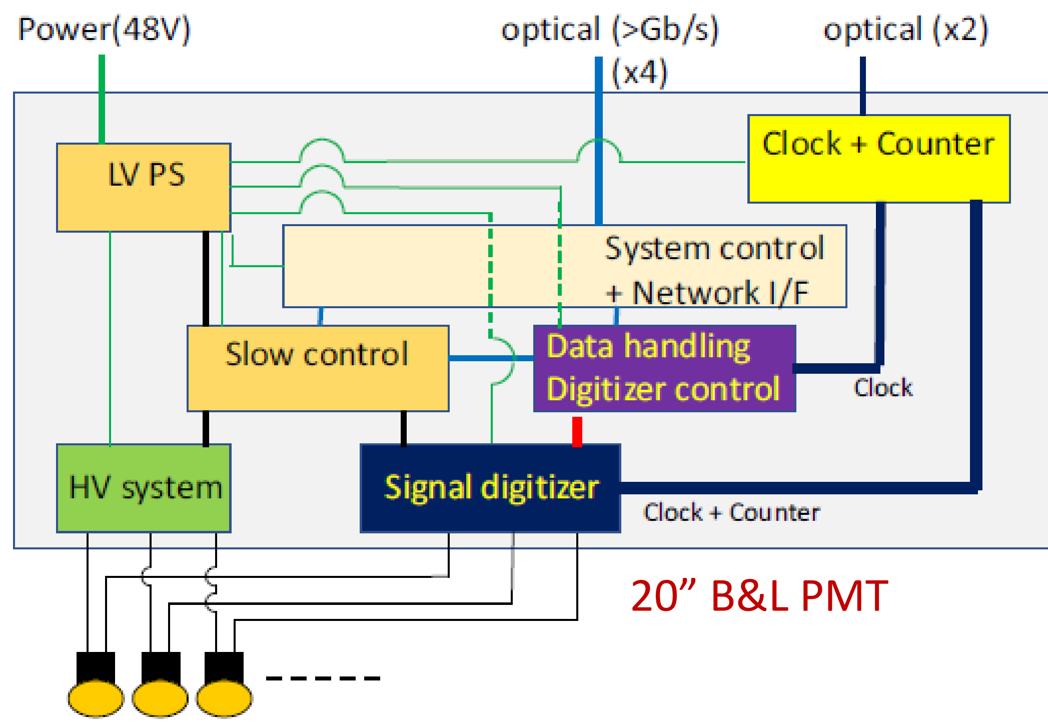


- Major task is to acquire analog signals from photosensors, digitize them (i.e. provide time and charge information) and send to downstream system for event selection and long term storage
- Tailored to the needs of the experiment
  - Requirements come from physics (general rule – better than photosensor)
- One should avoid overdesigning in order to keep costs under control

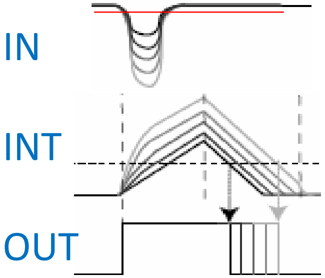
Front-End Electronics
Hyper-K and E61 case

- Self-triggering system
  - Digitize all photo-sensor signals above discriminator threshold
  - Send hit information to readout computers
  - Use software trigger for event selection & send them to offline system for storage
- Accurate clock synchronization and GPS
- Stable power supplies (photo-sensors, other systems)
- Front-end boards must provide high reliability (operation >20 years)
  - Underwater electronics - non-serviceable after filling detector with water
- Low power due to requirements of water circulation inside the tank (mainly HK requirement)

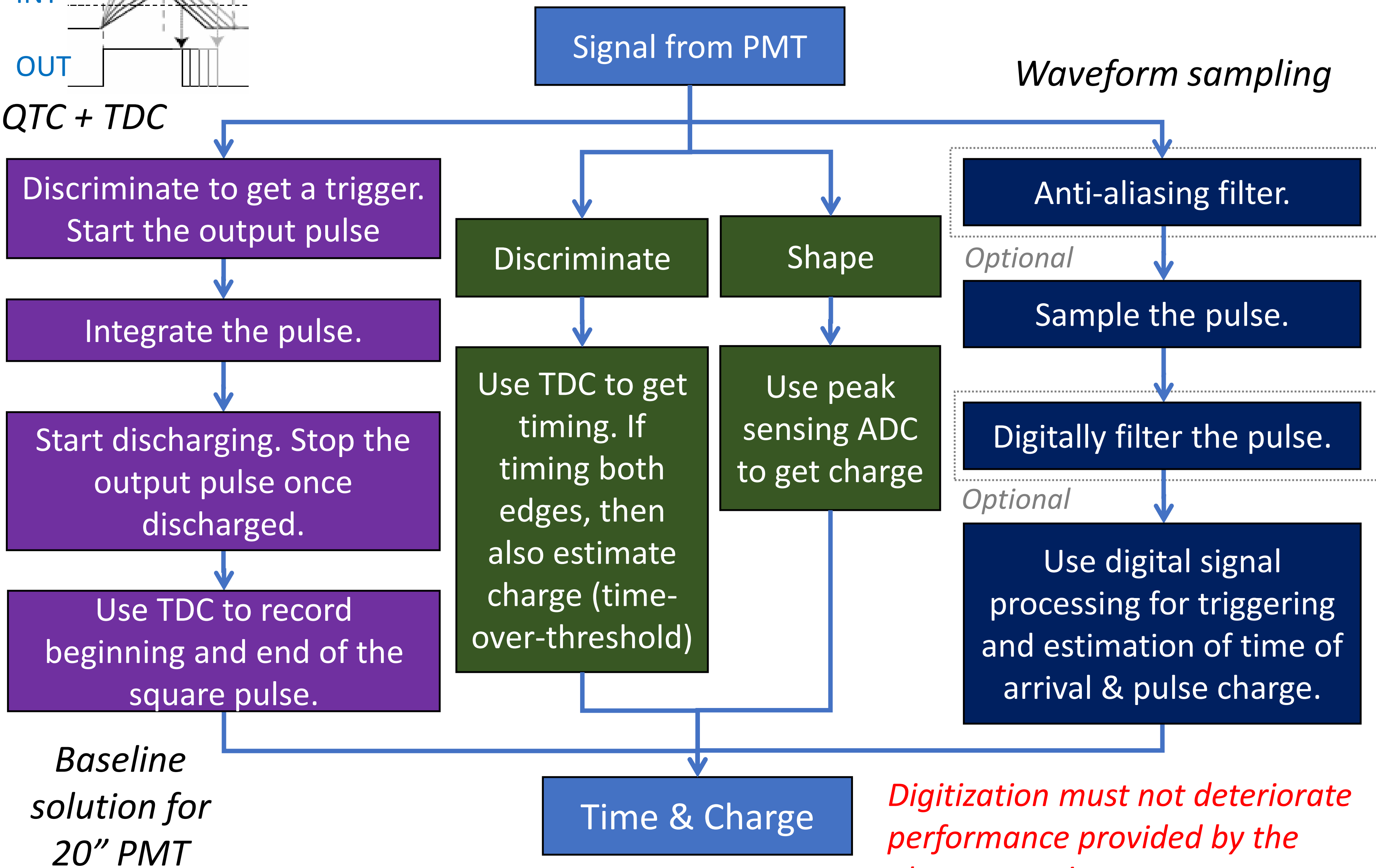




Digitization Options



QTC + TDC

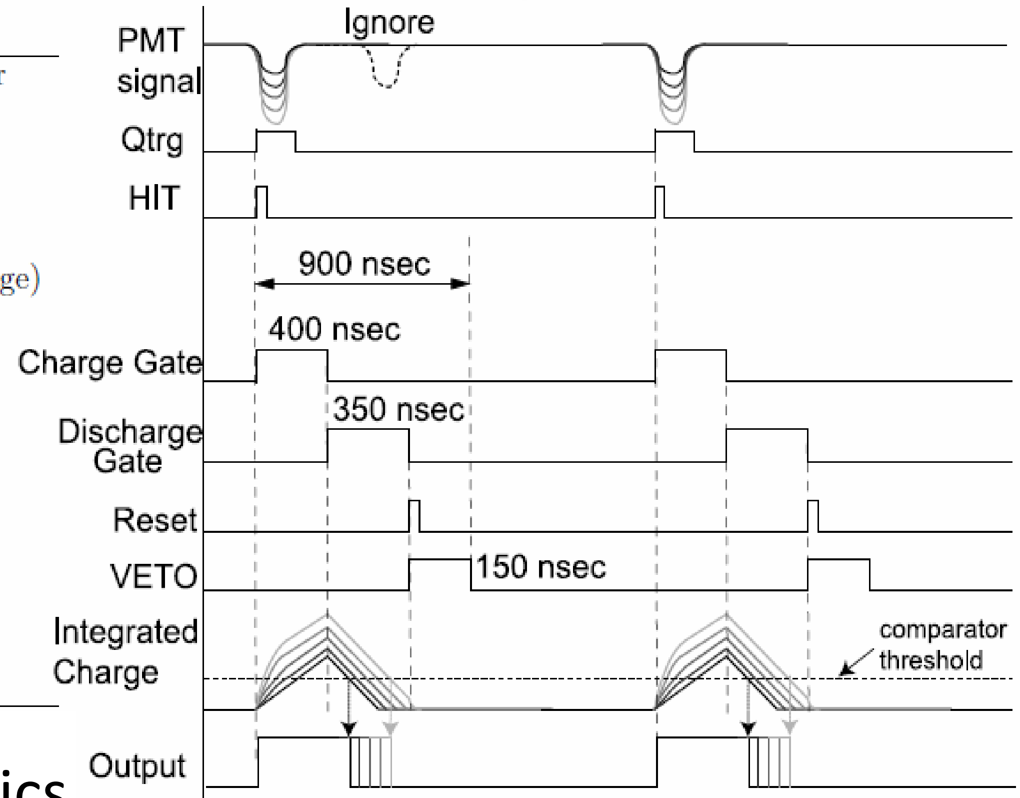


Digitization must not deteriorate performance provided by the photosensor!

Digitization – QTC + TDC

Table 1: Specification of the QTC

Type of trigger	self trigger by discriminator
Number of Input Channels	3
Processing Speed	$\sim$ 900 ns/cycle
Charge Integration Gate	400 ns
Number of Gains	3 (Ratio $1 : \frac{1}{7} : \frac{1}{49}$)
Discriminator Threshold	$-0.3 \sim -14$ mV (small range)
Charge Dynamic Range	$0.2 \sim 51$ pC (small)
	$1 \sim 357$ pC (medium)
	$5 \sim 2500$ pC (large)
Charge Resolution	$\sim$ 0.2 pC (small)
Integral (Non-)Linearity	$< \pm 1\%$
Timing Resolution	0.3 ns (2 pC, -3 mV)
	$<$ 0.1 ns ($>$ 100 pC)
Power Dissipation	$<$ 100 mW/ch
Process	0.35 μm CMOS
Package	100 pin CQFP

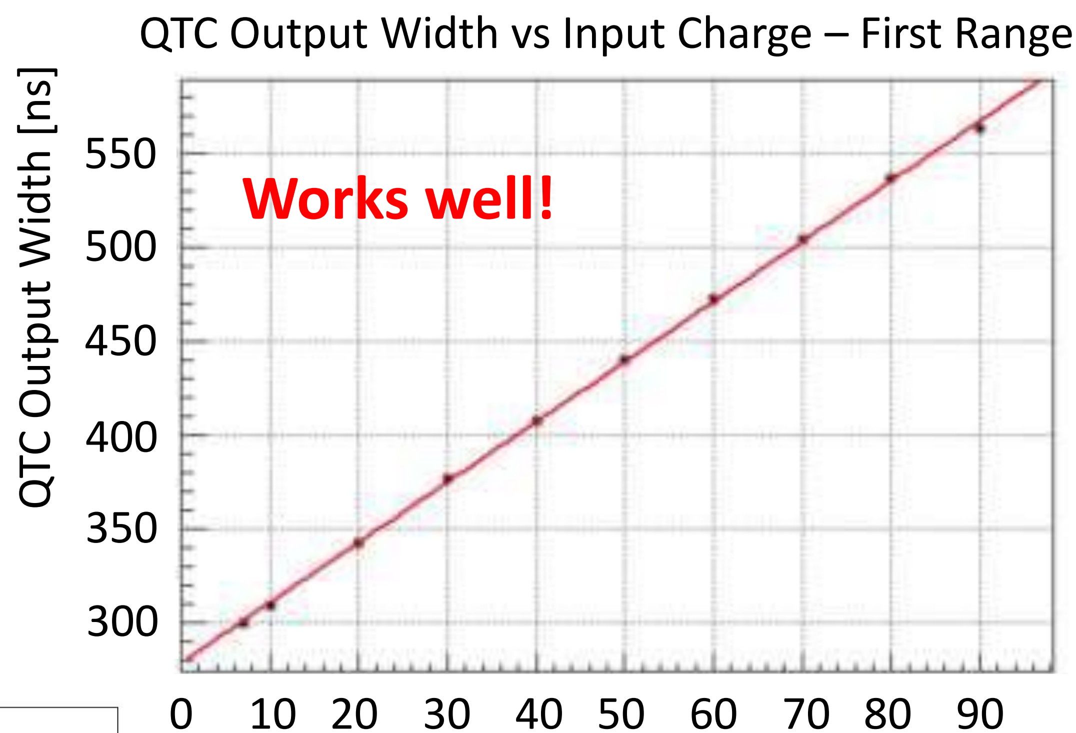


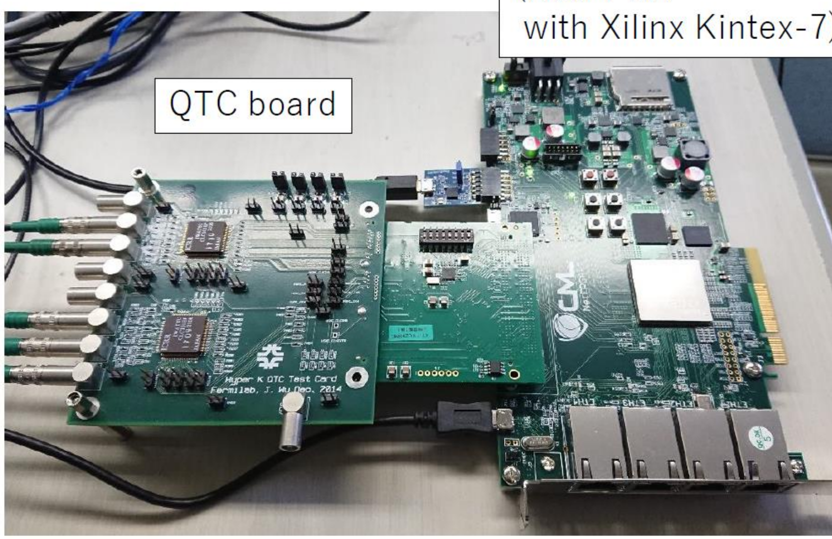
- Similar to SK-type electronics
  - Uses custom built QTC ASIC and external TDC
- 3 QTC channels per one PMT channel
  - Necessary to cover wide dynamic range (1250 p.e.)
- Process rule is CMOS 0.35 μm – still possible to manufacture the same chip. However, TDC chip is no longer available.

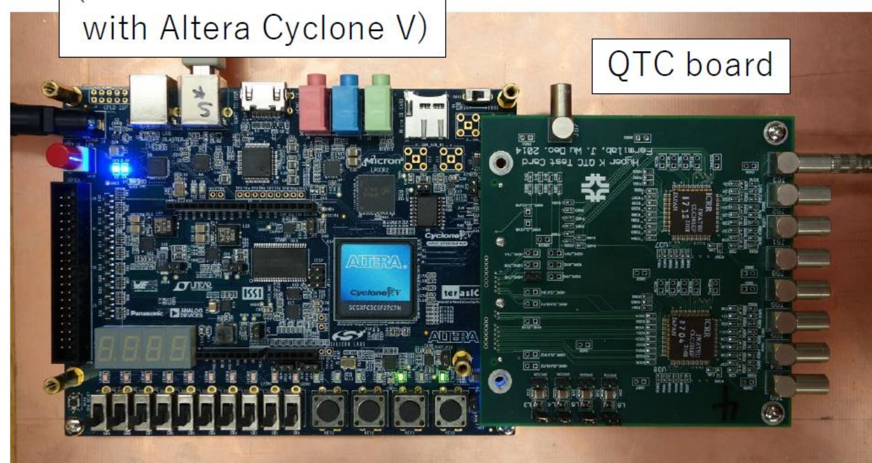
Works well in SK!
(QBEE board)

QTC + TDC

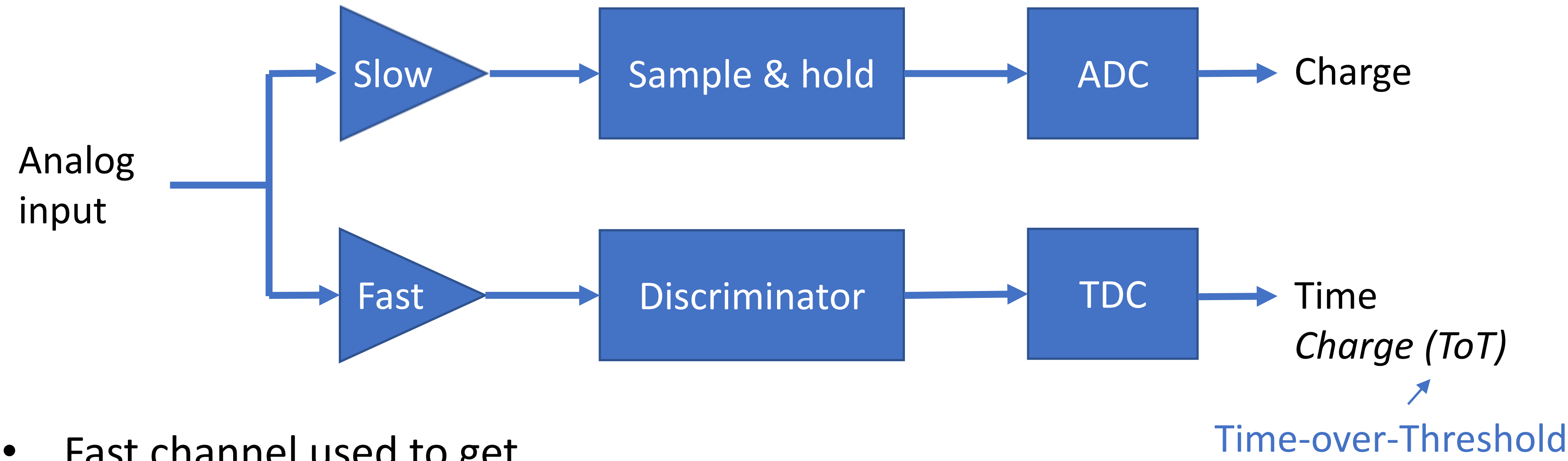
- Both QTC and TDC work well
- TDC performance: *(Y. Kataoka)*
  - DNL: ≈ 50 ps to 60 ps (σ)
  - INL: ≈ 60 ps to 70 ps (σ)
- Charge linearity (QTC + TDC):
  - +/- 1% up to 2000 p.e. when using low-gain, mid-gain and high-gain channels



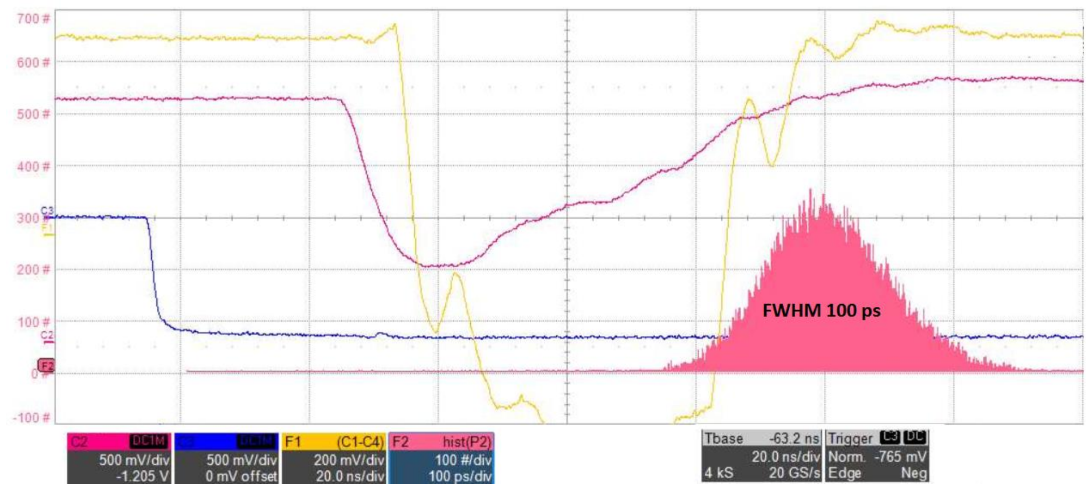




TDC + Peak Sensing ADC



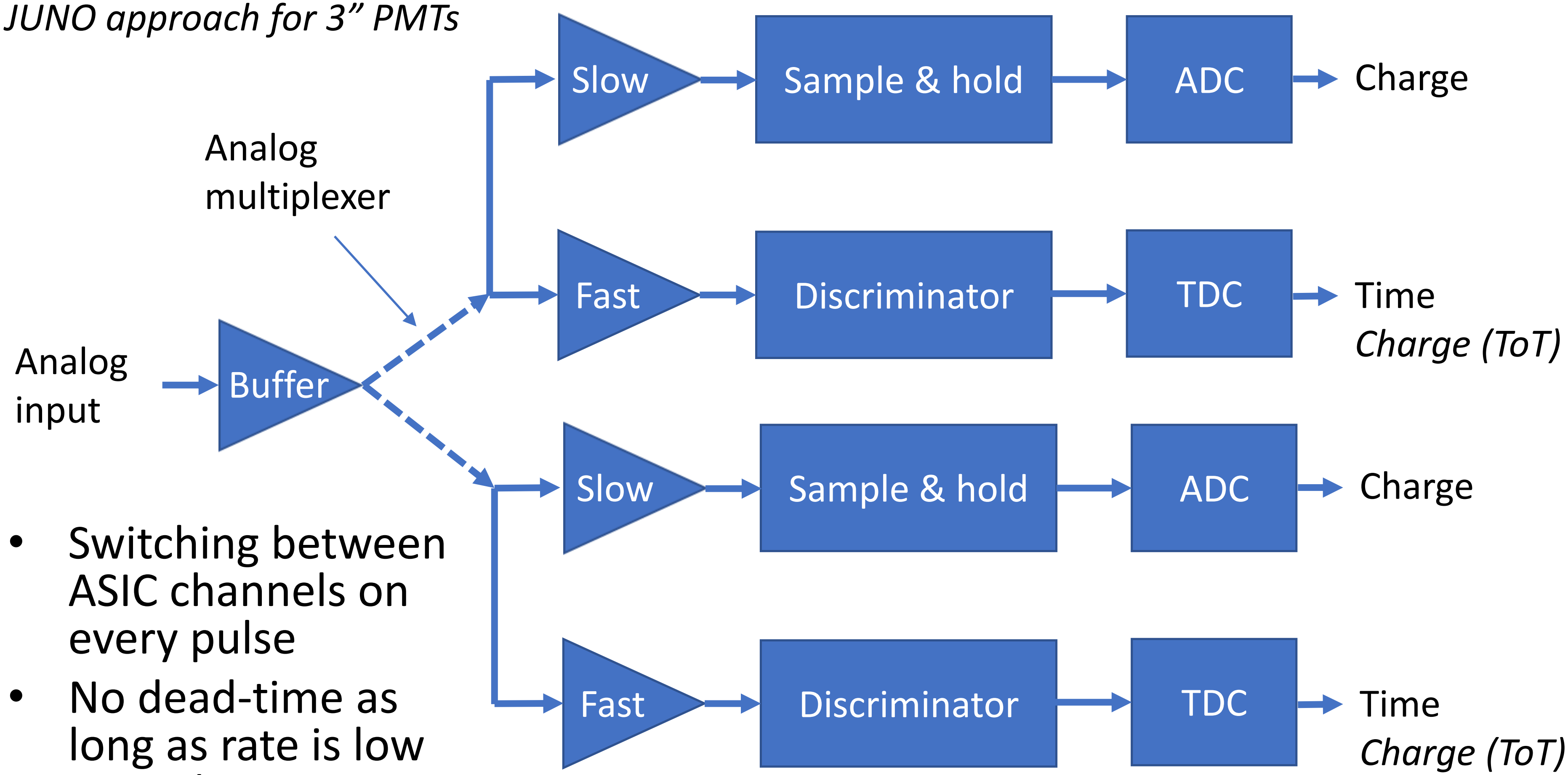
- Fast channel used to get timing
- Slow channel used to get charge
  - Significant bandwidth cut (i.e. low-pass with low frequency cut-off) → deterministic pulse shape regardless of input pulse shape
  - Direct correlation of amplitude with charge
- Introduces dead time



A. Evangelisti

TDC/Peak Sensing ADC – Ping-Pong

JUNO approach for 3" PMTs

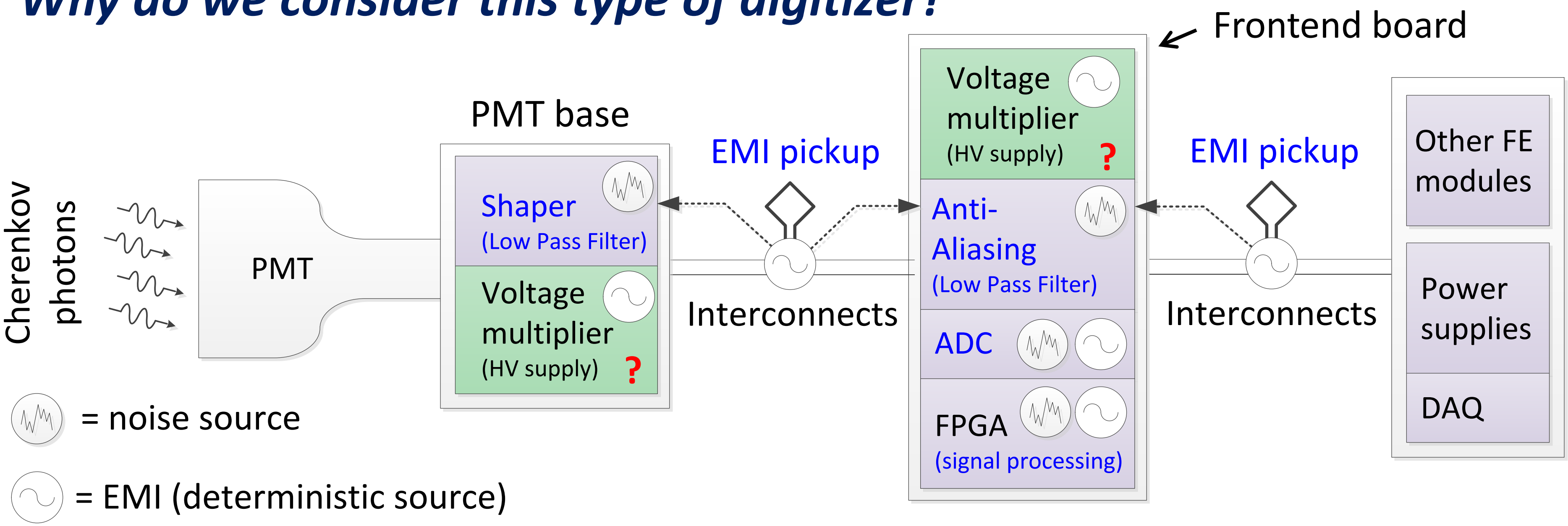


- Switching between ASIC channels on every pulse
- No dead-time as long as rate is low enough
- Feasible option when using ASIC (CATIROC)

Some capability to get charge even if two slow channels 'occupied', but resolution will deteriorate

Digitization – Waveform Sampling

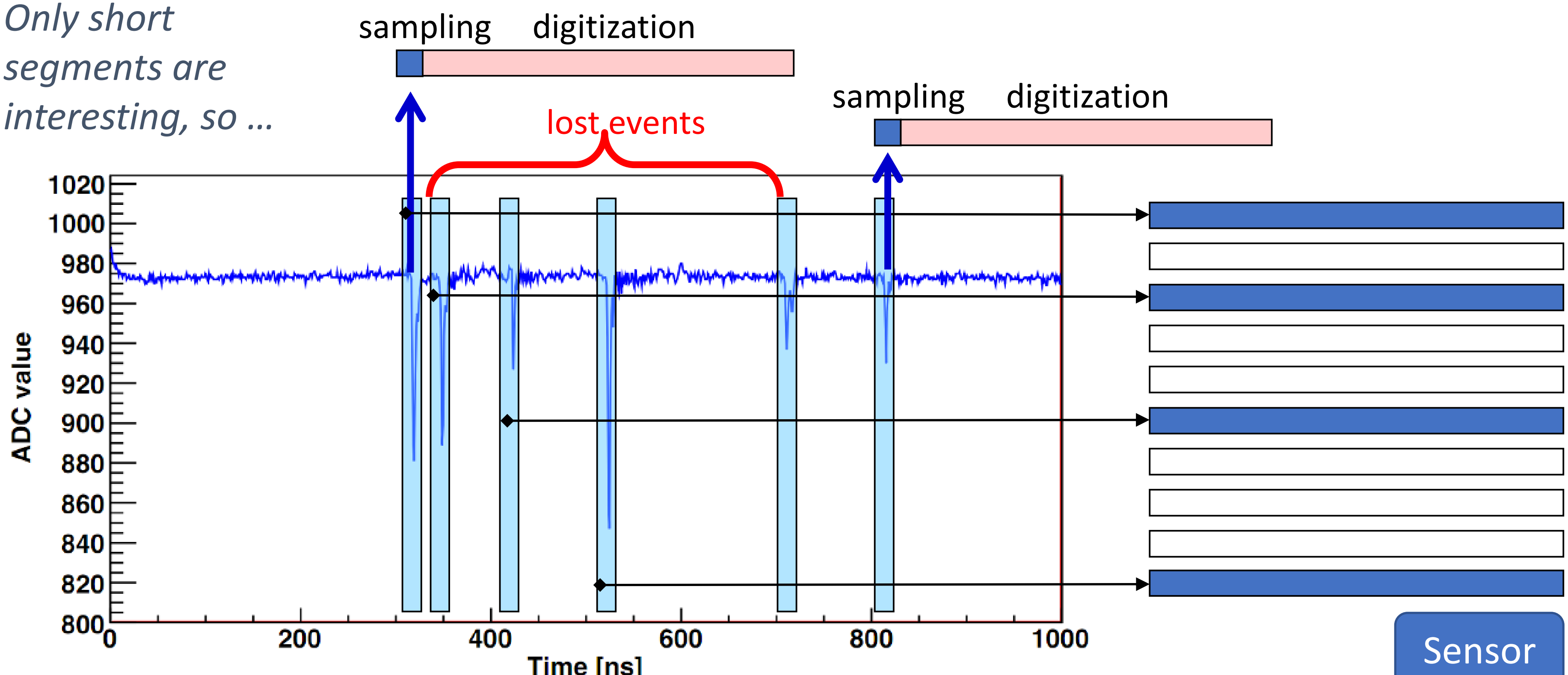
Why do we consider this type of digitizer?



- Possibility to implement completely dead-time free system.
  - Better ability to tag decay electrons that occur at short decay times and high muon energies.
  - E61 case – ability to disentangle in-bunch pile-up
- Pulse processing on-the-fly (i.e. send only time/charge – most of the time)
- Can subtract off periodic EMI by digital filters implemented in FPGA firmware.
- There is a price to pay: power consumption and cost (?).
  - We need to reduce both without affecting physics performance

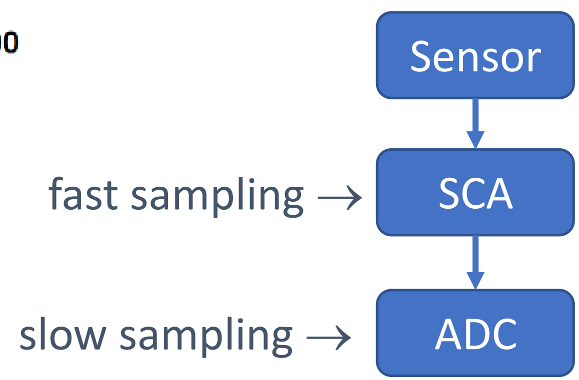
Lowering Power Consumption – Switched Capacitor Arrays (DRS4 example)

Only short segments are interesting, so …

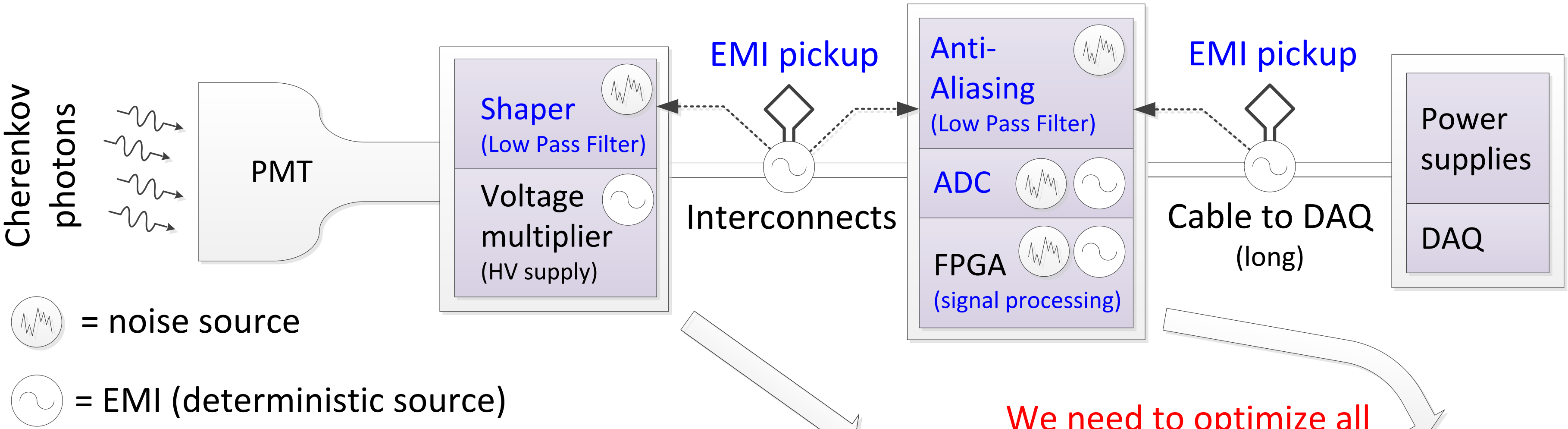


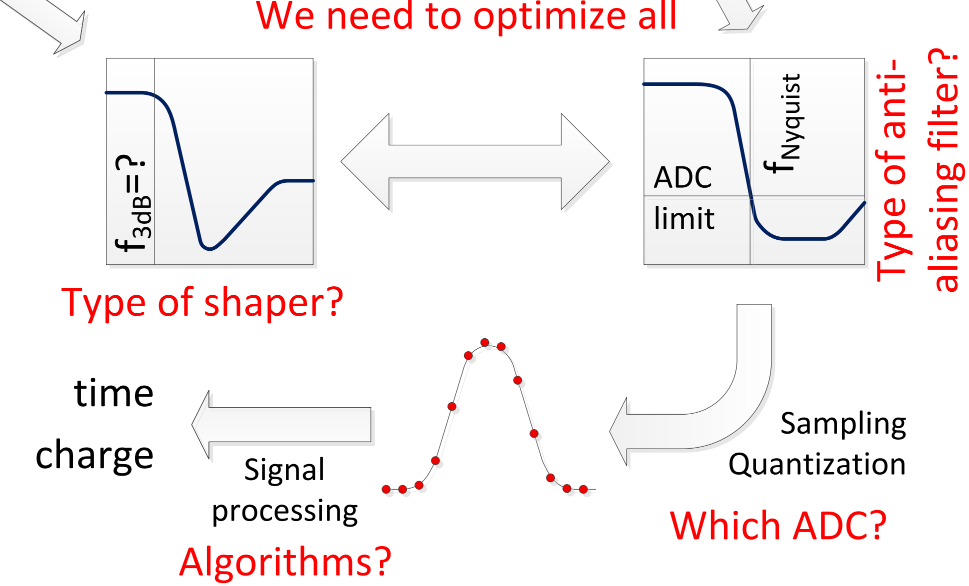
Avoiding dead time in capacitor arrays:

- Use chip with segmented memory
  - Latch only part of array, keep other parts active (DRS5 solution – not yet available)
- Use multiple arrays for single waveform



Lowering Power – Optimize Signal Chain





- Test various methods of time estimation:
  - Digital Constant Fraction Discriminator
  - Optimal Zero-Average FIR Filter
  - Matched Filter + Cross-correlation
  - Fits *(off-line processing only)*
- Develop and validate model of the full signal chain
  - Will allow exploration of various variants of shaper/ADC combinations without the need for many prototypes

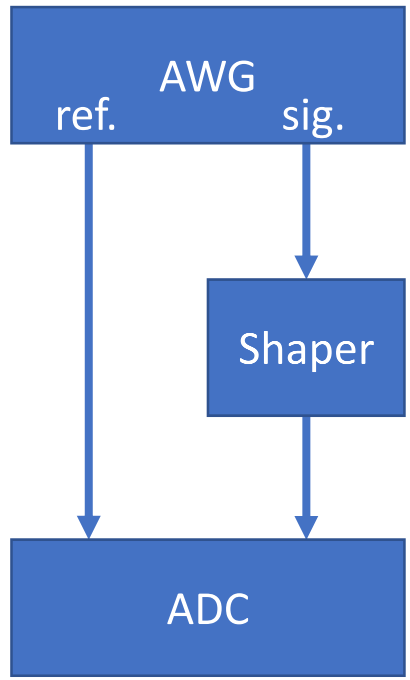
Sharing of signal processing between DAQ and FPGA?

Timing Resolution of Sampling Digitizers

PURPOSE OF THE STUDY:

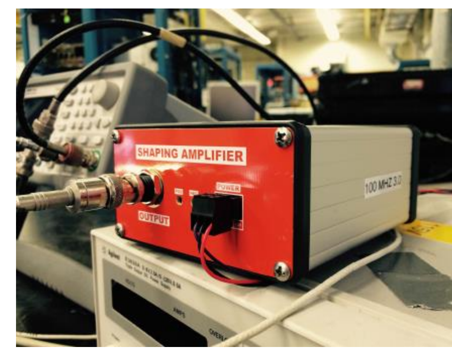
Determine how fast and how precise does a system needs to be to achieve given performance specs?

- Use AWG instead of PMT.
- Use large reference pulse (timing accuracy $\sigma \approx 10$ ps) and small, shaped signal pulse (1 mV ~ 100 mV).
- Apply signal processing methods and calculate time difference Δt between ref. and sig. channels.
- Repeat multiple times and compute RMS of Δt values.
- Two shapers:
  - 15 ns and 30 ns rise time (10% to 90%), 5-th order Bessel-type low-pass filters.



Agilent 33600A (1 GSPS/80 MHz)

Custom shapers



Commercial ADCs (CAEN)

DT5724 (100 MSPS/14b)

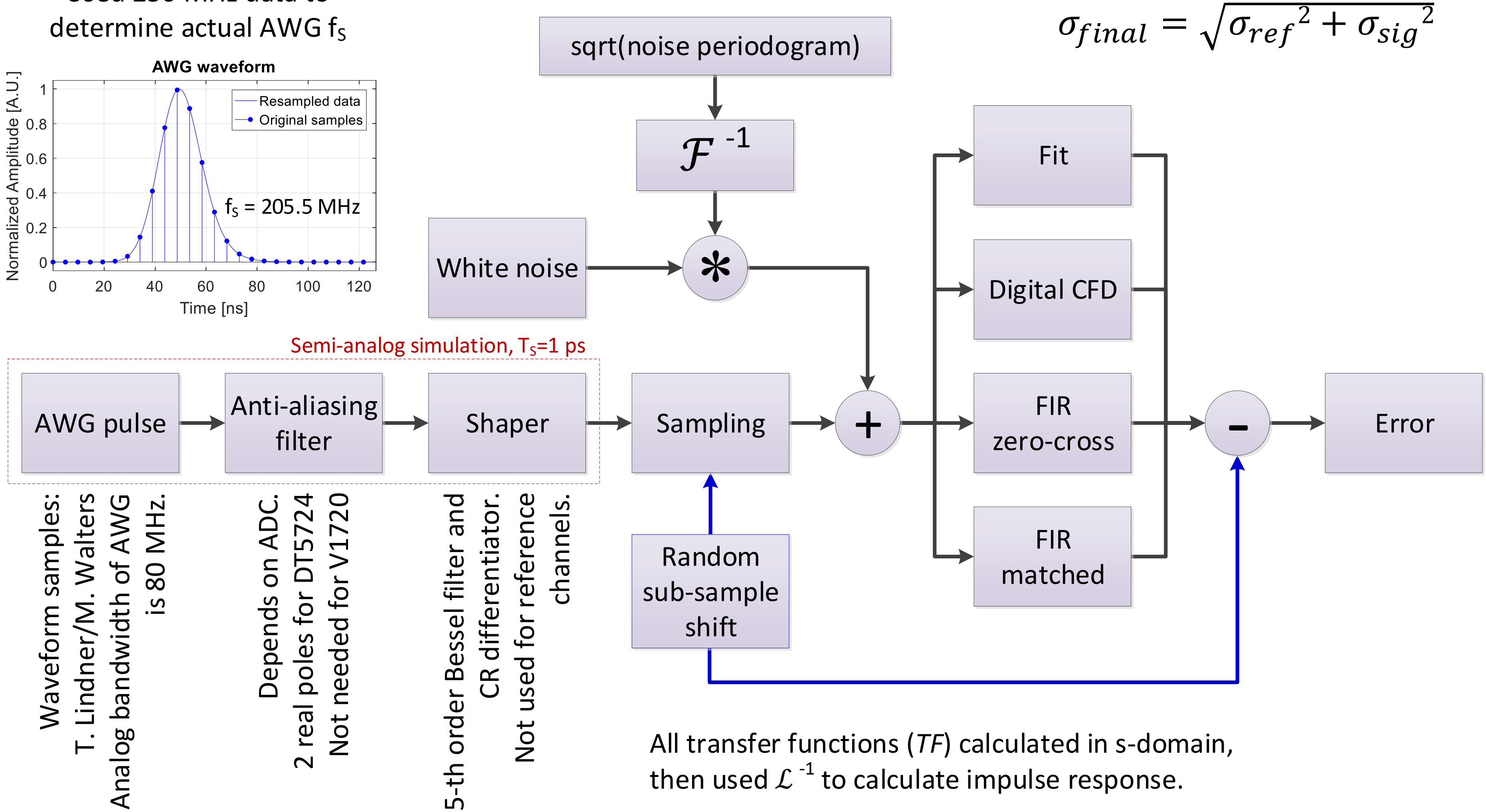
V1720 (250 MSPS/12b)

V1730 (500 MSPS/14b)

System Model (each channel)

Used 250 MHz data to determine actual AWG f_S

$$\sigma_{final} = \sqrt{\sigma_{ref}{}^2 + \sigma_{sig}{}^2}$$

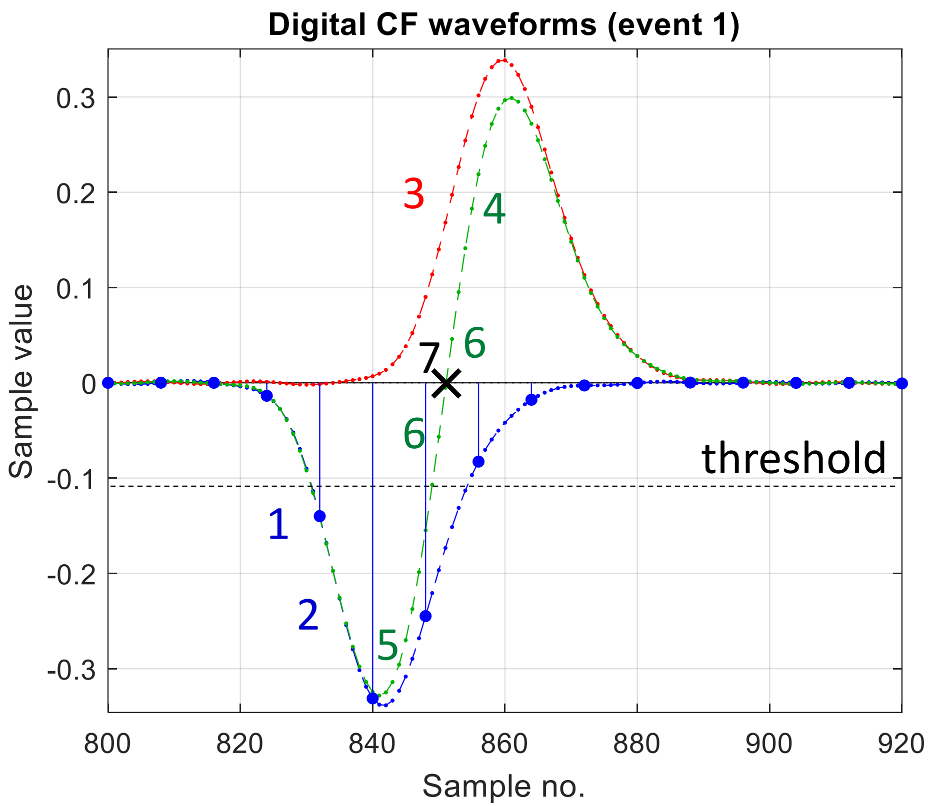


All transfer functions (*TF*) calculated in s-domain, then used $\mathcal{L}^{-1}$ to calculate impulse response.

Digital Constant Fraction Discriminator

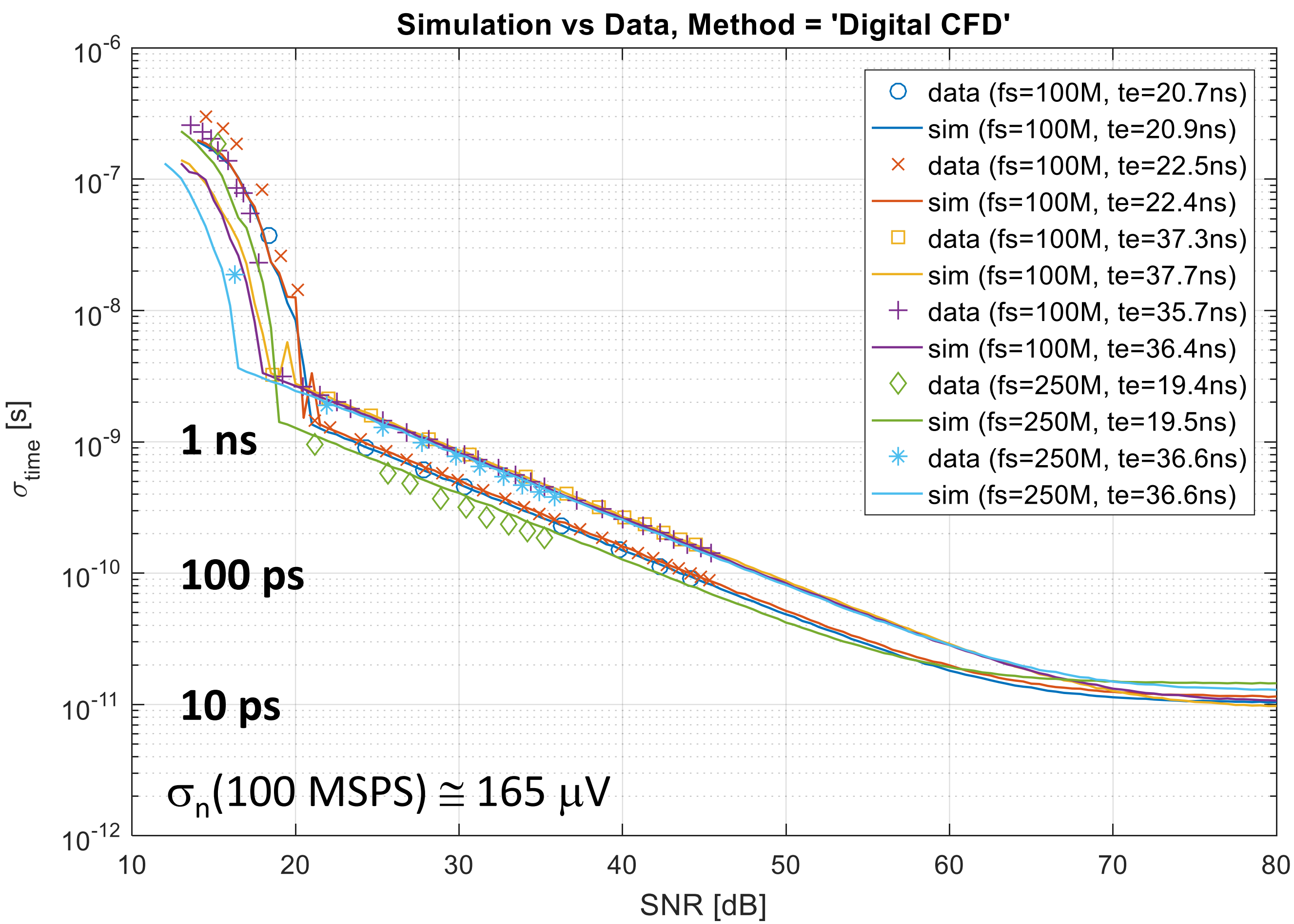
Processing via a digital CFD:

1. See if input waveform crosses threshold.
2. Interpolate input waveform via FFT to allow for sub-sample CFD delays (optional step).
3. Delay and invert the interpolated waveform.
4. Subtract the inverted waveform from the interpolated one.
5. Find the minimum of the constant-fraction waveform.
6. Find two samples that have opposite sign, but require that they are after the minimum from step 5.
7. Calculate coordinate of zero-crossing by using linear interpolation.



If sampling density is low and pulse shape is constant, then one can apply additional correction to account for non-linear waveform near the zero-crossing point.

Results – Digital CFD 1/2



mV → 0.5 1.7 5.2 16.5 52.3 165.3 523 1653

SNR ≥ 20 dB

Good match of model and data for 100 MHz ADC, slightly worse for 250 MHz ADC

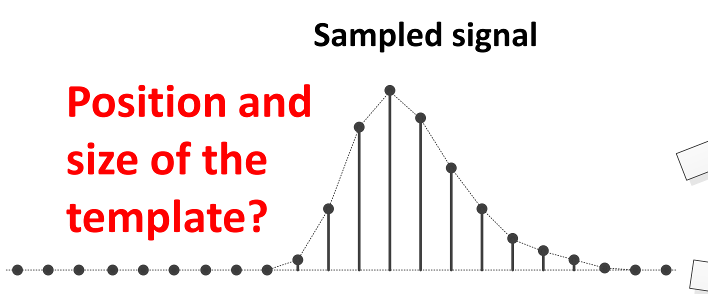
SNR < 20 dB

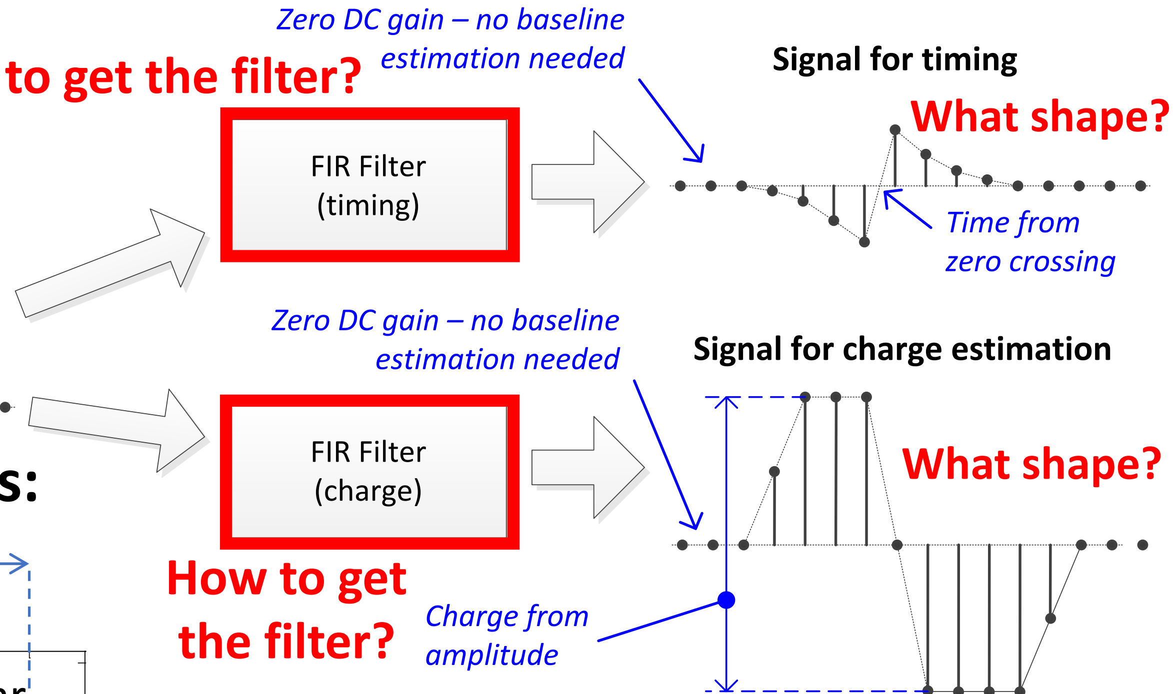
Poor match, data worse than model. Not a useful range anyway, as we need σ_{time} < 1 ns.

Timing resolution is proportional to

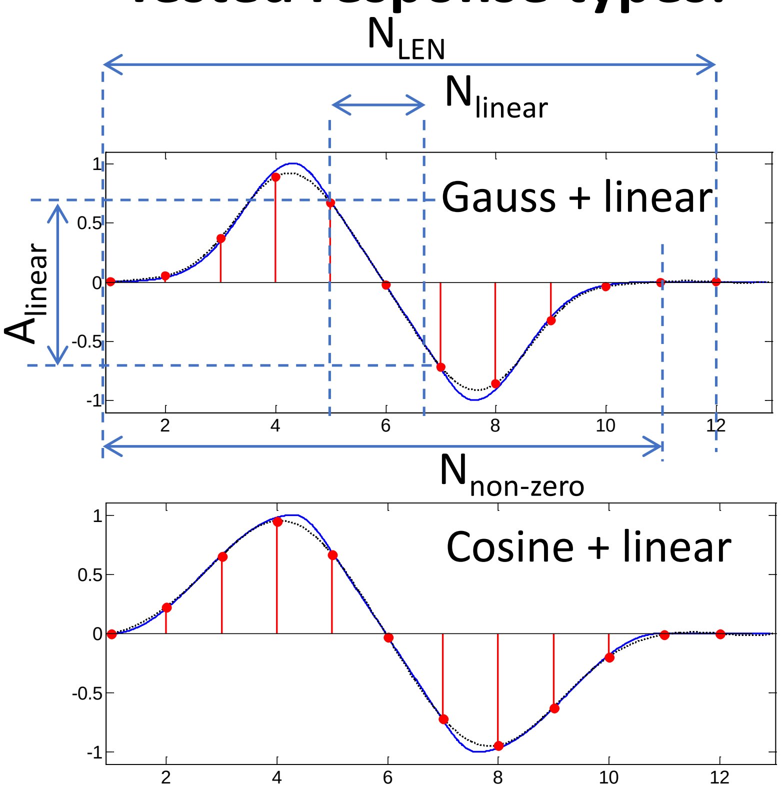
$$\frac{t_{rise}}{SNR}$$

FIR Filter Processing – DPLMS Method





Tested response types:

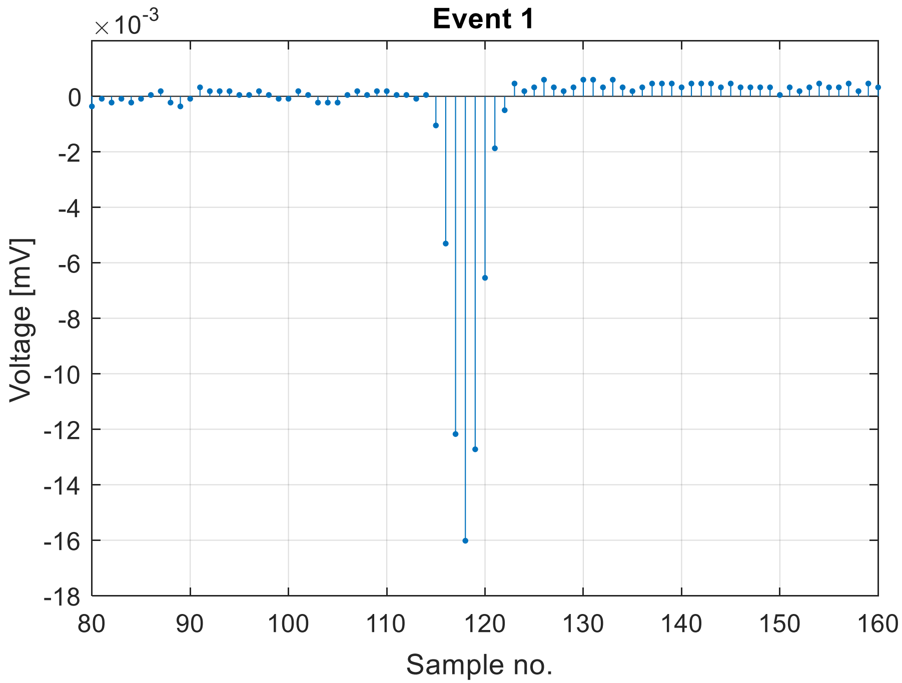


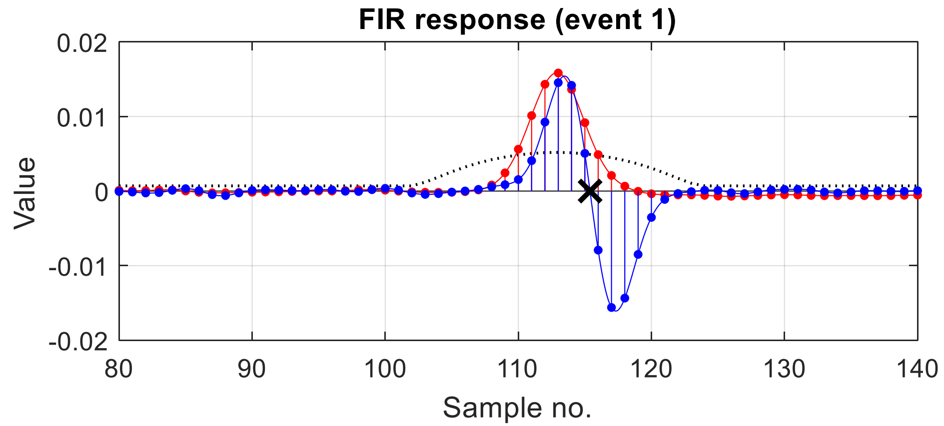
… or simply subtract pedestal and integrate.

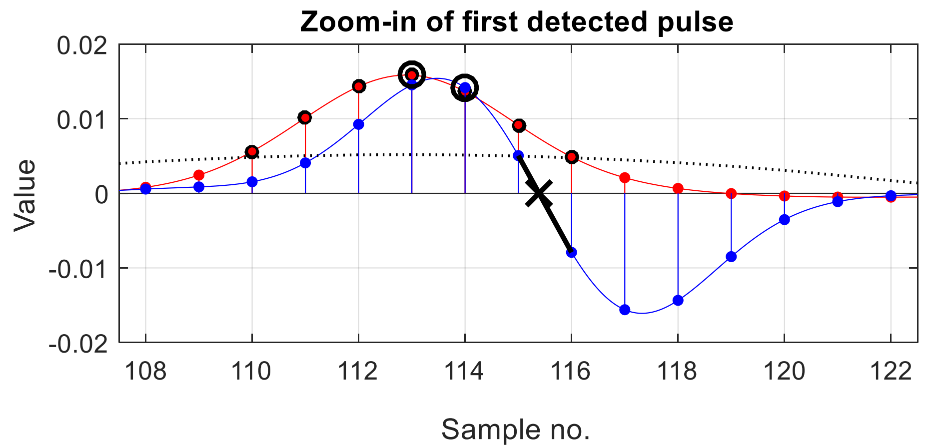
- FIR = Finite Impulse Response
- 'Black-box' approach → transform **known** input into desired output, don't care how.
- Arbitrary filter characteristic possible.
- Filter should be 'optimal' → **maximize SNR at detection time**.

Gatti E., et al., "*Digital Penalized LMS method for filter synthesis with arbitrary constraints and noise*", NIM A523, 167-185, 2004

FIR Processing

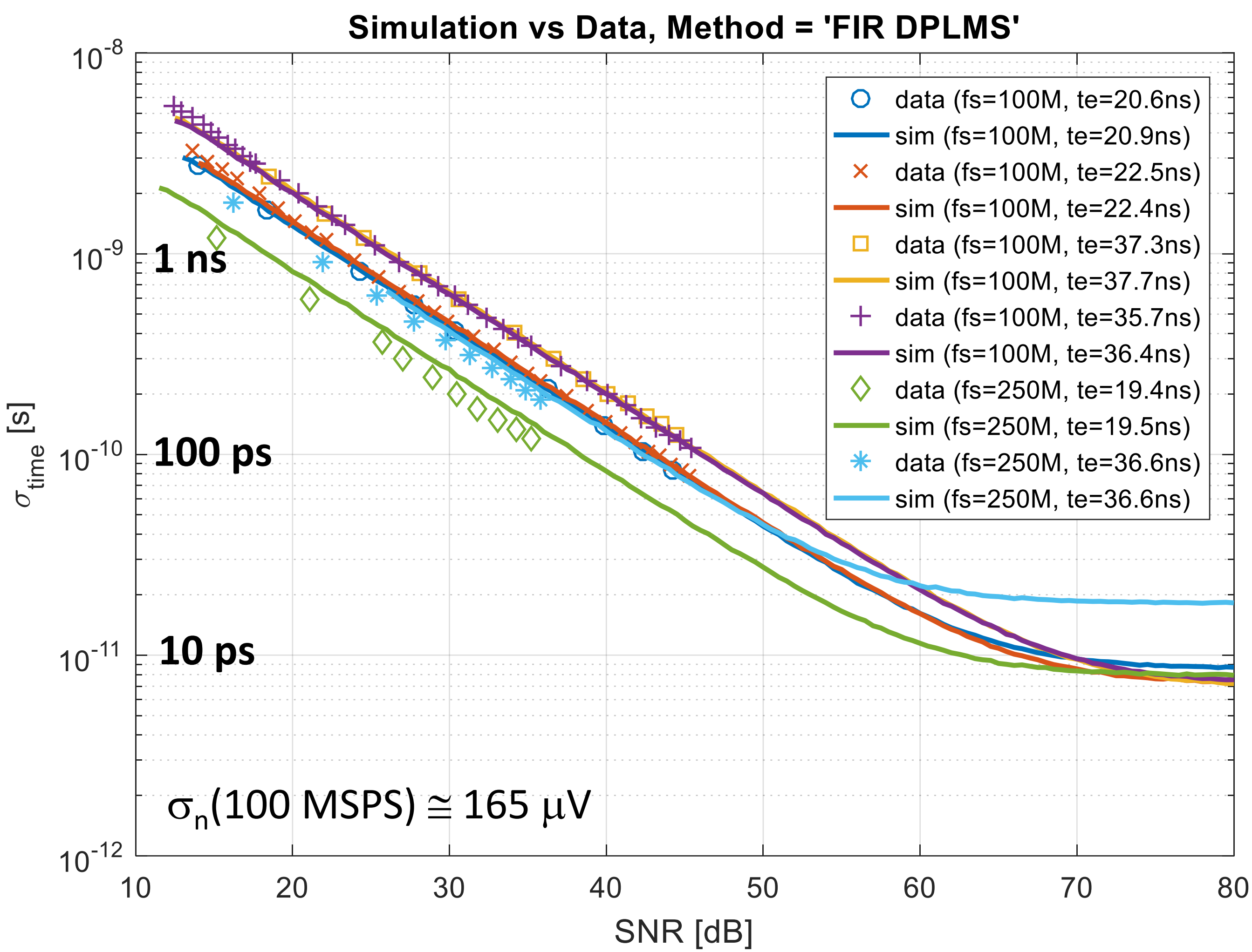






- Trigger on 'gate' filter response (red)
- Use adaptive threshold to prevent false positives (dotted black line)
- Timing using 'timing' filter response (blue)
- Apply correction to counteract non-linear shape of the waveform near zero-crossing.

Results – FIR DPLMS



mV → 0.5 1.7 5.2 16.5 52.3 165.3 523 1653

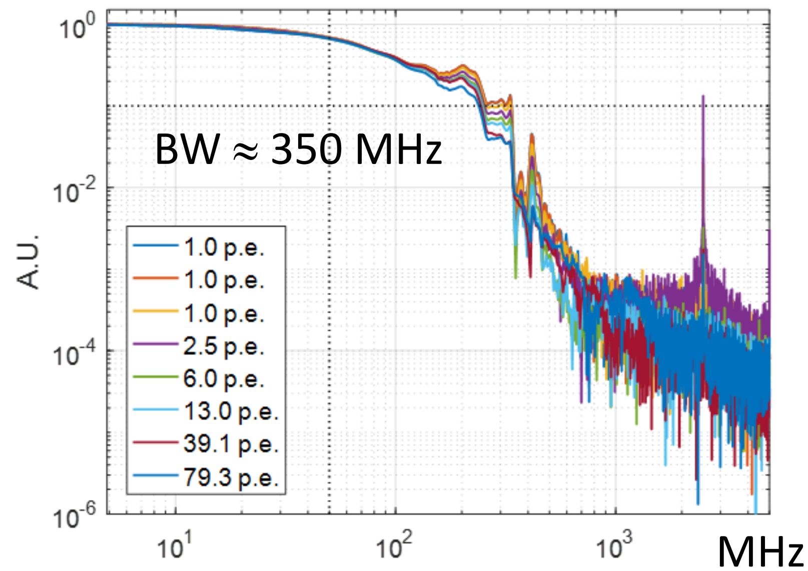
Good match of model and data for 100 MHz ADC, slightly worse for 250 MHz ADC

250 MHz data better than model – possibly due to some correlation which is not reflected by simulation.

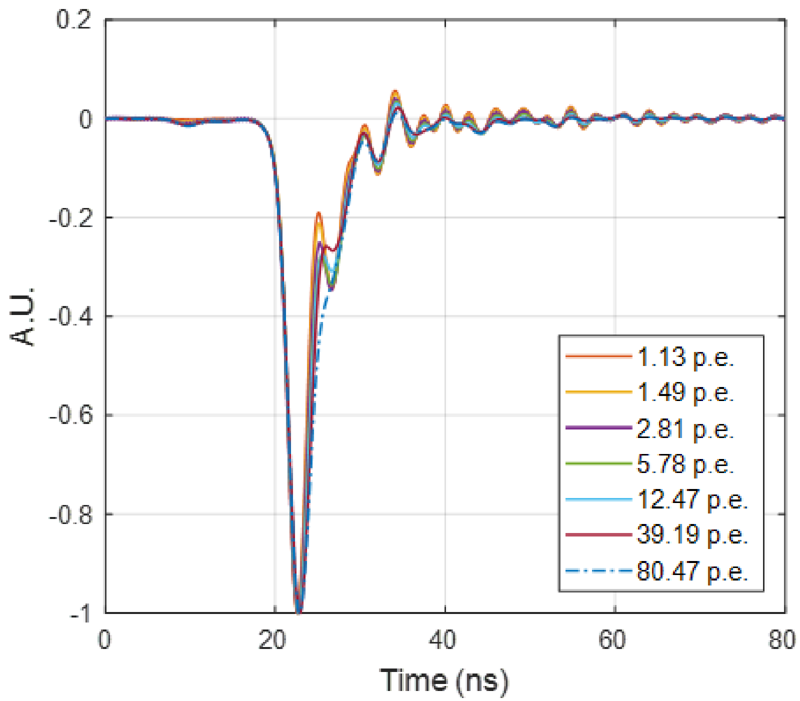
R14347 – Waveforms

- Small (but visible) dependence of waveform shape on PMT orientation wrt. Earth magnetic field
- Relatively large dependence of waveform shape on position of the light source on the photocathode
- $t_{rise} \in$ (1.9 ns, 3.0 ns), FWHM $\in$ (3.0 ns, 4.7 ns); both increase with PE level (expected)

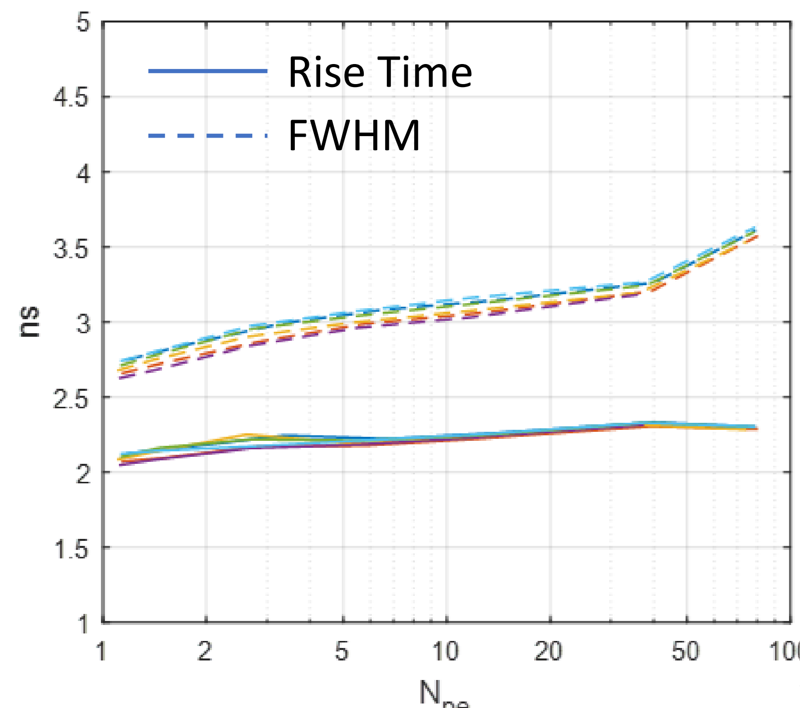
Normalized Amplitude Spectrum



Normalized templates

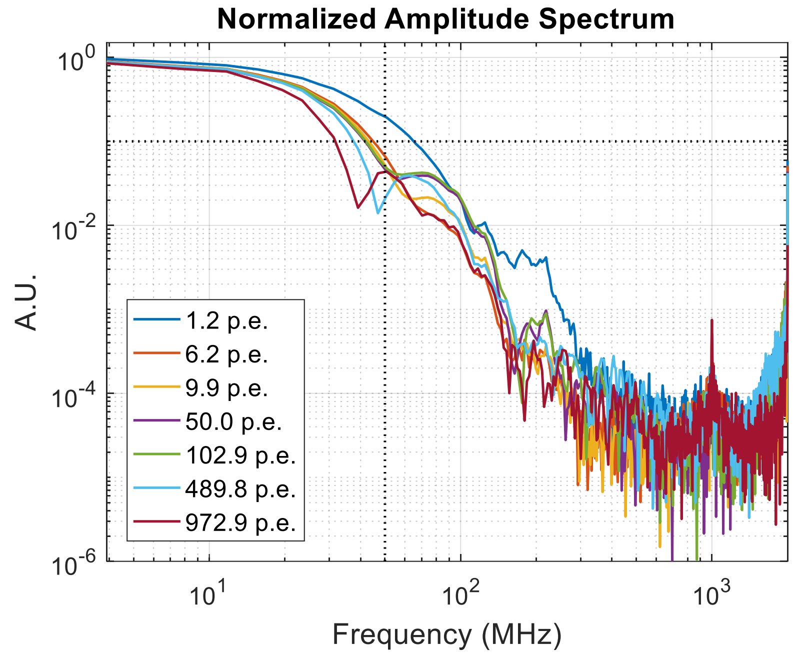


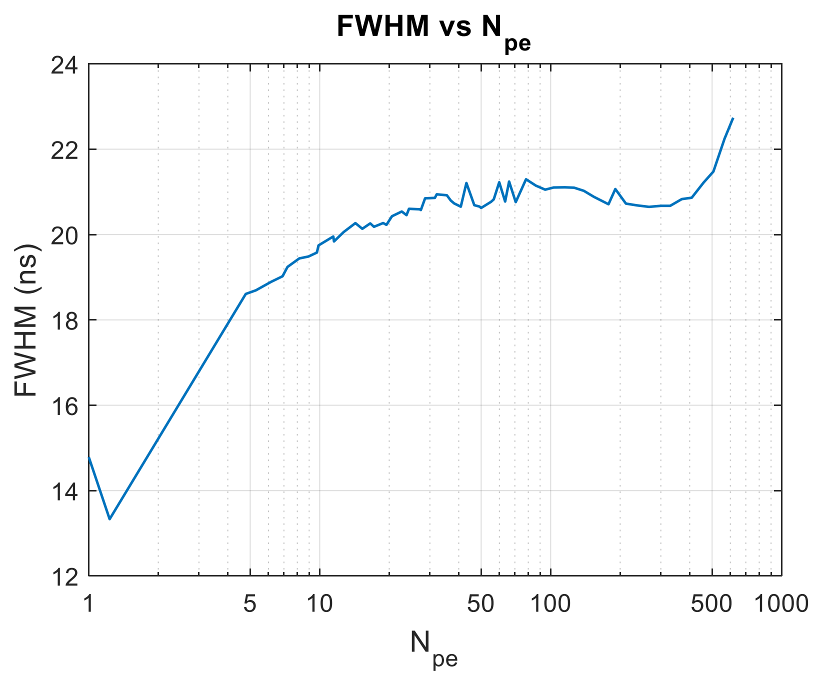
Rise Time & FWHM

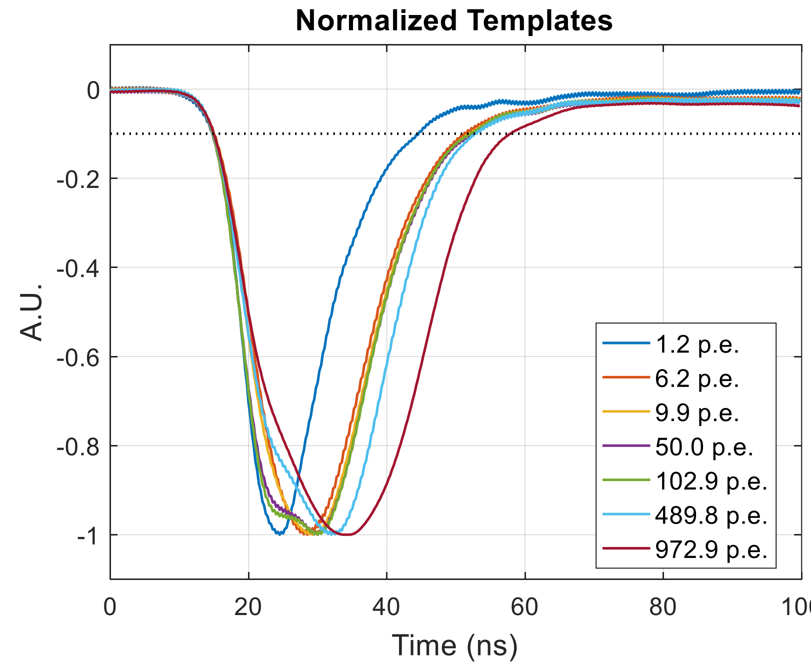


20” Box& Line – Waveforms

- See change in shape with p.e. level
    - Expected due to TTS
- Significant pulse broadening and edge deterioration for large p.e. levels
    - 1 p.e. $\rightarrow t_r \approx 3.5$ ns, multi-p.e. $\rightarrow t_r \approx 8$ ns
- Bandwidth is roughly 65 MHz for 1 p.e. pulses, down to 1% power level
    - Significantly lower for larger pulse





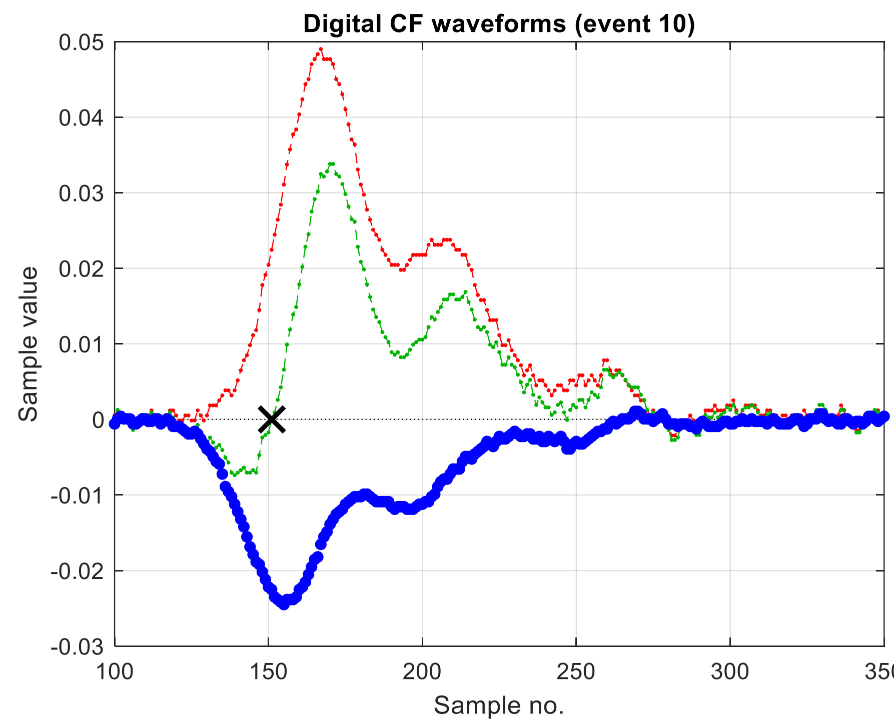


Waveform Samplers - Conclusions

- Pulse shape of PMT response changes with number of photons and position on the photocathode
- Need to foresee that in FIR-based methods the estimate may be completely wrong in case of non-standard shape (for ex. pile-up)
  - Need quality factor for each time/charge estimate
  - Should send full waveform for off-line processing

Revised time estimation

- Digital CFD – limit shift to leading Edge only
- For FIR-based method, need to parameterize impulse response of the filter wrt. charge



Significant increase in data rate – need efficient coding and possibly lossy waveform compression

Summary

- Front-end choice is a critical decision
- Various options of digitizers available, each with its advantages and disadvantages
- Cost-wise, waveform approach seems comparable to TDC/QTC or TDC/Peak-Sensing (at least in HK case), but power is still an issue
- HK/E61 case:
  - First prototypes of QTC+TDC approach already tested and working OK
  - Sample & hold approach also after first tests, on-going work
  - FADC prototypes foreseen for end of this year / beginning of 2019. Already have good electronics models.
  - Starting work on modifications to FIR-based waveform processing and data compression
  - Currently not-considering multiple time-over-threshold

BACKUP

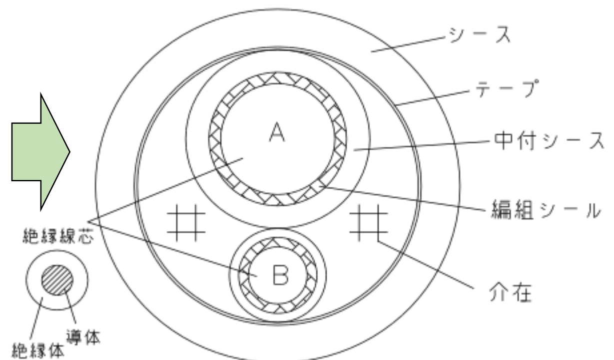
Waterproof Cable and Connector

Waterproof cable complex

Coaxial cable with
1-wire HV for SK

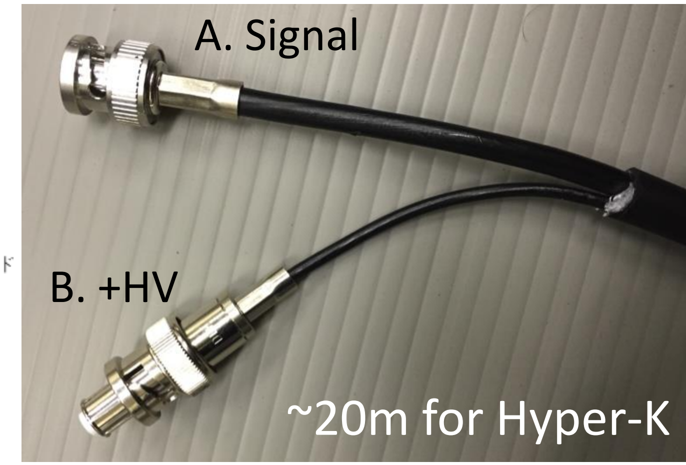
8.4 mmϕ, 68 g/m

Two coaxial cables for HK
PE sheath



9.4 mmϕ, 86 g/m

By Hamamatsu



Watertight connector (up to 100m water)

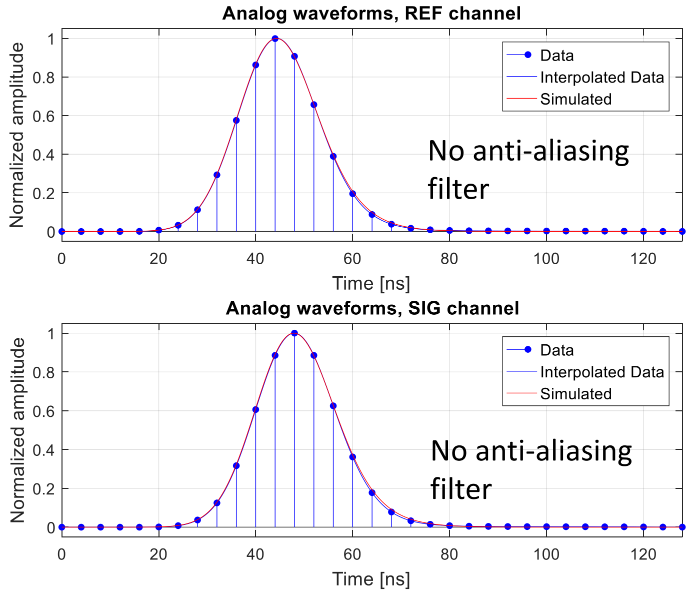
- Dedicated connector was developed.
- Connected to electronics case in water, and can be disconnected.

TNC coaxial signal (RG58C/U) + HV pins

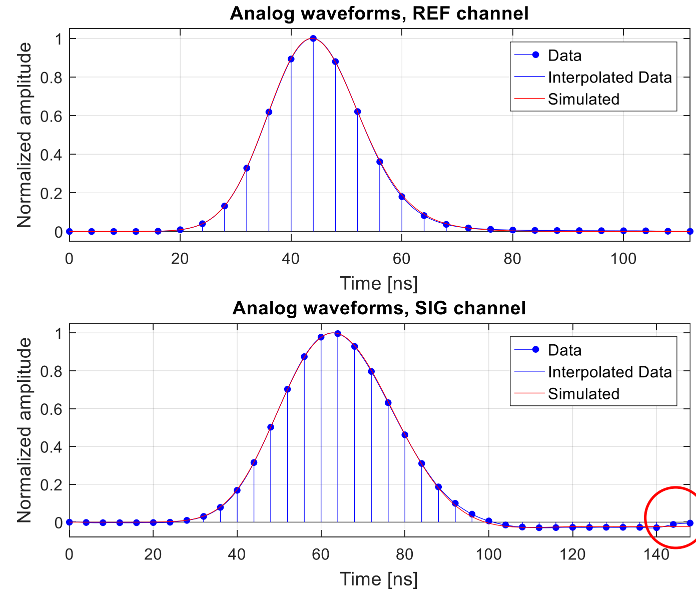
Improved noise shield and less failure of connection compared with SK.

Signal Models

250 MHz, CH1 = ref, CH2 = ref

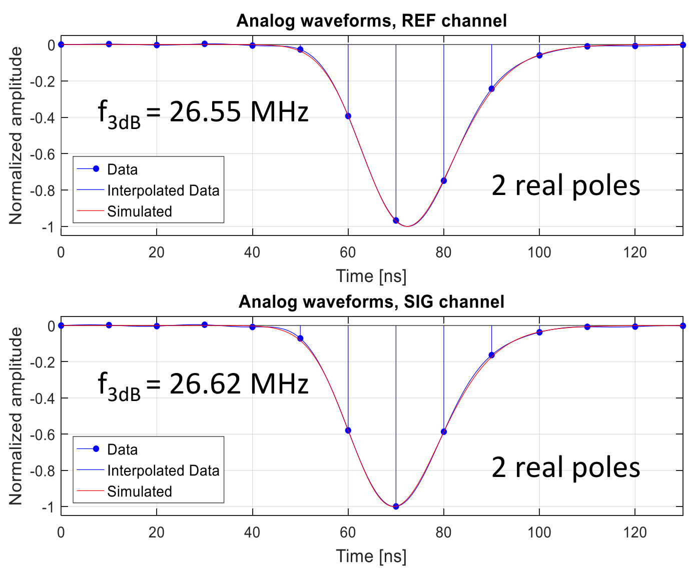


250 MHz, CH1 = ref, CH2 = sig (15 ns)

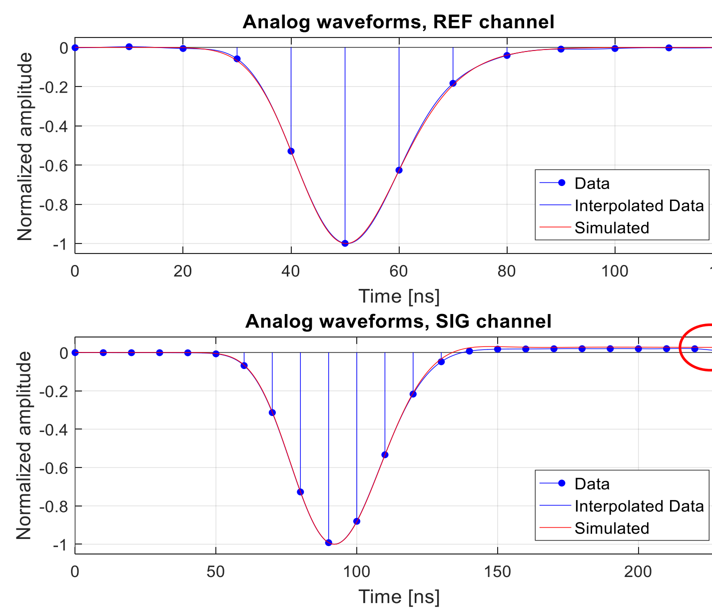


All pulses matched by FWHM

100 MHz, CH1 = ref, CH2 = ref

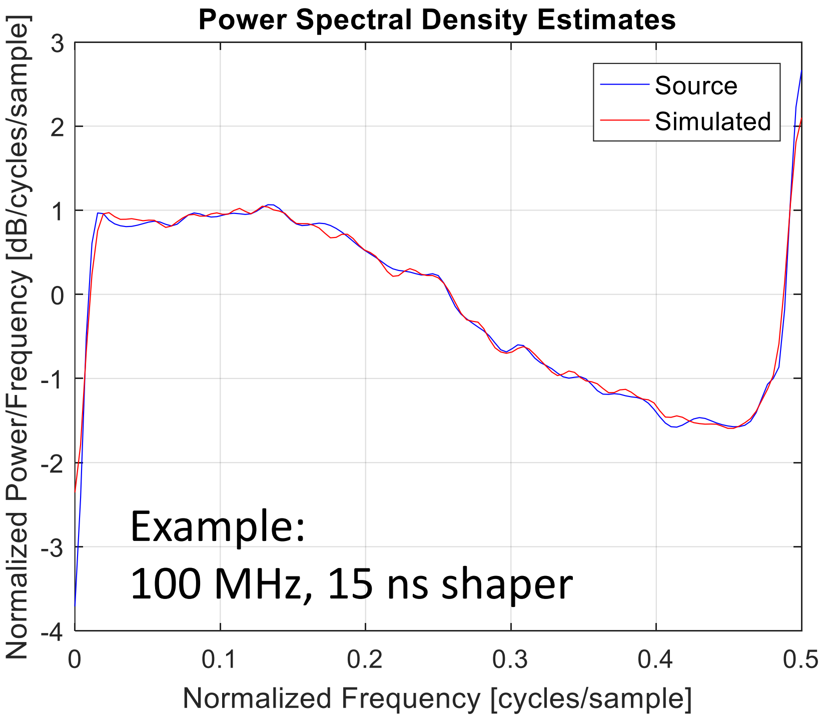


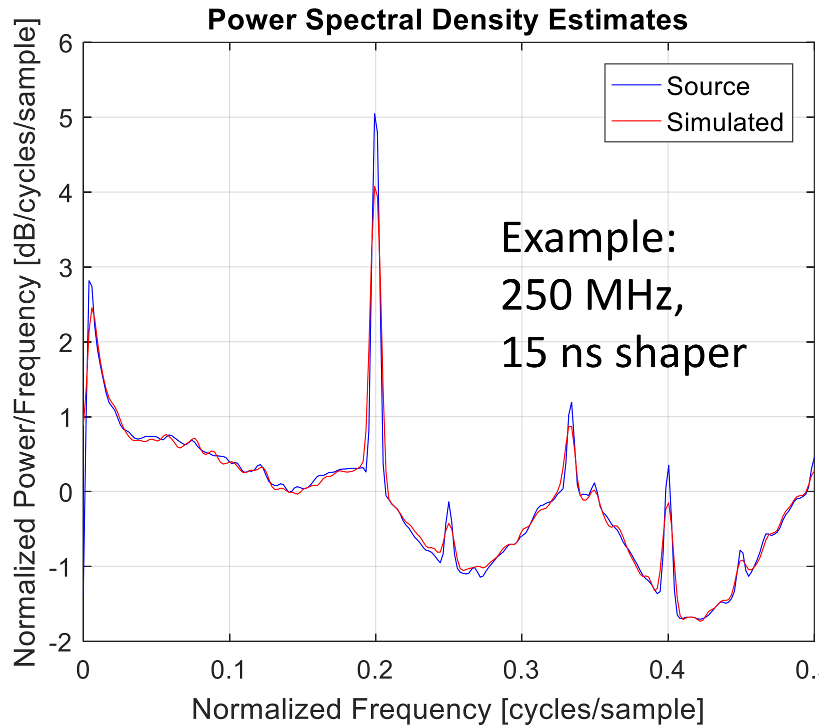
CH1 = ref, CH2 = sig (15 ns low power)



Interpolation artefacts

Noise models

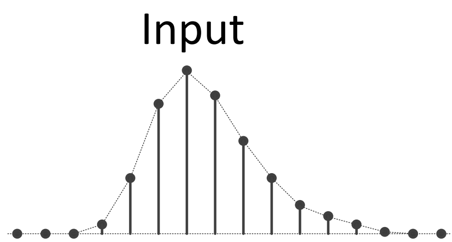


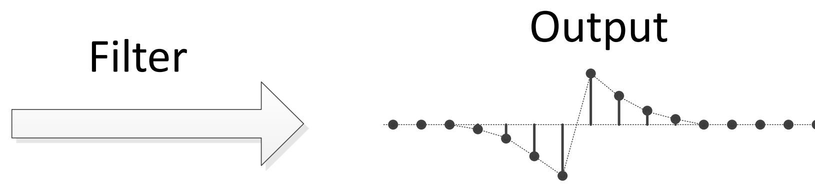


- Good match of simulated periodogram with an experimental one.
- Potential problem:
  - Some of the deterministic components (peaks in spectrum) do not have random phase, but are correlated to sampling clock.

Synthesizing FIR filter – Method 1

Digital Penalized LMS Method





input signal *noiseless signal (our template)* *stationary noise*

$$x[n] = x'[n] + x''[n]$$

number of filter taps *impulse response of the filter*

Filter is **linear**, so the output signal is:

$$y[n] = \sum_{l=0}^{N-1} h[l]\cdot x'[n-l] + \sum_{l=0}^{N-1} h[l]\cdot x''[n-l]$$

Therefore, we can deal with noise and signal components separately

Take multiple measurements, then:

Minimize overall variance of the response:

Sought filter

$$Var(y) = \boldsymbol{h}^{1,N}\cdot \boldsymbol{R}^{N,N}\cdot \boldsymbol{h}^{N,1}$$

Noise auto-covariance matrix

Minimize difference between filter response and our desired response

$$(E(y[k]-v_k))^2 = (\boldsymbol{h}^{1,N}\cdot \boldsymbol{x}'(k)^{N,1} - v_k)^2$$

Value of k-th sample of the response to x' *N past samples of x', starting from k*

Gatti E., et al., "*Digital Penalized LMS method for filter synthesis with arbitrary constraints and noise*", NIM A523, 167-185, 2004

Synthesizing FIR filter – Method 1 (cont.)

Digital Penalized LMS Method

Add additional constraints for frequency response, including gain at DC ...

Add constraints related to bit-gain (i.e. how well we are supposed to reject quantization noise) ...

Finally, build the error functional and minimize it:

$$Area(FIR) = \frac{Area(y)}{Area(x)}$$

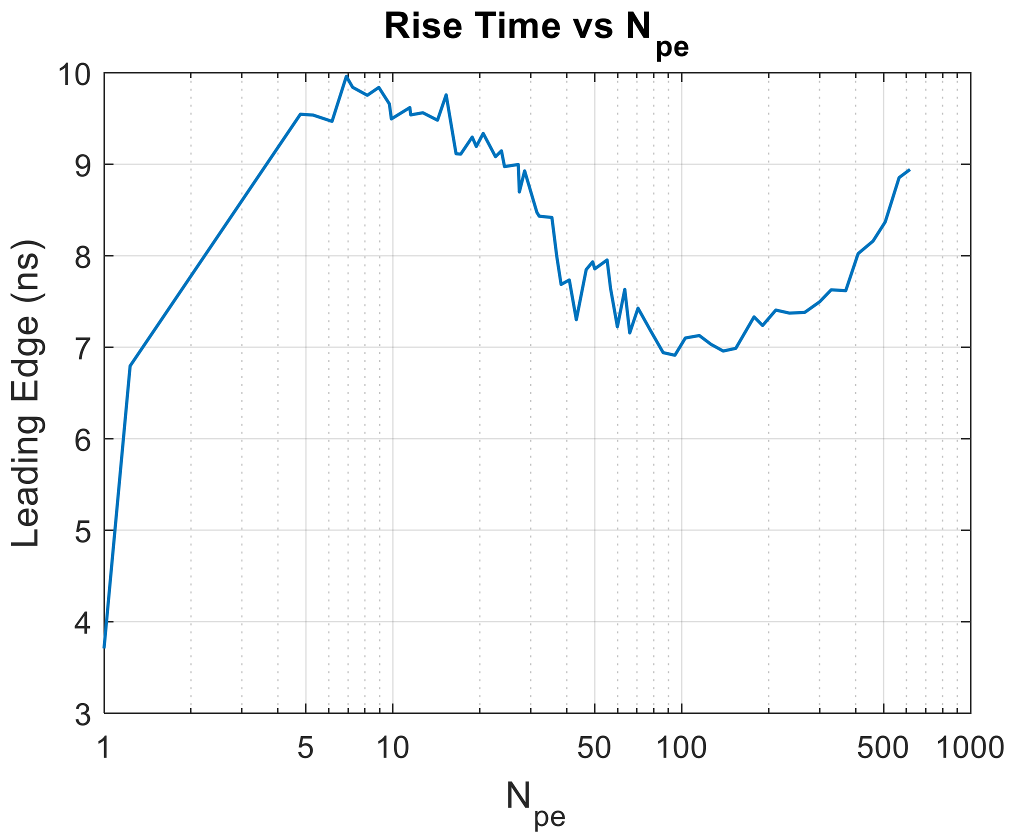
$$\varepsilon^2 = \overbrace{Var(y)}^{\text{Constraint for variance}} + \overbrace{\sum_{k=1}^{N} \alpha_k \cdot (E(y[k]) - v_k)^2}^{\text{Constraints for shape of response to pulse template}} + \overbrace{\sum_{l=1}^{L} \beta_l \cdot \left(|F\{\boldsymbol{h}\}|_{\omega_l}\right)^2}^{\text{Frequency constraints}}$$

$$+ \underbrace{\varphi \cdot (F\{\boldsymbol{h}\}_{\omega=0} - Area(FIR))^2}_{\text{DC gain (i.e. area) constraint}} + \underbrace{\gamma \cdot \sum_{n} (h[n])^2}_{\text{Bit-gain constraint}}$$

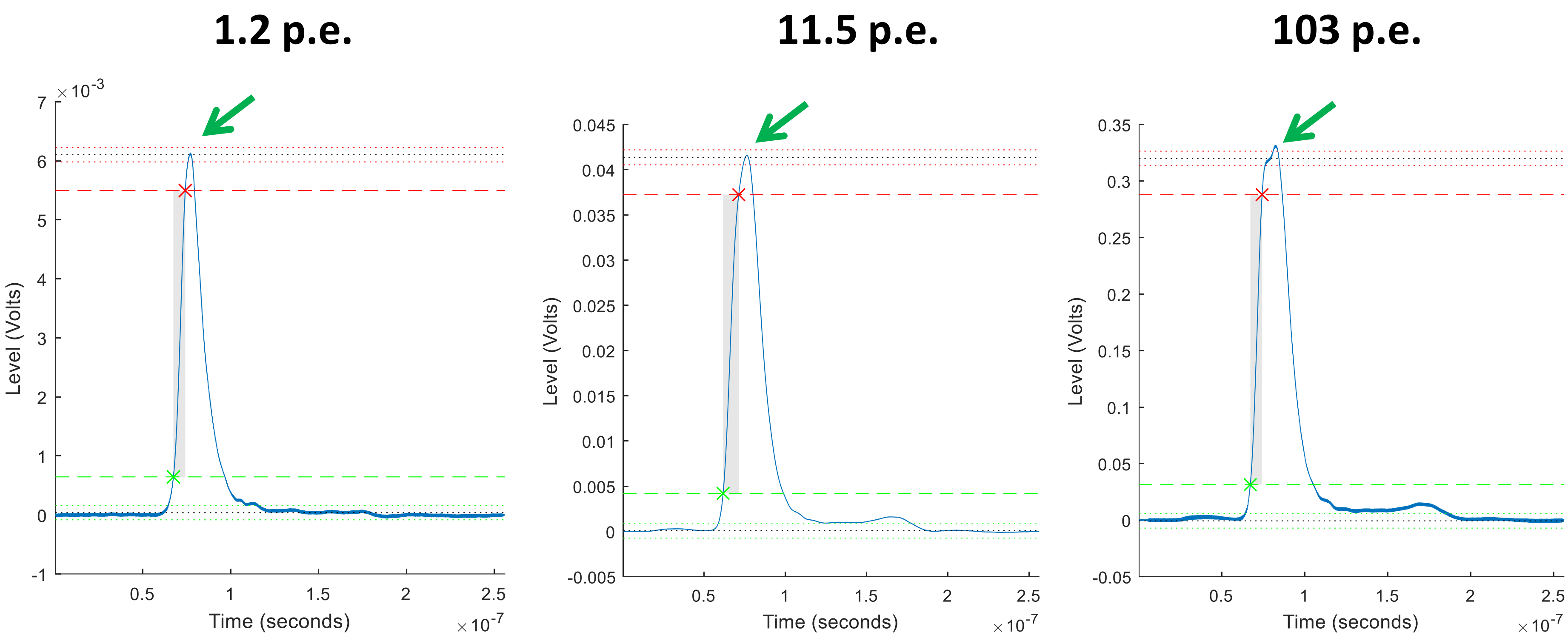
All components are square functions, so there exists a global minimum – just need to properly choose ***N***, $\vec{v}$, $\vec{\alpha}$, $\vec{\beta}$, φ and γ → papers don't say much about that

20” B&L Rise Time

- Significant changes in pulse rise time with p.e. level
- Very strange dependency – not sure if it comes from the PMT or also from the laser



20” B&L - Why Strange Rise Time Dependency?



- Clearly see peak broadening. For ≈10 p.e. level the ‘double peak’ effect manifests itself just as pulse broadening (hence longer edge), for larger p.e. level it is more clearly visible and we again see the edge of first pulse only.

20” B&L - Saturation

- Also checked maximum level and pulse shape in case PMT gets saturated
- The pulse amplitude can go above 6 volts.
- Also see significant change in shape.