ALICE PHOS Collaboration

**Configurable Electronics with Low Noise and 14-bit Dynamic Range for
Photodiode-based Photon Detectors**

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Abstract

We describe the principles and measured performance of custom configurable 32-channel shaper/digitizer Front End Electronics (FEE) cards with 14-bit dynamic range for gain-adjustable photon detectors. The electronics has been designed for the PHOS calorimeter of ALICE with avalanche photodiode (APD) readout operated at -25°C ambient temperature and a signal shaping time of 1 μ s. The electronics has also been adopted by the EMCal detector of ALICE with the same APD readout, but operated at an ambient temperature of +20°C and with a shaping time of 100ns. The CR-RC2 signal shapers on the FEE cards are implemented in discrete logic on a 10-layer board with two shaper sections for each input channel. The two shaper sections with gain ratio of 16:1 are digitized by 10-bit ADCs and provide an effective dynamic range of 14 bits. Gain adjustment for each individual APD is available through 32 bias voltage control registers of 10-bit range. The fixed gains and shaping times of the pole-zero compensated shapers are defined prior to FEE production by the values of a few R and C components. For trigger purposes, “fast OR” outputs with 12-bit dynamic range are available. FPGA based slave logic, combined with a USB processor supports a variety of remote control and monitoring features, including APD gain calibration. The measurements presented here for APDs at -25° C ambient temperature and 1 μ s shaping time achieve an average RMS noise level of 0.25 ADC counts or 290 electrons. The linearity over the dynamic range is better than 1%, as is the uniformity of shaping time and gain over 32 channels. Due to the excellent correspondence of the output pulse shape with offline fit, a differential timing resolution of less than 1.5 ns between channels has been achieved at ca. 2 GeV, i.e. at 1.5% of the dynamic range of PHOS.

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1. Photoelectron signal of PHOS and EMCal detectors

The PHOS detector [1] of ALICE is an electromagnetic calorimeter consisting of 17920 PbWO₄ crystals of equal size 180x22x22 mm³. The detector is operated at an ambient temperature of -25° C to enhance the scintillation light yield. Each crystal is attached to an Avalanche Photo-Diode (APD) of type S8148, produced by Hamamatsu Co, Japan. The APDs have an active area of 5x5 mm² and are mounted in a ceramic case of 9x11 mm².

The EMCal detector [2] of ALICE is a large acceptance lead/scintillator sampling calorimeter with “shashlik”-type readout via wavelength shifting fibers and consists of 12672 readout towers. The fiber bundle of each tower is read out with a S8148 APD operated at the ambient temperature of the ALICE experiment (+20° C).

Tests of the FEE electronics described here have been performed with S8148 APDs, excited with photon pulses which imitate the characteristics of PbWO₄ scintillation light via LEDs of type Kingbright L7104PCB with 470 nm peak wavelength. The number of APD photoelectrons per MeV at gain M=1 can be estimated from the PbWO₄ light yield measurements [1] taken with a photomultiplier XP 2262b at room temperature. By taking detector surfaces and quantum efficiencies of photomultiplier and APD in relation for a nominal PbWO₄ light-yield of 10 photoelectrons per MeV at +25 °C one obtains:

$$N_{APD} = N_{PMT} \times (S_{APD}/S_{PMT}) \times (E_{APD}/E_{PMT}) = 10 \times (25/484) \times (0.7/0.25) = 1.45e/MeV$$

At the PHOS operating temperature of -25° C the light-yield is about 3 times higher, hence

$$N_{APD(-25C, M=1)}^{PHOS} \sim 4.4 e/MeV$$

With a nominal APD gain for PHOS of M=50, the APD charge signal is:

$$N_{APD(-25C, M=50)}^{PHOS} \sim 220 e/MeV$$

Preliminary results from test beam measurements for the EMCal gave corresponding numbers for the EMCal with APD gain M=1 and M=30 of

$$N_{APD(25C, M=1)}^{EMC} \sim 4.4 e/MeV^1 \text{ and } N_{APD(25C, M=30)}^{EMC} \sim 132 e/MeV$$

With a 1.2pF charge capacitor of the preamplifier, the charge conversion gain is 0.83 V/pC and the preamplifier output step voltages for PHOS and EMCal at APD gain M=50 and M=30 are:

$$U_{step(-25C, M=50)}^{PHOS} \sim 29.2 \mu V/MeV \text{ and } U_{step(+20C, M=30)}^{EMC} \sim 17.5 \mu V/MeV, \text{ respectively.}$$

2. Dynamic range

The dynamic range of PHOS has been defined by the requirement of 80 GeV maximum single channel energy and 250 GeV maximum single channel energy for EMCal. Given the 16:1=2⁴ gain ratio of the high and low gain channels, the two overlapping 10-bit ADC ranges correspond to an effective 14-bit dynamic range.

High gain range: 4.9 MeV to 5 GeV (PHOS) and 15.3 MeV to 15.6 GeV (EMCal)

Low gain range: 78 MeV to 80 GeV (PHOS) and 248 MeV to 250 GeV (EMCal)

The minimum value corresponds to the Least Significant Bit (LSB) of the high-gain ADC. The digital resolution is given by the LSB value relative to a signal value of energy E. For the above PHOS ranges, the digital resolutions LSB/E=4.9/E [MeV] for high gain and 78/E [MeV] for low gain are shown in Fig. 1, compared to first testbeam data taken by PHOS with old electronics [3]. Apart from a small region above 5 GeV, the digital resolution lies always significantly below the targeted PHOS energy resolution at high energies of 3%/√E ⊕ 0.8%. The combined noise from 3x3 crystal areas of 7.5 MeV/ E (2.5 MeV single channel noise) lies closely above the digital resolution of the high gain shaper.

1. the same light yield figure as PHOS is accidental

Fig. 1. Digital resolution over PHOS dynamic range.

The intrinsic energy resolution of the EMCal is expected from simulations to be about $6.9\%/\sqrt{E} \oplus 1.4\%$ [2]. The relative contribution of the digital resolution to the total EMCal resolution is shown as bottom dual-range curve in Fig. 2. A 1% constant contribution due to relative calibration is assumed. The electronics noise has been assumed to be 2000 electrons per APD for 100 ns shaping time. The dotted curve represents the corresponding 3x3 electronics noise contribution term of 160 MeV/E which is negligible to the intrinsic resolution (top solid line).

Fig. 2. Various contributions to the EMCal energy resolution.

3. Shaper gain

The shaper gain is defined with the above choices since the maximum energies must correspond to the maximum analogue input ADC Voltage of 1.0 Volt. With a nominal preamplifier output of 29.2 μ V/MeV for PHOS with APD gain M=50, the two gain ranges are defined as:

High gain: LSB= 143 μ V (5.0 MeV), Umax= 0.146 Volt (5 GeV) -> **shaper gain = 6.85**

Low gain: LSB= 2.28 mV(78 MeV), Umax= 2.34 Volt (80 GeV) -> **shaper gain = 0.427**

And in the case of EMCal with the above nominal preamplifier output of 17.5 μ V/MeV and APD gain M=30, the two gain ranges are defined as:

High gain: LSB= 267 μ V (15.3 MeV), Umax= 0.273 Volt (15.6 GeV) -> **shaper gain = 3.66**

Low gain: LSB= 4.34 mV(248 MeV), Umax= 4.37 Volt (250 GeV) -> **shaper gain = 0.229**

The real implementation does not need to reproduce these exact gain values since the assumptions about light-yield are only nominal and may be adjusted via the APD gains.

4. Energy resolution

The energy resolution of the calorimeter is parameterized as [1]

$$\frac{\sigma}{E} = \sqrt{\frac{a^2}{E^2} + \frac{b^2}{E} + c^2}$$

with the following parameters:

- noise term a: parallel and serial noise from detector diode and preamplifier
- stochastic term b: shower fluctuations and APD excess noise
- constant term c: detection losses, inter-calibration of channels, non-uniformity, non-linearity, instabilities (temperature and bias)

In the case of PHOS, a fit to PHOS test beam data [3], gives the following preliminary values: a=13 MeV (3*3 crystals), b=3.58% GeV^{0.5} and c= 1.12%. For the high energy region E > 1 GeV, the fit parameters are compatible with

$$\frac{\sigma}{E} = \frac{3\%}{\sqrt{E}} \oplus 0.8\%$$

The interest of PHOS in a high signal/noise ratio is to minimize the noise term a in order to achieve a good energy resolution at low energies. In particular, the two-photon invariant mass peaks of the π^0 (134.97 MeV) and η (547.75 MeV) are important for yield measurements and for energy calibration of the calorimeter [4].

5. Channel readout

Fig.3 depicts the readout principles for one FEE channel. An APD is biased via a resistor R_p to a voltage which corresponds to the required gain. The charge produced by the photodiode is the product of number of photoelectrons and APD gain. As noted above for PHOS with gain M=50, the nominal detector charge is Q=220 e/MeV and Q=132 e/MeV EMCal with M=30. For further signal analysis, the charge dynamics is approximated as a delta function, although in reality, the scintillating light of the PbWO₄ crystals has a finite rise and decay time (of order 1ns and 10 ns, respectively). Ideally, the charge sensitive preamplifier (CSA) converts the delta charge into a voltage step of amplitude U=Q/C_f. In reality the step function has a finite risetime of order 20 ns which is also limited by the finite risetime of the CSA. For simplicity, the capacitor C_f is auto-discharged via a resistor R_{dis} with a time-constant which is very large (of order 100 μ s) compared to the time of measurement. The CSA and photodiode are physically close to the calorimeter, hence operated at the detector's ambient temperature.

Fig. 3. Photodiode and preamplifier, followed by dual gain shaper and ADC's.

The CSA step voltage signal is connected via a short cable to the shaper and digitizer of the FEE electronics where it is amplified by a low noise buffer, followed by two parallel CR-RC2 shapers of equal time constant and gain ratio of 16:1. The products of the buffer gain (g) and the shaper gains (HG, LG) correspond to the gains required for the given dynamic range. For each shaper section of high and low gain, there is one 10-bit ADC to digitize the data.

6. Shaper bandpass

The FEE shaper is a second-order bandpass filter for a central frequency of $f_c = 1/(2\pi * \tau)$ where τ is the shaping constant. It consists of a simple RC high-pass of -3dB/octave attenuation below f_c , followed by two low-pass filters with -6dB/octave above f_c . The combined bandpass enhances the Signal-over-Noise ratio by separating the “noise sea” from the signal’s main Fourier components. The choice of the cutoff frequency f_c (or shaping constant) defines the Signal-over-Noise ratio. The shaper’s Bode diagram of Fig.4 represents the frequency domain transfer characteristics, which is strictly linked to the shaper’s time domain behaviour or

output pulse shape discussed later. The Bode diagram of Fig.4 was obtained by simulating the circuit diagram of the shaper with Matlab. It depicts the high pass (dotted) and lowpass (solid red) transfer characteristics as well as the combined bandpass transfer curve (interrupted line). The peak frequency at $f_c = 160$ kHz corresponds with $f_c = 1/(2\pi \tau)$ to a shaping time of $\tau = 1 \mu s$.

Fig. 4. Bode diagram of the PHOS shaper bandpass for $\tau = 1 \mu s$.

7. Signal over noise

With the aim to achieve a good energy resolution also at low energies, the shaper was designed with a low-noise input buffer stage. The noise of cascaded amplifiers with N1 and N2 noise levels is calculated as

$$N_{\text{noise}} = \sqrt{N1^2 + \left(\frac{N2}{K1}\right)^2}$$

K1 is the amplification of the first buffer stage. By choosing $K1 > 1$ the contribution of the shaper noise N2 can be reduced by choosing a buffer with a low-noise figure N1.

The standard noise analysis [5] is depicted in Fig 5.

Fig. 5. Equivalent noise of detector and preamplifier, followed by the shaper

The noise components are:

- Current noise (parallel): Shottky $2qI_{\text{diode}}$ + shunt resistor thermal current $4kT/R_b$ + equivalent input current
- Voltage noise (serial) $4kTR_s$ thermal Johnson noise at the amplifier input
- $1/f$ noise

The Equivalent Noise Charge (ENC) at the output of the shaper is a function of the shaping time constant τ , of the absolute temperature T , of the parallel, serial, and constant noise sources, and of the detector capacitance. The current noise is proportional to the shaping time τ , the voltage noise is inversely proportional to τ , and the $1/f$ noise is independent of τ . With an estimate for $R_s = 2/3 * (1/g_m)$ where g_m is the forward transconductance of the JFET transistor of the CSA, the ENC as function of shaping time is given by:

$$ENC^2 = \frac{4kT}{q^2 \cdot R_b} \cdot F_p \cdot \tau + \frac{4kT}{q^2} \cdot \frac{2}{3} \cdot \frac{1}{g_m} \cdot F_s \cdot \frac{C_d^2}{\tau} + C_d^2 \cdot \text{const}$$

For the PHOS electronics the following values apply: $R_b = 20 \text{ MOhm}$, $g_m = 30 \text{ mS}$, and $C_d = 100 \text{ pF}$ (APD = 90 pF @ -25°C , JFET = 10 pF). The noise figures [6] for a CR-RC2 shaper are: $F_s = 0.84$ and $F_p = 0.63$.

The ENC noise in dependence on the shaping time τ (without the negligible constant $1/f$ noise term) is shown in Fig.6 for EMCal ($+20^\circ \text{C}$) and for PHOS (-25°C). The noise minimum lies at $2...3 \mu\text{s}$. At shorter shaping times, the voltage noise dominates while at higher shaping times, the current noise dominates. PHOS has nevertheless chosen a shaping time constant of $1 \mu\text{s}$ as a compromise for improved offline timing resolution (see section 17.4). The achieved average RMS noise level of 0.25 ADC counts (see section 16.3) corresponds to $39 \mu\text{V}$ at the output of the preamplifier, or with $N_{q, \text{noise}} = 1/q (U \times C_f)$ this corresponds to 292 noise electrons. Corrected for APD gain $M=1$, this corresponds to a noise level of ca. 6 primary photoelectrons.

In the case of the EMCal, the intrinsic resolution is worse than PHOS and so the electronics noise contribution is less important. As shown in Fig.2, an electronics noise contribution of as much as 2000

electrons per channel is not significant relative to the intrinsic resolution. Therefore EMCal has chosen a much shorter shaping time of 100 ns to suppress energy deposit from late event-related neutron background, and to reduce the data volume per channel. The calculated noise for EMCal would correspond to the upper curve (+20° C) of Fig. 6.

Fig. 6. Calculated ENC noise as function of shaping time for PHOS at -25° C and EmCal +20° C.

8. Shaper analysis

The Laplace representation of an n-th order shaper of type CR-RCn with step function input from a charge sensitive amplifier is a product of equivalent operators. With the Laplace equivalent “s” for the step voltage,

Fig. 7. Full diagram of the analogue stages of a CR-2RC shaper.

the preamplifier-shaper chain of Fig.7 can be transcribed with shaping constant τ_0 and shaper gain A into an Laplace operator $H(s)$ which represents the full chain of the FEE electronics analogue part:

$$H_{\text{CSA} + \text{shaper}}(s) = s \times \left[\frac{1}{1 + \frac{1}{R_{\text{dis}} C_f}} \right] \times \left[\frac{\tau_0}{1 + s\tau_0} \right] \times \left[\frac{A}{1 + s\tau_0} \right]^n$$

The first bracket term represents the “auto-discharge” combination $R_{\text{dis}} * C_f$ of the CSA. As shown below, this term can be eliminated by including the “pole-zero-cancellation” RC network with the inverse operator in the input of the shaper. The second bracket is the $C_p * R_d$ highpass (differentiator) and the third term are n equal lowpass (integrator) stages. The shaping time constants τ_0 of differentiator and integrators are equal. The general solution [6] of this Laplace operator in the time domain is a Gamma function of order n :

$$V_n(t) = \left[\frac{n^n Q \cdot A^n}{C_f \cdot n!} \right] \cdot \left[\frac{t}{\tau_p} \right]^n \cdot e^{-\frac{t}{\tau_p}}$$

The zero-crossing of the first derivative shows that the peak amplitude is proportional to the APD charge Q , hence to the scintillation light.

$$V_{\text{max}} = \frac{Q \cdot A^n \cdot n^n}{C_f \cdot n! \cdot e^n}$$

For the $n=2$ second-order CR-RC2 shaper, the pulse shape $V_n(t)$ in the time domain is described with a peaking time τ_p at V_{max} . The relation between peaking time and shaping time is $\tau_p = n * \tau_0$, hence for the FEE electronics with $n = 2$, the peaking time is twice the shaping time and the pulse shape is:

$$V_{\text{out}}(t) = \left[\frac{2Q \cdot A^2}{C_f} \right] \cdot \left[\frac{t - t_0}{\tau_p} \right]^2 \cdot e^{-2 \cdot \frac{t - t_0}{\tau_p}}$$

Please note that this solution is the time domain-equivalent representation of the frequency diagram shown previously in Fig.4.

9. Shaper implementation

The shaper highpass (differentiator) stage can simply be implemented as an RC filter with $R_0 * C_z = 1/(2\pi * f_c)$ at the input of the low noise buffer (Fig.8). In order to cancel the first bracket operator of the Laplace operator above, it is sufficient to add a resistor R_z in parallel to C_z and choose the combined time constant equal to $R_{\text{dis}} * C_f$. The final value of R_z is deviating slightly from the theoretical value; it was determined by measurement, i.e by adjusting the signal $V_{\text{out}}(t)$ to zero undershoot.

Fig. 8. Differentiator and pole-zero cancellation stage of the shaper input stage.

The combined resistor values of the differentiation and pole-zero cancellation are listed in Table 1 for two shaping times.

shaper constant τ_0	R_z	C_z	R_o
$\tau_0=2\mu\text{s}$	143 k	470 pF	9.1 k
$\tau_0=1\mu\text{s}$	143 k	470 pF	4.22 k

Table 1: Values of differentiator and pole-zero filter for 1 and 2 μs shaping time.

The effect of the CSA auto-discharge term is demonstrated in Fig. 9 by replacing the CSA output with a step pulse generator: the pole-zero cancellation filter which would normally compensate the signal overshoot produces the opposite undershoot. The fully compensated FEE pulse with CSA is shown later in Fig. 12.

Fig. 9. Oscilloscope output of uncompensated step pulser time response for two shaping times: 1 μs and 2 μs .

The shaper lowpass was implemented as a dual integrator stage using a single operational amplifier with multiple feedback architecture (MFB) as shown in Fig. 10. The values of the components are obtained as solution of the transfer equations, described in the following sections.

Fig. 10. Dual integrator circuit with multi-feedback architecture.

The dual integrator transfer function can be expressed [7] with gain K as a second order polynomial of the Laplace s operator:

$$H(s) = \frac{K}{1 + as + bs^2}$$

For a flat Bessel filter transfer characteristics of second order, the following conditions for the coefficients of the polynomial apply:

$$K = R_2/R_3 \quad (\text{gain of dual integrator for } f \ll f_c)$$

$$a = \omega_c C_1 (R_2 + R_3 + R_2 R_3 / R_1) = \mathbf{1.3617}$$

$$b = \omega_c^2 C_1 C_2 R_2 R_3 = \mathbf{0.618}$$

In order to avoid complex solutions the capacitor values must fulfil the condition:

$$C_2 \geq C_1 \cdot \frac{4b \cdot (1 - K)}{a^2}$$

The solutions for the resistances, given as a set of capacitor values and a shaper gain K for a cutoff frequency $f_c = 1/(2\pi \cdot \tau_0)$ with shaping time constant τ_0 , are shown below:

$$R_2 = \frac{aC_2 - \sqrt{a^2 \cdot C_2^2 - 4bC_1C_2 \cdot (1 - K)}}{4\pi f_c \cdot C_1C_2}$$

$$R_1 = \frac{R_2}{-K}$$

$$R_3 = \frac{b}{4\pi^2 f_c^2 \cdot C_1C_2R_2}$$

$$f_c = \frac{1}{2\pi \cdot \sqrt{R_2 \cdot R_3 \cdot C_1 \cdot C_2}} \cdot \frac{1}{FSF}$$

The frequency scaling factor is $FSF = \mathbf{1.2736}$. The values obtained for 100ns and 1 and 2 μs shaping time for high and low gain are shown in Table 2.

τ_0	R1	R2	R3	C1	C2	K
2 μs (high gain)	649 Ω	4.42 k Ω	1.78 k Ω	150 pF	2.2 nF	3.35
2 μs (low gain)	10.5 k Ω	4.22 k Ω	5.90 k Ω	220 pF	470 pF	0.21
1 μs (high gain)	681 Ω	4.87 k Ω	1.96 k Ω	68 pF	1000 pF	3.35
1 μs (low gain)	4.02 k Ω	1.69 k Ω	5.36 k Ω	150 pF	470 pF	0.21
100 ns (high gain)	78.7 Ω	316 Ω	205 Ω	100 pF	1000 pF	1.95
100 ns (low gain)	590 Ω	147 Ω	442 Ω	220 pF	470 pF	0.12

Table 2. CR-RC2 shaper component values for three shaping time constants.

The full implementation of one shaper channel, including pole-zero-compensation, is depicted in Fig. 11 for the 2 μs shaping time ($= 4\mu s$ peaking time). A low-noise buffer of gain 2 with RC highpass as shown in Fig.8 is followed by two parallel, 2nd order lowpass filters of gain ratio 16:1, each implemented with the same Bessel filter transfer characteristics and with the same shaping time. Their differential output, required for the ADCs, is implemented by inverting and non-inverting drivers. Analogue anti-aliasing filters of ca. 4 MHz cutoff frequency are inserted in the input lines of the ADC in order to remove out-of-band components, due to digital ADC switching noise.

Fig. 12. Offline high gain ADC counts versus samples [100 ns]. The red line is a Gamma-2 fit.

After pedestal correction, the signal peak amplitude of this event is 410 ADC counts and the peaking time corresponds to $2 \mu\text{s}$. Since the dynamic range definition of PHOS implies that 5 MeV corresponds to 1 ADC count, the peak amplitude corresponds to 2.05 GeV.

12. FEE prototype response to electrons and protons

The first FEE card prototypes were tested with a triggered electron beam, directed to a 16 by 16 PbWO_4 crystal matrix [4] in August 2004, and with mixed beams of protons, pions, and electrons with momenta of 4 to 120 GeV/c directed to an 8 by 8 array of EMCal towers in November 2005. Fig. 13 shows the digitized pulse shape output with 10MHz sampling (100ns/channel) for an electron event in PHOS (Right) and a proton event in EMCal (Left). The different peaking times of $2 \mu\text{s}$ and 200ns, respectively, corresponding to the different shaping times are apparent.

Fig. 13. FEE electronic pulse shapes (100ns/channel) with testbeam: Left-EMCal, Right-PHOS.

The test beam results confirm the FEE gains and shaping times and allow extracting the light yields from the calibrations in terms of MeV/ADC count and the overall energy resolution including the FEE noise contribution. Gain adjustments within a factor 2 are easily accomplished via APD bias adjustment.

13. FEE card

The architecture of FEE electronics within the PHOS readout and trigger system has been described in [13] and [14]. The photograph of Fig. 14 depicts the 10 layer FEE card with all its details. The total power consumption is ca. 5 Watt.

Fig. 14 Photograph of the 32 channel FEE card

Special attention was given to a hierarchical power filtering and careful layout of ground planes for analogue, digital, and high voltage sections, avoiding coupling of digital noise into the analogue shaper sections. With temperature, voltage and current monitoring, individual FEE cards within a closed detector volume can be disabled in case of power or temperature problems. The cards are designed for operation inside water-cooled copper cassettes. For this purpose, the FEE component layout leaves room for inlet and outlet water pipes and contains holes for mounting screws of the surrounding cassette.

14. Test environment

A test environment with APD diodes requires ideally the same conditions as exist in the calorimeter, i.e. a stable ambient temperature and a pulsed photon light source with similar wavelength and timing

Fig. 15. Test environment in the Wuhan PHOS laboratory at CCNU.

characteristics as the scintillating light produced by electromagnetic showers in PbWO_4 crystals. The corresponding test setup of the PHOS laboratory in Wuhan is shown as a simplified diagram in Fig 15.

The inputs of the FEE card are connected to an array of APD/CSA in a testbox which is kept inside a refrigerator at a temperature of -25°C in order to mimic the PHOS operating conditions. A pulse generator is used to trigger an avalanche LED pulser, containing a LED with 470 nm peak spectral emission. The emitted light-pulse is distributed via optical fibers to the APD array. All APD's have been set via bias voltage registers to a gain value which generates the same amplitude for the common light input. Upon reception of the trigger output from the trigger pulse generator, the Readout Control Unit (RCU) sends a timing strobe to all ALTRO chips to store ADC samples of all selected channels into their sample buffers. Thereafter, these buffers are transmitted, via RCU and the optical fiber DDL link to a Data Acquisition Computer running the DATE software. The ROOT analysis tools are used to sort the sampled ADC data into histograms and to perform user-defined data analysis. The results are shown and discussed in the following chapters.

14.1 LED pulser

An avalanche LED pulser [15] was built for the test setup to generate up to 10^{10} photons of 470 nm for distribution via optical fibers with a 1-2 ns light risetime. Fig 16 shows a ~ 20 Volt LED pulse amplitude of < 1 ns risetime and the APD testbox with optical fibers, fanned out from the pulsed LED.

Fig. 16. Left: LED voltage pulse at the pins of a Kingbright L7104PCB / Right: APD test box with optical fibers facing each APD.

The charge/voltage step response of the CSA, which corresponds to an (attenuated) photon flash from a pulse like Fig 16, transmitted over ca. 50 cm of optical fiber to the APDs, is depicted in Fig. 17. The LED (Kingbright L7104PCB) is specified for typically 350 milli-candela @ 20 mA. When measured over a series

resistor, a pulse corresponding to 10A peak current can be converted over lumen and human eye correction factors into ca. $8 \cdot 10^9$ photons. The step function of the attenuated light flash arriving over optical fibers at the APDs, shows a slight overshoot which is due to the test-setup. This effect is not observed with real particles in the testbeam.

Fig. 17. CSA response to a LED pulse.

15. Fast analogue OR

A threshold trigger can be based on analogue sums generated on the FEE cards from 2*2 input channels. A trigger region unit (TRU) has been designed [16] to receive all fast OR signals from 14 FEE cards and to generate charge sums over 4*4 sliding windows over 448 channels. In order to generate an analogue sum from 4 input channels, the input signals on the FEE cards are split (compare Fig. 11) into semi-gaussian energy channels and into fast-OR trigger channels with 12 bit dynamic range. The fast OR's are generated by a simple RC shaper with 100 ns pulse width.

16. Measured FEE characteristics (PHOS 1 μ s shaping time)

The measurements presented here for the PHOS detector with the choice of 1 μ s shaping time are results of the ROOT offline analysis with data generated via the described test system and read out via a computer for offline analysis.

16.1 Linearity and gain over dynamic range

The theoretically linear relation between the peak amplitude of the Gamma-2 function and the APD charge (input voltage step amplitude) was confirmed by the measurement of Fig 18.

Fig. 18. PHOS linearity measurement over dynamic range.

The lower curve represents the full low gain range (2.3 mV...2.39 V input step) the upper line shows the high gain range up to 0.149 Volt, extrapolated to input step voltages below 10 mV. The voltage levels below 10 mV could not be measured precisely with our oscilloscopes. The two shaper gains can be derived from the slopes (i.e. p1 coefficient) after converting ADC counts in Voltage according to 1ADC bit = 1V/1024=0.97656 V:

- High gain, measured = **6.7 +- 0.051**
- Low gain, measured = **0.397 +-0.001**

These values correspond with the design values defined in section 2. The gain ratio of 16.87 is close to the value obtained later in section 16.5 by direct measurement.

16.2 APD gain calibration for PHOS

The gain spread of the APDs for a common bias voltage of 340 Volt was measured in a PHOS testbeam October 2004 with the first FEE prototypes with 2 GeV electrons. Individual gain control via 10-bit bias voltage registers [17] on the FEE cards allows for controlled gain calibration of 0.20Volt per bit in the range 210-400Volt. The measurement of Fig.19 shows the relation between APD bias Voltage and remote controlled DAC settings:

$$V_{\text{bias}} = 209.9 + 0.2022 * \text{DAC-counts}$$

Fig. 19. Bias register counts versus APD bias voltage.

With a gain dependence on the voltage of $1/M \times (dM/dV) = 3.3\% / \text{Volt}$ for gain $M=50$ [18] and a bias voltage increment 0.20Volt/bit, the gain resolution ($T=\text{constant}$) is 0.66%. Fig. 20 shows the gain distribution before and after FEE gain calibration, i.e. after individual bias adjustment via bias control registers. The un-calibrated variations amount to more than 300%. After (non-exhaustive) hardware pre-calibration the variation was reduced to ca. 5%.

Fig. 20. APD gain spread at -25°C with 2 GeV particles (left) and after hardware pre-calibration (right).

A direct determination of the APD gain versus bias voltage correspondence of individual APDs requires knowledge of their signal amplitude for gain $M=1$. By measuring the signal amplitude at gain $M=1$, and amplified by the desired gain, one obtains the corresponding bias voltage. Such a measurement is shown in Fig. 21 over the full bias voltage range for two APD's. This determination of the gain $M=1$ is however not very precise and depends also on the light wavelength. As described in [18], APD gain $M=1$ should be

measured at a bias Voltage of 10 Volt due to the large increase of diode capacitance at low voltages. With this, the nominal gain of $M=50$ is reached at ca. 350 Volt for the two APD's.

Fig. 21. APD gain curve for two APDs at -25 C. Gain 50 is reached at around 350 Volt if gain $M=1$ is determined at 10 Volt.

In view however that an inter-calibration of better than 1% is aimed for in the constant term of the energy resolution term of the PHOS and EMCaL detectors, this method of gain calibration is only useful for pre-calibration, while inter-calibration to the <1% level requires calibration with electromagnetic showers.

16.3 Pedestal and RMS measurement

The pedestal and RMS measurements are taken from N pre-samples of the digitized signal as shown in Fig. 12, measured over a sample of events. The RMS for a single event is calculated as

$$\text{RMS}(N) = \sqrt{\frac{1}{N} \sum_{i=1}^N (y_i - \text{Mean})^2}$$

where Mean is the average of the N samples. With $N = 10$ pre-samples of signals like shown in Fig. 12 the measurement in the laboratory with the described PHOS test setup gave the following results: For an unconnected FEE card, Fig 22 (left) shows the pedestal distribution over all 32 high gain FEE shaper channels (bottom) and their derived RMS noise (top) for 2050 events. The constant spread of pedestal levels is normal, due to component tolerances. The uniform RMS noise level is around 0.22 ADC counts. After connecting FEE channels 24...31 to 8 APD detectors in an absolute dark environment with gain setting $M=50$ at -25°C the noise distribution taken from pre-samples of LED pulses is a little more disturbed as shown in Fig 22 (right). There is a clear RMS noise enhancement in the APD channels over the unconnected channels to an average of 0.23 ADC counts.

Fig. 22. left: Pedestal and RMS noise measurement for a FEE card without detector connected; top: RMS noise averaged over 2050 events; right: same with 8 APD's connected and 24 channels unconnected. Note that the pedestals are unchanged.

With the dynamic range definition of PHOS, 1 ADC counts corresponds to 5MeV, hence the single channel RMS noise for PHOS is 1.16 MeV. This figure is significantly better than measured with first FEE prototypes in the 2004 testbeam. For the final noise situation inside the PHOS module we assume 2.5 MeV as a more conservative noise figure, hence the PHOS noise term for a 3x3 crystal area can be assumed with new FEE electronics as $\sigma = 7.5$ MeV (compare section 4).

16.4 APD noise as function of bias voltage

As shown in section 7 the FET voltage noise dominates at shaping times below the minimum noise, hence the APD current has little influence at shaping times below $2\mu s$. However the APD capacitance is a strong function of the APD bias voltage with a strong increase of capacitance and noise towards low voltages. Towards high bias voltages, the excess noise, due to charge avalanche, increases the effective noise level.

The excess noise is a statistical process which broadens the signal resolution, hence belongs to the stochastic term “b” of the spectrometer’s energy resolution. Given an excess noise factor of $F=2.27$ for gain $M=50$ it can be shown that [18]

$$\frac{b}{\sqrt{E}} = \sqrt{\frac{F}{N_{pe}}}$$

With $N_{pe} = 4$ e-/ MeV and $F=2.27$ @ ($M=50$), the lower limit of the single channel stochastic term at 1 GeV can be approximated as $b > 2.3 \text{ \% GeV}^{1/2}$.

Fig. 23 depicts an all inclusive noise measurement versus APD bias voltage for 8 APDs at the PHOS ambient temperature of $-25^{\circ}C$. The effect of excess noise can be seen for some APDs already above 360 Volt. The average noise at 350 Volt ($M=50$) corresponds here to 0.3 ADC counts (1.5 MeV).

Fig. 23. Measurement of effective noise (ADC counts) versus APD bias voltage for 8 APDs, read out and digitized by FEE electronics.

16.5 Gain ratio of high /low gain

The design gain ratio for FEE cards (high gain / low gain= 16/1) of the two shapers sections cannot be implemented exactly as 16/1. The 32-channel gain ratio measurement of Fig. 24 reports a value of 16.83 which effectively increases the dynamic range to 14.075 bits. This measurement also demonstrates that excellent uniformity has been achieved with the discrete component shapers of the FEE cards.

Fig. 24. Uniformity of gain ratio High/Low

16.6 Peaking time measurement

The design peaking time of the second order PHOS shapers with $1\mu\text{s}$ shaping time is $2\tau = 2\mu\text{s}$. Fig. 25 shows the measurement taken with 8 APD's pulsed by a single LED flash. The measured mean value is $2.175 \pm 0.01\mu\text{s}$.

Fig. 25. Peaking time measurement of the PHOS FEE shaper with $1\mu\text{s}$ shaping time.

17. Timing resolution

Great attention was given to the PHOS requirement for a timing resolution of the order 1 to 2ns, for time-of-flight cuts that will be used to discriminate against anti-neutron background, predominantly around 2 GeV. The timing resolution achievable with semi-gaussian shapers is a weak function of noise and shaping time but a strong function of the signal amplitude (ADC bits) and of the sampling frequency. Considerable improvement was gained by using the proper offline method and by using a LED pulser with light risetime of the same order or better than the targeted resolution.

The intrinsic timing resolution of PbWO_4 crystals is reported as 0.13ns [19] and timing measurements in testbeam [4] report a constant timing resolution of 0.5 ns above 1.5 GeV. The only affordable way of measuring time-of-flight in each individual FEE channel is by offline determination of the time reference obtained from the Gamma-2 fit. In order to achieve the 1ns level of timing resolution, it is important to first determine the peaking time constant of each individual channel and then use these values for the Gamma-2 fit.

17.1 Timing resolution of FEE card

The method applied for measuring the differential timing resolution of the FEE alone is depicted in Fig.26. A step voltage of equivalent energy E is connected to the inputs of two or more shapers. The differences of the time references t_1 and t_2 of the offline fits are compared and the differences are accumulated in a histogram. With sufficient statistics, the variance of the difference plot $\sigma(t_2-t_1)$ is a measure of the differential timing resolution. Assuming that both channels have the same resolution, the timing resolution is $\sigma/\sqrt{2}$.

Fig. 26. Principle of measuring differential time resolution measurement.

The measurement of the differential time resolution of 2 FEE channels without detector, using a step pulser of 2GeV equivalent energy is shown in Fig 27. With 2050 events and a sigma of 0.712, the differential time resolution of PHOS FEE electronics at 2 GeV is 0.5 ns.

Fig. 27. Offline measurement timing resolution of pulsed FEE channels.

The energy dependence of offline timing resolution measurement is shown in Fig. 28 for pulsed FEE channels. It is due to the digital resolution limit of the dynamic range, as shown in Fig.1. In the limit of 10-bit signals (1024 counts), the FEE electronics can resolve 300 ps at a shaping time of 1 μ s. This result hints that for future upgrades, the digital resolution should be increased to a dynamic range of 16-bits.

Fig. 28. Timing resolution dependence on equivalent energy (ADC counts).

17.2 Timing resolution with APD and LED source

The timing resolution with particles and APD diodes can be measured in good approximation by operating the APD's in the same bias, temperature, and connectivity environment as in the experiment and by pulsing a LED of similar wavelength and timing characteristics as the scintillation light of a PbWO_4 detector. The test environment described previously provides this possibility. The timing resolution measurement is depicted in Fig 29 (high gain) and Fig 30 (low gain). The equivalent LED light energy was ca. 2 GeV (compare Fig. 12 for the amplitude /energy equivalence of Run 157).

Fig. 29. Histogram of high gain timing resolution at ca. 2GeV with LED pulser.

The differential time resolution of the high gain shaper is 1.5 ns and of the low gain shaper is 6.22 ns. The combined high/low timing resolution for 1 μ s shaping time at 2 GeV equivalent energy is:

$$\tau_{\text{res}} (1\mu\text{s}) = 1.455 \pm 0.026 \text{ ns}$$

The difference in resolution with and without APD/preamplifier is .1 ns. This difference is due to the limited timing resolution of the APD (order 400 ps) and the finite risetime of the LED light.

Fig. 30. Histogram of timing resolution measured with the low-gain shaper channel. The resolution is worse compared to the high gain channel due to the low digital resolution at 2 GeV.

17.3 Timing resolution dependence on sampling frequency

The timing resolution obtained with a LED source is shown in Fig 31 as a function of ADC sampling frequency. The plateau above 1 MHz sampling frequency complies with the Nyquist theorem that a sampled signal is fully defined by sampling it at twice the signal bandwidth.

As shown in Fig. 4, the -10db signal bandwidth above f_c is ca 600 kHz for 1 μ s shaping time. With 10 MHz sampling rate, semi-gaussian signals for 1 μ s shaping time are by far over-sampled. Unless over-sampling [20] methods can be used offline to improve timing resolution (by gaining more effective bits), a sampling frequency of 2 MHz is sufficient for 1 μ s shaping time.

Fig. 31. Measurement of timing kresolution dependence on the ADC sampling frequency.

17.4 Comparative measurement with $2\mu\text{s}$ shaping time

As depicted in Fig. 6, a shaping time of $2\mu\text{s}$ should result in even lower RMS noise. Fig. 32 confirms this for $2\mu\text{s}$ shaping time by a measurement under exactly the same conditions as Fig. 22. Obviously the ENC noise from the shaper reaches the noise limit of the ADC at ca 0.2 ADC counts.

Fig. 32. Comparative measurement of Fig.21 but with $2\mu\text{s}$ shaping time ($4\mu\text{s}$ peaking time)

The reduced noise with $2\mu\text{s}$ shaping time entails however a worse timing resolution of 1.76 ns as measured in Fig. 33 with a LED source under the same conditions as Fig. 29. The conclusion of the two comparative measurements is that the RMS noise reduction has less importance for PHOS than the ca. 17 % improved timing resolution, and therefore a $1\mu\text{s}$ shaping time has been adopted.

Fig. 33. Comparative measurement of Fig. 28 with $2\mu\text{s}$ shaping time.

18. Conclusions and Outlook

The 32-channel low noise FEE electronics presented here has been designed for the high resolution PHOS electromagnetic calorimeter of ALICE with a dynamic range of 14-bits. The design performance for the PHOS calorimeter has been verified by measurements in the dedicated test facilities at CCNU and HUST in Wuhan, China. With cooled APD's and a shaping time of 1 μ s, the 0.25 ADC count RMS noise level corresponds to 290 electrons at gain $M=50$ and the offline timing resolution is better than 1.5 ns. The configurable shaper properties and the hardware gain calibration feature allow for adaptation to other photon detectors. The EMCAL project of ALICE [2] has adopted this electronics with 100 ns shaping time. An upgrade towards a 16-bit dynamic range for better timing and noise performance and a direct, optical readout concept [21] via 1 Gbit/s network protocols is under consideration for new applications.

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