MONOLITHIC ELECTRONICS
FOR NUCLEAR AND HIGH-ENERGY PHYSICS EXPERIMENTS

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MONOLITHIC ELECTRONICS FOR NUCLEAR AND HIGH-ENERGY PHYSICS EXPERIMENTS

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Abstract

Electronic instrumentation for large fixed-target and collider experiments is rapidly moving to the use of highly integrated electronics wherever it is cost effective. This trend is aided by the development of circuit building blocks useful for nuclear and high-energy physics instrumentation and has accelerated recently with the development of monolithic silicon chips with multiple functions on one substrate. Examples of recent developments are given, together with remarks on the rationale for use of monolithic electronics and economic considerations.

INTRODUCTION

A major consideration in the design of large experiments is the functionality and cost of the readout electronics required for the detector under consideration. Achieving large solid-angle coverage while retaining adequate granularity and detector thickness can often result in channel counts in excess of 100,000 and even 1 million for detectors of the scale required for SSC and LHC. Collider experiments have to ensure good reliability and significant self-diagnostic and calibration capabilities because of their scale, general inaccessibility for test and repair, and long operation periods; such capabilities obviously benefit fixed-target experiments also. The press to handle ever-higher luminosities to study high-mass, high- pt and/or rare processes has led to the development of fully pipelined architectures, which are extremely demanding in component count due to the deep buffers and extensive control logic required by them.

These considerations have spurred the development of multistage monolithic electronics, both in bipolar and CMOS processes, for use in present-day experiments. These circuits offer several advantages, including:

1. Reduced space requirements – this is particularly relevant both for silicon micro-strip detectors and for hermetic fine-grained calorimeters.
2. Reduced power consumption – 1 million channels operating at 1 watt/channel (typical for CAMAC or FASTBUS) for a half year costs \$0.24M at \$60/MWh and results in a severe cooling problem; CMOS electronics of considerable complexity can achieve power dissipations less than one-tenth this amount.
3. Complexity – circuits with as many as a dozen gain ranges are possible, as are circuits with shaping, discrimination, timing, calibration, and LVL-1 trigger all incorporated onto one substrate. The "wiring" is also built into the chip, improving reliability.

4. Elimination of delay cables – typical LVL-1 trigger formation delays are 500 ns, corresponding to 100 m of RG-58 coaxial cable, resulting in one mega-meter of delay cable even in a modest calorimeter of 10K channels; this can be replaced by monolithic analog storage units.
5. Cost – less than \$50/channel is possible for a system which includes preamp, shaper, discriminator, TAC, ADC, output buffer, calibration, and trigger formation; this must, however, be tempered with the observation that the development costs of monolithic electronics tends to restrict their deployment to detectors with more than 10K channels.

The development of monolithic electronics in the nuclear and high-energy physics communities stretches back well over ten years, as evidenced, for example, by the early efforts for detectors such as the PEP-4 TPC at PEP and the SLD detector at the SLC. Considerable impetus came from groups developing the necessarily miniature electronics to read out silicon microstrip detectors at FNAL and CERN, followed by the large efforts for the LEP, HERA, TRISTAN, and Tevatron detectors. The field expanded considerably as a result of the R&D efforts for SSC, RHIC, and now the LHC, with their emphasis on large channel counts, pipelined architectures, and extensive functionality. Examples are given in Table I below.

Table I. Examples of monolithic electronics developed for NP/HEP experiments

Analog Sampling/Delay Lines

LVL-1 trigger delay	DESY, Nevis, LBL, ORNL
Sampling for TPCs	LBL

Silicon-Strip Detectors

Early efforts: e.g., CDF vertex	CERN, FNAL, LBL, ORNL
Present: preamp/discr/trigger/ AMU/ADC	CERN, FNAL, LBL, ORNL

Time-Projection Chambers

Sampling for TPCs	LBL, Munich
Preamp/shaper/FADC for TEC/TRD	BNL

Calorimetry

Preamp/shapers	BNL, FNAL, CERN
Integrators	ORNL, CERN
AMU	LBL, Nevis, DESY, ORNL, CERN
TAC/ADC	Nevis, LBL, ORNL, CERN

Table I. Continued

Drift/Straw Chambers

Fast preamp/shaper/discr	U. Penn
Multihit TDC	KEK, LeCroy, CERN, U. Paris

MWPCs

Preamp/shaper/discr/latch	LeCroy/Yale
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Pad Chambers

Preamp, rad-hard devices	BNL, UT/ORNL
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Speciality

CFD/ LE discriminators	LeCroy, CERN, ORNL
Trigger summing	many
Maskable calibration	many

Superconducting (!)

Josephson junction (Nb, Nb-nitride) tunneling devices 100 GHz counters, (F)ADCs, etc.	Hypres Corp., Elmsford, NY
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EXAMPLES OF RECENT CIRCUITS

ASD-8 Chip (University of Pennsylvania)

This is a combined preamplifier, shaper, and discriminator eight-channel chip developed for drift chambers. The circuit is implemented in a bipolar process in order to achieve the very fast rise and fall times needed for precise tracking and good two-track separation using drift chambers and straw tubes. The initial development was aimed at the straw-tube tracker of the SDC experiment at SSC. A block diagram [1] is shown in Fig. 1. The design goals of the chip include:

- 0.75-ns time resolution (which translates to 50-100 microns space-point resolution, depending on chamber gas)
- Double pulse resolution of 20 ns (i.e., 1- to 2-mm spatial separation)
- Semigaussian shaping to minimize baseline shifts
- Pole-zero and ion-drift tail cancellation
- Threshold programmable from 1-10fC
- Programmable disable mask
- Power dissipation of less than 10 mW/channel
- Radiation resistance to better than 1 MRad and 10^{14} n/cm²

Figure 1: Block diagram of the ASD-8 circuit.

These goals are met. Cross-talk of order only 1% even at 1.5fC threshold has been observed. A sample pulse shape observed from an early version is shown in Fig. 2.

Figure 2: ASD-8 second shaper response for a 1fC impulse input.

TMC-Series Pipelined TDCs

A series of pipelined TDCs has been developed at KEK, again with initial impetus from the SDC-tracking effort. A diagram [2] of the architecture of these circuits is shown in Fig. 3. The concept underlying this design is to use a high-speed shift register technique to determine when a discriminator signal occurs with an accuracy of better than 1 ns. Instead of a 1-GHz clock (which would require use of GaAs technology), actual implementation in a CMOS process uses a series of rows of memory cells. Successive (logical) rows of 32 cells each are selected by a 30- to 40-MHz clock, and within the

period of this clock, the phase of the detector signal is determined by comparing it to a copy of the clock signal sent down a delay line with taps at 1-ns intervals. The pattern of fired cells thus records the times of the leading and trailing edges of the discriminator signal from the detector. This pattern is stored in a dual port memory some 128 cells deep to allow for LVL-1 trigger delay of up to 4 μ s. Patterns selected by a valid trigger for readout are encoded to conserve on the number of output pins on the 4-channel chip. Both leading and trailing edges are reported, together with a tag bit to speed up later sparse-data scans and zero-suppression operations.

Figure 3: Architecture of TMC-series of pipelined TDCs: (a) high-speed shift register; (b) time memory cell.

The chip incorporates a phase-locked loop and feedback to ensure long-term stability of the time base relative to an external global reference clock signal. This also permits a variation of the operating frequency from 30-40 MHz, which allows the same chip to be used for a range of beam-crossing frequencies as would be encountered at, e.g., different colliders. The latest versions of this chip achieve better than 0.3-ns resolution and provide LVL-1 trigger delays in excess of 4 μ s.

Broad-Range Integrator for Calorimetry

A standard problem in calorimetry at colliders is obtaining a system with 8-9 bits precision at any given point in an overall dynamic range of 18-22 bits. A novel technique for dealing with this in the case of calorimeters read out by photomultipliers has been developed at FNAL for the scintillating-tile calorimeter proposed for SDC. This technique takes advantage of the fact that photomultipliers are good current sources

over a large dynamic range and implements a current-sharing technique to achieve 10 bits of pre-ranging in a pipelined fashion. A block diagram is shown in Fig. 4 (Ref. [3]).

The anode signal from a PMT is fed into a binary-weighted, stabilized and balanced, current splitter which covers ten binary steps. The split currents are integrated on a bank of capacitors, whose voltages are then compared in parallel to a set of ten reference voltages. The resulting pattern is then encoded to form a four-bit exponent of the ultimate ADC output and simultaneously drives an analog multiplexer. The multiplexer selects the appropriate stored charge to be digitized by a FADC of 8 to 12 bits to provide the mantissa of the ADC output.

Actual implementation required use of a four-stage pipelined scheme in order to allow for settling times for the various logical steps shown in Fig. 4. The clock for the pipeline proceeds at the full speed of the SSC (or LHC) beam-crossing clock, meaning that (assuming a sufficiently fast FADC is employed) fully-digitized data from the calorimeter can be prepared with as many as 12 bits precision over a 22-bit range at the full beam-crossing rate. The utility of such a scheme in forming fast, accurate, pipelined triggers to look for jets and high- p_T particles in a collider or fixed-target environment is clear.

Figure 4: Simplified block diagram of broad-range integrator.

Electronics for a Lead-Glass Calorimeter

A pair of chips have been developed to read out a 10K-element lead-glass electromagnetic calorimeter used as the photon detector in CERN experiment WA98. Each glass block, or "module," is read out by means of its own photomultiplier tube. The anode pulses from these photomultipliers are observed to have a nearly Gaussian shape, with 10-ns rise and 10-ns fall (10% to 90% amplitude), and a small-amplitude tail extending to nearly 300 ns after the event. The measured pulse-height resolution behavior of a 3 x 3 group of modules can be described as $\frac{\sigma_E}{E} = \frac{6\%}{\sqrt{E(\text{GeV})}} + 1\%$ as measured with electrons from 0.1 to 70 GeV. The timing resolution of a module has been measured to be δ_t (rms) ~ 280 ps at 2-3 GeV for electrons.

The electronics were designed to provide the following: (1) the 300-400 ns of LVL-1 trigger delay required in WA98, without the use of delay cable, (2) digitization of the integrated anode charge with 10 bits precision over a 13-bit range, (3) timing information with 100-ps electronics resolution or better, (4) maskable calibration pulsers for the amplitude and timing channels, and (5) cluster-trigger with adjustable energy threshold. The custom part of the implementation consisted of two ASICS: a "Preamp" chip and an AMU/TAC/ADC ("ATA") chip, described below.

The principle of the pulse height measurements is to sample at a fixed frequency the output of a circuit which integrates the anode current as a function of time, and then to store those samples in a switched-capacitor analog memory unit (AMU) pending receipt of a positive trigger decision from external logic in the experiment. Receipt of a valid trigger causes a sample recorded before the event (the "pre-" sample) and one recorded following the event (the "post-" sample) to be selected in succession from the AMU, digitized by the ADC, and the results sent to the data acquisition system. The integrator output is sampled, and thus the AMU is written, at 20 MHz in the present system. The AMU is 16 cells deep, implying that the trigger decision must be received no more than 750 ns after the event.

The timing channel was designed with a constant-fraction discriminator in order to provide walk of ≤ 300 ps over a dynamic range in pulse height of 100:1. This pulse-height range is set to correspond to a photon energy range of 50 MeV to 5 GeV in actual operation. The TAC is operated in common-start mode, in order to avoid having channels rendered dead because of an uninteresting pulse occurring since the previous accepted event but prior to the next accepted event. This required development of a delay, stable to 1 part in 10^3 , for the discriminator output which could provide the 300- to 400-ns delay required for LVL-1 trigger formation.

The module cluster trigger circuits are designed to form first a disjoint set of 2×2 sums of module pulse heights, i.e., produce a single nonoverlapping tiling of the 10K modules. These 2×2 sums are then combined into a set of overlapping 4×4 sums in order to provide full trigger coverage for photons even if they should land near edges of modules or of the 2×2 regions themselves. This trigger is implemented in a current-summing mode and includes a programmable threshold controlled by a 6-bit DAC, all included on the "Preamp" chip. Board-level traces and board-to-board jumpers are included to form trigger sums which extend beyond the confines of a single ASIC or even a given electronics board.

The Preamp ASIC includes eight channels of charge-integrating amplifiers followed by x1 and x16 amplifiers which in turn drive the two AMU inputs and thus provide the dual-ranging ADC behavior desired. There are also eight timing filter amplifiers each followed by a constant fraction discriminator. Each channel has a maskable, triggerable calibration pulser set by a 6-bit on-chip DAC and is also connected to a multiplexer used for observation of the input anode signal. A block diagram of the Preamp ASIC is shown in Fig. 5.

The constant fraction discriminator uses on-chip, lumped-element shaping to remove the need for external delay cables. Auto-zeroing is included to eliminate any need for walk adjustment. The circuit is optimized for the 10-ns rise/fall times of the FEU-84 PMTs used, so these features remove the need for any adjustment other than setting a firing threshold, which is done using one of the six-bit DACs included on the Preamp

chip. The walk is observed to be ± 300 ps over the entire 100:1 dynamic range of -30 mV to -3 V for which the discriminator is designed. Over limited dynamic ranges of 6:1, walk less than 100 ps has been observed.

Figure 5: Block diagram of lead-glass Preamp chip.

The AMU is a voltage-write, voltage-read type. It operates at 20 MHz (writing) and is capable of a 25-ns write to a given cell. The typical memory "gain" of the AMU is 0.956. Cell-to-cell pedestal variation of 3.8-mV rms has been measured for a single pipeline of the prototype AMU, using addressing driven by an FPGA. This corresponds to 0.8 LSB for 10-bit operation. The measured integral nonlinearity is $\pm 0.05\%$.

The TAC range is adjustable from 50 ns to 1000 ns using an external control voltage. The noise has only been measured to date using a 10-bit ADC; all events (for in excess of 10,000 counts) fell into one ADC channel for this measurement, indicating that at

the ± 2 sigma level, the rms noise is of order 25 ps or less. The TAC resets to better than 1 LSB (out of 10-bits) in 50 ns.

The ADC is a multichannel Wilkinson-type architecture with 10-bit counters. The ramp generator and clock are shared among the 24 channels. The clock is operated at 100 MHz nominally, but has been tested successfully up to 170 MHz. The integral nonlinearity over the full range is 0.2%, and the measured differential nonlinearity is 33%, for an rms of 0.6 channels. The design allows a straightforward extension to 12 bits, faster clocks, and more channels.

COSTING CONSIDERATIONS

Development of such circuits is not for the faint of heart. A new design can require 2-3 man-years of effort, and a modification can require 2-6 man-months to complete. Better than half the effort is commonly invested in testing the design to see if it meets specifications; the capability to do this has to be built in to the design, as probing all points inside the chip is not possible. An investment in the necessary computer-aided design tools is mandatory; fortunately, public-domain tool sets exist.

Prototyping costs depend heavily on the process used. Bipolar and radiation-hard processes typically require the user to shoulder the full cost of a run, which can approach \$100K. CMOS designs which are not rad-hard can be prototyped using the MOSIS facility sponsored by DARPA, or may be handled by silicon foundries which offer such a service. The MOSIS facility is designed to encourage student participation from universities and can cost as little as a few hundred dollars for four prototype 4 mm² chips. A disadvantage lies in slow turn-around, but it does provide access to several foundries. Private commercial arrangements can cost between \$3 and \$15K for prototypes from 4 mm² to 64 mm² in size but offer reliable turn-around time.

Production costs at a silicon foundry for large runs typically have nonrecurring costs of \$30-45K for masks, \$20-30K for probing, and \$30K for a minimum set of ten 4-inch-diameter wafers. A given wafer produces anywhere from 50-400 raw die, depending on die size. However, once a project's requirements result in a need for more than ten wafers, economies of scale set in very rapidly: extra wafers cost typically \$1K each, and the number ordered can increase without limit as long as the foundry's production can keep up. Adding packaging and testing of the final product (which is mandatory, as yields for 16-mm² die, for example, are typically only 50%), the marginal cost of a die can fall to well under \$50, to as low as \$2 each (packaged) for small die. This then can lead to per channel marginal costs of \$1-\$10, which is extremely attractive, more so when the functionality per chip described above is considered.

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