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COORDINATE MEASUREMENT BY CURRENT DIVISION: A SIMPLE SYSTEM FOR MULTICHANNEL DETECTORS*

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ABSTRACT

We describe a simple current-division scheme based on an unorthodox connection of the 733 amplifier chip: a common-base input can be obtained by grounding the normal inputs and driving a gain control terminal. With ganged multiwire chambers detecting cosmic rays in the proportional region, under realistic operating conditions (charge ~ 2 picocoulombs per gap), the resolution is about 2% of full length. Calibration procedures are discussed, as are the effects of changing various parameters.

I. INTRODUCTION

In the current-division method (see Fig. 1 for a rough illustration) the coordinate x of a track is measured by observing the current pulses (L,R) observed at the left and right ends of the sense wires. Specifically, x is proportional to the asymmetry

$$A \equiv \frac{L - R}{L + R}$$

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of the pulses. The error in x depends on electronic noise which limits the accuracy of the pulse-height measurement. This fact has stimulated considerable development of low-noise, low-input-impedance preamps. (The input impedance must be low compared with the resistance of the sense wire or, when N wires are ganged for economy, with the parallel equivalent resistance.)

In a large multichannel detector, however, the noise of the "bare" preamp is outweighed by other effects:

1. Thermal noise in the ganged sense wires.
2. Additional noise introduced by the capacitance of the ganged sense wires, and by cables from the sense wires to the preamp.
3. Ambient electrical noise.
4. Residual (after filtering) time structure of the pulses which may lead to an effective difference in the times at which L and R are sampled.
5. Reduction of the observed asymmetry by the finite high-voltage coupling capacitors (if required).

We are considering a modest addition to the MAC detector, namely adding current-division to the outer (muon tagging) calorimeter, to aid in tracking. In connection with this we decided to re-examine the current division problem and go to the limit in preamp simplicity: A single 733 chip which could then easily be mounted right on the detector. We have tested this idea using cosmic rays in a small, self-triggered bench system. The circuits are indeed simple and inexpensive and seem

to perform comparably to more complicated systems (given the overall requirements).

II. APPARATUS

A. Detector

Figure 1 shows the detector used in the tests. Coincidences were observed between four layers, each consisting of four cells ganged horizontally. The chambers are the same ones used in the MAC calorimeter, made from closed aluminum extrusions. Other relevant parameters are:

Sense Wire:	40 micron stainless steel
	Effective length 88" (223 cm)
	Resistance 1400 ohms (for four wires, 350 Ω)
Gas:	80% Argon, 20% CO ₂
HV:	2100 volts

Under these conditions, minimum ionizing tracks at approximately normal incidence gave a total charge of about 1.5 pcoul (most probable) or 2 pcoul (mean) per gap crossing (Fig. 8).

B. Electronics

1. Overview

We wished to use the data-logging hardware and software which had been developed for MAC, but without changing critical settings (such as gate width and timing); therefore, the apparatus for these tests was made reasonably self-contained. Figure 2 is a simplified diagram of the front end electronics and Fig. 3, of the remainder. Eight preamps (four L and four R) drove eight receivers through about 120 ft. of

bundled, group-shielded twisted pair. (Preamp details are given later.) Since this is a high-gain system (roughly 0.5 volts per picocoulomb) it is prone to oscillate and care must be taken with the shielding. In particular, we found that when individually shielded twisted pairs were used to connect to the main cable, it was important to connect the shield directly to the ground of its own preamp.

We wished to trigger on the coincidence of at least two hits so L and R were added for each layer and the four sums were fed to four timing discriminators. These used leading-edge extrapolation¹ to generate a timing pulse, 50 nsec wide and starting 400 nsec after the front corner of the signal pulse (see Fig. 4). Each discriminator output was a current source, so by simply bussing these into a common resistor one obtained a quantized signal whose level reflected the multiplicity of the coincidence.

The entire MAC data-logging machinery depends on the periodic nature of the PEP ring (interactions in a given region occur every 2.44 microseconds). In order to use this machinery we therefore simulated the PEP cycle (that is, generated appropriate gates and strobes every 2.5 microsec) and lay in wait for cosmic rays which happened to occur at the right time so that the pulses were sampled correctly (with a total tolerance of 50 nsec, the discriminator output width). This reduced the counting rate by a factor $2500/50 = 50$ but still left about 1.5 Hz which was sufficient. A "multipurpose" NIM module (Fig. 3) was built to do all this. Figure 4 shows the timing of various pulses. A 400 kHz oscillator triggered A and B gate generators (the sample-and-hold is described later) and a strobe pulse. The discriminator multiplicity

level was fed to a comparator set to the desired multiplicity, whose output was sampled at strobe time; if true, a dead time generator was triggered whose output killed all following gates (to prevent the latched data from being erased) and alerted the data-logging machinery (to read the data).

Timing monitor output was found to be essential in setting up the gate timing. This consisted of markers for the "PEP" pulse, the strobe, and the trailing edges of the A and B gates all superimposed on any desired analog output pulse from the receivers. This single signal is then delayed in a long cable and observed on a scope triggered by the "event" pulse (which came after the gates of interest!). Correct timing is shown in trace of Fig. 4: The A and B gates are adjusted to equal width (to minimize sample-and-hold pedestal drift), and the strobe is set so that the A gate closes shortly before the pulse starts to rise (in the worst case) and the B gate closes shortly after the pulse starts to fall (in the worst case). (If the B gate closed at the top of the pulse, the setup would be rather sensitive to time structure.)

Finally, a linear pulse generator was provided to calibrate the gains of the eight channels. This pulser was in random phase with the "PEP oscillator", so the pulser data accurately reflected the effect of the 50 nsec timing window on a monochromatic pulse.

2. Preamplifier and Receiver

Figure 5 shows the preamp and receiver. The 743 is a video amplifier, popular, inexpensive and evidently very rugged. It uses an emitter-coupled front end followed by a balanced output stage.

The usual inputs are the bases (pins 1 and 2) of the input transistors. However, pins 4 and 9, normally used to set the gain by means of emitter degeneration, connect directly to the emitters of the input transistors. Therefore one can ground pins 1 and 2 and drive either 4 or 9 to obtain a current amplifier with common-base input. Noise is about the same as that of a bipolar transistor at the same collector current (3 mA) and input impedance is about 20 ohms. In our setup noise was dominated by the detector (equivalent resistance of the ganged sense wires was 350 ohms) since the "white" baseline noise observed on a scope approximately doubled when the sense wires were connected to the input.

The preamp is further simplified by using the output leads to supply power. This saves putting in low-voltage buses at the detector, and does not increase the component count since decoupling would be needed anyway.

The receiver consists of another 733, used conventionally. The current signal is integrated by the capacitor between pins 1 and 2 (in conjunction with the 1K resistors). In our tests, since we were dealing only with minimum-ionizing pulses covering a small dynamic range, we jumpered for the maximum gain of 400. However, the 733 only puts out 2.5 volts, so if the data cover a large dynamic range (as in a shower counter) it might be necessary to reduce the gain and add a post-amp capable of driving the sample-and-hold to a greater voltage. Incidentally, the overall gain spread in our eight nominally identical channels was about 20%. This was corrected in the offline analysis without any attempt to trim it out in the hardware.

The preamp/receiver features balanced drive and reception

of the long (120 ft.) cables from the detector to the counting house, without transformers.

3. Sample-and-Hold

We used our standard MAC sample-and-hold circuits (shown somewhat simplified in Fig. 6). This circuit puts out the difference between voltages stored on two memory capacitors, one disconnected from the input just before the pulse starts and the second, shortly after the pulse peaks. This arrangement has many advantages: cancellation of gate feedthrough and hence, excellent pedestal stability; cancellation of parasitic currents and hence, very slow "droop" and; perhaps most important here, sensitivity only to baseline changes between times A and B. (In effect, low frequency noise is eliminated.)

III. PROCEDURE AND ANALYSIS

A. General

Four data sets were taken, each an overnight run with about 20,000 events. Conditions are summarized below:

<u>Run #</u>	<u>Coupling Capacitance</u>	<u>Integrating Capacitance</u>	<u>Cable Shielding</u>
2279	0.01 μ f	200 pf	group shielded
2282	0.02 μ f	200 pf	group shielded
2301	0.02 μ f	200 pf	individual shielding to main cable, then group
2303	0.02 μ f	100 pf	individual shielding to main cable, then group

We used a twofold (rather than fourfold) coincidence requirement to obtain some events where any given chamber was missed by the

track, in order to sample "pedestal" outputs (gate but no signal).

Calibrate pulser runs were taken in conjunction with each data run, to determine the effective preamp gains as discussed below.

To simplify the following discussion, we summarize very briefly the main differences between the runs: going from 0.01 to 0.02 μ farad coupling capacitors widened the asymmetry distribution as expected, and improved the resolution; changing the shield configuration visibly improved the rejection of external interference but had no measurable effect on the resolution and finally, reducing the integrating capacitor (and shortening the gates a bit to correspond) further increased the width of the asymmetry distribution but tended to spoil the resolution, as well as requiring more critical timing (because of the faster pulse fall-off). Thus the middle two runs are the best and our performance statements refer to them.

B. Determination of Constants

The circuit theory of current division is developed in the Appendix. Viewed simply as a mathematical problem, it can be parametrized in various ways, but separating the effects into those caused by the electronics (gain and pedestal) and those caused by "extra" impedances around the current-division loop gives the best physical insight into the problem.

1. Pedestal and Gain

First we wish to correct the data so that they appear to be coming from ideal charge amplifiers (equal gain and no gate feedthrough). The corrected charge (arbitrary units) is

$$\text{charge} = \text{gain} * (\text{data} - \text{pedestai}) \quad (1)$$

The pedestal for each channel may be read from a histogram of the raw data (Fig. 7) plotted on a very fine scale (2 mV/channel). The width ($\sigma_{\text{PED}} \approx 9$ mV) is dominated by thermal noise in the sense wires.

The eight gains are measured with pulser runs in which equal charges are fed to each preamp; to explore the entire dynamic range these charges are increased in linear increments. The computer histograms the sum ($L + R$) and the asymmetry $(L - R)/(L + R)$ for each event. The sum distributions are broad because of the 50 nsec timing range combined with the fact that the pulses are sampled on the falling slope. Also, the raw sum distributions appear at different values for each L,R pair because of different preamp gains. The raw asymmetry distributions are narrow (because the timing jitter cancels out to first order) but are not peaked at zero (because of different gains). If $\text{oldsum}(i)$ and $\text{oldasy}(i)$ are respectively the sum and asymmetry observed for the uncorrected data, and if xnorm is an arbitrary number then letting

$$\begin{aligned} \text{gain}(L) &= \text{xnorm} / (1. + \text{oldasy}(i)) / \text{oldsum}(i) \\ \text{gain}(R) &= \text{xnorm} / (1. - \text{oldasy}(i)) / \text{oldsum}(i) \end{aligned} \quad (2)$$

will yield corrected data with all pairs having sum xnorm and asymmetry zero. (The sum distributions will still be broad, but will all look the same.)

2. Corrections for Extra Resistance

If the effects of the coupling capacitors and the preamp input impedance were negligible the coordinate x (Fig. 1) would be given by:

$$x = -L * A \quad (3)$$

where $L \equiv$ half-length of the sense wire, and $A \equiv$ observed pulse asymmetry. (The minus in Eq. 3 arises from a conflict between the conventional definitions of increasing x and of asymmetry.) If these extra impedances are present, we show in the Appendix that (assuming the preamp input impedances are equal) the coordinate is still linear in the asymmetry:

$$x = -L (a + b * A) \quad (4)$$

where $a = (s - u)/\rho$

$$b = 1 + (2t + s + u)/\rho \quad (5)$$

s = equivalent resistance due to left coupling cap.

u = equivalent resistance due to right coupling cap.

t = preamp input resistance.

ρ = parallel equivalent resistance of the ganged sense wires

(here $1400/4 = 350$ ohms).

In practice, constants a and b are determined empirically, than the values of s , u , t may be inferred. The best way would be to fit tracks already measured by an independent, high-resolution detector. Though this can usually be done in a full-fledged experiment, in our test we had no such detector. Therefore we used the observed cutoff points of the raw asymmetry spectrum from each plane, to infer a and b for that pair. If the half-max points of the asymmetry for the i^{th} plane are $a_{\min}(i)$, $a_{\max}(i)$ then using in Eq. 4

$$\begin{aligned} b(i) &= 2. / (a_{\max}(i) - a_{\min}(i)) \\ a(i) &= -1. - b(i) * a_{\min}(i) \end{aligned} \quad (6)$$

will yield an observed x distribution running from $-L$ to L .

In principle one ought to use the end points of the asymmetry distribution in Eq. 6 and the correct half-length in Eq. 3; in practice the end points are hard to determine so we use the asymmetry at half max and an "adjusted" half-length in Eq. 3. Thus the final x distribution (Fig. 10) fits our preconceptions about the dropoff of cosmic-ray acceptance at the ends, combined with the finite x resolution, but we cannot be sure the length scale is exactly right. To put it another way, our apparatus by itself cannot measure the precise effective length of the sense wires; for that, an independent detector or a well-collimated beam would be necessary. But none of this appreciably affects our conclusions about resolution.

B. Event Analysis

After the constants were determined by a few initial passes through the data, the runs were finally analyzed processing each event as follows:

1. Raw data were converted to charge (pedestal and gain corrections).
2. Charge sum and asymmetry were calculated for each L-R pair.
3. Events with sum less than 0.1 volt or greater than 2 volts (for any pair) were dropped. 60% of the events survived this cut.
4. For surviving events, four x coordinates were calculated (Eq. 3).
5. A straight line $x(z)$ was fitted to the four points.

The residual ($x_{\text{meas}} - x_{\text{fit}}$) was calculated for each point.

IV. RESULTS AND DISCUSSION

Figure 8 is a histogram of the charge sum (L + R) averaged over the four chambers. Events above 2.0 were cut out of the track analysis because they probably correspond to a very large track angle or an energetic delta-ray traveling down the chamber.

Figure 9 shows the final calibration functions determined for the runs with 0.01 and 0.02 microfarad coupling capacitors, to give an impression of the overall scatter of the eight channels. The ideal calibration function is also shown (dashed line). The scatter is probably due to the spread in values of the coupling capacitors. The fact that the constant is always negative is probably an artifact; it could be caused (given our analysis procedure) by the calibrate pulse to one end being systematically a little larger than the other end. This would not affect our conclusions about resolution.

Figure 10 is the distribution of the x intercept of the fitted tracks (i.e., the best value of the track position at the center of the stack). As mentioned earlier, the length scale may be slightly wrong.

The residuals for each chamber were checked for shifts with x which would indicate that the a 's and b 's were wrong. Such shifts were found to be considerably smaller than the width of the residual distribution. A typical residual distribution (chamber 2, residuals for the entire length of the chamber) is shown in Fig. 11 (1 cm/channel). In general, the residuals tend to be smaller for chambers 1 and 4 (because straight-line fits to few data points tend to favor the ends),

and they are about 1.8 times narrower near the center of each chamber than at the ends. Error analysis leads one to expect a factor $\sqrt{2} = 1.4$.

Finally, the most efficient measure of the scatter in a parent population after one has performed a linear fit is

$$\sigma = \sqrt{\frac{\sum x^2}{n-2}} = \sqrt{\frac{\sum \bar{x}^2}{2}}$$

and Fig. 12 is a histogram of this quantity, taken over the entire chamber. (As with the residuals, it is about 1.8 times sharper in the middle of the chamber than at the ends.) The quoted resolution of the system is somewhat a matter of taste; if one takes the most probable σ , it is 1.75/220 or slightly better than 1%; if one takes the average σ , it is 5.44/220 or 2.5%.

We should mention that the observed width of the pedestal distributions (a measure of system noise), the width of the pulser asymmetry distributions, and the width of the observed residuals are all more-or-less consistent, about 2.2%, and this agrees fairly well with the mean σ .

V. EFFECT OF COUPLING CAPACITORS

After taking some preliminary data with 0.01 microfarad coupling capacitors it became obvious that the effect of extra resistance was large: The raw asymmetry distribution was far narrower than the ideal $(-1 \rightarrow 1)$. Since from direct measurements (i.e., observing the output while shunting the preamp input with test resistances) we knew that the preamp input impedance was only 20 ohms, we suspected the coupling

capacitors. This was confirmed by running with two different values and extrapolating the results to infinite coupling capacitance.

In the Appendix we show that the coupling capacitance can be replaced by an effective impedance τ/C where C is the capacitance and τ is an effective time having to do with pulse time structure, integrating filter time constant and sampling time. Rewriting Eq. 5 in terms of the average effective impedance $(s + u)/2$ we obtain

$$b = 1 + (2\tau + 2(s + u)/2)/\rho$$

$$= 1 + (2\tau + \tau/C)/\rho$$

$$\text{or} \quad b = (1 + \frac{2\tau}{\rho}) + (\frac{2\tau}{\rho}) \frac{1}{C}$$

In the following Table we list b as a function of $1/C$, entering the average of the four b 's obtained for each of the two runs (2279 and 2282):

<u>C</u>	<u>1/C</u>	<u>$\bar{b}$</u>
0.01	100	1.85
0.02	50	1.47
(∞)	0	(1.09)

From the extrapolation:

$$1.09 = 1 + 25/s$$

$$\tau = 350 * (1.09 - 1)/2 = 16 \text{ ohms}$$

in fair agreement with the measured preamp impedance, and

$$(1.85 - 1.47)/50 = 2\tau/\rho$$

$$\tau = 350 * 0.0076/2 = 1.33 \text{ } \mu\text{sec}$$

Thus the coupling capacitors dominate the situation. Furthermore, this effect is intrinsically nastier than the preamp impedance because coupling capacitors have wide tolerances (typically -20%, +80%) and because the time structure of the pulse, which could change from event to event, enters in. Coupling capacitors cannot be made arbitrarily large because they are bulky, expensive and hard to get in sizes larger than 0.01, and because the greater energy stored in the capacitor is more likely to destroy the preamp in the event of a momentary arc. The value 0.02 seems a reasonable compromise for this system.

In principle the coupling capacitor can be avoided by running the sense wires at DC ground and the surrounding structure at high voltage. While this is quite practical for small detectors or geometries where field-shaping wires are used rather than solid structures, in the case of really large conventional detectors it becomes extremely awkward mechanically. In addition, the decoupling capacitors which are then required from the outer case to ground, to provide a pulse current return, tend to introduce crosstalk between channels. Also, a separate, high-voltage insulated wire must then be provided from one end to the other of the sense wire to complete the current-division loop. (Although MAC was originally built with the shower chamber sense wires at ground, these effects proved sufficiently annoying that we changed over to sense wires at HV.)

VI. ACKNOWLEDGEMENT

The test used hardware and software set up by many members of the MAC collaboration. In particular, however, thanks are due to Jim Johnson for setting up the test chambers, and Henry Band and Dave Wiser for help with the programs.

APPENDIX

Circuit Theory of Current DivisionA. Effect of Coupling Capacitors

If the detector wires are at high voltage, one must use coupling capacitors, and there are good reasons for keeping them small. In this case, they tend to reduce the pulse asymmetry below what would be observed in an ideal setup. In this section we show that this effect can be replaced by an equivalent impedance. Consider the setup shown below:

Here we have assumed that the coordinate x is measured from the center of the wire; also, L is the half-length of the wire, R its half-resistance,

C is the coupling capacitor at each end, and r the preamp input resistance. y is the current flowing to the right and i, the total pulse current. Kirchoff's current law is taken care of by the notation, and the voltage law gives

$$y(1 - \frac{x}{L})R + \frac{1}{C} \int_0^t y dt + yr = (i-y)(1 + \frac{x}{L})R + \frac{1}{C} \int_0^t (i-y)dt + (i-v)r$$

Physically, the capacitance effect arises as follows: Suppose that more current is flowing to the right so that one would expect the right amplifier signal to be larger. This also implies that the right capacitor will be charging faster than the left one; this creates a reverse voltage (represented by the integrals in the equation) which will be larger on the right than on the left, tending to reduce the asymmetry. The total effect has something to do with: (a) The time structure of the current pulse itself; (b) The time at which the amplifier output is sampled and (c) Any additional filtering in the amplifier. However, assuming these effects are the same at the two ends, one can replace the integrals:

$$\frac{1}{C} \int_0^t y dt \rightarrow y(\frac{\tau}{C}) = y r_c$$

$$\frac{1}{C} \int_0^t (i-y)dt \rightarrow (i-y) \frac{\tau}{C} = (i-y) r_c$$

Here τ is an effective time: the form factor of the pulse. It is clear that it would be difficult to predict τ from first principles, and probably somewhat useless since actual pulse shapes are ragged and would be very hard to parametrize. Empirically, we simply want to find a value

of C which is large enough to reduce the effect to an acceptable level; this is best done by experiment.

B. Full Analysis

Our diagram reflects the fact that several (N) wires may be ganged, and that the various "parasitic" resistances may be different at the two ends. s , u represent the effective resistance of the capacitors, and t , v that of the preamps; otherwise the notation is the same as before. We now have two voltage equations:

$$y(1 - \frac{x}{L})R + z(u+v) = (i-y)(1 + \frac{x}{L})R + (i-z)(s+t)$$

$$z(u+v) = (y-z) \frac{2R}{N-1} + (i-z)(s+t)$$

Solving these for the current z :

$$z = \frac{N(s+t) + (1+x/L)R}{N(s+t+u+v) + 2R} \cdot i \quad (1)$$

z is related to the observed asymmetry by

$$A = \frac{L-R}{L+R} = \frac{(i-z)t - zv}{(i-z)t + zv}$$

which can be rewritten

$$z = \frac{t(1-A)}{(1+A)v + (1-A)t} \cdot i \quad (2)$$

Combining Eq. 1 and 2 we find x in terms of the asymmetry:

$$\frac{x}{L} = \frac{((N/R)(s+t+u+v) + 2) t(1-A)}{t+v+A(v-t)} - \left(\frac{N}{R}\right)(s+t) - 1$$

The preamp input impedances are small to begin with and it is an excellent approximation to take them equal ($t = v$) in which case x is linear in the asymmetry:

$$x = -L(a + bA)$$

with coefficients

$$a = (s - u)/(2R/N)$$

$$b = 1 + (2t + s + u)/(2R/N)$$

That is, a reflects the difference between the capacitive effects while b reflects the sum of all parasitic effects around the loop. ($2R/N$ is just a complicated way of writing the parallel equivalent resistance of the N wires, $2*700/4 = 350$ ohms in our setup.) In the main text, we

interpret the observed b coefficient to show that our results with two different coupling capacitances are consistent. It would be unwise to take the a coefficient too seriously since a very small error in determining the preamp gains could result in a fictitious value of a.

REFERENCE

1. R. Pittschalk, "Timing Discriminator Uses Leading-Edge Extrapolation," Nuclear Instruments and Methods, to be published.

FIGURE CAPTIONS

- Fig. 1. Proportional chamber setup used in tests.
- Fig. 2. "Front-end" electronics (block diagram).
- Fig. 3. Multipurpose NIM module (simplified diagram).
- Fig. 4. Timing diagram.
- Fig. 5. Preamplifier and receiver.
- Fig. 6. Sample-and-Hold circuit (one channel).
- Fig. 7. Distribution of "pedestal" outputs in channel 5, that is, events in which that channel was not hit by the track. Width is dominated by "white noise" in the preamplifier, including the contribution of the ganged sense wires. True signal peak would appear far to the right.
- Fig. 8. Distribution of the average sum (left plus right) of the charge received by the eight preamps. (Raw data have been corrected so that the four pairs have the same gain.)
- Fig. 9. Measured calibration functions (see text) for 0.01 fd and 0.02 fd coupling capacitors. Also shown are the functions that would be obtained for infinite coupling capacitance (preamp input impedance remains) and in the ideal case (no extra impedances).
- Fig. 10. Distribution of the x intercept of fitted tracks (coordinate where the track crosses the midplane). The overall length normalization is somewhat arbitrary (see text).
- Fig. 11. Distribution of the difference between the measured and fitted track for the second chamber down, including data from the entire chamber.
- Fig. 12. Distribution of the estimate of the standard deviation from each event, also over the entire chamber.

Fig. 1

Fig. 2

Fig. 3

Fig. 4

Fig. 5

Fig. 6

Fig. 8