A nuclear life-time measuring system has been constructed using solid state semiconductor radiation detectors at (near) room temperatures for measurements in the sub-nanosecond range by the method of "self-comparison centroid shift" (31). The data is accumulated with the aid of a computer controlled pulse height analyser operating in shared real-time access mode (85), on-line with three other independent experiments. An overlap Time-to-ampitude Converter has been developed operating in the range 0 - 90 ns; the differential linearity is ± 4%, which is an improvement on that of a borrowed commercial T.A.C. operating in the same range. A method of calibrating T.A.C.s with the aid of signals from a sampling oscilloscope has been developed.

The nuclear life-time investigated is the $f^{5/2}$ excited state of Pb$^{207}$ (from Bi$^{207}$). This life-time has been measured a number of times previously, the most recent result being $129 ± 1$ ps (34); we have measured it to be $128 ± 90$ ps. This large statistical error is due to the very weak source used and to the relatively poor time resolution of solid state detectors in this context.

In the final experimental system, a thin fast transmission detector picks off the required time information to be fed to the T.A.C. whilst a thick detector immediately behind provides the required energy information on the residual radiation using slow coincidence techniques. If necessary, slow signals from the thin and thick detectors could be recombined to provide (nearly) the original energy resolution when dealing with complicated decay schemes. To permit a wide range of energy to be accepted in the thin detectors, yet with good time resolution maintained, pulse height compensation to the output of the fast T.A.C. is provided.

A theoretical introduction and a review of the electronic techniques and experimental problems in measuring short nuclear life-times, in particular using solid state detectors, is included. The thesis concludes with an appraisal of the whole system and suggestions for further modifications. An appendix contains the relevant computer programs written in PAL 3 for on-line operation of the pulse height analysis system.
THESIS

submitted by

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for the degree of

Doctor of Philosophy

University of Edinburgh

1972
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Summary

A nuclear life-time measuring system has been constructed using solid state semiconductor radiation detectors at (near) room temperatures for measurements in the sub-nanosecond time range by the method of "self-comparison centroid shift". The data is accumulated with the aid of a computer controlled pulse height analyser operating in shared real-time access mode, on-line with three other independent experiments. An overlap Time-to-Amplitude Converter has been developed operating in the range 0 - 90 ns; the differential linearity is ±4%, which is an improvement on that of a borrowed commercial T.A.C. operating in the same range. A method of calibrating T.A.C.s with the aid of signals from a sampling oscilloscope has been developed.

The nuclear life-time investigated is the $f_{5/2}$ excited state of $^{207}$Pb (from $^{207}$Bi). This life-time has been measured a number of times previously, the most recent result being $129 \pm 1$ ps $^{(34)}$, we have measured it to be $128 \pm 80$ ps. This large statistical error is due to the very weak source used and to the relatively poor time resolution of solid state detectors in this context.

In the final experimental system, a thin fast transmission detector picks off the required time information to be fed to the T.A.C. whilst a thick detector immediately behind provides the required energy information on the residual radiation using slow coincidence techniques. If necessary, slow signals from the thin and thick detectors could be recombined to provide (nearly) the original energy resolution when dealing with complicated decay schemes. To permit a wide range of energy to be accepted in the thin detectors, yet good time resolution maintained, pulse height compensation to the output of the fast
T.A.C. is provided.

A theoretical introduction and a review of the electronic techniques and experimental problems in measuring short nuclear life-times, in particular using solid state detectors, is included. The thesis concludes with an appraisal of the whole system and suggestions for further modifications. An appendix contains the relevant computer programs written in PAL 3 for on-line operation of the pulse height analysis system.
# CONTENTS

## CHAPTER 1  INTRODUCTION

1:1 Importance of Life-time Measurements  1
1:2 Models of the Nucleus  3
1:3 Probability of Emission for Electromagnetic Radiation  4
1:4 Single Particle Estimation for Electromagnetic De-excitation  6
1:5 Present Experiment  10

## CHAPTER 2  LIFE-TIME MEASUREMENT OF NUCLEAR EXCITED STATES

2:1 Introduction  12
2:2 Method of Delayed Coincidence  13
2:3 Coincidence Circuit  14
2:4 Single Channel Delayed Coincidence Technique  17
2:5 Multichannel Time Analyzers  19
2:6 Time-to-Amplitude Converters  21
2:7 Auxiliary Circuits  26
2:8 Derivation of Timing Signals  30
2:9 Methods of Time Pick-off  34
2:10 Radiation Detectors for Timing Experiments  36
2:11 Present Life-time Measuring System  41

## CHAPTER 3  ORIGINAL EXPERIMENTAL SYSTEM

3:1 Introduction  45
3:2 Detectors, Housing and Vacuum System  46
CHAPTER 4  PULSE HEIGHT ANALYSIS SYSTEM

4:1 Introduction 55
4:2 Hardware 56
4:3 Interfacing of NS 627 to PDP-8 56
4:4 Supervisor Program 57
4:5 Simple Pulse Height Analysis Program 58
4:6 Auxiliary Programs for Pulse Height Analysis System 62

CHAPTER 5  EXPERIMENTAL WORK

5:1 Preliminary Experimental Work 64
5:2 The Attempt to Measure the Half-life of the First Excited State of Pb$^{207}$ with the Original Experimental System 71
5:3 Modification of the Original Experiment 75

CHAPTER 6  RESULTS AND CONCLUSIONS

6:1 General Observations on the NS 627 / PDP-8 Pulse Height Analyser 85
6:2 Measurement of the Life-time of the $f_{5/2}$ State of Pb$^{207}$ 86
6:3 Further Modifications and Improvements to the Present System 93

REFERENCES
APPENDIX

Pulse Height Analysis Control Programs

Numerical Data PUNCH Output from Computer Memory

Subroutine to Calculate CENTROID of Single Peaked Time Spectrum
There are two main measurable parameters for all radioactive decay processes, the disintegration energy ($E$) and the disintegration constant ($\lambda$). Accurate experimental measurements of these quantities provide the basic means of testing the validity or otherwise of the various theoretical models of the nucleus. In this thesis we are concerned with the measurement of short nuclear life-times within the region $10^{-9}$ to $10^{-12}$ second (the mean life of a disintegration being the inverse of the disintegration constant).

The majority of nuclear states which decay with this order of life-time do so by electromagnetic transitions. These electromagnetic modes of decay are:

1. Gamma emission, the emission of an electromagnetic photon,
2. Internal Conversion of an orbital electron, and
3. Internal Pair Production.

Internal Conversion occurs because of the finite probability of an orbital electron being near the nucleus; then the excitation energy is passed to the electron, which is emitted with the decay energy minus the binding energy of the orbital electron. We define the internal conversion coefficient as

$$ a = \frac{N_e}{N_\gamma} \quad (1) $$

where $N_e$ is the number of conversion electrons emitted per second and $N_\gamma$ is
the number of gamma-rays emitted per second. The values of \( a_K, a_L, a_{\\lambda\nu}, \) etc., for the different orbital shells are similarly defined, their sum being the total internal coefficient \( a \). Theoretical estimations of internal conversion coefficients have been made by Rose\(^1\), Sliv and Band\(^2\), and Hager and Seltzer\(^3\). The experimental determination of internal conversion coefficients is of particular importance in nuclear spectroscopy as they provide one of the few methods of estimating the multipolarity and type of a transition - by comparison of the conversion coefficients for the different atomic shells.

Internal Pair Production is only possible if the energy available for decay is greater than the 1.02 Mev energy required to create an electron-positron pair.

Another mode of decay from states with life-times of the order of interest to this project is that of "long-range" alpha particle emission. In a very few cases where the ground state of a nuclide, such as Po\(^{212}\), has a very short mean life \( (3\times10^{-7}\text{ sec.}) \) before decaying by alpha emission, in this case to Pb\(^{208}\), there is a finite probability of alpha decay competing with gamma emission from the excited states of the parent nuclide. These particles then are emitted with the excitation energy plus the energy normally available for the groundstate alpha particle emission. In such a case, where two or more emissions are competing, the total transition probability (which is the only parameter experimentally measurable) is the sum of the individual partial transition probabilities, viz.,

\[
\lambda = \lambda_\alpha + \lambda_y
\]  

\( \lambda \) Knowing the branching ratio \( \frac{\lambda_\alpha}{\lambda_y} \), this equation may be used to determine the individual partial decay constants. In electromagnetic decays, the
reduced transition probability depends sensitively on the wave-functions of the initial and final states; thus experimental measurements of λ and comparison with theoretical predictions on the basis of any particular model should yield valuable information about these wave-functions.

The quantity of most use is the "enhancement" or "retardation" factor, which is the ratio of the experimental half-life to the theoretically predicted value from the "single particle" estimate. Deviation from unity of this ratio must be explained by some acceptable alternative model of the nucleus, which may lead to additional selection rules to account for otherwise unexplained retardation or (in the case of some E2 transitions) enhancement of transition rates.

Two main models are used to describe the nucleus - the Shell (or Hartree) model(4) and the Collective model(5). The Nilsson model(6) describes a compromise model between the two extremes to explain the properties of nuclides exhibiting a mixture of both structures.

The Shell model envisages the nucleons, only weakly coupled to one another, moving in orbits with characteristic quantum numbers $l^2$(orbital angular momentum) and $s^2$(spin $=\pm \frac{1}{2}$) in a spherical potential well due to all other nucleons. According to Pauli's exclusion principle, the energy levels(characterised by the quantum numbers) are filled successively with protons and with neutrons independently. The "magic numbers" denote shell closures and the central potential is arranged to accommodate these experimentally
determined values. When an excited nucleus falls to a lower state, only one nucleon is considered to release angular momentum and change orbit. This model satisfactorily accounts for the properties of the groundstate and lower excited states of many nuclides in the neighbourhood of closed shells, but fails in attempting a description of the excited states for nuclides in the rare earth region and for those of high "Z". To account for the properties of these, Bohr and Mottelson treat the nucleus as an incompressible liquid drop, in which energy is stored in the form of vibrational and rotational states. Here the nucleons strongly interact and move in collective orderly displacements. Low-lying states of excited nuclei with large deformations exhibit the low frequency rotational motion, called "rotational bands". Nuclei of nearly spherical shape display vibrational levels; these correspond to waves moving across the surface of the nucleus. De-excitation of vibrational and rotational states corresponds to a change in frequency of the collective oscillations of all the nucleons. The Nilsson model attempts to predict the energy levels of single particle orbits in a deformed potential well corresponding to surface waves generated by the remaining nucleons, as in the Collective model.

1:3 The Probability of Emission for Electromagnetic Radiation

Electromagnetic radiations are classified into Electric (E) and Magnetic (M) transitions of Multipole Order $\ell^2$, where $\ell$ is the angular momentum carried off by the emitted photon. Conservation of angular momentum and parity lead to the following selection rules for allowed transitions:
(a) \[ |(I_1 - I_F)| < 1 < I_1 + I_F \]  

where \( I_1 \) is the angular momentum of the initial state and \( I_F \) is the angular momentum of the final state.

(b) \[ \Delta \pi = \pi_1 \times \pi_F = (-1)^{\Delta \pi} \] for \( E1 \) transitions  
    \[ = (-1)^{\Delta \pi} \] for \( M1 \) transitions

where \( \Delta \pi \) is the parity change, \( \pi_i \) is the parity of the initial state and \( \pi_f \) is the parity of the final state.

The transition probability for any electromagnetic radiation is given by (7)

\[ \lambda(k)_{E1} = \frac{8 \pi (2l+1)}{L[(2l+1)!!]} \frac{1}{\hbar} k^{2l+1} \lambda_0(k; I_1 \rightarrow I_F) \]  

for Electric (\( E1 \)) transitions and

\[ \lambda(k)_{M1} = \frac{8 \pi (2l+1)}{L[(2l+1)!!]} \frac{1}{\hbar} k^{2l+1} \lambda_0(M2; I_1 \rightarrow I_F) \]  

for Magnetic (\( M1 \)) transitions, where

\( L = \) angular momentum of the emitted photon,

\( (2l+1)!! = 1 \cdot 3 \cdot 5 \cdots (2l+1) \)

\( k = \) wave number of the emitted photon,

\( \lambda_0 = \) reduced transition probability for the radiation defined in the subsequent brackets.
\[
\begin{align*}
\left( \frac{\partial^2 n}{\partial z^2} \right)_{E^2} - \frac{E^2}{E_0} \left( \frac{\partial^2 n}{\partial z^2} \right)_{E_0} &= A \\
\left( \frac{\partial^2 n}{\partial z^2} \right)_{E^2} - \frac{E^2}{E_0} \left( \frac{\partial^2 n}{\partial z^2} \right)_{E_0} &= B
\end{align*}
\]

For electrostatic radiation and

\[
\left( \frac{\partial^2 n}{\partial z^2} \right)_{E^2} - \frac{E^2}{E_0} \left( \frac{\partial^2 n}{\partial z^2} \right)_{E_0} = C
\]

Results:

The single particle expression for the electrostatic field excitation

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The single particle expression for the electrostatic field excitation

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for Magnetic radiation, where
\[ \lambda_0 = 10^{24} \text{ sec}^{-1} \]
\[ k_0 = 10^{-13} \text{ cm} \]
\[ E_0 = 197 \text{ MeV} \]
\[ E = \text{the decay energy} \]
\[ R = 1.2 \times 10^{-13} \text{ cm} = \text{nuclear radius} \]

In the above equations \( \lambda_{M1} \) and \( \lambda_{M2} \) refer, of course, to the partial decay constants, only and correction must be made when competing processes occur.

Fig. 1 shows a plot of the expected mean lives of M1 and E2 gamma transitions from Mosskowski’s estimation.(9)

From experimental results(10), it would appear that all E1 life-times are longer than theoretically predicted, varying between a hindrance of 1 and \( 10^6 \). M1 life-times also are generally longer, but clustering around 100. E2 transitions, however, unlike other higher multipole transitions are often enhanced – generally by a factor of 2 to 500. These experimental measurements, though, are being compared with a very simple “single particle” picture of the nucleus; but, despite this, there is sufficiently good agreement in the description of excited states of nuclei exhibiting shell structure to make assignments to spin levels (i.e. s, p, d, f, …..etc.) within the nucleus. The poor agreement in the case of E1 transitions is attributed to the fact that most low energy E1 transitions are likely to involve at least one rather complex state(10). The dramatic enhancement of E2 transitions in the rare earth region and the region above Pb 208 is attributed to the large nuclear deformations from spherical symmetry with accompanying large quadrupole moments. Here the nucleus behaves as a spheroidal rotor exhibiting
Estimated mean lives of M1 and E2 gamma-transitions, based on single particle theory: 2 values of Z are shown (10 and 32), from Goldhaber and Sunyar (10).

Fig. 1
\[ g^* \gamma = \gamma \]

As shown, the experimental points are expected to be close to the theoretical values. In the \( G \) region, the total cross-section and the transition matrix element are in good agreement with the predicted values. The reduced transition probability is given by

\[ \frac{\hbar^2 \gamma}{2} \left( \frac{1}{E} \frac{dE}{d\gamma} \right) = \frac{\hbar^2 \gamma}{2} \left( \frac{1}{\gamma} \frac{d\gamma}{dE} \right) \]

The even-even nucleus is studied in detail by the experimental state and the quantum state of an excited state. On this model, the reduced transition probability for \( \Delta I = 1 \) is

\[ (I+1) \frac{\hbar^2 \gamma}{2} = (I)^* \text{ total} \]

The strong coupling model, with even states, agrees well with the experimental states described by the Born and Wigner model.
A semilog plot of half-life versus energy for nuclei with \( N \) and \( Z \) both even.

From Enge (14)
In the usual notation, as the table in Fig. 3 shows, the expression

\[
\Theta = \exp \left[ \frac{-2\pi\frac{z}{(1-\alpha^2)^{1/2}}}{\sqrt{1-z^2}} \right] \text{or} \quad \frac{x}{1-\alpha^2} = (\alpha) \Delta
\]

(114) numerically evaluated, the particle must be absorbed at the origin of the coordinate system of the scattered particle. The energy of the particle parameter is \( E \) and the parameter momentum of the scattered particle parameter is \( \mathcal{P} \). By

\[\frac{\text{denominator momentum}}{\text{numerator momentum}} + \text{denominator momentum} \times \alpha \text{} \frac{\text{numerator momentum}}{\text{denominator momentum}} = 0\]

we have the reduced mass \( m \.

\[
F \left[ \frac{\pi(z)}{\Delta(1-\alpha^2)^{1/2}} + (\alpha) \Delta \right] \left( \frac{\pi(z)}{\Delta(1-\alpha^2)^{1/2}} \right) = \mathcal{F}
\]

(115) and where \( \mathcal{F} \) is the numerically evaluated camps of the scattered particle. Coulomb potential barrier equals the energy of the scattered particle at the point at which the

\[
F \left[ \frac{\pi(z)}{\Delta(1-\alpha^2)^{1/2}} + (\alpha) \Delta \right] \left( \frac{\pi(z)}{\Delta(1-\alpha^2)^{1/2}} \right) = \mathcal{F}
\]

(116) where \( \mathcal{F} \) and \( \mathcal{F}_H \) are the co-ordinate distances from the origin to the

\[(\alpha) \left\{ \frac{\pi(z)}{\Delta(1-\alpha^2)^{1/2}} \right\} \text{exce} = \beta
\]

that points. The energy of the

\[\text{potential} \quad \mathcal{F} = \beta
\]

(117) where \( \beta \) reduced density constant for the density equation and
Barrier penetration factors for higher orbital angular momentum states

\[ Z = 90, \; E = 4.5 \text{ MeV}. \]

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<tr>
<td>( \frac{B_l}{B_0} )</td>
<td>1</td>
<td>0.84</td>
<td>0.60</td>
<td>0.36</td>
<td>0.18</td>
<td>0.078</td>
<td>0.023</td>
</tr>
</tbody>
</table>

*Fig. 3*
small for low values of $l$, but is much larger for higher values of $l^{(16)}$.

The value of the reduced decay constant may be estimated from a naive classical picture of a single alpha particle inside the nucleus as

$$\lambda_0 = \frac{v}{2R} \cdot P$$

(xiv)

where $\frac{v}{2R}$ is the number of times a single particle travelling at speed $v$ appears at the nuclear surface $R$, and $P$ is given by

$$P \approx 25 \left(\frac{4}{\Lambda}\right)^3$$

Alternatively, the Geiger-Nuttall plot will give the value.

Experimentally, the value of $\lambda_0$ may be determined from long-range alpha emission (e.g. $^{212}$Po excited states) by measuring the branching ratio $\frac{\lambda_0}{\lambda_x}$ and the experimental life-time of the state ($\lambda$). Then the reduced probability for alpha emission ($\lambda_x$) is determined from

$$\lambda = \lambda_x + \lambda_y$$

(xv)

and hence $\lambda_0$ may be found.

1.5 The Present Experiment

The decay scheme for $^{207}$Bi to $^{207}$Pb is shown in fig. 4. The nuclear excited states of $^{207}$Pb are generally described as shell model single neutron hole states, $p_{\frac{1}{2}}, p_{\frac{3}{2}}, f_{\frac{5}{2}}, i$ and $f_{\frac{7}{2}}$. The life-time of the $f_{\frac{5}{2}} \rightarrow p_{\frac{1}{2}}$ transition has been measured a number of times$^{(17)}$. But values seem to vary
from Körner et al. (17)

The Decay Scheme for Bismuth$^{207}$
between 90 ± 30 picoseconds and 134 ± 9 picoseconds, so a re-investigation would seem necessary to determine which group of results is the more correct. And since a ready-made Bi$^{207}$ source was at hand, it seemed a suitable nuclear life-time to measure by way of testing the operation of our solid state detector life-time measuring system and the computer controlled pulse height analyser - developed in the course of this project. The two gamma transitions (106$^\text{Kev.}$, $\frac{1}{2} \rightarrow \frac{1}{2}$ and 570 Kev., $\frac{3}{2} \rightarrow \frac{1}{2}$) are detected by their K - conversion electrons (approx. 10% conversion) which appear in the radiation spectrum as well resolved mono-energetic peaks.
CHAPTER 2

LIFE-TIME MEASUREMENTS OF NUCLEAR EXCITED STATES

2:1 Introduction

The methods of life-time measurements roughly may be divided into the following groups,

1. Recoil and Doppler Shift
2. Nuclear Resonance Fluorescence
3. Mössbauer Effect
4. Coulomb Excitation
5. Inelastic Electron Scattering
6. Mono-energetic Positron Emission
7. Observation of X-ray Satellites
8. Delayed Coincidence

The methods employed in groups 1-7 are adequately reviewed elsewhere(18),(19) so we shall restrict our survey to group 8 - the method of delayed coincidence. In this literature survey, the historical development of the electronic methods of life-time measurement will be reviewed, followed by an analysis of the practical problems encountered when setting up a system. The analysis is performed in the reverse direction to the progress of the electronic pulses through the system, because this is the normal direction when "trouble-shooting" and making improvements. The chapter ends with a discussion of the design of this experiment.
The Method of Delayed Coincidence

It was J. C. Jacobsen in 1934 (20) who first successfully performed a life-time measurement using a delayed coincidence technique. He used the delightful method of two Geiger counters connected, via amplifiers, to moving iron oscillographs, attached to which were small mirrors. These were set to rotate in a mutually perpendicular direction such that a beam of light could be deflected in two dimensions. The subsequent trace from such a beam was written onto photographic film moving at 1 cm/sec. In front of one of the counters was placed an aluminium foil to absorb the alpha particles - thus counting only beta rays in that channel; the other counter detected both alphas and betas. From the traces, Jacobsen deduced the life-time of RaC' (Po²¹⁴) to be $2 \times 10^{-7} \pm 50\%$.

But this is of historical importance only as the crossed oscillographs were quickly superseded by the electronic coincidence circuit. Jacobsen's method, however, has been adapted by replacing the oscillographs with the cathode ray oscilloscope (21).

The method of delayed coincidence utilizes a circuit which registers coincidences between input pulses from two radiation detectors in conjunction with a variable delay (T) inserted into either channel. If $x_1$ and $x_2$ be two radiations emitted in cascade and B the excited state whose mean-life ($\tau$) is to be determined (see fig. 5), then, in a typical coincidence experiment, one detector registers the radiation $x_1$ from the parent nuclide A, whilst the other registers radiation $x_2$ from the daughter nuclide B. By varying the
delay \((T)\), the change in coincidence rate may be investigated. If there is no natural \(\text{delay} \) between the emitted radiations i.e. a "prompt" coincidence, then a plot of coincidence rate vs. time delay \((T)\) gives the prompt resolution curve, as shown in fig. 6. For extremely small natural delays, the lifetime (according to Bay\(^{(22)}\) and Newton\(^{(23)}\)) may be determined from the shift of the centroid between a true prompt coincidence plot and the natural delay coincidence plot. If the state B has a mean life greater than the resolving time of the system, the experiment provides the time distribution \(\exp(-T/\tau)\) and is called the delayed resolution curve. A logarithmic plot of this gives a straight line slope from which the mean life may be determined directly.

2:3 The Coincidence Circuit

Both the introduction of the principle of coincidence methods and the first electronic coincidence circuit are attributed to Bothe and his collaborators\(^{(21)}\). His circuit consisted of a tetrode valve whose grids were connected to positive pulses from ionization chambers. Only when grids received a signal could the anode current flow (fig. 7a) generating an output pulse across the anode load resistance. It is essentially a "series" circuit as it operates as two switches in series, which must both be closed to allow the passage of current (fig. 7b). This type of circuit is limited in the number of coincidence inputs by the number of electrodes within the valve.

A more flexible circuit (fig. 7c) was introduced in 1930 by Rossi\(^{(24)}\),
Fig. 6

P(x) prompt and F(x) delayed coincidence resolution curves for a radiation of Hg^{199} having a half life of $2.35 \times 10^{-9}$ seconds. Diagram (a) logarithmic scale, diagram (b) linear scale. Chance coincidences have been subtracted.

from Bell (21)
Operation of the Meunier Coincidence Circuit
A negative input pulse at A will not pass to the input diodes; a positive pulse at B will pass the same signal to all 3 inputs of the difference amplifiers and will thus produce no output from either. Pulses A and B in coincidence will generate a large positive pulse at "3", a small positive pulse at "2" and no pulse at "1". This combination will produce an output from both difference amplifiers and coincidence unit.
which operates as a number of switches in parallel; all must be open to stop the flow of current through the anode load resistor. With this configuration, any number of coincidence elements may be incorporated as desired. Since valves and, to a greater extent, transistors and diodes are not perfect switches - in that they have a finite resistance when "off" - it is necessary to remove the small output pulses corresponding to "single" events by applying a suitable bias level at the output. It is of interest to note early workers did not have scalers to record coincidences - they listened through head-phones, recording the clicks with pencil and paper.

A third type of coincidence circuit is the balanced circuit (fig. 8); two such circuits are the Bay balanced coincidence selector and the Mannie differential coincidence selector. The former operates as follows: a negative input pulse at $B$ by itself will be stopped by the high impedance of diode $D_2$; a negative input pulse at $A$ will simultaneously pass through $C_2$ and $D_2$ via $r_b$ to earth and through $D_1$ and $C_1$ via $r$ to earth. The potential difference between $P_1$ and $P_2$ will thus remain zero. When two pulses arrive in coincidence, pulse $B$ blocks the passage of pulse $A$ at diode $D_2$ creating a potential difference across $R$; thus an output signal occurs. All balanced circuits have two features in common. Firstly, when carefully balanced they can accept input signals with a wide range of amplitudes and can successfully reject single events without resorting to pulse amplitude standardization. Secondly they are sensitive to small signals because the signal need only be large enough to establish slight non-linearity in a circuit element and do not have to switch currents completely on or off.

The resolving time of a coincidence is defined as the minimum time...
Fig. 8

(A) Biyar balanced coincidence circuit.

(B) Mauier differential coincidence circuit.

(C) \( d_3 \) and \( d_4 \) are tunnel diodes from Bell (26)
interval between the arrival of two input pulses which is required for them to be resolved as non-coincident. The resolving time of the original circuits was of the order of $10^{-3}$ sec., but nowadays coincidence modules are commercially available with resolving times of a few nanoseconds ($10^{-9}$ sec.). Fransini\(^{(25)}\) claims a resolving time of 1.5 nsec. for a fast coincidence circuit using tunnel diodes as switching elements (see fig. 8c). But ultimately the resolving time of a system is limited by the uncertainty with which the detector pulses characterize the arrival of the primary radiation. This uncertainty is twofold; firstly there is the statistical jitter in the leading edge of the pulse which is a direct function of the electronic noise and can only be improved by reducing the noise or risetime of the pulse, and secondly there is an amplitude-dependant uncertainty (walk) which is a function of the discriminator bias level with respect to the maximum pulse height. This may be reduced by the use of a "fast - slow" coincidence system (fig. 9). Each input signal, in addition to going to the fast coincidence circuit, goes to a single channel pulse height analyser which is set to accept only those pulses which fall in a narrow amplitude range. The output pulses from the two pulse height analysers are combined with the output of the fast coincidence unit in a slow triple coincidence unit which produces an output pulse only when the time and amplitude restrictions are satisfied simultaneously. This format is very common with present day delayed coincidence experiments where pulse height selection and high resolution are required to correlate two nuclear events out of a whole spectrum of radiations.
Fast - Slow Coincidence System
Early lifetime measurements were performed simply with a coincidence unit and a variable delay inserted into one channel. The "integral" method, developed by Feather and Dunworth (26), stretched the parent pulse to a known duration $T$ with a pulse shaping circuit whilst keeping the daughter pulse short. The coincidence counting rate then is given by $N = \exp(-T/\tau)$, where $\tau$ is the mean life of the excited state. For any particular value of $T$ the measured coincidence is the integral of the desired time distribution $N_0 \exp(-T/\tau)$ from $T=0$ to $T=T$. The original form of the time distribution is obtained by differentiating the measured curve or, in practice, by subtracting successive readings. This method was used to determine the half-lives of the alpha emitting series Po$^{216}$ ($\frac{1}{2} = 0.16$ sec.) to Po$^{212}$ ($\frac{1}{2} = 3 \times 10^{-7}$ sec.).

The differential delayed coincidence method was introduced by Jacobsen and Sigurgeirsson (27) in 1943 in an experiment on Po$^{214}$. The pulses from the counters were kept to a short length and the variable time delay in the parent pulse channel was inserted by a monostable circuit of variable length. As mentioned before, the coincidence count-rate in this configuration is given by $\exp(-T/\tau)$. The method was taken up by De Benedetti and Mc Gowan (28) in 1946, who applied it to some sixty nuclides with life-times in the microsecond region. In 1947 Rowlands (29) introduced the idea of using the parent pulses to trigger the sweep of an oscilloscope time base and the daughter pulse to be fed to the Y deflection plates, the resultant trace being photographed. However, he comments that this method is not as
accurate as the coincidence circuit method but that it might have potential for shorter life-time measurements.

The slow risetime (10^{-7} \text{sec.}) of pulses from ionization chambers set the limit of about 10^{-7} \text{sec.} on early life-time measurements. The advent of the photomultiplier tube in conjunction with fast scintillators reduced this limit by two orders of magnitude to about 10^{-9} \text{sec.} and Bay\textsuperscript{30}, replacing phosphors with Cerenkov counters (which have virtually zero risetime) claims a resolving time of less. At this time Bay and Newton introduced the idea of the "centroid shift" measurement of life-times which is capable of extracting measurements yet two orders of magnitude lower i.e. approx. 10^{-11} \text{sec.} But at present this seems to be the limit of the delayed coincidence technique for life-time measurements; perhaps the new photodiodes (Instrument Technology Ltd.) recently on the market, which claim risetimes of 100 picoseconds, will reduce the limit to 10^{-12} - 10^{-13} \text{sec.} Perhaps more sophisticated system stabilisation and data extraction, as is attempted here, will also reduce the limit to this region.

An improvement on the centroid shift method is that of "self comparison" first published by R. E. Bell et al.\textsuperscript{31} in which the delayed coincidence curve is compared not to a prompt curve but to its own inverse i.e. to the delayed curve obtained by reversing the roles of the counters. The centroid shift, in this case, is 2 \gamma. Using this method, the authors found the half-life of the 412 kev gamma transition of Hg\textsuperscript{198} following beta decay of Au\textsuperscript{198} to be 1.0 \pm 1.7 \times 10^{-11} \text{sec.} This method also incorporated the fast - slow coincidence technique, but as applied to beta-spectrometers - the energy
selection being effected magnetically within the spectrometer.

A novel, but rather inefficient, single channel method of time analysis is that of the "microwave" or "high frequency deflection" method described in a review article by Bonitz\(^{32}\). This is only applicable to the detection of charged particles as the principle of operation involves their deflection away from the detectors (beta-spectrometers) except at the one point in time when the electric gate, through which they must pass, is at zero potential. This gate is controlled by applying to it a high frequency voltage from a microwave generator. The phase of the gating signal in the parent pulse is successively retarded until coincidence occurs with the daughter particle in a slow coincidence unit. The electronic resolving time is then a function of the microwave frequency; the higher the frequency, the shorter the resolving time. Resolutions of the order of $10^{-11}$ sec. are claimed for this method, but data collection is extremely slow.

2.5 Multichannel Time Analysers

It is because of this very inefficient collection of data in single channel analysers that multichannel systems have been so successful. Whereas the time for data collection in the former method is proportional to the number of incremental steps in the delay measurement, the latter performs the entire operation in a single measurement, reducing amongst other things the effects of long term drift in the apparatus.

One such multichannel system was the "chronotron" used by Bell and Hinks\(^{33}\) in 1952. It was a multiscaling device using 10 scalers which
were successively stepped through at 1 microsecond intervals by a digital clock. The parent pulse initiated the sweep and the daughter pulse recorded its existence in whichever scaler was open at that particular time; so a delayed coincidence spectrum was accumulated, analysing each pair of pulses as they occurred. This method is only satisfactory for comparatively long life-time measurements (> 0.1 microsec.). For shorter times Lefevre and Russel(36) have developed a vernier chronotron. In this, two time marking pulses are triggered by the parent and daughter inputs and are circulated around delay loops of slightly differing time delays ($T_1$ and $T_2$). The number of circulations is counted by a scaler until a coincidence between the pulses halts the counting. Then the original time delay of the daughter after the parent is

$$\tau = n \Delta T$$

where $\tau = \text{mean life}$

$$n = \text{number of counted pulses}$$

$$\Delta T = T_1 - T_2$$

The number $n$ is then presented to a multichannel analyser: this presents a direct transition of time-to-digital information, thereby removing the intermediary steps of time-to-analogue then analogue-to-digital conversion as in the Time-to-amplitude Converter / Kicksorter arrangement.

But the Time-to-amplitude Converter (T.A.C.) is the most popular method because of the simplicity of coupling it to a standard multichannel pulse height analyser. With this combination the time interval between the two input pulses is converted to an output pulse, whose amplitude is proportional to the time interval, and analysed on a Kicksorter.
Multichannel pulse height analysers have now reached a high level of sophistication. When first constructed, they were no more than a bank of separate scalers coupled to an analogue-to-digital converter, where the count on each scaler had to be read and the spectrum manually plotted. Nowadays, the spectrum is displayed on an oscilloscope screen, plotted on a graph-plotter and the contents of the core typed or punched out for further analysis as desired. Other functions such as spectrum-stripping and integration may be performed by built-in fixed programs. And the number of subdivisions of analysis or channels is increasing dramatically. But so is the cost of these units and it is becoming increasingly popular to interface a specially designed A. B. C. for nucleonic work to the memory core of a small computer. This provides a multichannel pulse height analyser with a large number of channels combined with a flexibility in data processing unattainable in the standard kicksorter. This is probably where the improvements in the resolution of life-time measurements will next come.

2.6 Time-to-Amplitude Converters

The heart of the majority of life-time measurement experiments is the time-to-amplitude converter (T. A. C.). The accuracy of a time measurement depends upon that of the T. A. C. - its linearity, temperature dependence, overload characteristics and long-term stability. Very many forms of converter have been devised and their merits discussed (35), (36), (37), (38), (39) but all perform the same basic operation - that the converter gives an output pulse whose height is proportional to the time delay between two input pulses;
this is usually accomplished by charging a capacitor with a constant current for the period of time concerned. We may classify T. A. C. s into two main groups — the "start-stop" method and the "pulse overlap" method.

The "start-stop" converter is the form most popular amongst commercial units because of the ease of altering the time range. A simplified form is shown in fig. 10. In the quiescent state $T_1$ and $T_2$ are conducting. The parent or start pulse switches $T_2$ off, diverting the constant current supply to the capacitor $C$. The daughter or stop pulse switches $T_1$ off and the charging of $C$ ceases. Thus, ideally, the voltage at the gate of the F. E. T. transistor is

$$V = \frac{I \cdot \tau}{C}$$

where

$I = \text{charging current}$

$\tau = \text{period of time between start and stop pulses}$

$C = \text{capacitance of the condenser}$

To reset the circuit, both $T_1$ and $T_2$ are made conducting again, shorting the charge on the capacitor to earth. For very short time ranges, a high charging current is fed to a small capacitance; subsequent increase in time range is effected by switching in larger capacitors and/or reducing the charging current. The high input impedance of the F.E.T. transistor reduces to a minimum the leakage current from the capacitor whilst being charged. For short time ranges ($0 - 1 \mu\text{sec}$.) this is of little consequence but above this ($0 - 100 \mu\text{sec}$.) a large leakage current would lead to a nonlinear time
conversion at the top end of the range.

It is characteristic of every "start-stop" converter that a single input pulse, in most cases the start pulse, will produce an output signal. For this reason, some form of supervisory circuit must be introduced which suppresses this kind of output pulse and also discharges the capacitor before the advent of the next event. Variations of the "start-stop" theme are the "start-multiple stop" converter proposed by Kowalski (40) and a circuit by Culligan and Lipman (41) later improved by Orphir (42) which uses an inductance to store the time-proportional charge (17) before integration on a capacitor. Other converters are described in references (43, 44, 45, 46).

The "pulse overlap" converter receives two shaped square wave pulses from the two inputs and adds them, biasing the output to accept only coincident parts of the pulses. The resultant charge is integrated on a capacitor. A series coincidence circuit of the Bothe type was used by Bell and Green (47) with a 6SN6 gated beam valve; but nowadays, with transistors, the parallel or Rossi type of circuit is favoured. A typical pulse overlap circuit is that of Simms (48) shown in fig. 10b. Normally the transistors $T_1$ and $T_2$ are conducting, fed from a constant current source. If either of the transistors is switched off then the current flows through the other. Only when both are cut off is the current diverted to transistor $T_3$, where the charge is integrated on its own stray capacitance. By adjusting the biasing of $T_3$, excellent singles rejection is obtained.

But the pulse overlap converter on its own gives a double valued output i.e. it is possible to have two output pulses of the same height for two different configurations in time of the input pulses. Thus a further
supervisory circuit is required, usually a simple coincidence circuit controlling a gate through which the output pulse of the T. A. C. must pass. Careful juggling of delay times in the various channels will veto one half of the pulse overlapping process (to be described more fully in chapter 3).

This type of converter is simpler in construction than the "start-stop" version and its linearity for short time ranges appears to be better, however it is more cumbersome to alter the time range as it involves replacing delay cables and altering the input pulse widths. A general theory of time matching for these converters is given by Tao et al.

A novel variation on the overlap theme is a converter circuit by Weisberg. It gives a single valued conversion output by virtue of the hysteresis of a tunnel diode (see fig. 11). The start channel generates a short spike pulse whilst the stop channel generates a long square pulse. These signals are added producing the current waveform $I_1 + I_2$ as shown. In the idling condition the tunnel diode is in its "off" state (point 1 in fig. 11b); it is switched on to point 3 by the short spike, then relaxing to point 4 where it remains till the end of the wide pulse, when it returns to point 1. In this way the integrated voltage across the tunnel diode gives the single valued time conversion. The range of the converter can be altered simply by varying the length of the stop pulse. In Weisberg's circuit there is some non-linearity in the time conversion, but only at the extremities of the range - the linearity in the central region being tolerable.

Unusual forms of T. A. C.'s have been introduced by Siekman and by Cottini and Gatti. The former uses an oscilloscope time-base which is triggered by a start pulse; the stop pulse is superimposed upon the linear
Fig. 11

(A) START > I₁ > I₂ > INT > OUTPUT

(B) Pulse from "start" counter
Current output from "narrow" univibrator
Pulse from "stop" counter
Current output from "wide" univibrator
Total current input to tunnel diode
Voltage across tunnel diode

from Weissberg (5)
ramp of the X-deflection voltage, the resultant pulse height being analysed in the conventional manner (see fig. 12a). No added supervisory circuit is required as the X-deflection voltage resets itself at the end of its sweep. This method has potential for a "start-multiple stop" unit. The second system utilises the vernier principle for time expansion in conjunction with a standard "start-stop" converter (see fig. 12b). Two oscillators of slightly differing periods \( T_1 \) and \( T_2 \) are triggered by the start and stop(1) pulses respectively. If the time interval between the input pulses is \( T \) then the two oscillators will arrive in phase with one another after a time given by,

\[
T' = \frac{T T_2}{\Delta T}
\]

where \( \Delta T = T_1 - T_2 \) with \( T_1 \neq T_2 \) and \( T_1 > T_2 \).

A discriminator triggers when the resultant signal exceeds the threshold level, generating the stop(2) pulse for the time converter (which itself had been started by the start pulse). This converter, however, has not been much used — presumably because of its large dead-time.

In an early attempt to attack the everpresent problem of instrumental drift especially in time converters, the differential time-to-amplitude converter was conceived. The system used by Berlovich (54) was based on the use of two converters situated on a variable delay line operating in reverse to one another (fig. 12c). The time profiles obtained are displayed in two halves of a Kicksorter (by virtue of their position on the variable delay line); the delayed coincidence curves are then displaced with respect to the prompt curves in opposite directions, whilst the instrumental effects
Fig. 12

(A) channel 1

(B) from Cottini + Gatti (53)

(C) from Berlovich (54)

A – calibration curve; B – variable delay line, (a) and (b) – positions of converter; C – relative positions of prompt and delayed curves in cases (a) and (b).
displace them in the same direction. Sen and Patro (55) modify this idea by summing the two T. A. C. output signals in a difference amplifier before analysis by a kicksorter.

There are many varieties of time-to-amplitude converter, but none of them is perfectly linear or perfectly stable. For short time measurements the "pulse overlap" type seems to give the best differential linearity; for long time measurements the "start-stop" unit is easier to construct and has better integral linearity. For any particular experiment, one must select the converter which is best suited to that purpose.

2:7 Auxiliary Circuits

(a) Pulse Height Compensators

When using leading edge discrimination for timing signals from anything other than monoenergetic radiation, some form of pulse height compensation is required, otherwise the amplitude-dependant time walk at the discriminators will broaden the time spectrum unnecessarily. The pulse height compensator, as its name implies, adds to the output signal of the time converter a correction signal which endeavours to compensate for this amplitude-dependant time shift. A number of successful circuits have been devised, all utilizing the signals from the slow channels to provide the compensation. The circuit of Rodda et al. (56) takes the amplitude of the slow channel pulses, via attenuators, and simply adds one channel to and subtracts the other channel from the T. A. C. output pulse (see fig. 13a). Thus if a pulse from the detector in channel 2 were larger than average, it would stop the time
Fig. 13

from Thieberger (52)

from Radka et al. (56)
converter prematurely giving too small an output pulse; but the pulse in the slow channel would be larger than average and thus compensate for the deficiency of the time converter pulse.

This compensator gives a first order correction only, whereas the method used by Thieberger(57) attempts a more complete correction. In it he uses a second time-to-amplitude converter, operating from the slow channels, which is started, in effect, by the stop pulse of the fast converter and vice-versa (see fig. 13b). Therefore the second slow converter has a reverse centroid shift vs. pulse height characteristic compared to the fast one and is utilized to cancel any time-dependant walk in the first. An advantage of this over the previous system is that single correcting pulses cannot arrive at the Kielesorter to produce unwanted broadening or shift of the time spectrum. This type of compensator is not easy to adjust, but, when correctly set, the combination is supposed to be superior to zero cross-over timing techniques (which will be discussed later in the chapter).

A modification of the first compensator comes from Jaklevich et al.(58) who, instead of using a simple linear compensation signal from the slow channels, have a logarithmic signal. This was in an effort to introduce a better correction for a system with large volume Ge(Li) detectors to detect a wide energy range of gamma radiation; it gave an improvement in the time resolution of a factor of 2 in most cases. Thieberger and Harms-Ringdahl(59) tackled the same problem using neither a linear nor a logarithmic correction amplifier but one whose output is an adjustable nonlinear function of the slow channel signals.
(b) Anti-pile up systems

A further cause of time spectrum broadening is pulse pile up; if the countrate is high then the pulses in the slow channels will tend to ride up on one another, distorting their true shape. This is particularly troublesome for the time measurement by the slope method of lifetimes slightly larger than the prompt response slope. The anti-pile up circuit eliminates this by vetoing information connected with the second pile up pulse. Two forms of circuit are shown in fig. 14a. The first, after Schwarzschild(36) eliminates the unwanted pulses from the slow channels by producing a 4 microsecond block in the system after each pulse.

However, this does not eliminate events in which a fast coincidence is followed by another pulse. These pulses also contribute to the background. The circuit in fig. 14a ii, after Welsberg, rejects such pulses.

A pile up detector has two time characteristics - the pile up resolution time, which should be as short as possible, and the blocking time, which should be equal to the recovery time of the slow channels. There are several ways of obtaining anti-pile up control and these are reviewed by Ogata et al. (49)

If one uses zero crossover timing (i.e. triggering a discriminator at the point when a double differentiated pulse crosses through zero potential, thereby generating an amplitude-independent time mark) at high countrates, then an anti-pile up circuit is essential as the time of crossing at the zero level can be shifted by the presence of another pulse too close-by. But a system with good pulse height compensation and leading edge discrimination needs no anti-pile up circuit since a weaker source may be used.
(A)

(i) P.M. 1
  DYNODE → AMP
  DISC
  OR 4 μsec BLOCK
  0.15 μsec DELAY
  ANTI
  SLOW TRIPLE COIN
  GATE OUTPUT
  
  P.M. 2
  DYNODE → AMP
  DISC
  SINGLE CHANNEL PHA

(ii) P.M. 1
  PILE UP DETECTOR
  OR CIRCUIT

P.M. 2
  PILE UP DETECTOR
  COIN
  DELAY
  ANTICOIN CIRCUIT
  GATE OUTPUT

from Schwarcz (36)

(B)

from Simmes et al. (63)

Fig. 14
(e) Stabilization

Instability due to thermal variation is a problem present in all electronic systems, although, by good design, most of it can be removed. But unless some form of stabilization is incorporated, a time spectrum may suffer from broadening and, even worse, a shift in the centroid. For best stability thermostatic control is the real answer but, failing that, there are other methods. Rota et al. (60) and Weisberg and Berko (61) took data alternatively from a prompt and a delayed source which was mounted on a wheel and changed at set times, the information being stored in alternate halves of the Kicksorter. A feedback system introduced by Bell (25) and used by Weaver (62) holds a reference event in the time spectrum (usually supplied by a pulser) constant with respect to the Kicksorter, thereby stabilizing the whole system excepting the photomultiplier tubes. This technique has long been used in straightforward energy spectroscopy.

In certain special cases, where the nuclear decay scheme involves a triple cascade, one of which may be taken as prompt, a third detection line is incorporated and the prompt and delayed spectra simultaneously recorded in different halves of a Kicksorter. The method was first published by Simms et al. (63) but is reputed to have been in use before this. Fig. 14b shows the relevant parts of the decay scheme of $^{42}$K. The life-time of the $0^+$ state was to be measured; the $2^+$ state life-time was assumed sufficiently short as to be considered prompt. One channel detected the beta continuum feeding the $0^+$ state and the $2^+$ state, the second channel detected the $1.5$ Mev. gamma rays and the third channel detected the $0.31$ Mev. gamma rays. When a triple coincidence ($p_1, \gamma_1, \gamma_2$) occurred, the time signal, which
contained the delay of the 0+ state, was routed to one part of the kicksorter. When only a double coincidence (\(B_2\) and \(Y_2\)) occurred, the time signal containing the prompt delay of the 2+ state was routed to the other part. The centroids of the distributions in the two analyser sections were compared to derive the life-time. This method was extended by Schwarzschild\(^{36}\) to the use of two separate active nuclides within the one source, one with a prompt emission e.g. Na\(^{22}\), the other with the delayed emission. However, unless the radiations emitted by both nuclides are similar in energy and pulse shape, problems of accurate timing will still arise.

2.6 Derivation of Timing Signals

To obtain precise time information about nuclear events, logic pulses marking their occurrence must be generated. This is accomplished by some form of fast timing discriminator, the different types of which may be listed as follows,

1. Leading Edge
2. Zero Cross-over
3. Constant Fraction
4. Amplitude and Rise-time Compensated

(a) Leading edge timing

This was the original method of generating a timing pulse. With valves, the signal from the detector amplifier was fed to a high gain sharp cut-off tube which gave an output pulse approximately square shaped. With transistor
circuitry, the input pulse, upon exceeding a set threshold level, triggers a monostable device giving out a square wave pulse. To reduce the uncertainty with which the output pulse characterizes the timed event, it should have as short a rise-time as possible. Very fast pulses have been obtained from avalanche transistors but they suffer from a long recovery time. A popular means of generating a fast timing pulse is with tunnel diodes; these have a very fast switching time and can tolerate large signals in both directions. But in the sub-nanosecond region tunnel diodes suffer from excessive time walk as shown in fig. 1b. Being charge sensitive devices, they require a minimum charge \( q_{\text{min}} \) to turn them on; and the time delay before switching occurs is a function of the charge overdrive. A current pulse \( I_1 \) presenting a charge \( q_1 \) generates an output pulse \( q_{0v} \) delayed by \( t_{d1} \); whereas a smaller pulse \( I_2 \) presenting a charge \( q_2 \) generates an identical output pulse \( q_{0v} \) but delayed by \( t_{d2} \); so the total walk \( W \) is the sum of the amplitude dependant walk \( W_{st} \) and the walk \( (t_{d2} - t_{d1}) \) inherent in the tunnel diode operation. This problem is discussed by Grunberg and Tepper (64) who overcome it by isolating the input signal from the tunnel diode with a "snap-off" diode; this is a storage diode which delivers an approximately constant charge, independent of current, to the tunnel diode thus eliminating this excessive walk.

(b) Zero cross-over timing

As mentioned earlier, there are two main sources of timing error - jitter, the timing uncertainty due to the electronic noise of the system, and walk, the timing variation due to amplitude- and rise-time variations.
Rough picture of TD behavior.

a) Input current pulses; b) Output amplitude.

- $t_{d1}$ - delay for a large charge overdrive $Q_{0y}$
- $t_{d2}$ - delay for a small charge overdrive $Q_{0y}$
- $W_r$ - real "walk" of output pulse

From Gurnberg and Tupper (64)
at the discriminator threshold (see fig. 15a). Jitter cannot be removed, only reduced by filtering and cooling. However, walk can largely be overcome by the technique of zero crossover timing; the input signal is differentiated to form a bipolar pulse and the discriminator triggered when it crosses through the zero point which, to a first approximation, is independent of pulse amplitude. But setting a discriminator level to trigger at the zero point requires some form of supervisory circuit to inhibit continuous triggering on noise. Either the discriminator level can be slightly off-set above or below the noise level, in which case there will be a small amount of walk associated with small amplitude pulses or a second discriminator, triggering on the leading edge of the input pulse, can be employed. The output from this second discriminator can be used to set the first from some non-zero level to the zero point in readiness for the return swing of the pulse through zero. This is the method commonly used in commercial zero crossover discriminators. Weiberg and Lefevre(65) remove the zero crossing point from the noise level by adding a pedestal signal (see fig. 16a). Two signals corresponding to an event are taken from the anode and the last dynode of a photomultiplier tube. The anode signal is double differentiated; the dynode signal generates the pedestal by triggering a fast discriminator. These two signals are summed and fed to a tunnel diode discriminator set to trigger at the same level as the height as the pedestal.

For slow shaped input signals zero crossing timing gives excellent results by way of the elimination of time walk but, for fast pulses of varying profiles and amplitudes such as is obtained from radiation detectors, leading edge timing with pulse height compensation is better.
Pulse Shape and Time Consideration for the Constant Fraction Timing Technique

from Ortec Catalogue (27)
(c) Constant fraction timing

This is a hybrid form of leading edge and zero cross-over timing. Various results from timing experiments indicate that a minimum time resolution is obtained for a fractional level of triggering of about 10% of the peak pulse height; it has also been suggested that optimum time resolution should be obtainable with a circuit which triggers at a constant fraction of the peak pulse height. Gedcke and McDonald (66) have developed a constant fraction timing discriminator using an attenuation-subtract technique. As shown in fig. 16a, the input signal (a) is divided in two, one part being attenuated (b) and the other part delayed and inverted (c); these are summed to give (d). Zero crossing discrimination is performed, giving a timing mark which is free from walk yet effectively triggering on a constant fraction of the maximum pulse height. Gedcke's circuit claims a walk of less than 120 picoseconds for a dynamic range of 100:1. This circuit is very satisfactory only if the shape of the input pulse is identical for all amplitudes.

(d) Amplitude and rise-time compensated timing

This differs from constant fraction timing only in that the delay inserted into the inverted signal is much shorter; in so doing, variations in rise-time, assuming them to be merely slope changes, are compensated. Fig. 17a, taken from the Ortec Catalogue (67), describes the operation of their constant fraction timing discriminator. \( V_1 \) and \( V_2 \) in diagram(a) represent two input signals of different rise-times but same maximum value. With a delay of only \( t_d \) in the inverted section of the signal, both \( V_1 \) and \( V_2 \) pass through zero at the same point in time. Figs. (c)+(d) merely show the limits set on the
**A**

**Input Signal**

- **Voltage**
  - Time \( t = 0 \)
  - \( t_{\text{max}} \)
  - \( t_{\text{min}} \)
- **Input Limit**
  - Signal must not exceed \( V \) before this time
- **Shaped Signal for Timing**
  - **Zero Crossing Time**
- **Minimum Input Limit**
  - Signal must not exceed 1 V before this time
- **Maximum Input Limit**
  - Signal must exceed lower level disc. by this time

From Ortec Catalogue (67)

---

**B**

Charge pulse shapes corresponding to single interactions at various positions in the depletion region of a planar Ge(Li) detector.

- **a**—Charge pulse due to an interaction in the middle of the depletion region (solid) and near the middle (dotted);
- **b**—a delayed;
- **c**—a inverted and attenuated;
- **d**—sum of b and c.

From Chase (69)
input signals for the particular Ortec unit. For signals from Ge(Li) detectors which have systematic variations in pulse rise-time profile — not just the slope — Chase(68) performs the same operation (see fig. 17b) but this time the triggering level is not at the zero point but at some point on the leading edge of the pulse through which all rise-time profiles pass at the same point in time. This level is theoretically computed for a particular detector assuming very simple variations in pulse profile.

Methods of Time Pick-off

Photomultipliers (and scintillators) present very little problem in supplying a suitably fast output pulse for timing purposes. To extract a faster pulse the output may be taken from an earlier dynode in the chain, but the signal is still of sufficient size to trigger a timing discriminator with little further amplification. Output pulses in the nanosecond region are usual. Solid state detectors, on the other hand, although having superior energy resolution, generate such a small pulse that great care must be taken to use it effectively; and the charge collection time within the detector is long by P. M. tube/scintillator assembly standards.

With solid state detectors it is usual to extract the pulse along two electronic chains. A slow charge sensitive preamplifier registers energy whilst a fast voltage sensitive preamplifier registers time; the former gives good energy resolution but poor timing (due to the capacitative input impedance of the preamplifier) whereas the latter gives good time
Fig. 18 from Alberigi Quaranta et al. (69)
(a) Introduction

To a first approximation,

\[ dt = T_r \cdot \frac{dE}{E} \]

where \( dt \) = uncertainty in time measurement

\( T_r \) = rise-time of pulse

\( dE \) = energy resolution of that pulse

\( E \)

For best timing conditions, then, one would like very low noise combined with fast rise-time in a detector. With the exception of thin detectors (used for heavy particle detection), the photomultiplier/scintillator gives better time resolution than the solid state detector - because of its much faster rise-time; the good energy resolution does not compensate for poor rise-time in the latter. But there may be other considerations necessary in the choice of system. The solid state detector, just because of its superior energy resolution, is more capable of investigating an event in a background of many other events which the photomultiplier/scintillator could not resolve. It also lends itself more readily to feedback stabilization since the amplitude of the pulse from the detector is not drastically affected by drift in the voltage supply - as is the case with P. M. tubes. When performing time measurements by the method of centroid shift, it is preferable
to have a good method of stabilisation than the best time resolution.

(b) The Photomultiplier/Scintillator Combination

The various photomultiplier systems have been reviewed by Ogata et al.\(^{(49)}\); but the main sources of limitation of resolution may be listed as follows,

1. Jitter in energy transfer to scintillator
2. Fluctuations in fluorescence and phosphorescence
3. Effective light output to photocathode of P. M. tube
4. Time spread between electrons travelling through successive stages
5. Delayed photo-electric effect
6. Inter-stage interference

Fast phosphors such as NE 102 and Natom 136 give good timing response with fast low noise P. M. tubes such as the 56 AVP. The RCA 7004,5 and the Philips XP 1020 tubes were specifically developed for timing operations and are capable of 100 ps. resolution on \(^{60}\)Co gamma rays. The idea has been suggested \(^{(72)}\) of using the glass window of the new fast photodiodes, recently on the market, as Čerenkov counters (with virtually zero rise-time) to produce a very narrow time resolution curve.

(c) The Solid State Detector

The use of solid state detectors in timing systems has been discussed in some detail by Alberigi Quaranta\(^{(69)}\). The criterion for solid state detectors still applies — low noise and short rise-time. However there are a number of parameters which must be taken into consideration before selecting a detector for a timing experiment. Firstly one must decide which type of
detector is most suitable for the radiation in question and secondly one must investigate the inter-relationship of rise-time and energy resolution with respect to the different parameters.

The two main forms of solid state semiconductor detector are,  
1. the p-n junction or surface barrier, and  
2. the p-i-n junction or lithium drift.

Both can be made from germanium or silicon. Surface barrier silicon detectors are preferred for measuring particle energy because they can operate at room temperature (and therefore no cryostatic housing for the radiation to penetrate) and have a very thin window or dead-space in front of the sensitive layer. Silicon lithium drift detectors are suitable for low energy photon and electron detection and can be operated conveniently at room temperature. Germanium is more suitable for measuring gamma radiation because of its higher atomic number (and thus stopping power); but the high thermal leakage current at room temperature requires it to be cooled. Lithium drifted germanium must be operated around 77°C to reduce the mobility of the lithium ions and prevent them from drifting out.

The rise-time of a pulse from a solid state detector depends upon the speed of collection of the charge liberated by the radiation in the sensitive volume of the detector; the charge is produced in the form of electron-hole pairs which are swept to the respective electrodes by a suitable bias voltage.

The stopping time of the radiation within the detector may be considered negligible ($\sim 10^{-12}$ sec.) in comparison to the other contributory factors to the rise-time. The collection time for the charge carriers is determined by,  
1. the position at which they are formed,  
2. the magnitude and distribution of the electric field,
3. the respective carrier mobilities,
4. the "plasma delay" time (for heavy ionizing particles).

The exact theoretical process of charge collection is not fully understood and there is some controversy as to whether Hamo's theorem\(^{(73,74)}\) or the Energy Balance equation\(^{(75,76,77)}\) should be employed in its calculation. In practical terms the difference in result is negligible and, when choosing a detector for an experiment, one may use the rough guide that the rise-time is approximately\(^{(78)}\)

\[
t = \frac{d^2}{\mu V}
\]

where
- \(t\) = the collection time,
- \(d\) = the thickness of the sensitive layer,
- \(\mu\) = the mobility of the carrier,
- \(V\) = the applied voltage

To maintain a short rise-time, then, the depletion layer should be no thicker than is required to stop the incoming radiation (and possibly slightly less)\(^{(79)}\). An increase in the bias voltage, in an attempt to reduce the charge collection time, will lead eventually to a velocity saturation level, above which the mobility of the charges will not increase. This saturation voltage varies to some extent with material and temperature, but is in the region \(10^3 - 10^5\) volts/cm. As the electrons and holes have differing mobilities the leading edge of the output pulse will be made up of two components - a fast rising edge due to the electrons and a slower rising edge due to the holes. For
timing systems, triggering on the faster edge - in practice 10 - 20% of the maximum pulse height - gives best time resolution. If very high energy resolution is required, then the detector's electrical capacity must be taken into consideration and reduced to a minimum concomitant with other requirements. For heavy ionising particles the charge collection is retarded by the highly conductive plasma of electrons and holes generated along the tracks. The process is not fully understood (69), but Moszynski and Bengtson (80) have shown the time delay and excess plasma jitter to be associated with the start of the charge collection and both to be inversely proportional to the electric field within the detector (despite the bias voltage being above that for velocity saturation). The time delay is experimentally defined by the operation,
\[
t_{\text{meas.}} = \sqrt{t_r^2 + t_d^2 + t_p^2}
\]
where
- \( t_{\text{meas.}} \) = rise-time of measured voltage pulse,
- \( t_r \) = charge collection time only,
- \( t_d \) = rise-time of detector equivalent and voltage amplifier circuit,
- \( t_p \) = plasma time

and it can be of the order of nanoseconds (69), with the excess jitter of the order of \( 10^{-11} \) sec. (80). This would seem to set the ultimate limit on the resolution of the time distribution for semiconductor detectors for particle measurement.

A more detailed appraisal of detectors, their construction and operation, may be had from various sources (67, 68, 78, 31), but for timing experiments there
is considerable room for latitude depending upon the energy resolution and rise-time requirements of the particular system.

2:11 The Present Life-time Measuring System

The brief of this project was twofold,

1. to set up a pulse height analysis system coupled on-line to a computer operating in "shared real-time access" mode, in cooperation with the other on-line user/research teams.

2. to set up a system for measuring sub-nanosecond nuclear life-times using solid state semiconductor detectors and the computerized pulse height analysis system.

A discussion of the former is to be found in chapter 4. Here, we will concern ourselves with the reasoning behind the design of our life-time measuring system.

The system as a whole is required to provide good energy resolution and good timing characteristics: this is best attained by performing energy selection with one chain of slow electronics and extracting time information along a separate (fast) chain of electronics, culminating in suitable "fast-slow" coincidence circuitry to permit data collection of only the relevant information.

(a) The detector assembly

As the system is to be used for alpha- and low energy beta- detection, any window through which the radiation must pass must be as thin as possible.
Detector assemblies maintained at liquid nitrogen temperatures, which could not be opened to the atmosphere, would present considerable difficulty in introducing and refurbishing a fast decaying source; and, at the time of setting up the experiment, there were no detectors available on the market which could be cycled between liquid nitrogen- and room temperatures. So it was decided to use silicon detectors at room temperature in a chamber which was easily accessible. However, a refrigeration unit was on hand so it was incorporated into the system to stabilize and maintain the temperature of the detectors just below room temperature (thus reducing electronic noise). To permit refrigerated operation and reduce energy degradation in alpha particles, the whole was operated in vacuo in a specially constructed vacuum chamber.

(b) Method of time pick-off

Bell(82) in a comparison of leading edge and cross-over pick-off timing techniques showed that leading edge timing will always be superior statistically. A suitable fast leading edge time pick-off unit appeared to be the Ortec 260 unit in conjunction with the 403A control unit. So leading edge timing, backed by pulse height compensation, was selected. No suitable fast voltage preamplifier was available to augment weak timing signals for the Ortec 260 unit, so it was decided to construct such a unit in the laboratory. A further disadvantage of cooled detectors would have been the need to cool the input F. E. T. transistors in both the charge sensitive and voltage sensitive preamplifiers; this would have introduced constructional problems in the
detector head assembly.

(a) Time-to-amplitude converter

In a review paper by Ogata et al. \(^{(49)}\) it was noted that the pulse overlap type of time-to-amplitude converter appeared to give better differential linearity for short time ranges than the start-stop type. As there was no pulse overlap converter commercially available, it was decided to develop one for this experiment to operate in the range 0 - 50 nanoseconds.

(b) Pulse height compensation

Since leading edge timing was selected, it must be followed by some form of pulse height compensation. The type of system which seemed to give best compensation was one by Thieberger \(^{(57)}\) which utilized a second time-to-amplitude converter operating on the slow channels in the reverse direction to the fast converter. It was a simple matter to feed a slow converter with square wave pulses from a commercial timing discriminator and sum the outputs of the two converters.

(c) Data collection

To stabilize the timing chain it was intended to use some form of feedback stabilization by injecting a standard event at the beginning of the system - as is common practice in energy spectroscopy. The Northern Scientific NS 627 A. D. C. provided the facility of up to 8192 channel digitization combined with a built-in stabilizing system and so seemed the best choice of A. D. C. to couple to the PDP/6 computer for our purposes.
(f) Other equipment

Where possible, commercial modules were used for standard operations within the timing system and were bought from a local manufacturer (Nuclear Enterprises Ltd.) so as to have prompt service in case of malfunction. It was convenient (almost essential) to have at hand a small pulse height analyser (Laben 400 channel) as well as the computer system for preliminary setting-up purposes and for monitoring any drift in the energy channels during live timing runs.

A full technical description of our system, as initially set-up, follows in the next chapter.
CHAPTER 3

ORIGIINAL EXPERIMENTAL SYSTEM

3:1 Introduction

In this chapter a full description of the life-time measuring system, as originally constructed, is given – with the exception of the computer controlled pulse height analysis system, which is dealt with in chapter 4. Later, in chapter 5, we shall describe the modifications that had to be carried out in the light of measurements performed with this system.

Fig. 19 shows a block diagram of the complete electronic system. The nuclear radiation from source (S) is detected by solid state (silicon) detectors \( D_1 \) and \( D_2 \), operating in vacuo at \(-5^\circ\text{C}\). The bias voltages to the detectors are supplied by two high voltage units (H.T.). The fast leading edge of the signal from the detectors is picked off by a fast voltage amplifier (V/S amp.) before the whole charge is passed to a charge sensitive preamplifier (C/S amp.). The time mark generation is performed by an Ortec 260 time pick-off unit (T.P.O.), fed by the fast voltage preamplifier. The time mark signals from each channel go, via Ortec 403A (T.P.O. Control) units, to the two inputs of the home-built time-to-amplitude converter (Fast T.A.C.), the output of which goes through a 500 nanosec. delay cable before reaching the summing amplifier (Sum). The slow signals collected in the charge sensitive preamplifier (C/S Amp.) are amplified (Amp.) and used for energy selection by a single channel analyser (S.C.A.). As mentioned in chapter 2,
### Key to Diagram 12.

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>P/G</td>
<td>Pulse generator, Tektronix 111</td>
</tr>
<tr>
<td>S</td>
<td>Radioactive source</td>
</tr>
<tr>
<td>D₁, D₂</td>
<td>Radiation detectors</td>
</tr>
<tr>
<td>H.T.</td>
<td>Bias supply for detectors, NE 4605</td>
</tr>
<tr>
<td>C/S amp.</td>
<td>Charge sensitive preamplifier, NE 5267</td>
</tr>
<tr>
<td>V/S amp.</td>
<td>Fast voltage sensitive preamplifier, home-built</td>
</tr>
<tr>
<td>T.P.O.</td>
<td>Ortec 260 time pick-off unit</td>
</tr>
<tr>
<td>T.P.O. Control</td>
<td>Ortec 403 control unit</td>
</tr>
<tr>
<td>Amp.</td>
<td>Slow pulse amplifier, NE 4603</td>
</tr>
<tr>
<td>Fast T.A.C.</td>
<td>Fast time-to-amplitude converter, home-built</td>
</tr>
<tr>
<td>Trig.</td>
<td>Monostable circuit to trigger timing discriminator, home-built</td>
</tr>
<tr>
<td>Sum</td>
<td>Summing amplifier, home-built</td>
</tr>
<tr>
<td>T.A.C.</td>
<td>Slow time-to-amplitude converter, home-built</td>
</tr>
<tr>
<td>S.C.A.</td>
<td>Single channel pulse height analyser, NE 4602</td>
</tr>
<tr>
<td>T. Disc.</td>
<td>Timing discriminator, NE 4616</td>
</tr>
<tr>
<td>Disc.</td>
<td>Integral discriminator, NE 4623</td>
</tr>
<tr>
<td>Coinc.</td>
<td>Triple coincidence unit, NE 4620</td>
</tr>
<tr>
<td>P.H. Analyser</td>
<td>Computerised multichannel pulse height analyser, NS 627</td>
</tr>
</tbody>
</table>

* denotes a delay cable
General view of nuclear life-time measuring system
where leading edge timing is used, some form of pulse height compensation is required. This is performed in this system by the use of a second time converter operating in reverse (time wise) to the fast time converter. Time mark signals are generated from the leading edge of the slow pulses by a timing discriminator (T. Disc.) and the pulses from both channels fed to the slow time converter (T. A. C.). The outputs from the two time converters are summed in a summing amplifier (Sum) and analysed by a multichannel pulse height analyser (P.H. Analyser). To select only the required signals for analysis, the input to the kicker is gated by a triple coincidence (Coinc.) circuit whose inputs are derived from the two slow channel analysers (S. C. A.) and an integral discriminator (Disc) at the output of the summing amplifier.

To ensure that pulse height compensation only occurs when a genuine output from the fast T. A. C. appears, a discriminator (Trig.) sets one of the timing discriminators ready for action. To generate a stabilising peak in the time spectrum, a pulse generator (P/G) sends a signal into each channel, one delayed slightly behind the other to offset the peak within the time spectrum.

3.2 Detectors, Housing and Vacuum System

A 1/2 mm. depletion layer silicon (lithium drift) detector (NE 200-2A) and a 2 mm. depletion layer silicon (lithium drift) detector (NE 200-2A) were used for the 976 KeV. and 482 KeV. K-conversion electrons of Bi$. These detector thicknesses gave a suitable compromise between good energy resolution of the 976 KeV. electrons and a fast charge collection. The source, on a thin foil, was sandwiched between the two detectors and the whole assembly mounted on
a cold finger inside the vacuum system (see fig. 20). The detectors were
designed to work at room temperature, but operation at a steady -5°C gave a
reduction in noise and eliminated the chance of any spurious effects due to
thermal variations. A bias of +200 volts was applied through a 10 MΩ resistance
to the 200-½A detector and +700 volts to the 200-2A detector; this is about
twice the recommended voltage but gave an improvement in time resolution
(by increasing the rise-time) whilst not noticeably deteriorating the slow
channel energy resolution. Leads from the detectors were passed through the
vacuum chamber via O-ring sealed F. E. T. plugs. The external leads to the
fast amplifiers are 5 cms. in length and to the charge sensitive amplifiers
a further 5 cms., being as short as physically possible.

The vacuum system, shown schematically in fig. 21, consists of standard
units — a Speedivac oil vapour diffusion pump (DP), model EO 2, backed by
a rotary pump (RP), model 1 SC 50 A. The vacuum chamber can be isolated from
the pumping system by a butterfly valve (BV); a roughing line is also
provided. The pressures are monitored in the chamber by a Penning Gauge (Pen)
and in the backing line by a Pirani Gauge (Pir). A pressure of 3 × 10⁻⁵ torr
is easily maintained and the time to reduce the pressure below 10⁻⁵ torr is
less than 5 minutes. The cold finger (C) is attached to a container (F) inside
the vacuum chamber; Anti-freeze (A) is pumped through the container by a
Grant's pumping refrigeration unit, type LB 4. This unit thermostatically
maintains the temperature to within ½°C.

For safety there is an oil vapour trap (T) containing flowing cold water
(W) to condense any vapour which might otherwise damage the detectors and a
magnetic valve (SV) is incorporated to guard against mains failure. The various
Key to diagram

D₁ and D₂ - detectors
S. H. - source holder
S - spacers
H - brass detector holder
C - co-axial connector leads
C. F. - copper cold finger
vaccum lines are controlled by manual valves (V).

3:3 Fast Electronics

(a) Fast preamplifiers

The electronic resolution of the Ortec 260 time pick-off units, used on their own, is equivalent to 600 KeV.; thus, to observe particles of this order of energy a fast low noise preamplifier is required. The method of time pick-off in the Ortec 260 units is performed by passing the signal from the detector to the charge sensitive preamplifier through a specially designed transformer which senses the current flowing (see fig. 18a-d). The current in the secondary winding is passed to the input of a multistage amplifier before discrimination by a tunnel diode monostable circuit. Sensing the current flowing to a detector is indeed the way to obtain the fastest rise-time pulse (69), but it does not necessarily give the best time resolution (as defined in section 2:10a). Sherman et al. (71) have tackled this problem and designed a low noise F. E. T. input preamplifier which, although sensing the collection of charge at the detector rather than the current flowing to the detector and hence a slower rise-time pulse, gives a superior time resolution. Two such units were constructed (fig. 22) and they operate as follows. The fast and slow systems are joined in parallel by an inductance (4.70 µH) which delays the transfer of charge into the slow charge sensitive preamplifier whilst generating a voltage pulse at the input of the fast preamplifier. The inductance is large enough to allow full voltage amplitude at the F. E. T. of the fast amplifier before the pulse decays as the charge transfers to the
a. +24 V

Fig. 22

resistors in ohms
 capacitors in microfarads, unless stated

FAST PREAMPLIFIER  Fig. 22
slow channel. The 4.7 kΩ resistance in parallel serves to eliminate ringing due to the inductance but it unfortunately contributes a noticeable amount of noise to the slow channel. In an effort to reduce this, the resistance was removed and the value of the inductance decreased until the ringing ceased. However, this gave a ½ decrease in signal voltage in the fast preamplifier, therefore it was decided to retain a larger fast channel signal at the expense of the slight (10 kev.) energy degradation in the slow channel. The resistance (R) in the source of the F. E. T. was set for a current of 8 mA through the transistor — here, 19 ohms.

Across the output line of the fast preamplifier was inserted an integrating capacitor. According to Douglas et al. (83), the insertion of a low pass filter whose time constant is of the same order as that of the passing pulse provides optimum time resolution. This was found to be so, experimentally, as shown in the table in fig. 23. The capacitor also helps to damp out any unwanted residual ringing in the circuit.

The preamplifiers were constructed on double sided printed circuit board, laying out the components with extreme care in order to maintain the fast (2 nanoseconds) pulse rise-time and to keep ringing to a minimum. The output was fed to the current transformer in the Ortec 260 unit, the other end being terminated by a 50 ohm resistor (see fig. 24b). By varying the value of the plug-in resistor (R) at the input of the 260 unit, the gain of the fast amplifier chain could be controlled.

The 260 unit consists of a fast amplifier followed by a tunnel diode discriminator whose triggering level is controlled by the Ortec 403A unit. The square wave output from the 260 unit was found to vary in length with
Low pass filtering of signals in fast voltage amplifier from 200-μA detector

<table>
<thead>
<tr>
<th>Integrating capacitor (pF)</th>
<th>Time constant 2.2 RC (μS)</th>
<th>Energy spread ΔE (Kev.)</th>
<th>Rise-time R (nS)</th>
<th>Time spread ΔE/R 1 Mev. (nS)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>-</td>
<td>130 ± 10</td>
<td>35 ± 5</td>
<td>4.5 ± 1</td>
</tr>
<tr>
<td>200</td>
<td>22</td>
<td>100 &quot;</td>
<td>40 &quot;</td>
<td>4.0 &quot;</td>
</tr>
<tr>
<td>600</td>
<td>66</td>
<td>100 &quot;</td>
<td>75 &quot;</td>
<td>7.5 &quot;</td>
</tr>
</tbody>
</table>
Fast Tunnel Diode Monostable

Coupling in Ortec 260 time pick-off unit
To provide a standard square pulse to feed the overlap type time-to-amplitude converter, a further tunnel diode monostable circuit was inserted after the 260 unit. The circuit, shown schematically in fig. 24a, was a modification of that used by Clausen and Nainan (24b). The length of the output pulse from the Ortec 260 unit was 20 nS; the monostable was set by a suitable value of the inductance L to generate a pulse of 80 nS. The monostable circuit was constructed on a small printed circuit board and mounted inside the 260 casing.

(b) Time-to-amplitude converters

Both time-to-amplitude converters are of the pulse overlap type and, as mentioned in chapter 2, give a doubled output. Therefore, to obtain a single value from the fast time converter (the signals to the slow time converter are gated before conversion), the output signal is passed through a linear gate (NE 5730) controlled by a fast coincidence circuit as shown in fig. 27.

The time-to-amplitude converters are based on a design by Simms (48); the circuit of the fast T.A.C. is shown in fig. 25. The two input transistors, T₁ and T₂, are biased so as to be conducting in their quiescent state with transistor T₃ cut off. When either T₁ or T₂ is cut off by a negative input pulse, from the 80 nS monostable in the 260/405A combination, all the current from the constant current generator T₄ flows through the other transistor and T₃ remains off. When both T₁ and T₂ are cut off, the current is switched to T₃ and integrated on the 330 pF capacitor at T₃. The output from the collector, going to emitter follower T₅, has an amplitude, \[ V = I \cdot \Delta t / C \]
Fig. 25

resistors in ohms
-capacitors in microfarads, unless stated.

T₁, T₂, T₃, T₄ ~ P346A
T₅ ~ 2N1195

-24V

-6V

-6V

FAST T.A.C.
Fig. 26

Fast Coincidence Unit

-24V

3.3K

6.2K

150K

22 pF

1N994

V405A

-12V

.1

470

1N3717

22

2N3602

1K

2K

500

3V

6.8K

L

-24V

P346A

i/p 51

2K

51

100

1K

.1

6V

0.001

resistors in ohms
capacitors in microfarads,
unless stated

L - to give 500 ns pulse
Fig. 27
where \( I \) = current supplied by \( T_4 \),
\( C_B \) = capacitance at the collector of \( T_3 \),
\( \Delta t \) = period of overlap of the input pulses.

Assuming the input pulses to be flat topped, then \( V \propto \Delta t \).

The fast coincidence circuit (fig. 26) controlling the linear gate (normally held open) was designed and built along the same lines as the fast T. A. G., except that the current switched from the input P346A transistors causes a tunnel diode 1N3717 to trigger, generating the gating pulse. The coincidence circuit sorts out the order of overlap by vetoing, in the linear gate, one of the nodes of overlap.

The timing sequence of the overlap process may be understood from figs. 27 and 28. Fig. 27 gives the various (deliberate) delays inserted between the electronic modules in the two fast timing chains; from each 403A fan-out unit, one of the two fast negative pulses goes to the coincidence unit and the other to the T. A. G. such that a signal travelling along channel alpha will reach the coincidence circuit 48 ns after reaching the T. A. G. and a signal travelling along channel beta will reach the coincidence circuit 34 ns before reaching the T. A. G.. Fig. 28a shows the time relationship of a pulse in channel beta arriving at the T. A. G. approximately 20 ns before a pulse from channel alpha. The overlap of beta before alpha produces an output (o/p) pulse whose height is \( V \) and which is not vetoed by the coincidence unit. Fig. 28b, on the other hand, shows the other mode of overlap - the situation where an output pulse \( V \) is generated by the input pulse from channel alpha arriving approximately 20 ns before a pulse from channel beta; here the output pulse \( V \) is vetoed by a coincidence.
occurring at the coincidence unit.

To maintain speed and reduce ringing (thereby improving conversion linearity), the fast T. A. C. circuit was built on double-sided high frequency printed circuit board with careful attention paid to component lay-out. As the T. A. C. conversion does depend on the shapes of the incoming pulses, once calibrated, the interconnecting cables from the 260 units to the T. A. C. were not exchanged. Any delay required in either channel is inserted between the fast amplifier and the input of the 260 unit.

3:4  **Slow Electronics**

(a) Time-to-amplitude converter

The slow T. A. C., whose circuit is given in fig. 29, is almost identical to the fast T. A. C. except that it operates on a time scale approximately 20 times greater and with positive logic. The output is integrated to 0.5 μs rise-time in the output emitter follower for feeding to the summing amplifier. The input pulses to the slow T. A. C. come from the delayed (variable) output of the timing discriminators; these trigger on the leading edges of the signals from the respective slow channels. The length of output pulse from both discriminators was modified to 400 nS duration and the variable delay to 0 - 500 nS. The timing discriminator in channel beta is externally triggered by the output of the fast T. A. C. This removes unwanted single pulses from the slow T. A. C.
resistors in ohms

capacitors in microfarads,

unless stated

+24V

25 µF

T1 T2 T3 T4 ~ 2N1195

T5 ~ 2N1711
(b) Summing amplifier

The signal from the fast T. A. C. goes via a 500 ns delay cable to the "fast T. A. C." input of the summing amplifier (see circuit in fig. 30); the signal from the slow T. A. C. goes to the "slow T. A. C." input. The former signal passes through a two stage amplifier and a 0.5 μs rise-time shaper and then to the summing unit; the latter signal, which has already been shaped to 0.5 μs rise-time, passes through a one stage amplifier to the summing unit. There the two signals are added, inverted and fed to an output emitter follower, whence the composite signal is analysed by the pulse height analysis system.

(c) Commercial equipment

The rest of the slow electronics, apart from the computerized pulse height analysis system, is made up of standard N. I. M. units manufactured by Nuclear Enterprises Ltd. They comprise the following:

Detector bias supply NE 4605
Charge sensitive preamplifier NE 5287
Pulse amplifier NE 4603
Pulse height analyser NE 4602
Timing discriminator NE 4616
Integral discriminator NE 4623
Coincidence unit NE 4620
Scaler/timer NE 4612
Summing Amplifier

Fig. 30

 Resistors in ohms
 Capacitors in microfarads, unless stated
(a) Radiation pulse simulator

To help set up the experiment, an "electronic prompt source" was constructed (see fig. 31). This takes a negative square waveform from a Lyons PG 32 pulse generator and shapes it to simulate "live" radiation pulses. After shaping, in order not to alter the rise-time with variation of the 500 ohm potentiometer, the signal is passed through a double emitter follower which presents to the potentiometer an impedance of approximately 50 kΩ; the estimated rise-time variation through the full range of the potentiometer is less than 0.005% (i.e. 2 pS in 40 ns/raise-time).

(b) Logic converter

Because the Leben Kicksorter requires negative pulses of -4 volts and >½ μs wide to operate its coincidence/anticoincidence circuitry, the NS 627 A. D. C. requires positive pulses of +5 volts and >5 μs wide and the output pulses from the N. I. M. coincidence unit are +5 volts and ½ μs wide, a fan-out logic converter was constructed. This accepts any pulse >+5 volts and converts it into +5 volts at one output and -5 volts at the other, both 5 μs wide.

(c) Temperature stabilization

Fortunately the laboratory in which the present experiment was conducted had thermostatically controlled heating; thus the temperature was maintained between 25°C and 27°C.
Fig. 31

Radiation Pulse Simulator

Resistors in ohms
Capacitors in microfarads,
Unless stated
CHAPTER 4

PULSE HEIGHT ANALYSIS SYSTEM

4.1 Introduction

As part of this project, a pulse height analysis system was set up using "shared real time access to a small computer" (85). High resolution automated data acquisition systems (in our case a multichannel analyser) are so costly that it is more economical for a group of potential users to pool resources and purchase one small computer; to this they may collectively interface their respective experiments and have all and more than the advantages of a standard system. However, it does require a sophisticated Supervisor program (86); but once written, programming for the individual experiments is not difficult. The experiments in this laboratory currently interfaced to the computer are,

1. Wind tunnel control and data logger
2. Multichannel pulse height analyser
3. Vanguard particle track scanning machine
4. D. M. A. C. graph measuring system

As well, when operating on-line and off-line, program editing and computing facilities are available.

We shall limit ourselves to the relevant aspects of this particular project, as a fuller description has been given elsewhere (85).
The computer used is a Digital PDP/8 with 8K core store; peripheral devices to the central processor unit (C. P. U.) include 2 Teletypes, a high speed paper tape reader/punch, 3 D.D.C.-tape transports, an A/D and D/A converter and an oscilloscope. The front end of the pulse height analysis system, in a separate room, comprises a Northern Scientific 6192 channel A. D. C. type NS 627, an instruction console, a Tektronix storage oscilloscope type TM 554 and a Bryan 21001 graph plotter. A block diagram of the system is shown in fig. 32. Photographs of the computer and the front end are included.

The NS 627 analyser was chosen because of its internal digital stabilization. The operation of the stabilizer is fully described in the manual; but, briefly, an adjustable window looks onto a standard peak within the spectrum to be analysed and maintains the overall gain and the zero position of the whole system (not just the A. D. C.) constant with respect to that peak. In this experiment a standard time peak is generated by injecting two pulses from a pulse generator into the inputs of the preamplifiers (see fig. 19). They are slightly delayed with respect to one another so that the standard peak on which the system is to be locked lies in the upper part of the spectrum and out of the way of the live measurements.

4:3 **Interfacing of NS 627 to PDP/8**

There are two methods of interfacing an A. D. C. to a computer, either by "data break" or by "program interrupt". In the former, the core of the
"Front end" of computerized pulse height analyser
computer is connected directly to the A. D. C. such that the data by-passes the C. P. U. In the latter, the A. D. C. is coupled to a buffer which, on instruction, can be read by the C. P. U. The former allows very fast data collection but is restricting in that the locations for data acquisition are fixed; and the interface is costly. The latter is restricting on the speed of data acquisition (8 K p.p.s.) but permits a high degree of flexibility, the locations of the controlling program and the data store being controlled entirely by software; and the interface is relatively cheap. We use "program interrupt".

The drawings for the interface, which was designed and built in the laboratory, are shown in the included photographs; the interface wiring is given in the table in fig. 35. The 12 bit address of the NS 627 A. D. C. is coupled to the PDP/8 buffer by 25 yds. of 15 core screened cable; the STORE and FETCH commands occupy two of the remaining leads. Another 15 core screened cable interconnects the Push Button, 5 Instruction Switch bits, a -12 volt power line and 6 relay/indicator light lines to the computer. The X, Y, Z signals for the storage oscilloscope come through separate low loss coaxial cables; the graph plotter is brought into circuit simply by switching (by relay from the instruction console) the X and Y oscilloscope lines to the X and Y input terminals of the graph plotter. It is then treated by the program as a very slow sweeping oscilloscope.

4:14 Supervisor Program

In order to operate several experiments simultaneously it is necessary
Computer interface drawings
Computer interface drawings

PH.A. CONTROL
### Interfacing Table for MS 627 / PDP-8

#### Cable 1

<table>
<thead>
<tr>
<th>A.D.C. Logic</th>
<th>A.D.C. Plug</th>
<th>Cable Code</th>
<th>PDP/8 Interface</th>
</tr>
</thead>
<tbody>
<tr>
<td>DB 1</td>
<td>H</td>
<td>Light green</td>
<td>F24 D</td>
</tr>
<tr>
<td>&quot; 2</td>
<td>E</td>
<td>Black</td>
<td>&quot; E</td>
</tr>
<tr>
<td>&quot; 3</td>
<td>G</td>
<td>Brown</td>
<td>&quot; H</td>
</tr>
<tr>
<td>&quot; 4</td>
<td>A</td>
<td>Red</td>
<td>&quot; K</td>
</tr>
<tr>
<td>&quot; 5</td>
<td>B</td>
<td>Orange</td>
<td>&quot; M</td>
</tr>
<tr>
<td>&quot; 6</td>
<td>D</td>
<td>Yellow</td>
<td>&quot; P</td>
</tr>
<tr>
<td>&quot; 7</td>
<td>F</td>
<td>Dark green</td>
<td>&quot; S</td>
</tr>
<tr>
<td>&quot; 8</td>
<td>J</td>
<td>Blue</td>
<td>&quot; T</td>
</tr>
<tr>
<td>&quot; 9</td>
<td>L</td>
<td>Gray</td>
<td>&quot; V</td>
</tr>
<tr>
<td>&quot; 10</td>
<td>N</td>
<td>Red/green</td>
<td>F25 D</td>
</tr>
<tr>
<td>&quot; 11</td>
<td>R</td>
<td>Red/yellow</td>
<td>&quot; E</td>
</tr>
<tr>
<td>&quot; 12</td>
<td>T</td>
<td>Red/blue</td>
<td>&quot; H</td>
</tr>
<tr>
<td>STORE</td>
<td>Y</td>
<td>Pink</td>
<td>&quot; K</td>
</tr>
<tr>
<td>FETCH/CLEAR</td>
<td>W</td>
<td>White</td>
<td>&quot; M</td>
</tr>
<tr>
<td>EARTH</td>
<td>K,M</td>
<td>Shielding</td>
<td>Chassis</td>
</tr>
</tbody>
</table>

#### Cable 2

<table>
<thead>
<tr>
<th>A.D.C. Element</th>
<th>Compartoscode</th>
<th>Cable Code</th>
<th>PDP/8 Interface</th>
</tr>
</thead>
<tbody>
<tr>
<td>Push Button</td>
<td></td>
<td>Violet</td>
<td>F30 S,U</td>
</tr>
<tr>
<td>Switch 1</td>
<td></td>
<td>Dark green</td>
<td>F25 P</td>
</tr>
<tr>
<td>&quot; 2</td>
<td></td>
<td>Yellow</td>
<td>&quot; S</td>
</tr>
<tr>
<td>&quot; 3</td>
<td></td>
<td>Gray</td>
<td>&quot; T</td>
</tr>
<tr>
<td>-15V Power</td>
<td></td>
<td>Red/blue</td>
<td>&quot; V</td>
</tr>
<tr>
<td>Relay</td>
<td>10G</td>
<td>Red</td>
<td>F32 D</td>
</tr>
<tr>
<td>Light C</td>
<td>11G</td>
<td>White</td>
<td>&quot; H</td>
</tr>
<tr>
<td>&quot; D</td>
<td>12G</td>
<td>Green</td>
<td>&quot; K</td>
</tr>
<tr>
<td>&quot; E</td>
<td>13G</td>
<td>Brown</td>
<td>&quot; M</td>
</tr>
<tr>
<td>&quot; F</td>
<td>14G</td>
<td>Pink</td>
<td>&quot; P</td>
</tr>
</tbody>
</table>
to have the Supervisor Program. The essential functions of this are to provide the routines for multiprogramming and to handle the input/output for the various programs in an efficient way. In addition, it provides routines needed by more than one user. The advantages of running a program under such a system are that the facilities developed for the system as a whole are immediately available to all users. With very little programming effort, we have access to the high speed reader/punch or magnetic D.E.C.-tape at the central computer.

Multiprogramming is handled by keeping a queue of pointers to active programs. Each program must call the routine SUSPEND after running for not more than 1/20 th second. This action causes suspension of the program until all other programs have had a chance to run (for not more than 1/20 th sec. each) after which, the program is reactivated at the point from which it suspended itself. In effect, calling SUSPEND causes a pointer to the program to be placed at the end of the program queue and the program indicated by the first word in the queue to be entered (see fig. 34). This method of multiprogramming is particularly useful on a small machine which has no special multiprogramming hardware.

Simple Pulse Height Analysis Program

This program was written to test the system and to provide the basic essentials of a standard pulse height analyser. Although the A. D. C. has a potential of 8192 channels, it is restricted (for the sake of other on-line users) to 1024 channels, being quite sufficient for general use. The whole program is in two sections - the part which does the actual communication
Queue

Execute Program
and shift queue

1) Program A
2) Program B
3) Program C
n) Program A'

Supervisor Queue System

Fig. 34
with the A. D. C., accepting numbers and incrementing the appropriate channels, and the part which controls the overall functions of the system as an analyser. The former is an integral part of the Supervisor and is permanently held in core; it functions irrespective of the control program when an interrupt at the A. D. C. occurs; but, in processing, it tests a number of conditions set by the control program e.g. overflow? date-disable? etc. If any one of these conditions is not satisfied, the number is rejected returning the computer to normal operation. Only when the P. N. A. control program is in core is the system usable as a pulse height analyser.

Being a standard program, a copy of the P. N. A. control program is held on magnetic D.E.C.-tape and with the following instructions via the Teletype may be brought down into core,

\[
\begin{align*}
E &: 7400 : X : 4477 \\
N &: X : 4027 \\
N &: X : 7400 \\
I &: 7400
\end{align*}
\]

and to end the program, the following instruction,

\[
E &: 7403 : 4450
\]

Here, X refers to the previous contents of the address being examined printed out by the machine. Once the program is loaded into the computer, all further instructions are given from the Instruction console beside the A. D. C.
by setting the appropriate switch pattern and depressing the Push Button.
The instruction patterns on the switch register are as follows,

<table>
<thead>
<tr>
<th>Instruction</th>
<th>Pattern</th>
</tr>
</thead>
<tbody>
<tr>
<td>Enable using double precision</td>
<td>0000</td>
</tr>
<tr>
<td>Enable using single precision</td>
<td>1000</td>
</tr>
<tr>
<td>Disable</td>
<td>0100</td>
</tr>
<tr>
<td>Test</td>
<td>1100</td>
</tr>
<tr>
<td>Erase</td>
<td>0010</td>
</tr>
<tr>
<td>Plot</td>
<td>0111</td>
</tr>
<tr>
<td>Scope on/off</td>
<td>1110</td>
</tr>
</tbody>
</table>

The Enable/Disable commands tell the computer to accept or reject data from the A. D. C. The Enable/double precision splits the memory into two halves using corresponding channels in the second half as overflow counters for the first, thus increasing the maximum count from 4095 to 8388607 i.e. approx $10^7$. This, however, halves the resolution to 512 channels. The Test command generates a St. Andrew's cross in the memory for aligning up the oscilloscope display. The Erase command sets all data locations to zero. The Plot command requires 3 successive depressions of the Push Button: the first holds the display routine in readiness at channel 1, the second initiates a very slow sweep of the memory and the third restores normal oscilloscope display. The Scope on/off command provides the facility of writing the display on the oscilloscope in storage mode, then shutting down the time consuming display routine. This is useful where spectrum accumulation is slow and only intermittent inspection is required. The current mode of operation is
shown by illumination of one or more of the indicator lights.

A flow chart of the P. H. A. control program is given in fig. 35. In the idle state the program tests the Push Button, displays a section of the spectrum and returns control to Supervisor (as displaying the spectrum is one of the few routines which requires a large amount of C. P. U. time, the display is split into sufficiently small sections to avoid stealing time from other users; also, it is interleaved to reduce flicker). If the Button is depressed, the program reads the switch register and branches to the appropriate routine. Being a lengthy process, the Plot routine is broken up by a SUSPEND instruction between each point plotted.

The actual taking of data from the A. D. C. is done under INTERRUPT. As soon as an address in the A. D. C. buffer is ready for transfer, a STORE command is generated causing an INTERRUPT in the computer. The computer ceases normal operation to search out and process the interrupting device; on completion, it passes a CLEAR/FETCH command back to the A. D. C. which in turn resets itself in preparation for the next analysis. In order to give every priority to INTERRUPTs from fast devices such as the A. D. C., the major part of the processing of INTERRUPTs from slow devices (such as the Teletype) is done by a Secondary Interrupt Processor Routine (see fig. 36), which itself may be interrupted by a further hardware INTERRUPT. In this way, two priority classes of INTERRUPT are established — a high and a low — without the expense or complication of a hardware Automatic Priority Interrupt system.
Fig. 35

[SUSPEND]

(F/B Flag set?) \rightarrow (Scope switch on?) \rightarrow [Disp. Spectrum]

: Read and do instruction

Plot Routine

[Disp. point 1]

Start \rightarrow [SUSPEND]

Stop \rightarrow (Plot flag set?)

Erase \rightarrow [Disp. next point]

Test \rightarrow (End of spectrum?)

Scope Switch

P.H.A. Control
Interrupt Procedure

Fig. 36
Since core space is at a premium when operating in shared real time access mode, once a spectrum has been accumulated, the P. H. A. control program may be deleted from core to make space for auxiliary routines such as the PUNCH and CENTROID routines. The PUNCH routine punches out the channel contents along with channel identification in octal numbers. This is converted into typescript off-line on the spare Teletype. The CENTROID routine calculates the centroid position of a single (time) spectrum, printing the answer out on the on-line Teletype. It firstly calculates,

\[
\text{DENOM} = \sum_{0}^{1023} n_x \quad \text{where } n_x = \text{no. of counts in channel } x,
\]

storing the number in octal double precision. Secondly it calculates,

\[
\text{NUM} = \sum_{0}^{1023} n_x \cdot x \quad \text{where } x = \text{the channel number},
\]

storing the number in octal triple precision. Both numbers are converted to decimal and printed out. It would be wasteful to attempt to calculate

\[
\text{CENTROID} = \frac{\sum_{0}^{1023} n_x \cdot x}{\sum_{0}^{1023} n_x}
\]

on the computer, as it is much simpler to perform the calculation on a 15 place desk-top calculator.

These auxiliary programs are all initiated by loading the relevant binary
tape into the high speed reader and typing.

I

I : 7200

The pulse height analysis program may easily be returned to the machine from magnetic tape as described above.

A further standard program which is used is the BINARY PUNCH. One of the problems with most nuclear spectroscopy experiments is that they require long periods of time to collect a significant number of live counts. Our PDP/8 computer is often used twice a day for communication with the Regional Computing Centre's IBM 360/50, which uses up the entire core space of the PDP/8. The BINARY PUNCH program extracts from the core the partially completed spectrum onto paper tape; then, after the communications run, when the Supervisor has been reinstated, the spectrum may be restored to core from the paper tape. The instructions for the whole operation are as follows: insert the BINARY PUNCH binary tape into the high speed reader and type,

I : 7400

CTRL T

F 5600 (Computer address of channel 1 of data)

F 7577 ( " " " 1024 " )

ALT MODE

and the data tape is produced. The contents are restored to core by loading
the data tape into the high speed reader and simply typing.

To return to standard pulse height analysis, the F. H. A. control program must be reinsered (as before), since the BINARY PUNCH program overwrites it. A listing of all the programs developed for this project are given at the end of this thesis in the appendix.
(a) Calibration of time-to-amplitude converters

There are three main methods of calibrating time-to-amplitude converters. They are:

1. Delay insertion
2. Statistical coincidence
3. Use of Sampling Oscilloscope

The first of these is the simplest; a pulse from a pulse generator is divided and passed along the two fast channels to the T. A. C. (see fig. 37a). Different delays either by electronic means or by delay cable are inserted into one of the fast channels and the resultant output pulse is measured. The accuracy here is limited to the error in measuring the inserted delay.

The statistical coincidence method or "white spectrum" method (fig. 37b) feeds a regular pulse train, whose interval is greater than the range of the T. A. C., into one channel and a random pulse train, such as would be generated by triggering on nuclear radiation or noise, into the other. There is an equal probability of any period of time delay occurring between the stop and start pulses, and so there is no preferred pulse height from the T. A. C. if its conversion is truly linear \(^{38,39}\). If this signal is fed into a multi-channel pulse height analyser, a flat or "white spectrum" should appear.
However, the method has some disadvantages; firstly, for a non-linear converter, it gives results which are not easily interpretable and, secondly, it does not provide an absolute time calibration. But it does give a very sensitive test of the differential linearity.

To understand the method of calibration using a sampling oscilloscope, it is necessary to appreciate the operation of the sampling unit itself. On an oscilloscope one can increase the speed of complete pulse display to a certain limit only — this limit being determined by the speed of the amplifying circuitry and the cathode ray tube within the instrument. Beyond this, one must use the sampling technique in conjunction with a train of identical input pulses. Basically, an instantaneous voltage is sampled in successive steps from each pulse and the whole reconstructed on the screen. This is achieved as follows (see fig. 38). The horizontal sweep is produced by a slow staircase voltage which is incremented each time a sample is to be displayed. One excursion of the input signal causes the trigger circuit to perform one cycle of the sampling operation, producing one dot on the display. The cycle starts when the trigger circuit recognises an excursion of the triggering signal and unclamps a fast ramp generator which produces a rundown voltage to be compared to the slow staircase voltage. The resulting comparison pulse is sent to the vertical system as a sampling-drive pulse and to the staircase circuit as a staircase advance pulse. The sampling circuit then samples the input signal and advances the staircase generator by one step. The output from the sampling memory (which holds the instantaneously sampled voltage level until changed by the next sample) is applied to the vertical amplifier and the new staircase output level is applied to the
horizontal deflection of the oscilloscope. Each subsequent triggering event initiates the same sequence of events but, since the staircase voltage moves down one step each time, the fast ramp has to run slightly farther each time before a comparison pulse is produced. In this way, the sampling event is delayed by successively longer intervals with respect to the triggering point. It is this sampling pulse which is used to stop the T. A. C. and the input signal which is used to start it, thereby giving successive integral intervals of time difference between the stop and start signals. The T. A. C. output, when fed into a multichannel pulse height analyser, gives a series of peaks corresponding to the interval of time stepped by the sampler. On some Kicksorters such as the Laben 400 Spectroscope, the integrating facility permits an insight into the integral linearity. Using the sampling oscilloscope method of calibration, Cho and Bohm(94) claim an accuracy of a few picoseconds.

There are other methods similar to this where the sampling 'scope is replaced by two fast pulse generators running slightly out of phase with one another. These likewise produce a train of start-stop pulses with successively increasing time intervals(68).

The method of calibration used here is somewhat simpler than that of Cho and Bohm(94). As the T. A. C. is used in conjunction with the time pick-off units, standard Tektronix oscilloscope probes were plugged into the inputs of the 260 units and used, as probes, to take off the signals from the relevant points (X and Y in diagram in fig. 38) within the sampling unit. In order not to upset the operation of triggering and of the ramp generators the start pulse was taken from point X (fig. 38) where the input signal is
Calibration Output from Fast T. A. C.

Calibration Output from Slow T. A. C.

Fig. 39
Fast T.A.C.
Calibration

Counts

Pulse Height

0 - 4 Volts

4 - 8 Volts

Peaks at 1 nanosecond intervals

Fig. 40
FAST T.A.C. CALIBRATION

Pulse Height

Channel No.

Delay

60 nS
about to be sampled, and the stop pulse from point Y, where the sample-drive pulse enters the diode sampling bridge. A Tektronix type 111 pulse generator provided a fast input signal of approx. 0.5 volts to initiate the sampling cycles. Photographs of a typical output from the two T. A. C.'s, when displayed on a slow sweeping oscilloscope, using the above method are shown in fig. 39; in the fast T. A. C., each successive output pulse corresponds to an increment of 1 nS. in the input time delay; in the slow T. A. C. the increment is 20 nS. A graphical plot of the fast T. A. C. output, via the multichannel analyser is given in fig. 40. Fig. 41 gives an integral plot of the delay against output pulse height.

(b) The Ortec 403A fault

On performing the above calibration procedure, it was noticed that the T. A. C. voltage/time conversion varied with discrimination level and also with count rate in the Ortec 260 unit. This was traced to a faulty circuit design in the 403A control units. Referring to the circuit in fig. 42, the -12 volt supply is fed to the unit through a 10 ohm safety resistor (as are all the supplies), with no stabilizing element (e.g. Zener diode) to maintain a fixed voltage on side A of this resistor. Thus, when a pulse passes through the unit, the various monostables go into action drawing current from the -12 volt supply and thereby depressing the potential at point A. Since the tunnel diode (in the 260 unit) discriminating current is also taken from point A, the larger the current drawn through the resistance chain, the lower the voltage at A; and the higher the count-rate (e.g. by lowering the discrimination level) the lower will fall the potential at A. So the discriminator
Fig. 42
level would appear to be non-linear and a function of the count-rate. This effect is not large and thus not readily noticeable but, when looking for very small time measurements, any such variation could give incorrect results. The fault was remedied by shorting out the four 10 ohm safety resistors in each unit.

(c) Development of the present time-to-amplitude converter

As mentioned in chapter 2, the pulse overlap type of converter is reported to be superior in differential linearity to the start-stop converter for short time measurements. Therefore it was decided to construct a pulse overlap unit and compare its characteristics to those of a commercial unit of the start-stop type for the same time range (a Nuclear Enterprises NE 4645 converter, kindly loaned by the firm). A time-to-amplitude converter, published by Simms (A8), was built and tested. This unit was found to have rather poor linearity and suffer badly from thermal drift. The latter was traced to the fact that the original circuit used the stray capacitance of a transistor on which to integrate the time proportional charge; this was partially remedied by adding a 330 pF capacitor in parallel with the stray capacitance (see circuit diagram of T. A. C. in fig. 25). To reduce the remaining thermal drift, a temperature compensation circuit was incorporated into the circuit; two diodes were placed in series with the potential divider which normally clamps the base of the constant current generator. The temperature coefficient (i.e. variation of output pulse height with temperature) for this circuit was measured to be +0.02% per °C. Final stabilization, whilst in operation, is effected by the NS 627 digital stabilizer in the A. D. C.
The action of the time-to-amplitude converter essentially depends upon some electronic switch (transistor or diode) being switched on (or off) to pass current to a capacitor for the period of time, being measured, before being switched off (or on) - or the current rerouted by some other switch. Now all fast switching circuits are prone to ringing, i.e. they do not generate a perfect step function, but slightly overshoot and return to the second state by a decaying oscillation (as shown in fig. 4.5). If the length of this oscillation is of the same order as the period of time being measured, it then constitutes a non-linearity in charging current and hence a non-linearity in time conversion. The problem, as it seems, is that it is virtually impossible to construct a time-to-amplitude converter for short time measurements which is perfectly linear.

In the start-stop converter, the transistors, in order to give good linearity for longer time measurements, are overdriven from fully on to fully off; this leads to a comparatively large amount of ringing for the first 50 ns, or so and thus poor linearity in this range. So the train of thought followed was, instead of overdriving the transistors from one state to the other, to switch them from just on to just off with a small but clean square wave pulse. But this passes the onus of non-ringing back to the circuit generating the (ideally) square wave pulses. Fortunately the fast negative pulses from the Ortec 403A unit, when correctly terminated, approach the ideal of a square wave with very little ringing, so these pulses, whose length was dictated by the fast tunnel diode monostable just before the 403A unit, were taken directly to the input of the time-to-amplitude converter. The pulses were adjusted to be approximately 1 volt high, simply by altering a few resistors within the 403A unit).
Fig. 43

Ideal

Practical

Charging Current

T. A. C. Output

T. A. C. Operation
A second drawback of Simms' circuit was a large (15 KΩ) resistor connected to a 150 volt supply which served as a constant current generator. This generated approx. 1.5 watts of heat which disturbed the components nearby. It was replaced by a constant current generator circuit into which was incorporated the thermal drift compensator.

In fast electronic circuits, the layout of components is of paramount importance. After several circuits were built and modified, the best converter was obtained by building it on high frequency double sided printed circuit board and endeavouring to make the path of the signal as short as physically possible. Combined with careful termination of cables and stringent A.C. decoupling throughout, this eventually gave a time-to-amplitude converter whose differential linearity, as measured by the "statistical white noise" method, was an improvement on that of the commercial unit measured under identical conditions. The results of the two may be compared in fig. 44; to be fair to the commercial unit, the "white noise" spectrum of the range 0 - 500 ns. is also included. An integral calibration of this time converter is presented in fig. 41 with the actual digital values of the 1 nanosecond time intervals, as measured by the sampling 'scope method, given in the table in fig. 45.

The slow converter was built on the same lines as the fast converter except that the output conversion gain is continuously variable by a multi-turn potentiometer for the accurate setting up of pulse height compensation. A graph of conversion gain vs. potentiometer setting is given in fig. 46, and a differential "white noise" spectrum given in fig. 47.
Differential Spectra

**COMMERCIAL T.A.C.**

Counts

0 50 nS

**FAST T.A.C.**

Counts

0 50 nS

**COMMERCIAL T.A.C.**

Counts

0 500 nS

Fig. 44
**Fig. 45**

**CALIBRATION OF FAST T: A.C.**

NS 627 A.C. settings:  
Conversion gain: 0 - 2048 channels (0 - 8 volts)  
Input: 0 - 10 volts  
Zero level: 10%; 010

Time mark separation = 1 ns.

<table>
<thead>
<tr>
<th>Channel Number</th>
<th>Differential no. of chans</th>
<th>Channel Number</th>
<th>Differential no. of chans</th>
<th>Channel Number</th>
<th>Differential no. of chans</th>
</tr>
</thead>
<tbody>
<tr>
<td>42</td>
<td>-</td>
<td>471</td>
<td>25</td>
<td>914</td>
<td>23</td>
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<td>66</td>
<td>24</td>
<td>494</td>
<td>23</td>
<td>937</td>
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<td>90</td>
<td>24</td>
<td>517</td>
<td>23</td>
<td>959</td>
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<td>115</td>
<td>25</td>
<td>541</td>
<td>24</td>
<td>982</td>
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<td>136</td>
<td>21</td>
<td>564</td>
<td>23</td>
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<td>157</td>
<td>21</td>
<td>588</td>
<td>24</td>
<td>1028</td>
<td>23</td>
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<td>178</td>
<td>21</td>
<td>612</td>
<td>24</td>
<td>1051</td>
<td>23</td>
</tr>
<tr>
<td>197</td>
<td>19</td>
<td>635</td>
<td>23</td>
<td>1074</td>
<td>23</td>
</tr>
<tr>
<td>218</td>
<td>21</td>
<td>658</td>
<td>23</td>
<td>1097</td>
<td>25</td>
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<tr>
<td>239</td>
<td>21</td>
<td>681</td>
<td>23</td>
<td>1119</td>
<td>22</td>
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<tr>
<td>263</td>
<td>24</td>
<td>704</td>
<td>23</td>
<td>1143</td>
<td>23</td>
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<tr>
<td>286</td>
<td>23</td>
<td>728</td>
<td>24</td>
<td>1166</td>
<td>23</td>
</tr>
<tr>
<td>309</td>
<td>23</td>
<td>751</td>
<td>23</td>
<td>1187</td>
<td>24</td>
</tr>
<tr>
<td>333</td>
<td>24</td>
<td>775</td>
<td>24</td>
<td>1211</td>
<td>24</td>
</tr>
<tr>
<td>355</td>
<td>22</td>
<td>797</td>
<td>22</td>
<td>1235</td>
<td>24</td>
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<tr>
<td>377</td>
<td>22</td>
<td>821</td>
<td>23</td>
<td>1257</td>
<td>22</td>
</tr>
<tr>
<td>400</td>
<td>23</td>
<td>844</td>
<td>23</td>
<td>1277</td>
<td>20</td>
</tr>
<tr>
<td>424</td>
<td>24</td>
<td>867</td>
<td>23</td>
<td></td>
<td></td>
</tr>
<tr>
<td>446</td>
<td>22</td>
<td>891</td>
<td>23</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Differential linearity = ± 1 channel over used range i.e. ± 4%
Fig. 47

**Slow TAC - Differential Spectrum**

![Graph showing a slow TAC differential spectrum with counts on the vertical axis and time in nanoseconds on the horizontal axis. The graph shows a flat baseline followed by a sharp drop at 400 ns.](image-url)
(a) Introduction

In this section we will describe the experimental procedure followed in an attempt to measure a short nuclear life-time by the method of "self-comparison of the centroid shift"; the following section will describe the modifications that were made subsequently to the apparatus in the light of the initial results. The timing was performed by triggering the fast discriminators on the leading edge of the 1 Mev. and 500 Kev. (approx) conversion electron pulses from the radiation detectors, the triggering levels being set such that the ratio of the discriminator level to the mean average maximum height of pulses for each conversion electron was the same.

Firstly the parameters in the slow channels of the system were set. The bias voltages to the detectors were set deliberately high to attain as fast a rise-time as possible in the fast channels, yet not so high as to seriously deteriorate the energy resolution in the slow channels. The amplifier integration and differentiation constants were optimised experimentally and the gain adjusted in both channels so that the 500 Kev. and the 1 Mev. conversion electron peaks sat at channels 156 and 336 respectively on the Laben Kicksorter with the back-bias set at 0.30 (this was used as a visual check on drift during live runs). The slow amplifier output of the channel not being monitored by the Laben was terminated with a 6.8 K. resistor; this being the empirically measured input impedance of the Kicksorter. The single channel analysers were set to accept the two conversion electron peaks by
gating the input of the Kicksorter with the output of the analysers. The slow coincidence unit resolving time was set to 1 µs and the delays in each channel were set to \( \frac{1}{2} \) µs. The centroids of the two energy peaks accepted by the single channel analysers were determined - on the Laben by using the Integration facility and on the NS 627/PDP-8 by using the CENTROID program. The various experimental settings for the whole system are tabulated in fig. 46.

The procedure for setting the fast channel parameters was rather involved (especially as it was discovered that the discriminator level in the Ortec 260 unit was in no way linear), so the process will be described in detail. The purpose was to set the fast discriminators to trigger on a constant fraction of the full pulse height for the different energies being detected. This was determined by viewing the output pulses from the fast time-to-amplitude converter, one input channel of which was being triggered at a fixed point in time whilst the fast discriminator level in the other channel was being investigated - and then vice-versa. The operations are as follows.

\[(b) \text{ Matching pulses from the Simulator to live radiation pulses}\]

By gating the slow channel spectrum to the Laben Kicksorter with the output pulses from the corresponding fast discriminator, the triggering level in terms of energy as a function of setting could be viewed. Two settings were noted, one for approximately 300 KeV and one for approximately 150 KeV. Then four live timing experiments were performed, accumulating the time spectra in the pulse height analyser using the following discriminator settings,
# Experimental Settings for 207 Bi Life-Time Measurement Attempt

<table>
<thead>
<tr>
<th>Detector</th>
<th>Channel alpha</th>
<th>Channel beta</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>NE 200-2A</td>
<td>NE 200-2A</td>
</tr>
<tr>
<td>Bias</td>
<td>+700 V</td>
<td>+200 V</td>
</tr>
<tr>
<td>Preampl. gain</td>
<td>x1</td>
<td>x1</td>
</tr>
<tr>
<td>Slow amp. gain</td>
<td>128</td>
<td>256</td>
</tr>
<tr>
<td>Integration</td>
<td>0.4 μs</td>
<td>0.4 μs</td>
</tr>
<tr>
<td>Differentiation 1</td>
<td>0.4 μs</td>
<td>0.4 μs</td>
</tr>
<tr>
<td>Differentiation 2</td>
<td>0.4 μs</td>
<td>0.4 μs</td>
</tr>
<tr>
<td>Fast disc. level</td>
<td>500 Kev. 200</td>
<td>242</td>
</tr>
<tr>
<td></td>
<td>1 Kev. 800</td>
<td>900</td>
</tr>
<tr>
<td>Single channel</td>
<td>△E 036</td>
<td>036</td>
</tr>
<tr>
<td></td>
<td>E 500 Kev. 332</td>
<td>324</td>
</tr>
<tr>
<td>analyser</td>
<td>E 1 Kev. 701</td>
<td>657</td>
</tr>
</tbody>
</table>
This gave a measure of the rise-times of the 1 Mev. live pulses in each channel. Replacing the live radiation with the Simulator, the rise-time of the pulse from the latter was adjusted as near possible to give the same time shift for the two fast triggering levels (viz., 150 KeV. and 300 KeV. levels).

(c) The accurate determination of the constant fraction triggering levels on the fast discriminators.

A block diagram of the configuration of equipment required for this operation is shown in fig. 49. One output from the Lyons PG 23 pulse generator goes to one input of the time-to-amplitude converter via an Ortec 260 unit; the other output goes via the Simulator and Ortec 260 unit (whose level is to
be set) to the other input of the time-to-amplitude converter. From the second branch of the Simulator, the signal goes to the Kicksorter via an amplifier.

Viewing the output of the fast amplifier with a Tektronix 585A oscilloscope (2 ns rise-time), the Simulator pulse height was adjusted to be approximately the same height as the 1 MeV. radiation pulses. Then the gain of the slow amplifier (to the Kicksorter) was adjusted so that the output pulse fell into the channel (in the Kicksorter) corresponding to the energy centroid of the 1 MeV. radiation peak. So now, by altering the gain control on the Simulator, we have a pulse source whose variable pulse height could be accurately determined and set (relative to the approximate 1 MeV. setting). The centroid of the time spectrum was determined for a 1 MeV. simulated pulse with the fast discriminator triggering at approximately 300 KeV. (but whose dial setting was accurately noted); the pulse height from the Simulator was altered to coincide with the centroid of the 500 KeV. peak and the triggering level of the fast discriminator varied until the resultant time peak coincided with the time peak for the 1 MeV. pulse. This whole process was repeated for the other channel, thus providing the two required constant fraction triggering positions for each channel.

(d) Live runs

To perform a live run, the Tektronix 111 pulse generator which generated the stabilizing peak was set to 10 c.p.s. and the NS 627 A. D. C. stabilizer was "locked in" so that the peak appeared just off the upper end of the time spectrum, i.e. no counts from the pulser were recorded in the 1024
channels of the memory. The numerous analogue controls on the electronics were set to detect the 1 Mev. conversion electron peak in channel alpha and the 500 Kev. peak in channel beta for the first run - and vice-versa for the second run. Altering the fast discriminator levels for the second run, of course, shifted the position of the stabilizing peak; this had to be reset to maintain the previous gain of the system. During a live run, the energy spectra in the slow channels were monitored on the Laben kicsorter and any drift was manually corrected.

5:3 Modification of the Original Experiment

(a) Introduction

The experimental procedure described above consistently gave centroid shifts of about 2 nanoseconds, corresponding to the 500 Kev. electrons being emitted approximately 1 nsec. after the 1 Mev. electrons. The cause of this erroneous result is not fully understood, but is attributed to (a) the difficulty in accurately setting the constant fraction triggering levels in the Ortec time pick-off units and to (b) probable asymmetry of the peaks in the time spectrum. To show that the life-time is not, in fact, of this order, a single time spectrum was accumulated and a logarithmic plot made, as shown in fig. 50. If the life-time were 1 - 2 nsec., it would appear as a straight line with a slope of this value on the left hand side (delayed coincidence slope) of the plot. Due to the poor resolution of this time peak, we can only say that the
Time Spectrum of 1st excited state of Pb$^{207}$ (from Bi$^{207}$)

- Delayed Coincidence Slope: 720 pS
- Prompt Slope (R.H.S.): 640 pS
- Full Width Half Max.: 3.2 ns
- 1 Channel equivalent to: 78 pS

Fig. 50

Counts

Channel No.
Life-time is less than 720 psec.

The main difficulty identified in performing life-time measurements by the centroid shift method is that the output pulse shapes from semiconductor detectors for various energies of the incoming radiation are not identical, but depend on the type of detector, its dimensions, the energy of the radiation and the angle of incidence.

(b) The pulse shape from a P-I-N detector

In the ideal P-I-N detector, the electric field is constant throughout the intrinsic region. This is achieved, during manufacture, by injecting lithium atoms into the semiconductor material (silicon) to equalise the donor and acceptor concentrations, so reducing the residual space charge to zero.

With reference to the diagram in fig. 51, we consider of an electron-hole pair in the intrinsic region of a p-i-n detector of depletion depth \( d \), with an applied bias voltage \( V \) (the electric field \( E \) being \( \frac{V}{d} \)). If the electron-hole pair is created at a position \( x_0 \) then, after time \( t \), the electron will have moved to \( x \) where its velocity is given by:

\[
v = \frac{dx}{dt} = -u_e \cdot E
\]

\[
= -u_e \cdot \frac{V}{d}
\]

where \( u_e \) is the mobility of an electron in silicon. By integration then,

\[
x = x_0 - u_e \cdot \frac{V}{d} \cdot t
\]
According to Ramo's Theorem, the magnitude of the charge induced at the electrodes by the motion of the electron within the detector is:

\[ q_{e} = \frac{q}{d} \cdot \frac{x_{0} - x}{d} \]

where \( q \) is the electronic charge.

Therefore,

\[ q_{e} = q \cdot \frac{u_{e} V}{d^{2}} \cdot t \]

Similarly for the charge induced by the motion of a hole,

\[ q_{h} = q \cdot \frac{u_{h} V}{d^{2}} \cdot t \]

where \( u_{h} \) is the mobility of a hole in silicon and is approximately one third the mobility on an electron.

The time to collect these carriers at their respective electrodes from a point \( x_{0} \) is given by,

\[ t_{e} = \frac{x_{0} d}{u_{e} V} \quad \text{and} \quad t_{h} = \frac{(d - x_{0}) \cdot d}{u_{h} V} \]

Thus the profile of the leading edge of the resultant output pulse will have two distinct slopes as shown in fig. 51b, the relative contributions from the two slopes depending on the position of creation within the sensitive layer of the detector.
Fig. 51
We now consider a very simplified picture (fig. 52) of two radiations (500 keV and 1 MeV electrons) entering a 2 mm. Si(Li) detector, as in the experiment. According to the range-energy chart for silicon, a 500 keV electron will penetrate approximately 0.6 mm, whilst a 1 MeV electron will penetrate 1.6 mm. Assuming all the energy to be deposited very near to the end-point, the resultant output pulse profile for the two radiations may be estimated from the above theory; the profiles so calculated are shown in fig. 52b. From the two profiles it may be seen that their respective times of crossing a constant fraction discriminator level \(V_{\text{disc.}}\), which, in the original experiment, was \(\frac{1}{2}V_{\text{max.}}\), may well differ \((t_1\) and \(t_{11}\)). Since the actual process for the different radiations within the detector is far from simple or uniform, it is very possible for a timing error of a nanosecond to appear in pulses with a rise-time of around 50 nsec. Our solution to this problem is to perform the time measurement between two identical fast pulses extracted by means of thin transmission detectors and perform energy selection on the residual radiation passing to the thick detectors.

(c) The final experimental system.

From the range-energy chart for silicon a 1 MeV electron passing through a 200 micron thick silicon (surface barrier) detector will deposit a minimum of 72 keV of energy and a 500 keV electron will deposit 76 keV of energy. So, for timing purposes, the energy distinction (and hence any pulse profile difference) between the two radiations is removed. The residual energy of the radiation is absorbed by the thick detectors (positioned immediately behind each transmission detector) providing the requisite energy discrimination.
A block diagram of the new system is shown in fig. 53. The additional electronic modules required for the thick detector signal processing are from the Nuclear Enterprises Edinburgh Series (marked with "x" on the diagram). The two detection chains (thick detectors and thin detectors) are completely separate electronically with the exception of the combined coincidence circuitry (Coinc. 3) at the end of each chain. Physically, the four detectors are stacked upon one another, all mounted on the cold finger, as shown in fig. 54. The detectors were separated by small chips of insulating tape to prevent constricted air pockets from physically damaging the thin detectors during the pumping down or bleeding of the vacuum system. The vacuum chamber remains unaltered and is as described before.

The signal from each thick detector $D_1$ and $D_4$ ($\frac{1}{2}$ mm and 2 mm Si(Li)) pass through a charge sensitive preamplifier (c/a amp; NE 5287A) and a pulse amplifier (amp; NE 5256) to two single channel analysers (S.C.A.; NE 5159), one set to trigger on 500 kev electrons and the other set to trigger on 1 Mev electrons. Coincidence circuits 1 and 2 generate an output pulse when a coincidence occurs between a 1 Mev radiation pulse in one channel and a 500 kev radiation pulse in the other - or vice-versa. Both coincidence outputs go, via isolating inputs, to the same input line of coincidence circuit 3. Coincidence circuit 4 generates an output pulse of 80 microseconds, which also goes to bit 9 of the parallel address from the NS G27 analyser to the computer and has the effect of directing any signal from the T.A.C. into one of two halves of the data array space in the computer depending on the presence or absence of a pulse from "Coinc.1". Thus a time spectrum of 1 Mev electrons in channel alpha coincident with 500 kev electrons in channel beta is collected.
Key as for Diagram 19.
Key to Diagram:

- **D₁, D₄** - Thick Si(Li) Detectors
- **D₂, D₃** - Thin S.B. Detectors
- **H₁, H₄** - Detector Holders
- **T** - Coaxial Terminals to Detectors
- **M** - Detector Head Mount
- **F** - Cold Finger
- **S, P** - Bismuth Source mounted on Plate P
- **L** - Top Detector Location Collar

*Fig. 54*
in the first 512 channels of memory and vice-versa in the second 512 channels. The accumulation of both modes in the same run helps to eliminate any instrumental drift as both spectra should be equally affected.

The thin transmission detectors D2 and D3 were specially made (by Nuclear Enterprises) from 500 ohm-cm resistivity silicon and are 200 microns thick. They are run with a bias of 350 volts, just giving full depletion, and at a temperature of -20°C (as are all the detectors). At this temperature the noise contribution is approximately the same as from the first F.E.T. at the input of the fast preamplifier (v/s amp). Theoretically the pulse rise-time from these thin 500 ohm-cm detectors is 500 psec. - which is faster than the electronic circuitry (2 nsec. rise-time, but integrated to 10 nsec. to reduce random noise). The signal from the detector is amplified by the fast voltage sensitive preamplifier, fed to the Ortec 260 unit (T.P.O.) for time pick-off and then to the Time-to-Amplitude Converter (Fast T.A.C.). Instead of being terminated immediately after the input transformer of the Ortec 260 unit, the fast signal is passed to a terminated input of a slow pulse amplifier (Amp) after which energy selection is performed by the single channel analyser (S.C.A.) and passed to the coincidence circuitry. The pulse height compensation circuitry is incorporated as before, thus permitting the detection of radiation of a wider range of energies without deteriorating the time spectrum resolution. This has the advantage of increasing the countrate which, in this experiment, is important as the strength of the Bi207 source is only 1 microcurie. The compensated signal from the summing amplifier (Sum) passes through a 2 microsecond delay line to the MS 627 analyser to ensure that the coincidence gate
is opened well before its arrival.

From previous literature\textsuperscript{69,71} it had been inferred that the best time resolution may be obtained by using separate fast and slow preamplifiers for time and energy information respectively. Experimenting with various preamplifier configurations has led to the conclusion that only the fast voltage sensitive preamplifier is necessary coupled, via the time pick-off unit, to a slow pulse amplifier as shown in fig. 55c. In the parallel configuration of preamplifiers used in the original experiment (fig. 55b), noise from the input F.E.T.$^3$ of the charge sensitive preamplifier and from the resistor $R$ (there to damp oscillations from $L$) passes into the fast voltage sensitive preamplifier - and likewise with the charge sensitive preamplifier. A résumé of this experimentation is given in the table in fig. 55(d). (d) Setting up the final experiment

After the detector head and source were installed within the vacuum chamber, the system was pumped down to $3 \times 10^{-5}$ torr and the temperature of the cold finger slowly reduced to $-20^\circ$C over a period of 24 hours. Then the bias voltages to the detectors were applied in small steps to reduce the risk of electrical breakdown within the semiconductor material. The gains and pulse shaping constants of the slow amplifiers were set to give optimum energy resolution and the single channel analysers were adjusted to the required discrimination levels. Since the level of noise in the fast electronic chain (the fast preamplifier and the 260 unit) is not much lower than the size of the requisite live pulses, the fast discriminator levels were set down in the noise giving a triggering rate of between $10^5$ and $10^6$ c.p.s.
A

D → c/s preamp. → Amp. → S

B

D → c/s preamp. → Amp. → S

D → L → R

v/s preamp. → Ortec 260 → F

C

D → v/s preamp. → Amp. → S

v/s preamp. → Ortec 260

D - Detector
S - Slow channel output
F - Fast channel output
L, R - see text

Fig. 55
**Table of Energy Resolutions for Various Preamplifier Configurations**

shown in Fig. 55

<table>
<thead>
<tr>
<th>Configuration</th>
<th>Energy Resolution at 3°C</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Fast Channel</td>
</tr>
<tr>
<td>(a) c/a Preampl only</td>
<td>-</td>
</tr>
<tr>
<td>(b) c/a Preampl and v/a Preampl in parallel</td>
<td>100 ±10 Kev</td>
</tr>
<tr>
<td>(c) v/a Preampl only</td>
<td>50 ±10 Kev</td>
</tr>
</tbody>
</table>

Fig. 55(ii)
A table giving these settings and other relevant information is given in fig. 56.

After initial experimentation, because of the high noise level in the fast channels, it was found more convenient to accept radiation from the transmission detectors in the energy range 164 - 375 keV using pulse height compensation, rather than 80 - 100 keV without compensation as was originally intended since this improved the resolution in the live time spectra. This pulse height compensation system operates so that any variation in output pulse from the fast T.A.C. due to amplitude variation at the fast discriminators is cancelled by a similar variation in the opposite direction from the output of the slow T.A.C.\textsuperscript{57}. In the fast T.A.C., channel alpha provides the delayed overlapping pulse therefore, in the slow T.A.C., channel beta is set to provide the delayed overlapping pulse by adjusting the delay time potentiometer on the timing discriminator. To reduce chance coincidence counts from the slow T.A.C. the prompt output from channel alpha timing discriminator (T. Disc) is fed to the trigger input of channel beta timing discriminator which is on "external trigger". To set up the pulse height compensation initially, the rise-time ($t_{\text{fast}}$) of the pulses entering the Ortec 260 unit was estimated on an oscilloscope (Tektronix 585A, rise-time 2.2 nsec.); then the rise-time ($t_{\text{slow}}$) of the pulses in the slow amplifiers feeding the timing discriminators was measured. The ratio of the time conversion ($\frac{dv}{dt}$) of the two T.A.C.'s was adjusted according to the relationship\textsuperscript{57},

\[
\frac{t_{\text{fast}}}{t_{\text{slow}}} = \frac{\frac{dv}{dt}_{\text{slow}}}{\frac{dv}{dt}_{\text{fast}}}
\]
### Experimental Settings and Information

<table>
<thead>
<tr>
<th></th>
<th>CHANNEL ALPHA</th>
<th>CHANNEL BETA</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>2 mm Thick Detector</td>
<td>½ mm Thick Detector</td>
</tr>
<tr>
<td>Bias Voltage</td>
<td>+200 V</td>
<td>+350 V</td>
</tr>
<tr>
<td>Preamplifier Gain</td>
<td>×5</td>
<td>-</td>
</tr>
<tr>
<td>Preamp Integ. Const.</td>
<td>-</td>
<td>10±2 nS</td>
</tr>
<tr>
<td>Slow Amp. Gain</td>
<td>×580</td>
<td>×256</td>
</tr>
<tr>
<td>&quot; &quot; Integration</td>
<td>Min.</td>
<td>0.1</td>
</tr>
<tr>
<td>&quot; &quot; Differentiation</td>
<td>0.2</td>
<td>0.8</td>
</tr>
<tr>
<td>S. C. Analyser 1</td>
<td>100±500 Kev</td>
<td>-</td>
</tr>
<tr>
<td>&quot; &quot; 2</td>
<td>600±1000 Kev</td>
<td>164±373 Kev</td>
</tr>
<tr>
<td>Timing Disc. Bias</td>
<td>-</td>
<td>50 Kev</td>
</tr>
<tr>
<td>&quot; &quot; Delay</td>
<td>-</td>
<td>0</td>
</tr>
<tr>
<td>Fast Energy Resolution</td>
<td>-</td>
<td>70±20 Kev</td>
</tr>
<tr>
<td>Slow Energy Resolution</td>
<td>20 Kev</td>
<td>23 Kev</td>
</tr>
</tbody>
</table>

Fig. 56
To set the system more accurately, live time spectra from the thin detectors were accumulated; firstly, the energy range of 164 - 266 keV of the spectrum from the channel beta thin detector was divided into two using two single channel analysers, such that coincidences from energies of 164 - 266 keV were routed to the second half of the pulse height analyser memory and energies of 266 - 373 keV were routed to the first half. If there were perfect pulse height compensation, then there should be no relative shift between the centroids of the time spectra. Several live runs were performed varying the value of $\frac{dV}{dt}$ (set by the multi-turn potentiometer on the slow T.A.C.) until there was minimal shift between the two time centroids. The setting and amount of residual shift were noted and are given in tables "a" and "c" in fig. 57; the mean average energy, as defined by,

$$E_{\text{average}} = \frac{\sum n_i E_i}{\sum n_i}$$

(where $n_i$ is the number of radiation counts with energy $E_i$ summed over the energy range 164 - 266 keV for the lower half or over 266 - 373 keV for the upper half) of each half of the slow energy spectrum was determined using the "integration" facility on the Laben Kicksorter; the results are given in table "b" of fig. 57. With this $\frac{dV}{dt}$ setting on the slow T.A.C., the roles of channels alpha and beta were interchanged; now channel alpha energy spectrum was divided into two halves (164 - 266 keV and 266 - 373 keV) and corresponding live time spectra were accumulated simultaneously, the difference in centroids being noted (table "c" in fig. 57). Also, for correction purposes in conjunction with the above data, the mean value of the energy deposited in the thin detector
Information for setting up Pulse Height Compensator

(a) Rise time of fast pulses ($t_{fast}$) = 10±2 ns
   " " slow " ($t_{slow}$) = 350±50 ns

Time conversion of fast T.A.C. = 11.0 nsec/volt

Setting for slow T.A.C. time conversion = 0.091 V/nS
   = 0.0022 V/nS
   = 0580 setting

(b) Mean Average Energies computed as described in the text

<table>
<thead>
<tr>
<th>Channel Alpha</th>
<th>Channel Beta</th>
</tr>
</thead>
<tbody>
<tr>
<td>$E_{av}$ for 164 - 266 Kev</td>
<td>202±2 Kev</td>
</tr>
<tr>
<td>$E_{av}$ for 266 - 373 Kev</td>
<td>310±2 Kev</td>
</tr>
<tr>
<td>$E_{av}$ for 1 Mev radiation in range 164 - 373 Kev.</td>
<td>222±2 Kev</td>
</tr>
<tr>
<td>$E_{av}$ for 500 Kev radiation in range 164 - 373 Kev.</td>
<td>217±2 Kev</td>
</tr>
</tbody>
</table>

(c) Measured Time Centroids for pulse height compensation set at 0580.

- $E_{av}$ = 310 Kev
- $E_{av}$ = 202 Kev

Time Spectrum Centroids, ch.alpha
   " " " " , ch.beta

Fig. 57
in the energy range 164 - 373 Kev by the 1 Mev electrons and by the 500 Kev electrons was determined (table "b" in fig. 57).

Various energy spectra are given in figs. 58, 59, 60, 61 & 62 all taken on the Laben Kicksorter. Fig. 58 shows a typical ungated spectrum from the thin transmission detectors. Fig. 59 shows two spectra from a transmission detector gated by its associated thick detector; (a) is gated by radiation received in the thick detector in the range 600 - 1000 Kev and (b) by radiation in the range 10 - 500 Kev. Fig. 60 shows a typical ungated spectrum deposited in the 1/2 mm Si(Li) detector; when this spectrum is gated with a limited bandwidth of energies in the transmission detector, the resolution is improved, but obviously the count rate is reduced as is seen in fig. 61. And, finally, fig. 62 shows a spectrum of Bi 207 detected in the 1/2 mm Si(Li) detector without the transmission detector in front.

The settings for the NS 627 analyser when measuring the time spectra are given in the table in fig. 63. The digital stabilization facility (by locking into a standard peak) was not used this time, as any instrumental drift was cancelled by the simultaneous collection of the two requisite time spectra. The time peak was adjusted, by using the zero level shift, to fall in the centre of its respective half of the memory.
Spectrum from 200 micron Si surface barrier transmission detector.
(a) Spectrum of energy deposited in transmission detector by 1 Mev. electron.
(b) Spectrum of energy deposited in transmission detector by $\frac{1}{2}$ Mev. electron.
Fig. 60

Spectrum from \( \frac{1}{2} \) mm. Si(Li) detector with transmission detector in front.
Spectrum from $\frac{1}{8}$ mm Si(Li) detector gated by radiation deposited in the transmission detector in the energy range 75 - 105 Kev.
NS-627 A.D.C. Settings

Input : 0 - 10 Volts Bipolar

Conversion Gain : 204.8 Channels

Group Size : 512 Channels

Zero Level : 0 - 100% ; Approx. 150

\( u_i, L_i, D_i \) : 500

\( L_i, H_i, D_i \) : 030 ; 0 - 100%

A.D.C./PDP-8 : Enable ; Single Precision

Gain Control Section not used.
Zero Control Section not used.
The operation of the Pulse Height Analysis System under shared real-time access to the PDP-8 computer has not been entirely satisfactory. Unfortunately, the system in this laboratory has been overdeveloped and the situation has been reached where the computer is used so much that the various operators are queuing to use its facilities. One of the main drawbacks is that the PDP-8 is fully commandeered for up to three separate hourly sessions daily for the purpose of communicating data from other on-line experiments to the Regional Computing Centre; thus, not only are three hours of potential use of the analyser lost, but any possibility of long uninterrupted runs is ruled out.

Even if the computer were operated 24 hours per day under the Supervisor Program (i.e., in shared mode), there is too much interaction between different experiments. With two research groups trying to display their information on two separate oscilloscopes, the large amount of C.P.U. time required to perform these continuous displays causes severe flicker—especially as 1024 channels are being displayed on the P.H.A. screen. Having only one high-speed reader/punch for data input/output and only one Teletype for communicating instructions causes a bottleneck when more than one group requires these
facilities. If any hardware modifications for one research group have to be carried out, then the machine is out of commission for all other users.

As an academic exercise in stretching the capabilities of one small computer, the project has been successful. As a standard tool for nuclear physics research, a pulse height analyser operating under shared real-time access is a waste of time. It would be far better to have one small computer system, e.g. processor, Teletype, oscilloscope and interfacing, which is fully committed to an A/D/C. for pulse height analysis only or, alternatively, a standard multichannel analyser whose data output is connected directly to a computer on which any complex processing may be performed off-line.

6:2 The measurement of the life-time of the $f_{2/3}$ state of Pb$^{207}$

(a) Data processing procedure on the PDP-8

The relevant data appear in the computer memory in two adjacent array spaces - Data Field 1, addresses 5600-6577 and 6600-7577. At the end of each complete run this information is written onto magnetic DEC-tape using the following program inserted via the Teletype keyboard, the tape deck being set to channel 4.

E: 7403 4450 ..... stops P.H.A. program

E: 7200 4476 ..... call data read/write routine

N (for next) 0000 ..... (or any keyword code number)

N 43n1 ..... type n = 3 for read or n = 5 for write

N 7212 ..... address containing tape starting block number
With the data written onto magnetic tape, the array space is erased i.e. all addresses are set to zero, with the P.H.A. program. Using the above program, with the relevant alterations, the first four blocks (200 addresses per block) are read back from tape into the first half of the array space, leaving the second half blank. To reduce inaccuracies in the centroid measurement due to counts lying in channels well off the requisite time peak, a digital window of 12 ns (approx) is imposed upon the spectrum as follows. The position of the centre of the peak is estimated from the oscilloscope display; then, using the BINARY PUNCH program, the desired range of data (+ 240 channels from the peak centre) is extracted onto paper tape. The array space is erased and the data on the paper tape is reinserted. The centroid of this "trimmed" time peak is found using the CENTROID program. Then the second four blocks of data from the magnetic tape are brought down to overwrite the first half of the data array, still leaving the second half blank. The same digital window is
applied to this time peak and its centroid determined in the same manner.

(b) Calculation of result

Here we calculate the life-time measurement with correction for imperfect pulse height compensation for the final run performed with this system.

The uncorrected centroid shift between the two accumulated time spectra was found to be 5.74 channels, corresponding to the 1 Mev electron being emitted before the 500 Kev electron, as shown in the following table.

<table>
<thead>
<tr>
<th>Array 1</th>
<th>Array 2</th>
</tr>
</thead>
<tbody>
<tr>
<td>1 Mev - Ch. Alpha</td>
<td>500 Kev - Ch. Alpha</td>
</tr>
<tr>
<td>500 Kev - Ch. Beta</td>
<td>1 Mev - Ch. Beta</td>
</tr>
</tbody>
</table>

\[
\begin{align*}
\sum n_i E_i & = 166556 & 143101 \\
\sum n_i & = 755 \text{ counts} & 666 \text{ counts} \\
\text{Centroid} & = 220.60 \text{ ch.} & 214.86 \text{ ch.}
\end{align*}
\]

From the data given in the table in fig. 57, we calculate the correction to this. The average energy difference between radiation deposited by 1 Mev electrons and by 500 Kev electrons in channel alpha is \(- 11 \pm 4\) Kev (1 Mev electrons depositing lower average energy than 500 Kev electrons) and in channel beta is \(0 \pm 4\) Kev. From the time shifts given in table (c) of fig. 57, an energy
shift from $202 \pm 2$ keV to $310 \pm 2$ keV - a difference of $108 \pm 4$ keV - gives a time shift of $-2.4 \pm 1.0$ channels in channel alpha. Therefore a difference of $11 \pm 4$ keV will give a correction value of $0.2 \pm 0.2$ channels. Fortuitously, in channel beta, the value for the average energy deposited by 1 MeV electrons and 500 keV electrons is the same, therefore no correction, only a quotation of error ($\pm 0.1$ channels) is required. Because the difference in average energies deposited by the 1 MeV and the 500 keV electrons in the thin detectors is small and because the variation of the centroid of the time spectrum is small for a large energy variation (due to the efficient operation of the pulse height compensator), we have considered first order corrections only, i.e. we have assumed an approximately linear relationship between time shift and energy shift.

The following table shows the appropriate correction with its corresponding live centroid measurement from which the value of the life-time is determined.

<table>
<thead>
<tr>
<th>Array 1</th>
<th>Array 2</th>
</tr>
</thead>
<tbody>
<tr>
<td>Centroid (from above)</td>
<td>220.60 ch.</td>
</tr>
<tr>
<td>Correction value</td>
<td>$+0.2 \pm 0.2$ ch.</td>
</tr>
<tr>
<td>Corrected centroid</td>
<td>220.84 ch.</td>
</tr>
<tr>
<td>Statistical error</td>
<td>$E_1 = \pm 2.56$ ch.</td>
</tr>
</tbody>
</table>
But 1 channel represents a time shift of $4.3 \pm 0.4$ ps, therefore,

\[ 2 \tau = 257 \pm 161 \text{ ps} \]

Therefore,

\[ \tau = 128 \pm 80 \text{ ps} \]

The main experimental error in the measurement has been the large statistical error arising from the very low count-rate (approx. 3 counts per hour per mode) which is given by,

\[ \tau = \frac{E \pm 0.4 \text{ ps}}{\sqrt{N}} \]

where $W$ = full width at half maximum (i.e. 3.8 ns) and $N$ = total number of counts. Thus the total statistical error in measuring this life-time is

\[ \pm \sqrt{\frac{E_1^2 + E_2^2}{N}} \]

So the life-time of the first excited state of Pb$^{207}$ as determined from this system is $\tau = 128 \pm 80$ picoseconds. Fig. 64 shows a typical single time peak obtained with the above settings; the F.W.H.M. is measured to be 3.8 ns. On replacing the present source with a source approximately three times as strong, a subsequent measurement by the same method gave a value of,

\[ \tau = 149 \pm 62 \text{ ps} \]
Fast Coincidence Spectrum of 1 Mev electrons and 500 keV electrons from Bi$_{207}$, both depositing 164-373 keV of energy in the thin detectors using pulse height compensation.

F.W.H.M. = 3.8 nS
(c) Problems encountered in setting up the final experimental system

Before obtaining the final results, there were, of course, many abortive attempts - each one uncovering yet another malfunction of the system. Here we list the main difficulties encountered.

Accurate timing of all analogue and logic pulses with respect to one another throughout the system is of paramount importance. It must be remembered that integral discriminators trigger on the leading edge of an input pulse whereas single channel analysers trigger on the falling edge, also that slow pulse amplifiers have a finite propagation time. So there must be a sufficient delay inserted between the output of the fast T.A.C.s and the summing amplifier (Sum) to ensure that the outputs of the two T.A.C.s combine to give a clean composite signal which is acceptable to the multichannel analyser. And, to ensure that all logic operations controlling the input gate of the analyser are performed well in advance of the arrival of the signal from the T.A.C.s, a 2 microsecond delay line was inserted at the analyser input. It was found that if the input pulse and gating pulse were too close in time, then the analysis of the pulse height became unreliable.

Initially, when setting the thick detector pulse height analyser windows, too large an energy range was accepted. This caused logic pulses to be generated for both 1 Mev and 500 KeV electrons in the one analyser, so that the time information from the T.A.C.s was not being directed to its correct sector of the computer memory. The result of this was too small a centroid shift. This error was corrected by carefully calculating the energy range which must be accepted by each single channel analyser in view of the energy range accepted by the thin detectors.
A disconcerting effect which appeared on the time spectrum was a double peak instead of the desired single peak. This was traced to ringing in the trailing edge of the fast pulses being fed to the Ortec 260 time pick-off unit. The double peak was removed by shorting the output lead of the time pick-off transformer (see fig. 55), which goes to the terminated input of the slow pulse amplifier, to earth via a 270 pF capacitor. The capacitor acts as a high pass filter which increases the current of the fast leading edge through the transformer. Any subsequent ringing on the slow trailing edge which might trigger the discriminator is reduced with respect to the height of the leading edge. The capacitor further shorts out any reflections back along the cable from the slow amplifier input terminal.

The recent power cuts (Feb. 1972) have caused the noise level in one of the detectors to increase twofold, so increasing the F.W.H.M. of the time spectra. This deterioration is attributed to the continual cycling from room temperature (+25°C) to the temperature at which the detectors normally operate (−20°C) when the refrigeration unit was without power. There is not much that can be done to remedy this except to replace the detector.

Finally, temperature instability of equipment in the laboratory has been combated by installing an electronically controlled fan heater which maintains the temperature to ±1°C.
Further Modifications and Improvements to the Present System

If the life-time measurement were to be performed between two radiations in a very complex decay scheme such that the superior energy resolving power of semiconductor detectors was essential, then the original (or nearly so) energy spectrum may be reconstructed by the addition of the output pulses from the slow amplifiers of both the thin and thick detectors. To prevent distortion of the energy spectrum due to the difference in solid angles subtended by the two detectors, a linear gate at the output of the mixed signals controlled by the rear thick detector must be incorporated. Then only radiation absorbed by both detectors will be registered by subsequent analyzing systems.

In essence, the limit of short life-time measurement by the present method is superior to the slope method (whose limit is approx. 1/5 th of the F.W.H.M. of the time spectrum). But to justify its use, the accuracy of measurement must also be of the same order or less; and this depends on the care and precision with which the various controls are set. A decrease in the F.W.H.M. of the time spectrum will inherently improve the accuracy of measurement - and this may be done in two ways. Firstly the energy of radiation "creamed off" by the transmission detectors might be increased, which would require the use of thicker detectors, say 300 - 400 microns thick. A factor of 2 (by utilizing energies of 300 - 400 Kev) could be achieved in the time resolution, but the difference in energy spectra deposited by the 1 Kev and 500 Kev electrons of Bi²⁰⁷ would become more noticeable so reducing the original anomy of the two spectra in the thin detectors. Secondly, the noise of the whole system might be reduced
by cooling the detectors, the input P.E.T.s and the input bias resistors of
the fast voltage preamplifiers to liquid nitrogen temperatures (78⁰K). This
would require extensive modifications to the vacuum chamber and cooling system.
The detectors, supplied by Nuclear Enterprises, are only guaranteed down to
-20⁰C - the present temperature; special detectors for liquid nitrogen operation
would be required or, alternatively, an investigation into how quickly the
present detectors may be cooled to liquid nitrogen temperatures with no ill
effects. Apparently it is the rapid cooling of these detectors which cracks
the insulation varnish and so damages the detectors. Since it is the bias
resistor at the input P.E.T. which is the main source of noise with respect to
the fast time resolution[75], cooling to 78⁰K would only reduce the noise by
a factor of 2 approximately (the noise amplitude in a conductor being proportional
to the square root of the absolute temperature)[93].

If the F.W.H.M. of the present time spectra were so reduced to less than
1 ns, then we might reasonably expect life-time measurements in the region of
10⁻¹¹ - 10⁻¹² sec to be within the scope of this system. And such a system
would be suitable for investigating short nuclear life-time in alpha emitters,
e.g. Thorium C, exhibiting fine structure, coincidences being accumulated
between the alpha particles and the conversion electrons from the de-exciting
gamma rays.
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APPENDIX

Computer programs for the NS 627 - PDP/8 on-line pulse height analyser.

(These programs are written in PAL III, being the mnemonic machine code used with the PDP/8 computer)
THREE POINT DISPLAY WITH
SINGLE AND DOUBLE PRECISION
AND SCOPE SWITCHING

/MAIN PROGRAM

SCOPE=132
MNCHAN=134
BASE=135
ONDATA=136
OVRFLO=137
CDF=6201
DXL=6053
DYL=6063
DIX=6054
RLON=6131
RLOF=6132
PAFC=6114
PBSF=6121
PBRR=6122
SSC1=6144
SSC2=6142

SUSPEND=4420 /ALLS OTHER PROGRAMS TO RUN

*7400
STA;DCA OVRFLO
START;CDF
SUSPEND
TAD OVRFLO;SMA CLA;JMP STOP
PBSF;JMP DISPLAY
TAD BASE;DCA MEMLC2
TAD MNCHAN;DCA MFULL2
PBRR;TAD SET;DCA INSTR
TAD 1 INSTR;DCA INSTR
JMP 1 INSTR /GO TO INSTRUCTION SUBROUTINE

SET,+1
ENABL2/0000
ENABLE /1000
DISABL /0100
TEST /1100
ERASE /0010
ANALYZ /1010
PLOT /0110
SCOPEC /1110
TEST, CLA CLL
JMS CONREL: 0012
STA
ERASE, CLL
CDF 10
CLA CML: /DURING ERASE, AC=0 AND LINK CLEAR
/DURING TEST, AC#0 LINK IS COMPLEMENTED
DCA I MEMLC2
TAD I MEMLC2
SZL
TAD STEP2
ISZ MEMLC2
ISZ MFULL2
JMP +7
JMP START

ENABLE2,
STA: DCA PREC2; JMP +3
ENABLE: CLA
DCA PREC2
JMS CONREL: 0014
STA: DCA ONDATA
PAFC: JMP START
PREC2: 0

DISPLAY,
TAD SCOPEF; SMA CLA; JMP START /TEST SCOPE ENABLED
SSC2: CDF 10
TAD (7700; DCA MFULL2
RPT, TAD STEP3/SET X INCREMENT
TAD XC; DXL; DCA XC
TAD I MEMLC3
CLL RTR
DYL; SNA; JMP NXT; DIX
IAC; DYL; DIX; CLA
TAD XC; IAC; DXL; DIX; CLA
NXT; ISZ MEMLC3; ISZ MEMLC3
ISZ MFULL2; JMP RPT
SSC1
ISZ COUNT
JMP START
TAD BASE; DCA MEMLC3
TAD STEP3; CIA; DCA XC/SET -X INCREMENT
TAD SPLIT3; DCA COUNT
ISZ COUNT2; JMP START
ISZ XC; ISZ MEMLC3
STA CLL RAL; DCA COUNT2
CDF: TAD PREC2; SZA CLA; JMP PRECDB
JMP START
CALL ROUTINE DE TAPE

1) SPLIT3, -10
2) STEP3, 2
3) MEMLC3, 0
4) COUNT2, -1
5) COUNT1, -1
6) XC, 0

STEP2, 10
MEMLC2, 0
MFULL2, 0
INSTR, 0

*7200
CONREL, 0
TAD RELAY; RLOF
TAD I CONREL; DCA RELAY
TAD RELAY; RLON
ISZ CONREL
JMP I CONREL
RELAY, 0

PRECD,,
TAD BASE; DCA LOC1
TAD MCHAN; STL RAR; CIA; TAD BASE; DCA LOC2
TAD MCHAN; STL RAR; DCA COU4
CDF 10
93, TAD I LOC1; SNA CLA; JMP 02
STL CLA RAR; T0; TAD I LOC1; ION; DCA I LOC1
ISZ I LOC2; JMP 02
CDF; DCA OVRFLO; DCA ONDATA; JMP START
02, ISZ LOC1; ISZ LOC2
ISZ COU4; JMP 03
CDF; JMP START

LOC1, 0
LOC2, 0
COU4, 0

STOP, JMS CONREL: 0011
STA; DCA OVRFLO
JMP START

SCOPEC; TAD SCOPEF; CMA; DCA SCOPEF
JMP START

DISABLE; DCA ONDATA
JMS CONREL: 0013; JMP START

ANALYZ; TAD RELAY; SZA; RLOF
4477; 4026; 7400 / CALL ROUTINE OFF TAPE
PLOT,TAD BASE;DCA MEMLC4
TAD MNCHAN;DCA MFULL4
SSC2
DXL;DYL
SUSPEND;6123;JMP #2; /WAIT FOR INTERRUPT
SKP CLA
AGN,TAD XC4;DXL;IAC;DCA XC4
CDF 10;TAD 1 MEMLC4;CDF;CLL RTR;DYL
ISZ DEL;JMP #1 /18 MSEC DELAY
ISZ DEL;JMP #1 /18 MSEC DELAY
SUSPEND
ISZ MEMLC4;ISZ MFULL4;JMP AGN
SUSPEND;6123;JMP #2 /WAIT FOR INTERRUPT
SSC1;JMP START

DEL,0
XC4,0
MEMLC4,0
MFULL4,0

$
CDF=6201
BASE=135
MCHNAN=134
*7200
/MEMORY OUTPUT FROM HIGH-SPEED PUNCH
START,CLA CLL
4440/CLAIM PUNCH
DCA CHNUM/SET CHANNEL NUMBER TO ZERO
TAD BASE
DCA MEMLOC/SET STARTING LOCATION
TAD MCHNAN
DCA MFULL/SET TOTAL NUMBER OF CHANNELS
NEWLIN,JMS CRLF
TAD M10/SET PRINT FORMAT
DCA COUNT3
TAD CHNUM
4443/PUNCH OCTAL NUMBER
JMS SPACE
JMS SPACE
NEWNUM,CDF 10/CHANGE TO FIELD 1
TAD I MEMLOC/READ CONTENTS OF CHANNEL CDF
4443/PUNCH OCTAL NUMBER
ISZ CHNUM
ISZ MEMLOC
ISZ MFULL
JMP +2
JMP OUTCOM
ISZ COUNT3
JMP NEWNUM
JMP NEWLIN
OUTCOM,JMS CRLF
4442/FREE PUNCH
4450/STOP UNDER SUPERVISOR
CRLF,0
TAD K215;4441;TAD K212;4441/PRINT CRLF
JMP I CRLF
K215,215
K212,212
SPACE,0
TAD K240;4441/PRINT SPACE
JMP I SPACE
K240,240
CHNUM,0
COUNT3,0
MEMLOC,0
MFULL,0
M10,-10
$
SUBROUTINE TO CALCULATE CENTROID OF SINGLE PEAKED TIME SPECTRUM

*7200
START TAD MNCHAN/FROM 0134
DCA MEMFUL
TAD BASE/FROM 0135
DCA MEMLOC/SET CHANNEL COUNTERS
CDF 10
TAD I MEMLOC
CLL
TAD SUM1
DCA SUM1/SUM ALL COUNTS IN SPECTRUM
SZL
ISZ SUM2
ISZ MEMLOC
ISZ MEMFUL
JMP -8
CDF
SUSPEND
01 TAD MINUS12/CONVERT OCTAL TO DECIMAL
DCA BUFFUL
TAD LASTDIG
DCA BUFLOC/OCTAL NUMBER 3
02 TAD SUM3
M6L
DVI
TENA,0012
DCA RMNDR
MQA
DCA SUM3
TAD SUM2/OCTAL NUMBER2
M6L
TAD RMNDR
DVI
TENB,0012
DCA RMNDR
MQA
DCA SUM2
TAD SUM1/OCTAL NUMBER1
M6L
TAD RMNDR
DVI
TENC,0012
DCA RMNDR
MQA
DCA SUM1
TAD RMNDR
TAD ISOCONST
DCA 1 BUFLOC
CLA CMA
TAD BUFLOC DCA BUFLOC
ISZ BUFFUL
JMP 92
TAD MINUS19
DCA PRINTFUL
TAD STARTBUF
DCA PRINTLOC
TAD I PRINTLOC
4457/TYE SYMBOL
ISZ PRINTLOC
ISZ PRINTFUL
JMP -4
E=TAD MNCHAN
DCA MEMFUL
TAD BASE
DCA MEMLOC
DCA SUM1
DCA SUM2
DCA SUM3/CLEAR NUMBER LOCATIONS
93,CDF 10
ISZ MULT
TAD I MEMLOC
MQL
MUY
MULT>0000/CALCULATE COUNT X CHANNEL
CLL
DCA SUMTEMP
MQA
TAD SUM1
DCA SUM1
SZL
ISZ SUM2
SKP
ISZ SUM3
CLL
TAD SUM2
TAD SUMTEMP
DCA SUM2
SZL
ISZ SUM3
ISZ MEMLOC
GDF
SUSPEND
JMP Q4
SUMTEMP, 0
PRINTLOC, 0
PRINTFUL, 0
STARTBUF, 7343
MINUS19, 7755
ISOCONST, 0060
RMNDR, 0
BUF1, 0304/LETTER D
BUF2, 0305/LETTER E
BUF3, 0316/LETTER N
BUF4, 0256/FULL STOP
BUF5, 0275/ = SIGN
BUF6, 0/LOCATION FOR NUMBER TO BE STORED
0000
0000
0000
0000
0000
0000
0000
0000
0000
0000
0000
0000
0000
0000
0000
0000
0000
0000
0000
0000
0000
B1UF18, 0215/CARRET
BUF19, 0212/LINE FEED
BUFLOC, 0
BUFFUL, 0
LASTDIG, 7363
MINUS12, 7764
SUM1, 0
SUM2, 0
SUM3, 0
MEMLOC, 0
MEMFUL, 0
Q 4, ISZ MEMFUL
JMP I 003
TAD LETTRN
DCA BUFF1
TAD LETTRU
DCA BUFF3
TAD LETTRM
DCA BUFF3/TYPE OUT ...NUM
TAD STOPCODE
DCA I END
TAD STOPCODE
DCA I BEGIN
JMP I 001
001, 01
BEGIN, START
END, E
STOPCODE, 4450
LETTRN, 0316
LETTRU, 0325
LETTRM, 0315
BUFF1, 0
BUFF2, 0
BUFF3, 0
003, 03