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INTERNATIONAL CENTRE FOR THEORETICAL PHYSICS
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ICFA INTERNATIONAL COMMITTEE FOR FUTURE ACCELERATORS

ISTITUTO NAZIONALE DI FISICA NUCLEARE INFN

H4.SMR. 394/6

SECOND ICFA SCHOOL ON INSTRUMENTATION IN ELEMENTARY PARTICLE PHYSICS

12- 23 JUNE 1989

# SILICON DETECTOR AND SIGNAL PROCESSING EXPERIMENT

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# Silicon Detector and Signal Processing Experiment

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August 1988

## Introduction

Before discussing the details of the experiment a brief summary of the relevant features of silicon detectors and signal processing for minimising noise is given. I hope that this will provide a review of the important results and link together the experiment with the more detailed explanations given in the lectures. If the lectures have reached the topics covered here it will be necessary only to read quickly through the preliminaries; otherwise more attention will be required.

## The diode detector

Modern silicon detectors are mostly based around the p-n junction diode, and the detectors to be used for this experiment have been fabricated in that form. We outline below the simple model which is usually used to describe the p-n diode, and which is adequate for a description of the main features of a particle detector.

Silicon particle detectors are always fabricated as *diodes*. This may initially seem surprising since the energy loss of a charged particle or photon in silicon is independent of the state of the detector material and it would be simple to apply an electric field to a piece of detector material sufficient to sweep ionisation electrons across it. The reason of course is that silicon is a semiconductor and an electric field applied across a silicon layer would generate a sizeable direct current which would have undesirable consequences. Apart from detector heating the flow of current would lead to *noise*, large enough to obscure the small signals we expect to observe, from fluctuations in the number of charge carriers (shot noise).

### Detector material

Detector grade silicon comes in the form of high purity, and thus high resistivity, crystals. The crystals are doped with small amounts of impurities which lead to extra, loosely bound, *electrons* in *n-type* material (dopants As,P,..) or *holes* in *p-type* material (dopants Bo,Al,..); either p-type or n-type silicon can be used, although n-type is perhaps more common. In crystalline materials the individual atomic energy levels merge into bands, and the outermost electrons are contained in two bands, called the valence and conduction bands, with an energy gap between them where no available energy levels exist. Electrons in the conduction band are mobile and can move freely though the crystal under the influcence of an electric field while those in the valence band are fixed. The opposite is true for holes. In a diode detector the conduction band is essentially empty of electrons and the valence band of holes - therefore only when electrons and holes enter the bands via ionisation does any current flow.

This is usually fabricated by implanting a thin layer of p-type atoms into the surface of an n-type wafer. The density of p-type doping centres in the surface layers is many times higher than the density of n-type centres in the bulk crystal.

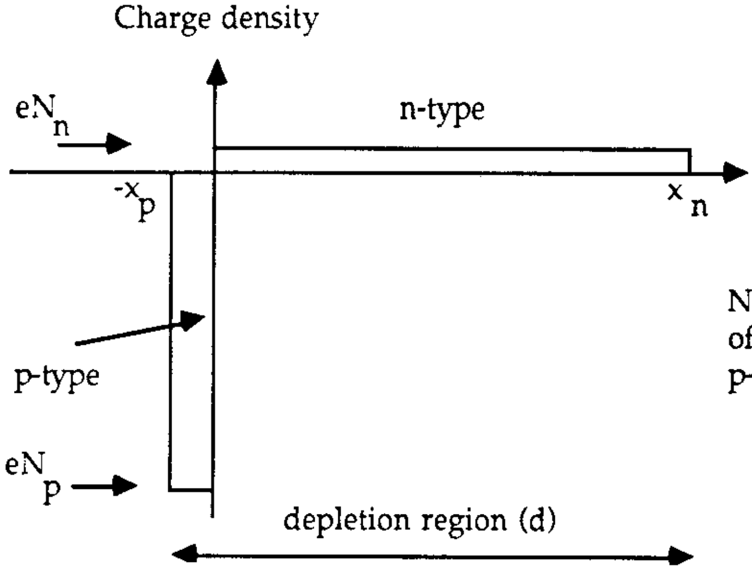
We can understand the properties of the junction best by considering the unreal situation where the n- and p-type layers are originally separated and then brought into contact. Although both pieces of material are electrically neutral the p-type side of the junction will contain a high density of mobile holes whereas the n-type side will have a high density of mobile electrons. Thermal equilibrium will rapidly be established during which time the the mobile electrons and holes will flow across the junction boundary and combine. [Vacant energy levels in the valence band of the p-type material (holes) will attract electrons at higher energy levels from the conduction band in the n-type material]. As this process continues a *space charge* in the region of the junction will build up, caused by the exposure of the fixed ionic charges of the doping atoms. The space charge in the junction leads to a (reverse bias) electric field which eventually will be large enough to prevent any further flow of charges.

The region of space charge will thus be almost devoid of mobile charge carriers, and is known as the *depletion region*. It can be enlarged by the application of further reverse voltage to the junction. It will extend further into the less heavily doped n-type material than into the p-type region, and its electrical properties can be described well by a simple one-dimensional model.

**Table 1 - Summary of silicon properties**

| | |
|---|---|
| Intrinsic doping concentration ($n_i$) | $1.45\ 10^{10}\ cm^{-3}$ |
| Band gap (Eg) | 1.1 eV |
| Density of electrons in conduction band ($n_i$) at temperature T | $\alpha\ e^{-Eg/2kT}$ |
| Energy required to create an electron-hole pair | 3.6 eV |
| Electron mobility ($\mu_e$) | ~ 1500 $cm^2$/volt.s |
| Hole mobility ($\mu_h$) | ~ 450 $cm^2$/volt.s |
| Dielectric constant ($\varepsilon$) | 1.05 pF/cm |
| Most probable energy loss of minimum ionising particle | ~ 26 keV/ 100 $\mu$m<br>~ 7200 electron-hole pairs |

The space charge distribution, $\rho(x)$,in the neighbourhood of the junction is shown below

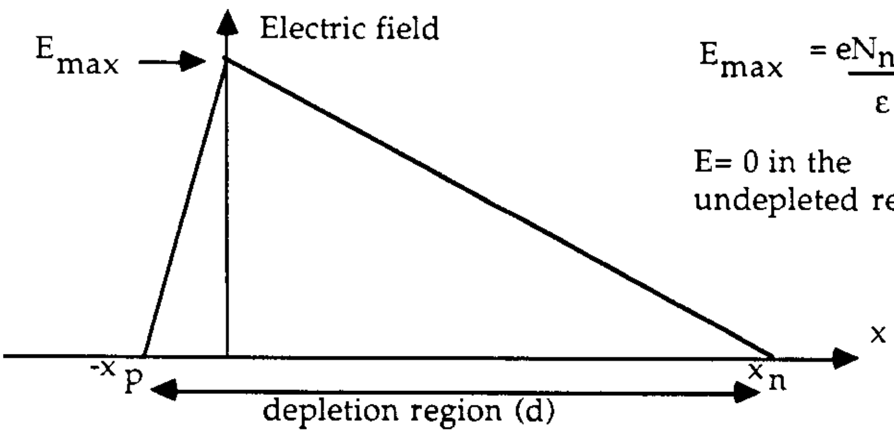



$N_n$ and $N_p$ are the densities of doping atoms in the n and p-type regions.

Overall charge neutrality requires $N_p x_p = N_n x_n$.
The electric field and potential are calculated from Poisson's equation

$$d^2\phi/dx^2 = -\rho(x)/\varepsilon$$

leading to a field as shown




$$E_{max} = \frac{eN_n d}{\varepsilon}$$

E= 0 in the undepleted regions

The potential difference across the junction can be obtained by inspection (the area under the triangle) to be :

$$V + \phi_{bi} = \Delta\phi = E_{max}\, d/2 = eN_n d^2/2\varepsilon$$

where V is the applied voltage and $\phi_{bi}$ the built in diode voltage (typically a fraction of a volt). Note that the thickness of the depleted region varies with the square root of the applied voltage and inversely with the square root of the doping concentration, which is an important reason for choosing high resistivity material to manufacture detectors.

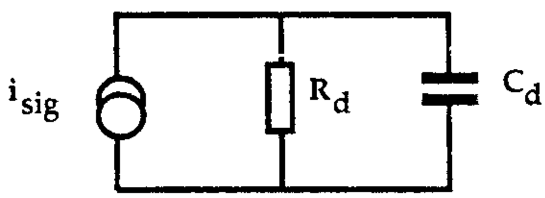
In addition the junction behaves like a parallel plate capacitor with a thickness the size of the depletion region so, for area A,

$$C = \varepsilon A/d \quad \sim V^{-1/2}$$

The device is a slightly leaky capacitor and will usually have small dark current flowing across it (perhaps a few nA). This leads to the equivalent circuit for the detector.

## Equivalent circuit of detector

The detector can be viewed to a very good approximation as a source of current (the signal) in parallel with a capacitance and a resistor. Thus :



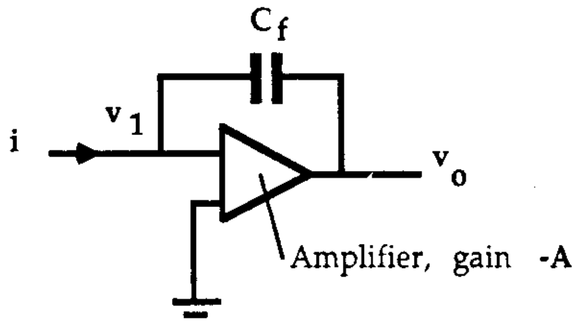

Since the d.c. leakage current is small $R_d$ is very large and is generally ignored. The detector capacitance is that of the depleted p-n junction and therefore varies with the applied bias. Part A of the experiment will verify this picture of the diode detector.

Amplifier and Signal Processing

The amplifying chain used for this experiment is of the type conventionally used with silicon detectors - it is a *charge sensitive preamplifier* followed by a *pulse shaping filter amplifier.*

**Charge sensitive preamplifier**

The preamplifier can be viewed as a high gain, inverting amplifier (see Appendix B) with a capacitor in the feedback loop



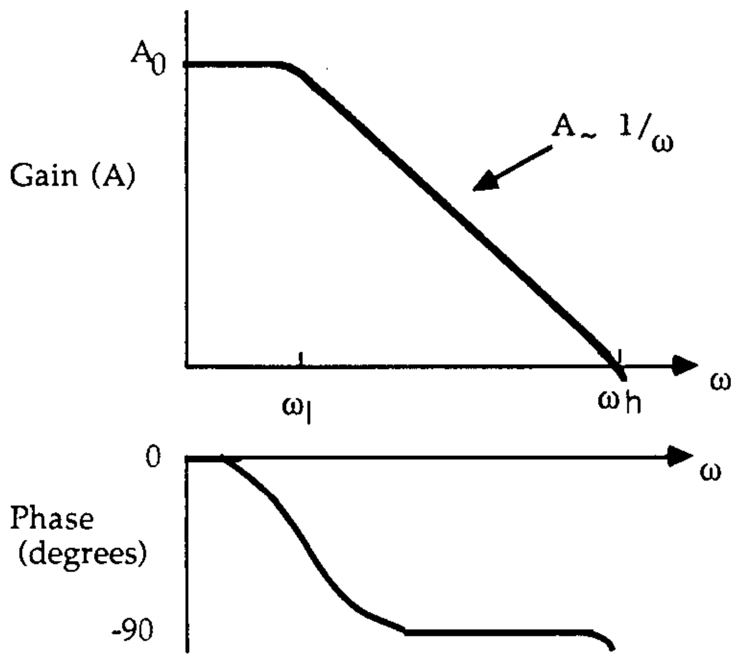

It has a very high input impedance without the feedback loop connected and therefore draws almost no current. The circuit equations are

$$v_0 = -Av_1 \qquad \text{and} \qquad v_1 - v_0 = Q/C_f$$

where Q is the charge deposited by the signal on the feedback capacitor. Therefore

$$v_0 = -AQ/(A+1)C_f \approx -Q/C_f \qquad \text{for } A >> 1$$

Generally therefore $C_f$ is small (1pF is typical) for maximum sensitivity. In reality the gain, A, of the amplifier is a function of frequency. Typical behaviour is shown in the curve below, plotted on a log-log scale:



Two points are of special interest on this plot, the unity gain frequency $\omega_h$ and the frequency $\omega_l$ where the gain decline begins. The amplifier "d.c gain" is

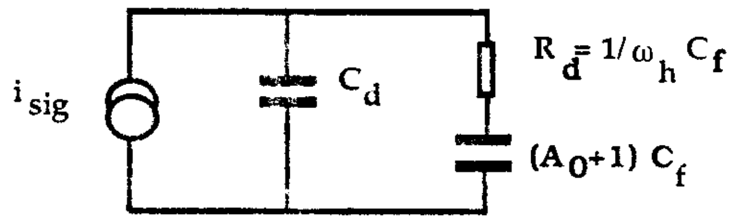
$$A_0 = \omega_h/\omega_l$$

The phase shift between the (inverted) output voltage ($-v_{out}$) and $v_{in}$ is zero at frequencies where the gain is frequency independent, and -90 degrees where the gain decreases inversely with frequency. Thus the charge sensitive amplifier appears to a signal like a large capacitor in series with a resistor, which can be seen by considering the behaviour of

$$v_1 = i/(A+1)\omega C_f$$

at high and low frequencies. The capacitance arises from the frequency independent gain, and the resistance from the frequency dependent gain in conjunction with the feedback capacitance $C_f$.

We can now draw the equivalent circuit of the detector and the preamplifier.




If $(A_0+1)C_f >> C_d$ most of the signal is <u>eventually</u> transferred into the amplifier,i.e. onto the feedback capacitance, and the output voltage becomes

$$v_0 \approx -Q/C_f \qquad \text{for } A_0C_f >> C_d$$

The <u>rise time</u> of the output voltage ( i.e. the time required for the charge transfer) is determined by the time constant

$$\tau_r = C_dR_d = C_d/\omega_hC_f$$

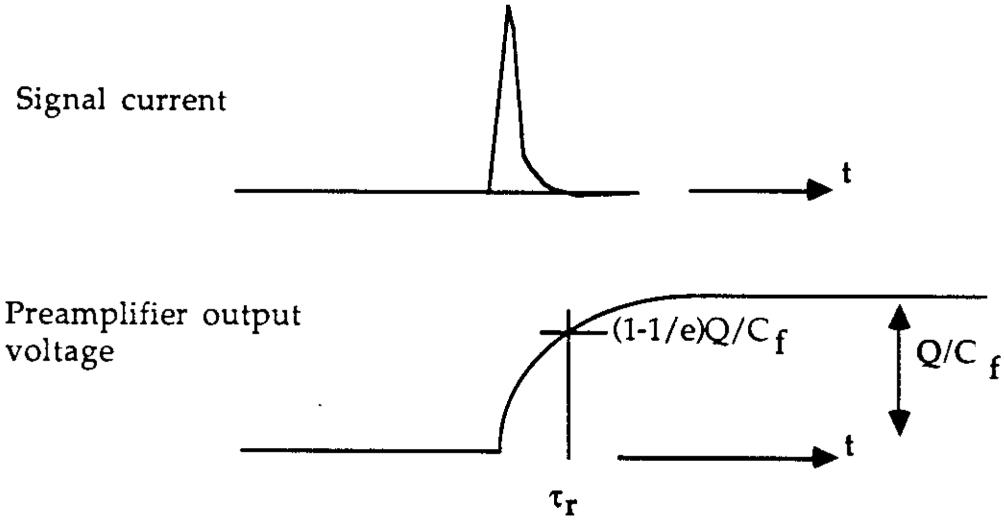
For a fast signal we regard the current pulse as an impulse

$$i_{sig}(t) = Q\,\delta(t)$$

which can be shown to contain an equal mixture of all frequencies, i.e. the signal is *white*. The preamplifier output voltage is then,

$$v_o(t) = -(Q/C_f)\,(1-\exp(-t/\tau_r))$$

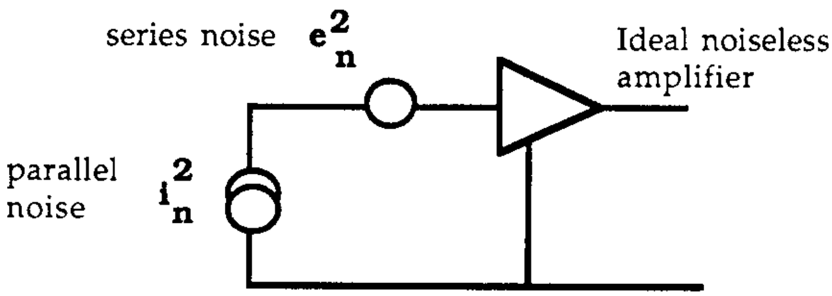
and some revealing observations about the detector and the amplifier can be made by measuring $\tau_r$ as a function of $C_d$ or $C_f$.




In practice the feedback loop of the preamplifier usually also includes a large resistor in parallel with the capacitor so that the preamplifier output pulse actually decays exponentially from its peak value with a long time constant.

**Noise and Amplification**

The problem of noise in an amplifier system is greatly simplified by considering an ideal, noiseless amplifier with noise sources placed at its input. In general it is helpful to consider the noise originating in two separate sources, one in parallel with the amplifier input, the other in series.




$i^2$ and $e^2$ represent mean squared values of current and voltage fluctuations per unit frequency interval. We always use squared quantities, $i_n^2$ or $e_n^2$, to perform analysis of noise amplification essentially because we are concerned with the magnitude of fluctuations about an average level.

The *parallel noise* originates from shot noise in leakage currents (I) and thermal noise in resistances (R) in parallel with the input. Thus

$$i_n^2 = (2eI + 4kT/R)$$

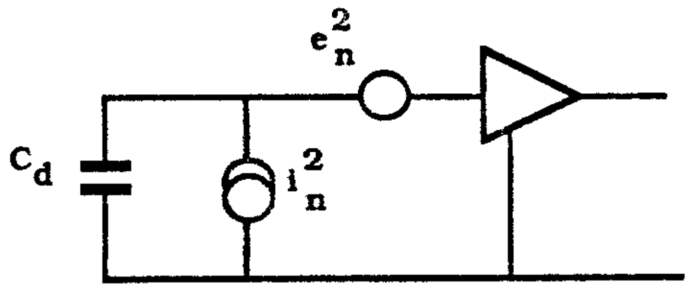
Usually R is chosen to be large enough that leakage current shot noise dominates.

The *series noise* actually originates in thermal fluctuations of current flowing in the conducting channel of the Field Effect Transistor (FET) at the first stage of the preamplifier but appears like a voltage source when placed at the amplifier input.

$$e_n^2 = 4kT\,(2/3g_m)$$

where $g_m$ is a quantity known as the FET transconductance which has dimensions of 1/resistance and typically might have a value of ~ 1/200Ω.

Both of these noise sources are *white* - however when a detector is connected the total noise at the input can be affected by the presence of the detector capacitance.




One can then show that the total effective noise at the amplifier input is (expressed as current noise)

$$I_n^2 = i_n^2 + e_n^2\omega^2 C_d^2$$

which is *no longer white* because of the fact that $C_d$ represents a frequency dependent impedance.

**Pulse shaping.**

We can now see that some kind of pulse shaping is required for two main reasons:

(a) the output from the charge sensitive preamplifier is not a very practical shape for handling multiple pulses, and

(b) the system is not optimised for best signal to noise performance. The fact that the signal is white and the noise has a strong frequency dependence suggests that we can improve things by selecting a band of frequencies to transmit using a pulse shaping, or filter, amplifier.

The pulse shaping employed in the amplifier used in this experiment is not the one which produces the very best signal to noise ratio but one which can be realised practically in quite a simple way. The circuit looks like two well known filters, a high pass and a low pass, in series following the preamplifier. The overall effect is to produce a pulse shape which can be written as

$$f(t) = \alpha Q\,(t/\tau)\,e^{-t/\tau}$$

Q being the input signal charge
α a constant

where τ is the time constant of both filters.

The noise performance of the system is characterised by a quantity referred to as the Equivalent Noise Charge (ENC) which is defined as the signal which would generate an output voltage of the same magnitude as the r.m.s. noise in the system. In practice one would normally require a signal several times larger than the ENC for observation since when viewing on an oscilloscope a signal must be ~3x $\sigma_{noise}$ for a triggered scope, or ~5 x $\sigma_{noise}$ for an untriggered scope for the signal to be readily visible.

The ENC can be written

$$\sigma^2_{noise} = ENC^2 = \mathbf{a}\, i_n^2 \tau + \mathbf{b}\, e_n^2 C_d^2/\tau + \mathbf{c}$$

which is a general result for any filter characterised by time constant $\tau$, though **a**, **b** and **c** are different for each filter. For the filters used in this experiment **a** = **b** $\approx$ 1.0

From the values of $i_n$ and $e_n$ given above we can see

$$\sigma^2_{noise} = \mathbf{a}\, 2eI\tau + \mathbf{b}.\, 4kT\,(2/3g_m).\, C_d^2/\tau + \mathbf{c}$$

and we can predict that the observed noise will depend on I, $C_d$ and $\tau$. In addition it can be seen that for a given value of I and $C_d$ there will be an optimum choice of $\tau$. We are going to attempt to observe this effect in Part B of the experiment.

The detector used for the experiment is a silicon photodiode of area 1 $cm^2$. It is constructed as shown below in a (not to scale) section

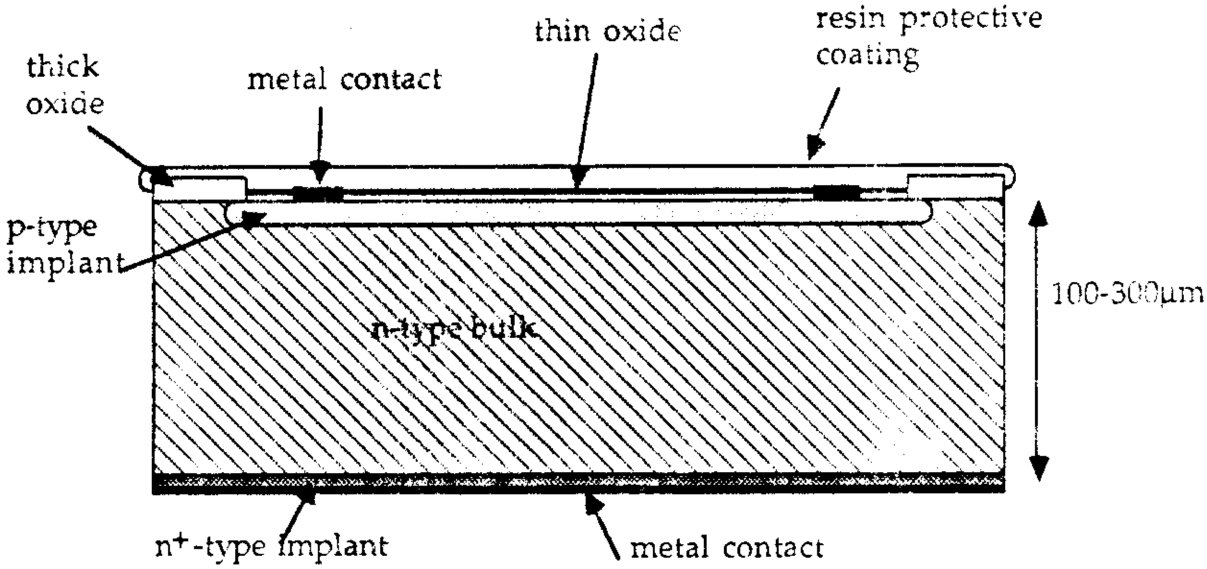



**Table 2 - Summary of detector properties**

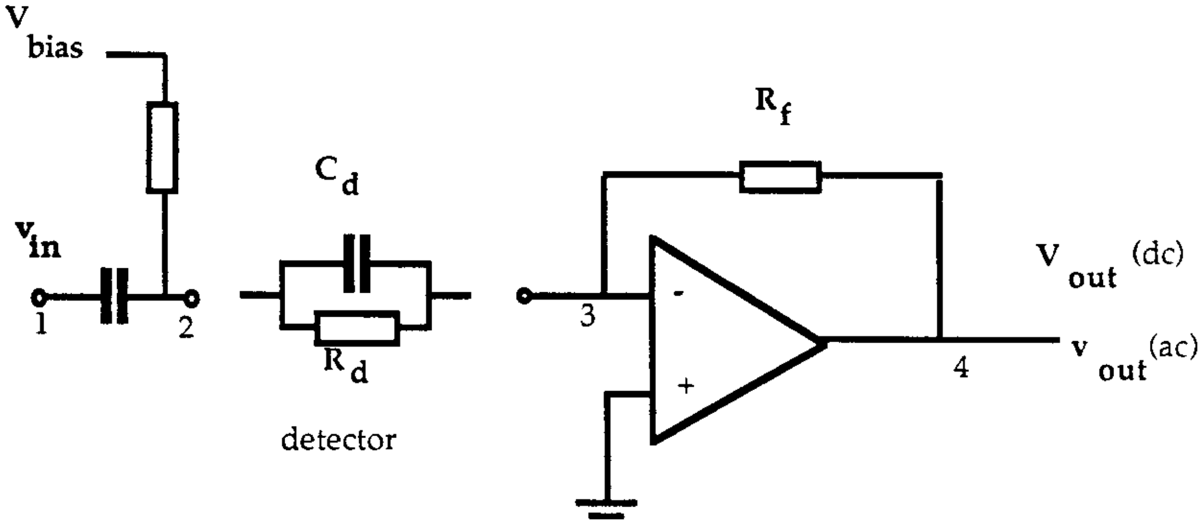
| | |
|---|---|
| Resistivity of bulk n-type material with doping concentration $N_n$ | $\Omega = 1/e\mu_e N_n$ |
| Built in potential | $\phi_{bi} = (kT/e)\ln(N_p/n_i) + (kT/e)\ln(N_n/n_i)$ |
| | [ $kT/e = 0.026$ volts at $300^oK$] |
| Depletion depth at applied voltage V | $d = \sqrt{2\varepsilon(V+\phi_{bi})/eN_n}$ |
| | $= \sqrt{2\varepsilon\Omega\mu\,(V+\phi_{bi})}$ |
| Maximum field | $E_{max} \approx eN_n d/\varepsilon = 2V/d$ |
| Junction capacitance per unit area | $C = \varepsilon/d$ |
| Carrier velocity in field E | $v = \mu E$ |

PART A - Measurement of detector characteristics

When operating a new silicon detector it is normal to make observations of its leakage current and capacitance at several voltages over the operating range. This is for two reasons:

- to establish the voltage required to deplete the detector to a suitable depth,
- to determine the two parameters, I and C, which determine the noise performance of the system so that the best choice of amplifier can be made.

We shall measure the leakage current ($I_{in}$) and capacitance ($C_d$) with the aid of the simple circuit sketched below




The circuit forms an amplifier with a gain $G = -R_f/Z_d$ where $Z_d$ is the impedance of $C_d$ and $R_d$ in parallel (Appendix B).

Thus for a sinusoidally varying input voltage ($v_{in}$) of frequency f

$$|G| = |v_{out}/v_{in}| = 2\pi f\, C_d\, R_f \qquad \text{if } R_d \text{ is large}$$

and for the d.c. voltage at the output $|V_{out}| = I_{in}R_f$

The circuit is already prepared with the oscillator included. All that is necessary is to record $v_{in}$, $v_{out}$, and $V_{out}$ for values of $V_{bias}$.

The component values are set to $R_f = 1\ M\Omega$ and $f = 1kHz$. $v_{in}$ is adjustable but should be small compared to $V_{bias}$ so $v_{in} = 0.1$ volt would be a suitable value, and also makes G easy to calculate.

The diode is already mounted in a metal box to screen it from light and to provide electrical shielding. It should be possible simply to connect it to the measuring device and make the required observations. Using the graphs attached to this script it should be possible to make the some deductions from the measurements you make.

Measurement programme

*To avoid making lots of calculations some figures have been prepared and attached to the end of this manuscript - refer to them as you proceed with the measurements below.*

**1)** With $v_{in} \approx 0.1$volt measure $v_{out}$ and $V_{out}$ for values of $V_{bias}$ from ~0 volts to ~50 volts. Calculate C (Fig.1) and $I_{dark}$ and plot the results on the log-log graph paper provided (remember to make a subtraction of the stray capacitance due to the cables used to connect the diode). Without making too many measurements try and obtain good coverage of the entire range, with several observations below about 2 volts. You will then be able to see from your graphs the behaviour of C versus V, and I versus V.

2) The plot of most interest is the C-V plot. First verify that the capacitance-voltage behaviour is as expected and then answer the questions below:

| | |
|---|---|
| What is the depletion voltage of the detector? | (Fig.2) |
| What is the thickness of the photodiode? | (Fig.2) |
| What is the resistivity of the n-type silicon used to to make the diode? | (Fig.3) |
| What is the doping density ($N_D$)? | (Fig.4) |

3) Using the value of $N_D$ estimate the part of the built-in junction voltage due to the n-type component of the junction (Fig.5). Try to make an estimate of the total built-in junction voltage from the measured capacitance at low bias. If you do this carefully it should be possible to calculate the doping density on the p-type side of the junction and also the thickness of the p-type region. How do these compare with the approximations used for the p-n junction model discussed above?

4) The average electric field in the diode at full depletion can also be calculated and from this the velocity of the charge carriers. Thence an estimate of the charge collection time can be made.

5) From the discussions above you are now also in a position to determine the likely noise of the system you will use for the second part of the experiment. First, from the observations of I and C you have made, you should be able to identify which of the two contributions to the noise is likely to be dominant (values of $\tau$ from 0.5µsec to 10µsec will be used). You can also make an estimate of the optimum shaping time constant and the absolute value of the ENC. Calculate ENC in units of electrons (in round figures making any necessary approximations).

Use $\quad ENC^2 \approx 4kTR_sC^2/\tau + 2eI\tau \quad$ coulombs$^2$,

or $\quad ENC^2 \approx (4kT/e)(R_sC^2/e\tau) + 2I\tau/e \quad$ electrons$^2$.

Assume $R_s \sim 160\Omega$ ; the electronic charge $e = 1.6\ 10^{-19}$ coulombs and $(kT/e) \approx 0.025$ volts at room temperature.

In the second part of the experiment the photodiode is to be used as a particle detector, and to relate the observations made in part A to the operation of the detector-amplifier combination. We shall start with some qualitative observations before attempting to make some more quantitative measurements.

The detector is connected to the bias power supply via the preamplifier. Bias voltages up to 50-60 volts should be adequate. The preamplifier output signal can be directed, using the multiposition switch on top of it, to one of five inputs to the shaping amplifier. Each of the five amplifier channels shapes the pulse using an RC-CR (differentiating-integrating) filter with a different time constant ($\tau \sim 0.5, 1, 2, 5$ and $10\mu$sec). Ideally this gives rise to a unipolar output pulse peaking at $t = \tau$.

**1)** First examine the pulse shape produced by the filter amplifier by injecting a test pulse into the preamplifier; the appropriate input is indicated. The diode should not be connected for this observation. The test pulse is a square voltage step which places a small charge on a test capacitor. The sensitivity of the system can be calibrated using this signal; if V is the amplitude of the voltage step and $C_t$ is the test capacitance then the effective charge at the preamplifier input is $Q_{test} = C_t V$ (see Appendix A). Some useful equivalences here are

1mV x 1pF = 1 femtocoulomb = 6250 electrons = 22.5 keV Si.

[*Note*: in actual practice such a calibration of the system should be done with the detector present under bias. Can you see why this should be so? ]

**2)** Select the amplifier channel so that the shortest time constant is used and disconnect the test pulse from the preamplifier. Put the scope on AUTO trigger and look at the trace showing the noise in the system. You will probably need the scope on its most sensitive voltage range. The traces you see are the noise inherent in the system due to the amplifier only and since $C_{in} \sim 0$ this should be quite small. It is hard to judge the true r.m.s. value from this observation since it depends on the intensity of the scope and other factors.

**3)** Now connect the detector but leave the bias voltage at around 0 volts. The noise trace on the scope should have increased considerably due to the effect of the large capacitance at the preamplifier input. You may like to vary the time base of the scope to observe the noise over different intervals. Slowly increase the bias, and continue to observe the display - the effect of increasing the depletion volume should be quite evident. Try making the same observations using a different channel of the amplifier and compare them with your expectations from the predictions above. We shall try and make the observations quantitative but before doing that let us examine the response of the detector to particles.

**4)** The particles which are to be observed are $\beta$-electrons from a strontium radioactive source. The most energetic electrons will have an energy of ~2.2 MeV and are easily able to penetrate the silicon layer. However many of the electrons will have much less energy than this and some will stop in the silicon depositing all their kinetic energy (for example the range of a 200 keV electron is ~180$\mu$m). This means that it is hard to make a precise calculation using an electron signal as a calibration since there is no unique energy deposition.

Nevertheless you should be able to see clear effects as you increase the bias voltage on the detector. Set the scope to trigger on positive going pulses. Again start from zero bias voltage and examine the signal pulses using different time constants. In this case too the effect of increasing the size of the depletion region on both the signal and the noise should be evident. Can you draw any conclusions about the best amplifier time constant to use for best signal/noise discrimination? Does this change as the bias is increased?

5) We can now try to make the measurement of noise in the system more precise. For really accurate estimation we could measure the voltage at the output with a true r.m.s. voltmeter, but a simpler method which will be adequate for our purpose is described.

First remove the source and connect the amplifier output to both channels of the oscilloscope using a T-connector at both inputs. Terminate the second input in 50Ω. Set the scope to sweep alternately between the two channels using either ALT or CHOP settings. Two noise traces will be visible, separated by a dark band. Select the same voltage scale on both channels - sensitive enough to make the signals occupy about one vertical division. Now adjust the vertical position of one of the traces so that the separating dark band just disappears. Now remove the signal. The separation between the two traces should represent twice the r.m.s. noise level ($2\sigma$) which is a consequence of superposing two gaussian distributions.

Try and repeat the adjustment a couple of times to check that it is repeatable. It may take a bit of practice to master, but it should be possible to reproduce your estimates with about 10% accuracy.

6) When you are satisfied with your measurements estimate $\sigma_{noise}$ for several values of bias voltage on the diode. Repeat the observations for two or three time constants of the filter amplifier and plot the results as r.m.s noise versus bias voltage. Compare them with your expectations from the discussion of amplifier noise above. Try and use the test pulse calibration to convert the measurements of r.m.s voltage into equivalent noise charge. It is also interesting to examine the variation of $\sigma_{noise}$ with time constant, or peaking time of the amplifier.

Another method of estimating the r.m.s. noise is to use a pulse height analyser which produces a digital output proportional to the maximum voltage level at the input within a preselected time interval, or gate. We shall do this for both β electron signals and noise, but since it is only practical to provide one pulse height analyser this will be done in collaboration with the demonstrators. It should be possible to provide even clearer evidence of the response of the diode at different operating voltages.

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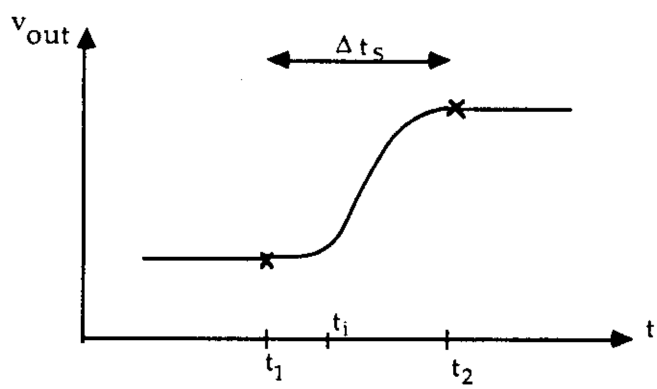
In the discussion above it was implied that pulse shaping is essential to achieve optimum measurements of signals in the presence of noise. Actually another method for signal measurement exist which can produce similar noise reduction by relying not on filtering the preamplifier signals with a pulse shaper but on *sampling* the preamplifier output voltage at selected times. This is a method implemented in many modern VLSI amplifiers as it is particularly appropriate using technology where transistors and capacitors are easy to build accurately and reproducibly but resistors are much harder to construct. This is true in some important integrated circuit fabrication techniques such as CMOS. The sampling method is also ideally suited to colliding beam experiments where the exact time of a possible interaction is precisely defined by the crossing of opposing bunches of particles in the accelerator.

The essence of the method, known as correlated double sampling, is easy to understand and is illustrated below:




The output from a charge sensitive preamplifier is given by $v_{out}$ and the expected arrival time of an interaction time is accurately known to be $t_i$.

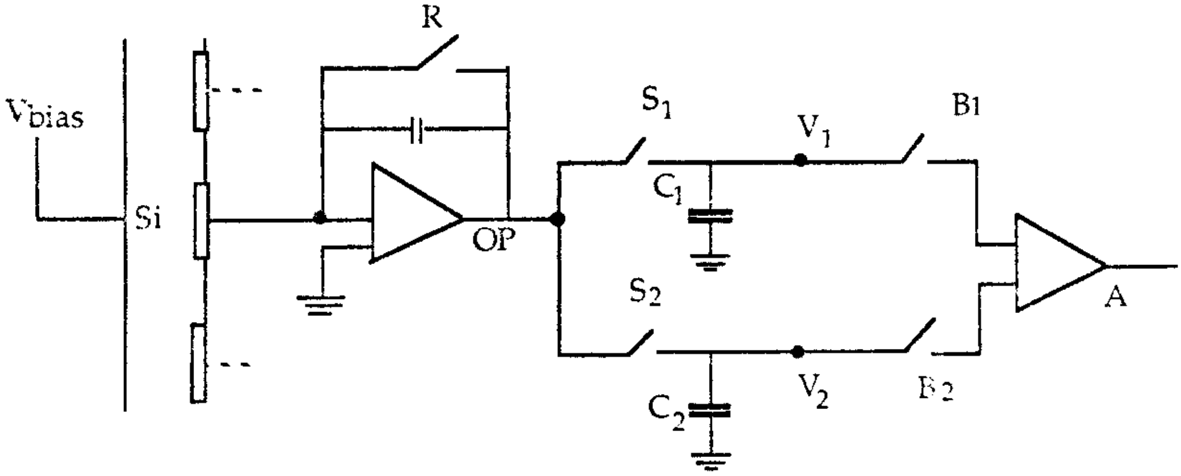
Therefore we simply sample the output voltage level before and after $t_i$ and measure the difference to determine whether or not a signal was detected. Our next question will be to define precisely when to take the samples to optimise the measurement. We should expect that, having done no pulse shaping, noise from the preamplifier will be visible and a measurement of a voltage difference between $t_1$ and $t_2$ will not alone be significant.

Intuitively we expect fluctuations due to shot noise in detector leakage currents to influence the observed voltage difference, particularly if the interval between samples is long. We also know that the rise time of the preamplifier is not zero and, in fact, is controlled by the details of its design. Therefore $\Delta t_s$ cannot be made too short.

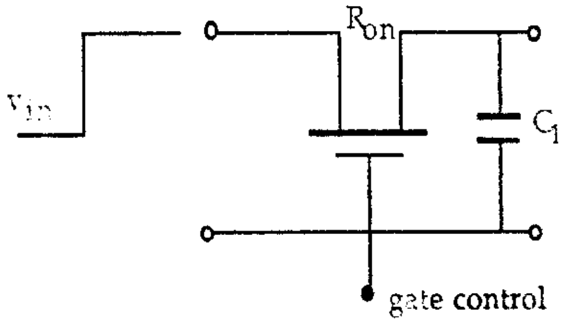
A more detailed analysis reveals that thermal noise in the input FET is limited only by the bandwidth of the preamplifier - ie related to rise time. Shot noise can be easily calculated from the leakage current using Poisson statistics. However 1/f noise is not independent of sampling time and can be somewhat reduced by increasing the sampling interval. A simple formula is not easy to write down to express all these facts.

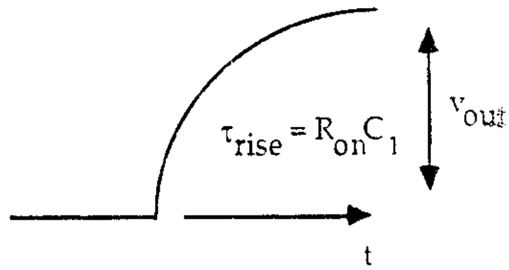
As an extension to the experiment we hope to demonstrate the use of a modern integrated circuit amplifier to read out a microstrip detector. The version we will use is a CMOS Microplex chip designed by Rutherford Appleton Lab. As the experiment will be done in collaboration with the demonstrators we give below an outline, leaving the details for practical demonstration.

The basic elements of each unit of the system are shown below:

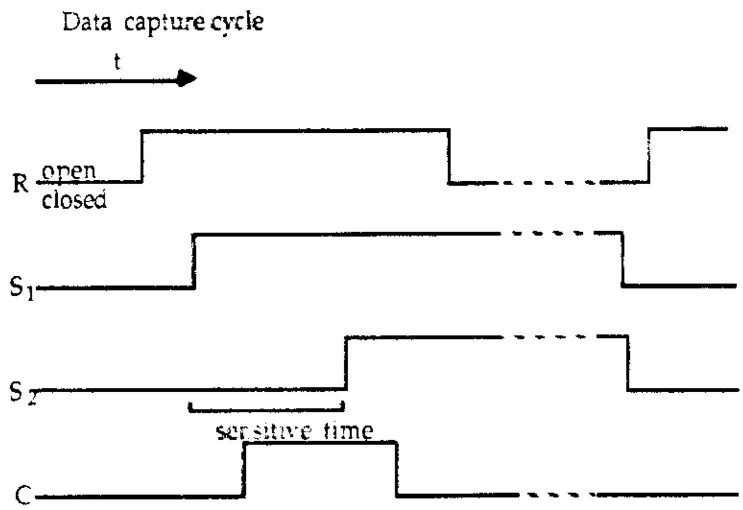



In addition to the normal charge-sensitive preamplifier there are a number of important switches. These are actually FETs (Field Effect Transistors) which can be switched from a conducting to a non-conducting state by application of a voltage to the gate.




Switch R acts to reset the system by discharging the preamplifier feedback capacitor; it is opened a few microseconds before an interaction is expected. Since it is located at a sensitive point in the circuit and its opening injects some charge into the system input it is necessary to allow sufficient time to permit switching transients to die away before any sampling takes place. The switch sequence is sketched below.



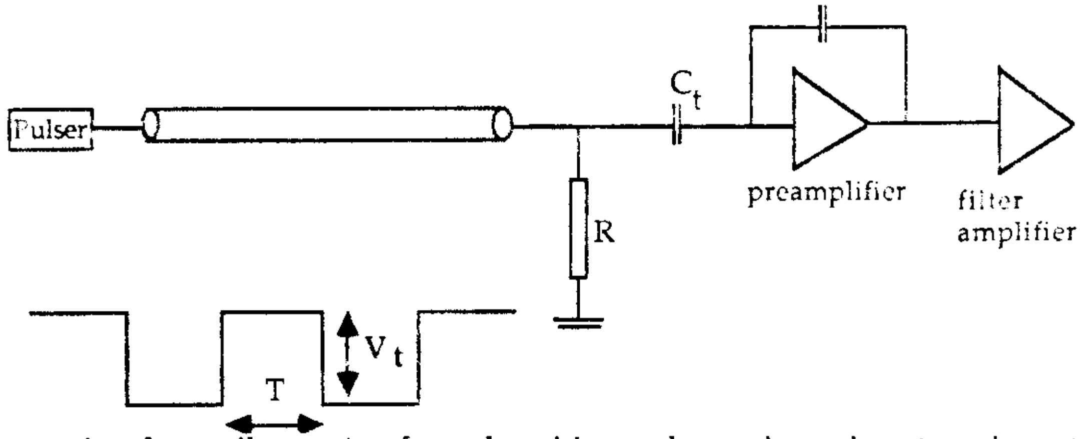
The switches S1,S2 and storage capacitors C1 and C2 perform the correlated double sampling. S1 and S2 were originally closed but once transients from R have disappeared S1 is opened, storing the voltage level at OP on C1 prior to the interaction. Several microseconds later S2 is also opened, now storing the level at OP on C2 *after* an interaction should have occurred. To measure the signal observed it is now necessary to measure the difference between voltages V1 and V2. This is achieved by simultaneously closing switches B1 and B2. Voltages V1 and V2 are then applied to an off -chip differential amplifier; there is one of these for all 128 channels on the microplex and the switches B1 and B2 are controlled by a shift register which steps sequentially from channel to channel at a fixed time interval.

All the functions are controlled by a driver box manufactured as a NIM module with front panel adjustment of all the relative timings. Calibrate signals, C, can be applied to all the 128 channels in groups of 4 channels (ie 1,5,9..../2,6,10,..../etc) which charge an on-chip test capacitor on every channel.

In the laboratory session we hope to see the efforts of varying relative timings; the most interesting are the sampling interval and timing within the gate of the signal.

We also expect to observe the detection of β electrons in the silicon microstrip though this is much harder than with a pulse shaping amplifier because of the small sensitive area of each strip and the random arrival of signals. With an LED pulser we can probe the detector in a more predictable way using the sensitivity of the silicon to infrared light since the transparent oxide gaps between strips will allow light to enter the detector.

There are a number of ways in which to calibrate the energy response of a detector-amplifier system. One of the most useful is to apply a voltage step of a known amplitude to a capacitor of defined value at the input of the amplifier. This is usually done by means of an arrangement like that sketched below:




The input signal, usually a train of equal positive and negative going steps, is sent from a pulser via a cable terminated in its characteristic impedance, R. Even if the cable is too short to be considered as a transmission line it can be detached and terminated in the same resistance at a scope (i.e. 50Ω here) for proper observation of the signal.

On arrival of a positive voltage step at the cable end the test capacitor is charged by a very rapid flow of current which simulates the signal response of a detector. The amount of charge placed on the capacitor is simply

$$Q_t = C_t.V_t$$

Since the preamplifier input is a virtual ground point (see Appendix B) and the decay time constant of the feedback is very long this charge remains on the capacitor until it is removed by a negative voltage step. At this time (T) the capacitor is discharged by an equal, opposite rapid flow of current. The system response to this series of positive and negative current flows is therefore a series of positive and negative going train of pulses, each one of which represents the impulse response of the amplifier system.

Since the charge injected into the system can be well defined, ie

number. of electrons = $C_tV_t/e$

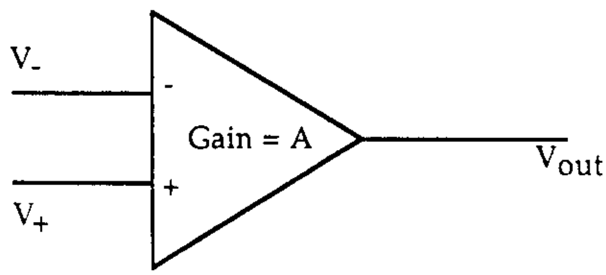
or equivalent energy = $C_tV_t.\varepsilon/e$ with $\varepsilon$=3.6eV

then the response of the system can be calibrated by measuring the amplitude, at its peak, of the output pulse $v_{out}$. The value of the ratio $v_{out}/Q$ is an *arbitrary* quantity (ie its absolute value has no significance) which depends on the *arbitrary* gain of the system. However it can now be used to estimate the sizes of signals from interactions in the detector in useful units of either numbers of electron-hole pairs or energy deposited.

As in all such arguments some assumptions have been made which can be questioned, such as linearity of the system, or treated as experimental problems. What is the requirement on pulse repetition rate, ie the interval T between pulses?

It is helpful to be able to analyse some of the circuits used in this experiment and some useful results can be obtained with only an elementary knowledge of electronics. The operational amplifier is a well defined building block that is used in many circuits and its properties can be exploited without detailed knowledge of its internal structure. An excellent introduction to it, and much other valuable electronics, is contained in the book *The Art of Electronics* by P.Horowitz and W.Hill published by Cambridge University Press.

The operational amplifier (op. amp) is a very high gain amplifier with a single output but dual, differential inputs. In other words the voltage at the output is the product of the voltage difference between the inputs multiplied by the gain.




$$V_{out} = A\,(V_+ - V_-)$$

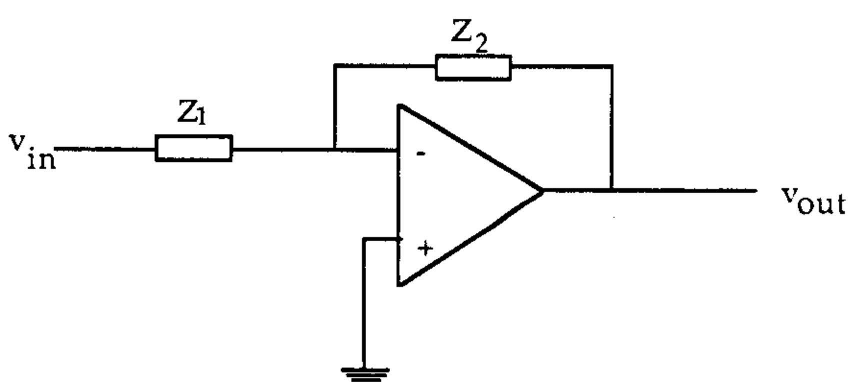
In reality the op-amp is never used in this simple way but is almost always used in a feedback mode where some fraction of the output voltage is fed back to the inverting ($v_-$) input where it is subtracted from the input signal at this point. This is called negative feedback and explains why the inputs are labelled as + and -: the output goes positive when the non-inverting input ($v_+$) goes more positive than the inverting input($v_-$), and vice-versa.

In practice the gain of an op-amp is made so high and they are designed to draw such small currents at the inputs that circuits with them can be analysed with two simple rules:

*i) the output attempts to set itself so that the voltage difference between the inputs is zero*

*ii) the inputs draw no current*

Applying these rules to the circuit below:



$$v_- = v_+ = 0$$

$$(v_{in} - v_-) / (v_- - v_{out}) = Z_1/Z_2$$

$$\rightarrow v_{out} / v_{in} = -Z_2/Z_1$$

In the first part of the experiment this result was used to derive the performance of the C-V analyser circuit; it can also be used to understand the behaviour of the charge sensitive preamplifier as explained in the text. In that case

$$v_{in} / Z_1 = i_{in} \qquad \text{and} \quad Z_2 = 1/j\omega C_f$$

$$v_{out} = -i_{in}/j\omega C_f$$

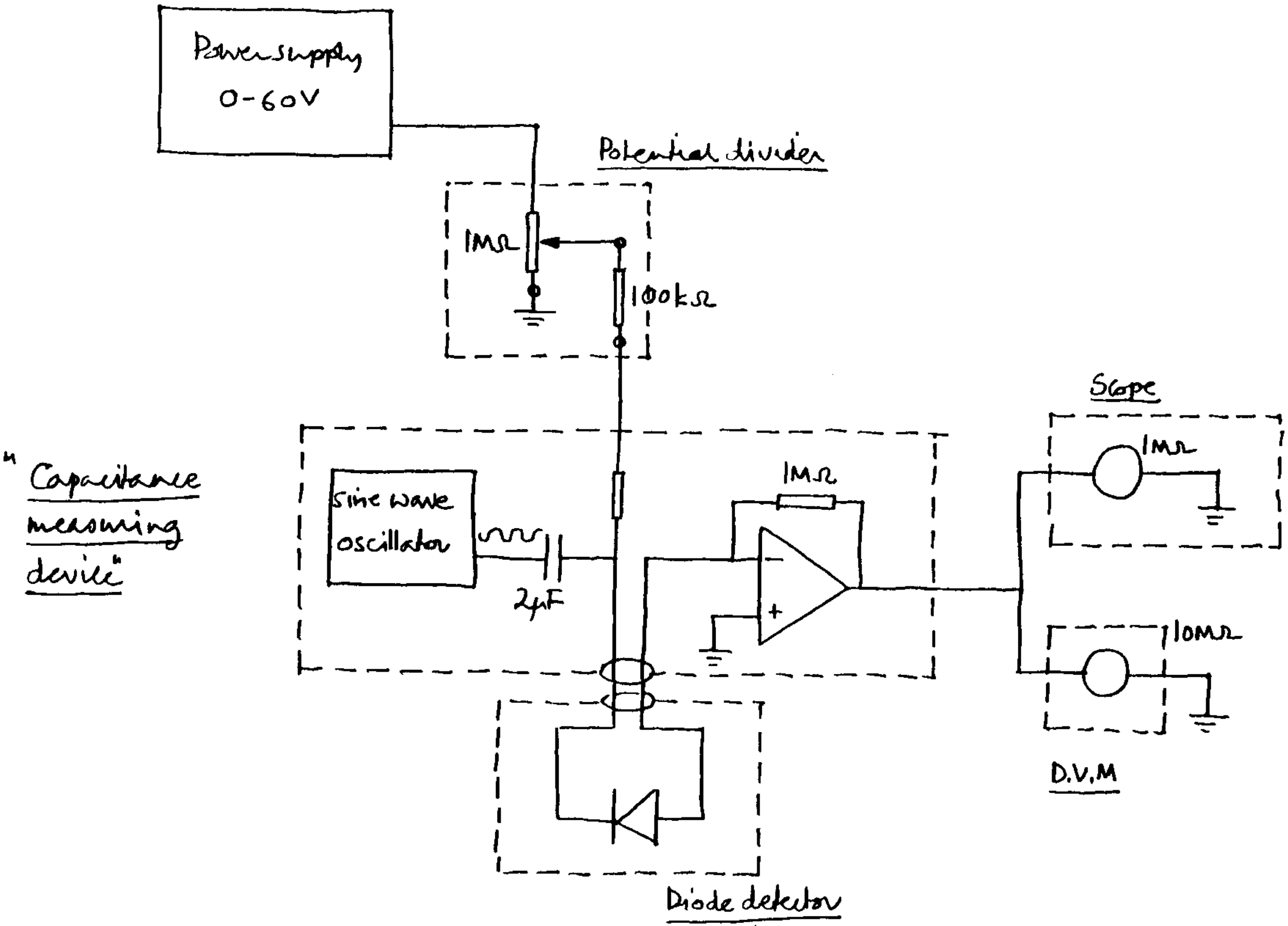
This is the same as the result given above but expressed in a slightly different fashion. In fact, to be consistent, we should recognise that we are working in the frequency domain rather than the time domain and we therefore write more explicitly.

$$v_{out}(\omega) = -i_{in}(\omega)/j\omega C_f = -Q/j\omega C_f \qquad \text{(for a } \delta \text{ function input)}$$

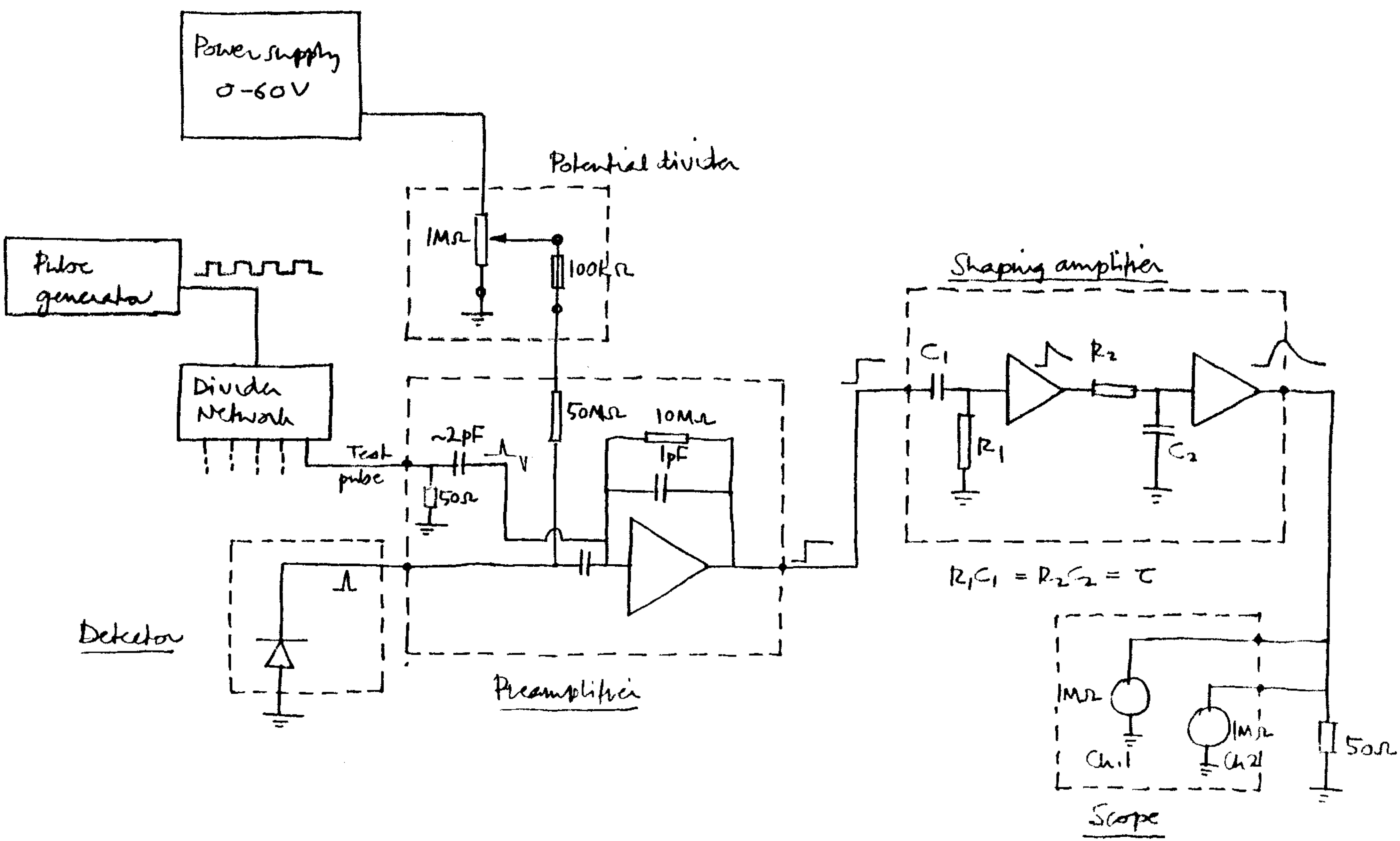
which is recognisable as the Fourier transform of a step function in time.

In reality the ideal op-amp does not exist and the details of the performance of these circuits are affected by the deviations from ideal behaviour. One of these, the effect on the rise time of the amplifier, was mentioned in the text. Other characteristics can be understood with the aid of more knowledge of the internal structure, at the transistor level, of some of the elements of the op.amp. Hopefully this brief introduction will encourage you to look a bit more into this aspect of the circuits you are using.

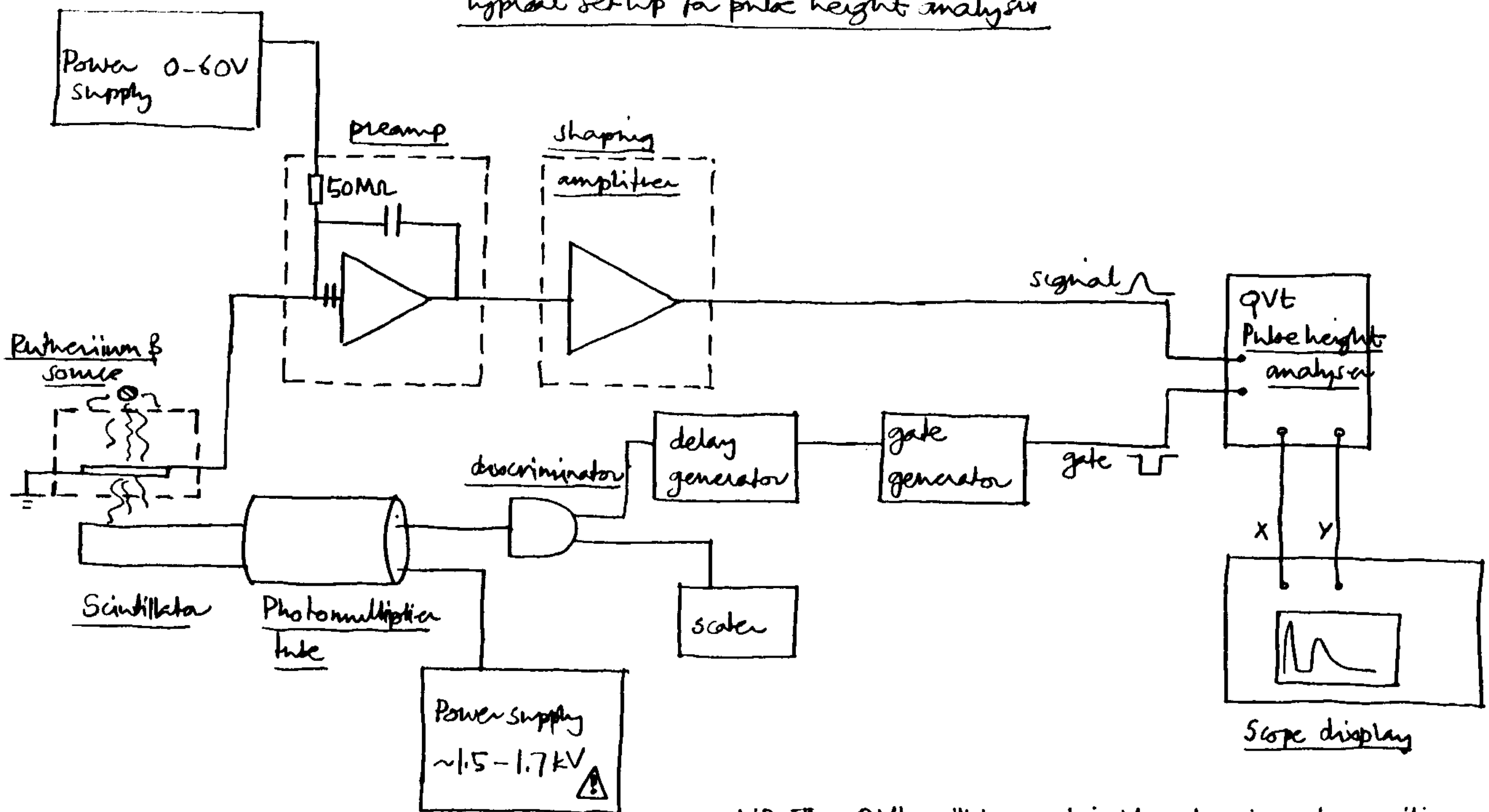
# Setup for Capacitance/Leakage current measurements



Basic setup for signal & noise measurements

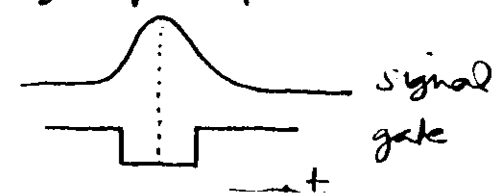


Typical set up for pulse height analysis




NB The QVt will be used in V mode, ie peak sensitive to positive going pulses. The gate should be delayed so that it encloses the signal pulse peak.

ie



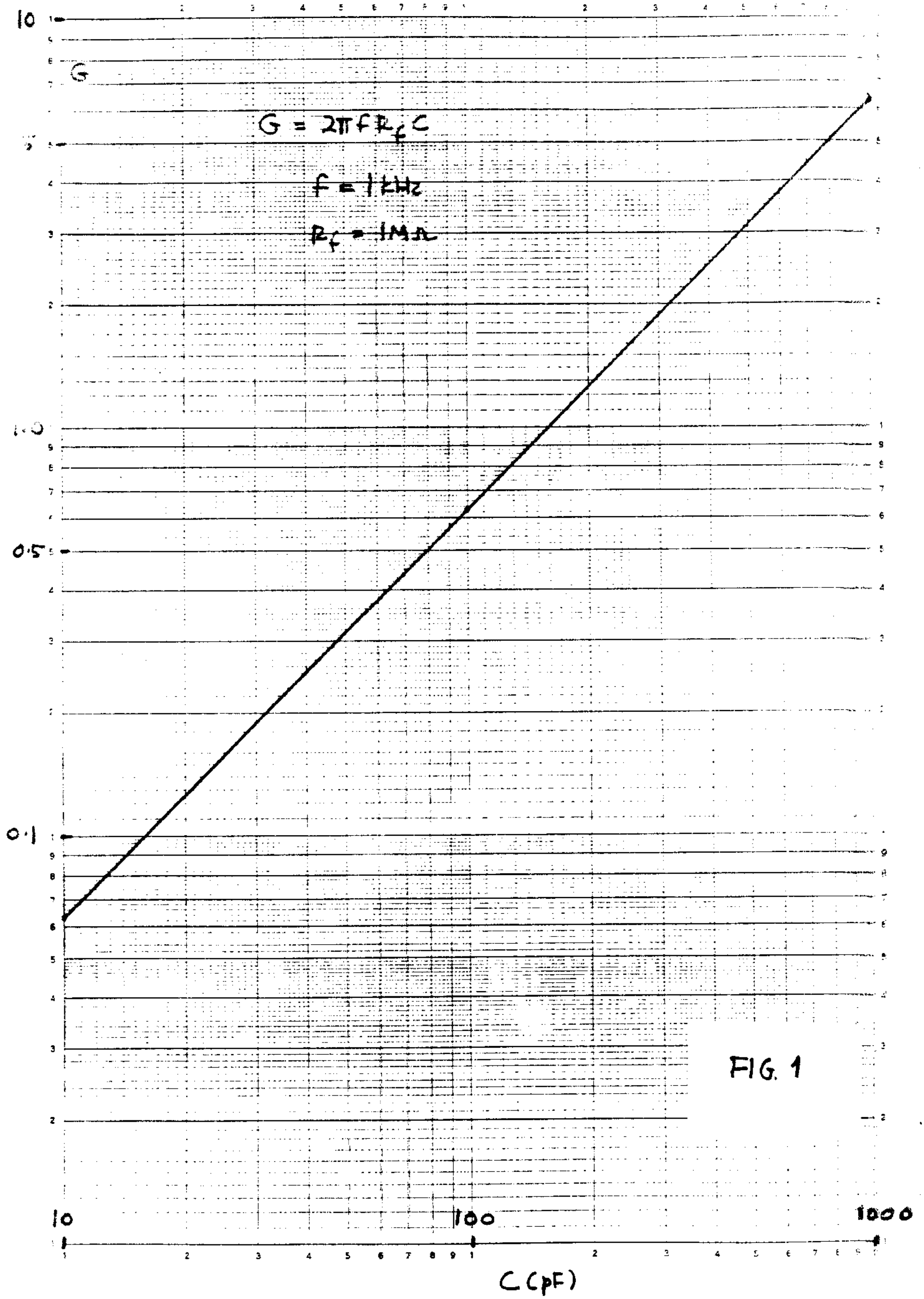
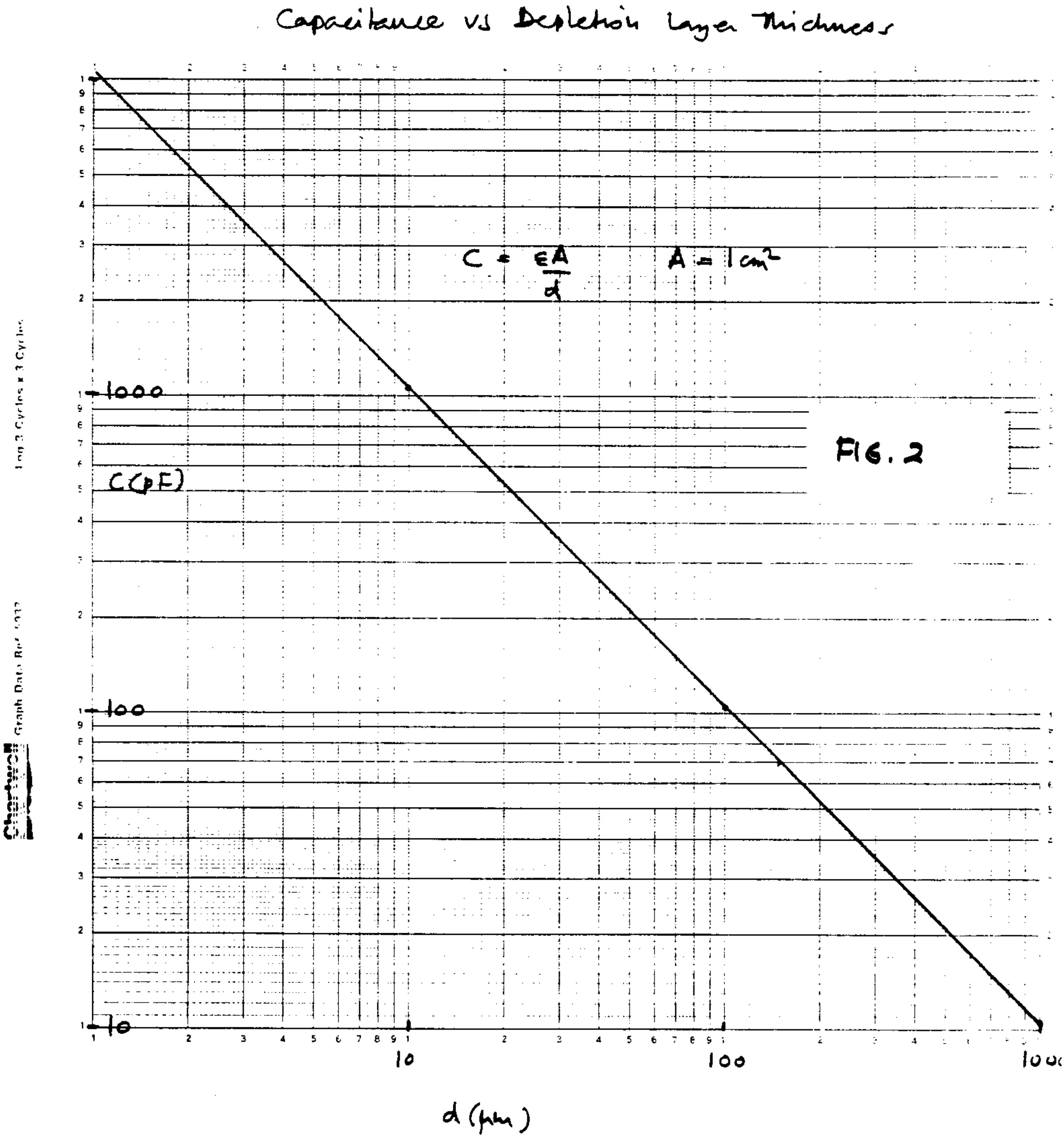



FIG. 1



Capacitance vs Depletion Layer Thickness
$C = \frac{\epsilon A}{d}$
$A = 1 cm^2$
1000
100
10
C(pF)
10
100
1000
d (μm)
Graph Data Ref. 5232
Log 3 Cycles x 3 Cycles


FIG. 2

Depletion depth vs Voltage and Resistivity

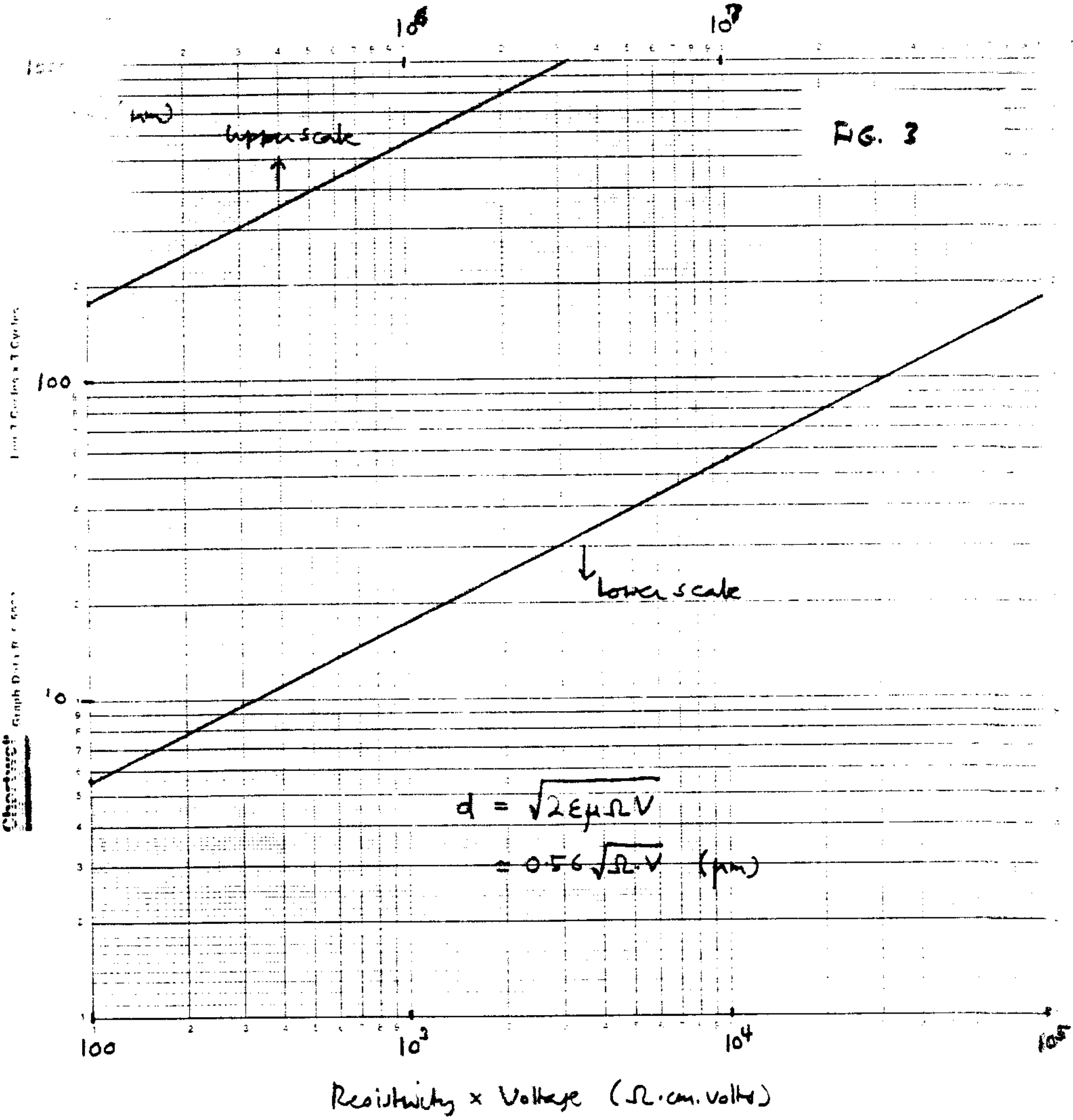



FIG. 3

Resistivity vs Doping Concentration

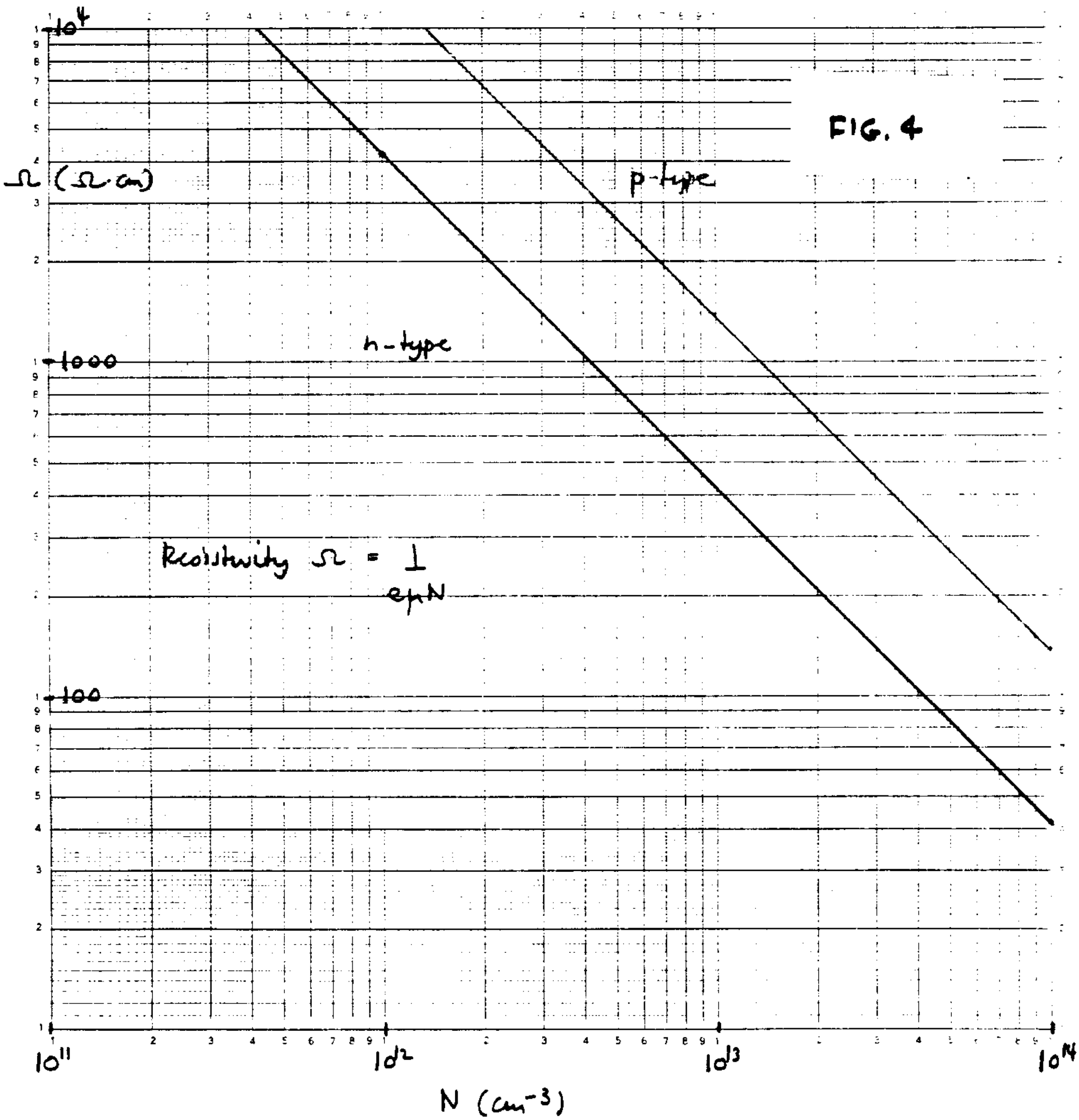


FIG. 5

$\phi = \frac{kT}{e} \ln\left(\frac{N}{n_i}\right)$ at $T \simeq 300°K$