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Contents

Abstract	3
I. Introduction	4
II. Amplifier Signal-to-Noise Ratio	5
A. Choice of Input Tube and Shunt Input Resistance for Best Resolution	7
B. Optimum Amplifier Time Constant	8
C. Summary of this Section	9
III. Detector Parameters	9
A. Summary of this Section	11
IV. The Guard Ring Detector	12
A. Experimental Results	13
B. Summary of this Section	15
V. Energy Resolution as Particle Detector	15
VI. Construction Details of Guard Ring Counter	17
Appendices	
A. Derivation of Detector Formulae	20
B. Derivation of Noise Formulae	23
1. Sources of Noise	23
2. Effect of Amplifier Bandwidth on Signal and Noise at Output	25
C. Special Electronic Circuits	27
1. Low-Noise Preamplifier	27
2. Biased Amplifier	28
Acknowledgments	29

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ABSTRACT

The theoretical limits of noise in detector-amplifier combinations are discussed and related to the bulk properties of the semiconductor (lifetime and resistivity). A new detector structure which includes a guard ring as an integral part of the detector is described and its effect in eliminating surface leakage is discussed. Experimental results in good agreement with theory for detector leakage and noise resolution are presented. The residual surface effects not eliminated by the guard ring are shown to be important in very-low-leakage detectors, and the results of surface treatments to reduce these effects are briefly mentioned.

Observations of the energy resolution of these detectors using various types of particles indicate that factors other than noise are important. Surface imperfections such as are encountered in polished surfaces may cause poor resolution when heavily ionizing particles are detected. Multiple peaks have also been observed in a spectra obtained with diffused junctions, possibly due to other surface effects. A brief description is given of the special electronic circuits used in these measurements.

Techniques used in producing the guard-ring counter are described in detail. These employ gaseous diffusion to produce a thin surface junction, followed by etching through a photo-resistant mask to produce the desired geometry.

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I. INTRODUCTION

Semiconductor junctions are finding increased use as radiation detectors. Although they are likely to be valuable replacements for existing types of detectors, possibly their main importance lies in the field of particle energy determination. Because absorption of a given amount of energy from the radiation produces about 10 times as much ionization in solids of interest as in gases used in ionization chambers, energy resolution can be improved by a similar factor. Existing methods of achieving energy resolutions comparable to the possible performance of silicon detectors are expensive and essentially use low geometric counting efficiencies. Therefore it is quite possible that experiments hitherto considered impossible will become practicable through use of these detectors. Their importance may be increased by their inherent short collection time, of great interest in the detection and timing of minimum ionizing particles.

The investigation described here was undertaken to determine the fundamental limits of performance of this type of detector. Literature has appeared describing various types of junction detector, some relying on surface barriers, while others use diffused junctions similar to those employed in commercial semiconductor devices. It has been our experience that the behavior of these simple detectors is unreliable. Excellent results do occur occasionally for no apparent reason, but the method of manufacture seems to be poorly controlled, and the detectors are seriously affected by their environment. For example,

some detectors operate well in vacuum but not in air, while in other cases the reverse occurs. This example demonstrates that a major problem in these detectors arises at the edge of the junction where the p-n transition reaches a surface whose properties depend upon surface contaminants.

A major objective in our work was therefore to develop a device in which the surface problem is reduced. The suggestion that a guard ring be applied to a semiconductor junction edge is not new. However, because a suitable geometry was lacking, guard rings have not been used prior to the work described here. In practice, our guard ring is more complicated in its behavior than the simple insulator guard ring. This is discussed in detail later, but at this stage it is adequate to note that surface effects can be reduced to negligibility by careful design. Once this is achieved, the bulk properties of the semiconductor become the limit to performance. In particular, the lifetime of the minority carrier is of great interest.

Relating the electrical parameters of the junction detector to its energy resolution, when used as a particle detector, involves consideration of the spread in output pulse amplitude from the detector—pulse amplifier system. Since electrical noise may in some circumstances be the primary limitation, the amount and behavior of this noise must be investigated under various conditions.

II. AMPLIFIER SIGNAL-TO-NOISE RATIO

A junction detector may be looked upon as a solid ionization chamber, and it is natural to consider how the electrical parameters compare with those of the gaseous ionization chamber. For noise calculations, ionization chambers may be regarded as very-high-impedance charge sources shunted by capacity. The junction detector acts in a similar way. The detector impedance is somewhat lower than that of an ionization chamber, and its capacity is greater. Also, a significant leakage current flows across the junction. Despite these slight differences of

behavior the general signal-noise considerations are the same as those developed for ionization chamber amplifiers, but the optimum design parameters of the system may differ with the particular detector. Formulae similar to those developed previously for ionization chamber amplifiers¹ are presented in Table I. These formulae assume that the pulse amplifier contains single integrating and differentiating circuits having equal time constants, and that the detector collection time is very small compared with the amplifier time constant. Although a slight improvement in signal to noise can be obtained by using more complex pulse-shaping networks, the case considered here achieves almost the ideal result and lends itself to simple adjustment. This is essential if theory and experiment are to be compared. Note that the results given in Table I assume that the ambient temperature is 25°C.

In Table I, note that the tube-flicker-effect contribution to total noise is independent of the actual value of amplifier time constant. Calculation also shows that it is small compared with the other terms in the practical case. Note, also, that any other noise having a $1/f$ frequency dependence (as the flicker effect) would also result in a noise component independent of amplifier time constant. In our work we have assumed that no such $1/f$ noise apart from tube flicker noise is present; the justification for this assumption will be illustrated later when experimental results are presented. However, $1/f$ noise may be produced by semiconductor surface effects. The fact that we do not observe a $1/f$ noise term (i. e., independent of amplifier time constant) indicates that our surface effects are very small.

¹These formulae are derived in Appendix B. They are similar to ones developed in A. B. Gillespie, Signal, Noise and Resolution in Nuclear Counter Amplifiers (Pergamon Press, London, 1953).

The remaining terms in Table I all demonstrate some dependence on amplifier time constant. Tube shot noise² is inversely proportional to amplifier time constant, while grid current noise², detector leakage current noise,² and resistance noise² are all linearly proportional to amplifier time constant. Thus, an optimum amplifier time constant exists at which the signal-to-noise ratio has its greatest value. This optimum time constant lies generally in the range 0.1 μ sec to 2 μ sec, depending upon the system parameters.

Figure 1 presents curves calculated from Table I illustrating the variation of the mean square noise with the various parameters in the equations. The curves assume a tube grid current of 2×10^{-9} amp and a total shunt input resistance of 5 M Ω , each easily attained in practice. For short amplifier time constants, tube shot noise is dominant, whereas detector leakage-current noise is dominant for long amplifier time constants.

A. Choice of Input Tube and Shunt Input Resistance for Best Resolution

In an ionization chamber amplifier, the grid current of the input tube, the shunt input resistance, and the tube mutual conductance determine the optimum circuit, since the detector exhibits no leakage. However, semiconductor junction detectors at present exhibit leakages much larger than the input-tube grid current. Therefore, grid current becomes less important and we choose a tube with a higher mutual conductance and grid current than would be optimum for ionization chamber uses. Measurements reported here use 417A input tubes operated at an anode voltage of 100 v and 10 ma current. Under these conditions the grid current is about 2×10^{-9} amp and mutual conductance 16 ma/v.

Table I may be used to determine a suitable value of input shunt resistance. Equating detector leakage current noise² to input resistance noise² permits the choice of R so as to make its noise contribution much smaller than that resulting from detector leakage. In our detectors $i_L > 25 \text{ m}\mu\text{a}$ under normal operating conditions, which means that a shunt impedance $> 5 \text{ M}\Omega$ is desirable. The curves in Fig. 1 are calculated for $R = 5 \text{ M}\Omega$. Ideally, a larger value of R should be used.

B. Optimum Amplifier Time Constant

Table I also permits ready practical calculation of the optimum amplifier time constant. Since tube shot noise and detector leakage current noise are dominant, the remaining terms may be neglected. Making this assumption, we have

$$\tau_{\text{opt}} = \frac{0.35C}{\sqrt{g_m i_L}} \quad (1)$$

By using this value of amplifier time constant, the optimum total mean square noise may be calculated. In the following equation, flicker effect, resistance noise, and grid current noise are neglected:

$$\begin{aligned} \langle \text{noise}^2 \rangle &= 4 \times 10^{-2} \frac{C^2}{g_m \tau_{\text{opt}}} \text{ kev}^2 \\ &= 0.11 C \sqrt{\frac{i_L}{g_m}} \text{ kev}^2 \end{aligned} \quad (2)$$

For example:

$$\text{If } g_m = 16 \text{ ma/v,}$$

$$C = 80 \text{ pf (typical in our experiments),}$$

$$i_L = 50 \text{ m}\mu\text{ amp; using 1-cm diam detector,}$$

$$\tau_{\text{opt}} = 1.0 \text{ }\mu\text{sec}$$

$$\langle \text{noise}^2 \rangle = 15 \text{ kev}^2,$$

$$\text{Full width at half max of resolution curve} = 9.2 \text{ kev.}$$

This is in good agreement with the curves of Fig. 1, if one considers that the curves include all noise contributions whereas the simplified calculations include only tube shot noise and detector leakage current noise.

C. Summary of this Section

Equations (1) and (2) express the theoretical optimum noise conditions in a detector-amplifier system. The curves of Fig. 1 show the dependence of noise on amplifier time constant. The shape of these curves is used in the following work to show that the noise frequency spectrum agrees with this theory. Detector surface noise and detector series resistance noise have been omitted in this discussion. Our experience indicates that these are usually negligible in well-designed detectors, but the additional contribution due to series resistance can be calculated without difficulty if required.

III. DETECTOR PARAMETERS

The work described is confined to junctions manufactured by diffusing a donor impurity (phosphorus) into high resistivity p-type silicon. Phosphorus surface concentration, diffusion temperature and time were chosen to produce a very thin skin of highly doped n-type silicon on the face of the p-type wafer.

Table II presents formulae for calculating the depletion layer width, capacity, diffusion and generation currents in the types of junction used in this investigation. Appendix A shows how these formulae are derived from those appearing in the literature and gives general relationships which may be applied to any situation.

To illustrate the order of magnitude of the detector parameters, and using the formulae given in Table I to calculate the predicted characteristics of a 1-cm-diam detector using 1500 ohm-cm p-type material having a equivalent² minority carrier lifetime of 500 μ sec, and carrying out

². This is not the normal minority carrier lifetime-see note in Appendix 1.

the calculation for an applied voltage of 200v, we have

$$\text{Detector area} = 0.8 \text{ cm}^2,$$

$$\text{Depletion layer width} = 176 \mu,$$

$$\text{Detector capacity} = 48 \text{ pf},$$

$$\text{Generation current} = 34 \text{ m}\mu\text{a},$$

$$\text{Diffusion current} = 0.6 \text{ m}\mu\text{a},$$

(taking t = wafer thickness = 35 mils = 900 microns).

Note that the diffusion current is very small compared with the current caused by generation of carriers at trapping centers in the depletion layer. In a well-designed detector this will always be so, and for this reason, we neglect discussion of diffusion currents.

By using Eq. (2), which relates noise to detector characteristics, with the results of Table II the following mean square noise at optimum amplifier time constant for the system can be derived:

$$\langle \text{noise}^2 \rangle = \left[\frac{A}{\tau_0 g_m} \right]^{1/2} 0.7 C_{in} (\rho v)^{1/4} + \frac{2.2 \times 10^4 A}{(\rho v)^{1/4}} \text{ kev}^2 \quad (3)$$

Where A = detector area in cm^2 , C_{in} = input capacity of the system apart from the detector, and the remaining constants are as defined in Tables I and II.

The first term in Eq. (2) is proportional to $(\rho v)^{1/4}$, while the second is inversely proportional to this quantity. Therefore an optimum operation voltage must exist from the point of view of signal to noise. This voltage is given by

$$v(\text{opt}) = \frac{10^9 A^2}{\rho C_{in}^2} \text{ volts} \quad (4)$$

Inserting this into Eq. (3) the best performance we can expect with a given semiconductor material and input tube in the amplifier is

$$\langle \text{noise}^2 \rangle (\text{opt}) = 250 A \sqrt{\frac{C_{\text{in}}}{\tau_0 g_m}}$$

Using these equations to predict optimum performance for the type of detector used in the earlier example:

$$A = 0.8 \text{ cm}^2,$$

$$\rho = 1500 \text{ } \Omega\text{cm},$$

$$\tau_0 = 500 \text{ } \mu\text{sec},$$

$$g_m = 16 \text{ ma/v},$$

$$C_{\text{in}} = 25 \text{ pf},$$

$$\langle \text{noise}^2 \rangle (\text{opt}) = 11.4 \text{ kev}^2,$$

$$\text{Equivalent full width at half max} = 7.8 \text{ kev},$$

$$\text{Optimum applied voltage} = 615 \text{ v},$$

$$\text{Detector leakage current} = 72 \text{ m}\mu\text{a}$$

$$\text{Detector capacity} = 35 \text{ pf},$$

$$\text{Optimum amplifier time constant} = 0.62 \text{ } \mu\text{sec}.$$

A. Summary of this Section

Equations were developed permitting theoretical study of the ultimate noise performance of the detector-amplifier system. According to Eq. (5), ultimate performance depends upon the lifetime of minority carriers in the material, and upon the input tube mutual conductance and its capacity. The resistivity of the material did not appear directly in this equation.

As an intermediate step in the development of the noise relationship, the leakage current of the detector as a function of resistivity and carrier lifetime in the bulk semiconductor were studied. We will now deal with practical detectors and use detector leakage current, capacity and noise to test agreement between theory and practice.

IV. THE GUARD RING DETECTOR

In order to realize practically the theoretical behavior predicted in the foregoing sections, it is clear that surface leakage effects must be eliminated. Leakage current values predicted in Table II are one or two orders of magnitude lower than is commonly measured in simple junction detectors. To decrease surface leakage as far as possible, the guard ring structure shown in Fig. 2 was devised. Phosphorus was diffused into p-type silicon, then the wafer was etched through a photoresist mask to prevent etching in the detector and guard ring areas on the front face. This resulted in breaking of the n-type skin on the front face into two areas; a central area used as the detector, surrounded by a guard ring area. The space between guard ring and detector should be as small as practicable; we used 2.5 mils spacing in the work described here.

In some respects the behavior of the guard ring resembles that of the conventional insulator guard ring. Since no or only very small voltage exists between central areas and guard ring, no average current can flow out across the surface from the central region. However, although this applies for any guard-ring-to-central-region spacing, it should be obvious that it is not the only condition for eliminating surface noise. Indeed, if the spacing is large, one might expect surface noise due to the edge of the central junction region to be increased by the presence of the guard ring, although average surface leakage out from the central region might still be zero. This anomaly can be explained by assuming equal and opposite currents to flow from central region to guard ring and vice-versa. Since each current contributes its noise component, the guard ring has, in fact, increased the surface noise instead of reducing it. This argument suggests that the guard-ring-to-central-region spacing should be very small. A little consideration shows that the spacing should be made very small compared with the depletion layer width at the operating voltage of the detector.

In this case the depletion layer "pinches off" the space between guard ring and central region as the applied voltage is increased beyond a few volts. Therefore, the surface of the etched ring rises in potential as the applied voltage is increased.

Surface leakage in p-n junctions is generally attributed to three causes:

- (a) Ionic conduction on the outside of an oxide film.
- (b) Surface channeling produced by an inversion layer on the surface of the p- or n-type material. This results in ohmic behavior in the reverse voltage characteristic of the junction.
- (c) Injection at the surface into the bulk of the material.

The guard ring entirely eliminates (a) and (b), but the possibility of (c) still remains. The magnitude of (c) depends upon the surface treatment of the region between guard ring and detector, but our measurements indicate that in all cases leakage current is much lower in a guard-ring detector than in the simple detector where (a) and (b) are likely to be present.

A. Experimental Results

Several guard ring counters were constructed, and, in general, the measurements on these counters agreed with theory. Leakage current measurements were carried out by using the potentiometer arrangement shown in Fig. 3, ensuring that measurement was performed with the guard ring and detector areas at the same potential. It also permitted measurement of the impedance across the surface of the etched ring which shunts the detector load and may thereby increase the system noise.

A typical plot of detector capacity is shown in Fig. 4, and the leakage current is shown in Fig. 5. The improvement obtained by using the guard ring is apparent, since in this case the guard-ring leakage current was nearly $10\mu\text{a}$ at 200 v and a simple type of detector with the same surface area would presumably also exhibit several μa leakage.

The detector capacity obeyed an inverse-square-root law with voltage, as predicted by Table II. This implies that the depletion layer width was proportional to $\sqrt{V}$. However, the detector leakage current obeyed the expected $\sqrt{V}$ law only at high voltages. The departure at low voltages is believed due to injection effects where the p-n junction reaches the surface. This belief is strongly supported by the observation that operating the counter in different environments produced considerable change in the low voltage part of the curves in Fig. 5. Small traces of chlorine (known to produce a p-type surface) caused a large increase in the leakage current particularly at low voltages, while traces of ammonia (known to produce an n-type surface) caused the current at low voltages to fall almost to the value expected on the basis of a $\sqrt{V}$ dependence. Therefore, to obtain the ultimate performance a surface treatment resulting in a stable, very light "p" or "n" surface in the etched ring is necessary. No quite satisfactory surface treatment has yet been found, but silicone resins and oxide film growth seem promising.

Measurements of the noise of a detector-amplifier system made by using the detector whose characteristics appear in Figs. 4 and 5 are shown in Fig. 6. Appendix C describes the low-noise preamplifier and biased amplifier units used for these measurements. The curve shown in Fig. 6 agrees very well with that predicted from Fig. 1, which indicates that the equations given in Table I accurately represent all noise sources and that no additional significant source of noise is present.

The leakage current curve of Fig. 7, rather than Fig. 5, shows the best result. The current in Fig. 5 corresponds to an electron lifetime of 130 μsec while that of Fig. 7 corresponds to one of 850 μsec . As yet we have insufficient experience to indicate the best treatment, but the spread in lifetimes we apparently encountered showed that improvement may be possible. Incidentally, the wafer used in the detector of Fig. 5 was cooled quickly by withdrawing into air at room

temperature after diffusion whereas that of Fig. 7 was slowly cooled; however, the increase in lifetime in the latter case is probably fortuitous. One point to note is that no satisfactory method seems to exist for measuring minority carrier lifetime in thin wafers of high resistivity material.

Surface treatment of the etched ring affects the impedance measured between guard ring and detector as well as influencing carrier injection. The n-type surfaces tend to cause a low impedance, but guard ring and detector are virtually isolated if a p-type surface is produced. If the surface is only lightly n-type, impedance is small when a low voltage is applied to the detector, but increasing the applied voltage depletes the surface and impedance increases. Thus, as stated above, a light n-or p-type surface is ideal. A typical impedance variation with applied detector voltage for what we believe to be a light n-type surface is shown in Fig. 8.

B. Summary of this Section

Experimental results on guard ring detectors show good agreement with theory for high applied voltages, but discrepancies arise at low voltages due to surface effects in the etched ring. Surface treatments to overcome these effects are being studied.

V. ENERGY RESOLUTION AS PARTICLE DETECTOR

The main purpose here has been to evaluate electrical noise in a detector amplifier system. It has been shown that, if surface problems are eliminated, reasonable agreement exists between theory using measured values of detector leakage and practical noise measurements. However, other limits to the particle energy resolution of practical detectors may exist.

In general, our energy-resolution experiments using β particles indicate that the limit to β -particle resolution is electrical noise and it appears that the

equations derived in the earlier part of the paper may be used to determine the resolution. We measured an energy resolution (full width at half max) of 10 kev, using conversion electrons from an Hg^{203} source and a 0.8-cm^2 detector.

On the other hand, the measured resolution figures for α particles were much larger than is accounted for by electrical noise. The best resolution we obtained on 6-Mev α particles was 22 kev (full width at half max), and, in many cases, resolutions much worse than this were observed. Several early detectors were produced with mechanically polished surfaces and, since surface damage caused by the polishing appeared to penetrate 0.5μ or more below the surface, it seemed reasonable to attribute some α -particle pulse-spread to absorption in this damaged layer. Recent detectors have used smooth etched surfaces, but the α -particle resolution was not greatly improved by this. Several detectors exhibited multiple peaks in α spectra when only a single peak should have been present. The reason for this multiple peaking is not clear, but it is likely that the poor resolution of even the best detectors is partly due to unresolved multiple peaks. Observation that the spectrum resolution was improved by using long amplifier time constants indicated that the multiple peaks were due to slight changes in charge collection time at different points on the detector surface. This may have been due to a trapping phenomenon.

An intensive investigation of effects just under the detector surface is required before the ultimate α -particle resolution can be realized. This presumably also applies to other heavily ionizing particles.

VI. CONSTRUCTION DETAILS OF GUARD RING COUNTER

All guard ring counters constructed have used phosphorus diffusion into 1500- or 5000-ohm-cm p-type silicon and have had identical guard ring geometries. Wafers 15 to 40 mils thick were cut parallel to the 1,1,1 plane from single optically oriented 2-cm - diam silicon crystals. Saw damage was removed by lapping several mils from each side with "Lapmaster" #1950 grinding compound. One side of each wafer was masked with PiceinSM sealing wax and 2 to 5 mils etched from the other side with a polishing HNO_3 -HF mixture (10 parts HNO_3 to one part HF). The etched wafers were then cleaned successively in trichloroethylene, methyl alcohol, and deionized water, in preparation for phosphorus diffusion.

Phosphorus diffusion was performed in the quartz-lined 2-in. -diam tube furnace at 900°C . The phosphorus source was P_2O_5 from a furnace-pre-heated zone held at 210 to 220°C , and carried with about 2 liters per minute of dry nitrogen gas obtained by boiling liquid nitrogen. The P_2O_5 source was aged for at least 1 hr prior to diffusion. A typical diffusion schedule is 5 min at 900°C , followed by turning off the furnace and allowing the wafers to anneal. This diffusion schedule was intended to give a calculated junction depth of about $0.05\ \mu$. The junction depth has not been measured, but controlled etching tests showed that it was less than $0.5\ \mu$. The sheet resistance of the diffused layer is found by four-point probe measurement to be 4 to 5 ohms per square, which indicates a surface phosphorus concentration of between 5×10^{21} and 10^{22} phosphorus atoms per cm^2 .³ The phosphosilicate glass grown on the silicon was not thick enough to interfere with resistivity measurements.

A P^+ back-contact was made for each wafer by removing the phosphorus diffusion from the lapped side by hand-lapping, then by alloying with aluminum

³If the sheet resistance and junction depth are known, the surface concentration can be found from curves given by G. Bakenstross, Bell System Tech. J. (May 1958), pp. 699-709.

or gold-gallium. In each case, about $2\ \mu$ of the contact metal was applied by vacuum vaporization and the alloying performed in dry nitrogen at 600°C for Al and 500°C for Au-Ga.

The guard-ring structure was formed by a modified photoengraving technique. All counters made to date have used the same engraving negative or stencil. The stencil was made on a glass Kodalith plate by a 20-fold reduction of a line drawing that gave a guard ring diam of 0.6 in., a counting area diameter of 0.4 in., and an interelectrode spacing of 2.5 mils. Several attempts were made to use Kodak Photo Resist (KPR) as the etching mask, without success. In all cases the resist was penetrated in about 30 sec with even the weakest silicon etches. The etching procedure which has proved most effective was a mask with Kodak Metal Etch Resist (KMER) followed by etching in silver-glycol.⁴ One part of KMER was diluted with two parts of thinner and applied according to the manufacturer's literature. Exposure time was 12 min at 12 in. from a 100-watt mercury arc, with the sensitized wafer and stencil held in contact by a vacuum frame. It is important that the counter be washed with HF just prior to resist application, or the etch will quickly undercut the resist.

Etching was carried out in silver-glycol etch maintained at 25°C in a water bath for times ranging from 5 to 30 min, giving etching depths ranging from 0.5 to $3\ \mu$. No significant differences in performance were noted that were a function of the etching depth. The etching time routinely used was 10 min, which gave $1\ \mu$ depth. Before etching, the back contact was protected with Picein wax, as both Al-Si and Au-Si eutectics are soluble in the etchant.

After etching, the counter was washed with deionized water and then methyl alcohol and the resist was removed with trichlorethylene on a cotton swab. A final wash with deionized water followed by drying with a blast of dry nitrogen completed the procedure required before testing the counter.

⁴For the formula and method of using silver-glycol etch, see Transistor Technology, Vol III, F. J. Biondi, Ed. (D. Van Nostrand, Princeton, N. J. 1958), p. 158.

Several attempts have been made to protect the junction edges with various resins, waxes, and varnishes, but our experience is that these only delayed the absorption of ionic contaminants. The most successful material yet used has been Dow-Corning XR-6-2044 Silicone resin.

APPENDICES

A. Derivation of Detector Formulae

The general formula for depletion-layer width in an abrupt semiconductor junction is given in all textbooks on the subject as

$$W = 1.05 \times 10^{-6} \left[\frac{E(V_0 - V)}{2\pi q(N_a + N_d)} \right]^{1/2} \left[\left(\frac{N_a}{N_d} \right)^{1/2} + \left(\frac{N_d}{N_a} \right)^{1/2} \right] \text{ cm},$$

where:

- E = dielectric constant of material,
- V_0 = built-in potential barrier at junction,
- q = electronic charge,
- V = applied voltage (negative for reverse bias),
- N_a = acceptor concentration on "p" side of junction,
- N_d = donor concentration on "n" side of junction.

In a junction where the "n" side is much more heavily doped than the "p" side, and where the applied reverse bias is much greater than V_0 ,

$$W = 1.05 \times 10^{-6} \left[\frac{EV}{2\pi q N_a} \right]^{1/2} \text{ cm}.$$

The acceptor concentration is related to the resistivity of the material and to carrier mobility by the equation

$$N_a = \frac{1}{q \mu_h \rho}$$

where ρ = resistivity of p-type material, μ_h = hole mobility.

Therefore,
$$W = 1.05 \times 10^{-6} \left[\frac{EV \cdot \rho \cdot \mu_h}{2\pi} \right]^{1/2}.$$

For silicon: $E = 12$, and $\mu_h = 480 \text{ cm}^2 \cdot \text{sec}^{-1} \cdot \text{V}^{-1}$

So, $W = 0.32 (\rho V)^{1/2}.$

A similar equation can be derived where the "p" type material is much more heavily doped than the "n" type material.

To determine the junction capacity we regard the junction as a parallel-plate capacitor with plate spacing W :

$$C = \frac{AE}{4\pi W} = 3.3 \times 10^4 (\rho V)^{-1/2} \rho / \text{cm}^2,$$

To determine the leakage current of the junction, we refer to the work of Sah, Noyce, and Shockley.⁵ They show the junction leakage as containing one component due to diffusion of carriers into the depletion layer, and another due to generation of carriers at trapping centers in the depletion layer itself. The probability of generation by direct excitation of electrons from valency to conduction bands is very small, and the trapping centers are necessary to explain the observed lifetimes in silicon. Sah, Noyce, and Shockley derive the following equations for the two leakage-current components:

$$I_g = \frac{q \times n_i \times W}{2 \tau_0} \text{ per cm}^2,$$

$$I_d = \frac{q \times n \times l}{\tau_0} \text{ per cm}^2,$$

where n_i = number of holes or electrons in intrinsic material,

l = distance from which carriers may diffuse into the depletion layer (the diffusion length in a large volume sample),

τ_0 = minority-carrier lifetime,

n = number of minority carriers in the material,

I_g = generation current,

I_d = diffusion current.

⁵C. T. Sah, R. N. Noyce, and W. Shockley, Carrier Generation and Recombination in p-n Junctions, Proc. I. R. E. 45, 1228 (1957).

A value of 1.5×10^{10} per cm^3 is assigned to n_i for silicon.⁶ Inserting the value of W derived earlier, we have

$$I_g = 38 \frac{(\rho V)^{1/2}}{\tau_0} \text{ m}\mu\text{ amp/cm}^2.$$

The value to use for n in the diffusion current equation is that for electrons (n_e) in the p region of the detector (if the high-resistivity material is p -type.)

$$\text{We have } n_e \times n_h = n_i^2,$$

$$n_e = \frac{n_i^2}{n_h} = n_i^2 \times p \mu_h \times q.$$

Including this in the equation for I_d and putting in values for the constants, we have

$$I_d = 2.75 \frac{\rho \times l}{\tau_0} \text{ m}\mu\text{ amp/cm}^2.$$

⁶See E.M. Conwell, Properties of Silicon and Germanium, Proc. I.R.E. 46. 1281, (1958).

Note added in proof.

A private communication received from T.R. Kohler and E.S. Rittner points out an error in this appendix. Sah, Noyce and Shockley assume that the trap level and the Fermi level coincide in position in order to derive the equations relating I_g and I_d to the carrier lifetime. This is not the normal situation and this fact accounts for the unusually long lifetimes indicated by our measurements. We may regard the lifetime τ_0 used in this report as an equivalent lifetime which would apply if the trap and Fermi levels did coincide. The relationship between τ_0 and the true carrier lifetime in the material can then be used to deduce something about the level of the trapping centers.

B. Derivation of Noise Formulae

Several sources of noise are present in a detector-amplifier combination, and the frequency dependence of each may differ from that of the others. Since the amplifier shaping networks limit its frequency response, computation of the noise output of the amplifier involves integration of the product $N(\omega) \times G(\omega)$ ⁷ in the range $\omega = 0$ to $\omega = \infty$ for each source of noise. Individual noise outputs may then be compounded into a single rms noise output by summing the mean square values of all individual terms and taking the square root of the total.

The noise output in itself is unimportant, the quantity of real interest being the signal-to-noise ratio, usually expressed by quoting the input signal charge required to produce an output signal equal to the rms noise. To calculate this, we consider the effect of the amplifier-shaping networks on signal as well as on noise. In outline, this is the procedure used to derive the formulae appearing in Table I. To simplify, we have considered only that case in which the integrating and differentiating networks in the amplifier have the same value of time constant τ .

1. Sources of Noise

(a) Tube Shot Noise. It is usual to represent this by including an equivalent resistance R_{eq} in series with the grid of the tube. For a triode, $R_{eq} = \frac{2.5}{g_m}$. The mean square noise generated by this resistor at the input in a small bandwidth is given by

$$\langle e_T^2 \rangle = 4 KT R_{eq} \Delta f = \frac{10 KT}{g_m} \Delta f = \frac{5 KT}{\pi g_m} \Delta \omega.$$

(b) Tube Flicker Noise. This is caused by fluctuations in the cathode of the tube and the mean square noise due to this source obeys a $1/f$ law.

Thus $\langle e_F^2 \rangle = \frac{A}{f} \Delta f = A \frac{\Delta \omega}{\omega}$, where A is a constant found experimentally to be approximately 10^{-13} .

⁷ $N(\omega)$ = frequency-dependent noise input term; $G(\omega)$ = frequency-dependent amplifier gain, where $\omega = 2 \pi f$.

(c) Tube Grid Current Noise. The tube grid current is subject to the usual statistical fluctuations:

$$\langle i_G^2 \rangle = 2q \times i_g \Delta f .$$

In the frequency range passed by the amplifier, if the detector load resistance and tube grid leak are made large in value, the impedance in which grid current fluctuations can develop voltage is almost entirely capacitive (due to the total capacity of tube and detector), and we have

$$\langle e_G^2 \rangle = \frac{2 \times i_g}{\omega^2 C^2} \Delta f = \frac{q \cdot i_g}{\pi \omega^2 C^2} \Delta \omega .$$

(d) Detector Leakage Noise. This, like the grid current of the tube, is subject to statistical fluctuations, resulting in a noise given by

$$\langle e_L^2 \rangle = \frac{q \times i_L}{\pi \omega^2 C^2} \Delta \omega .$$

(e) Input Resistance Noise. Shunt resistance in the input circuit produces resistance noise, and a portion of this noise appears across the input capacity of the amplifier. If shunt resistance is large, the current mean square noise produced by it is given by

$$\langle i_R^2 \rangle = \frac{4KT}{R} \Delta f = \frac{2KT}{\pi R} \Delta \omega, \text{ where } T \text{ is absolute temperature.}$$

The noise voltage² produced across the input capacity is therefore

$$\langle e_R^2 \rangle = \frac{2KT}{\pi R} \Delta \omega \cdot \frac{1}{\omega^2 C^2} = \frac{2KT}{\pi \omega^2 RC^2} \Delta \omega .$$

(f) Total Noise at Input to Amplifier. Summing the noise terms described above, we have

$$\begin{aligned} \langle e_I^2 \rangle &= \langle e_T^2 \rangle + \langle e_F^2 \rangle + \langle e_G^2 \rangle + \langle e_R^2 \rangle \\ &= \left[\frac{5Kt}{\pi g_m} + \frac{10^{-13}}{\omega} + \frac{1}{\pi \omega^2 C^2} \left(q(i_g + i_L) + \frac{2KT}{R} \right) \right] \Delta \omega . \end{aligned}$$

2. Effect of Amplifier Bandwidth on Signal and Noise at Output

The gain of an amplifier having equal integrating and differentiating circuits varies with frequency

$$G = G_0 \frac{2\pi f \tau}{1 + 4\pi^2 f^2 \tau^2} = G_0 \frac{\omega \tau}{1 + \omega^2 \tau^2}$$

Thus the mean square noise at the amplifier output is given by

$$\langle e_0^2 \rangle = \int_{\omega=0}^{\omega=\infty} G_0^2 \left[\frac{5KT}{g_m} \frac{\omega^2 \tau^2}{(1 + \omega^2 \tau^2)^2} + \frac{10^{-13} \omega \tau^2}{(1 + \omega^2 \tau^2)^2} + \frac{\tau^2}{C^2 (1 + \omega^2 \tau^2)^2} \left((i_g + i_L) + \frac{2KT}{R} \right) \right] d\omega.$$

Note that each of the three major terms in the integral has a different frequency dependence.

Performing the integration yields the result

$$\langle e_0^2 \rangle = G_0^2 \left[\frac{1.25 KT}{g_m \tau} + 0.5 \times 10^{-13} + \frac{\tau}{4C^2} \left(q(i_g + i_L) + \frac{2KT}{R} \right) \right]$$

If the input signal consists of a very small pulse of charge, the output signal produced by the amplifier is given by

$$\text{Output Signal} = \frac{Q}{2.73} \times \frac{G_0}{C}$$

where Q is the charge produced by the ionizing event in the detector.

Thus the detector charge input producing an output signal equal to the rms noise level at the output is given by

$$\langle Q_{\text{eff}}^2 \rangle = 8 C^2 \left[1.25 \frac{KT}{g_m \tau} + 0.5 \times 10^{-13} + \frac{\tau}{4C^2} \left(q(i_g + i_L) + \frac{2KT}{R} \right) \right]$$

Putting $q = 1.6 \times 10^{-19}$ coulombs

$$\frac{KT}{q} = 25 \text{ mv}$$

t in $\mu \text{ sec}$

R in $M\Omega$

i_g in $m\mu\text{amp}$

i_L in $m\mu\text{amp}$

g_m in ma/v ,

we have

$$\langle Q_{eff}^2 \rangle = 4 \times 10^{-35} \frac{C^2}{g_m \tau} + 4 \times 10^{-37} C^2 + 3.2 \times 10^{-34} \tau (i_g + i_L) + 1.6 \times 10^{-32} \frac{\tau}{R}$$

coulombs².

The various contributions listed in Table I are items in this equation. In order to express the input signal equivalent to noise in terms of energy absorbed from the incident radiation it is necessary to establish a value for the energy absorption per hole electron pair produced in the detector. The generally accepted value for silicon is 3.6 ev/hole-electron pair.

$$\begin{aligned} \langle E_{eff}^2 \rangle &= \text{Mean square energy spread due to noise} \\ &= \frac{\langle Q_{eff}^2 \rangle}{1.6^2 \times 10^{38}} \times 3.6^2 = 5 \times 10^{38} \times \langle Q_{eff}^2 \rangle \end{aligned}$$

If $\langle E_{eff}^2 \rangle$ is expressed in kev² we have

$$\langle E_{eff}^2 \rangle = 5 \times 10^{32} \langle Q_{eff}^2 \rangle \text{ kev}^2.$$

This equation is used to convert the values in the first column of Table I to those in second column.

C. Special Electronic Circuits

A standard linear pulse amplifier (Berkeley Model V) was used for noise and resolution measurements. This was preceded by a special low-noise preamplifier and followed by a biased amplifier unit. These two units are described below. The main amplifier was modified to contain equal integrating and differentiating time constants, the actual value of the time constant being controlled by a switch giving values of 0.2, 0.5, 1, 2, and 5 μsec . The gain stability of this system was not really adequate for accurate α -particle resolution measurements, but in the experiments described here, a standard pulse was always fed into the input of the preamplifier and the spectrum of the reference pulses was examined to determine that the gain drift during any particular experiment was negligible. We are now developing a more stable amplifier.

1. Low-Noise Preamplifier

The schematic of this unit is shown in Fig. 9. A 417A tube V1 is used in a cascode connection with tube V2(a) to give good noise performance. Tubes V1, V2(a), and V2(b) are arranged as a feedback integrator, capacitor C5 acting as the integrator capacitor. The anode load of V2(a) is "bootstrapped" from the cathode of V2(b) to increase the feedback loop gain, thereby increasing the effective capacity (looking into the input of the amplifier) to a large value (about 500 pf). The output signal from the anode of V2(b) feeds the White cathode-follower V3(a,b) which provides a low-output-impedance driver for the cable connecting the preamplifier to the main amplifier.

Note that an integrator arrangement is preferred to a conventional amplifier. This is convenient, since the output from the integrator is less dependent on detector capacity than it would be with a conventional amplifier. The integrator is slightly worse with respect to noise, since the integrating capacitor (C5) must be added to the tube input capacity used in the noise calculation. However, we feel the convenience of the integrator arrangement more than offsets its slightly degraded noise performance

2. Biased Amplifier

The purpose of this unit, which follows the main pulse amplifier, is to provide a method of subtracting a selected dc voltage from all pulses at the amplifier output and of linearly amplifying the part of pulses exceeding the dc voltage by preselected factors. The particular unit described here also shapes the pulses suitable for feeding any common type of multichannel pulse-amplitude analyzer.

The range of bias voltages available in the circuit shown in Fig. 10 is 0 to +50 v. Postbias gains of $\times 1$, $\times 2$, $\times 5$, and $\times 10$ are provided. The signal obtained at the output of the unit is of almost constant amplitude for a 1- μ sec period at the peak of the pulse. Performance of the unit was satisfactory when fed with input pulses produced by an equal-time-constant amplifier having time constant values in the range 0.2 μ sec to 5 μ sec.

In the circuit of Fig. 10, diode CR1 is back-biased by the bias supply, and only those pulses exceeding the bias cause it to conduct. Diode CR2, combined with the capacitors in the grid circuit of V2, lengthens the pulse and shapes it for analysis. Diodes CR3 and CR4 are included to compensate for the effects of temperature changes on the voltage drops in CR1 and CR2.

The signal appearing at the grid of V2 is amplified by the feedback amplifier containing V2 and V3. Switch S1 controls the degree of feedback and therefore the gain. Tube V4 acts as an output cathode follower providing a low-impedance-output source.

The method used to achieve the flat-topped pulse at the grid of V2 is of interest. A normal diode lengthener containing a single capacitor to ground would produce an exponential fall in potential at the grid of V2. It may be shown that by applying feedback through R8, and by splitting the lengthener capacitor into the series combination C4 and C5, a pulse having almost a flat top for a large fraction of its total recovery time can be produced. The value of R8 must be chosen to achieve this result.

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Table I

Noise contributions from various sources			
Noise Source	Input Equivalent ms noise (coulombs ²)	Input equivalent ms noise (Kev ²)	Constants
Tube shot noise	$4 \times 10^{-35} \frac{C^2}{g_m \tau}$	$2 \times 10^{-2} \frac{C^2}{g_m \tau}$	C = total input capacity, picofarads (p) i_g = tube grid current (m μ amp)
Tube flicker	$4 \times 10^{-37} C^2$	$2 \times 10^{-4} C^2$	g_m = tube mutual conductance (ma v) τ = amplifier time constant (μ sec)
Grid Current	$3.2 \times 10^{-34} i_g \tau$	$1.6 \times 10^{-1} i_g \tau$	R = total input shunt R (C $\times$ R assumed $\gg \tau$)
Detector leakage	$3.2 \times 10^{-34} i_L \tau$	$1.6 \times 10^{-1} i_L \tau$	i_L = detector leakage (m μ a amp)
Input resistance unit	$1.6 \times 10^{-32} \frac{\tau}{R}$	$8 \frac{\tau}{R}$	

(a) Third column gives equivalent energy absorbed from an incident particle, assuming 3.6 ev/hole electron pair (correct for silicon).

(b) Full width at half maximum of a resolution curve is approximated by taking the square root of the sum of contributions in column 3 and multiplying by 2.3.

Table II

Detector Formulae

Depletion Layer Width	$W = 0.32 (\rho v)^{1/2} \text{ microns}$
Detector Capacity	$C_d = 3.3 \times 10^4 (\rho v)^{-1/2} \rho f/cm^2$
Generation Current	$I_g = 38 \frac{(\rho v)^{1/2}}{\tau_0} \text{ m}\mu\text{amp/cm}^2$
Diffusion Current	$I_d = 2.75 \frac{\rho \cdot l}{\tau_0} \text{ m}\mu\text{amp/cm}^2$

Here W = depletion layer width,

ρ = resistivity of bulk material (ohm-cm.),

V = applied voltage,

C_d = detector capacity,

I_g = generation current,

τ_0 = minority carrier lifetime in bulk material (μsec),

I_d = diffusion current,

l = thickness of material from which diffusion may occur (cm).

FIGURE LEGENDS

- Fig. 1. Noise variation with amplifier T.C.
- Fig. 2. Guard ring detector.
- Fig. 3. Test circuit for leakage measurements.
- Fig. 4. Typical capacity-voltage relationship.
- Fig. 5. Typical leakage current-voltage relationship.
- Fig. 6. Typical noise variation with amplifier time constant.
- Fig. 7. Best leakage current curve yet attained.
- Fig. 8. Interelectrode impedance characteristic (n-type surface).
- Fig. 9. Low-noise preamplifier for S-C detectors.
- Fig. 10. Biased amplifier for α -particle resolution measurements.

MU-21766

Fig. 1

MU-21767

Fig. 2

MU - 21768

Fig. 3

MU-21769

Fig. 4

MU-21770

Fig. 5

MU-21771

Fig. 6

MU-21772

Fig. 7

MU-21773

Fig. 8

MU-21942

Fig. 9

MU-21941

Fig. 10

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