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ABSTRACT

A new type of preamplifier is described that promises a significant improvement in the energy resolution of semiconductor detector spectrometers. While other applications of the principles are anticipated, this paper deals only with their use for X-ray fluorescence spectroscopy. We have achieved a total electronic contribution to the resolution of about 150 eV (FWHM - silicon) with further small improvements considered likely in the near future. This includes the detector leakage noise contribution. A conventional system using components of the same inherent quality exhibits a resolution of nearly 300 eV.

I. INTRODUCTION

Although the energy resolution obtained at high energies in semiconductor detector nuclear spectrometers is determined mainly by statistical fluctuations of charge production in the detector, electronic noise in the input circuit of the pulse amplifier becomes the major limitation at low energies. While the statistical limitations in detectors are of a fundamental nature, and detector research for high-energy applications is therefore confined to attempting to realise the ultimate statistically possible resolution, the noise performance of preamplifiers has not yet approached a fundamental limit. The resolution capabilities of semiconductor spectrometers at very low energies have improved from about 2 keV (FWHM) in 1965 to 0.3 keV in late 1968. While the larger part of this improvement can be attributed to the availability of low-noise FET's, a steady improvement due to improved techniques must also be given credit.

Work on high-resolution preamplifiers using FET's has involved large expenditures of effort in the selection of FET's to give the best performance. This work has been complicated by a number of factors but the most important is that the noise performance of a given FET under the operating conditions which apply in nuclear preamplifiers bears very little relationship to the normally measured and specified parameters. In part, this results from the high source impedance feeding the FET, but we also achieve some noise reduction by reducing the operating temperature below that commonly used in conventional FET applications, and this carries us outside the normal temperature bounds of manufacturers' interest. However, general guidelines, such as using high mutual conductance (and inevitably high capacity) FET's for high capacity detectors, and vice-versa for low capacity detectors, are commonly employed. It is also observed that the best absolute performance is achieved using FET's that perform well at the lowest temperature. Performance differs considerably between units of the same type designation, but made by different manufacturers, and between individual units from one manufacturer. With a given FET the performance depends upon the capacity loading on its input and every effort is made (consistent with the application) to minimize this capacitance. Adjustment of the FET temperature for optimum temperature is also desirable.

The preamplifiers referred to in this paper use Texas Instrument 2N4416 FET's with the conditions optimized as stated above. The FET's which are now the best available for low capacity detectors, were selected and their performance would put them in the top 10% of the recent transistors of this type that we have measured. Lithium-drifted silicon detectors of our own manufacture are employed; they are 3mm thick 25mm^2 in area and exhibit measured capacities of about 1.3pF. They are made from silicon sufficiently free from trapping phenomena that no peak asymmetry is observed when tested with 1 MeV electrons at 77°K ($\approx$2.1 keV resolution). Consequently we can safely assume a complete lack of trapping effects on resolution for low energies. The detectors operate up to 800 volts and are used near to 77°K.

Work on FET's has overshadowed the smaller amount of effort expanded on testing high-valued resistors which are generally used as the dc feedback element in the charge-sensitive preamplifier configuration shown in fig. 1. The resistors degrade the system performance in several ways:

1) The resistor contributes normal thermal resistor noise. Operation at 77°K can reduce this type of noise but presently available resistors vary in a rather unpredictable manner with temperature. Many change value radically at low temperatures, and in some cases the changes are permanent.

2) In order to reduce the resistor noise contribution, it is necessary to use a very high value1). However, the optimum frequency band for the filter network required to give the best signal to noise ratio in these applications lies in the 100 KHz region. All high-valued resistors exhibit changes in both the resistive and reactive components of their impedance if the frequency exceeds a few KHz. The variation in the resistive part of the impedance for a typical (nominal) 1000 megohm resistor with frequency is shown in fig. 2.

The reduced value at high frequencies causes an increase in the noise contribution of this component*. The physical reasons for the changes in resistors with frequency are not clear, but distributed capacitance effects are certainly partly responsible and there are also likely to be end contact effects in some resistors. It is reasonable to suppose that the problems inherent in the high-resistivity films used to make these resistors will prohibit the manufacture of well-behaved high-valued resistors at high frequencies in the forseeable future.

3) Extraneous noise of an unexplained nature is known to be produced by these resistors$^{2)}$. This is not surprising as even low-valued carbon resistors are known to exhibit so-called current noise, and many of the high-valued resistors are made of pyrolitic carbon films.

4) By its very presence, the resistor increases stray capacity to ground on the input of the FET. We have seen earlier that minimizing this capacitance is important if the ultimate energy resolution is to be achieved, and therefore it is obvious that the presence of the resistor degrades performance.

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*It should also be noted that the variation in resistor value with frequency causes the response of the charge-sensitive stage of fig. 1 to depart from the single time-constant behaviour which might be anticipated. This results in problems at high counting-rates in all semiconductor detector spectrometer systems, since pole-zero cancellation methods of correcting for the response of the front-end stage are then inadequate. Consequently, opto-electronic feedback may be used to alleviate these problems too. This will be discussed in a later paper.

These considerations led to the thought of avoiding the resistor altogether. This could be accomplished by using the open-circuit gate approach of Elad[3)] but at the cost of losing the feedback considered necessary by nearly all workers in the field. The following section shows how the feedback can be accomplished while avoiding the use of high-valued resistors, by employing light as the feedback coupling element.

II. OPTO-ELECTRONIC FEEDBACK

Fig. 3 shows the arrangement used in the new preamplifier. If it is compared with the conventional arrangement of fig. 1, we see that the light coupling replaces the feedback resistor R_f. Fortunately, the recently available light-emitting diodes give light intensities almost proportional to the current through the diode[*]. This means that the relationships shown in fig. 3 apply, assuming that the resistance R_s is much larger than the incremental impedance of the light-emitting diode over its operating range. The parameter ϕ represents a coupling coefficient involved in the light production and coupling process. It is easily adjustable over a wide range of values by suitably positioning the light source with reference to the photodiode, by the use of absorbers or light pipes and by other methods.

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[*] We have tested a number of diode types including GaAs diodes which emit light in the infra-red region and Ga-As-P diodes and GaP diodes emitting in the red region of the spectrum. None obeyed a linear law when the light output was plotted against current, but a power law (Light $\propto I^n$) with $1 < n < 2$ applied in all cases. Over a limited range of currents an approximate linear law is assumed in this paper.

In fig. 3, the current into the FET gate from the photodiode is given by: $i = V_o/R_f'$ where $R_f' = R_s/\emptyset$.

The light feedback therefore acts like a feedback resistor. A typical value of R_s might be 100 ohms while $\emptyset$ can easily be adjusted over the range from 10^{-6} to 10^{-10}, giving values of R_f' ranging from 10^8 to 10^{12} ohms. Moreover, the light coupling is practically instantaneous, the response of both diodes is measured in nanoseconds and the value of R_f' is therefore independant of frequency, at least up to many MHz.

In the arrangement shown in fig. 3, the photodiode adds capacity to the input gate, and in this respect it suffers from the same disadvantage as the resistor. In practice the role of the photodiode can be assumed by the drain-gate junction diode in the FET* thereby avoiding this disadvantage in low-energy systems. Use of the FET in this manner necessitates opening its case so we have taken the opportunity to completely remove the FET from its case and to mount it in an assembly which reduces capacity on the gate lead. Furthermore, it has proven possible to incorporate the feedback capacitor as an integral part of the FET mounting thereby further reducing stray capacity to ground. Mechanical details of this assembly will be described later in the paper.

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* While good diodes exhibit excellent frequency response to light, we observe a long time-constant component in the FET response, particularly to IR light. This may be attributed to diffusion of charge from the bulk semiconductor. For low-energy systems this is of no importance, but if this technique is used to replace resistors in high-energy systems, it is possible that a red-emitting light must be used and/or a well-behaved photodiode.

The arrangement of fig. 3 may be thought of as a servo loop which maintains a mean current in the photodiode equal to the leakage current of the silicon radiation detector thereby stabilizing the dc operating point of the stage. However, statistical fluctuations in the detector leakage and those in the photodiode are not correlated so the total noise is just $\sqrt{2}$ times that which would arise from detector leakage alone. This constitutes the shunt noise in the input circuit while the FET channel noise is the principal series noise source. Calculations and tests show that all other noise sources can be made to contribute only very minor amounts of noise.

During tests of the new circuit, we observed that many FET's produce excess noise if the drain-gate voltage exceeds a few volts. The effect is also observed if the gate current is measured as a function of drain voltage with the FET operated in its normal mode i.e. not with source disconnected as is usually done in measurements of gate leakage, but with current flowing through the FET. Here a rapid increase in the gate current at drain voltages greater than a certain value is seen. We attribute this to the onset of an avalanche process in the channel with collection of the holes at the gate. It is analogous to positive ion collection at the grid of a vacuum tube. As the avalanche voltage exhibits a positive temperature coefficient, the critical voltage becomes smaller as the temperature is reduced, so the the anomalous situation can arise whereby the gate current increases at low temperatures. This applies only if the drain voltage exceeds a certain value and leads to the optimum temperature for noise being dependant on drain voltage. These considerations indicate that the drain voltage should always be lower than the critical value but not too low, as this would increase the drain-gate capacitance. These factors, which have not previously been appreciated, have led to some confusion in earlier FET testing for preamplifier applications.

The complete preamplifier used in the tests described in the following paragraphs is shown in fig. 4. Its design is in most respects similar to a conventional charge-sensitive unit employing a cascode stage following the FET. The output from the charge sensitive loop feeds the light-emitting diode whose light is directed onto the FET. A light-emitting diode is also pointed at the silicon detector to permit adjustment of the detector leakage. This is only essential if the drain-gate leakage exceeds the detector leakage (an unusual situation), but it is useful to be able to vary the detector leakage in tests on the system. We should note that very effective light shielding between the feedback light and the detector must be provided to prevent the loop exhibiting overall positive feedback as the large size of the detector makes it much more light sensitive than the FET.

III. EXPERIMENTAL RESULTS

A meaningful comparison between the new preamplifier and a conventional one can best be made by measuring the noise in the two systems as a function of the time-constant of an equal RC integrator-differentiator filter. If the $\overline{\text{noise}^2}$ is plotted versus the time-constant on a log-log plot, a straight line relationship is expected at short times where series input circuit noise is dominant, and again a long times where shunt noise is dominant. Such a plot is given in fig. 5. The FET's used in the two systems had previously been measured as having nearly equal performance in a conventional preamplifier. We can therefore regard the two curves in fig. 5 as representing a fair comparison between the two preamplifiers. Inspection of fig. 5 shows that the two systems give similar results at short time-constants although the reduced stray capacity in the light coupled system makes its performance superior. At long time constants, however, the superiority of the light coupled system is very obvious. Moreover, the slope of the line for the new preamplifier at short time constants is correct.

In order to make a valid comparison of the two systems in actual spectrum measurements, some decision must be made as to the type of pulse shaping to be used. We have chosen to use a Gaussian shaper which is part of our high-rate amplifier system4). For the conventional preamplifier the peaking time of the Gaussian was 2.25 microseconds, which is nearly optimum for this system. For the new preamplifier we have used a peaking time of 7 microseconds; although this is shorter than the optimum time, and improved results could be realised using a larger value, it was felt that this compromise between good noise performance and good rate performance would be fairly representative of normal use. The user can, of course, choose his own compromise depending on circumstances.

The comparison in performance between the two units for Mn and Np X-rays is shown in figs. 6 and 7. We need hardly add words to the picture conveyed in these figures! However, we should point out that many pulse-height analyzers will not permit good measurements on such long pulses and modifications to the analyzer may be necessary to realise these results.

It is unfortunate that the improved electronic resolution is not fully seen on the X-ray peaks of figs. 6 and 7 as the statistics of charge production in the detector are quite significant at even these low energies with such a good electronic system. Fig. 8 shows the calculated effect of detector statistics on systems exhibiting electronic resolutions of 125, 150, 175 and 200 eV. Electronic measurements show that the fundamental resolution of our present system is about 150 eV; it becomes about 192 eV at the 6 keV of the Mn line, and about 260 eV at the 20 keV of the main Np lines. Fig. 8 assumes a Fano factor of 0.12, a value which is subject to some question but which represents a fair estimate based on present experimental data obtained at much higher energies (1 MeV electrons). A number of measured resolutions for this X-ray system are shown in fig. 8 and these points are seen to fit reasonably well on the curve. This indicates that the assumed Fano factor 0.12 cannot be much in error. The system should permit an accurate determination of the Fano factor for silicon and we intend to do this work in the near future.

These numbers suggest that further improvements in the electronics will do little to improve experiments except at very low energies, and therefore for elements of low atomic number. Of course, this is an area of great potential interest as it is the region of the dominant constituents of biological material. Furthermore, we should remember that electronic noise not only limits the energy resolution but also limits the minimum energy that can be detected. A FWHM resolution of 150 eV indicates that noise counts over 300 to 400 eV are rare; therefore X-rays of nitrogen and possibly carbon should be detectable. Of course the problem of getting them into the detector must be faced before this can be done.

A few words must be said on the counting-rate performance of the unit. The design uses a value of 10^{-9} for the light coupling coefficient $\emptyset$ and the maximum light-diode current is 50mA. The maximum permitted detector current (average) is therefore 5×10^{-11}A which corresponds to a counting-rate of about 3×10^4 counts/sec. of 30 keV events. Actually the random nature of the counts reduces this rate to under 10^4 counts/sec. which is quite adequate for most X-ray applications. If the coupling coefficient $\emptyset$ is increased, a higher counting rate can be accomodated but the standing current in the light-source must be reduced accordingly. This should be avoided if possible as the light output becomes a more non-linear function of current at low currents. The minimum current level can be increased by using the light pointed at the detector to increase its leakage - but this gives increased noise, which may be tolerable in some circumstances.

As the counting-rate is limited mainly by the dc shift in the preamplifier, pulse-shaping of longer time constants is quite practicable. Measurements using a Gaussian shape peaking at 15 μsec. have given resolution of less than 140 eV FWHM in this system.

IV. MECHANICAL DESIGN

The front end of the system is mounted on a standard "dipstick" type of liquid nitrogen cooling dewar. This type of dewar was preferred over the LN tube cooling system due to its lower microphonics. Fig. 9 shows the mechanical arrangement of the front-end. A prime consideration is to maintain the lowest possible temperature on the detector since its leakage current is a dominant noise factor at long time-constants. Boron nitride is used as the electrical insulator for the detector voltage because its good thermal conductivity reduces losses in the cooling path to the detector. The two light sources are mounted directly on the cold member - this is possible as their efficiency increases at low temperatures - and lucite light-pipes convey the light to the detector and FET. The available light far exceeds our requirements so the surfaces of the lucite pipes are painted black to absorb light and the ends facing the detector and FET are lapped to a matte finish. As mentioned earlier, light shields are also provided to restrict the light to its intended paths.

Removal of the FET from its metal case has proved to be no problem. The FET chip, which mounts directly on the gate lead in the T.I. 2N4416, is removed from the metal can with the source and drain gold wires still attached. It is then mounted in the boron nitride mount shown in Fig. 9. Boron nitride was chosen for this mount due to its good thermal conductivity, which is necessary if adequate cooling of the FET is to be achieved. A wire firmly attached to the top of the mount provides a 0.2 pF capacity to the gate lead, and this is used as the feedback capacitor. The FET mount is mechanically supported from the cold member by a stainless steel tube which gives a controlled thermal resistance in the cooling path for the FET. A heater contained in the FET end of the tube is used to vary the FET temperature. The possibility of using a chip version of the FET was considered. However, the ability to test and select FET's before removing from the case seemed important. The final mount is an integral one which permits testing of the FET before inserting it in the holder.

Some simplifications in this design are expected after sufficient measurements have been made to show that certain items can be omitted without loss in performance. A slight reduction in stray capacitance, with consequent improvement in resolution, can probably be accomplished by the removal of insulating material where experience shows it serves no useful purpose.

V. CONCLUSION

The use of opto-electronic feedback appears to result in a very significant improvement in the energy resolution capabilities of semiconductor detector spectrometers used at low energies. While the system looks complicated, our experience is that the assembly and setting-up of the system present no difficulties. In fact they involve less difficulty than selecting high-valued resistors for use in high-resolution systems. The ultimate in performance still requires excellent FET's but performance better than that of conventional preamplifiers can be achieved with very little FET selection. The saving of effort due to this could well be more important than the improved ultimate performance capability.

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Several co-workers have given freely of their advice and services in this project. We are particularly indebted to W. L. Hansen who contributed to the ideas which led to the thought of using a light source and photodiode as a feedback element. J. Anderson and D. Clark built much of the hardware and carried out some of the tests. S. Wright, D. Landis, E. Rivet and R. Pehl were all responsible for helping in certain phases of the work and R. Lothrop and H. Sommer made the silicon detectors used in the tests.

FOOTNOTE AND REFERENCES

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2. V. Radeka, State of the Art of Low-Noise Amplifiers for Semiconductor Radiation Detectors, Brookhaven National Laboratory, BNL-12748 (1968).

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FIGURE CAPTIONS

Fig. 1. Block diagram of a conventional charge-sensitive preamplifier.

Fig. 2. Frequency effects in high-valued resistors.

Fig. 3. Block diagram of opto-electronic feedback preamplifier.

Fig. 4. Complete schematic of the preamplifier.

Fig. 5. Comparison of time-constant behaviour of the two types of preamplifier. Measured with a pulse-shaper containing equal RC integrator and differentiator.

Fig. 6. a) Mn X-rays - performance of new unit.
b) Mn X-rays - performance of conventional preamplifier.

Fig. 7. a) Np X-rays - performance of new unit.
b) Np X-rays - performance of conventional preamplifier
The results in figs. 6 (a) and 7 (a) were obtained using a Gaussian pulse-shaper peaking at about 7 microseconds while those of figs. 6 (b) and 7 (b) used a Gaussian peaked at 2.5 μsec. (nearly optimum for this system).

Fig. 8. The behaviour of resolution with energy due to combined electronic noise and detector charge statistics. The Fano factor is assumed to be 0.12. Some experimental measured points on the new system are shown.

Fig. 9. Mechanical details of the front-end assembly.

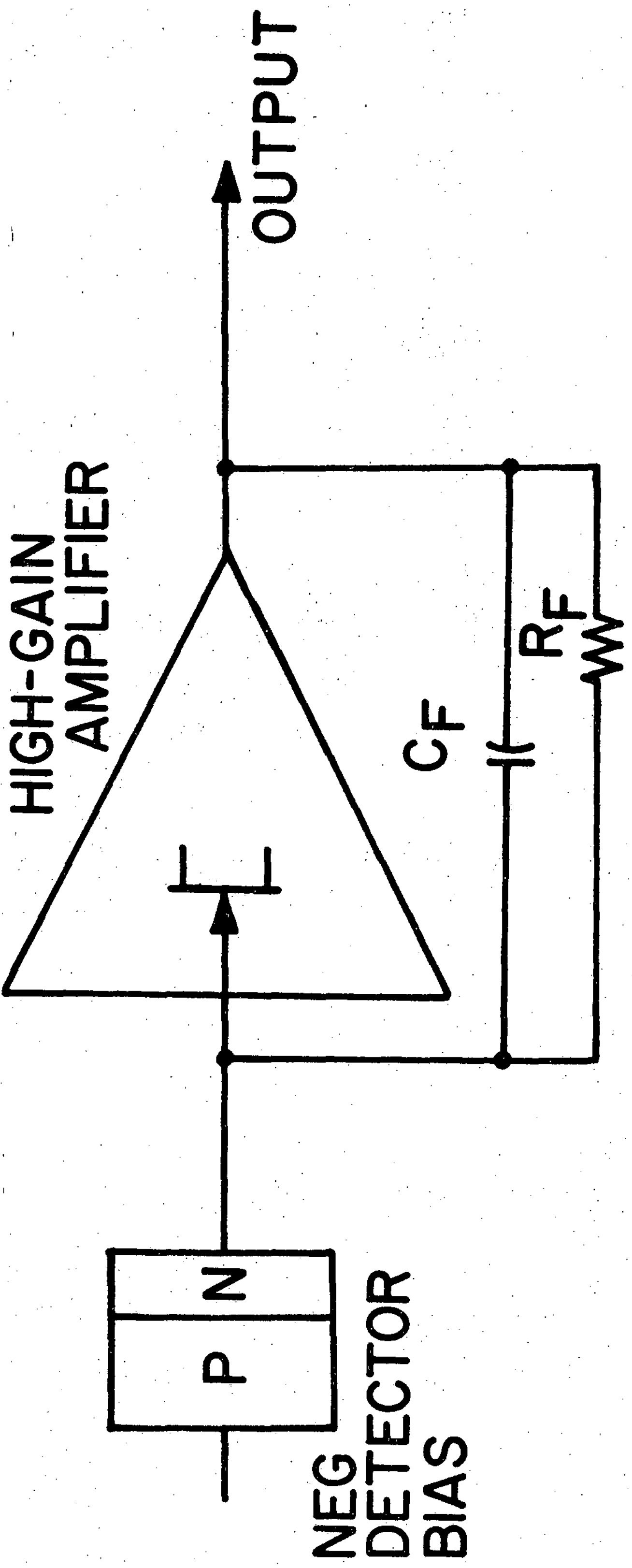

FIG.1:

XBL 692-169

XBL 692-170

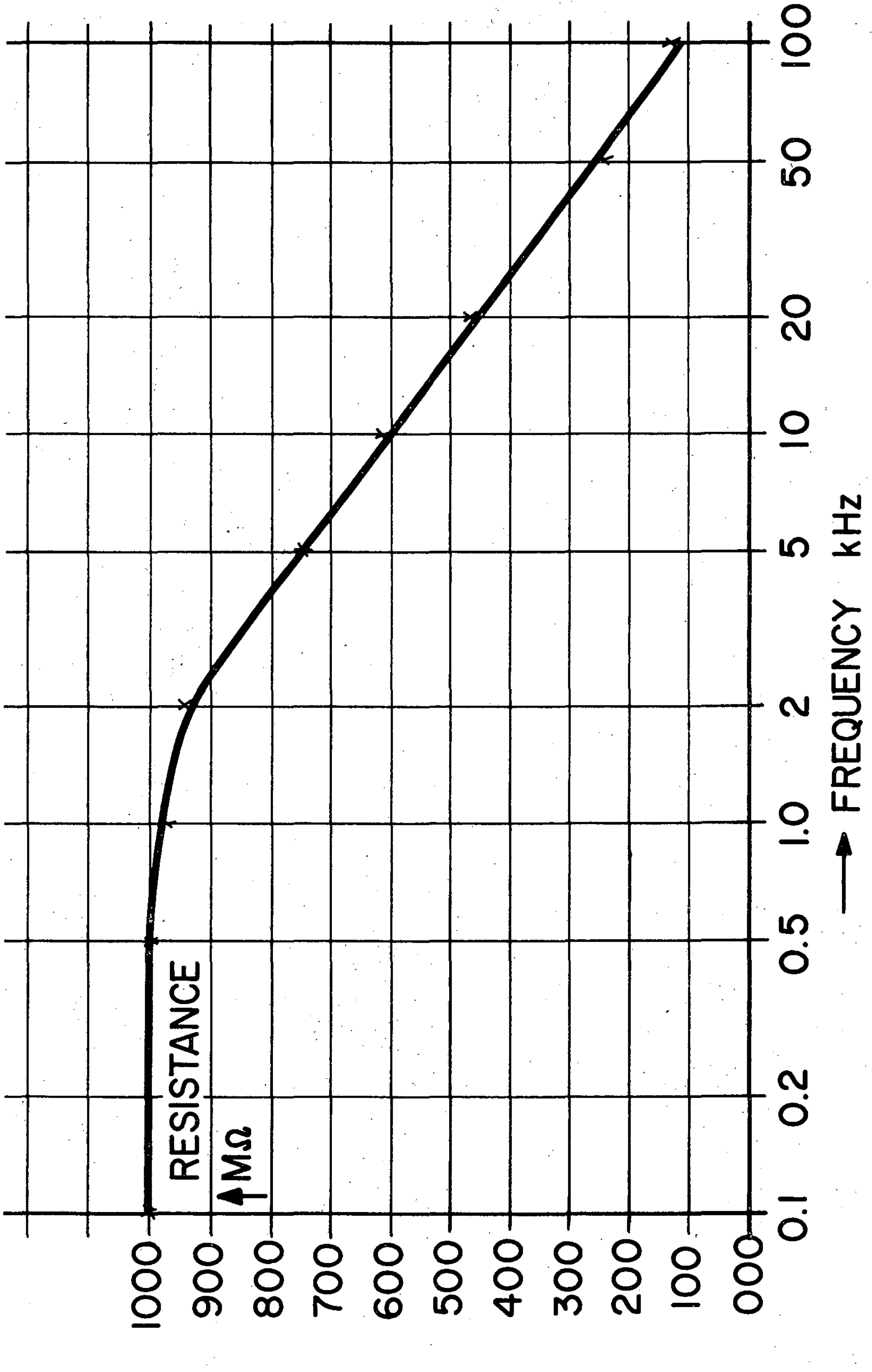

FIG. 2

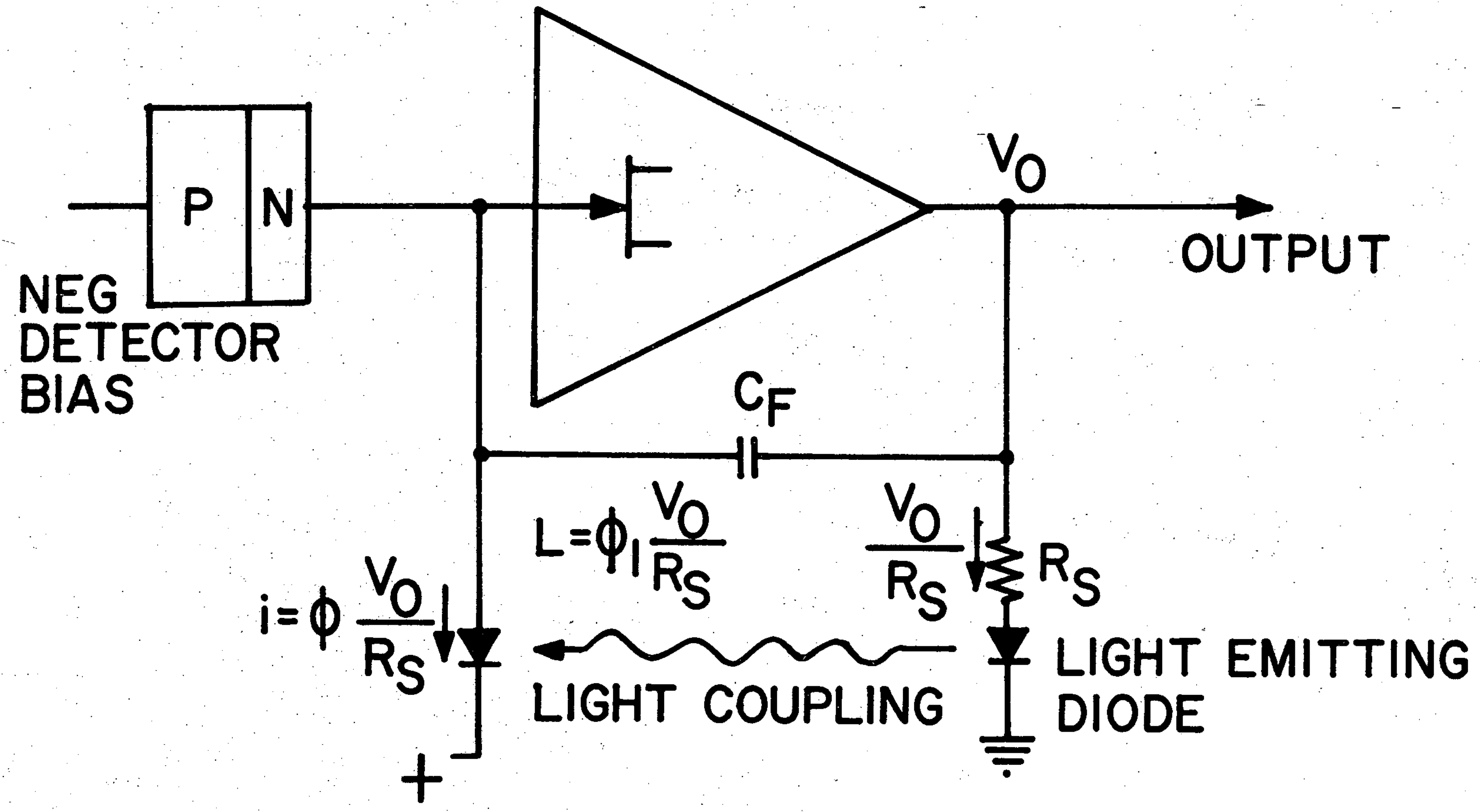

FIG. 3:

XBL 692-171

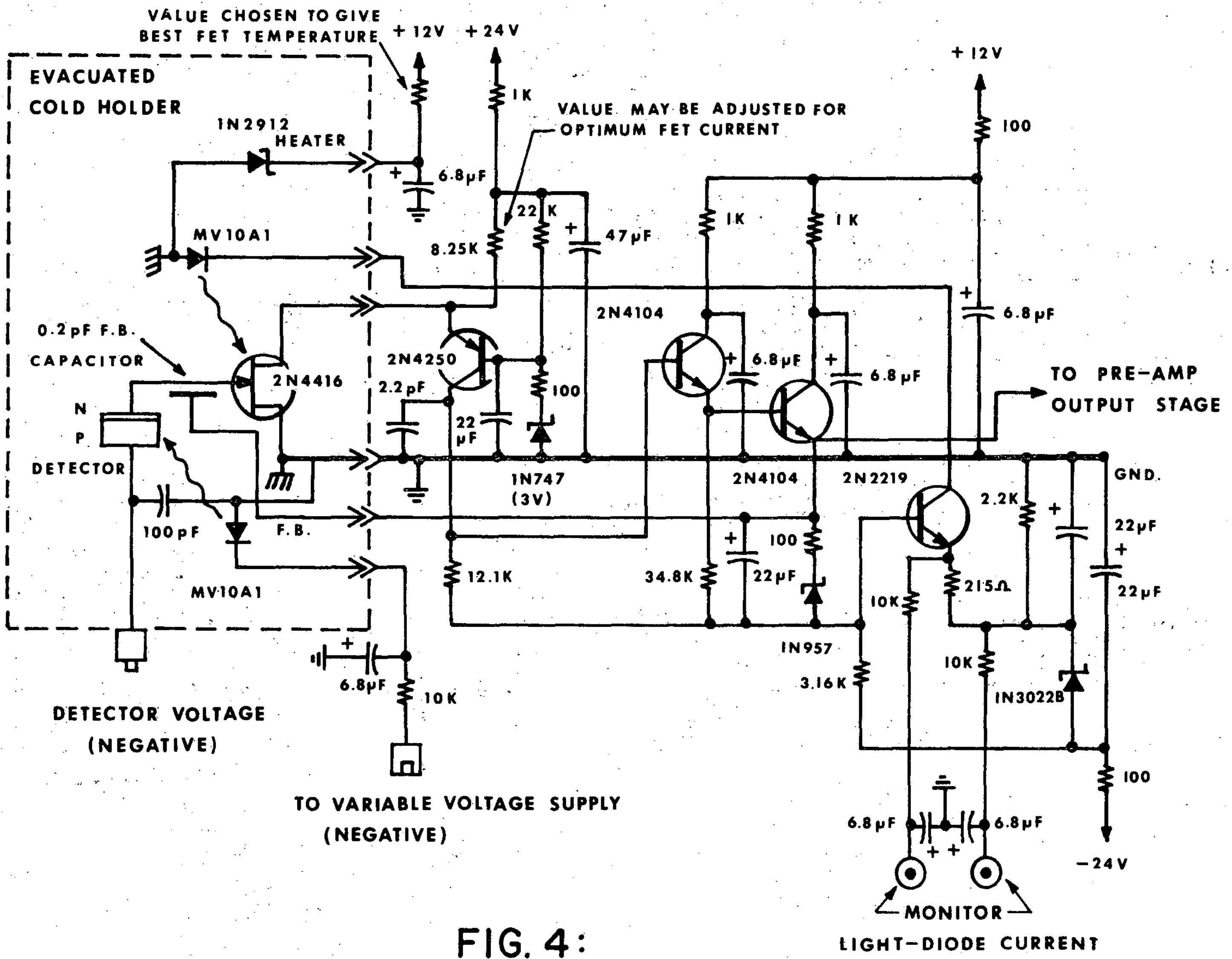

FIG. 4:

XBL 692-172

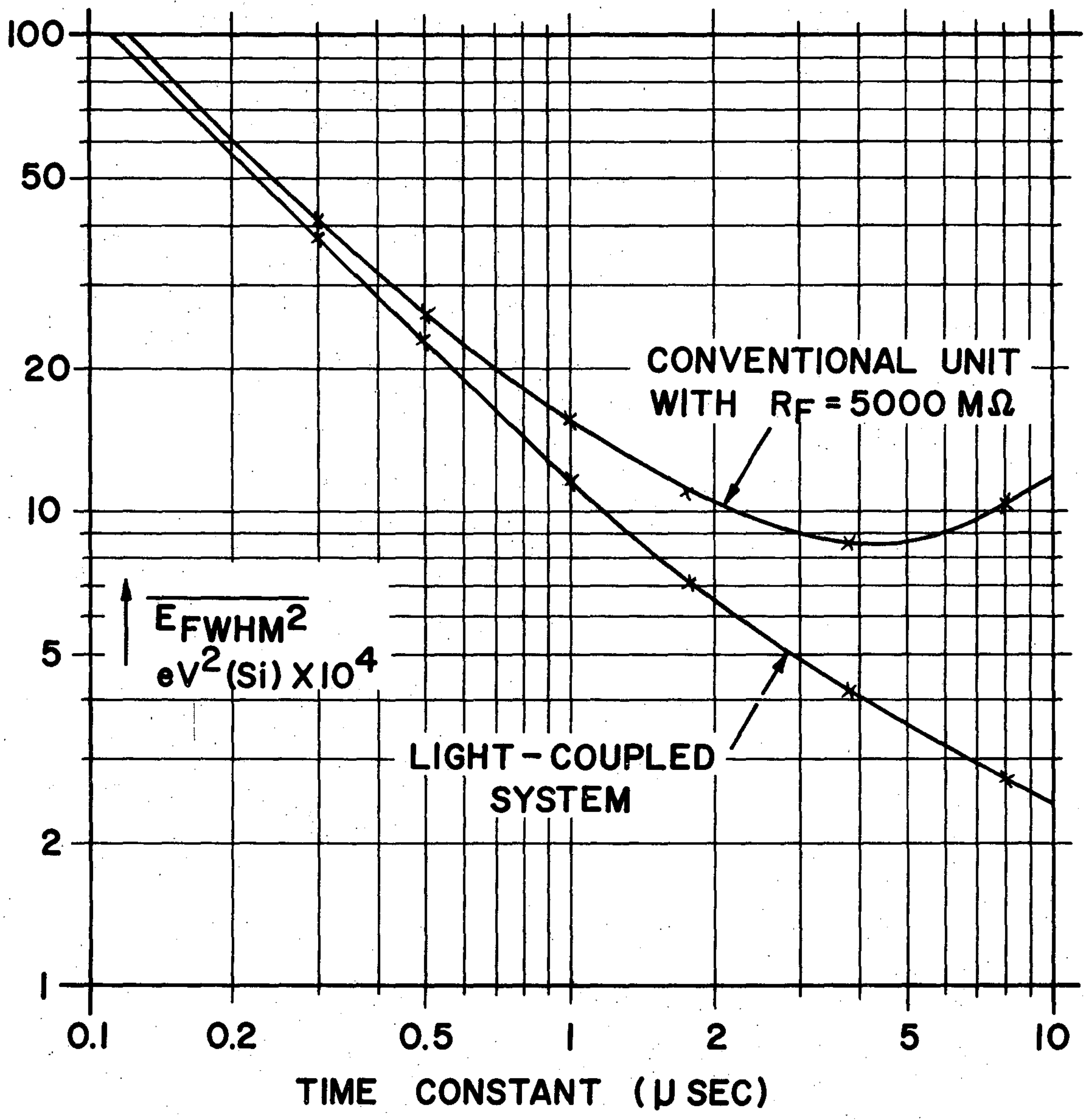

FIG. 5:

XBL 692-173

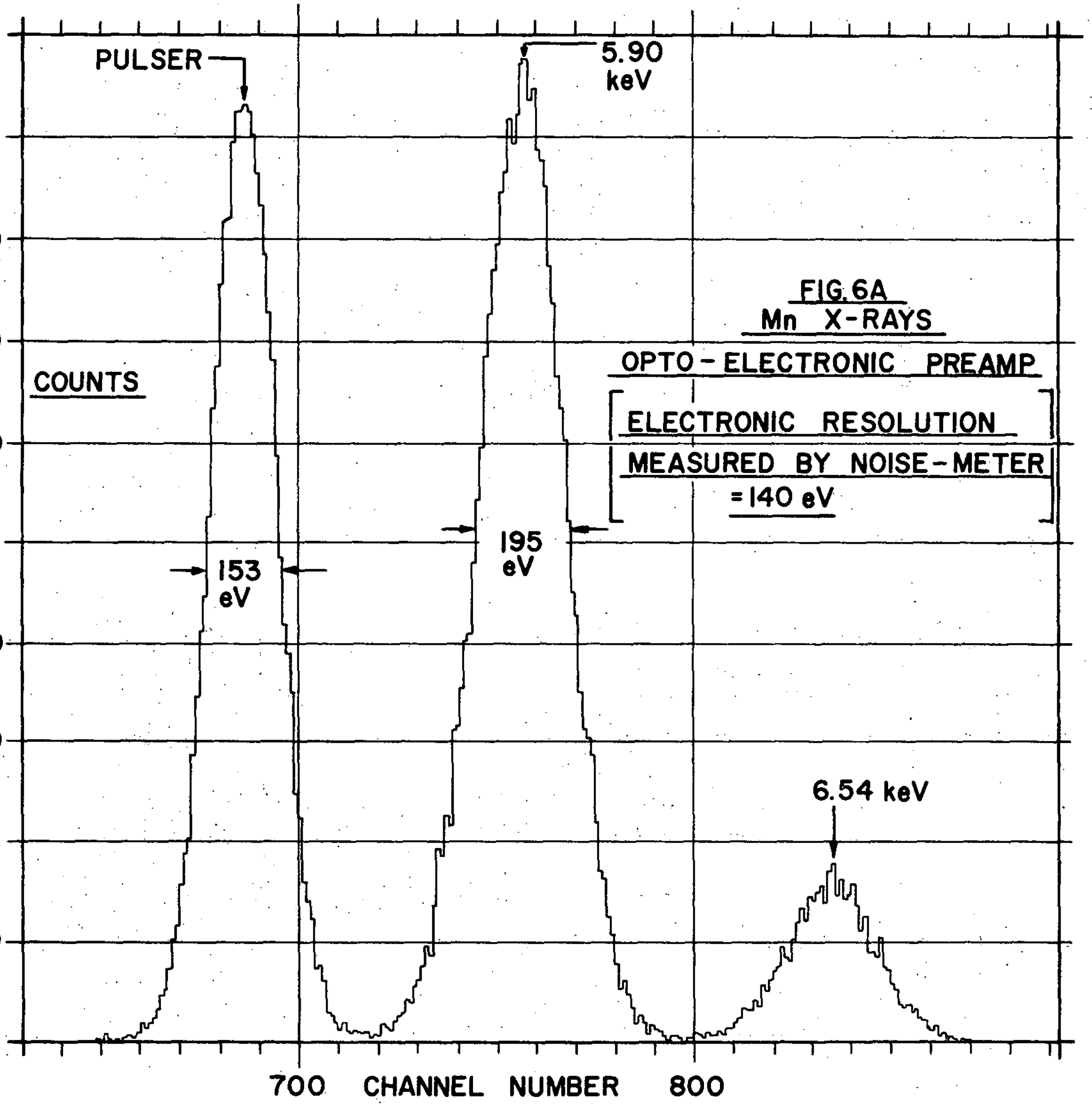

FIG. 6A
Mn X-RAYS
OPTO-ELECTRONIC PREAMP
[ELECTRONIC RESOLUTION
MEASURED BY NOISE-METER
= 140 eV]

XBL 692-174

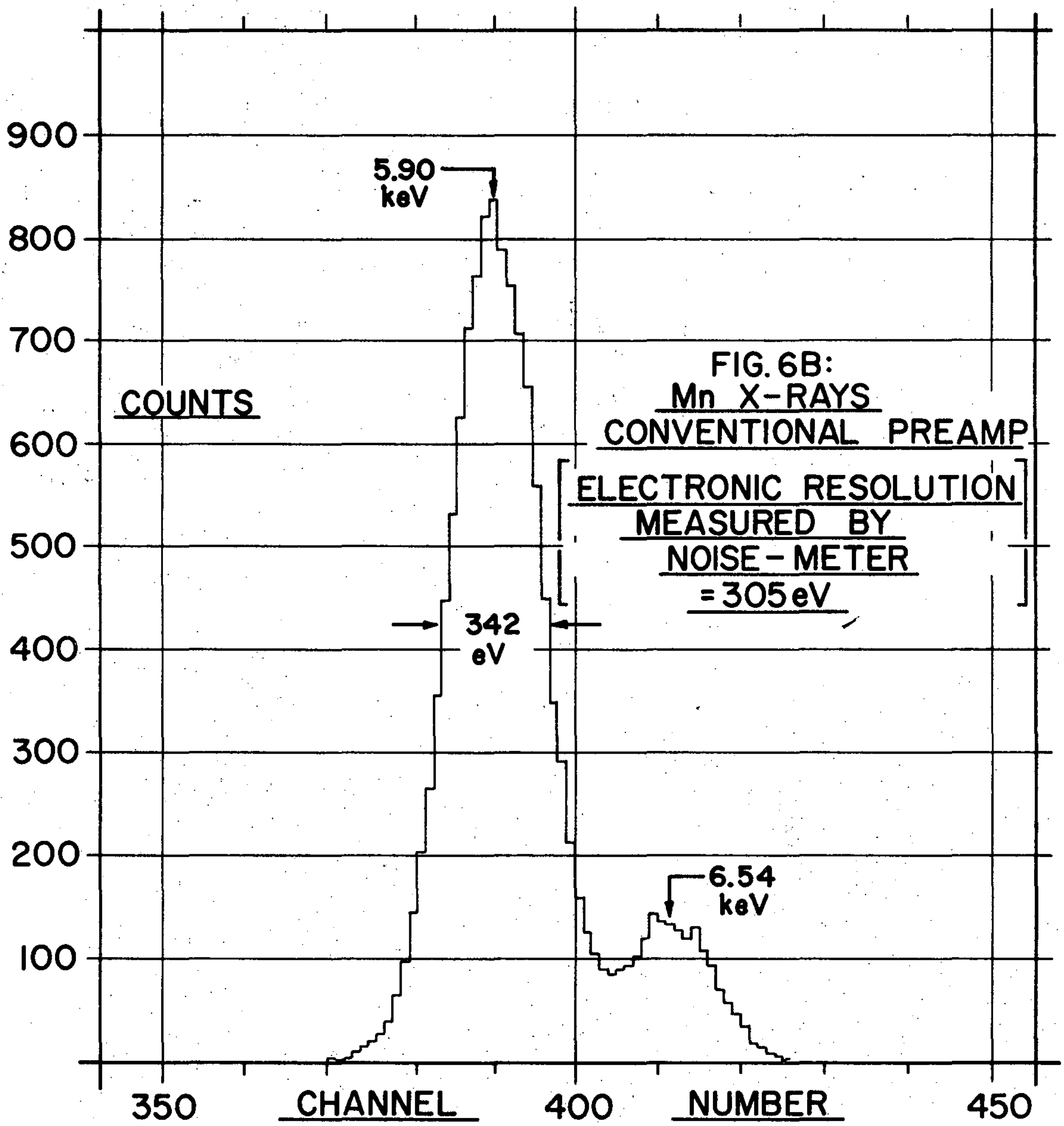

FIG. 6B: Mn X-RAYS CONVENTIONAL PREAMP

[ELECTRONIC RESOLUTION MEASURED BY NOISE-METER =305 eV]

XBL 692-175

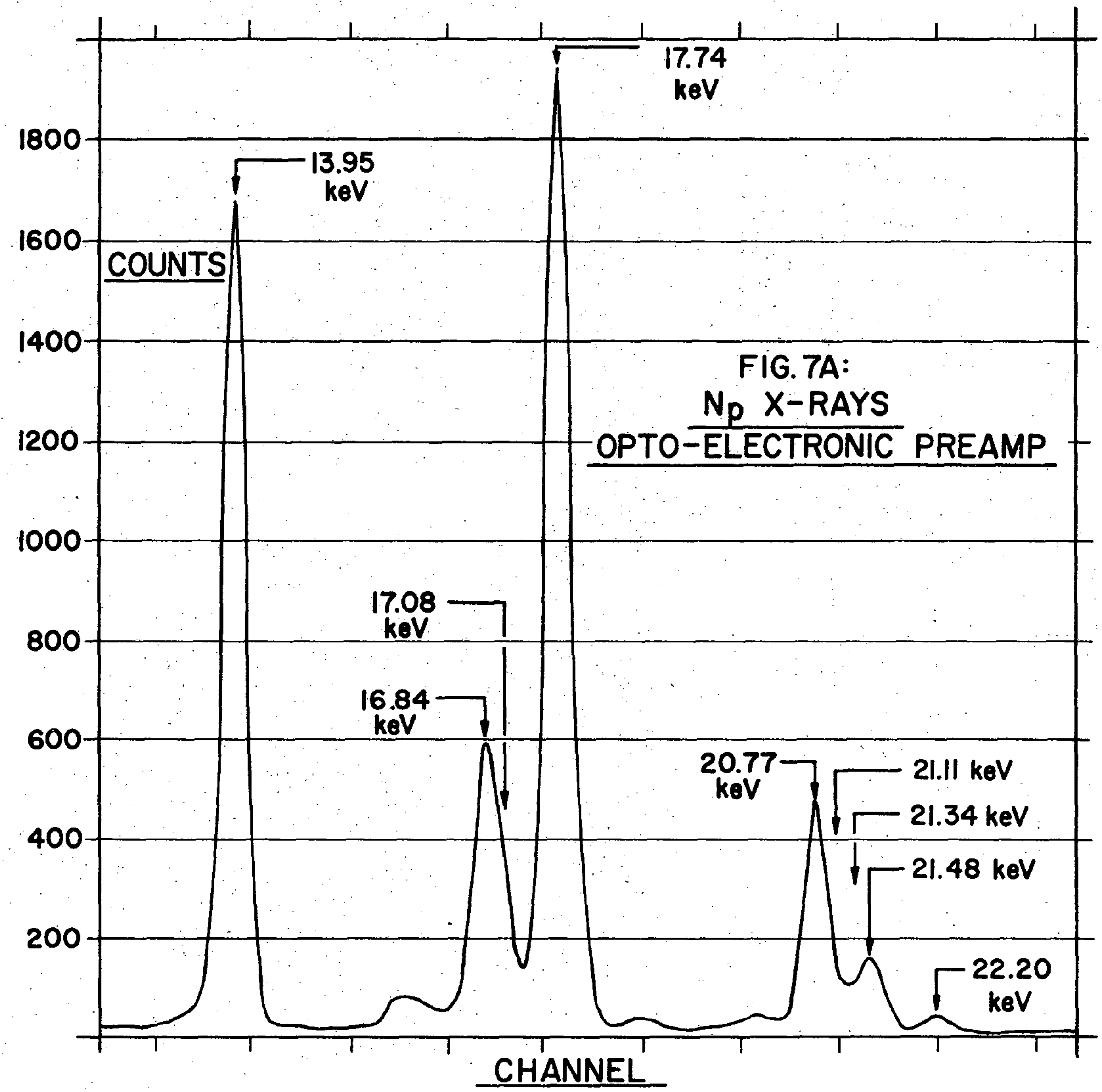

FIG. 7A: Np X-RAYS OPTO-ELECTRONIC PREAMP

XBL 692-176

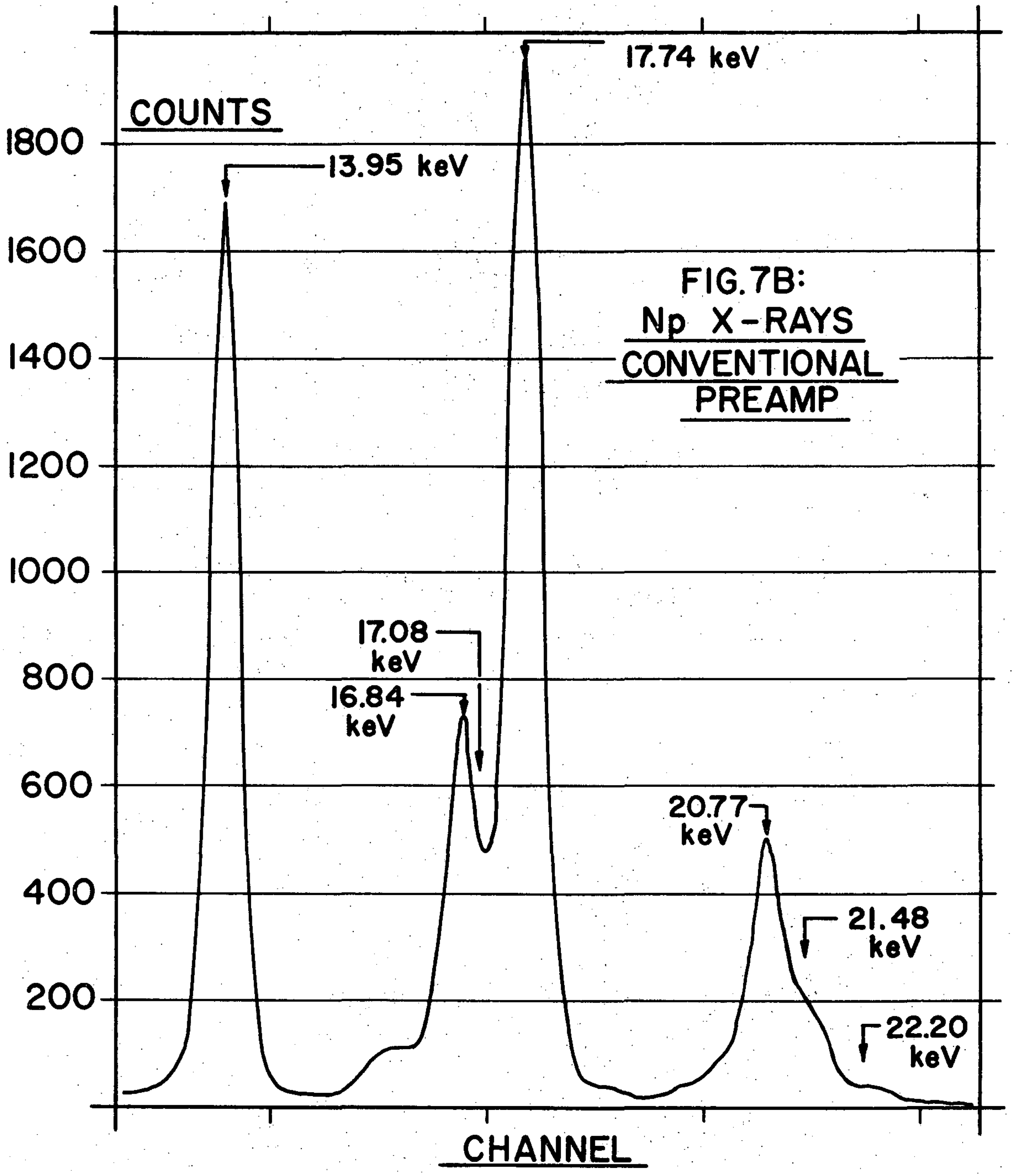

FIG. 7B: Np X-RAYS CONVENTIONAL PREAMP

XBL 692-177

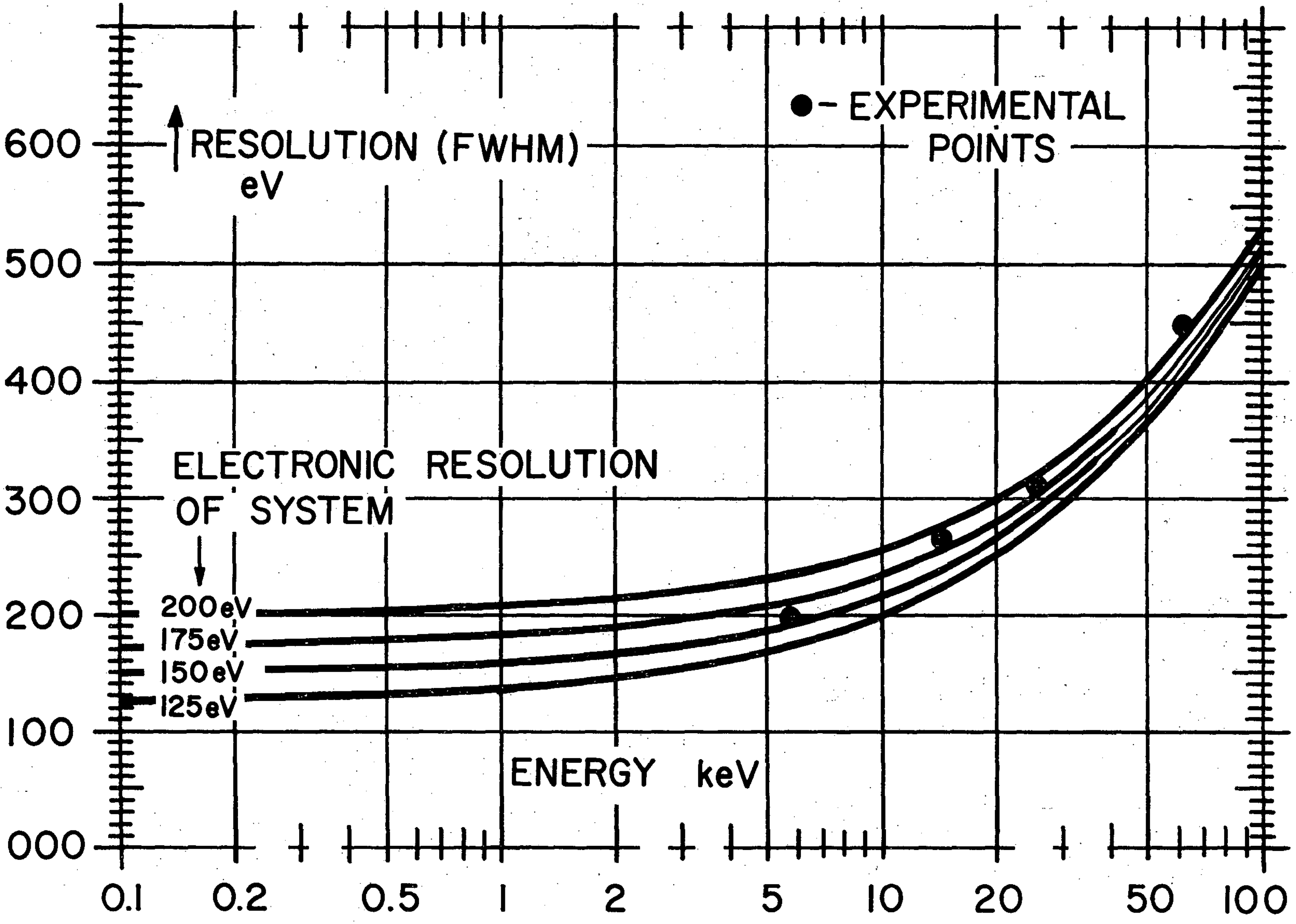

FIG. 8:

XBL 692-178

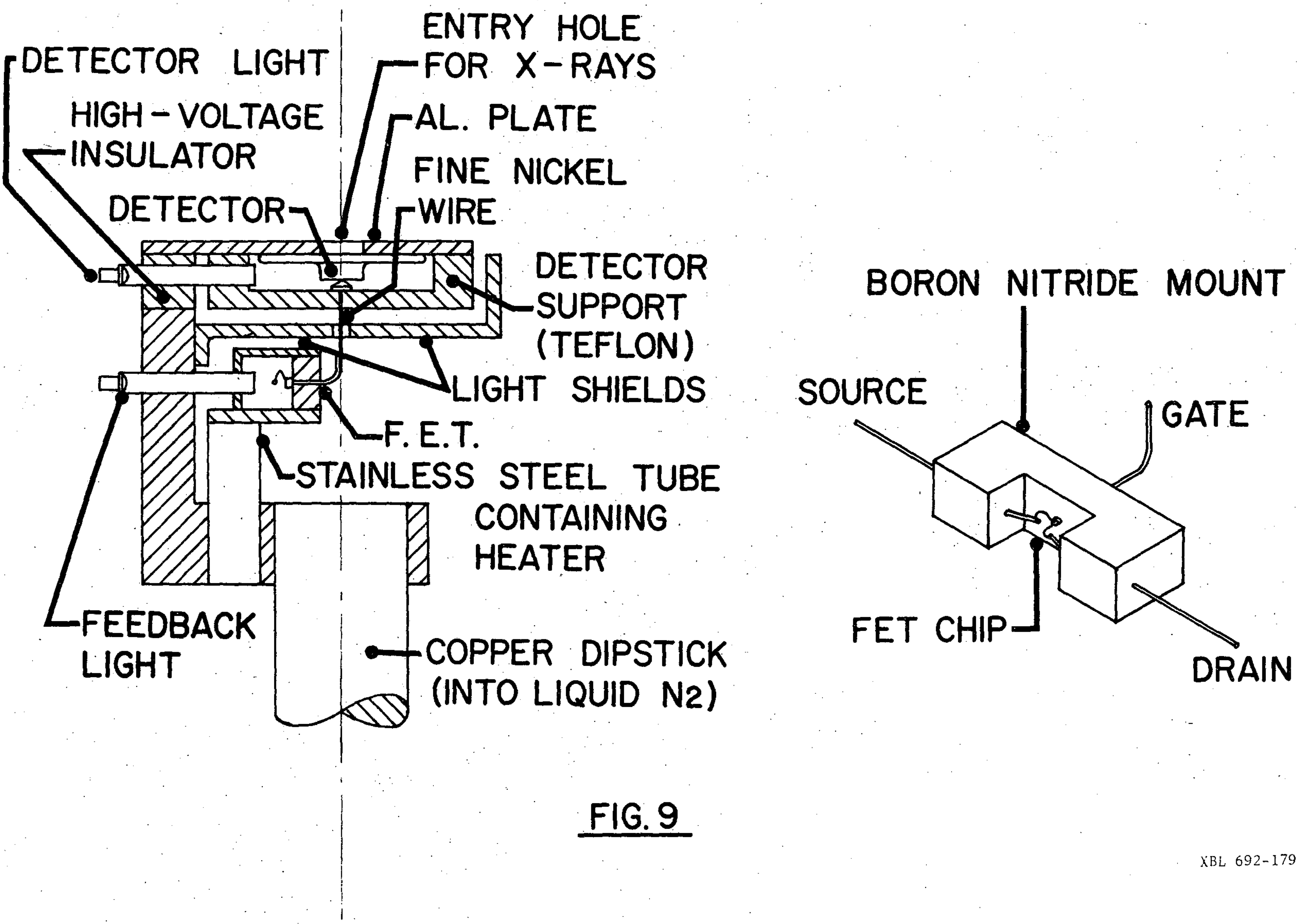

FIG. 9

XBL 692-179

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