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Model of a Fission Counter System for Optimizing Performance at High Gamma Dose Rates

V. K. Paré

OAK RIDGE NATIONAL LABORATORY
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MODEL OF A FISSION COUNTER SYSTEM FOR OPTIMIZING PERFORMANCE AT HIGH GAMMA DOSE RATES

V. K. Paré

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November, 1978

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HIGHLIGHTS

MODEL OF A FISSION COUNTER SYSTEM FOR OPTIMIZING PERFORMANCE AT HIGH GAMMA DOSE RATES

V. K. Paré

A mathematical model has been developed for predicting the ability of a fission counter to operate at a high gamma dose rate. There are several potential applications in reactor programs for detectors that can accurately measure neutron fluxes in the range 0.1 to 1.0 nv(th) while subjected to gamma dose rates of 10^4 to more than 10^6 R/hr. Fission counters for such service have up to now been designed on the basis of laborious testing of prototypes combined with extrapolation from existing designs. The objective of this work is to make it possible to arrive at an optimum design without expensive testing.

In this model, the complete system of counter, cables, preamplifier, main amplifier, filters, and discriminator is considered, including the effects of interaction of counter capacitance with cable and preamplifier impedance. Electrical noise theory is used to calculate the frequency spectrum of the interfering signal generated by pileup of gamma pulses, to add to it a contribution from electronic noise, and to obtain the rate at which the total noise signal triggers the discriminator. The shapes and widths of neutron and gamma pulses are calculated from the electron drift velocity and electrode spacing. All frequency response functions — both inherent and designed-in — in the counter and electronics channel are included in the computation of the neutron pulse heights and noise spectra at the input to the discriminator. The final result is a figure of merit, which is the ratio of the median neutron pulse height to the discriminator setting needed to limit to an acceptable level the false triggering caused by gamma pileup and electronic noise. The calculation relies on experimental data for quantities such as the average ionization per event, the gamma interaction probability, and the spectral density of electronic noise. Simple scaling laws are used to estimate the approximate variation of these quantities with the design parameters of the counter.

A subsequent report will describe the determination of experimental parameters from and comparison of the model's predictions with results from measurements with a test fission counter.

I. INTRODUCTION

There are several potential applications in reactor programs for sensitive neutron detectors that must operate in the pulse-counting mode in regions of high gamma activity. In-vessel low-level flux monitors (LLFMs), such as those to be installed in the Fast Flux Test Facility (FFTF) reactor, require a sensitivity of $\sim$1.0 count sec^{-1} $[nv(th)]^{-1}$ at $>10^6$ R/hr. In other liquid-metal, fast breeder reactor applications, the anticipated gamma dose rate is "only" 10^4 to 10^5 R/hr, but the counting sensitivity must be at least 10 counts sec^{-1} $[nv(th)]^{-1}$. These high-sensitivity applications could include failed-fuel monitors that detect delayed neutrons emitted from fission product precursors in the coolant stream, ex-vessel LLFMs (as proposed for the Clinch River Breeder Reactor), and criticality and assay monitors for fuel reprocessing plants.

Fission counters are unique in their ability to count neutrons at high gamma dose rates; nevertheless, they require careful design and optimization to meet the requirements described above. The capability of a fission counter to reject gamma background depends on design parameters of the electrodes and filling gas and also on the parameters of the electronics channel to which the counter is connected. Fission counters have been successfully designed for the FFTF LLFM operating conditions,[1-4] and the basic theory of the gamma irradiation effects has been worked out.[5]

A counter system model containing many of the elements of the present one has been used to study the effects of filtering in the counting channel and to estimate the performance of high-sensitivity fission counters.[6,7] Nevertheless, there has remained a need for a comprehensive systems engineering approach; that is, a calculation procedure that generates theoretically valid and internally consistent pulse shapes and noise spectra for any design configuration and enables a designer to specify all of the relevant parameters of a counter-channel system so as to obtain a quantitative prediction of its gamma rejection capability. The objective of this work was to make such a systems approach available for the first time. In this paper we will first specify the model of the system on which the analysis is based, summarize the basic theory involved, and give an overview of the procedure. Then the procedure will be described in detail, and its implementation on a computer will be discussed.

Much more is involved than constructing the model and developing the computer programs. It is also necessary to validate the model and to obtain some of the numerical parameters empirically by matching the calculated results with data on an actual fission counter system. For this purpose a test fission counter system was built at Oak Ridge National Laboratory (ORNL). In this system many design variables, including electrode spacing, can be varied easily so that one can evaluate their effects, make comparisons with theory, and develop scaling laws. This work has been completed and the results will be described in a forthcoming paper. This paper is concerned with outlining the basic theory of the model and the sequence of calculation, and with discussing some aspects of the calculation that must be understood clearly if valid results are to be obtained.

II. MODEL

We wish to study the performance of a neutron detection system consisting of a fission counter, a connecting cable, a preamplifier, main amplifiers, filters, and an integral discriminator, as shown in Fig. 1.

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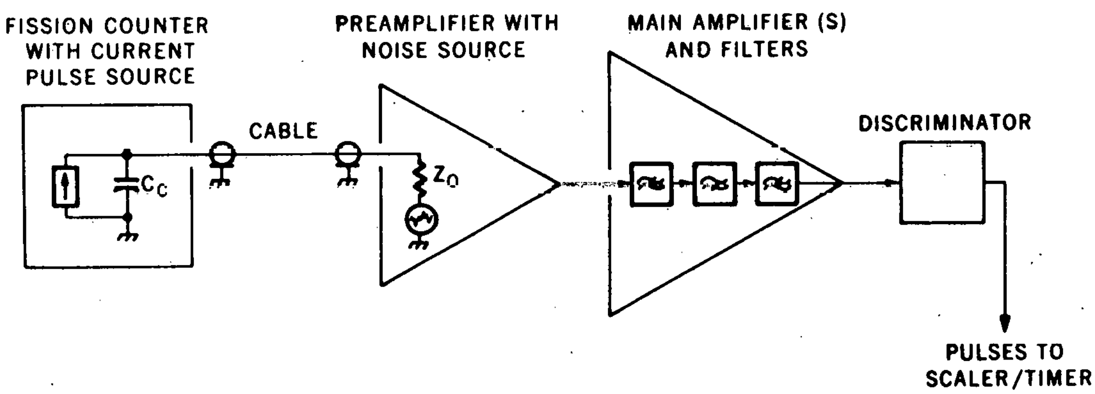

Fig. 1. Configuration of neutron detection system.

The counter is operated in the electron-collection, current-pulse mode, and the input impedance, Z_0, of the preamplifier[8] matches the characteristic impedance of the cable. The counter is exposed to simultaneous neutron and gamma radiation; in the intended operating regime the neutron (fission fragment) pulses occur at a sufficiently low rate so that they rarely overlap and are thus processed individually by the system. The gamma (secondary electron) pulses are two or three orders of magnitude smaller than the neutron pulses and could easily be rejected by the discriminator if they occurred individually. However, in the applications toward which this analysis is directed, the gamma pulse rate will be of the order of 10^{11} to 10^{12} pulses/sec, and multiple overlap will occur.[9] The result will be a fluctuating current which will be called gamma pileup noise. Because of the high multiplicity (10^4 to 10^5) of overlap, some of the peaks of this noise current are as high as typical neutron pulses; consequently, false triggering of the discriminator occurs. The rate of triggering by gamma pileup noise can be reduced by raising the discriminator setting, but a serious loss of neutron counting sensitivity may result. The objective of design optimization is to minimize this loss.

The analysis must also include the inherent electronic noise generated in the input stage of the preamplifier, since experiments have shown that even at high gamma dose rates it contributes appreciably to the fluctuating signal at the input to the discriminator.[1] Since the frequency spectrum of preamplifier noise may differ greatly from that of gamma pileup noise, the presence of the former may have important effects on the values of optimum design parameters. A further possible source of noise is pileup of alpha pulses occurring in the fission counter. This effect is negligible in fission counters containing $\sim$2 g of electromagnetically enriched uranium with <0.3% ^{234}U; however, it may have to be considered in the design of a high-sensitivity fission counter which might contain 20 g or more of the same material. In the present analysis, alpha pileup will be neglected; however, its inclusion would be straightforward.

All of these signals pass through various filtering elements in the system before they reach the discriminator input. Typically, low-pass ("integrating") and high-pass ("differentiating") filters are added by design, usually in conjunction with the main amplifiers, to control the bandwidth of the system. If these filters are not present or are not the only limiting factors, it may be necessary to take into account the frequency response characteristics of the amplifiers themselves. As illustrated in Fig. 1, some of the charge collected from an ionizing event in the counter is stored temporarily in the electrode capacitance C_C before flowing in the resistive impedance Z_0 of the matched cable and preamplifier. An analysis, considering that electron collection represents a current source, shows that the combination of C_C and Z_0 constitutes a low-pass filter. This configuration will be called the "C_CZ_0" filter; if the electrode area--and hence C_C--is large, this filter can have a serious effect on the system's performance. An additional filtering effect may occur in the cable* connecting the counter and preamplifier. This cable may be quite long and may have to be designed to withstand a severe temperature and radiation environment. Thus its signal transmission characteristics may be less than ideal, such that it functions as a low-pass filter. Both of these last two filtering elements are unique in that they modify signals from the counter but do not affect the preamplifier noise.

III. THEORY AND OVERVIEW

The objective of this analysis is to define and calculate a quantitative measure of the extent to which the various types of noise interfere with the counting of neutron pulses. For this purpose it is necessary to know the forms of the signal components after they have passed through the system to the discriminator input, which is where the neutron pulses are separated from the noise. In calculating these results, we will assume that the system is a cascade of linear amplifiers and lumped linear circuit elements.

*If a differential preamplifier is used, there will be two parallel cables.

Obtaining the shape and height of a neutron pulse at the discriminator input, given its shape at the counter electrodes, requires only standard circuit theory, although the actual calculation is somewhat complex. Handling the noise components and finding the rate at which they trigger the discriminator appears much less straightforward. Fortunately, as pointed out by Dayal,[5] the two necessary theorems can be found in the work of Rice.[10] The first of these states that if a transient waveform (that is, a pulse) whose Fourier transform is F(f) occurs randomly (that is, with Poisson counting statistics) at an average rate ν, the superposition of these pulses constitutes Gaussian noise whose auto-power spectral density (APSD) is given by*

$$G(f) = 2\nu |F(f)|^2 . \qquad (1)$$

If the magnitude of F(f) is random, the formula is still valid if the average of $|F(f)|^2$ is used.

The second theorem gives the average rate R(D) at which the level of a Gaussian noise signal, whose APSD is G(f), increases through the value D:

$$R(D) = f_R e^{-\frac{D^2}{2\sigma^2}} , \qquad (2)$$

where

$$\sigma^2 = \int_0^\infty G(f)df \qquad (3)$$

and

$$f_R^2 = \frac{\int_0^\infty G(f)f^2 df}{\int_0^\infty G(f)df} . \qquad (4)$$

---

*A glossary of symbols is given in Sect. VII.

Here, σ is the rms level of the noise signal, and f_R is the rms frequency of the spectral distribution of the signal. Of course, R(D) is the rate at which the noise signal would trigger a discriminator set at the level D.

The procedure to be used in this analysis can be outlined briefly as follows: Starting with a single gamma current pulse in the counter, standard methods are used to calculate the shape of the same pulse after it has propagated to the discriminator input. Its Fourier transform is taken, and Eq. (1) is used to calculate the APSD $G_{\gamma D}(f)$ of the gamma pileup noise at that point. The quantity ν in Eq. (1) is the average rate at which gamma (secondary electron) pulses occur and is proportional to the gamma dose rate. The measured APSD of the inherent output noise of the preamplifier is modified by calculation, using the frequency response function of the main amplifier-filter system, to obtain the APSD $G_{PD}(f)$ of the preamplifier noise at the input to the discriminator. The APSD used in Eqs. (3) and (4) is

$$G_{TD}(f) = G_{\gamma D}(f) + G_{PD}(f) \ . \tag{5}$$

In actual operation of the system, the discriminator will be set at a level D_1 that will suppress the spurious counts from gamma pileup and electronic noise to a negligible level, which we will take to be $R(D_1) = 1$ count/sec. From Eq. (2), the required setting is

$$D_1 = \sigma\sqrt{2\ell n f_R} \ , \tag{6}$$

where σ and f_R are obtained from Eqs. (3) and (4). In another calculation, the shape $V_{ND}(t)$ of a neutron pulse at the discriminator input is obtained by the same method used for individual gamma pulses. If the maximum value, V_M, of $V_{ND}(t)$ is extracted, one can form the ratio

$$F_M \equiv V_M/D_1 \ , \tag{7}$$

which can be considered as a quantitative figure of merit for the performance of the complete counter-channel system when the counter is subjected to intense gamma radiation.

This definition is chosen because the model, in its present form, does not deal with pulse height distributions. Rather, it requires as input data a typical gamma pulse and a typical neutron pulse, which may differ in shape as well as in magnitude. As was pointed out in connection with Eqs. (1-4), the effect of gamma pileup depends entirely on the APSD of the pileup noise, which is an average over the gamma pulse height distribution. Thus a typical gamma pulse will give valid results if it yields the correct APSD.

The concept of a typical neutron pulse is less well justified, since the counter system processes neutron pulses individually rather than responding to their average effect. The concept will be valid under the following plausible assumption concerning the neutron counting rate C as a function of discriminator setting D: that C is the same function of D/V_M for all counter system designs to which the model is applied. The function $C(D/V_M)$ is the integral pulse height distribution in units of V_M; under the above assumption C(1)/C(0) is a fixed fraction.* The assumption will be satisfied if all neutron pulses have, as they occur in the counter, the same shape when normalized to the same height and plotted as a function of time in units of the electron collection time. Since $C(D/V_M)$ increases monotonically as D/V_M decreases, and the system will be operated at $D = D_1$, the design having the largest value of F_M will have the highest neutron counting rate at the gamma dose rate for which D_1 was calculated.

Clearly, the values of F_M for two designs will indicate which is better, but not how much better; to obtain the latter information, one must use the function $C(D/V_M)$. Since the present calculation procedure does not incorporate a means of calculating $C(D/V_M)$, the function has to be obtained from measurements on an actual counter. Furthermore,

*For instance, if C(1)/C(0) = 0.5, V_M is the height, at the discriminator input, of a median-size neutron pulse occurring in the counter.

the present procedure cannot be used to evaluate changes in design parameters, such as the thickness of fissile material coating on the electrodes, that affect the form of $C(D/V_M)$.

IV. DESCRIPTION OF CALCULATION PROCEDURE

A. Development of Formulas

Before we describe the calculation procedure in detail, it is necessary to discuss some requirements and assumptions that affect the forms of the formulas used and the sequence of the calculation. Many numerical parameters are required, most of which can be specified by the designer or calculated easily from existing experimental data. Some parameters in this category are the spacing and area of the counter electrodes, the time constants of the filters in the electronics channel, and the electron drift velocity in the filling gas. There are several other parameters, however, which are extremely difficult either to measure directly or to calculate. As a practical matter, they must be determined by fitting results of the calculation to data on an actual counter. These parameters are q_N, the charge collected in the counter from a typical or median neutron event; $q_{\gamma R}$, the rms value of the charge collected from a gamma event; G_{PO}, the amplitude of the APSD of electronic noise referred to the preamplifier input; and K, the proportionality factor between the gamma dose rate R_γ and the gamma event rate ν, defined according to

$$\nu = KR_\gamma \quad . \tag{8}$$

The calculation is organized so that the formulas are developed in normalized form up to the point where these parameters can be obtained by fitting the formulas to experimental data on a counter. Once these parameters are obtained for one counter, they can be scaled according to design parameters such as electrode area and gas pressure to obtain good estimates for the values that would apply to a counter one wishes to design.

Another possible design parameter, the electronic gain in the amplifiers, is not specifically included in the calculation. Rather, signals are considered to propagate through the system as though the various inherent and designed-in lumped filter networks were connected together through unity-gain buffer amplifiers. This convention is consistent with the measurement procedures used at ORNL, in which the gain of the counting channel is calibrated by applying a known step current to the input of the preamplifier, and discriminator settings are expressed in arbitrary absolute units representing the height of current pulses at the preamplifier input.* In this calculation, signal amplitudes at every point in the counting channel are expressed either in these units or in the normalized form mentioned above. The pulse height units are discussed specifically in Sect. V.B.

The current pulse induced on the counter electrodes following an ionization event can be written in the form

$$I = \frac{q}{T}\, y\left(\frac{t}{T}\right) , \tag{9}$$

where t is the time, T is the electron collection time, q is the total electronic charge generated in the event, and y(t/T) is a dimensionless function. If the product of gas pressure and electrode spacing is sufficiently small that the ionizing particles do not lose a large fraction of their energy in the gas, the ionization density will be approximately uniform from one electrode to another, and the current pulse will have the form

$$y\left(\frac{t}{T}\right) = 1 - \frac{t}{T} \quad (0 \leqslant t \leqslant T) ,$$

$$= 0 \text{ otherwise} . \tag{10}$$

In general, the ionization density will not be uniform, particularly for fission fragments, and an analysis of possible trajectories and ionization distributions must be made in order to obtain appropriate time functions

*This calibration is described in Appendix B.

for the pulses resulting from gamma and neutron events. The construction of these functions will be described in a subsequent paper.

The normalized form of Eq. (9) is

$$I_T = \frac{1}{T}\, y\left(\frac{t}{T}\right) . \qquad (11)$$

This function has units of reciprocal time; its Fourier transform S(fT) is dimensionless. The collection time, T, is the ratio of electrode spacing to drift velocity. The latter can be obtained from published data once the gas composition and pressure, the collecting voltage, and the electrode spacing have been specified.

The first filter encountered by the current pulse is the $C_C Z_0$ filter mentioned previously; its frequency response function is

$$H_C(f) = \frac{1}{1 + j2\pi f C_C Z_0} . \qquad (12)$$

The impedance Z_0 is specified in the design. The capacitance C_C can be calculated from the specified electrode area and spacing, with suitable allowance for stray capacitance. The fission counters being developed at ORNL for power reactor applications have a guarded configuration, with two signal cables connected to the inputs of a differential preamplifier.[8] An expression for the time constant of the $C_C Z_0$ filter in this case is derived in Appendix A.

It was pointed out earlier that the pulse may also undergo filtering because of frequency-dependent attenuation in the cable to the preamplifier. The cable's frequency response function $H_A(f)$ must, in general, be obtained by laboratory measurements on the intended type of cable. The frequency response function of the electronic circuits will be denoted by $H_E(f)$. This function will be the product of several functions representing the filters incorporated to control the bandwidth, plus any additional filter or amplifier response functions that are needed to give an adequate representation. Normally these functions will be of the single- or double-pole, low- or high-pass types.

Suppose $S_N(fT)$ and $S_\gamma(fT)$ are the normalized (dimensionless) Fourier transforms of, respectively, the neutron and gamma current pulses at the counter electrodes. When these pulses arrive at the discriminator, their transforms will be

$$F_{ND}(f) = S_N(fT)\ H_C(f)\ H_A(f)\ H_E(f) \tag{13A}$$

and

$$F_{\gamma D}(f) = S_\gamma(fT)\ H_C(f)\ H_A(f)\ H_E(f)\ . \tag{13B}$$

The inverse Fourier transform of $F_{ND}(f)$, $v_D(t)$, is the normalized time-domain form of the pulse at the discriminator; the neutron pulse in absolute form is

$$V_{ND}(t) = q_N v_D(t)\ , \tag{14}$$

and its height is

$$V_M = q_N v_{DM}\ , \tag{15}$$

where v_{DM} is the positive maximum of $v_D(t)$.

From Eqs. (1) and (8), the APSD of the gamma pileup noise at the discriminator input is

$$G_{\gamma D}(f) = 2KR_\gamma q_{\gamma R}^2 |F_{\gamma D}(f)|^2\ . \tag{16}$$

The normalized form of this function will be defined as

$$g_{\gamma D}(f) = |F_{\gamma D}(f)|^2\ . \tag{17}$$

The assumed linearity of the system makes it immaterial whether Eq. (1) is applied to the original gamma current pulses in the counter or whether it is applied, as is done here, at the input to the discriminator. Also, although the expressions in Eqs. (14-16) describe signals at the discriminator input, their absolute magnitudes are in terms of current at the preamplifier input. This scheme is consistent with the gain and calibration conventions specified previously, in the discussion preceding Eq. (9).

A first approximation to the APSD function of the preamplifier noise can be obtained by assuming that white noise generated at the input is filtered by the frequency response function of the preamplifier itself. A reasonable functional form would be

$$G_P(f) = G_{P0} g_P(f) = \frac{G_{P0}}{1 + (2\pi f \tau_P)^2} , \tag{18}$$

where τ_P is the time constant of the single-pole low-pass transfer function assumed for the preamplifier. To obtain a more accurate representation one could measure the actual noise spectrum of a preamplifier similar to those expected to be used in the system one is designing. The quantitative relationship between G_{P0} and the amplitude of the measured noise spectrum is discussed in Appendix B. After being filtered in the main amplifier system, the electronic noise at the discriminator input has an APSD given by

$$G_{PD}(f) = G_{P0} g_P(f) |H_E(f)|^2 . \tag{19}$$

Combining Eqs. (5), (16), and (19) enables one to write expressions for the integrals in Eqs. (3) and (4). More convenient forms will result if the following notation is defined:

$$J_{N0} = \int_0^\infty g_P(f) |H_E(f)|^2 df \tag{20}$$

$$J_{N2} = \int_0^\infty g_P(f) |H_E(f)|^2 f^2 df \tag{21}$$

$$J_{\gamma 0} = \int_0^\infty |F_{\gamma D}(f)|^2 df \tag{22}$$

$$J_{\gamma 2} = \int_0^{\infty} |F_{\gamma D}(f)|^2 f^2 df \ . \tag{23}$$

In terms of these normalized integrals, Eqs. (5), (3), and (4) become

$$\sigma^2 = G_{P0}J_{N0} + 2KR_\gamma q_{\gamma R}^2 J_{\gamma 0} \tag{24}$$

and

$$f_R^2 = \frac{1}{\sigma^2}\left(G_{P0}J_{N2} + 2KR_\gamma q_{\gamma R}^2 J_{\gamma 2}\right) \ . \tag{25}$$

These results are used to calculate D_1 according to Eq. (6), V_M is obtained from Eq. (15), and these yield the final result, the figure of merit [Eq. (7)].

B. Conditions for Reduction of Gamma Pileup Effect

If two simplifying assumptions are made, a formula for the figure of merit can be obtained that provides some insight into the conditions needed for optimum performance. The first assumption is that electronic noise is negligible; the second is that the neutron and gamma pulses have the same shape (though they differ greatly in magnitude) so that $F_{ND}(f) = F_{\gamma D}(f) = F_D(f)$. Combining the definition, Eq. (7), with Eqs. (6) and (15) yields

$$F_M = V_M/D_1 = \frac{q_N v_{DM}}{\sigma(2\ell n f_R)^{1/2}} \ , \tag{26}$$

and substituting Eq. (22) into Eq. (24) gives (assuming $G_{P0} = 0$)

$$\sigma^2 = 2KR_\gamma q_{\gamma R}^2 \int_0^{\infty} |F_D(f)|^2 df \ . \tag{27}$$

Since $v_D(t)$, whose maximum value is v_{DM}, is the inverse Fourier transform of $F_D(f)$, Parseval's formula[11] becomes

$$\int_0^\infty v_D^2(t)dt = 2\int_0^\infty |F_D(f)|^2 df \ , \tag{28}$$

and Eq. (27) can be written

$$\sigma^2 = KR_\gamma q_{\gamma R}^2 \int_0^\infty v_D^2(t)dt \ . \tag{29}$$

The quantity $(2\ell n f_R)^{1/2}$ varies only ~10% for a factor of 10 change in f_R; thus it can be taken to be a constant. Its value is 5.6 for the fission counter described in Ref. 1. With this simplification, substitution of Eq. (29) into Eq. (26) yields

$$F_M = 0.18(q_N/q_{\gamma R})(KR_\gamma)^{-1/2}T_W^{-1/2} \ , \tag{30}$$

where

$$T_W = v_{DM}^{-2}\int_0^\infty v_D^2(t)dt \tag{31}$$

can be considered as the effective width of a neutron or gamma pulse at the discriminator. That is, T_W is the width of a rectangular pulse that delivers the same energy to the discriminator input as the actual pulse $v_D(t)$.

An optimization formula having exactly the same form as Eq. (30) was derived by Robinson and Roux[9] by calculating the probability of multiple pileup of rectangular pulses, assumed as occurring in the counter. Equation (30) is a generalization of their result. It is applicable to pulses of any shape and embodies the recognition that the effects of gamma pulse pileup must be evaluated at the discriminator input.

The parameters in Eq. (30) are $q_N/q_{\gamma R}$, K, and T_W. The first of these has a large magnitude (~10^3) and is responsible for the superior gamma rejection performance of fission counters as compared with other types of neutron detector. However, there is little the designer can do to change the value of $q_N/q_{\gamma R}$, since it varies relatively little among the filling gas compositions that have adequate chemical stability and resistance to radiation damage, combined with a satisfactory electron drift velocity characteristic. Since drift velocity is a major factor affecting T_W, it tends to determine the choice of gas composition.

The factor K is proportional to electrode area and therefore to neutron counting sensitivity; thus, low-sensitivity fission counters can be made to have better gamma rejection than high-sensitivity ones. One could reduce K by employing electrode material and filling gas of low atomic number. However, the electrode material must be chosen to withstand the expected operating temperature and to ensure that the fissile coating will adhere reliably, and the choice of gas is, as mentioned above, severely limited.

For these reasons T_W is the parameter over which the designer has the most control. It is determined mainly by the electron drift velocity, the electrode spacing, and the bandwidth of the electronic system, including the effect of the $C_C Z_0$ filter. It is not useful to decrease T_W indefinitely, however, because the electronic noise, omitted from the derivation of Eq. (30), is by no means negligible. In fact the increase in bandwidth that goes with decreasing T_W tends to increase the integral J_{NO} in Eqs. (20) and (24) and therefore to increase the noise signal at the discriminator. The difficulty of balancing these factors is one of the principal reasons the present optimization theory was developed.

C. Determination of Parameters

The development of Eqs. (9-25) was accompanied by explanations of how values of most of the parameters appearing in them could be obtained either from the basic design values or from measurements on components. It was also pointed out that the quantities q_N, $q_{\gamma R}$, K, and G_{P0} can best be obtained from integral pulse height distributions obtained with a test fission counter exposed alternately to neutron and gamma radiation. The counter used for these measurements will be referred to as the reference counter. The procedure relies on the fact that v_{DM}, defined in connection with Eq. (15), and the integrals (20-23) can all be calculated for the reference counter without knowing any of the above four quantities. The type of data obtained from the reference counter is illustrated in Fig. 2, which shows integral pulse height distributions for neutrons, electronic noise, and four gamma dose rates for a development model high temperature fission counter.[1]

The first step is to establish V_{M0},* the value of V_M for the reference counter. This can be done by marking a point on the neutron integral pulse height distribution. If the point is chosen so that the corresponding count rate is half the extrapolated count rate at zero pulse height, V_{M0} is the median neutron pulse height at the discriminator. From Eq. (15), the median charge released in a neutron event is

$$q_{N0} = V_{M0}/v_{DM0} \ , \tag{32}$$

in which v_{DM0} is obtained from the theory, using the known parameters of the reference counter.

The second step is to fit Eq. (2) to the integral pulse height distributions for gamma and electronic noise. The most expeditious method is to convert the data to the form $\ell n(R)$ versus D^2 and do a linear least-squares fit; the resulting intercept is an estimate of $\ell n(f_{R0})$ and the slope is an estimate of $-1/2\sigma_0^2$. The smooth lines drawn through the gamma and electronic noise data in Fig. 2 were generated from the values of f_{R0} and σ_0 obtained by this method. Since both of these quantities are functions of gamma dose rate, R_γ, the third step is to do another least-squares fit of a straight line to the resulting values of σ_0^2 versus R_γ. According to Eq. (24), this line should have an intercept

$$I = G_{P00}J_{N00} \tag{33}$$

and a slope

$$S = 2K_0 q_{\gamma R0}^2 J_{\gamma 00} \ . \tag{34}$$

---

*Symbols applicable to the reference counter will be given an additional subscript 0.

Figure 3 shows this fitted line superimposed on the values of σ_0^2 obtained from the data of Fig. 2. It turns out that Eq. (2) does not always fit well to electronic noise data, and the fits may yield anomalous values of f_{R0}; it is possible that the electronic noise does not have a Gaussian amplitude distribution. For this reason, better results may be obtained if the electronic noise values are not used in fits of the type shown in Fig. 3. The final step is to use the theory to calculate J_{N00} and $J_{\gamma 00}$ from the known design parameters of the reference counter, and then to apply Eqs. (33) and (34):

$$G_{P00} = I/J_{N00} \tag{33A}$$

$$K_0 q_{\gamma R0}^2 = S/2J_{\gamma 00} \ . \tag{34A}$$

Although K_0 and $q_{\gamma R0}$ are not determined separately, there is no need to do so.

To obtain from these results the corresponding parameters for a counter one wishes to design, scaling laws can be used. The simplest schemes would be to assume that K is proportional to electrode area and that q_N and $q_{\gamma R}$ are proportional to the product of gas pressure and electrode spacing. Scaling laws with wider ranges of validity could be developed and verified by fitting the theory to data from several fission counters having widely varying design parameters. If the design counter is to use a different preamplifier than used in the reference counter, a conversion factor from G_{P00} to G_{P0} would have to be obtained by measuring the output noise spectra of both units.

V. COMPUTER IMPLEMENTATION

A. System Structure

A flow chart of the calculation procedure in its basic form is shown in Fig. 4. The chart is organized not only in logical sequence but also according to the type of data being processed. The heart of the calculation, called the "Fourier calculation sequence," consists of the

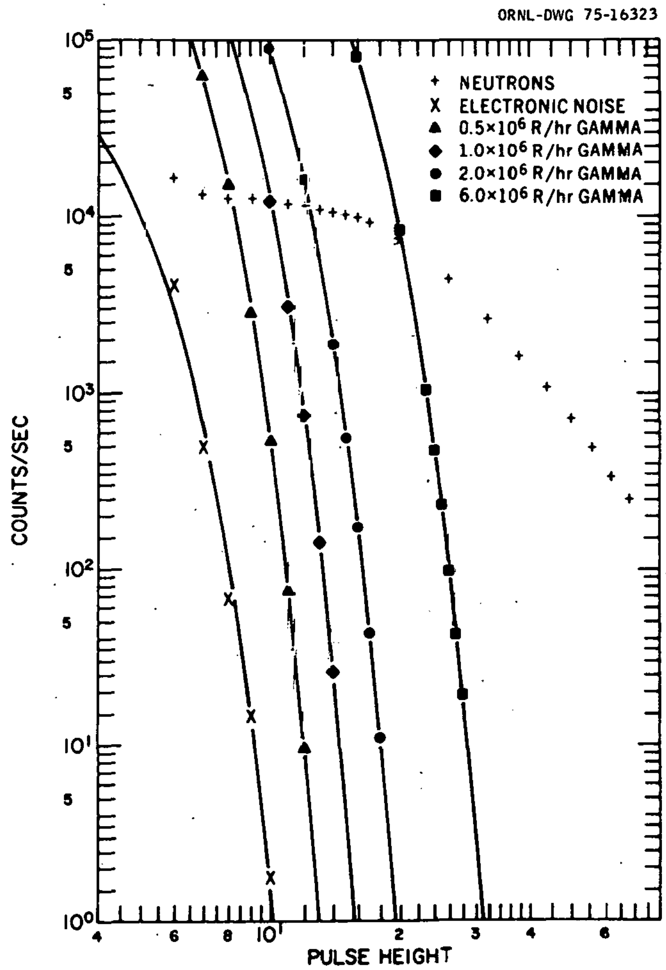

Fig. 2. Integral pulse height distributions for a fission counter. Smooth curves are fits of Eq. (2) to the gamma and electronic noise data.

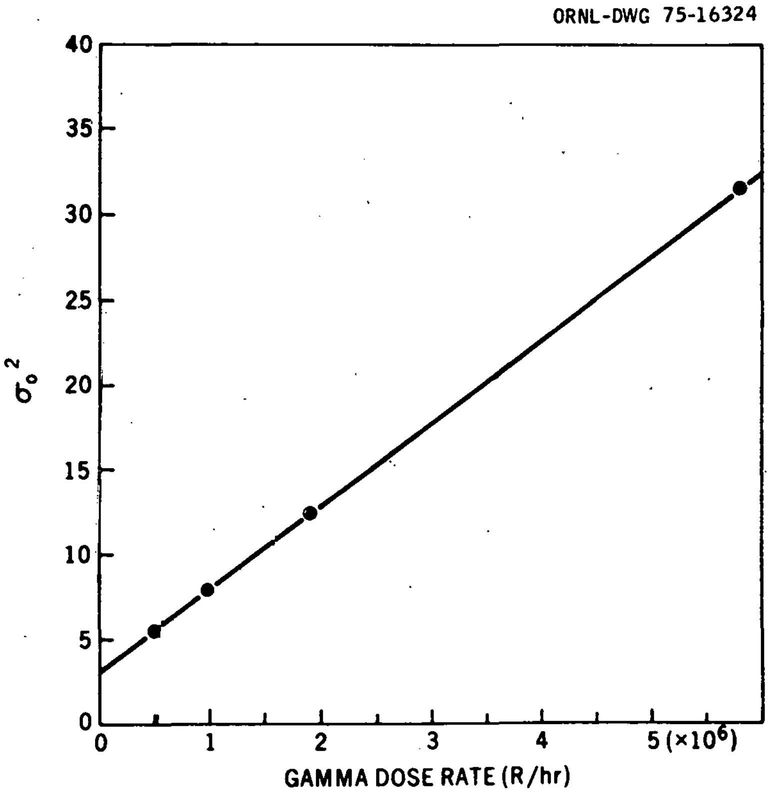

Fig. 3. Dose rate dependence of mean square level of gamma pileup plus electronic noise, using values derived from fits of Eq. (2) to data in Fig. 2. Straight line is a least-squares fit to the data shown.

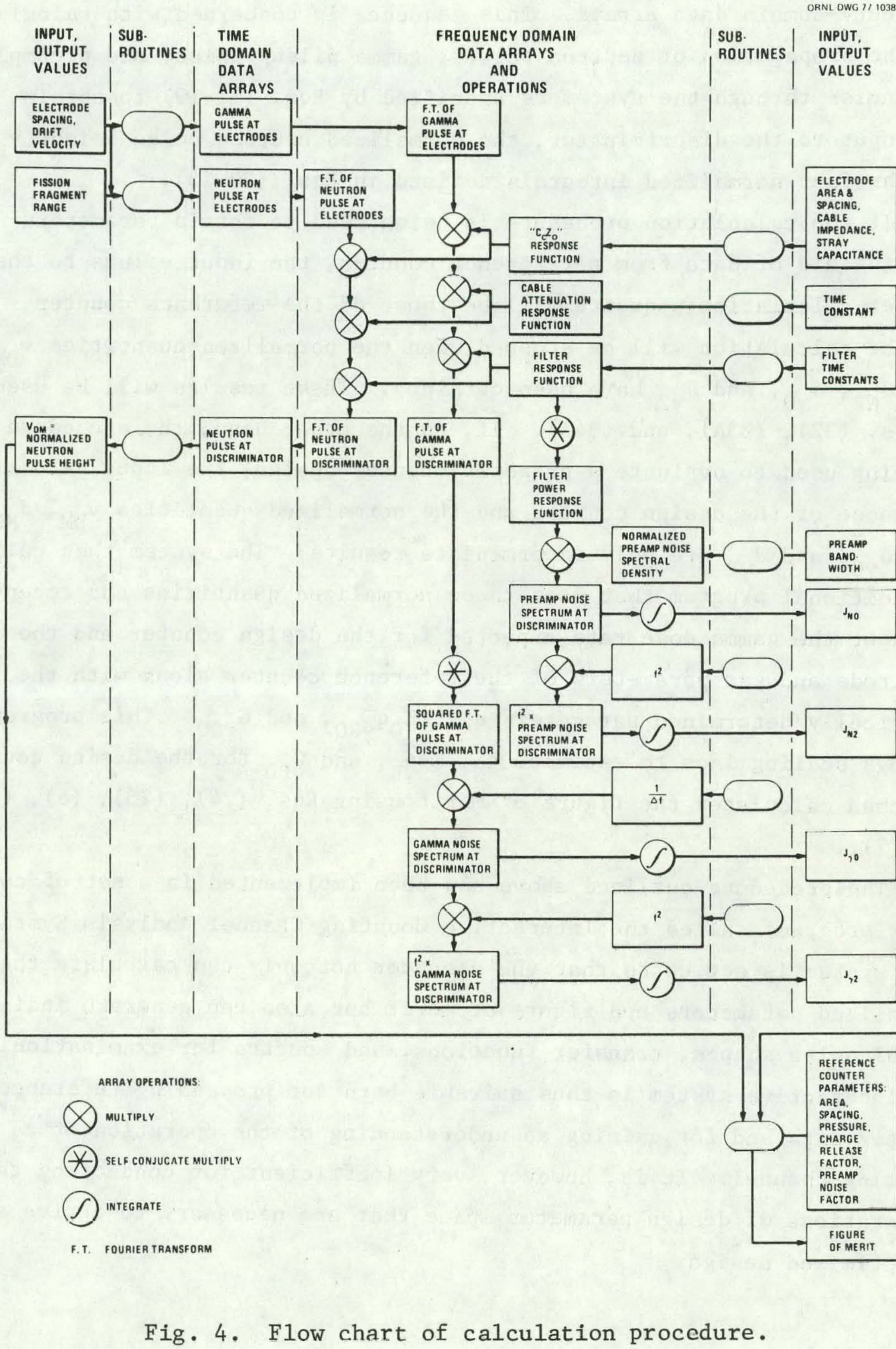

Fig. 4. Flow chart of calculation procedure.

generation, Fourier transformation, and manipulation of time- and frequency-domain data arrays. This sequence is concerned with calculating the propagation of neutron pulses, gamma pileup noise, and preamplifier noise through the system as specified by Eqs. (11-19) to obtain, at the input to the discriminator, the normalized neutron pulse height v_{DM} and the four normalized integrals defined in Eqs. (20-23).

If the calculation procedure is being used to obtain parameters with the aid of data from a reference counter, the input values to the Fourier calculation sequence will be those of the reference counter, and the calculation will be stopped when the normalized quantities v_{DM}, J_{N0}, J_{N2}, $J_{\gamma 0}$, and $J_{\gamma 2}$ have been obtained. These results will be used in Eqs. (32), (33A), and (34A). If, on the other hand, the procedure is being used to evaluate a proposed counter design, the input values are those of the design counter and the normalized quantities v_{DM}, J_{N0}, J_{N2}, $J_{\gamma 0}$, and $J_{\gamma 2}$ are only intermediate results. The system then calls an additional program that uses these normalized quantities and accepts as input the gamma dose rate expected for the design counter and the electrode and gas parameters of the reference counter along with the empirically determined parameters q_{N0}, $K_0 q^2_{\gamma R0}$, and G_{P00}. This program employs scaling laws to estimate q_N, $Kq^2_{\gamma R}$, and G_{P0} for the design counter, and then calculates the figure of merit using Eqs. (24), (25), (6), (15), and (7).

The procedure outlined above has been implemented in a set of computer programs called the Interactive Counting Channel Analysis System. This system is set up so that the operator not only can calculate the normalized parameters and figure of merit but also can generate individual pulse shapes, transfer functions, and spectra for examination. The interactive system is thus suitable both for processing reference counter data and for gaining an understanding of the operation of a counting channel. It is, however, very inefficient for conducting the explorations of design parameter space that are necessary to arrive at an optimized design.

To meet this latter need, a Production Counting Channel Analysis System has also been written and put into operation. This system facilitates exploration of parameter space by automatically incrementing, within preset limits, the electrode spacing, filter time constants, and gamma dose rate. The operator need specify only the limits and the number of increments for each. The filter time constants are derived from a reference time constant T_R and up to four ratios R_i, each representing a filter whose time constant is $T_R R_i$. The available filter types are one- and two-pole, low- and high-pass; one of each of the four possible types may be selected. The two-pole filters may be Butterworth, Bessel, or Dual RC ($Q = 1/\sqrt{2}$, $1/\sqrt{3}$, and $1/2$, respectively). Any or all of the five quantities T_R and R_i may be automatically incremented or held constant; thus the system can maintain constant fractional bandwidth by incrementing T_R or can vary the bandwidth by incrementing one or more of the R_i's. The system sets up increments that are equal on a linear scale for electrode spacing and on a logarithmic scale for T_R, the R_i's, and gamma dose rate.

To prepare input values for the interactive system, the operator selects a collecting voltage V, an electrode spacing D_E, and a gas pressure P and calculates the electric field $E = V/D_E$ and then the ratio $E/P = V/PD_E$. The drift velocity U is found from a plot of U vs E/P for the selected gas composition. For the production system, this scheme cannot be used, since D_E is to be automatically incremented. Instead, the operator selects U and E/P from the appropriate plot and chooses V. From these inputs, the system computes a quantity $(PD_E) = V/(E/P)$. Each time it selects D_E, it calculates $P = (PD_E)/D_E$.

The input parameters that represent the design counter are U, E/P, V, electrode area, estimated stray capacitances, the time constant for cable attenuation, the preamplifier input impedance, and the limits and numbers of increments for D_E, T_R, the R_i's, and gamma dose rate. The system also accepts the electrode area and spacing, the gas pressure, and the time constant τ_p in Eq. (18), along with the empirically determined parameters q_{N0}, $K_0 q^2_{\gamma R0}$, and G_{P00} of the reference counter; and uses them in the same way as the interactive system. Having accepted all these inputs, the production system prints them as output and proceeds to calculate and print intermediate and final results for each of the cases implied by the limits and numbers of increments specified for the seven automatically incremented quantities.

B. Time, Frequency, and Amplitude Scales

The time- and frequency-domain functions generated by the system exist as one-dimensional data arrays in the computer. Thus, time and frequency are represented by integer array subscripts, and the proportionality factors relating actual values to subscript values must be chosen so as to achieve the best possible resolution on both time and frequency scales. Furthermore, some of the functions are densities in the frequency domain, and care must be taken to assure that correct amplitudes are obtained when these functions are squared or integrated.

To express these considerations quantitatively, let

N be the number of elements in a time-domain array, and
Δt, the time interval between elements;
Δf, the frequency interval between elements, and
F_{max}, the maximum frequency in a frequency-domain array;
n, the array subscript for time-domain, and
m, the array subscript for frequency-domain arrays.

In the usual scheme of definitions,[12,13] with frequency restricted to nonnegative values, n varies from 0 to N - 1 and m from 0 to N/2. A Fourier transform has both real and imaginary parts for all elements except those at m = 0 and m = N/2, whose values are real only; thus the transform occupies the same amount of data space as the real time-domain function from which it was derived. The hypothetical sampling rate is $1/\Delta t$; by the sampling theorem,[12,13] the maximum frequency that can be represented is half this value, from which

$$\Delta t = 1/(2F_{max}) \ . \qquad (35)$$

Since there are N/2 frequency intervals,

$$\Delta f = 2\ F_{max}/N\ . \tag{36}$$

Combining these yields

$$\Delta f\ \Delta t = 1/N\ , \tag{37}$$

which expresses the tradeoff between time and frequency resolution.

To make this tradeoff more specific, assume that the most significant part of a time-domain function extends from $t = 0$ to $t = t_s$ and that the most significant part of its Fourier transform extends from $f = 0$ to $f = f_s$. The corresponding numbers of elements in each array are $n_s = t_s/\Delta t$ and $m_s = f_s/\Delta f$. Substitution into Eq. (37) yields

$$n_s m_s = N t_s f_s\ . \tag{38}$$

As an example, consider the idealized pulse defined by Eq. (10). The obvious choice for t_s is T; for f_s one can choose the frequency of the first zero of the real part of the Fourier transform. For the function of Eq. (10), this zero occurs at frequency 1/T; thus for this case $t_s f_s = 1$. It is generally true that $t_s f_s$ is of order unity for pulse-type time-domain functions. Hence, Eq. (38) reduces to

$$n_s m_s = N\ , \tag{39}$$

and the choice of scale that avoids excessively coarse resolution in either time or frequency is $n_s = m_s = \sqrt{N}$. Clearly, N must be large if the calculated functions are to be represented by a reasonable number of points, and, if it is, the "significant" part of each function will occupy only a small fraction of the array.

It should be kept in mind, however, that the calculation actually convolves a time-domain function such as that in Eq. (10) successively with the impulse response of several filters to obtain the time-domain pulse at the discriminator input. The latter pulse is much extended in time compared with the former. Thus, the condition $N >> n_s$ provides protection against the phenomenon of "wraparound"[12] which would occur if the time-domain functions occupied large fractions of their arrays.

The amplitude scales are derived from the expression for a current pulse, Eq. (9), and its normalized form, Eq. (11). The latter has dimensions of reciprocal time, and its Fourier transform is dimensionless. Thus, a normalized spectral density such as that given in Eq. (17) is also dimensionless. Some care must be exercised in computer programming, however, because when asked for a transform F(f), the discrete Fourier transform algorithm used in this work actually generates $\Delta f\ F(f)$. Furthermore the desired magnitude for the corresponding spectral density is $\Delta f|F(f)|^2$, in order that the correct magnitude and dimensionality are obtained when its frequency components are summed to generate integrals such as those in Eqs. (20-23). In Fig. 4 the gamma pileup noise spectrum at the discriminator is therefore shown as being obtained by squaring the Fourier transform of the gamma (secondary electron) pulse and dividing by Δf.

The magnitudes and units of the charge parameters q_N and $q_{\gamma R}$ are derived from the calibration procedure used in the ORNL neutron sensor development program. In this procedure, an arbitrary pulse height scale is established by recording the discriminator setting corresponding to a step-function current pulse of known height applied to the preamplifier input. The unit of pulse height, known as the PHS, represents a current of 38.28 nA. The dimensions of q_N and $q_{\gamma R}$ are PHS-μsec; the dimensions of spectral densities, such as defined in Eq. (16), are $(PHS)^2/MHz$. In Eq. (15), v_{DM} has dimensions of $(\mu sec)^{-1}$ [as does Eq. (11)], and V_M is expressed in PHS units.

C. Computer Requirements

The two, counting channel analysis systems could be implemented in any computer that has sufficient memory and processing speed and for which the required fast Fourier transform (FFT) and array manipulation software are available. The computer used in developing the systems is a Hewlett-Packard 5451B Fourier analyzer. Its principal elements are a 2100A computer with 32K of memory, a 5470B fast Fourier processor with 8K of memory, a disc drive, a CRT display, and a control keyboard. All FFT operations, as well as array manipulations such as multiplication and integration, are done in the fast processor, under control of programs loaded as required from the computer. The manufacturer's Fourier analysis software allows merger and interaction with user-written programs and uses the disc drive to store data arrays and complete core-loads of programs. However, it does not allow loading and automatic operation of partial core-loads as a true disc operating system would. An array size, N, of 4096 is used; thus the fast processor holds only two arrays. The operations shown in Fig. 4, therefore, require frequent temporary storage of intermediate arrays on the disc.

The 5451B Fourier analyzer is well suited for the interactive counting channel analysis system since it can be programmed interactively and the display can show pulse shapes, transforms, and spectra as they propagate through the counting channel. However, the production system, with its nested loop structure, more elaborate input and output formats, wider choice of filter functions, and ability to calculate a large number of cases, taxes both the processing speed and the memory addressing capability of the Fourier analyzer. The latter problem has had to be dealt with by dividing some of the programs into smaller modules and employing special loading procedures. With these modifications, the production system runs quite satisfactorily on this analyzer.

VI. ACKNOWLEDGEMENTS

In developing this fission counter system model, the author benefited greatly from discussions with J. T. De Lorenzo and K. H. Valentine, both of ORNL.

VII. GLOSSARY

C_C	Electrode capacitance
Z_O	Cable and preamplifier impedance
ν	Average gamma (secondary electron) event rate
R_γ	Gamma dose rate
K	Factor relating ν to R_γ
f	Frequency
σ	Standard deviation (rms value) of random signal
f_R	Rms freqency of random signal
D	Discriminator setting
D_1	Discriminator setting that limits average triggering rate due to noise to 1 count/sec
R(D)	Discriminator triggering rate
q	Electronic charge released in filling gas in a single ionizing event
q_N	Value of q for a typical or median neutron event
$q_{\gamma R}$	Rms value of q for gamma events
T	Electron collection time
I(t)	Current induced on counter electrodes by a single ionizing event, as a function of time, t
y(t)	Dimensionless function, with initial value of unity, having form of I(t)
$I_T(t)$	Normalized form of I(t)

Fourier Transforms:

F(f)	of unspecified function
S(fT)	of $I_T(t)$
$S_N(fT)$	of normalized neutron pulse in counter
$S_\gamma(fT)$	of normalized gamma pulse in counter
$F_{ND}(f)$	of normalized neutron pulse ADI*

*At discriminator input.

$F_{\gamma D}(f)$	of normalized gamma pulse ADI
$F_D(f)$	of normalized neutron or gamma pulse, assumed the same

Auto Power Spectral Densities (APSDs):

$G(f)$	of unspecified signal
$G_{\gamma D}(f)$	of gamma pileup noise ADI
$G_{PD}(f)$	of preamplifier noise ADI
$G_{TD}(f)$	of total noise ADI
$g_{\gamma D}(f)$	Normalized form of $G_{\gamma D}(f)$
$G_P(f)$	of preamplifier noise
$g_P(f)$	Normalized form of $G_P(f)$

Transfer Functions:

$H_c(f)$	Determined by C_C and Z_0
$H_A(f)$	of cable attenuation
$H_E(f)$	of preamplifier, main amplifiers, and filters
G_{PO}	Factor relating $G_P(f)$ to $g_P(f)$
τ_P	Time constant of single-pole roll-off of preamplifier noise spectrum
$V_{ND}(t)$	Neutron pulse ADI
V_M	Maximum of $V_{ND}(t)$
$v_D(t)$	Normalized form of $V_{ND}(t)$
v_{DM}	Maximum of $v_D(t)$
F_M	Figure of merit
$C(D/V_M)$	Counting rate as function of D and v_M
T_w	Effective width of pulse at discriminator
J_{N0}, J_{N2}, $J_{\gamma 0}$, $J_{\gamma 2}$	Integrals defined in Eqs. (20-23)

Additional subscript 0 indicates value derived from reference counter

I,S	Intercept and slope obtained in determining electronic and gamma noise parameters of reference counter
T_R	Reference filter time constant
R_i	Factor relating time constant of filter i to T_R
Q	Quality factor in filter transfer function
V	Collecting voltage
D_E	Electrode spacing
P	Gas pressure
E	Electric field between electrodes
U	Electron drift velocity
N	Number of elements in a time-domain array
Δt	Time interval between elements
Δf	Frequency interval between elements
F_{max}	Maximum frequency in a frequency-domain array
n	Subscript for time-domain array
m	Subscript for frequency-domain array
t_s	Time-domain extent of significant part of signal
n_s	Corresponding number of array elements
f_s	Frequency-domain extent of significant part of a spectrum
m_s	Corresponding number of array elements
PHS	Arbitrary calibration unit of pulse height

APPENDIX A

RISE TIME OF A COUNTER CONNECTED TO A DIFFERENTIAL CURRENT-PULSE AMPLIFIER

To minimize electromagnetic interference, reactor fission counters being developed at ORNL are built with both electrodes ungrounded and connected via cables of characteristic impedance Z_0 to the inputs, also of impedance Z_0, of a differential current-pulse preamplifier.[8] A circuit model of this arrangement is shown in Fig. 5. Here C_E is the capacitance between electrodes, and C_H and C_S are the stray capacitances of the electrodes to the case, which is connected to the preamplifier ground via the cable shields. The current pulse is represented by I_p, and E_d is the differential voltage output.

It is convenient to study this circuit in the frequency domain; its behavior is then specified by its transfer function

$$H_c(\omega) = \frac{E_d}{2Z_0 I_p} . \tag{A1}$$

Analysis of the network shows that

$$H_c(\omega) = \frac{1 + j\omega(\tau_s + \tau_H)/2}{1 - \omega^2(\tau_E\tau_H + \tau_E\tau_S + \tau_S\tau_H) + j\omega(2\tau_E + \tau_S + \tau_H)} , \tag{A2}$$

where

$$\begin{aligned} \tau_E &= Z_0 C_E \\ \tau_H &= Z_0 C_H \\ \tau_S &= Z_0 C_S . \end{aligned} \tag{A3}$$

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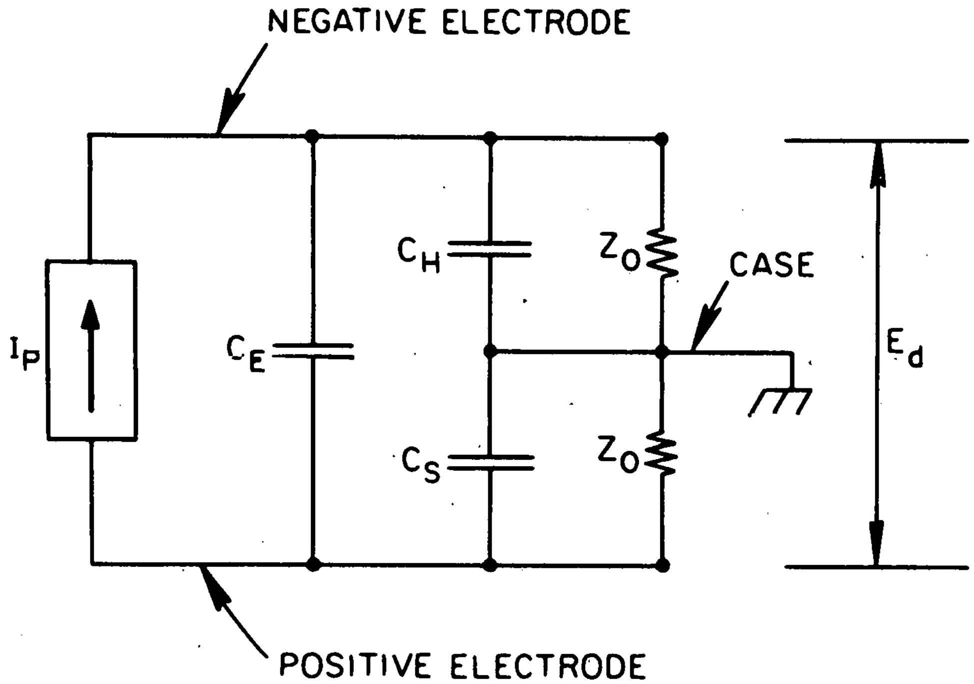

Fig. 5. Circuit model of guarded fission counter connected to differential preamplifier.

Since a long rise time is deleterious to the counter's performance, it will be designed so that C_H and C_S are much less than C_E. Thus, in the frequency range in which $H_e(\omega)$ has significant magnitude, $\omega(\tau_H + \tau_S)$ is small. Dividing the denominator of $H_c(\omega)$ by the numerator and dropping terms of order $\omega^2(\tau_S + \tau_H)^2/4$ or smaller yields

$$H_c(\omega) \cong 1/\{1 + j\omega\ [2\tau_E + \frac{1}{2}\ (\tau_S + \tau_H)]\}\ . \tag{A4}$$

Thus the circuit behaves as a single-pole, low-pass filter with a time constant

$$\tau_c = 2\tau_E + \frac{1}{2}\ (\tau_S + \tau_H)\ . \tag{A5}$$

It can be shown that this result is exact if $\tau_H = \tau_S$.

APPENDIX B

REPRESENTATIONS OF PREAMPLIFIER NOISE

In designing or specifying a preamplifier for a counter system, or in establishing input parameters for this counter system model, it may be useful to be able to relate preamplifier noise, as defined in electronics terminology, to the noise parameters defined in the counter system model and to the parameters of the integral pulse height distribution, which is the observable manifestation of the noise. This discussion shows how the formalism of the model can be used to establish these relationships and develops approximate formulas based on simplified representations of the preamplifier noise spectrum and of the frequency response of the counting channel.

An accepted specification of preamplifier noise is in terms of the equivalent noise current referred to the input. It is consistent with measurements on actual preamplifiers to assume that the power spectral density $G_P'(f)$ of this current [in units of $(pA)^2/Hz$] has a constant value G_{P0}' from zero frequency to a maximum frequency f_m, and is zero above f_m. This quantity is analogous to G_{P0}, defined in Eq. (18), But multiplies a different frequency function and is expressed in different units. A common measure of such noise is the rms noise current density,[5,14]

$$I_P = \sqrt{G_{P0}'} \; , \tag{B1}$$

in units of $pA/\sqrt{Hz}$.

The frequency response function of the channel will be denoted by H(f); it includes the electronic gains of the preamplifier and main amplifier(s) as well as the effects of the filters added by the designer to optimize system performance. Since H(f) relates discriminator input voltage (V) to preamplifier input current (pA), it has units of teraohms ($T\Omega = 10^{12}\ \Omega$).

The power spectral density of the preamplifier noise at the discriminator input is, in V^2/Hz,

$$G_D(f) = G_P'(f)|H(f)|^2 . \qquad (B2)$$

The rate at which this signal triggers the discriminator is given by Eq. (2), with σ and f_R given by

$$\sigma^2 = \int_0^\infty G_D(f)df \qquad (B3)$$

and

$$f_R^2 = I_2/\sigma^2 , \qquad (B4)$$

where

$$I_2 = \int_0^\infty G_D(f)f^2df . \qquad (B5)$$

Both σ and D are in volts and f_R has units of Hertz.

From the standpoint of evaluating and operating the counting channel, a significant quantity is D_1, the discriminator setting required to limit the spurious counting rate (due to noise) to 1 count/sec. The expression for D_1, Eq. (6), involves the integrals defined in Eqs. (B3) and (B5). Using Eq. (B2) and the representation of preamplifier noise defined in connection with Eq. (B1), we write these integrals as follows:

$$\sigma^2 = I_P^2 \int_0^{f_m} |H(f)|^2df \qquad (B6)$$

and

$$I_2 = I_P^2 \int_0^{f_m} |H(f)|^2f^2df . \qquad (B7)$$

Normally the filtering in a channel will consist of one each of single-pole, high- and low-pass filters, both having the same RC time constant τ. Then if A is the electronic gain without filtering, H(f) is given by

$$H(f) = A \frac{j2\pi f\tau}{(1 + j2\pi f\tau)^2} \tag{B8}$$

so that

$$|H(f)|^2 = A^2 \frac{(2\pi f\tau)^2}{[1 + (2\pi f\tau)^2]^2} . \tag{B9}$$

Substituting Eq. (B9) into Eqs. (B6) and (B7) yields

$$\sigma^2 = \frac{A^2 I_P^2}{2\pi\tau} \int_0^{x_m} \frac{x^2 dx}{(1 + x^2)^2} \tag{B10}$$

and

$$I_2 = \frac{A^2 I_P^2}{(2\pi\tau)^3} \int_0^{x_m} \frac{x^4 dx}{(1 + x^2)^2} , \tag{B11}$$

where

$$x = 2\pi f\tau \tag{B12a}$$

and

$$x_m = 2\pi f_m \tau . \tag{B12b}$$

Evaluating these integrals yields

$$\sigma^2 = \frac{A^2 I_P^2}{2\pi\tau} \left[- \frac{x_m}{2\left(1 + x_m^2\right)} + \frac{1}{2} \tan^{-1}\left(x_m\right) \right] \tag{B13}$$

and

$$I_2 = \frac{A^2 I_P^2}{(2\pi\tau)^3} \left[x_m + \frac{x_m}{2\left(1 + x_m^2\right)} - \frac{3}{2} \tan^{-1}\left(x_m\right) \right] . \qquad \text{(B14)}$$

Typical parameters for a system would be $\tau = 27$ nsec and $f_m \cong 60$ MHz, so that $x_m \cong 10$. Thus a good approximation is $\tan^{-1}(x_m) \cong \pi/2$, and the terms containing $1 + x_m^2$ in the denominator can be neglected. With these approximations one obtains

$$\sigma^2 = \frac{A^2 I_P^2}{8\tau} , \qquad \text{(B15)}$$

and, using Eq. (B4),

$$f_R^2 = \frac{x_m - \frac{3\pi}{4}}{\pi^3 \tau^2} . \qquad \text{(B16)}$$

Since D_1, as given by Eq. (6), is very insensitive to variations in f_R, it is sufficient to use the typical value $x_m = 10$, yielding

$$f_R^2 = 0.25/\tau^2 . \qquad \text{(B17)}$$

Substituting Eqs. (B15) and (B17) into Eq. (6) yields

$$D_1 = AI_P \left[\frac{-1.40 - 2 \ln \tau}{8\tau} \right]^{1/2} . \qquad \text{(B18)}$$

If τ is replaced by its equivalent τ_n in nanoseconds and if τ_n is in the typical range of 10 to 50 nsec, this expression can be approximated within 3% by

$$D_1 = 6.5 \times 10^4 \frac{AI_P}{\sqrt{\tau_n}} . \qquad \text{(B19)}$$

The above formula gives D_1 in terms of discriminator threshold in volts. Next we will show how to express D_1 in terms of ORNL absolute calibration ("PHS") units. The calibration is done by applying pulses of precisely known height to the preamplifier input, using a mercury relay pulser. These pulses have a rise time, T_{CP} (10% to 90%), of ∿12 nsec and a decay time of ∿300 μsec. Since the rise time is shorter than typical values of the filter time constant, τ, and the decay time is much greater, the system responds nearly as though the calibrating pulse were a step function.

It can be shown that applying a step function pulse to a system whose frequency response function is given by Eq. (B8) produces a pulse at the discriminator input, whose height is

$$D_{CP} = \frac{1}{e} A\alpha I_{CP} \quad , \tag{B20}$$

where e = 2.718..., I_{CP} is the amplitude of the calibrating pulse that corresponds to 1 PHS unit, and αI_{CP} is the actual height of the calibrating pulse. The ORNL calibration method employs the arbitrary definition

$$I_{CP} = 38.28 \text{ nA} = 38280 \text{ pA} \tag{B21}$$

and normally applies pulses whose amplitude is α = 36.11 PHS units. Clearly, D_{CP}/α is the height, in volts at the discriminator, of a calibrating pulse whose amplitude is 1 PHS unit. Thus, the discriminator setting given by Eq. (B19), expressed in PHS units, is

$$D_{1P} = \frac{D_1}{D_{CP}/\alpha} = \frac{6.5 \times 10^4 \dfrac{AI_P}{\sqrt{\tau_n}}}{\dfrac{1}{e} AI_{CP}} \quad ,$$

or

$$D_{1P} = \frac{1.767 \times 10^5}{\sqrt{\tau_n}} \cdot \frac{I_P}{I_{CP}} \quad . \tag{B22}$$

Substituting the definition in Eq. (B21) gives

$$D_{1P} = 4.616 \frac{I_P}{\sqrt{\tau_n}} . \qquad \text{(B23)}$$

Equation (B23) provides the relationship between the rms preamplifier noise current, I_p, and the discriminator setting, D_1 (in ORNL PHS units) required to limit to 1 count/sec the average spurious counting rate due to preamplifier noise in the absence of other signals or sources of noise. However, the noise in a preamplifier connected to a counter may not have the simple spectrum defined in connection with Eq. (B1). Thus, Eq. (B23) can be a useful guide, but should not be used quantitatively without a measurement of the actual noise spectrum.

References

1. V. K. Paré, W. T. Clay, J. T. De Lorenzo, and G. C. Guerrant, "Design Parameters and Test Results for a Fission Counter Intended for Operation Up to 750°F at High Gamma Dose Rates," *IEEE Trans. Nucl. Sci.* NS-22, 696 (February 1975).

2. L. D. Philipp, N. C. Hoitink, W. G. Spear, and M. R. Wood, "Effects of Gas Mixture, Electrode Spacing, Gas Pressure and Applied Voltage on the Gamma Performance of Fission Counters," *Nucl. Technol.* 20, 51 (1973).

3. L. D. Philipp, L. D. Muhlestein, and M. R. Wood, "Comparison of Gamma Irradiation Performance of Three Neutron Fission Counters," *Nucl. Technol.* 23, 80 (1974).

4. L. D. Philipp and N. C. Hoitink, "Parameter Variations for Improved Fission Counter Performance," HEDL-TME 75-17, Westinghouse Hanford Company (1975).

5. Y. Dayal, *Analytical Study of a Source Range Monitoring System for LMFBR*, GEAP-10718 (November 1972); Y. Dayal and L. C. Wimpee, *Trans. Am. Nucl. Soc.* 16, 236 (1973).

6. J. T. De Lorenzo, "Optimization of RC-CR Filters for Processing of Current Pulses from a Fission Counter Operating in a High Gamma Background," ORNL/TM-4928 (August 1975).

7. J. T. De Lorenzo, "Feasibility Study for a High-Sensitivity Neutron Detector," ORNL/TM-5129 (March 1976).

8. J. T. De Lorenzo, W. T. Clay, and G. C. Guerrant, "Differential Current Pulse Preamplifier for Fission Counters," *IEEE Trans. Nucl. Sci.* NS-21, 757-762 (1974).

9. D. P. Roux et al., "Neutron Sensor Development for Extreme Environments," ORNL/TM-3959 (December 1972).

10. S. O. Rice, "Mathematical Analysis of Random Noise," from selected papers on *Noise and Stochastic Processes*, pp. 133-294, Dover Publications, 1954.

11. A. Papoulis, "The Fourier Integral and Its Applications," p. 27, McGraw-Hill, New York, 1962.

12. William D. Stanley, "Digital Signal Processing," pp. 239-307, Reston, 1975.

13. Peter R. Roth, "Digital Fourier Analysis," and Agoston Z. Kiss, "A Calibrated Computer-Based Fourier Analyzer," Hewlett-Packard Journal 21, 2-20 (1970).

14. U. S. Atomic Energy Commission RDT Standard C 15-3T, "Current Pulse Preamplifiers for Use with Fission Counters," with Amendment 1 (June 1973).

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