SIGNAL, NOISE AND RESOLUTION IN POSITION-SENSITIVE DETECTORS*

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ABSTRACT

An analysis is presented of signal, noise and position resolution relations for some of the most interesting position-sensing methods. "Electronic cooling" of delay line terminations is introduced in order to reduce noise in the position-sensing with delay lines. A new method for terminating transmission lines and for "noiseless" damping which employs a capacitance in feedback is presented. It is shown that the position resolution for the charge division method with resistive electrodes is determined only by the electrode capacitance and not by the electrode resistance, if optimum filtering is used.

1. Introduction

There has been a great proliferation of various readout methods for position sensitive detectors. One would imagine that in each particular case the best solution is arrived at by 1) the detector design to maximize the significant signal, 2) reduction of noise at its physical source, and 3) optimum filtering of signal and noise. Many imaginative contributions have been made to the detector design. It can be stated, however, that the noise reduction and filtering have been neglected to the point where detector design and operation become critical in order to obtain a large enough signal to overcome noise.

In interpolating readouts the position resolution is determined by the noise which originates either in the position sensing medium or in the amplifier and other components external to it. In readouts for multiwire proportional chambers based on one amplifier per wire it is usually assumed that the signal is large and cost and size are dominant in design considerations. Recently applications have become of interest where the signal magnitude is limited by the detector design (smaller anode spacing), or where linear signal information is needed. In all cases involving semiconductor detectors, and gas or liquid ionization detectors with little or no gain, noise becomes important.

This paper concentrates on point 2), that is, reduction of noise at its physical source. Optimum filtering is determined by both signal and noise and by the information (amplitude or time) required so that a great variety of cases is possible. A simple approach based on filter weighting function analysis is given here, and the treatment of this subject is limited to several examples.

There are two prevalent interpolating readout methods in use and under development at present. One is based on a delay line as a position sensing medium.[1,2,3] The other is based on distributed R-C line.[4-9] The two are entirely different. In the first one the position-sensing medium is non-dissipative in principle, and the noise is generated external to it in the terminations and amplifiers. In this paper "electronic cooling" of delay line terminations is introduced in order to reduce noise. A new method for terminating transmission lines and for "noiseless" damping by a capacitance in feedback is presented.

An R-C line is a dissipative position-sensing medium. With proper choice of line and amplifier parameters the noise in signal amplifiers is negligible, and all the noise is generated in the position-sensing medium. Thus the noise and the position resolution are determined by the R-C line and by the filtering. It is shown here that for detectors with short charge collection times the position resolution is independent of the time required to process one event. Detectors with smaller resistance, better timing and energy resolution can be made with the same position resolution.

The case of single amplifier per wire (detector segment) is discussed with respect to noise limits, location of amplifiers, and sensitivity to spurious signals in interconnections.

2. Characterization of Signal and Noise

2.1 Signal

The main features of signals of the two principal kinds of ionization detectors are briefly described here. The elementary current and charge waveforms for gas multiplication proportional detectors and for planar semiconductor detectors are illustrated in Fig. 1. The current results from the motion of charge carriers in the electric field. Therefore, distinctly different waveforms result for the cylindrical geometry of the proportional counter than for the planar geometry of the semiconductor detector.

For point ionization in a proportional detector the electrons reach the multiplication region simultaneously. If the detector is operated in the proportional mode the charge multiplication takes place in the high field region near the center wire (anode). The electrons resulting from the multiplication are collected on the center wire after traversing a very small potential difference. Their contribution to the total observed charge is thus very small. In this case most of the signal is due to the sheath of positive ions moving toward the cathode. The current as a function of time is of the form:[10]

$$i(t) = \frac{Q_m}{2t_o \ln(b/a)} (1 + t/t_o)^{-1} \ , \qquad (1)$$

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and the charge

$$Q(t) = \frac{Q_m}{2\,\ell n(b/a)}\ \ell n\ (1 + t/t_o)\ \text{ for } \frac{t}{t_o} \le \left(\frac{b}{a}\right)^2 . \tag{2}$$

$Q_m = N_o\ A_g\ q_e$, N_o = number of electrons produced by ionization, A_g = gas gain, b and a are cathode and anode radius.[8] t_o is determined by the mobility of positive ions μ_p, b and a, and by the voltage between anode and cathode; $t_o = [a^2\ \ell n\ (b/a)]/2\ \mu_p V_o$.

For extended ionization tracks, the total signal is obtained by superposition of signals due to individual primary electrons arriving at different times to the multiplication region. This affects the early portion of current and charge waveforms, Fig. 1(b), and it is of great significance for timing. The logarithmic nature of the charge as a function of time has consequences in many applications. Where event timing resolution is required, only the early portion of the signal is useful, and thus very little charge is utilized. For measurements based on the quantity of charge (energy) longer filtering times are required. The whole function is rarely observed in practice, and the measurements of gas gain are usually biased due to the low frequency cutoff of filters used.

Multiwire proportional chambers are frequently operated in the saturated gain mode. It has been suggested[2,11] that the gas multiplication takes place far enough from the anode so that the avalanche electrons contribute significantly to the signal by an amount of charge which is collected in a much shorter time (10-20 nsec) than the charge due to positive ions. This mode cannot be used where (linear) information about the charge produced by an ionizing event is required.

Charge collection times in semiconductor position-sensitive detectors are usually very short (< 100 nsec), so that the signal current can be considered as an impulse in most cases.

2.2 Noise

Some simple noise relations are presented here in order to be able to determine the noise contributions by various circuit components. Detectors based on ionization represent a capacitive source of charge. The basic equivalent circuit of detector and amplifier is shown in Fig. 2. The signal is represented as a current source in parallel with the total input capacitance. The series noise is related to the amplification mechanism. The parallel noise is due to imperfections of the amplifier and the detector (leakage currents) and due to dissipative elements (represented by R_p) connected to the input. The noise is expressed in terms of equivalent noise charge. The noise at the input is important only if it contributes to the output of the filter, and the knowledge of the transfer function of the whole system is essential to determine the equivalent noise charge. The whole system from the detector to the output of the filter is described by its weighting function (which is the mirror image in time of the impulse response for time-invariant filters). A more detailed discussion of sources of noise is given in Ref. 12, and signal processing and noise calculations are described in Ref. 13.

We can think of the parallel noise as a random sequence of current impulses (delta functions). Similarly, we can think of the series noise as a random sequence of voltage impulses. The series noise can also be represented by a current generator in parallel with the input C-R network, as shown in Fig. 2(b). The current noise spectrum is obtained upon multiplication by $C_{in}\ (j\omega + 1/\tau_{in})$, where $\tau_{in} = C_{in} R_p$. Consequently each voltage impulse $v_o\delta(t)$ is converted into $v_o C_{in}\left[\delta'(t) + \frac{1}{\tau_{in}}\delta(t)\right]$, the sum of a doublet and an impulse. The mean square equivalent noise charge is obtained by adding up independent contributions to the filter output by all impulses and doublets, according to the Campbell's theorem. (It can be shown that the mean square contributions to the filter output by the pairs of impulses and doublets with the same time origin are uncorrelated, and thus the impulses and doublets due to the series noise can be treated as independent in the summation of their mean square contributions.) The equivalent noise charge for these two noise generators is given by the following relations,

$$\text{Series noise: } \overline{ENC_s^2} = \frac{1}{2}\,\overline{e_n^2}\,C_{in}^2\int_{-\infty}^{\infty}[w'(t)]^2 + \frac{1}{\tau_{in}^2}[w(t)]^2 dt \tag{3}$$

$$\text{Parallel noise: } \overline{ENC_p^2} = \frac{1}{2}\,\overline{i_n^2}\int_{-\infty}^{\infty}[w(t)]^2\ dt, \tag{4}$$

where: e_n = rms eq. series noise voltage per $Hz^{1/2}$

i_n = rms noise current per $Hz^{1/2}$ from parallel sources

$\tau_{in} = C_{in} R_p$ = input circuit time constant

e_n expressed in terms of series eq. noise resistance:

$$\overline{e_n^2} = 4\ kTR_s\ . \tag{5}$$

i_n can be generated as thermal noise in the resistors in parallel with the input,

$$\overline{i_n^2} = 4\ kT\frac{1}{R_p}\ ,\ \text{or} \tag{6}$$

as shot noise in the current I_o into the input (leakage currents, transistor base current, etc.),

$$\overline{i_n^2} = 2\ q_e\ I_o\ . \tag{7}$$

The noise spectrum of the charge amplifier is characterized by the noise-corner time constant, which can be expressed in terms of various noise parameters as,

$$\tau_c = C_{in}\frac{e_n}{i_n} = C_{in}\ (R_s\ R_p)^{1/2}\ , \tag{8}$$

$$\text{or, } \frac{\tau_c}{\tau_{in}} = \left(\frac{R_s}{R_p}\right)^{1/2}\ . \tag{9}$$

The weighting function is normalized to its maximum value so that the impulse signal (charge) from the detector is recorded with a weight of unity. The limits of integration, $-\infty$, ∞, imply that the integration is carried out for all non-zero values of w(t) and w′(t). The noise contribution of a specific part of the weighting function is determined by integration over that part only. The weighting function determines the magnitude and the relative importance of various noise contributions. For inspection of the filtering properties of any weighting function, it is necessary to evaluate the integrals:

$$I_1 = \int_{-\infty}^{\infty} [w'(t)]^2 \, dt = \frac{a_{F1}}{\tau_F} \qquad (10)$$

$$I_2 = \int_{-\infty}^{\infty} [w(t)]^2 \, dt = a_{F2}\tau_F \qquad (11)$$

τ_F is the "width parameter" of the weighting function in time, and a_{F1} and a_{F2} are nondimensional "form factors".

a_{F1} represents the effect of steep parts of the weighting function (the high frequency limit), and it determines the contribution of the series (or "doublet") noise. It is easy to compute these integrals for a given weighting function. For purposes of quick evaluation of noise and of the weighting function, a piecewise linear approximation of the weighting function can be used. A straight line should be fitted to the steepest parts of the function in order to obtain a conservative estimate of the series noise. The values of integrals I_1 and I_2 for various line segments are given in Fig. 3. Weighting functions can be so approximated with an accuracy of noise estimates of better than 10%. For the purpose of inspection of Eq. (3) we assume a unipolar triangular weighting function as shown in Fig. 2, and with $t_1 = t_2 = t_m$. Then the two integrals are

$$I_1 = \frac{2}{t_m} \qquad (12)$$

$$I_2 = \frac{2}{3}\, t_m \qquad (13)$$

Eqs. (3) and (4) can be written in the form,

$$\text{Series noise: } \overline{ENC_s^2} = \frac{\overline{e_n^2}\, C_{in}^2}{t_m}\left[1 + \frac{1}{3}\left(\frac{t_m}{\tau_{in}}\right)^2\right] \qquad (14)$$

$$\text{Parallel noise: } \overline{ENC_p^2} = \frac{1}{3}\,\overline{i_n^2}\, t_m \qquad (15)$$

If $t_m/\tau_{in} \ll 1$, Eq. (14) reduces to the classical case of the charge amplifier (R_p is very large),

$$\overline{ENC_s^2} = \frac{\overline{e_n^2}\, C_{in}^2}{t_m} \qquad (16)$$

In a number of instances with position-sensitive detectors $t_m/\tau_{in} \gg 1$. Let us assume, as an example, that $t_m/\tau_{in} = 2$. Then the two terms in Eq. (14) are:

$$\overline{ENC_s^2} = \frac{\overline{e_n^2}\, C_{in}^2}{t_m}(1 + 4/3) = 2.33\ \left(\overline{ENC_s^2}\right)_{min} \quad . \qquad (17)$$

In addition R_p generates parallel noise:

$$\overline{ENC_p^2} = \frac{1}{3}\left(\frac{t_m}{\tau_c}\right)^2 \frac{\overline{e_n^2}\, C_{in}^2}{t_m} \qquad (18)$$

Thus, if τ_{in} is smaller than the width of the weighting function, there is a significant increase in the contribution by the series noise, simply because the signal is attenuated before it even arrives at the amplifier.

The lowest noise can be achieved with no dissipative elements connected to the input, Eq. (16). The filtering or "pulse shaping" should be performed after amplification, so that no noise is added by dissipative filter components. The current from the detector can be measured by shaping (differentiating) the signal after amplification.

3. Transmission Line Termination with Electronically Cooled Resistance

The need for damping arises in three typical cases: 1) there is a resonant circuit in the input, or a transformer in conjunction with a capacitive signal source, and an aperiodic response is required; 2) waveform measurement from a capacitive signal source by "differentiation" of the charge signal at the input; and 3) transmission line termination. These cases are illustrated in Fig. 4(a). It is a common practice to use a resistor to achieve the required damping. In all these cases the required damping or terminating resistor is the only dissipative element in the input circuit, and its noise exceeds the amplifier noise.

It has long been recognized that damping can be achieved with resistance in feedback, for example to damp a galvanometer,[14] where the noise added by the "active resistance" is less than it would be by the physical resistor. Referring to Fig. 4(b), an inverting amplifier with a real gain G_o and a resistance R_f in feedback results in an apparent input resistance,

$$R_{in} = \frac{R_f}{|G_o|+1} \approx \frac{R_f}{|G_o|} \qquad (19)$$

R_f represents a noise source in parallel with the input, and its noise current is much smaller than it would be from a resistor of value R_{in}, Eqs. (4) and (6).

This technique is limited to low frequencies because the amplifier gain should be real. It also poses stability problems at higher frequencies, due to the phase shift in the amplifier and in the combination of the feedback resistance and the capacitance at the input.

The new approach is based on exploiting the natural properties of an elementary operational amplifier. For such an amplifier the gain expressed as a function of frequency is given by

$$G(j\omega) = -\frac{\omega_h}{j\omega} \quad , \tag{20}$$

where ω_h is angular frequency at which the gain becomes unity. It is assumed that any additional poles correspond to frequencies higher than ω_h, and that the gain reaches its d.c. value at a frequency lower than the lowest frequency of interest. As is well known for amplifiers where the gain is inversely proportional to frequency ("6 dB per octave"), there must be a 90^o phase lag associated with this dependence, according to Bode's law. This phase lag must be offset by a 90^o phase lead in feedback in order to achieve a real input impedance (resistance). Thus a capacitance is required in feedback in order to realize a damping resistance. Necessary relations can be derived from the basic feedback circuit in Fig. 4(c). It is simple to show that the input impedance is given by

$$Z_{in} = \frac{Z_f}{(-G+1)} \approx \frac{Z_f}{-G} \quad , \tag{21}$$

(where Z_{in}, Z_f and G are functions of either $j\omega$ or the Laplace transform variable s). We assume a capacitance in feedback, as shown in Fig. 4(d),

$$Z_f = \frac{1}{j\omega C_f} \quad . \tag{22}$$

The input impedance is then,

$$Z_{in} = \frac{1}{\omega_h C_f} = R_{in} \quad , \tag{23}$$

i.e., the input resistance is equal to the absolute value of the reactance in feedback at the unity-gain frequency of the amplifier. Thus, damping can be realized without connecting any dissipative elements to the input. The damping cannot be noiseless since it is realized by a physical amplifier. It is interesting to find the effective noise temperature of the damping resistor realized by feedback. Due to the feedback the equivalent series noise voltage appears at the input terminal,

$$v_n = -\frac{e_n}{1-\frac{1}{G}} \approx -e_n \quad .$$

Thus the open circuit mean square noise voltage per $H_z^{1/2}$ (the physical spectral density) at the input terminal is

$$\overline{v_n^2} = 4\,kTR_s \quad . \tag{24}$$

Since it appears to be generated by the source with an internal resistance R_{in}, it can be expressed in the form

$$\overline{v_n^2} = 4k\,\left(T\frac{R_s}{R_{in}}\right) R_{in} \tag{25}$$

or in the form of a current source,

$$\overline{i_n^2} = \overline{v_n^2}/R_{in}^2 = 4k\,\left(T\frac{R_s}{R_{in}}\right)\frac{1}{R_{in}} \quad . \tag{25a}$$

Thus the effective temperature of the damping resistance R_{in} is

$$T_f = \frac{R_s}{R_{in}}\,T \quad . \tag{26}$$

The improvement in noise over the physical ("warm") damping resistor depends on its value. The equivalent series noise resistance for field-effect transistors and bipolar transistors is in the range 50 - 200 Ω. If a low value of R_{in} is required, a transformer can be used. A high value of R_{in} is realized by feedback, and then transformed to the required low value R_{in1}, Fig. 4(e). The effective temperature is then,

$$T_f = \frac{R_s}{n^2 R_{in1}}\,T \quad . \tag{27}$$

The upper limit to the cooling factor $n^2 R_{in1}/R_s$ is determined by the practical considerations in the transformer design (low parasitic inductance and capacitance, low noise).

A basic circuit configuration for cooled damping with capacitance in feedback is shown in Fig. 5. This is a complementary cascode used in charge amplifiers.[12] The open loop gain is

$$G = g_m\,\frac{1}{j\omega C_o} \quad , \tag{28}$$

and the input resistance becomes

$$R_{in} = \frac{1}{g_m}\,\frac{C_o}{C_f} \quad . \tag{29}$$

In this discussion it was assumed that the gain is given by Eq. (20) and that in the region of interest $|G(j\omega)| \gg 1$. We now consider the complete expression for all frequencies. Assuming negligible loading of the amplifier output by the feedback network, the open loop gain is given by

$$G(j\omega) = -\,g_m\,\frac{1}{j\omega C_o + 1/R_o} \quad . \tag{30}$$

($1/R_o$ is the output conductance of the current source and the input conductance of the unity gain amplifier in Fig. 5). The input impedance for a capacitance in feedback ($R_f \to \infty$, $R_{fs} = 0$) is then,

$$Z_{in} = \frac{1}{g_m}\,\frac{C_o}{C_f}\left[(1 + 1/j\omega C_o R_o)^{-1} + j\omega/\omega_h\right]^{-1} \tag{31}$$

At frequencies where $\omega C_o R_o >> 1$,

$$Z_{in} \approx \frac{1}{g_m} \frac{C_o}{C_f} (1 + j\omega/\omega_h)^{-1} \quad , \tag{32}$$

or $Z_{in} \approx \left(\frac{1}{R_{in}} + j\omega C_f\right)$, where R_{in} is given by Eq. 29, and $\omega_h = g_m/C_o$. The term $j\omega C_f$ represents only the feedback capacitance. The important component of input impedance is the one due to the feedback,

$$Z_{in} = \frac{1}{g_m} \frac{C_o}{C_f} + \frac{1}{j\omega C_f(g_m R_o)} \quad . \tag{33}$$

This is a series combination of the resistance R_{in} and the capacitance $C_f g_m R_o$. In the frequency region of interest for damping the second term should be negligible, i.e. $\omega C_o R_o >> 1$.

We note parenthetically that Eq. (33) represents input impedance of the "operational integrator" or charge amplifier. It has not been generally recognized that in most cases the first term dominates.

It is assumed here that the biasing resistor R_f is large, so that its noise is negligible. It is interesting to note that R_f appears at the input as an inductance. By using Eq. (21), it follows:

$$L_{in} = \frac{C_o}{g_m} R_f = \frac{R_f}{\omega_h} \quad . \tag{34}$$

This is another way of interpreting the periodic impulse response of an operational amplifier with resistance in feedback and capacitance at the input. Damping resistance can be realized at all frequencies from dc to beyond ω_h by making $R_f C_f = R_o C_o$. This eliminates the imaginary term in Eq. (33).

The feedback network can be used to compensate for the effect of parasitic input capacitances. With R_{fs} in series with C_f in feedback,

$$Z_{in} = \frac{1}{C_f \omega_h} + j\omega \frac{R_{fs}}{\omega_h} \tag{35}$$

The second term represents an inductance in series with the resistance R_{in}, and it can be used for "high frequency peaking" in transmission line terminations.

It was assumed throughout that additional poles correspond to frequencies higher than ω_h ($\omega_1 \geq 4\ \omega_h$), which can be achieved with high frequency transistors. An aperiodic response can be achieved more easily with fewer additional poles in the loop, if a short rise time is required. In such a case feedback is taken from the output of the cascode, as shown in Fig. 5, rather than from the output of the unity gain amplifier.

The output rise time for impulse signals is determined by the time constant,

$$\tau_r = R_{in} C_{in} = \frac{1}{g_m} \frac{C_o}{C_f} C_{in} \tag{36}$$

as is well known for charge amplifiers.[12]

Damping or resistive termination by capacitance in feedback results in an amplifier with a real input impedance which integrates the input current. Further pulse shaping (filtering) should take into account this integration. The current waveform can be obtained by differentiation.

4. Position Sensing with Delay Lines

Position sensing with delay lines is based on conversion of position information into a time delay. In this method position sensing electrodes of the detector are connected at uniform spacing to a delay line. The signal can be coupled either capacitively, or directly.[2,3,15] Direct coupling can be achieved also by making the detector electrodes a part of the helical coil of the delay line.[2,3] If cathodes of a multiwire proportional chamber, or strips on either side of a semiconductor detector, are connected to the delay line, the well known interpolating property results. The image of the signal charge is seen by several electrodes, and by determining the centroid of the composite signal propagating along the delay line, the position can be determined with an error smaller than the wire spacing or the projected length of an inclined particle track. This method, where the position appears as the difference between the arrival times of the signals at the ends of the delay line, is illustrated in Fig. 6(a).

The delay line is a non-dissipative position sensing medium in principle. The noise in this system is generated in the terminations and in the amplifiers. The position resolution is ultimately limited by the statistics of the spatial charge distribution in the detector. The electronic noise determines the magnitude of the signal required to achieve this resolution.

We derive first the expression for noise in the conventional system with "warm" resistance terminations. The signal and noise sources for the circuit in Fig. 6(a) are shown in the equivalent circuit in Fig. 6(b). We use the results of the analysis in Section 2. With a terminated delay line the impedance of the input circuit is real, and clearly the time constant $\tau_{in} = C_{in} Z_o/2$ will be much smaller than the width of the weighting function. In this case the first term in the expression (3) for series noise becomes negligible, and we have for the noise from the amplifier,

$$\overline{ENC_s^2} = \frac{1}{2} \overline{e_n^2} \frac{1}{(Z_o/2)^2} I_2 \doteq 8\ kT \frac{R_s}{Z_o^2} I_2 \quad . \tag{37}$$

The noise from both terminations is in parallel with the input. Using Eqs. (4) and (6) with the values from Fig. 6(b), the noise from the terminations is:

$$\overline{ENC_p^2} = 4\ kT \frac{1}{Z_o}\ I_2 \quad . \tag{38}$$

The total noise for "warm terminations":

$$\overline{ENC^2} = 4\ kT \frac{1}{Z_o}\ (1 + 2\frac{R_s}{Z_o}) \int_{-\infty}^{\infty}[w(t)]^2\ dt \tag{39}$$

We now consider "cooled" terminations introduced in the preceding section. The delay line and the preamplifier circuit is then as in Fig. 6(c). Referring to Eqs. (3), (10) and the equivalent circuit in Fig. 2, we have again $\tau_{in} \ll \tau_F$, and we substitute Z_o for R_p. The remaining second term in Eq. (3) represents the equivalent noise charge due to the series noise of one amplifier,

$$\overline{ENC_{s1}^2} = \frac{1}{2}\ \overline{e_n^2}\frac{1}{Z_o^2}\ \ I_2 = 2\ kT\frac{R_s}{Z_o^2}\ \ I_2 \quad . \tag{40}$$

The active termination at the other end adds an equal amount of noise due to the series noise of its amplifier. As shown in the preceding section, a transformer can be used if a low value of input resistance is required. This is equivalent to dividing $\overline{e_n^2}$ by n^2. The total equivalent noise charge is then,

for electronically "cooled" terminations:

$$\overline{ENC_s^2} = 4k\left(T\frac{R_s}{n^2 Z_o}\right)\frac{1}{Z_o}\int_{-\infty}^{\infty}[w(t)]^2\ dt \quad . \tag{41}$$

The ratio of noise temperatures for the two cases gives the improvement factor due to cooling,

$$\eta_f^2 = \frac{T_w}{T_f} = \frac{n^2 Z_o}{R_s}\left(1 + 2\frac{R_s}{Z_o}\right) \tag{42}$$

Z_o = characteristic impedance of the delay line,
R_s = eq. series noise resistance of the amplifier,
n = transformation ratio between the delay line and the amplifier.

η_f represents the reduction in rms noise. The signal amplitude can be decreased by η_f while giving the same position resolution as with "warm" terminations. Let us assume, as an example, Z_o = 1000 Ω, n = 1, R_s = 50 Ω. Then

$$\eta_f \geq \sqrt{22} = 4.7$$

R_s is determined by the input amplifying device. A higher ratio can be achieved for delay lines with higher characteristic impedance Z_o. For low impedance delay lines a transformer is necessary. (Practical improvement factor for existing systems with warm terminations may be higher since in many cases the amplifiers used have not been designed for low noise.)

What is the relative position resolution with the delay line position sensing? The resolution is determined by several factors: the detector design, particle type, energy, angle of incidence, etc. As the length of the delay line is increased, the dispersion in the delay line limits the timing resolution, and the resolution becomes determined largely by the signal-to-noise ratio. The best resolution in this case is achieved with an optimum filter for timing. This filter should be matched to both the signal and noise. The noise in this case is white. The optimum filter for timing is the derivative of the optimum filter for amplitude measurement. The optimum filter for amplitude measurement is in this case (with white noise) a mirror image in time of the signal waveform from the delay line. This is illustrated in Fig. 7 on a piecewise linear approximation of an odd current function. The optimum filter for timing results from minimization of the ratio of noise to the slope of the signal at the output of the filter. The filter weighting function is generally bipolar and the maximum slope is at zero crossing, which results in well-known antiwalk properties. As is known from the filter theory, the minimum ratio for a given signal and noise is given by,

$$\frac{\overline{v_n^2}}{(d\ v_{out}/dt)^2} = (\delta t)^2 = \frac{W_o}{A^2\int_{-\infty}^{\infty}[f'(t)]^2\ dt} \quad , \tag{43}$$

where W_o = white noise spectral density, A = signal amplitude, and f(t) is the signal waveform.

Let us assume that the current signal at the output of the delay line is approximated by a gaussian, described by the total charge and by the full width at half maximum,

$$i(t) = \frac{2.35}{\sqrt{2\pi}}\ \frac{Q_s}{t_{FW}}\ \exp\left[-2.76\ \left(\frac{t}{t_{FW}}\right)^2\right] \quad . \tag{44}$$

The optimum weighting function is a derivative of the gaussian. The noise current spectral density in the case of cooled terminations is determined from the relations (40) and (41) assuming no filtering, i.e. $I_2 = 1$. Then we can calculate the variance in timing from eqs. (43) and (44) as

$$\overline{(\delta t)^2} = 0.55\ \frac{\overline{e_n^2}\ t_{FW}^3}{Z_o^2\ Q_s^2} \quad . \tag{45}$$

We can define the quality factor of the delay line as the ratio of the total delay to the time dispersion t_{FW},

$$\Theta_D = \frac{t_D}{t_{FW}} \quad . \tag{46}$$

The relative position resolution is determined by the timing error from the two outputs. The noise from the two outputs is uncorrelated for timing, since the noise from one amplifier is added with a delay t_D to the other one.

Thus,

$$\frac{\delta \ell}{\ell} = \sqrt{2}\,\frac{\delta t}{t_D} \quad . \tag{47}$$

With (45) and (46) it follows:

$$\frac{\delta \ell}{\ell} = 2.46\,\frac{1}{\Theta_D}\,\frac{e_n}{Z_o Q_s}\,t_{FW}^{1/2} \quad [\text{FWHM}] \quad . \tag{48}$$

Let us assume, as an example, $\Theta_D = 15$, $Z_o = 500\ \Omega$, $e_n = 10^{-9}\ \text{V/Hz}^{1/2}$, $t_{FW} = 10^{-7}$ sec.

To achieve a resolution $\delta\ell/\ell = 10^{-3}$, a signal is required, $Q_s = 1.04 \times 10^{-13}$ C $= 6.5 \times 10^5$ e. Without electronic cooling a signal of about 3×10^6 e would be required.

Cooled terminations are simple to realize. They involve only a judicious use of a few amplifier components which had to be used also in the case of warm terminations. An amplifier used as a cooled termination with a multiwire proportional detector[15] is shown in Fig. 8. The amplifier is intended for low impedance delay lines, $Z_o = 200$ and $500\ \Omega$. Two bipolar transistors are used in parallel to reduce the effect of base and emitter bulk resistances and achieve $R_s \approx 30\ \Omega$. Above $Z_o = 500\ \Omega$ field-effect transistors should be used, since the base current noise becomes comparable to the (series) emitter current noise. An amplifier for damping in the range above 500 Ω is shown in Fig. 9. The value of R_{in} is adjusted by trimming C_o. g_m is maintained constant by the dc feedback control of the drain (collector) current. In this analysis we have assumed that the noise due to the losses in the delay line is negligible. We found this to be the case for the lumped parameter lines we measured. The noise measurement can easily be performed in the system according to Fig. 6(c). Delay line is replaced by a short connection between the amplifiers, which leaves the same relation of impedances in the circuit. There should be no difference in noise if the delay line noise is negligible.

5. Position Sensing with Resistive Electrodes

Resistive eletrodes have been used in various forms for position sensing in gas proportional detectors and in semiconductor detectors. There are two basically different methods for position sensing with resistive electrodes. Due to the electrode capacitance the resistive electrode represents in both cases a diffusive R-C line. The two methods are illustrated in Fig. 10. In the charge division method[4-7, 17,18] the position is determined from the ratio of charge flowing out of the ends of the resistive electrode terminated into low impedances. In the so-called rise time method[7,8,9] the resistive electrode represents a part of an infinite diffusive line (if it is terminated into its characteristic impedance). The position is determined from the difference in the signal diffusion time from the point where the charge is collected. In the charge division method the electrode capacitance is a nuisance - it limits the resolution for all variables of interest: energy, position and event arrival time as well as the resolving time. In the diffusion time method the electrode capacitance is an essential element. With the electrode resistance it determines the relation of the diffusion time to the position. In both methods signal processing in relation to R-C line parameters is critical in order to achieve the best position resolution, linearity of the output signal with the particle position, and maximum event rates. We outline here an analysis from which the optimum signal processing and the limits of performance can be determined.

All the performance parameters can be expressed in terms of the R-C line parameters,

$$\begin{aligned} R_D &= R\,\ell \ , \\ C_D &= C\,\ell \ , \\ \tau_D &= R_D\,C_D = RC\,\ell^2 \ , \end{aligned} \tag{49}$$

where R and C are resistance and capacitance per unit length, ℓ is the length of the line, and τ_D = the product of the total line resistance and capacitance is the line "time constant."

We first consider the requirement for linearity of the output signal with particle position. Due to the diffusive nature of the line, it takes time after a charge impulse is delivered until the ratio A/(A+B) $\approx x/\ell$ is established.

The charge at one end of the electrode (connected to zero impedance at both ends) as a function of time and position x/ℓ is given by,[7]

$$\frac{Q(t,x)}{Q_s} = 1 - \frac{x}{\ell} - \sum_{m=1}^{\infty} \frac{2}{m\pi}\,\text{Si}\left(\frac{m\pi x}{\ell}\right) \exp\left(-\frac{m^2\pi^2}{\tau_D}\,t\right) \quad . \tag{50}$$

The dominant time constant of the series for transient phenomena in the line is τ_D/π^2. For linear relation between Q(t,x) and the position, the sum in Eq. (50) should be negligible. The time required for the position nonlinearity to be less than 0.2% is,

$$t \geq \frac{1}{2}\,\tau_D \quad . \tag{51}$$

Since the signal waveform Q(t,x) is position-dependent, we have to consider the problem of variable ballistic deficit in the selection of optimum filtering. In such a case a trapezoidal weighting function[16] gives the best signal-to-noise ratio with minimum resolving time. The form of this function is determined by the relation of the noise generated in the resistive electrode and the amplifier (series) noise.

For $t \geq \frac{1}{2}\tau_D$ the admittance of the resistive electrode shorted at one end can be approximated by two components,[7]

$$Y = \frac{1}{R_D} + j\omega\,\frac{C_D}{3} \quad . \tag{52}$$

Then the equivalent circuit in Fig. 2 can be used for an approximate noise analysis, where R_D generates the "parallel noise" and CD/3 determines the contribution by the amplifier series noise. The noise corner time constant is for this case,

$$\tau_C = \left(\frac{C_D}{3} + C_A\right)\left(R_D R_s\right)^{1/2} \tag{53}$$

Since the diffusive line represents a dissipative position sensing medium, and its noise is inherently present with the signal, we should at least not add any other noise. The make the amplifier series noise negligible, one can conclude from relations (3), (4), (10) and (11) that the following conditions should be satisfied:

for noise sources: $$\frac{R_s}{R_D} \ll 1 \tag{54}$$

for the filter weighting function: $$\left(\frac{I_2}{I_1}\right)^{1/2} \gg \tau_C \tag{55}$$

The condition (55) emphasizes that the weighting function must not have infinitely steep parts. On the basis of this condition we can determine the ratio of the sloped parts to the flat part for the filter weighting function for a given set of values R_s, C_A, R_D, C_D. Conversely, we can assume a trapezoidal weighting function with a minimum easily realizable ratio of the sloped parts to the flat part, and determine the condition for the detector and amplifier parameters. Referring to Fig. 3, we assume for an even, unipolar, trapezoidal function, $t_1 = t_3 = 0.2\ \tau_F$, $t_2 = 0.6\ \tau_F$, $b = 0$, where τ_F is the base width of the trapezoid. Then $I_1 = 10/\tau_F$, $I_2 = 0.733\ \tau_F$. The condition (55) with (53) becomes then a stricter condition for R_s/R_D,

$$\frac{R_s}{R_D} \ll 0.42\ (1 + 3\ C_A/C_D)^{-2} \quad . \tag{56}$$

The resistive electrode generates noise in parallel with the input, Eqs. (4), (6) and (11),

$$\overline{ENC_p^2} \approx 2\ kT \frac{1}{R_D}\ a_{F2}\tau_F \qquad \text{for } \tau_F \geq \frac{1}{2}\tau_D. \tag{57}$$

For the trapezoidal function assumed, $a_{F2} = 0.733$, and

$$\overline{ENC_p^2} \approx 1.47\ kT \frac{1}{R_D}\ \tau_F \qquad \text{for } \tau_F \geq \frac{1}{2}\tau_D \quad . \tag{57a}$$

If we set $\tau_F = 0.8\ \tau_D$ so that the flat top of the trapezoidal function is equal to $1/2\ \tau_D$, we get for the minimum noise from the resistive electrode,

$$\overline{ENC_p^2} \approx 1.17\ kT \frac{\tau_D}{R_D} = 1.17\ kT\ C_D \quad . \tag{58}$$

This noise adds to the position signal Q_A (the noise in $Q_A + Q_B$ is much smaller since for $\tau_F > \tau_D/2$ it is anti-correlated at the two ends of the electrode). The optimum position resolution (for position non-linearity < 0.2%) is then,

$$\frac{\Delta\ell}{\ell} \approx 2.35 \frac{ENC_p}{Q_s} = 2.54 \frac{(kTC)^{1/2}}{Q_s} \qquad \text{(FWHM)} \tag{59}$$

<u>The position resolution for the charge division method with optimum filtering is determined only by the electrode capacitance and not by the electrode resistance.</u> The resolving time is determined by both the electrode capacitance and the resistance, and it is equal to the width at the base of the weighting function, $\tau_F = 0.8\ \tau_D$.

In addition to the position resolution two other parameters are frequently of interest. These are the noise in the energy signal (the sum $Q_A + Q_B$) and the position dependent timing walk in the sum signal. The equivalent circuit for the common electrode in Fig. 10(c), and for the sum signals in Fig. 10(a) can be approximated by the detector capacitance C_D in series with a resistance $R_D/12$ (Ref. 7). The noise contribution to the sum signal by the resistive electrode can be determined from the equivalent circuit in Fig. 2 and from Eqs. (3) and (5), substituting $R_D/12$ for R_s, C_D for C_{in} and $R_p \to \infty$. This noise contribution increases with both R_D and C_D, and it <u>decreases</u> with the width of the weighting function. The optimum weighting function for the energy measurement has a narrower flat part and longer sloped parts than the optimum function for the position measurements. The position-dependent timing walk for the timing signal derived from the sum signal is about $\tau_C/10$.

The lower limit for τ_D is determined by the charge collection time in the detector. The charge division method is insensitive to charge collection time variations, and as can be shown, the charge collection time can be longer than τ_D and/or τ_F, while preserving the position signal linearity. The only consequence is a smaller signal in relation to the noise, and thus poorer position resolution. For the case where τ_D is shorter than the charge collection time, τ_F can be increased from the optimum value for an impulse signal, to optimize the position resolution.

For a given (smallest possible) capacitance C_D, R_D should be reduced until $\tau_D/2$ is equal to the detector charge collection time. This is in order to achieve the shortest resolving time (higher event rates), better energy resolution, and smaller timing walk.

We take as an example a semiconductor detector with a resistive electrode. With $\ell = 5$ cm, $C = 50$ pF, the charge required for a position resolution $\Delta\ell/\ell = 10^{-2}$ is $Q_s \geq 6.5 \times 10^5$ e, and $ENC_p = 3 \times 10^3$ rms e. With a charge collection time of less than 100 nsec, τ_D should be reduced to about 200 nsec; i.e., $R_D = 4 \times 10^3\ \Omega$. The filter, as discussed previously, is a trapezoidal filter with base width $\tau_F = 0.8\ \tau_D = 160$ nsec. Such a filter can be approximated by single delay line clipping and RC integrating networks.

As another example we take a proportional detector with single wire with capacitance C = 10 pF. In order to achieve a resolution $\Delta\ell/\ell = 10^{-3}$, a signal charge $Q_s \geq 2.8 \times 10^6$ is required. A careful inspection of the charge collection waveform is required to determine if the signal-to-noise ratio improves enough to warrant an increase in τ_F. This is due to the logarithmic nature of the charge collection with time, Fig. 1. If R_D is given (by the construction limitations of the detector) then for point ionization, an increase of τ_F above 0.8 τ_D would result in poorer resolution. In some cases of extended ionization (high pressure neutron detectors) the signal increase may exceed the noise increase. There is presently a gap in resistance values between metal alloy wires and carbon coated quartz filaments. A 1m long metal alloy wire may have a resistance of 2 to 4 x $10^3\ \Omega$, if it is to be of adequate diameter for the detector operation and for the strength. With $C \approx 10$ pF, this would give τ_D = 20 to 40 nsec. Such a detector would be capable of handling event rates of $10^6\ \text{sec}^{-1}$ with 2 to 4% pileup probability.

Some properties of the diffusion time method have been described previously.[8,9] An exact comparison of all the properties of the two methods has not been made as yet. There is some experimental evidence that the position resolution as limited by noise is the same. The resolving time for the charge division method, as shown in this section, is about 0.8 τ_D for optimum position resolution and linearity. The resolving time for the diffusion time method, as achieved in practice,[8,9,19] is several times τ_D. There are some practical advantages in each of the methods. The charge division requires a ratio digitizer, while the diffusion time method requires only a time digitizer. The former requires a simple preamplifier with low input impedance. These preamplifiers can be some distance away from the detector, and the connecting transmission lines can be terminated by the input resistance of the preamplifiers (as shown in Sections 3 and 4). The diffusion time method requires R-C line terminations into its characteristic impedance, and preamplifiers with high impedance and short connections to the detector. The diffusion time method requires uniformity of both the resistance and capacitance in the design of the detector.

The difficulties of performing division have been somewhat exaggerated in the literature. A fast and accurate (logarithmic) divider has been constructed and operated.[17] Another approach, first proposed by Miller, et al.,[18] is based on a dividing analog-to-digital converter. With this approach the circuits are likely to be simpler than with a separate divider.

6. Position Sensing for Multielectrode Detectors with One Amplifier per Electrode

In readouts for multiwire proportional chambers based on one amplifier per wire, and where only position information is required, it is usually assumed that the signal is large and that electronic noise in the amplifier is not a problem. Due to cost and size considerations in the design of these amplifiers, their charge sensitivity is low. Since noise is not of concern in such applications, the circuit techniques used result usually in higher noise than justified. The problem of inadequate sensitivity arises in cases where the signal magnitude is limited by the detector design (smaller anode spacing). A number of applications have become of interest where linear signal information is needed (detection of transition radiation, identification of particles by ionization loss, x-ray detectors in astrophysics). In all cases involving semiconductor detectors and gas or liquid ionization detectors with little or no gain, charge sensitivity and noise become important.

What is the lowest noise one can achieve? Assuming detectors with a low electrode capacitance (10 - 100 pF) no transformers can be used in practice for detector amplifier matching. The limit is determined by the series noise of the amplifying device, assuming that the noise from parallel sources is negligible. The equivalent noise charge is given by Eq. (16),

$$\overline{ENC_s^2} = \frac{\overline{e_n^2}\, C_{in}^2}{t_m} \tag{60}$$

An optimum weighting function was assumed. In this case this is a triangle with base width 2 t_m. In practice, a gaussian function can be achieved, with a small increase in noise for equal width. We assume C_{in} = 25 pF for the electrode (wire) + amplifier capacitance and a short resolving time t_m = 25 nsec. For better bipolar transistors and field-effect transistors $e_n \approx 10^{-9}\ \text{V/Hz}^{1/2}$. Then

$ENC_s \approx 10^3$ rms e.

The equivalent noise charge can be reduced by increasing the resolving time. For $t_m = 0.5\ \mu$sec, we obtain $ENC_s \approx 220$ rms e.

The next question is, can one use a bipolar transistor or must one use a field-effect transistor. This is determined by the relation of the resolving time to the noise corner time constant. To make the parallel noise negligible, the following condition must be satisfied.

$$t_m < \frac{1}{2}\tau_c = C_{in}\,(R_s R_p)^{1/2} \tag{61}$$

For the above example with t_m = 25 nsec, this requires that $R_p \geq 5 \times 10^4\ \Omega$. For the example with $t_m = 0.5\ \mu$sec, the noise resistance in parallel must be much higher, $R_p \geq 5 \times 10^6\ \Omega$. To be able to compare the noise contribution by the base current shot noise to that of a biasing resistor, we can use the "50 millivolt rule." It can be derived from Eqs. (6) and (7). If the shot noise due to a current with mean value I_o is equal to the thermal noise from resistance R_p at T = 295°K, then the "50 mV rule" for equal noise from R_p and I_o: $R_p I_o \approx 50$ mV. (62)

For a base current $I_{Bo} = 10^{-5}$ A, $R_p = 5 \times 10^4\ \Omega$, $R_p I_o \approx 0.5$ V, which means that the base current noise is higher than the noise from R_p and that it does not satisfy the condition (61).

Thus even at very short resolving times field-effect transistors give better noise performance. If ultimate noise performance is not required; bipolar transistors can be used. The noise can be determined from the relations (3) and (4). In general, with bipolar transistors the width of the weighting function should not be larger than necessary for processing of the signal otherwise an increased noise due to the parallel sources would result.

How simple can a low noise amplifier be? Fig. 9 illustrates a design of a low noise amplifier for high resolution spectrometry. However, if the requirements on the linearity, dynamic range and open loop gain are somewhat relaxed, simpler low noise circuits result. The simplest feedback charge amplifier is based on the configuration in Fig. 5, and it employs three transistors, as shown in Fig. 11(a).

The next question of current interest with multiwire proportional chambers is whether the preamplifiers can be remote to the detector. The equivalent noise charge increases linearly with capacitance at the input, so that much higher noise would result. The amplifier noise may still be sufficiently small if large signals are available. The main problem in this case is pickup noise from the surrounding electrical equipment. Ground loop decoupling and common mode rejection are ineffective with high impedance sources, since these are difficult to balance, and the noise signal induced is not shorted in the source. A circuit configuration for "minimum electronics at the detector" and with common mode rejection is shown in Fig. 11(b). The eq. series noise resistance of the input device is increased by only about 100 Ω. The circuit also illustrates transmission line termination with a transistor in the common base configuration.

Remote amplifiers usually require transmission line termination. Instead of wasting the signal in the termination resistance, the input impedance of the amplifier can be used. One method is based on the feedback amplifier with resistive input impedance ("cooled" termination), Fig. 1(c), as described in Section 3. The other is based on a common base configuration, Fig. 11(d). In both cases a series sending end terminating resistor reduces reflections at high frequencies.

7. Discussion

The most interesting results are the resolution limitations due to noise in the position-sensing methods with delay lines and with resistive electrodes, Eqs. (48) and (59). It appears that the position sensing with delay lines requires a smaller signal for equal relative position resolution. An important condition for this is that "cooled terminations" should be used to achieve the improvement given by Eq. (42). This conclusion applies to the direct coupling to the delay line. With capacitive coupling to a delay line, a substantial attenuation of the signal may occur. With "warm" resistance terminations and amplifiers not optimized with respect to noise, the advantage of delay line position sensing may disappear. An improvement, as with "cooled" terminations, is important since it extends the useful operating range of proportional detectors, and it may make possible simultaneous position and ionization loss measurements.

It should be emphasized that these two methods are different and not applicable in all cases, and that each poses different requirements on the electrode configuration and construction of the detector. The same applies to the position sensing methods as to the detectors - no single solution satisfies all different applications. The analysis presented is concerned with noise limitations. In some cases a more detailed consideration of the charge collection statistics may be required. For example, with delay line position sensing, optimum filtering may be dependent on the particle track inclination. In general, methods based on timing are more dependent on the charge collection statistics than is the charge division method.

One subject this paper avoids carefully is any comparison of readout methods for particular applications. The choice of the detector and the readout is primarily a question of optimum experiment design and then of the system design. To mention a few variables entering in such considerations: single event or multiple event detection, event rate, particle type and density of ionization, track inclination, detection efficiency, timing and energy resolution requirements, and last but not least, cost distribution among the detector, readout electronics and data processing software.

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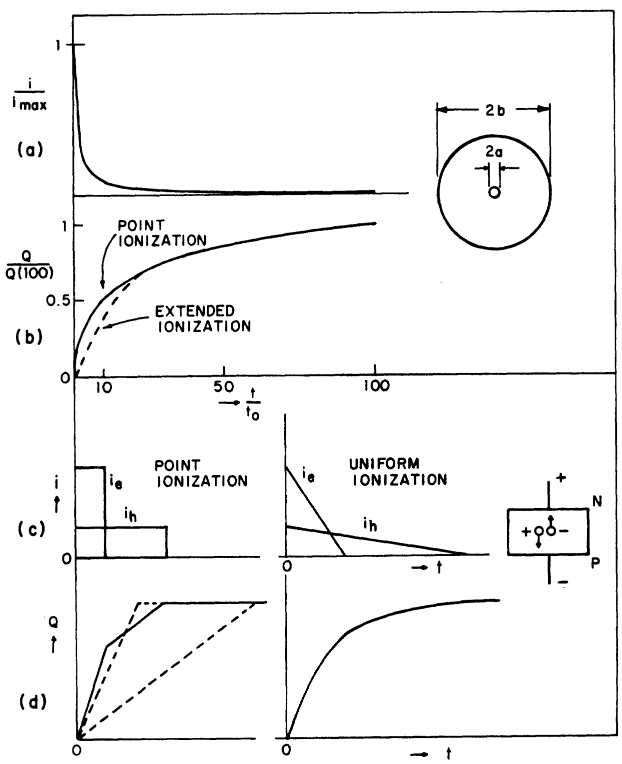


Fig. 1. Signal waveforms.

Proportional detector:

(a) current for point ionization

(b) charge for point ionization (solid) and for extended ionization (dashed)

Planar semiconductor detector (for point ionization in the middle and for uniform ionization between the electrodes):

(c) current

(d) charge

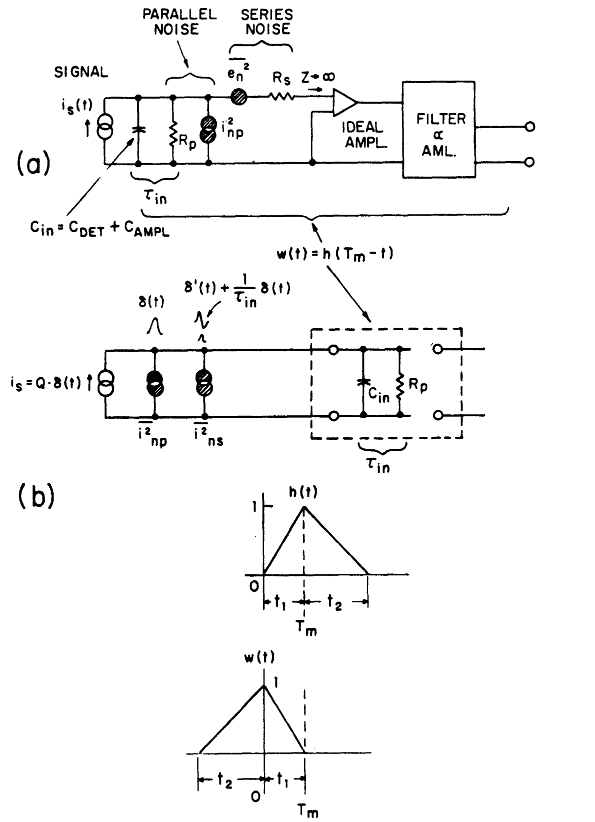
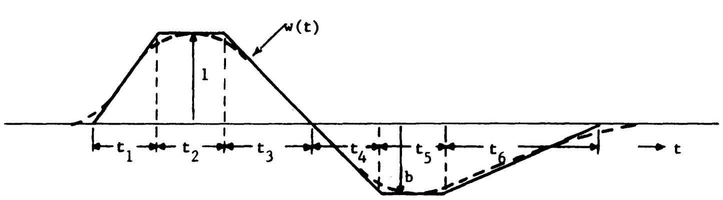


Fig. 2. Equivalent circuit of detector and amplifier for noise analysis.

(a) location of noise sources in the circuit

(b) eq. circuit with noise sources transformed to the input

C_{in} = detector capacitance + amplifier input capacitance.



$$I_1 = \frac{1}{t_1} + 0 + \frac{1}{t_3} + b^2\frac{1}{t_4} + 0 + b^2\frac{1}{t_6}$$

$$I_2 = \frac{t_1}{3} + t_2 + \frac{t_3}{3} + b^2\frac{t_4}{3} + b^2 t_5 + b^2\frac{t_6}{3}$$

Fig. 3. Evaluation of weighting function integrals, Eqs. (10) and (11) by piecewise linear approximation.

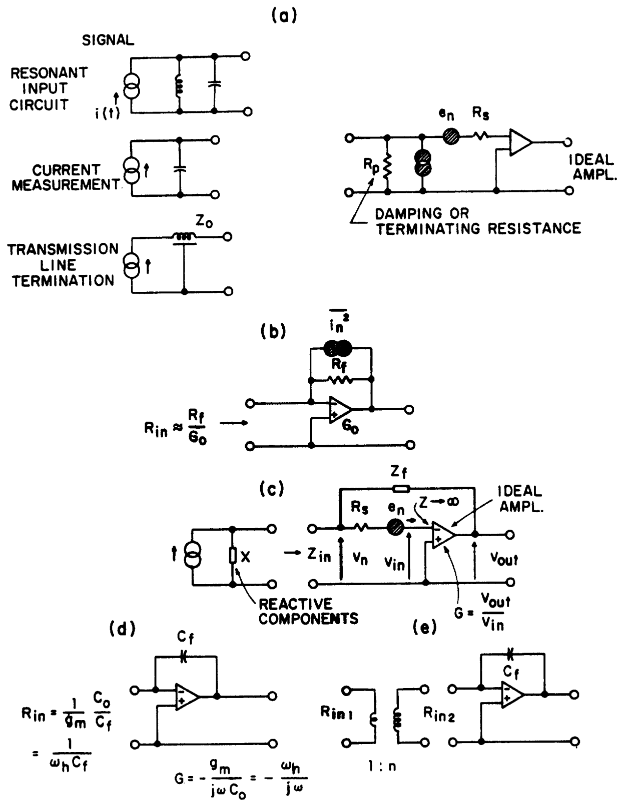


Fig. 4. Electronically "cooled" terminations and damping

(a) classical damping with resistor

(b) classical "cooled" damping with resistance in feedback.

(c) eq. circuit for analysis of "cooled" damping.

(d) "cooled" damping with capacitance in feedback.

(e) use of transformation to realize low values of "cooled" resistances.

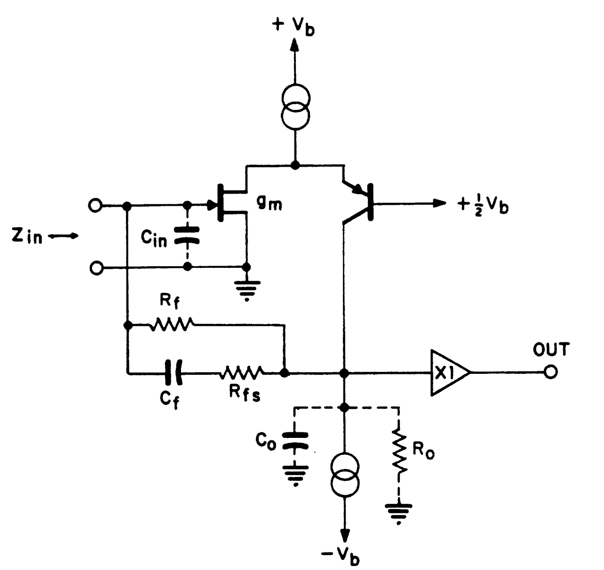


Fig. 5. Basic circuit configuration for "cooled" damping with capacitance.

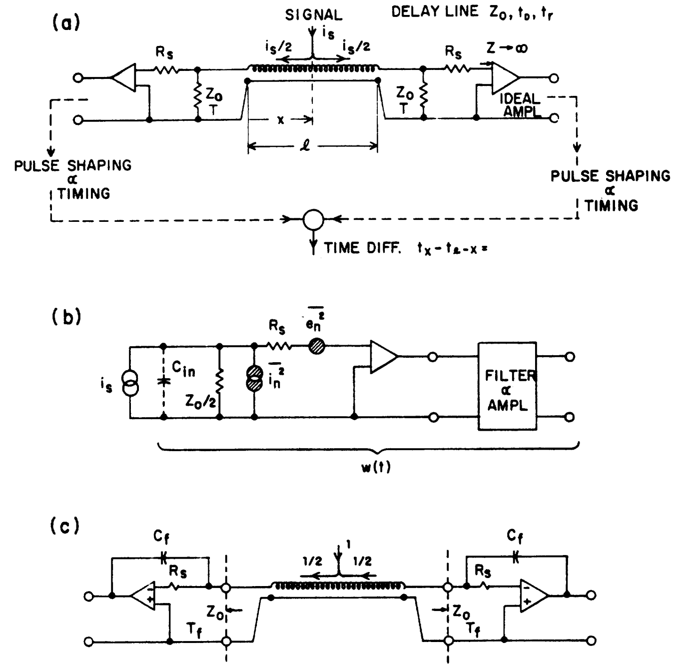


Fig. 6. Position sensing with delay lines.

(a) basic circuit with "warm" resistance terminations Z_o, T.
(b) eq. circuit of the "warm" terminations case for noise analysis.
(c) delay line position sensing with "cooled" terminations with capacitance in feedback.

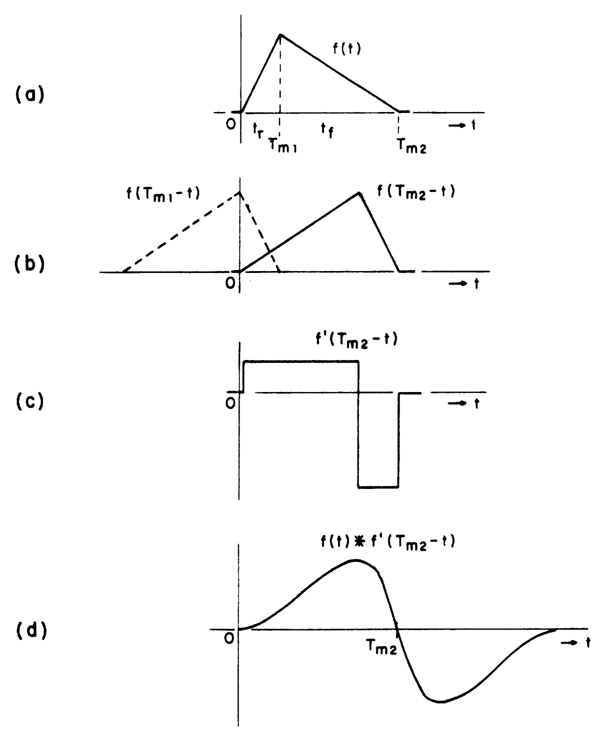


Fig. 7. Optimum filter for timing in presence of white noise (method of derivation).
(a) signal waveform

(b) optimum filter for amplitude measurements.

(c) optimum filter for timing - derivative of (b).

(d) output waveform.

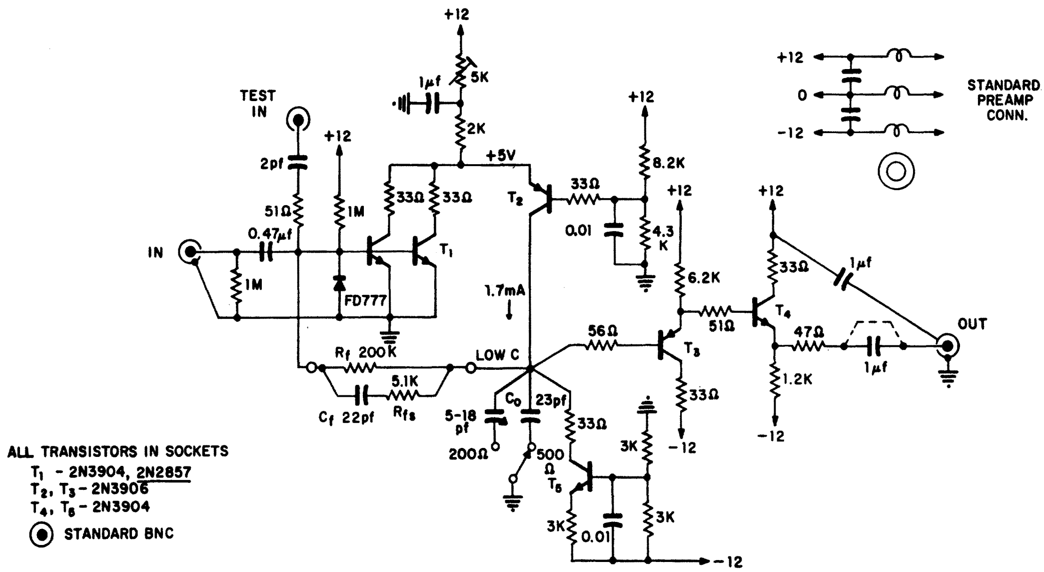


Fig. 8. An amplifier with bipolar transistor input for cooled damping or termination with R_{in} = 200 and 500 Ω. $R_{in} = \frac{1}{g_m}\,\frac{C_o}{C_f}$. Peaking at high frequencies may be used for delay lines with Z_o = 500 Ω. R_{fs} = 0 for cases where peaking is not necessary. Series eq. noise resistance $R_s \approx 30$ Ω.

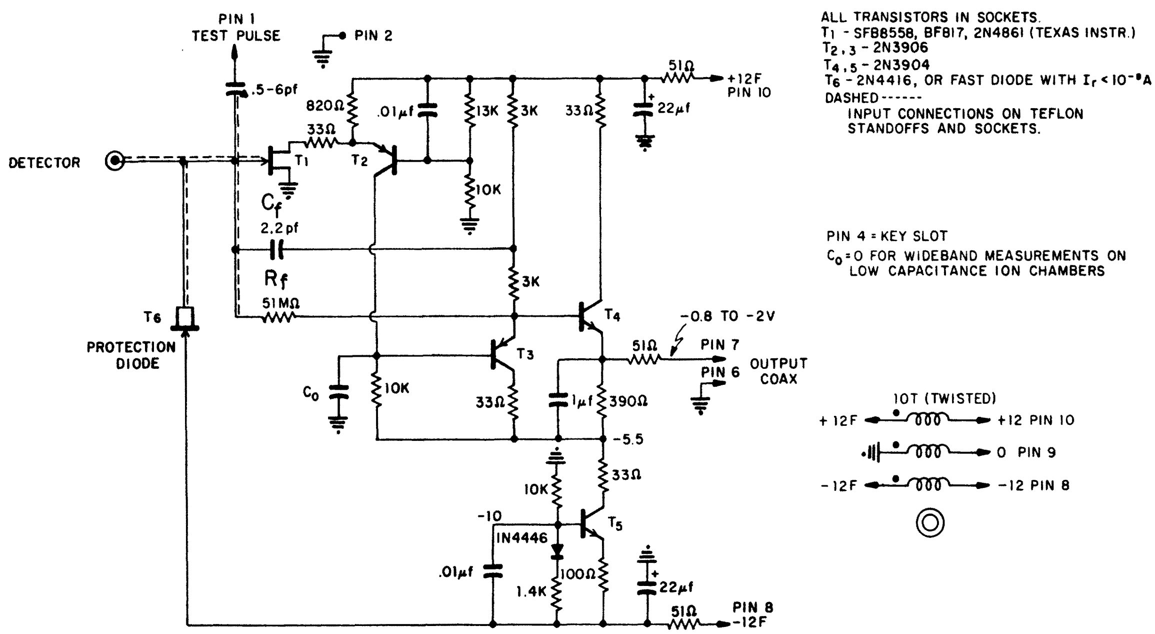


Fig. 9. An amplifier with field-effect transistor input for cooled damping in the range $R_{in} \geq 500$ Ω. $R_{in} = \frac{1}{g_m}\,\frac{C_o}{C_f/2}$. Series eq. noise resistance $R_s \approx 60$ Ω. Peaking at high frequencies may be achieved by adding a resistor in series with C_f.

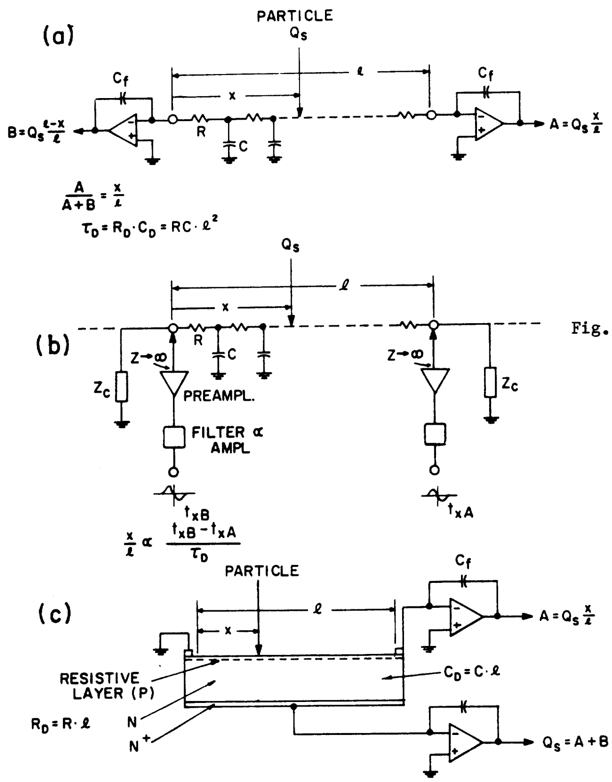


Fig. 10. Resistive electrode (R-C line) as position-sensing medium.

(a) charge division method.

(b) diffusion time method.

(c) alternative configuration for charge division illustrated for a semiconductor detector with a resistive layer.

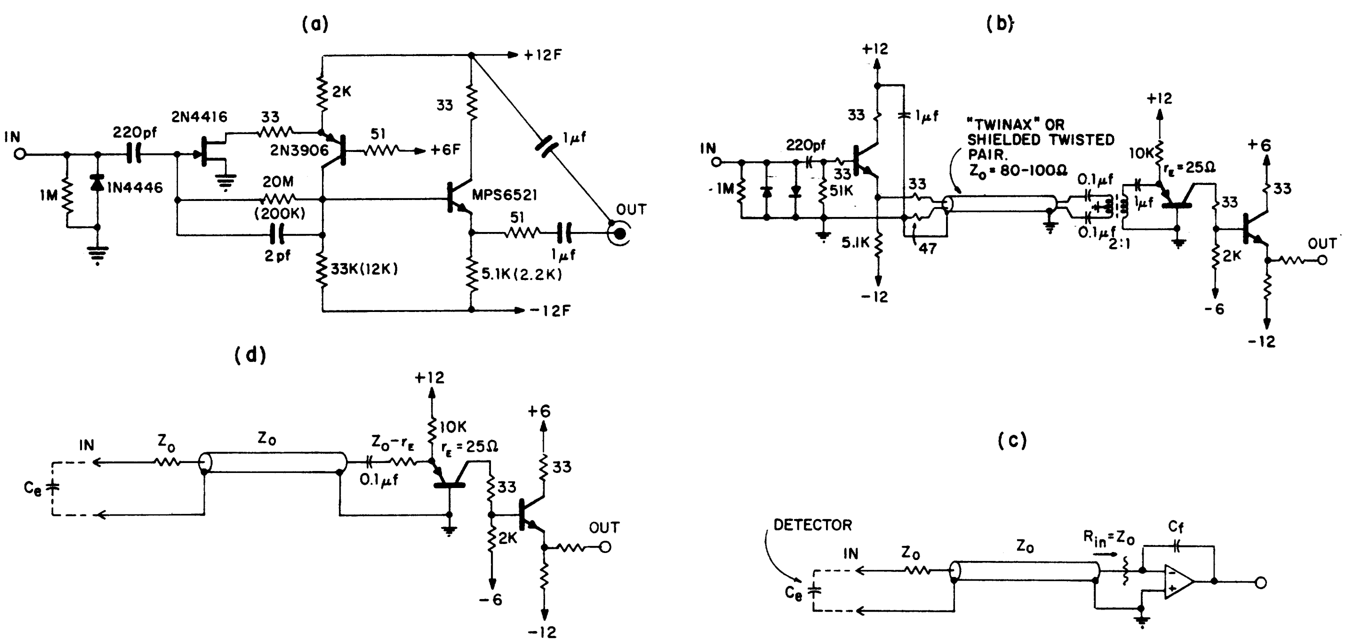


Fig. 11. Linear amplifier configurations for multielectrode detectors.

(a) a simple low-noise amplifier. Bipolar transistors can be used in place of the field-effect transistors when longer filter weighting function and minimum noise are not required (resistor values are indicated in parenthesis).

(b) the "minimum electronics at the detector" amplifier. Common mode rejection by transformer. Voltage gain from the input of the emitter follower to the output ≈ 20.

(c) remote amplifier. Termination by "cooled" resistance.

(d) remote amplifier. Termination by the emitter dynamic resistance.