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Minimum-Noise Pulse Shaping with New Double Delay-Line Filters in Nuclear Pulse Amplifiers*

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Three new double delay-line filters for pulse amplifiers were developed to circumvent the poor noise performance of conventional double delay-line filters. The operating principle of the new filters is the alteration of the noise spectrum stored in the delay lines prior to the signal arrival. With one of the new filters, a symmetrical bipolar output pulse was obtained that had a measured noise linewidth of 1.6 keV fwhm Si (189 rms electrons) for 5 pF external capacitance, 4.2 μsec delay-line length and 1.4 μsec *RC* integrating time-constant. The noise linewidth slope was 0.05 keV/pF.

INTRODUCTION

THE advantage of the bipolar output pulse of the double delay-line filter is often offset by the disadvantage of high ENV[1] (equivalent noise voltage). The delay-line bridge (Fig. 1) alters the output noise spectrum from the input section of a typical nuclear pulse preamplifier in such a manner that the average noise power per cps passed by a perfect low-pass network is increased as the cutoff frequency is increased and is decreased as the cutoff frequency is decreased. Thus, when a fast pulse is required, the ENV is especially high. Some understanding of this effect of the delay-line bridge on the noise spectrum can be acquired from the following discussion.

Consider the delay-line bridge shown in Fig. 1. Let the gains of the inverter and of the summing amplifier be set so that the pulse gain of the bridge is unity. Let the power-spectral density of the input noise be white, that is

$$N(\omega)=N(\text{mean-squared volts/rad/sec})=\text{const.} \quad (1)$$

If the delay line is disconnected, the output noise of the filter in the frequency interval $(0,\omega_h)$ will be

$$N_0(\text{mean-squared volts})=N\omega_h. \quad (2)$$

The voltage transfer function of the delay-line bridge is

$$T(j\omega)=1-\epsilon^{-j\omega\tau_d}, \quad (3)$$

where τ_d is the delay-line length in seconds. Therefore, with

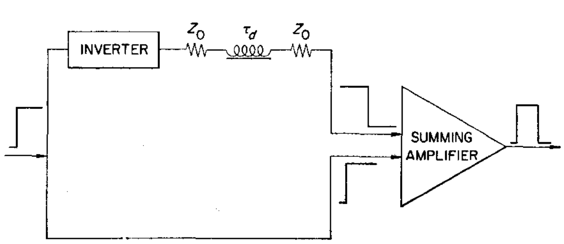


FIG. 1. Delay-line bridge.

* Research sponsored by the U. S. Atomic Energy Commission under contract with Union Carbide Corporation.

[1] The ENV is the ratio of the rms output noise voltage to the pulse gain of the filter. The magnitude of the pulse gain of the filter is the peak value of the output pulse when the input is a unit step function.

the delay line connected, the output noise in $(0,\omega_h)$ is given by

$$\begin{aligned} N_{01}&=N\int_0^{\omega_h}|1-\epsilon^{-j\omega\tau_d}|^2d\omega \\ &=(2N/\tau_d)(\omega_h\tau_d-\sin\omega_h\tau_d). \end{aligned} \quad (4)$$

Equation (4) shows that the output noise power in a low frequency interval $(\omega_h \ll 1/\tau_d)$ is reduced to values much less than $N\omega_h$, while as the interval extends into high frequencies $(\omega_h \gg 1/\tau_d)$, the output noise power approaches the value $2N\omega_h$.

When two delay-line bridges are cascaded, the pulse gain is again unity, and the output noise for disconnected delay lines is again $N_0=N\omega_h$ in the frequency interval $(0,\omega_h)$. With the delay lines connected, the output noise in $(0,\omega_h)$ is

$$\begin{aligned} N_{02}&=N\int_0^{\omega_h}|(1-\epsilon^{-j\omega\tau_d})^2|^2d\omega \\ &=(6N/\tau_d)[\tfrac{4}{3}\sin^3(\omega_h\tau_d/2)\cos(\omega_h\tau_d/2) \\ &\qquad +\omega_h\tau_d-\sin\omega_h\tau_d]. \end{aligned} \quad (5)$$

Taking the limits $\omega_h\to 0$ and $\omega_h\to\infty$ in Eq. (5) for τ_d finite and nonzero shows that at low frequencies $N_{02}<N\omega_h$ and at high frequencies $N_{02}\to 6N\omega_h$.

Thus, if the noise spectrum superposed on the input signal is white and the system bandwidth increases to accommodate higher frequencies (shorter risetime pulses), the ENV of a delay-line bridge increases by a factor that approaches the value $\sqrt{2}$, and the ENV of two cascaded bridges increases by a factor approaching the value $\sqrt{6}$. The situation is reversed at lower frequencies where both the single delay-line bridge and two cascaded bridges reduce the ENV.

The qualitative aspects of the above conclusions are also valid for other noise spectra. In particular, consider the spectrum expressed by

$$N(\omega)=K_n(1+k_n/\omega^2),$$

where K_n and k_n are constants. When the delay-line bridge operates on this spectrum, an increase in the high frequency power density and a decrease in the low frequency density is observed.

The alteration of the noise power-spectral density by the delay-line bridge is due to the delay term $\epsilon^{-j\omega\tau_d}$ in the transfer function. If the initiation of the signal pulse is at time $t=0$, the ENV of the delay-line bridge in the time interval $(0,\tau_d)$ is determined both by noise superposed on the input signal and by noise stored in the delay line prior to the signal arrival. It is shown in this paper that the proper modification of the power-spectral density of this stored noise spectrum effects a significant decrease in the ENV of a delay-line filter.

Three schemes were developed for modifying the noise spectrum stored in the delay lines. This paper presents a brief description of each scheme, some of the pertinent theoretical relationships, and several experimental results. Also, for completeness and comparison, the characteristics of conventional single and double delay-line filters are given. Detailed mathematical analysis and more extensive experimental results are published elsewhere.[2]

THEORY

A. Definitions and Symbolism

The theoretical results presented in this paper are based on the following assumptions: (1) the input signal to the filters is a step function of strength E, thus

$$e_s=EU(t) \text{ volts}; \tag{6}$$

(2) the input noise power-spectral density to the filters is given by

$$N(\omega)=K_n(1+k_n/\omega^2)\,(\text{mean squared volts/rad/sec}). \tag{7}$$

The noise spectrum $N(\omega)$ represents a general characterization of the noise output of input sections of both vacuum tube and semiconductor preamplifiers. For a charge-sensitive input section with a field-effect transistor used as the input device and fed by a semiconductor detector, the constant K_n is given by[2]

$$K_n=(2kT/\pi)G^2R_{n0}, \tag{8}$$

and the constant k_n by

$$k_n=Q_{n0}/R_{n0}, \tag{9}$$

where

$$R_{n0}=(1/C_f^2)[(C_d+C+C_f)^2R_n+C_{gs}^2(R_n/3)+RC_d^2+R_s(C_d+C_f)^2], \tag{10}$$

and

$$Q_{n0}=\frac{1}{C_f^2}\left[\frac{1}{R_g}+\frac{1}{R_f}+F(C_d+C+C_f)^2+(e/2kT)(I_2+I_2+I_{rd})\right]. \tag{11}$$

[2] T. V. Blalock, Oak Ridge National Laboratory Report No. ORNL-TM-1055 (March, 1965).

The symbols used in Eqs. (8), (10), and (11) are defined as follows: G—the voltage gain of a wideband amplifier that may be connected between the output of the input section and the input to the filter, k—Boltzmann's constant (1.38×10^{-23} J/°K), T—absolute temperature (°K), F—flicker noise constant [Ω (rad/sec)^2], where it is assumed that the flicker noise resistance of the FET channel is $R_{nf}=F/\omega^2$, e—electronic charge (1.6×10^{-19} C), I_1—gate reverse electron current (amperes), I_2—gate reverse hole current (amperes), I_{rd}—detector reverse shot-noise current (amperes).

Equations (10) and (11) can be modified easily to describe the noise spectral density of an input charge-sensitive section with a vacuum tube used as the input device. For this case R_n is the plate-current shot-noise resistance, I_1 is the negative component of grid current, and I_2 is the positive component of grid current; the term $C_{gs}^2R_n/3$ is omitted in Eq. (10), and the quantity F in Eq. (11) is usually negligible.

The following symbols will be employed in the filter discussions: s—Laplace transform variable, ω—angular frequency (rad/sec), τ_d—delay line length (sec), τ—RC time-constant (sec) of a low pass resistance-capacitance filter, λ—τ_d/τ is the pulse flatness parameter, t—time (sec), t_r—10–90% risetime (sec), g_p—pulse voltage gain of the filter (the magnitude of g_p is equal to the peak amplitude of the filter output pulse when the input is a unit step), N_0—noise power output (mean squared volts), ENV—equivalent noise voltage of the filter (volts),

$$\text{ENV}=(N_0)^{\frac{1}{2}}/g_p, \tag{12}$$

$e_0(t)$—filter output voltage (volts) as a function of time, and t_d—2% dwell time (sec) of the filter output pulse; this is the time for the output pulse to decrease to 2% of its peak value on the final return to the base line.

The subscript m attached to a quantity above denotes the value of the quantity when the minimum noise conditions are satisfied.

B. Conventional Single Delay-Line Filter

A conventional single delay-line filter (DL-RC) consists of a delay-line bridge cascaded through an isolating amplifier with a resistance-capacitance low pass network of time constant $\tau=RC$. The filter voltage transfer function is

$$T(s)=(1-\epsilon^{-\tau_d s})/(\tau s+1). \tag{13}$$

Thus, the response to a unit step voltage input is given by

$$e_0(t)=1-\epsilon^{-t/\tau}-U(t-\tau_d)(1-\epsilon^{-(t-\tau_d)/\tau}). \tag{14}$$

The time to the peak is

$$t_p=\tau_d, \tag{15}$$

and the pulse gain is

$$g_p=1-\epsilon^{-\lambda} \tag{16}$$

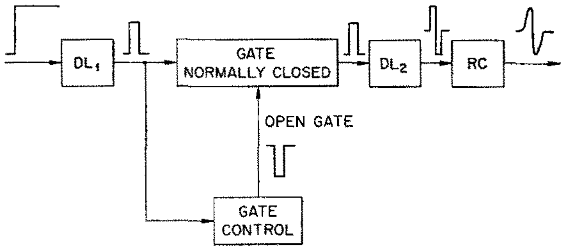


FIG. 2. Partially gated double delay-line filter.

The filter output noise is calculated from

$$N_0=\int_0^\infty N(\omega)T(j\omega)T^*(j\omega)d\omega. \tag{17}$$

The value of this integral for the single delay-line filter is

$$N_0=\pi K_n[k_n\tau_d+(1-\epsilon^{-\lambda})(1/\tau-k_n\tau)]. \tag{18}$$

With the values of Eqs. (16) and (18) for the pulse gain and output noise, respectively, the equivalent noise voltage is

$$\mathrm{ENV}=\frac{(\pi K_n)^{1/2}[k_n\tau_d+(1-\epsilon^{-\lambda})(1/\tau-k_n\tau)]^{1/2}}{1-\epsilon^{-\lambda}}. \tag{19}$$

The conditions for minimum ENV have been determined by Tsukuda,[3]

$$\lambda_m=1.036, \tag{20}$$

$$\tau_m=1.29\ (1/k_n)^{1/2}. \tag{21}$$

The minimum noise quantities corresponding to these conditions are

$$e_{pm}=0.645, \tag{22}$$

$$t_{pm}=1.34(1/k_n)^{1/2}, \tag{23}$$

$$t_{dm}=6.4(1/k_n)^{1/2}, \tag{24}$$

$$\mathrm{ENV}_m=2.75(K_n)^{1/2}(k_n)^{1/4}. \tag{25}$$

The ENV_m of the single delay-line filter is about 10% above the theoretical optimum value[4,5] and about 19% below the value[2,3] obtained with the simple RC-RC filter.

C. Conventional Double Delay-Line Filter

The conventional double delay-line filter consists of two cascaded delay-line bridges followed by a simple resistance-capacitance low pass network (RC integrator).

The voltage transfer function of the conventional $(\mathrm{DL})_2$-RC filter is

$$T(s)=(1-\epsilon^{-\tau_d s})^2/(\tau s+1) \tag{26}$$

Thus, the response to a unit step voltage input is given by

$$e_0(t)=1-\epsilon^{-t/\tau}-2U(t-\tau_d)(1-\epsilon^{-(t-\tau_d)/\tau}) \\ +U(t-2\tau_d)(1-\epsilon^{-(t-2\tau_d)/\tau}). \tag{27}$$

[3] M. Tsukuda, Nucl. Instr. Methods **14**, 241 (1962).
[4] C. H. Nowlin, J. L. Blankenship, and T. V. Blalock, Rev. Sci. Instr. **36**, 1063 (1965).
[5] K. Halbach, Helv. Phys. Acta **26**, 65 (1953).

The time to the first peak is given by Eq. (15), and the pulse gain is given by Eq. (16).

The equivalent noise voltage of the $(\mathrm{DL})_2$-RC filter is computed in the same manner as for the DL-RC filter. Its value is

$$\mathrm{ENV}=\frac{(\pi K_n)^{1/2}\{2k_n\tau_d+[(1-k_n\tau^2)/\tau](3-4\epsilon^{-\lambda}+\epsilon^{-2\lambda})\}^{1/2}}{1-\epsilon^{-\lambda}}. \tag{28}$$

The ENV of Eq. (28) approaches a minimum value of $2.69(K_n)^{1/2}(k_n)^{1/4}$ as τ approaches infinity.[4] In most situations it will probably be impractical to use the conventional $(\mathrm{DL})_2$-RC filter at this minimum ENV point because (1) a sufficiently high value of τ may be difficult to obtain, (2) the resulting triangular output pulse is unipolar, and (3) the minimum value of the ENV represents only about a 2% improvement over the DL-RC filter.

For comparison purposes, a minimum ENV for the $(\mathrm{DL})_2$-RC filter can be defined by arbitrarily fixing the flatness parameter λ and calculating the value of τ_m. Thus, according to Tsukuda,[3] if

$$\lambda_m=1, \tag{29}$$

then

$$\tau_m=2.2(1/k_n)^{1/2}. \tag{30}$$

With $\lambda_m=1$, the shape of the first half of the bipolar output pulse is very similar to that of the DL-RC filter output pulse.

The other quantites of interest at this quasi minimum-noise point are

$$e_{pm}=0.63, \tag{31}$$

$$t_{pm}=2.2(1/k_n)^{1/2}, \tag{32}$$

$$t_{dm}=12(1/k_n)^{1/2}, \tag{33}$$

$$\mathrm{ENV}_m=3.4(K_n)^{1/2}(k_n)^{1/4}. \tag{34}$$

At the quasi minimum-noise point, the ENV of the conventional $(\mathrm{DL})_2$-RC filter is about equal to that of the simple RC-RC filter.[3] However, if short time-constant (time constants less than $(1/k_n)^{1/2}$) constraints are imposed, the ENV becomes significantly higher than that of the RC-RC filter.

D. Partially Gated Double Delay-Line Filter

One scheme for modifying the noise spectrum stored in a double delay-line filter is shown schematically in Fig. 2. This partially gated $(\mathrm{DL})_2$-RC filter has a gate located between the two delay-line bridges. The gate is normally closed so that no portion of the filter input-noise spectrum is stored in the delay-line of the second bridge prior to signal arrival. The gate is opened on the leading edge and closed on the trailing edge of the pulse from the first delay-line bridge; consequently, the ENV of each half of the filter output bipolar pulse is the same as that of the conventional

DL-RC filter. Furthermore, the minimum noise conditions are unaffected by the second bridge and are therefore identical to those of the DL-RC filter. The sole function of the second bridge is to supply the second half of the bipolar pulse.

If the operation time of the gate is much less than the duration of the pulse from the first bridge, the response of the partially gated filter to a unit step input voltage is given by Eq. (27); the time to the first peak and the pulse gain are computed from Eqs. (15) and (16), respectively.

The ENV is obtained from Eq. (19). The conditions for minimum ENV are given by Eqs. (20) and (21), and the minimum noise quantities are[2]

$$e_{pm}=0.645, \tag{22}$$

$$t_{pm}=1.34(1/k_n)^{\frac{1}{2}}, \tag{23}$$

$$t_{dm}=7.16(1/k_n)^{\frac{1}{2}}, \tag{35}$$

$$\mathrm{ENV}_m=2.75(K_n)^{\frac{1}{2}}(k_n)^{\frac{1}{4}}. \tag{25}$$

Thus, at the minimum noise point, the improvements over the conventional double delay-line filter are the following: (1) the pulse gain increased by 2.4%, (2) the time to the pulse peak decreased by 39%, (3) the dwell time decreased by 40%, and (4) the ENV_m decreased by 19%.

At shorter time constants the decrease in ENV will be even greater. In the limit ($\tau \rightarrow 0$) the ENV of the partially gated $(\mathrm{DL})_2$-RC filter will be $1/\sqrt{3}$ times the ENV of the conventional filter, that is,

$$\lim_{\tau\rightarrow 0}\frac{\text{rhs Eq. (19)}}{\text{rhs Eq. (28)}}=\frac{1}{\sqrt{3}}, \tag{36}$$

where rhs means the right hand side of the equation. Thus, as τ approaches zero, the decrease in ENV by partial gating approaches 42.3%.

E. $(\mathrm{DL})_2$-RC Filter with Selective Noise Injection

The discussion in the Introduction indicated that the ENV of the conventional double delay-line filter was im-

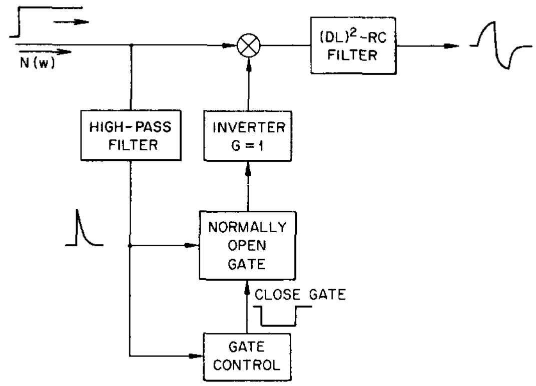


FIG. 3. Double delay-line filter with selective noise injection.

proved at low frequencies. This is due to negative correlation between the low frequency noise stored in the delay-lines and the low-frequency noise superposed on the signal pulse. Thus, the injection of the low frequency portion of the noise spectrum into the conventional $(\mathrm{DL})_2$-RC filter prior to the signal arrival should cause a considerable decrease in ENV.

The diagram of a system for accomplishing the low frequency noise injection is shown in Fig. 3. The high frequency part of the noise spectrum is selected by a high pass filter, fed through a normally open gate, inverted, and then is summed at the input of the conventional $(\mathrm{DL})_2$-RC filter. The leading edge of the signal pulse operates the gate control-trigger which causes the gate to close. This prevents degradation of the signal risetime. If the high pass filter is a simple resistance-capacitance network whose RC time constant is τ_1, and the voltage gain of the inverter is unity, then the noise spectrum input to the $(\mathrm{DL})_2$-RC filter prior to the signal arrival is given by

$$N_d(\omega)=N(\omega)\left[1-\frac{(\omega\tau_1)^2}{1+(\omega\tau_1)^2}\right]. \tag{37}$$

Thus, the noise spectrum stored in the delay lines will be composed predominantly of low-frequency components.

The ENV during the first half of the output pulse (Fig. 3) is[2]

$$(\mathrm{ENV})=\frac{(\pi K_n)^{\frac{1}{2}}}{(1-\epsilon^{-\lambda})}\left\{2k_n\lambda\tau+\frac{(1-k_n\tau^2)\{\tau_1[1+(\tau_1/2\tau)-2\epsilon^{-\lambda}+\epsilon^{-2\lambda}]-\tau(3-4\epsilon^{-\lambda}+\epsilon^{-2\lambda})\}+(1-k_n\tau_1^2)\tau_1(1.5-2\epsilon^{-\lambda\tau/\tau_1})}{(\tau_1^2-\tau^2)}\right\}^{\frac{1}{2}}. \tag{38}$$

The necessary condition for the ENV to have a minimum is that the partial derivatives of Eq. (38) must vanish at the minimum point. Taking the partial derivatives of Eq. (38) with respect to λ, τ, and τ_1 and then setting each derivative equal to zero yields a set of three equations whose simultaneous solution produces the conditions for minimum ENV. These conditions are[2]

$$\tau_{1m}=\tau_m, \tag{39}$$

$$\tau_m=1.14(1/k_n)^{\frac{1}{2}}, \tag{40}$$

$$\lambda=1.09. \tag{41}$$

The values of the quantities corresponding to the minimum ENV are

$$e_{pm}=0.664, \tag{42}$$

$$t_{pm}=1.24(1/k_n)^{\frac{1}{2}}, \tag{43}$$

$$t_{dm}=6.5(1/k_n)^{\frac{1}{2}}, \tag{44}$$

$$\mathrm{ENV}_m=3.1(K_n)^{\frac{1}{2}}(k_n)^{\frac{1}{4}}. \tag{45}$$

The ENV of Eq. (38) is a slowly varying function of pulse shape (the pulse shape is determined by λ and τ) in the vicinity of the minimum ENV point if τ_1 is properly selected for each change in λ and τ. Several sets of filter parameters that yield values of ENV within $\pm 1\%$ of the value given by Eq. (45) are listed below for $k_n=0.01$ μsec^{-2}:

λ	τ (μsec)	τ_1 (μsec)
0.8	16.0	9.0
1.0	13.0	10.5
1.2	10.0	12.0
1.4	8.5	13.5
1.6	8.0	15.0
1.8	7.0	16.5
2.0	6.0	18.0

For each value of λ above, the set (τ,τ_1) corresponds to a minimum ENV.

According to the preceding calculations, the noise injection scheme effects an appreciable improvement over the conventional $(\mathrm{DL})_2$-RC filter. At the minimum ENV point, the pulse gain has been increased by 5.4%, the ENV decreased by 8.8%, the time to pulse peak decreased by 43%, and the dwell time decreased by 46%.

The decrease in ENV is more pronounced at short time-constants, as was the case also for the partially gated filter. Thus,

$$\lim_{\tau\to 0}\frac{\text{rhs Eq. (38)}}{\text{rhs Eq. (28)}}=\frac{1}{\sqrt{6}}. \tag{46}$$

Consequently, as the pulse risetime becomes shorter $(\tau\to 0)$ the percentage decrease in ENV approaches the value 59.2%.

F. $(\mathrm{DL})_2$-RC Filter with Selective Noise Injection and Partial Gating

If a single delay-line filter could be improved by the injection scheme of Fig. 3, the partially gated filter discussed in Sec. D could be improved by low frequency noise injection into the first delay-line bridge. Thus, consider the DL-RC filter with selective noise injection. The ENV becomes[2]

$$(\mathrm{ENV})=\frac{(\pi K_n)^{\frac{1}{2}}}{(1-\epsilon^{-\lambda})}\left\{k_n\tau_d=\frac{(1-k_n\tau^2)[(\tau_1^2/2\tau)+(\tau_1-\tau)(1-\epsilon^{-\lambda})]-(\tau_1/2)(1-k_n\tau_1^2)}{\tau_1^2-\tau^2}\right\}^{\frac{1}{2}}. \tag{47}$$

It has been shown[2] that there is no point in the parameter space where the ENV of Eq. (47) has a value less than the minimum ENV of the conventional DL-RC filter. However, for certain parameter sets, the ENV of the injected filter is significantly lower than the ENV of the conventional filter. The necessary condition for decreasing the ENV of the DL-RC filter by injection is[2]

$$\tau<(1/k_n)^{\frac{1}{2}}. \tag{48}$$

Realization of the inequality (48) simultaneously with one of the following inequalities[2] is sufficient for reducing the ENV: For $\tau_1<\tau$,

$$(1-k_n\tau^2)[\tau_1/2\tau+(1-\tau_1/\tau)(1-\epsilon^{-\lambda})]>\tfrac{1}{2}(1-k_n\tau_1^2). \tag{49}$$

For $\tau_1=\tau$,

$$(1-3k_n\tau^2)>2\epsilon^{-\lambda}(1-k_n\tau^2). \tag{50}$$

For $\tau_1>\tau$,

$$(\tfrac{1}{2}+k_n\tau^2\epsilon^{-\lambda})>\epsilon^{-\lambda}+k_n\tau^2(1+\tau_1/2\tau). \tag{51}$$

As an example, let $k_n=0.01$ μsec^{-2}, $\lambda=3.0$, and $\tau=1.2$ μsec (a condition giving excellent pulse shape). Then Eq. (47) shows that the ENV for the injected filter is minimum at $\tau_1\approx 8$ μsec. This minimum value is about 19% lower than that of the conventional $(\tau_1=0)$ single delay-line filter.

The greatest decrease in ENV by injection occurs as τ approaches zero. The limit is

$$\lim_{\tau\to 0}\frac{\text{rhs Eq. (47)}}{\text{rhs Eq. (19)}}=\frac{1}{\sqrt{2}}. \tag{52}$$

Therefore, the greatest percentage decrease is 29.3%.

The essential conclusion expressed by inequalities (48), (49), (50), and (51) is that if the bipolar output pulse is constrained to have a fast risetime (τ small) and a flat top ($\lambda>1$), the noise injection scheme of Fig. 3 can be combined with the partial gating scheme of Fig. 2 to produce an ENV less than the value obtained with either scheme separately.

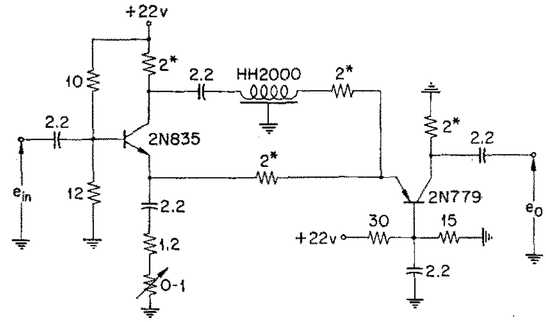


FIG. 4. Circuit of the delay-line bridge. *1% metal film resistor. Resistors in kilohms and capacitors in microfarads unless otherwise noted.

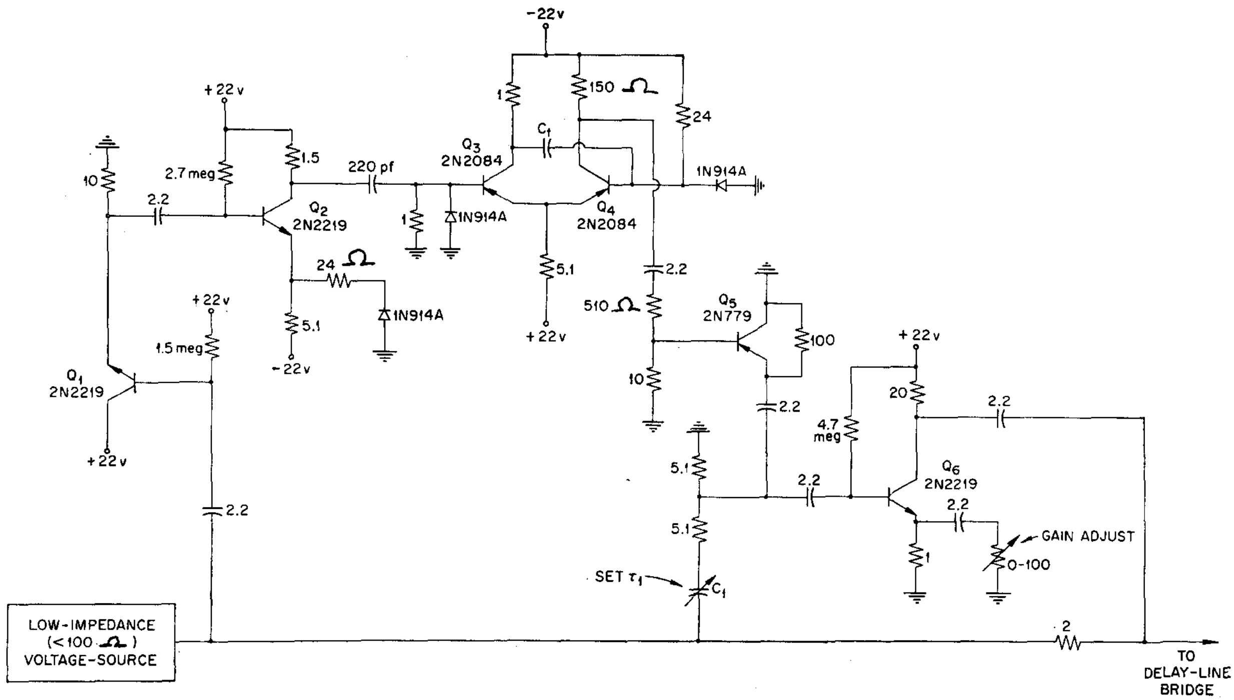
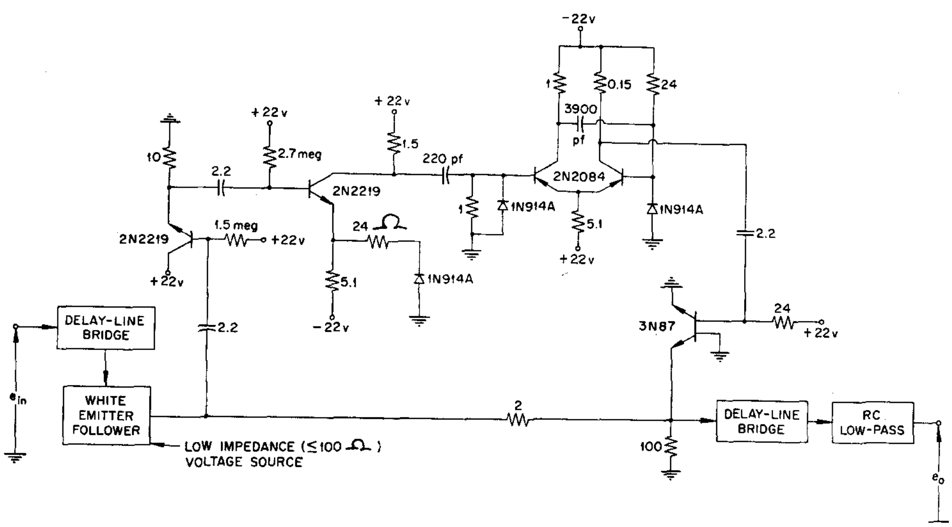


FIG. 5. Circuit of the noise injected double delay-line filter. Resistors in kilohms and capacitors in microfarads unless otherwise noted.

FIG. 6. Circuit of the partially gated double delay-line filter. Resistors in kilohms and capacitors in microfarads unless otherwise noted.



EXPERIMENTAL RESULTS

The characteristics of the new delay-line filters were obtained by measuring the noise linewidth (NLW) of an experimental preamplifier with a pulse height analyzer. The NLW is directly proportional to the ENV described above. Specifically, the NLW is the ENV divided by the charge gain from the preamplifier input to the filter input.

The input section of the preamplifier incorporates a field-effect transistor in its input stage. A temperature controller[2] was used to set the FET at its optimum temperature, which was about 125°K. The circuit details and the characteristics of the input section have been published elsewhere.[2,6]

The pulse shaping in the preamplifier was accomplished by both the conventional and the new double delay-line filters. The delay-line bridge circuit (based on a design by Hahn and Guiragossian[7]), the noise injecting circuit, and the partially gated $(DL)_2$-RC filter which were used in obtaining the experimental data are shown in Figs. 4, 5, and 6, respectively. These circuits were developed to test the concepts and are not offered as final designs.

The measured values of NLW are based on a silicon detector whose mean energy absorption per electron-hole pair formed is 3.6 eV. The equivalent noise charge in rms

[6] T. V. Blalock, IEEE Trans. Nucl. Sci. **NS-11**, No. 3, 365 (1964).

[7] J. Hahn and V. Guiragossian, IEEE Trans. Nucl. Sci. **NS-10**, No. 3, 44 (1963).

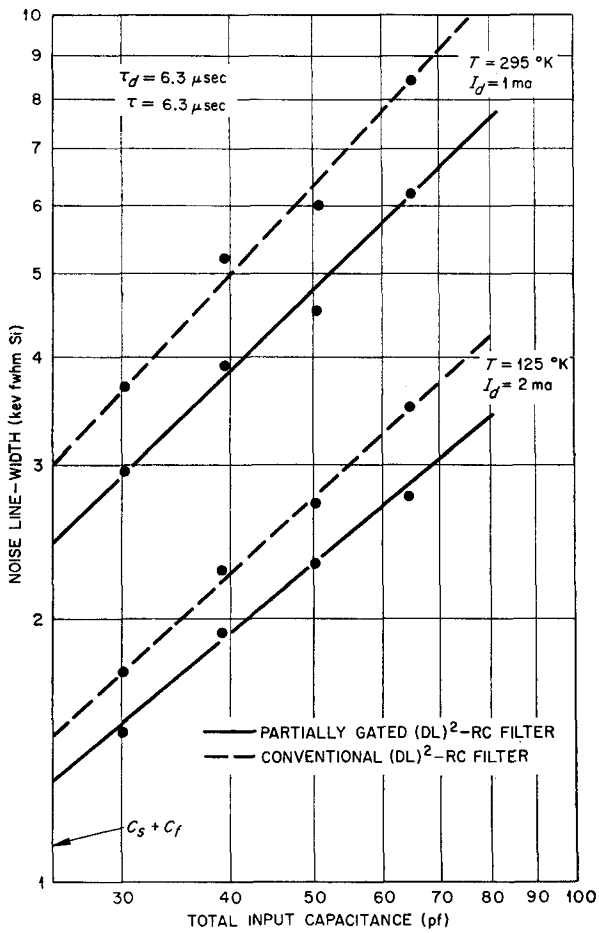


FIG. 7. Noise linewidths obtained with the partially gated and the conventional $(DL)_2$-RC filters for time constants near the minimum noise point.

electrons (ENC_e) can be obtained from the equation

$$ENC_e(\text{rms electrons}) = 118\ \text{NLW (keV fwhm Si)}. \quad (53)$$

Figure 7 shows the measured NLW of the conventional and the partially gated $(DL)_2$-RC filters for time constant values near the minimum noise point. Partial gating caused an improvement of approximately 20% in the NLW. The measured ratio of the NLW for the conventional filter to NLW of the gated filter was about 1.2. The calculated value from Eqs. (34) and (25) is $3.4/2.75 = 1.25$. For an external capacitance C_d of 5 pF and at an FET temperature of 125°K, the NLW of the gated filter was 1.5 keV fwhm Si. Under the same conditions, the NLW (Ref. 5) of the optimum RC-RC filter was 1.7 keV.

The filter time-constants for the data of Fig. 7 yield an output pulse which has unsatisfactory shape; the positive peak amplitude is not equal to the negative peak amplitude, and the risetime and dwell time are impractically long. Figure 8 shows the NLW of a pulse that has a more desirable shape. The measured rise time was 2.3 μsec, the time to the pulse peak was 4.2 μsec, and the dwell time was 15 μsec. The positive and negative peaks were equal in amplitude. The NLW of the conventional filter was much greater for this improved shape; however, the NLW of the partially gated filter was only slightly degraded. In other words, when the filter time-constants were shifted from the optimum values ($\tau_d = 6.3\ \mu$sec, $\tau = 6.3\ \mu$sec) to the values ($\tau_d = 4.2\ \mu$sec, $\tau = 1.4\ \mu$sec) for improved pulse shape, the NLW (FET temperature = 125°K, $C_d = 5$ pF) of the conventional filter increased from 1.7–2.5 keV while the NLW of the gated filter only increased from 1.5–1.6 keV. This was a 47% increase for the conventional filter but only a 7% increase for the gated filter. Thus the improvement in NLW by partial gating was 36%.

A greater improvement in NLW was measured at shorter time-constants. The output pulse of the filter used to obtain the data of Fig. 9 had an 0.8 μsec risetime, 2.2 μsec time to pulse peak, 6.1 μsec dwell time, and equal positive and negative peak amplitudes. The NLW of the gated filter was 2.2 keV at T = 125°K and with 5 pF external capacitance. The NLW of the conventional filter was 3.8 keV. Thus the improvement was about 42%.

The NLW of the noise injected filter is shown in Fig. 10. The high-pass filter RC time constant τ_1 was set to 13 μsec

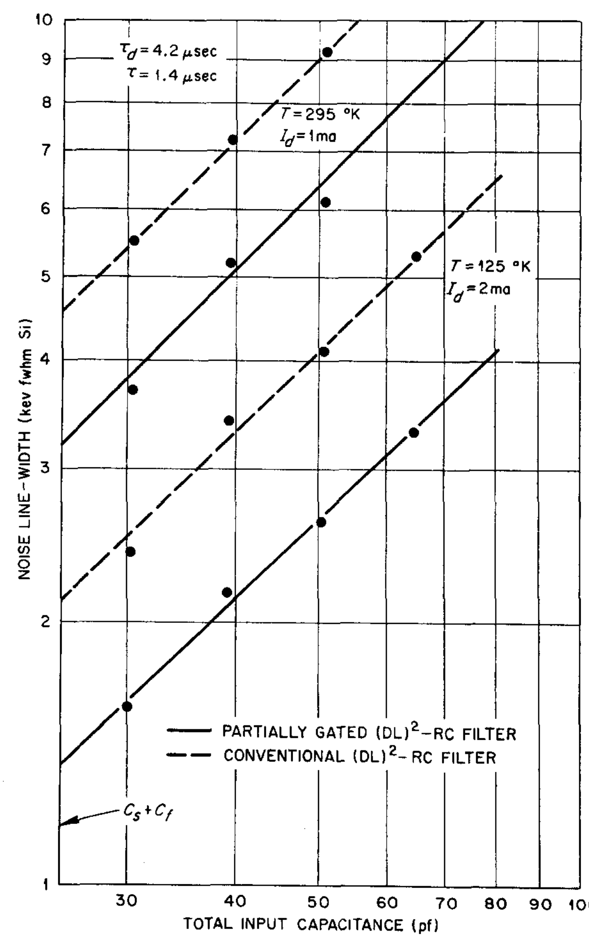


FIG. 8. Noise linewidths obtained with the partially gated and the conventional $(DL)_2$-RC filters for time constants that were selected for low noise and satisfactory pulse shape.

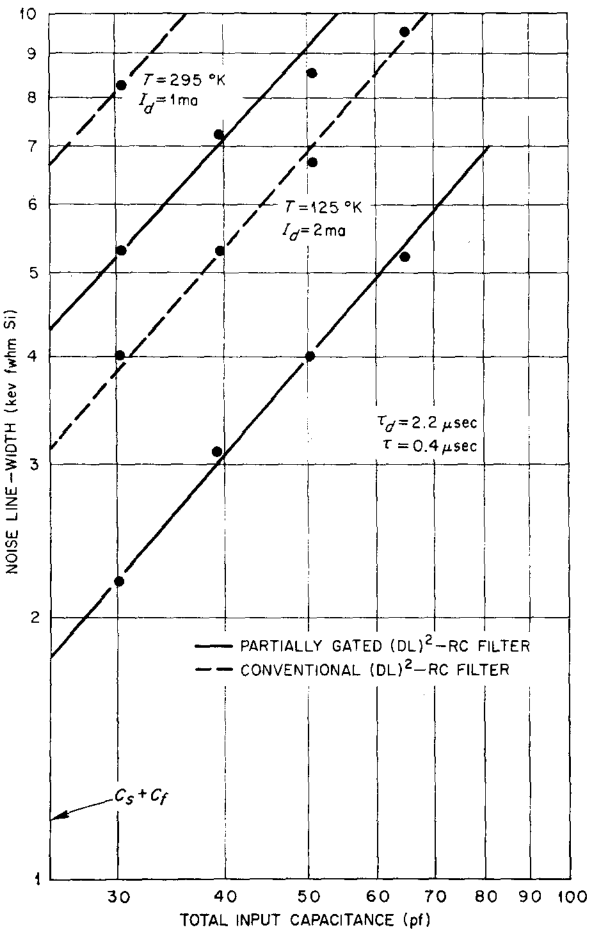


FIG. 9. Noise linewidths obtained with the partially gated and the conventional $(DL)_2$-RC filters for time constants selected to produce fast flat-topped pulses.

to give minimum ENV for $\lambda=3$ and $\tau=1.4\ \mu$sec. This value of τ_1 was selected from the numerical values of ENV calculated from Eq. (38) with a digital computer. For $\tau_d=4.2\ \mu$sec and $\tau=1.4\ \mu$sec ($\lambda=3$), this filter gave an NLW at T=295°K that was considerably lower than the value obtained with the partially gated filter (2.9 vs 3.8 keV at $C_d=5$ pF). The values at 125°K were slightly higher than those of the gated filter.

In principle, the noise injection scheme can be combined with the partial gating scheme to produce an improvement in NLW over either scheme separately if the filter time constants have certain constrained values [Eqs. (48), (49), (50), and (51)]. This concept was experimentally checked at T=125°K, $\tau_d=4.2\ \mu$sec, $\tau=0.4\ \mu$sec, and $\tau_1=7\ \mu$sec. The measured values of NLW ($C_d=5$ pF) were

$$\mathrm{NLW}\Big|_{\text{nongated}}^{\text{injected}}=2.4 \text{ keV fwhm Si},$$

$$\mathrm{NLW}\Big|_{\text{gated}}^{\text{noninjected}}=2.7 \text{ keV},$$

$$\mathrm{NLW}\Big|_{\text{gated}}^{\text{injected}}=2.1 \text{ keV}.$$

DISCUSSION

It has been shown analytically and experimentally that gating techniques can be employed to generate bipolar pulses whose signal-to-noise ratios are as high as those of the unipolar pulse obtained with a single delay-line filter and, consequently, superior to those obtained by resistance-capacitance shaping. Undoubtedly, some engineering problems will be encountered in applying the new gating concepts in practical amplifiers. The partial-gating scheme is probably the easiest to apply. In this case the active time of the gating circuitry is only for the duration of the rectangular output pulse from the first bridge. The active time of the gating circuitry in the injection scheme must be greater than the duration of the high-pass filter output pulse (Fig. 3) if a spurious signal pulse is to be avoided. The effects of gate nonlinearities, overload and high count rates, have not yet been studied in detail. Also, for very fast output pulses, the propagation time of the gating circuitry must be considered.

The use of gating techniques to improve signal-to-noise ratios in nuclear pulse amplifier is certainly not exhausted. Possibly, the bipolar pulse could be obtained by sampling methods, and thus eliminate the need for delay lines in

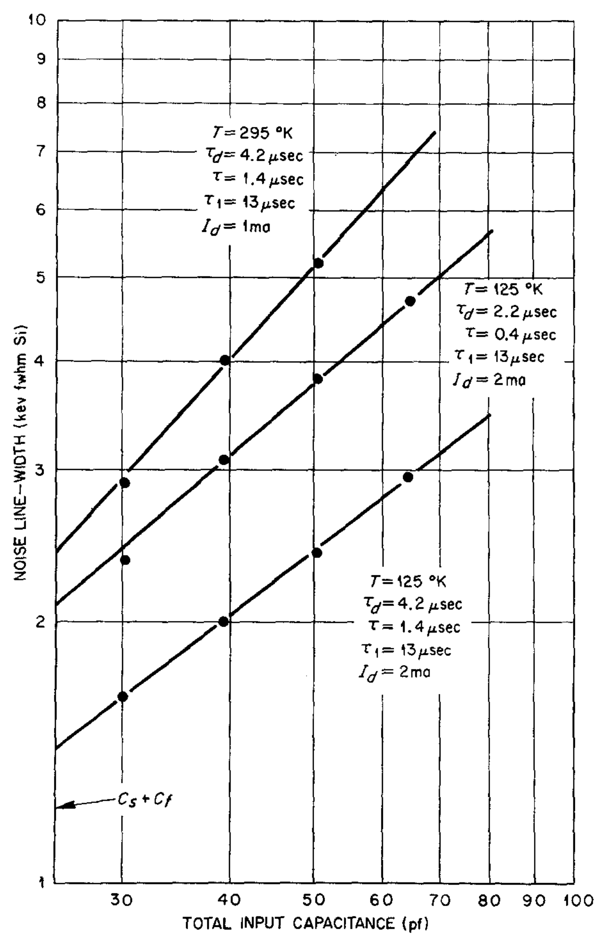


FIG. 10. Noise linewidths obtained with the noise injected $(DL)_2$-RC filter.

pulse-shaping networks. Another possibility is the application of gating techniques to lumped-element approximations of delay lines.

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Fixed Point Calibrations of Pressure Gauges*

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Manganin gauges were calibrated at five fixed points, and the resulting calibration compared with that obtained at a single fixed point, or at three fixed points. For most work three points are sufficient. The nonlinearity is 0.135%/kbar. Manganin gauges intercompared agree within one bar to 10 kbar. Pressures of the H_2O L-III-V, L-V-VI triple points and melting pressure of VI at 25°C were determined. It was not found possible to prepare a gold-chromium coil which was free of hysteresis.

INTRODUCTION

Both Bridgman[1] and Adams *et al.*[2] in their studies of the manganin resistance pressure gauge concluded that a properly prepared gauge was linear to 10 kbar, and thus could be calibrated by measuring its resistance at a single fixed point whose pressure was well known. However, Michels and Lenssen[3] encountered a nonlinearity of 0.25%/kbar in their studies with similar gauges; and subsequent extension of the pressure range by Bridgman caused him to re-examine and revise his opinion of the linearity of the gauges, and to start correcting for the nonlinearity by a two point calibration.[4] Recent work in this and other labs has uncovered further evidence of nonlinearities in the Bridgman 10 kbar scale,[5] and has been briefly reported in the literature.[6]

This study was undertaken to evaluate the validity of proposed procedures for calibration of manganin gauges in the 10 kbar region, as well as to shed some more light upon the effects of the individual winding technique, seasoning, hysteresis, etc., of both manganin and gold-chromium pressure gauges.

Table I. Coil descriptions.

Gauge type No.	Method of winding	Material used
1	loosely random wound, tied with thread	manganin, Au-Cr
2	tightly layer wound on fiber bobbin	manganin
3	2.38 mm helix supported on fired lava frame	manganin
4	wound loosely in grooves in 9.53 mm fiber rod	manganin, Au-Cr
5	3.175 mm helix supported in glass tube	Au-Cr

COIL PREPARATION

There were perhaps as many as 25 coils used in this investigation, and as a description of each would be superfluous, the types are listed in Table I. Coils of type 1 have been in regular use in this lab for a number of years, and represent the most commonly used winding technique. Type 2 coils were prepared for the primary purpose of determining the effects of a particular type of winding constraint on the coils; the very compact winding requiring substantially smaller volume within the pressure cell than other coils of equal resistance. Type 3 coil is similar to that described by Bowman and Johnson,[7] and represents a strain-free mounting. Type 4 requires the largest volume, but has minimal forming strains and is a strain-free mounting. Type 5 is similar to type 3, but the helix is larger, thus

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