CHAPTER IV

DISCUSSION AND CONCLUSION

The paper describes the development of amplifier and single channel analyzer with emphasis on the fast response of the amplifier and timing operation of the SCA using transistors and integrated circuits. The serious problems arising during the development is the inherent noise in the high slew rate operational amplifiers used in the amplifying section. Ultimately LM 318 operational amplifier is selected because of its optimum slew rate to noise ratio. Another important factor that should be considered in selecting the operational amplifier is its settling time. In many cases poor settling time can outweigh the advantages of having high slew rate and wide bandwidth. In such a case optimum performance can be achieved through external compensation.

In the SCA section, the output monostable multivibrator shapes the output signal to a pulse of 4 V. amplitude and approximately 0.5 µs. wide. The amplitude of this logic pulse is smaller than, than the one specified in page 31. However, its amplitude lies well above the logic "1" level of the digital IC's and can be used to drive other digital circuits without difficulty.

CHAPTER V

APPENDIX

CIRCUIT DESIGN

5.1 THE INVERTING AMPLIFIER [1,2,5]

The basic operational amplifier circuit is shown in Fig. 5.1. This circuit gives a closed-loop gain of R_2/R_1 when this ratio is small compared with the amplifier open-loop gain and output is out of phase by 180^o.

The input impedance is equal to R_1

R_3 should be chosen to be equal to the parallel combination of R_1 and R_2 to minimize the offset voltage error due to the bias current.

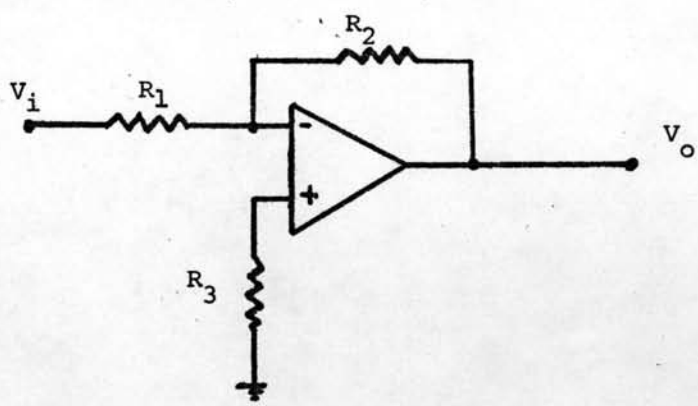


Fig. 5.1 Inverting Amplifier

$$V_o = -\frac{R_2}{R_1} V_i \quad , \qquad R_3 = R_1 // R_2$$

Refer to Fig. 2.2 $\quad \text{Gain} = \frac{R_8}{R_7} = 1$

Choose $R_8 = 1\ k\Omega \quad 1\ \%$

$\therefore \quad R_7 = 1\ k\Omega \quad 1\ \%$

$\therefore \quad R_9 = R_7 // R_8 = 500\ \Omega$

Choose $R_9 = 499\ \Omega \quad 1\ \%$

Refer to Figure 2.3 $\quad \text{Gain} = \dfrac{R_{17}}{R_{16}} = 4$

Choose $R_{17} = 4000\ \Omega \simeq 4020\ \Omega$

$\therefore \quad R_{16} = 1000\ \Omega$

$R_{18} = R_{17} // R_{16} \simeq 806\ \Omega$

Refer to Figure 2.3 & 2.4

$$\text{Gain} = \frac{V_o}{V_i} = \frac{R_f}{R_i}(\alpha)$$

α is given by fractional output voltage feed back. It varies from V_o to $\frac{V_o}{3}$ (Fine gain ranges from 1 to 3) See Fig. 2.4.

$R_f = R_{30} = 8060\ \Omega$

$R_i = R_{28} + R_{29} = 3000\ \Omega$ (For coarse gain the rotary switch position at 4, 8, 16 & 32)

and $R_i = R_{29} = 1500\ \Omega$ (For the position 64)

The gain of the first 4 positions.

if $\alpha = \frac{1}{3} \quad \therefore \quad \text{Gain} = \frac{8000}{3000} \times 3 = 8$

if $\alpha = 1 \quad \therefore \quad \text{Gain} = \frac{8000}{3000} = \frac{8}{3}$

The gain of the position 64

if $\alpha = \frac{1}{3} \quad \therefore \quad \text{Gain} = \frac{8000}{1500} \times 3 = 16$

if $\alpha = 1 \quad \therefore \quad \text{Gain} = \frac{8000}{1500} = \frac{16}{3}$

The gain of position 64 : position 32 = 16 : 8

= 2 : 1

Coarse gain setting : See Figure 2.4

$$\text{Current of attenuator} = \frac{V_o}{1500} = 4\ I \quad \text{amp.}$$

$$\therefore \quad I = \frac{V_o}{1500 \times 4}$$

$\therefore$ Voltage at :

$$V_{(4)} = 500\ I = 500 \times \frac{V_o}{1500 \times 4} = \frac{V_o}{12} \quad V$$

$$V_{(8)} = 1000\ I = 1000 \times \frac{V_o}{1500 \times 4} = \frac{V_o}{6} \quad V$$

$$V_{(16)} = 1000 \times 2I = 1000 \times \frac{2V_o}{1500 \times 4} = \frac{V_o}{3} \quad V$$

$$V_{(32)} = 1000 \times 4I = 1000 \times \frac{4V_o}{1500 \times 4} \quad \frac{2V_o}{3} \quad V$$

$$V_{(64)} = 1000 \times 4I \times 2$$

$$= 1000 \times \frac{8V_o}{1500 \times 4} = \frac{4V_o}{3} \quad V$$

(Gain of position 64 is twice times of position 32)

$\therefore$ Coarse gain range $= V_{(64)} : V_{(4)} = 16 : 1$

5.2 THE NON-INVERTING AMPLIFIER [1,2,5]

Fig. 5.2 shows a high input impedance non-inverting circuit. This circuit gives a closed-loop gain equal to the ratio of the sum of R_1 and R_2 to R_1.

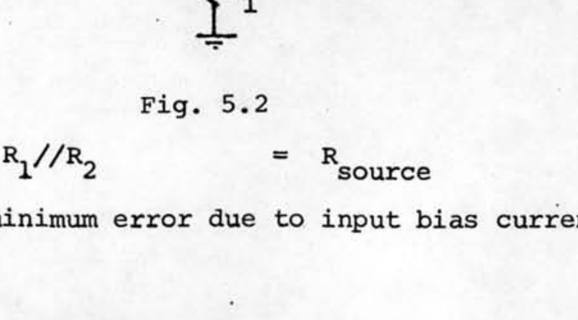


Fig. 5.2

$$\frac{V_o}{V_i} = \frac{R_1 + R_2}{R_1}$$

$$R_1 // R_2 = R_{source}$$

For minimum error due to input bias current

The primary differences between this connection and the inverting circuit are that the output is not inverted and that the input impedance is very high. If R_1 is very high, this circuit becomes the unity gain buffer circuit as shown in Fig. 2.1.

Refer to Fig. 2.6 Gain $= \frac{R_{55} + R_{56}}{R_{55}} \simeq 4$

Let $R_{56} = 10\ k\Omega$

$\therefore R_{55} = 3\ k\Omega$

IC-9 & IC-11 of Fig. 2.8 are the unity gain buffer circuits. The output voltage follows its input voltage.

5.3 THE DIFFERENCE AMPLIFIER [1,2,5]

The difference amplifier is the complement of the summing amplifier and allows the substraction of two voltages as, shown in Fig. 5.3.

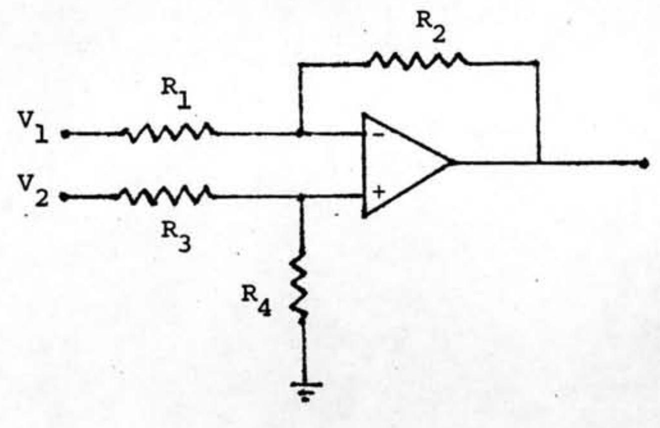


Fig. 5.3

$$V_o = \frac{(R_1+R_2)}{R_3+R_4} \frac{R_4}{R_1} V_2 - \frac{R_2}{R_1} V_1$$

For $R_1 = R_3$ and $R_2 = R_4$

$$\therefore \quad V_o = \frac{R_2}{R_1}(V_2 - V_1)$$

if $R_1 = R_2$

$$\therefore \quad V_o = V_2 - V_1$$

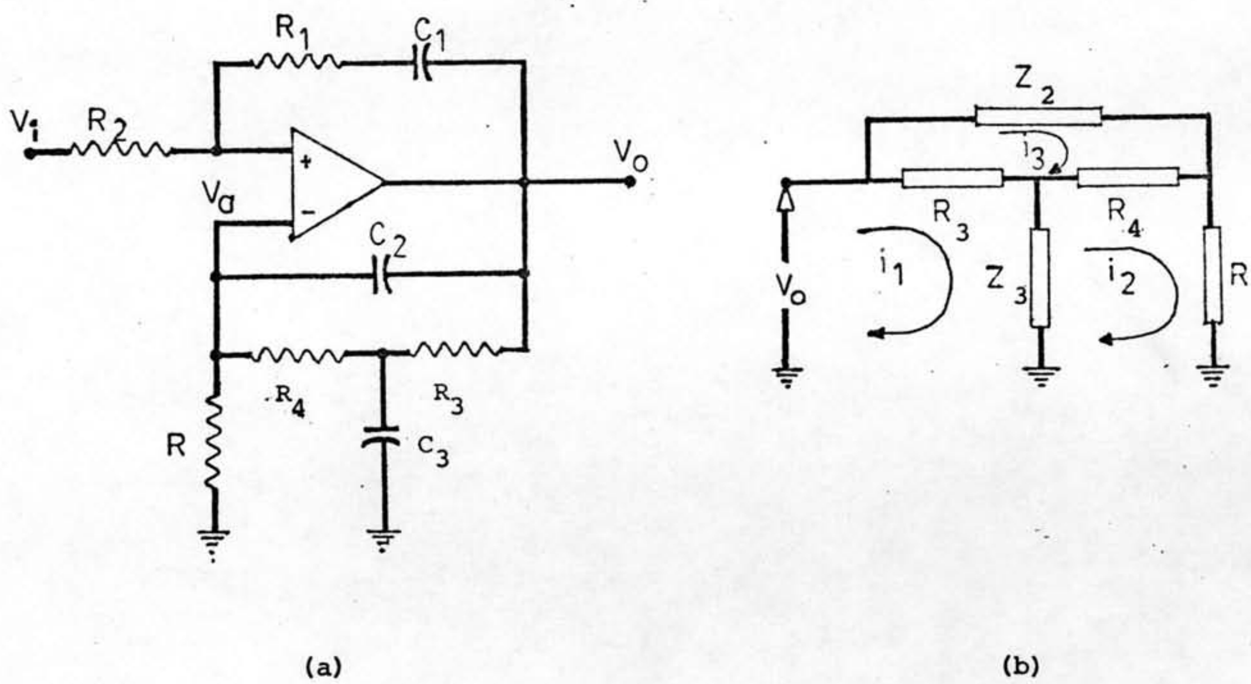
IC-10 of Fig. 2.8 is the difference amplifier

$$R_{92} = R_{93} = R_{96} = R_{97} = 1\ k$$

$$\therefore \quad V_o = -E - \Delta E = -(E + \Delta E)$$

(Sum of lower level and window control setting)

5.4 ACTIVE FILTER [3, 4, 6]



(a) (b)

Fig. 5.4 Active Filter

Refer to Fig. 5.4 (b)

$$V_o = (R_3 + Z_3)I_1 - Z_3I_2 - R_3I_3 \qquad (1)$$

$$0 = -Z_3I_1 + (R + Z_3 + R_4)I_2 - R_4I_3 \qquad (2)$$

$$0 = -R_3I_1 - R_4I_2 + (R_3 + R_4 + Z_2)I_3 \qquad (3)$$

$$\begin{bmatrix} V_o \\ 0 \\ 0 \end{bmatrix} = \begin{bmatrix} R_3 + Z_3 & - Z_3 & - R_3 \\ -Z_3 & R + Z_3 + R_4 & - R_4 \\ -R_3 & -R_4 & R_3 + R_4 + Z_2 \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \\ I_3 \end{bmatrix}$$

$$\Delta = \begin{vmatrix} R_3 + Z_3 & - Z_3 & - R_3 \\ -Z_3 & R + Z_3 + R_4 & - R_4 \\ -R_3 & -R_4 & R_3 + R_4 + Z_2 \end{vmatrix}$$

$$= (R_3+Z_3)\ (R+R_4+Z_3)\ (R_3+R_4+Z_2) - R_3R_4Z_3 - R_3R_4Z_3$$
$$-R_3^{\ 2}\ (R+R_4+Z_3) - R_4^{\ 2}(R_3+Z_3) - Z_3^{\ 2}(R_3+R_4+Z_2)$$

$$= RR_3R_4 + RR_3Z_3 + RR_4Z_3 + RR_3Z_2 + R_3R_4Z_2 + R_3Z_2Z_3$$
$$+ RZ_2Z_3 + R_4Z_2Z_3$$

$$I_2 = \frac{\begin{bmatrix} R_3 + Z_3 & V_o & - R_3 \\ -Z_3 & 0 & - R_4 \\ -R_3 & 0 & R_3+R_4+Z_2 \end{bmatrix}}{\Delta}$$

$$I_2 = \frac{R_3R_4Vo + Z_3(R_3+R_4+Z_2)Vo}{\triangle}$$

$$Va = I_2R = \frac{R_3R_4 + Z_3(R_3+R_4+Z_2) \quad RVo}{\triangle}$$

$$\frac{Vo}{Va} = \frac{RR_3R_4+RR_3Z_3+RR_4Z_3+RR_3Z_2+R_3R_4Z_2+R_3Z_2Z_3+RZ_2Z_3+R_4Z_2Z_3}{RR_3R_4+RZ_3\ (R_3+R_4+Z_2)}$$

Let $R_3 = aR$

$R_4 = 2aR$

$$\therefore \frac{Vo}{Va} = Ao = \frac{2a^2R^3+aR^2Z_3+2aR^2Z_3+aR^2Z_2+2a^2R^2Z_2+aRZ_2Z_3+RZ_2Z_3+2aRZ_2Z_3}{2a^2R^3+3aR^2Z_3+RZ_2Z_3}$$

$$= \frac{2a^2R^3+3aR^2Z_3+(a+2a^2)R^2Z_2+(3a+1)RZ_2Z_3}{2a^2R^3+3aR^2Z_3+RZ_2Z_3}$$

$$= 1 + \frac{(a+2a^2)RZ_2 + 3aZ_2Z_3}{2a^2R^2+3aRZ_3+Z_2Z_3} \tag{4}$$

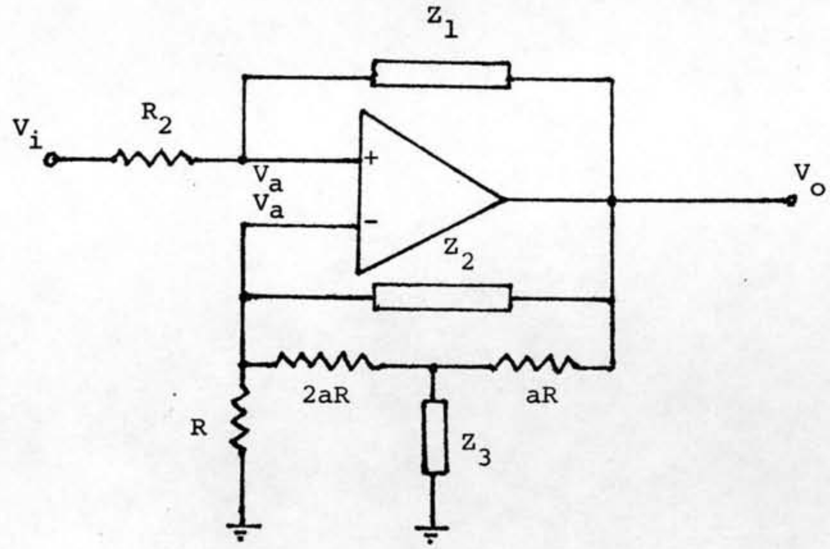


Fig. 5.5

Refer to Fig. 5.5

$$\frac{Vi-Va}{R_2} + \frac{Vo-Va}{Z_1} = 0 \qquad (5)$$

$$\frac{Vi}{R_2} - \frac{Vo}{AoR_2} + \frac{Vo}{Z_1} - \frac{Vo}{AoZ_1} = 0$$

$$\frac{Vi}{R_2} = \frac{Vo}{AoR_2} - \frac{Vo}{Z_1} + \frac{Vo}{AOZ_1}$$

$$\frac{Vi}{R_2} = \frac{Vo(Z_1-AoR_2+R_2)}{AoR_2Z_1}$$

$$\therefore \quad \frac{Vo}{Vi} = \frac{AoZ_1}{Z_1+R_2(1-Ao)}$$

$$= \frac{\left[1 + \frac{(a+2a^2)RZ_2+3aZ_2Z_3}{2a^2R^2+3aRZ_3+Z_2Z_3}\right] Z_1}{Z_1-R_2\left[\frac{(a+2a^2)RZ_2+3aZ_2Z_3}{2a^2R^2+3aRZ_3+Z_2Z_3}\right]}$$

$$\frac{Vo}{Vi} = \frac{\left[2a^2R^2+3aRZ_3+Z_2Z_3+(a+2a^2)RZ_2+3aZ_2Z_3\right] Z_1}{Z_1(2a^2R^2+3aRZ_2+Z_2Z_3)-(a+2a^2)RR_2Z_2-3aR_2Z_2Z_3}$$

$$= \frac{2a^2R^2Z_1+3aRZ_1Z_3+Z_1Z_2Z_3+(a+2a^2)RZ_1Z_2+3aZ_1Z_2Z_3}{2a^2R^2Z_1+3aRZ_1Z_3+Z_1Z_2Z_3-(a+2a^2)RR_2Z_2-3aR_2Z_2Z_3}$$

$$Z_1 = R_1+\frac{1}{SC_1}, \qquad Z_2 = \frac{1}{SC_2}, \qquad Z_3 = \frac{1}{SC_3}$$

$$\frac{Vo}{Vi} = \frac{\frac{2a^2R^2(SR_1C_1+1)}{SC_1} + \frac{3aR(SR_1C_1+1)}{S^2C_1C_3} + \frac{(SR_1C_1+1)}{S^3C_1C_2C_3} + \frac{(a+2a^2)(SR_1C_1+1)R}{S^2C_1C_2} + \frac{3a(SR_1C_1+1)}{S^3C_1C_2C_3}}{\frac{2a^2R^2(SR_1C_1+1)}{SC_1} + \frac{3aR(SR_1C_1+1)}{S^2C_1C_3} + \frac{(SR_1C_1+1)}{S^3C_1C_2C_3} - \frac{(a+2a^2)RR_2}{SC_2} - \frac{3aR_2}{S^2C_2C_3}}$$

$$= \frac{2a^2R^2S^2C_2C_3(SR_1C_1+1)+3aRSC_2(SR_1C_1+1)+SR_1C_1+(a+2a^2)RSC_3(SR_1C_1+1)+3a(SR_1C_1+1)}{2a^2R^2S^2C_2C_3(SR_1C_1+1)+3aRSC_2(SR_1C_1+1)+SR_1C_1+1-(a+2a^2)RR_2S^2C_1C_3-3aR_2SC_1}$$

$$\frac{Vo}{Vi} = \frac{S^3 2a^2R^2R_1C_1C_2C_3 + S^2\left[2a^2R^2C_2C_3 + 3aRR_1C_1C_2 + (a+2a^2)RR_1C_1C_3\right] + S\left[3aRC_2 + R_1C_1 + (a+2a^2)RC_3 + 3aR_1C_1\right] + (3a+1)}{S^3 2a^2R^2R_1C_1C_2C_3 + S^2\left[2a^2R^2C_2C_3 + 3aRR_1C_1C_2 - (a+2a^2)RR_2C_2C_3\right] + S(R_1C_1 + 3aRC_2 - 3aR_2C_1) + 1}$$

$$= \frac{S^3 + S^2\left[\frac{2a^2R^2C_2C_3 + 3aRR_1C_1C_2 + (a+2a^2)RR_1C_1C_3}{2a^2R^2R_1C_1C_2C_3}\right] + S\left[\frac{3aRC_2 + R_1C_1 + (a+2a^2)RC_3 + 3aR_1C_1}{2a^2R^2R_1C_1C_2C_3}\right] + \frac{(3a+1)}{2a^2R^2R_1C_1C_2C_3}}{S^3 + S^2\left[\frac{2a^2R^2C_2C_3 + 3aRR_1C_1C_2 - (a+2a^2)RR_2C_2C_3}{2a^2R^2R_1C_1C_2C_3}\right] + S\left[\frac{R_1C_1 + 3aRC_2 - 3aR_2C_1}{2a^2R^2R_1C_1C_2C_3}\right] + \frac{1}{2a^2R^2R_1C_1C_2C_3}}$$

Let $C_1 = C_3 = C, \qquad C_2 = 2C$

$$\frac{Vo}{Vi} = \frac{S^3 + S^2\left[\frac{4a^2R^2C^2 + 6aRR_1C^2 + (a+2a^2)RR_1C^2}{4a^2R^2R_1C^3}\right] + S\left[\frac{6aRC + R_1C + (a+2a^2)RC + 3aR_1C}{4a^2R^3R_1C^3}\right] + \frac{3a+1}{4a^2R^3R_1C}}{S^3 + S^2\left[\frac{4a^2R^2C^2 + 6aRR_1C^2 - 2(a+2a^2)RR_2C^2}{4a^2R^2R_1C^2}\right] + S\left[\frac{R_1C + 6aRC - 3aR_2C}{4a^2R^3R_1C^3}\right] + \frac{1}{4a^2R^2R_1C^3}}$$

For the simplest 3rd. power form expression

Let $R_1 = 2aR, \qquad R_2 = bR$

$$\frac{Vo}{Vi} = \frac{S^3 + \frac{S^2(18a^2+4a^3)R^2C^2}{8a^3R^3C^3} + \frac{S(9a+8a^2)RC}{8a^3R^3C^3} + \frac{3a+1}{8a^3R^3C^3}}{S^3 + S^2\left[\frac{16a^2-2b(a+2a^2)}{8a^3R^3C^3}\right]R^2C^2 + \frac{S(8a-3ab)RC}{8a^3R^3C^3} + \frac{1}{8a^3R^3C^3}} \tag{6}$$

Let us consider to the denominator of equation (6)

$$S^3 + S^2\left[\frac{16a^2-2b(a+2a^2)}{8a^3RC}\right] + \frac{S(8a-3ab)}{8a^3R^2C^2} + \frac{1}{8a^3R^3C^3} \tag{7}$$

The expression form of $\left[S + \frac{1}{2aRC}\right]^3 =$

$$S^3 + \frac{3S^2}{2aRC} + \frac{3S}{(2aRC)^2} + \frac{1}{8a^3R^3C^3} \tag{8}$$

By comparing the coefficient of equation (7) and (8)

$$b \simeq \frac{3}{4}$$

$$\therefore \frac{Vo}{Vi} = \frac{S^3 + S^2\frac{(18a^2+4a^3)R^2C^2}{8a^3R^3C^3} + \frac{S(9a+8a^2)RC}{8a^3R^3C^3} + \frac{(3a+1)}{8a^3R^3C^3}}{(S + \frac{1}{2aRC})^3} \quad (9)$$

Gain at d.c. = 10

$\therefore \quad \frac{Vo}{Vi}$ (at S = 0) $\quad 3a + 1 = 10$

$\therefore \quad a = 3$

$\therefore$ The complete equation is

$$\frac{Vo}{Vi} = \frac{S^3 + \frac{270S^2R^2C^2}{(6RC)^3} + \frac{99SRC}{(6RC)^3} + \frac{10}{(6RC)^3}}{(S + \frac{1}{6RC})^3} \quad (10)$$

The complete circuit is shown in Figure 5.6

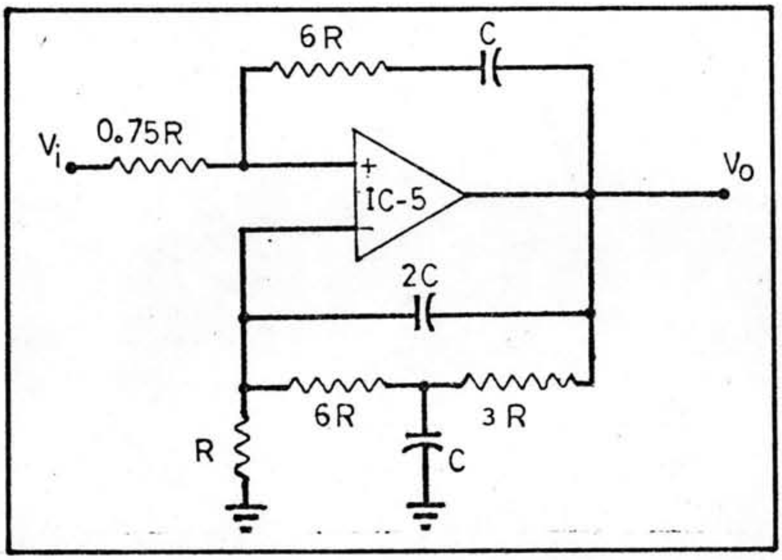


Fig. 5.6 Active Filter

Time constant		=	0.5	μs
6RC		=	0.5	μs
If R = 1 kΩ	∴ C	=	82	pF
Time constant		=	2	μs
6RC		=	2	μs
If R = 1 kΩ	∴ C	=	330	pF

Fig. 5.6 is an integrating amplifier composed of an I.C. operation amplifier IC-5 and integrating network R_{47}, R_{48}, R_{49}, R_{50}, R_{51}, C_{19}, C_{21} and C_{23}.

The output of this amplifier goes to a final differention network R_{54} and C_{25} which is also the input to the final amplifier.

From experiment near Gaussian pulse shaping, for 0.5 microsecond time constant of near Gaussian pulse.

R_{47}	=	750	ohms
R_{48}	=	5.62	k ohms
R_{49}	=	1	k ohm
R_{50}	=	6.04	k ohms
R_{51}	=	3.01	k ohms
C_{19}	=	27	pF
C_{21}	=	68	pF
C_{23}	=	39	pF
R_{54}	=	715	ohms
C_{25}	=	500	pF

For 2 microseconds time constant of near Gaussian pulse, all resistors are the same value, but capacitor networks are changed by the internal connection.

C_{20}	=	150	pF
C_{22}	=	270	pF
C_{24}	=	110	pF
C_{26}	=	2000	pF

5.5 POLE-ZERO CANCELLATION [3, 7]

Pole-zero cancellation is a method for eliminating pulse after the first differentiating network. The technique employed is described by the waveforms and equations shown in Figs. 5.7 and 5.8. In an amplifier without pole-zero cancellation the exponential tail on the preamplifier output signal (usually 50 to 500 μs.) causes an undershoot whose peak amplitude is roughly.

$$\frac{\text{Undershoot amplitude}}{\text{Differentiated pulse amplitude}} = \frac{\text{Differentiation time constant}}{\text{Preamplifier pulse decay time}}$$

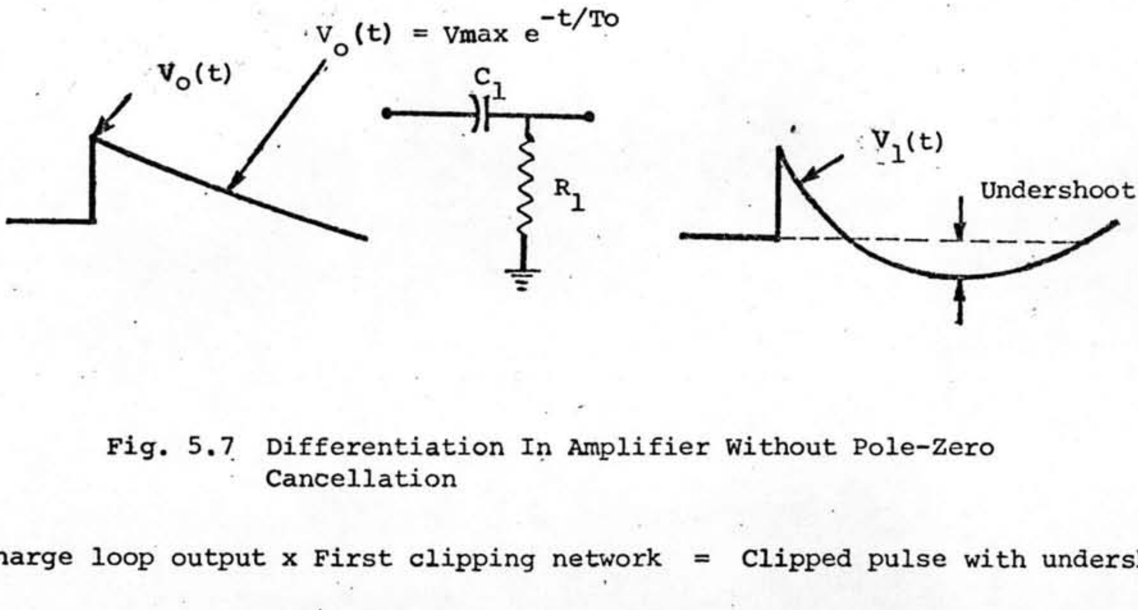


Fig. 5.7 Differentiation In Amplifier Without Pole-Zero Cancellation

Charge loop output x First clipping network = Clipped pulse with undershoot

Equation

$$Vmax\ e^{-t/T_o} \times G(t) = V_1(t)$$

Laplace Transform :

$$Vmax \times \frac{1}{S + \frac{1}{T_o}} \times \frac{S}{S + \frac{1}{R_1 C_1}} = V_1(S)$$

Let $T_1 = R_1C_1$

Therefore

$$\frac{V_{max}}{T_o - T_1}\left(T_o\, e^{-t/T_1} - T_1 e^{-t/T_o}\right) = V_1(t)$$

$0 \leq K \leq 1$

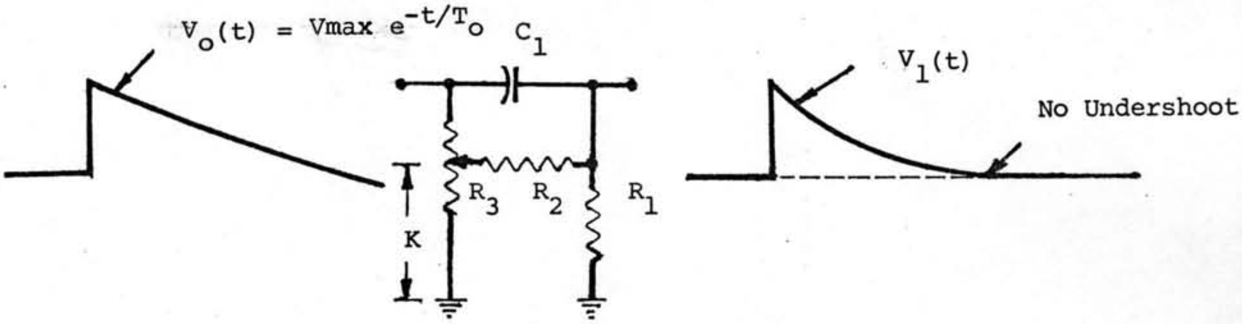


Fig. 5.8 Differentiation (clipping) in Pole-Zero Cancelled Amplifier

Charge loop output x Pole-zero cancelled clipping network = Clipped pulse with undershoot

Equation

$$V_{max}\, e^{-t/To} \times G(t) = V_1(t)$$

Laplace Transform;

$$V_{max} \frac{1}{S + \frac{1}{To}} \times \frac{S + \frac{K}{R_2C_1}}{S + \frac{R_1 + R_2}{R_1R_2C_1}} = V_1(S)$$

Pole zero cancel by letting

$$S + \frac{1}{To} = S + \frac{K}{R_2C_1}$$

$$\text{Or} \quad \frac{Vmax}{S + \frac{R_1 + R_2}{R_1 R_2 C_1}} = \frac{Vmax}{S + \frac{1}{R_p C_1}} = V_1(S)$$

$$\text{Where} \quad R_p = \frac{R_1 R_2}{R_1 + R_2}$$

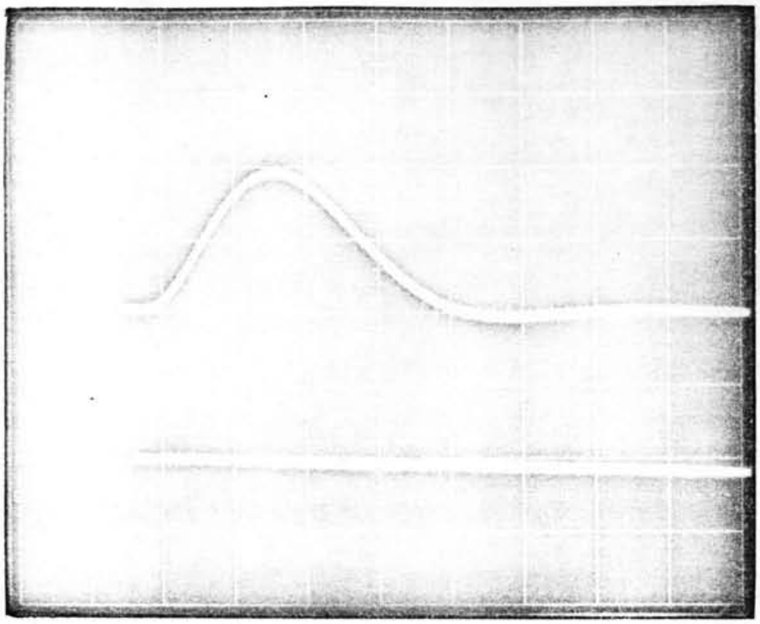
$$\therefore \quad Vmax\, e^{\frac{-t}{R_p C_1}} = V_1(t)$$

For a 1 µs. differentiation time and a 50 µs. preamplifier pulse decay time the maximum undershoot is 2% and decays with a 50 µs. time constant. Under overload conditions this undershoot is often sufficiently large to saturate the amplifier during a considerable portion of the undershoot, causing exessive dead time. This effect can be reduced by increasing the preamplifier pulse decay time (which generally reduces the counting rate capabilities of the preamplifier) or by compensating for the undershoot by using pole-zero cancellation.

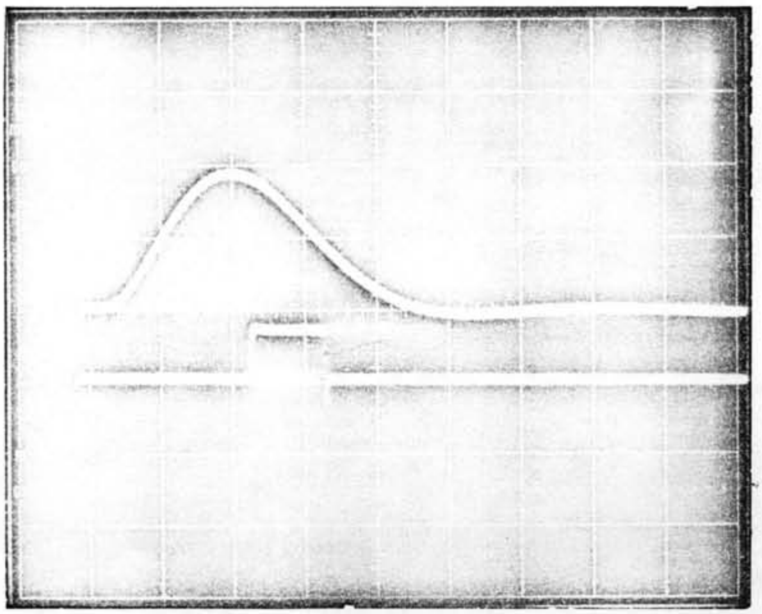
Pole-zero cancellation is accomplished by the network shown in Fig. 5.8. The pole $S + \frac{1}{T_o}$ due to the preamplifier pulse decay is cancelled by the zero $S + \frac{K}{R_2 C_1}$ of the network. In effect, the dc path across the differentiation capacitor adds an attenuated replica of the preamplifier pulse to just cancel the negative undershoot of the clipping network.

TESTING RESULTS

GAUSSIAN PULSE OUTPUT 0.5 μs. TIME CONSTANT

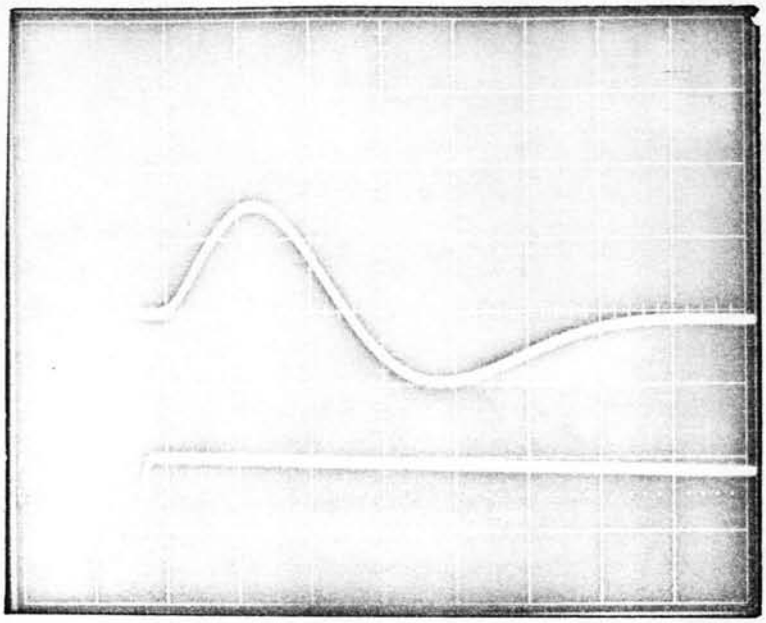


UPPER BEAM : 5V./DIV. AMPLIFIER OUTPUT (UNIPOLAR)
LOWER BEAM : 20mV./DIV. INPUT PULSE
TIME : 0.5 μs./DIV.

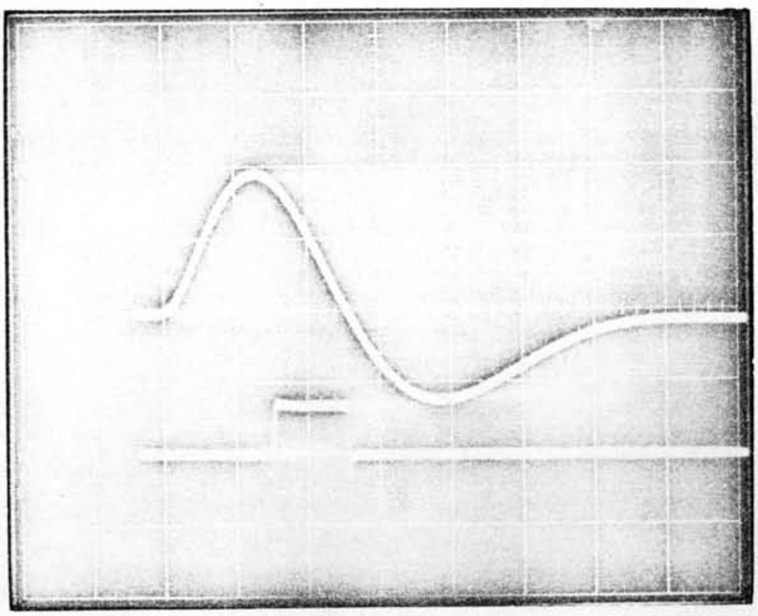


UPPER BEAM : 5V./DIV. AMPLIFIER OUTPUT (UNIPOLAR)
LOWER BEAM : 5V./DIV. SCA. OUTPUT
TIME : 0.5 μs./DIV.

GAUSSIAN PULSE OUTPUT 0.5 μs. TIME CONSTANT

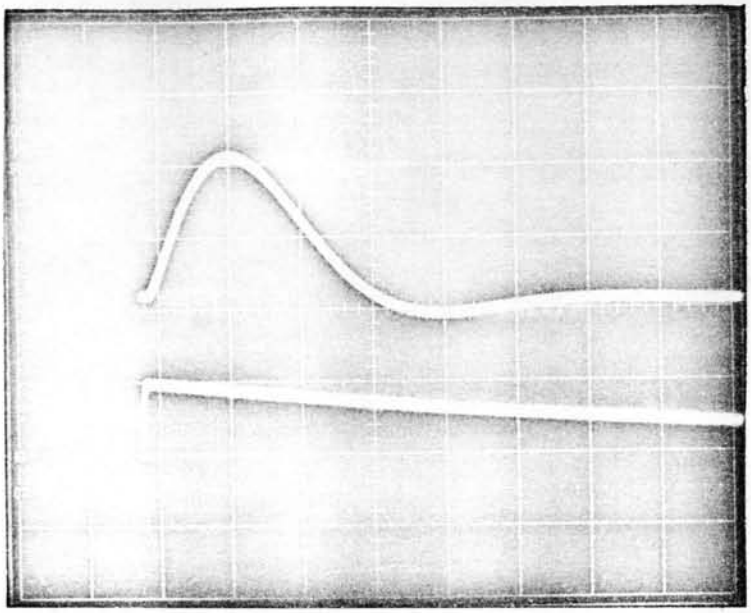


UPPER BEAM : 5V./DIV. AMPLIFIER OUTPUT (BIPOLAR)
LOWER BEAM : 20mV./DIV. INPUT PULSE
TIME : 0.5 μs./DIV.

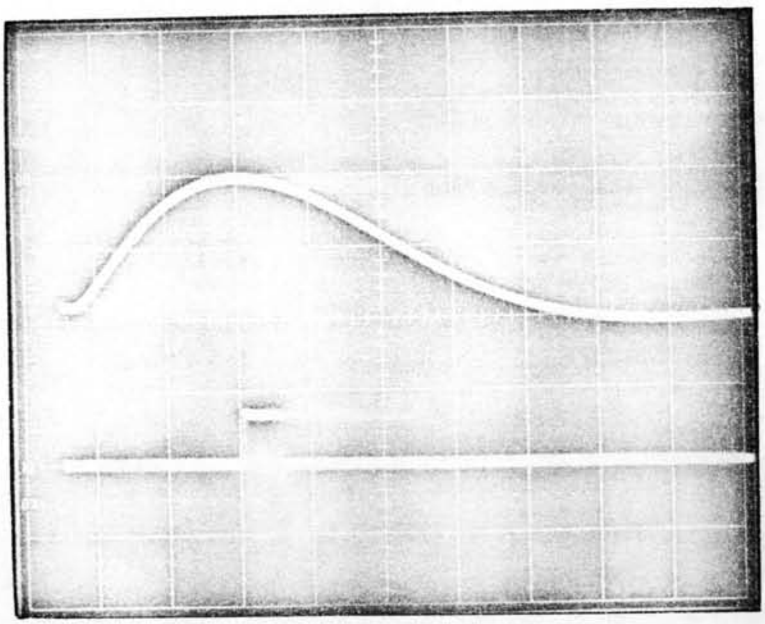


UPPER BEAM : 5V./DIV. AMPLIFIER OUTPUT (BIPOLAR)
LOWER BEAM : 5V./DIV. SCA. OUTPUT
TIME : 0.5 μs./DIV.

GAUSSIAN PULSE 2 μs. TIME CONSTANT

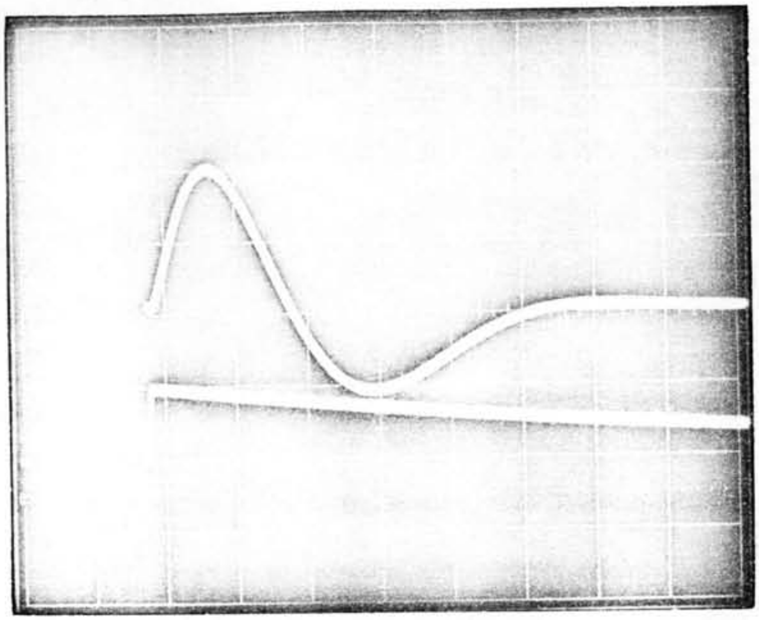


UPPER BEAM : 5V./DIV. AMPLIFIER OUTPUT (UNIPOLAR)
LOWER BEAM : 20mV./DIV. INPUT PULSE
TIME : 2 μs./DIV.

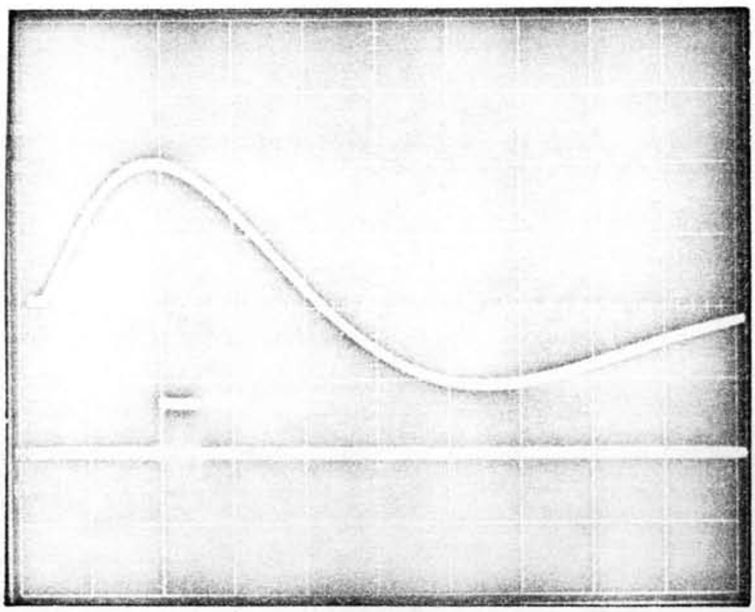


UPPER BEAM : 5V./DIV. AMPLIFIER OUTPUT (UNIPOLAR)
LOWER BEAM : 5V./DIV. SCA. OUTPUT
TIME : 1 μs./DIV.

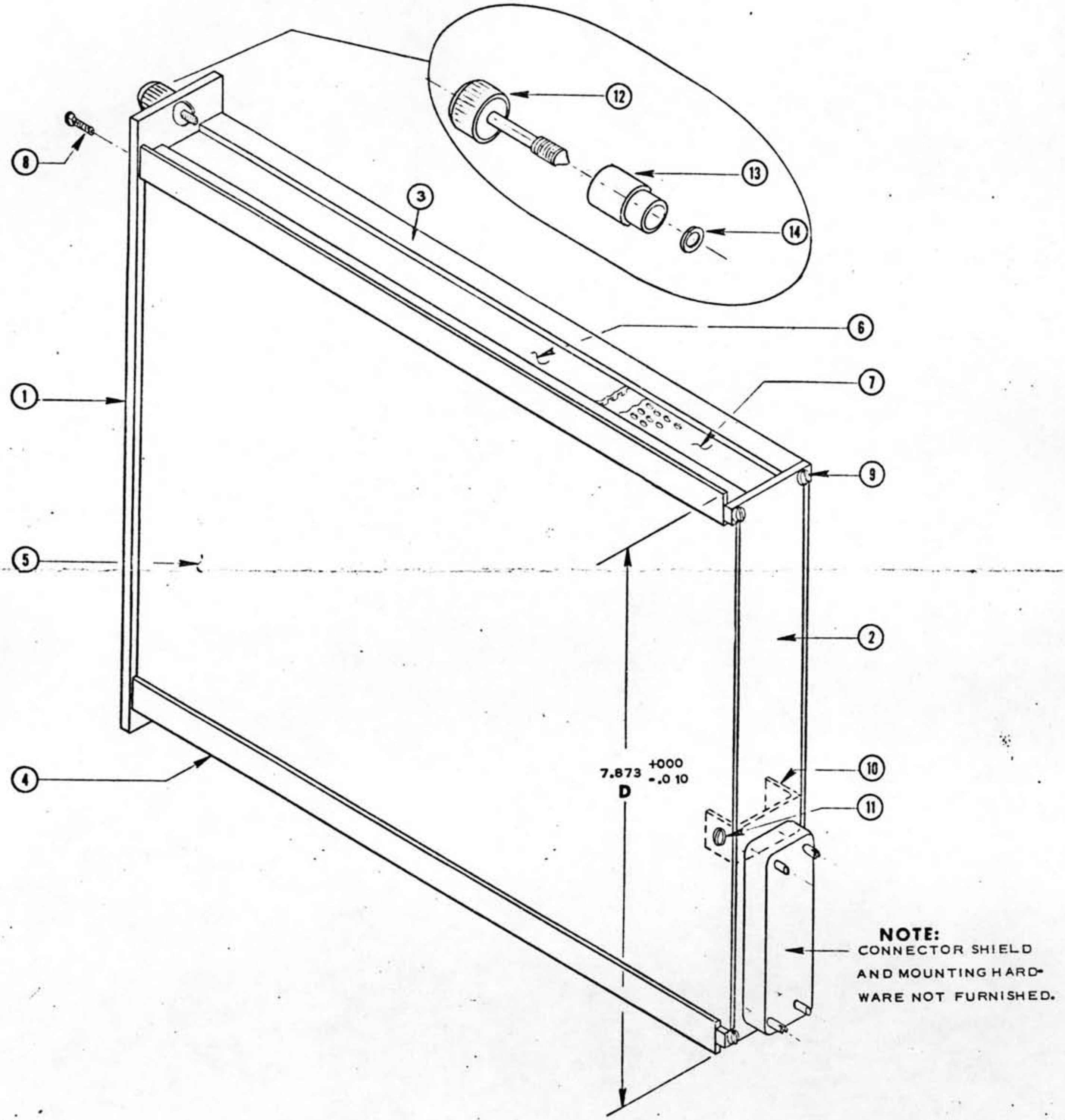
GAUSSIAN PULSE 2 μs. TIME CONSTANT



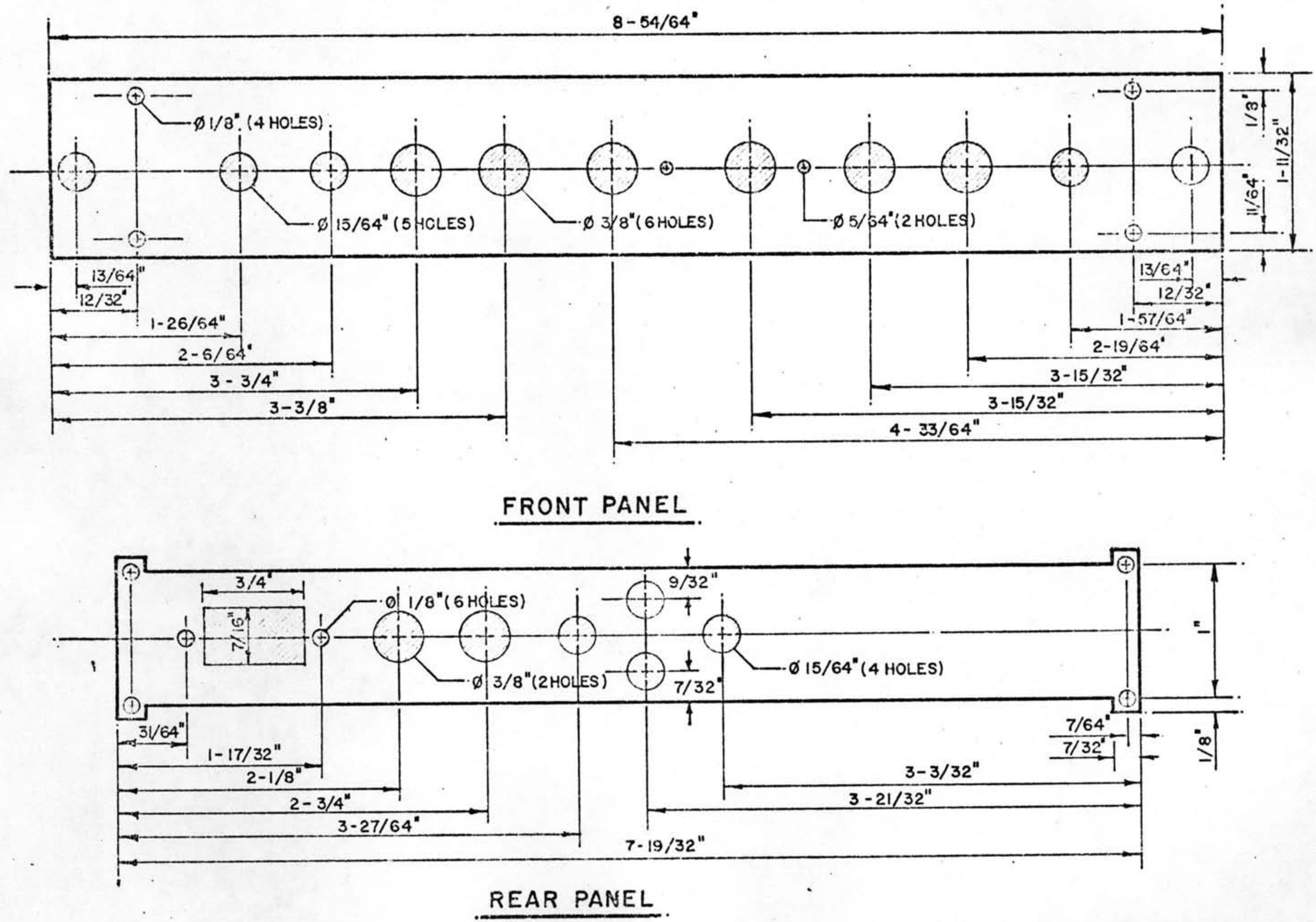
UPPER BEAM : 5V./DIV. AMPLIFIER OUTPUT (BIPOLAR)
LOWER BEAM : 20 mV./DIV. INPUT PULSE
TIME : 2 μs./DIV.



UPPER BEAM : 5V./DIV. AMPLIFIER OUTPUT (BIPOLAR)
LOWER BEAM : 5V./DIV. SCA. OUTPUT
TIME : 1 μs./DIV.



NIM MODULE ASSEMBLY



FRONT PANEL AND REAR PANEL

หน้า COPPER ด้านหลัง