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A Fast Versatile Preamplifier for Use in Particle Physics

by

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A FAST VERSATILE PREAMPLIFIER FOR USE IN PARTICLE PHYSICS

by

Jan K. Studebaker

ABSTRACT

A preamplifier with outstanding characteristics is described. It is built around an NE-592 monolithic integrated circuit. The amplifier has a negative output (0 to -5 V, typical), inverting as well as non-inverting inputs, excellent linearity, easily tailored gain and input impedance, 50 Ω drive capability (delay line clipping may be used) and a total assembled cost of less than \$25. It is readily adaptable to many detection schemes.

I. INTRODUCTION

The need for an inexpensive, yet versatile, linear amplifier of moderate gain has been felt by many users of the Clinton P. Anderson Meson Physics Facility (LAMPF). After careful consideration of their typical requirements, a design has been carried to fruition which incorporates these specified parameters.

II. GENERAL DESIGN GOALS

The most difficult aspect of this project was the determination of important design criteria. The following list was used as a design guide. The items are listed in order of estimated overall importance. These goals were:

- A. Good linearity (2% or better)
- B. 2 kV input transient protection
- C. Fast rise and fall times (10 ns or less)
- D. High reliability
- E. Output short circuit protection
- F. NIM and CAMAC compatible supply voltages
- G. Inverting and noninverting inputs
- H. Readily variable gain (up to 100)
- I. Large output swing (0 to -4 V necessary)
- J. Small physical size
- K. Lowest possible assembled cost
- L. Lowest possible input noise
- M. Variable input impedance (50 Ω to 1k Ω)

N. Current source output for delay line differentiation

O. Direct coupled input and output

P. Low output offset voltage (≤ 20 mV)

Q. High rate capability (1 MHz or greater)

R. No rate effect

S. Low temperature drift of output level and gain

T. Low power consumption (2 W or less)

U. Positive output capability

V. Maximum pulse width ≥ 500 ns.

It is evident that the exact ordering of priorities is difficult and subject to various users' needs. The cost limitation is as responsible for the ordering as is any other major factor.

III. AMPLIFIER DESIGN PHILOSOPHY

The amplifier's design was in many ways a cost/performance trade-off (refer to Sec. IV). However, most of the design criteria were equalled or bettered in the final version.

In order to keep cost, size, and power consumption to a minimum, I decided to utilize integrated circuitry (I.C.s) wherever possible.

A. Input Circuit

I checked all available I.C.s and finally chose the NE-592¹ as a near ideal low-level, high-gain bandwidth, linear amplifier on which to base my design. It is a modern version of the venerable

μ A733 which has proven its worth over a period of several years. The NE-592's only major difference lies in its dual front end current source. This availability allows the user greater freedom in the selection of gain and function (e.g., the gain may be varied by changing one resistor and the amplifier is easily converted to an integrator, filter, or differentiator; see Sec. VI.B.8. The NE-592 helps us to meet many of the design criteria - II. A, C, D, G, H, J, K, L, M, Q, R, T, V).

B. Output Circuit

Unfortunately, no fast linear I.C. is available which can serve as a high current transconductance output stage. It was decided that discrete components arranged in a classical bucking current source and sink combination would be most advantageous to the project. It was further determined (empirically) that a fully dc coupled amplifier would be extremely difficult to temperature stabilize by normal means. Thus, a unique feedback stabilized output stage (i.e., servoamplifier) was incorporated to control dc level. The dc level is sensed by a high gain, low bandwidth amplifier (μ A741) which is contained within the output current sink. It automatically senses and nulls dc level shifts due to temperature, component aging, and rate effect. Because it compensates for rate effect, it allows the use of coupling capacitors in all previous stages. It has the added benefit that no adjustments are required to bring the output to its proper (0.0 V) operating point. This unique output stage further allows us to meet the following design criteria: II. A, C, D, E, F, I, N, O, P, Q, S, V.

IV. DESIGN TRADE-OFFS

As with any design, some trade-offs must be tolerated. The following trade-offs were made in order to facilitate the realization of a practical amplifier.

Physical size (II.J.) was larger than desired because a 100% I.C. design was impossible. Size was maintained to a very workable 4.455 cm x 8.290 cm printed circuit (P.C.) board. If a fast voltage comparator, such as the AM685 had been installed instead of a linear transconductance output circuit, the resulting amplifier/discriminator would have been smaller than my design, but more costly and less versatile. It, however, would have met many

users' present needs and would be an extremely desirable combination.

Cost (II.K.) was higher than hoped, but easily consistent with other design goals. Surely no major cost reductions are possible while maintaining equal or better performance. The overall cost/performance trade-off is impressively gratifying.

Direct current coupling (II.O.) is not utilized in the amplifier, but it is effectively present (to a great extent) because of the self-nulling output stage. The primary benefits of dc coupling are largely preserved, but the drawbacks are effectively eliminated. There is some rate effect (II.R. and III.B.), but much less than in a conventional ac coupled circuit (see Sec. VII.J.).

Positive output (II.U.) is not possible because of the servoamplifier (see Sec. V). The negative output is of the correct polarity for the vast majority of commercially available NIM hardware.

V. CIRCUIT DESIGN DETAILS

Figure 1 illustrates the final design manifestation of Secs. II, III, and IV. A complete parts list is included in the Appendix.

The amplifier is built on a full ground plane and is based on the fastest monolithic, linear gain block available, the NE-592A by Signetics. R14 controls gain by changing the voltage gain of the input differential pair. CR1, 2, 3, 4 provide input overvoltage protection. R2 and R4 limit the current and the dv/dt at the input, so as to protect against fast, high voltage (2 kV) transients. Q1 functions as an impedance converter/buffer. Q2 is arranged as a voltage controlled current source and Q3 operates as a feedback controlled (III.B.) current sink. The feedback (from the output to the positive input of the operational amplifier) adjusts for a zero volt (dc) output regardless of quiescent level shifts due to time, temperature, or component tolerance. Because of the high gain of the μ A741 operational amplifier, the relatively large negative pulses at the feedback input tend to rapidly limit or reverse bias the input transistor, thus limiting the gain at the time of the negative pulse transition. Therefore, to a first-order approximation, the amplifier is effectively out of the circuit at the time of the pulse. This simply means that it is operating with a nonlinear gain function and

NOTES:

1. R14 IS USED FOR GAIN TAILORING. (NECESSARY)
2. C11 IS FOR OVERSHOOT SUPPRESSION. (OPTIONAL)

Fig. 1. Fast versatile preamplifier for use in particle physics.

as such, cannot respond to the true average level of the output wave form. If it did, it would not be capable of taking out "rate effect" level shifts. Further, it is a very slow amplifier and cannot reproduce the fast pulses the amplifier is capable of; if it could, it would tend to cancel the output pulse with a pulse of the opposite polarity. This effect can be seen as "droop" in very long input pulses (see Fig. 6). The output current source (A2) has an effective voltage gain of two (with a nominal output impedance of 50 Ω) so that the limited voltage swing of the NE-592A is doubled. This gives rise to a 4- to 6-V output swing (negative only, from ground). Short circuits on the output have no effect because of current limiting in Q2.

VI. CONSTRUCTION, MOUNTING, AND APPLICATION

A. Construction and Mounting

The amplifier is built on a 4.455 cm x 8.890 cm P.C. board that may be readily mounted in any number of enclosures, including the Pomona #3303 die cast enclosure. This enclosure features premounted BNC connectors and grounding studs for very rapid

drop-in, solder-in P.C. board mounting. The grounding studs on the #3303 are readily soldered to the ground plane on the P.C. board. It is also readily adaptable to NIM and CAMAC modules as well as direct mounting in large phototube bases.

B. Application

This amplifier may be tailored to a number of common applications.

(1) The amplifier's differential inputs permit positive or negative pulses to be amplified to a universal NEGATIVE output.

The inverting input is connected to R1 and the noninverting input is connected to R3. R1 and R3 may be any reasonable value between zero and 2 k Ω .

(2) R2 and R4 may be deleted if minimum rise and fall times (VII.C.) are required.* CR1 through CR4 will still protect the input for normal overloads.

*The input is protected against 2 kV transients. If this protection is not necessary in a given application, it may be removed by shunting R2 and R4 with bare wire. This will improve the response time of the amplifier and effectively remove a response pole. Note, CR1 through CR4 should be retained for overvoltage protection.

(3) Gain may be changed by varying R14 (see VII.A. and Fig. 2), by varying the input impedance in current-to-voltage applications (e.g., phototube), or by connecting a CERMET potentiometer in place of R1 or R3 (this permits screwdriver adjust gain).

(4) The output should normally be terminated in 50Ω. It will drive very long cables and is adaptable to delay line clipping (e.g., zero crossing discrimination).

(5) Overshoot and frequency response may be tailored by adding an output capacitor (C11). Equations 1 and 2 closely approximate the added break point when C11 is used. The cable capacitance does not affect the frequency response if the cable is properly terminated.

$$B_w \approx \frac{0.0032}{C11} \quad (1)$$

$$tr \approx 110 C11 \quad (2)$$

(6) Capacitively coupled inputs (if desired) will not cause the normal "rate effect" phenomenon because of the actively nulled output stage (III.B. and V).

(7) The output as well as the inputs sit nominally at zero volts dc.

(8) Filters may be implemented by adding or substituting inductors and capacitors in series with R14.

a) Low pass filters are achieved by adding an inductance (e.g., miniductor) in series with R14. R14 will set the midband gain as usual (VII.A. and Fig. 2). The low end will be determined primarily by the servoamplifier limitations (VII.I. and Fig. 6).

b) High pass filters are implemented by adding a capacitor in series with R14. Again, R14 determines midband gain and the upper break point is determined by either C11 (Eqs. 1 and 2) or the maximum frequency response of the amplifier (see VII.C. and VI.B.2,5).

c) Bandpass filters are merely an inductor and a capacitor in series with R14.

d) Band reject filters require a parallel tuned circuit in series with R14.

VII. SPECIFICATIONS AND PERFORMANCE

The following specifications (A to J) and Fig. 2. give a quantitative representation of important performance parameters. The specifications are arranged in tabular format, so that they may be easily referenced. Figures 3, 4, 5 and 6 give digestible qualitative performance proof and indicate important application situations. Figure 6 is discussed in Sec. V.

A. Gain Versus R14 Values:

R14	Gain	R14	Gain
46	190	464	31
68	148	649	25
103	110	715	22
120	100	787	20
140	88	866	17.5
178	73	953	16
215	63	1000	15
261	53	1100	14
301	48	1270	12
348	40	1470	10

B. RMS Noise:

With Full Input Protection Circuitry

36.5 μV (RMS) wideband input noise (100 MHz bandwidth) or 3.65 nV/√Hz input noise.

With R2 and R4 Removed (200Ω Input Protection Resistors)

40 μV (RMS) wideband input noise (100 MHz bandwidth) or 4 nV/√Hz input noise.

C. Response Time: (See Figs. 3 to 6.)

With Full Input Protection Circuitry

Leading Edge ≈ 12 ns @ A=100 (3 V into 50Ω)

Trailing Edge ≈ 10 ns @ A=100 (3 V into 50Ω)

Leading Edge ≈ 8 ns @ A=50 (3 V into 50Ω)

Trailing Edge ≈ 9 ns @ A=50 (3 V into 50Ω)

With R2 and R4 Removed (200Ω Input Protection Resistors)

Leading Edge ≈ 6 ns @ A=50-100 (3 V into 50Ω)

Trailing Edge ≈ 7 ns @ A=50-100 (3 V into 50Ω)

D. Delay Time: (Input to Output @ 1/2 V)

With Full Input Protection

td ≈ 13 ns

With R2 and R4 Removed (200Ω Input Protection Resistors)

td ≈ 10 ns

E. Power Requirements:

+12 V @ 70 mA

-12 V @ 70 mA

F. Linearity: (A=100 or less)

≈ 2% from 0 to -4.5 V with -5 V saturation level

G. Output Offset Voltage:

$V_{OFS} \approx 5$ mV (no signal condition)

H. Output Drive Capability: (A=100)

Minimum output amplitude = 4 V into 50Ω

Typical output saturation level = 5 V into 50Ω

I. Maximum Pulse Width:

1 → 2 μs because of droop (≈5%) and because of overshoot (≈10%)

J. Offset Voltage Due to Repetition Rate:

+20 mV @ ≈ 3 MHz (25 ns FWHM output pulse width)

+40 mV @ ≈ 5 MHz (25 ns FWHM output pulse width)

NOTE: The above is true from 1 to 4 V output amplitude (into 50Ω).

For Long Pulses a duty factor limit is most meaningful.

+20 mV with 25% duty factor

+40 mV with 37% duty factor

VIII. PRODUCTION TEST PROCEDURES

A. Equipment Needed:

- (1) Fast pulse generator with following output characteristics:
 - a) 4 ns rise and fall
 - b) 10 to 100 mV output amplitude (adjustable)
 - c) Negative and positive outputs (not needed simultaneously)
 - d) Repetition rate of 10 kHz
- (2) 150 to 200 MHz oscilloscope
- (3) Positive and negative 12-V power supply
- (4) 50Ω line terminator (BNC)
- (5) Cables:
 - a) Power supply (3 conductor #20)
 - b) Two 50Ω coaxial cables with BNC connectors on one end and stripped carefully at the other.

B. Technique:

All connections to the amplifier must be made very carefully, with the shortest possible leads. This is a very fast amplifier and requires good technique.

- (1) Connect power supply cable between amplifier terminals and power supply. (Power supply OFF.)
- (2) Solder on stripped coax leads to POSITIVE input and to output (coax shields must be soldered to the P.C.'s ground plane).
- (3) Terminate output in 50Ω line terminator and connect to the vertical input of the oscilloscope.
- (4) Do not connect input coax to the pulse generator.
- (5) Make sure R14 is plugged in; its value should be 121Ω.
- (6) Turn on power supply.
- (7) Check output dc level; it should be within a few millivolts of ground; put aside any amplifier that has a large dc offset.
- (8) Check output noise; it should be about 2 full centimeters on the 5 mV scale. Regardless of amplitude, measure all amplifiers on the same sensitivity, using the same triggering mode, the same intensity and the same sweep speed, since these variables affect the "viewed" noise amplitude. Put aside any amplifier that is significantly different from the majority.
- (9) Connect the input coax cable to the pulse generator (negative output) and verify the 12 ns response time, the full 4 V output level, and look for severe distortions. Put aside all amplifiers that do not come close to the published "typical" specifications.

NOTE: If a 4 V pulse cannot be obtained, replace the NE-592A.
- (10) Turn off power supply.
- (11) Switch the input coax to the NEGATIVE input.
- (12) Switch the input coax to the POSITIVE output of the pulse generator.
- (13) Turn on power supply.
- (14) Verify test data, #9.
- (15) Turn off power supply.
- (16) Test other units as described above.

IX. TROUBLESHOOTING

Standard "fast" troubleshooting procedures are required for fault location. In general, start by a close visual inspection. If that fails, use careful signal tracing and component substitution.

REFERENCE

1. "Linear Product Specifications" in Signetics Integrated Circuits, (1972) pp. 6-91 to 6-96.

Fig. 2. Voltage gain vs local feedback (R14).

Fig. 3. Shows the amplifier's performance with a gain of 100 and full input protection.

Fig. 5. Shows the same data, but with the input pulse width adjusted so as to illustrate the shortest possible input pulse that results in FULL output amplitude.

Fig. 4. Shows the same data with R2 and R4 removed. Notice the expanded time base.

Fig. 6. Illustrates the output waveform when a long pulse is applied to the input. Notice the pulse "droop" and the induced baseline ripple.

APPENDIX
ELECTRICAL PARTS LIST

<u>Circuit Designation</u>	<u>Name of Part</u>	<u>Description</u>	<u>Quan. Req'd.</u>	<u>Circuit Designation</u>	<u>Name of Part</u>	<u>Description</u>	<u>Quan. Req'd.</u>
R1,3	Resistor	51 Ω , 1/4 W, carbon	2	C1,2,4,5,6,8,9,10	Capacitor	0.47 μ F redcap	8
R2,4	Resistor	200 Ω , 1/4 W, carbon	2	C3	Capacitor	0.22 μ F redcap	1
R5	Resistor	1k Ω , 1/2 W, carbon	1	C7	Capacitor	0.047 μ F redcap	1
R6	Resistor	180 Ω , 1/4 W, carbon	1	C11	Capacitor	27 pF capacitor	1
R7	Resistor	11k Ω , 1/4 W, carbon	1	Q1,2	Transistor	2N3300	2
R8	Resistor	820 Ω , 1/4 W, carbon	1	Q3	Transistor	2N3502	1
R9	Resistor	10 Ω , 1/4 W, carbon	1	VR1,2	Diode	5.1 V, 1 W, Zener 1N4733	2
R10	Resistor	100k Ω , 1/4 W, carbon	1	L1,2,3,4	Choke	Inductor, 0.47 μ H	4
R11	Resistor	150 Ω , 1/4 W, carbon	1		Integrated Circuit	NE592	1
R12	Resistor	24 Ω , 1/2 W, carbon	1		Integrated Circuit	μ A741	1
R13	Resistor	1k Ω , 1/4 W, carbon	1	Test Clips	Spring loaded (for R14)		2
R14	Resistor	121 Ω , 1/8 W, precision	1	Terminal Lug	Stake-in		5
CR1,2,3,4	Diode	1N4154	4	Socket	14 pin, IC		1
				Socket	8 pin, IC		1
				P.C. Board			1

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