INSTRUMENT REFERENCE BOOK

for the Tektronix type

O

operational amplifier
plug-in unit

For all serial numbers

MPI EXTRACT

MPI January 1964

O (Cont'd)		
Adapters tentatively scheduled for production in mid-1963*	FEN	11-30-62
"Bandwidth" confusion*	FEN	1-12-62
Basic characteristics and operations*	A-2078-4	2-63
Calibration procedure*		061-416
Characteristics*	A-2078-4	2-63
"Constant intensity" displays (also see 3-9-62 FEN for additional information)*	FEN	2-23-62
Demo kit containing adapter hardware*	FEN	11-30-62
Gating adapter, characteristics*	A-2183	8-63
Grid-current checker available*	FEN	7-27-62
Grid current spec changed (see correction, 5-25-62 FEN)*	FEN	5-11-62
Instruction Manual, composite, SN 101 up*		070-323
LF Reject circuits: theory and limitations*	FEN	7-12-63
Log Amp Adapter, characteristics*	A-2167-1	6-63
Log Amp Adapter, data sheet		061-667
Log Amp Adapter, Instruction Manual*		070-386
Maintenance spares list		061-498
Price increase to $525.00, effective March 1, 1963*	SPR-134A	2-11-63
Stopless pot mod*	FEN	5-11-62
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AC-coupler adapters available*	FEN	11-30-62
Accessories included*		061-499
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Adapter hardware for demo's*	FEN	11-30-62
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*Included within IRB.

SALES

Inter-City Mfg. Co., Inc.
St. Louis 11, Mo.

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CATALOG

PRICE INCREASES

SPR-134A 2-11-63

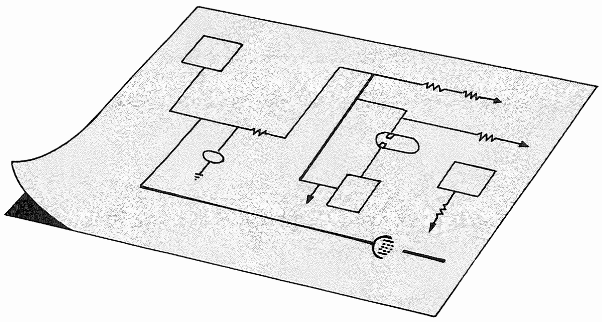
On March 1, 1963 the following instrument price increases will be effective:

Instrument Type	Catalog 21 Present price	New price as of 3-1-63
"O" Operational Amplifier	$475.00	$525.00

Any order dated prior to March 1, 1963 will be honored at the present Catalog Price.

Any order resulting from a quotation dated prior to March 1, 1963, will be honored at the present price. Quotations made on March 1st and later should be at the new price.

Customer Purchase Orders received in the Field *after* March 1, 1963, that did not result from a quotation at the present price, should show the new price or be referred back to the customer for correction.



SECTION 1

CHARACTERISTICS

General Information

The Type O Operational Amplifier Plug-In Unit consists of essentially three parts: a vertical preamplifier and two operational amplifiers. The vertical preamplifier can be used either as an independent oscilloscope preamplifier or to monitor the output of either of the operational amplifiers.

The operational amplifiers can be used for applications involving integration, differentiation, amplification by a constant factor, summation, and phase inversion (as well as many others; see Section 3). The output of one operational amplifier can be applied to the input of the second for combined operations.

The Type O Unit can be used with any of the Tektronix 530-, 540-, or 550-Series Oscilloscopes. It can also be used with the 580-Series Oscilloscopes in conjunction with the Type 81 or Type 81A Plug-In Adapters. The Type O Unit can be used with other oscilloscopes and devices through use of the Types 127, 132, or 133 Plug-In Power Supplies.

Vertical Preamplifier

Bandpass

Dc to 14 mc (3 db) in Tektronix Type 530-Series Oscilloscopes (except Types 532 and 536).

Dc to 25 mc (3 db) in Tektronix Types 540- and 580-Series Oscilloscopes, and the Type 555.

Risetime

Approximately 25 nsec in Type 530-Series Oscilloscopes (except Types 532 and 536).

Approximately 14 nsec in Type 540- and 580-Series Oscilloscopes, and the Type 555.

Vertical Deflection Factors

Nine calibrated steps provided: 0.05, 0.1, 0.2, 0.5, 1, 2, 5, 10, and 20 volts per centimeter. A variable uncalibrated control provides for continuous adjustment from 0.05 to 50 volts per centimeter.

Input Characteristics

Approximately 1 megohm paralleled by 47 pf.

Operational Amplifiers A and B

Open Loop Gain-Bandwidth Product

Approximately 15 mc. (Checked at 10 mc with 1-v input; see Chart 2-1.)

Open Loop DC Gain

When the input signal is applied to the —grid, with the +grid grounded, the gain is approximately —2500.

If the signal is applied to the +grid and the —grid is grounded, then the gain is approximately +2500.

Now if the signal is applied to both grids, the gain is found by using the formula:

$$A = \frac{e_o}{e_g\ (-\text{grid}) - e_g\ (+\text{grid})} = -2500$$

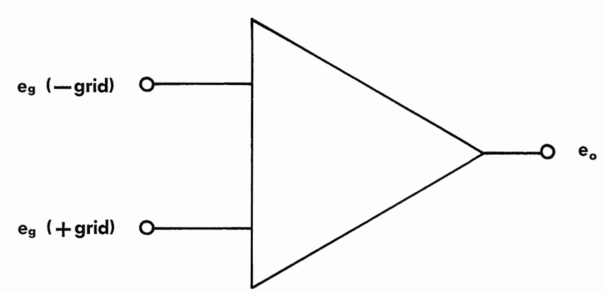


Fig. 1-1

Output Range

± 50 volts, ± 5 ma.

Output Impedance, at Front-Panel OUTPUT Connectors

Approximately 30 Ω at 1 mc for compensated unity-gain amplifier.

Noise

Typically 0.5 mv peak-to-peak, referred to input. Approximately 3 mv peak-to-peak output noise.

Drift

Typically less than 10 mv per hour (after warmup), referred to input.

Grid Current

Less than 0.5 nanoampere for both + and — input grid. Adjustable to less than 0.15 nanoampere for + grid, and less than 0.3 nanoampere for — grid.

Input Impedance

Selected by front-panel control. Values contained internally are: 0.01, 0.1, 0.2, 0.5, and 1 megohm; 10 pf*, 0.0001*, 0.001, 0.01, 0.1 and 1 μf, at $\pm 1\%$ (refer to calibration procedure). Other values may be connected externally.

Feedback Impedance

Same values and tolerances as the Input Impedances internally. Other values may be connected externally.

Signal Inputs

Signals may be connected to either the —grid (output inverted) or the +grid (output polarity same as input).

Feedback

Provision is made for permitting either positive or negative feedback.

Integration Low-Frequency Rejection

A low-frequency rejection circuit is provided to prevent undesired integration of dc components and dc drift from forcing the oscilloscope trace off the crt. It is also possible to reject line-frequency pickup and other low-frequency noise. Rejection will occur at about 1 cps or about 1 kc, depending on the setting of a front-panel control. The low-frequency rejection circuit may be switched in or out as desired.

* Individually adjustable.

Output DC Level

At ground potential. Output is adjusted to ground with a front-panel control.

Crosstalk Between Operational Amplifiers

Typically better than 400:1 under following conditions: Both Operational Amplifiers set for unity gain with $Z_i = Z_f = 1$ MEG, one amplifier driven with a capacitively-coupled oscilloscope Amplitude Calibrator signal of 100 volts ($\pm$50-volt square wave of about 1-μsec risetime). Output of other amplifier will not exceed 330 mv.

Other Characteristics

Construction

Aluminum-alloy chassis with Anodized panel.

Accessories

013-0048-01	2	ADAPTER, terminal
013-0049-00	2	SHIELD, terminal
103-0033-00	2	ADAPTER, BNC to binding post
012-0087-00	2	CORD, patch
070-0323-00	2	MANUAL, instruction

TYPE O

TEXT CORRECTION

Section 1 Characteristics

Page 1-2 Accessories

REMOVE:

103-0033-00	2	ADAPTER, BNC to Binding Post

This insert is placed in its appropriate position in your Product Reference Book and printed on colored paper to expedite retrieval. In a standard manual, it will be filed at the back of the manual.

MODIFIED PRODUCTS

Product	Mod	Description
0	113V	Add drag brakes.

Inter-Office Communication

To: Field Engineers
Field Maintenance Engineers
Regional & District Managers

Date: December 14, 1962

BEAVERTON

From: Field Training

Subject: O Unit Demo Adapter Kit

The accompanying write up on the O Unit Demo Adapter Kit will serve the purpose of aquainting you with several O Unit Demo Adapters for use as a sales tool.

Besides complete schematic information on four specific adapter circuits, you will find some good general information on the O Unit itself in the Question & Answer section.

The write up is one which accompanies a basic kit consisting of two blank boards, punched as for 013-048 and the necessary banana plugs and associated hardware for mounting same. These are available from Field Training.

Parts for the individual circuits may be obtained either through local sources or through regular Tek stock numbers.

Best regards,

Arch F. Brusch

Arch F. Brusch
Field Training Department

pm

O- UNIT

DEMO ADAPTER KIT

O-UNIT DEMO ADAPTER KIT

Contents:

2 - Blank boards, punched as for 013-048

12 - 134-014

Nuts and Hardware

Suggested Uses:

1. Construction of basic demonstration circuits, adequate to demonstrate:

   a. The principles of operation of more elaborate circuits, for specific applications.

   b. The general concept of O-Unit versatility.

   c. The use of external components for odd-values and trimming, for the three basic functions (amplification, differentiation, integration).

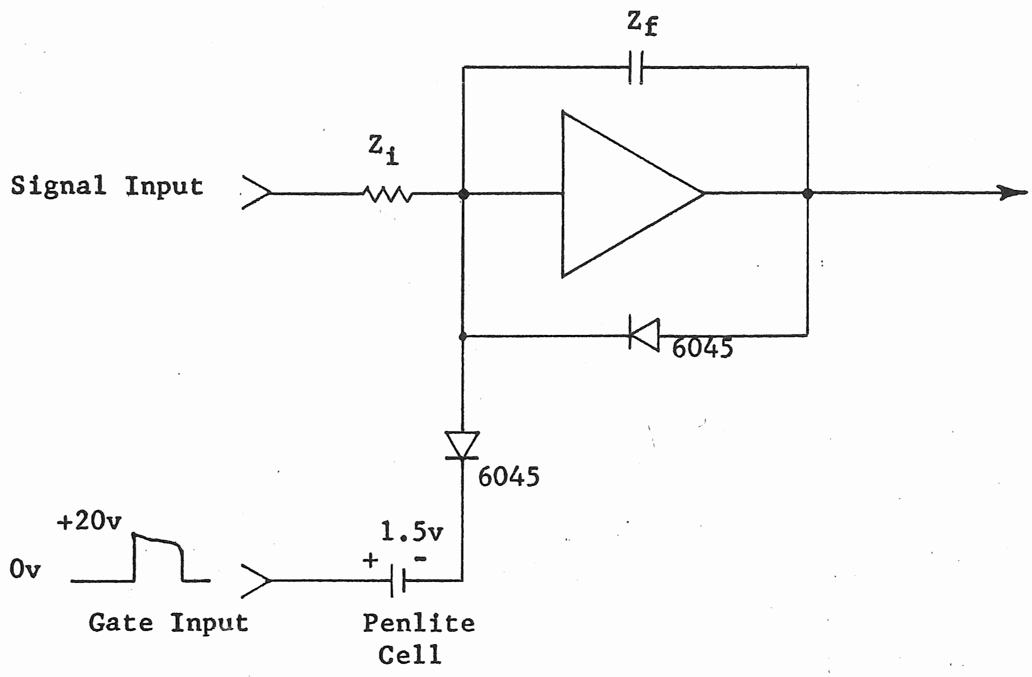
   d. The idea that external accessories are easy to build.

2. Use as a "quick change" foundation for the 013-048 adapter.

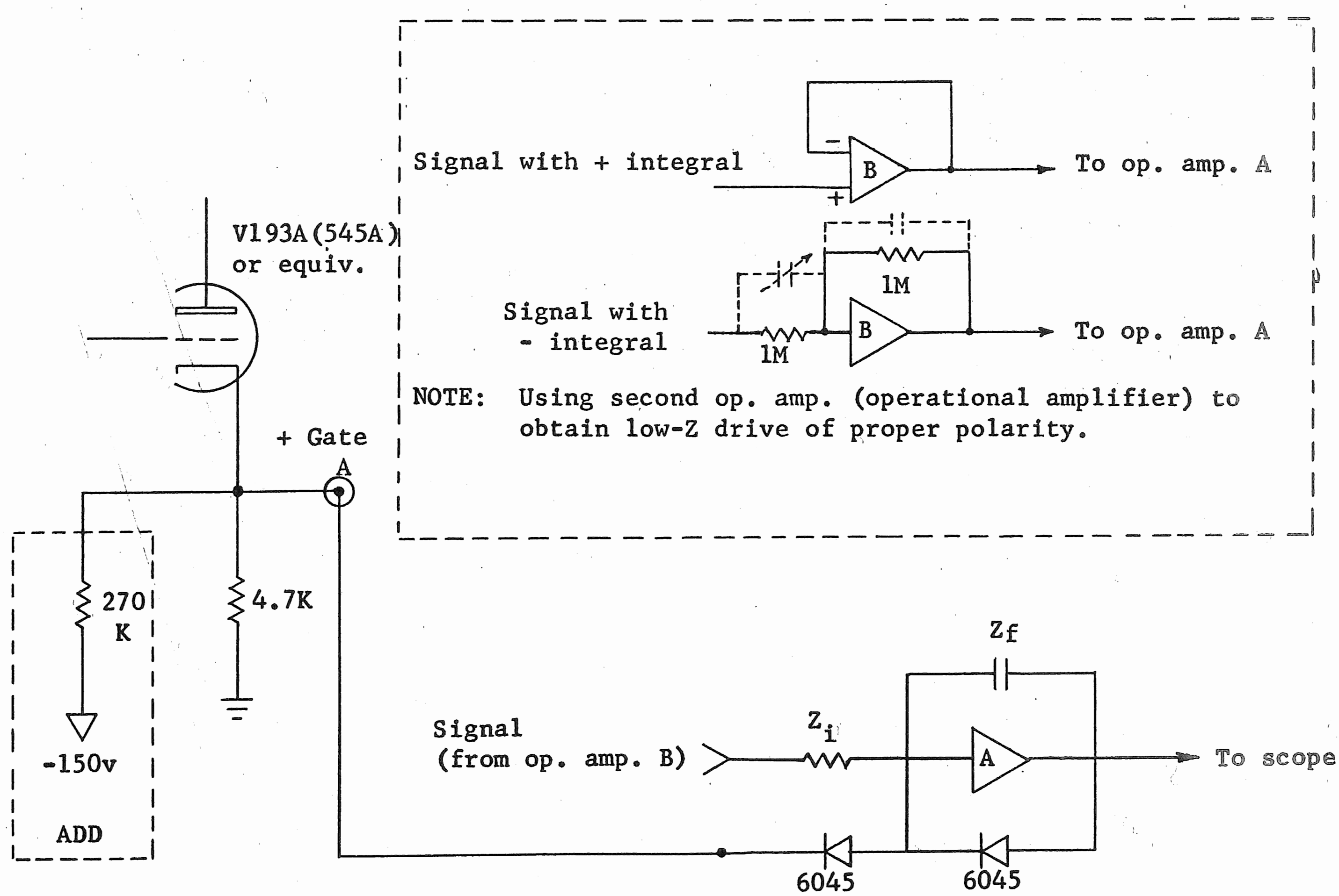
Suggested Circuits:

1. Gated integrator (+ going signals only)
2. Log amp adapter (+ going signals only)
3. Diode Leakage Tester
4. Slideback amplifier with large-signal compression. (Suggested accessory: calibrator rectifier adapter)
5. LF Function Generator

1. GATED INTEGRATOR
(+ Going Signals Only)



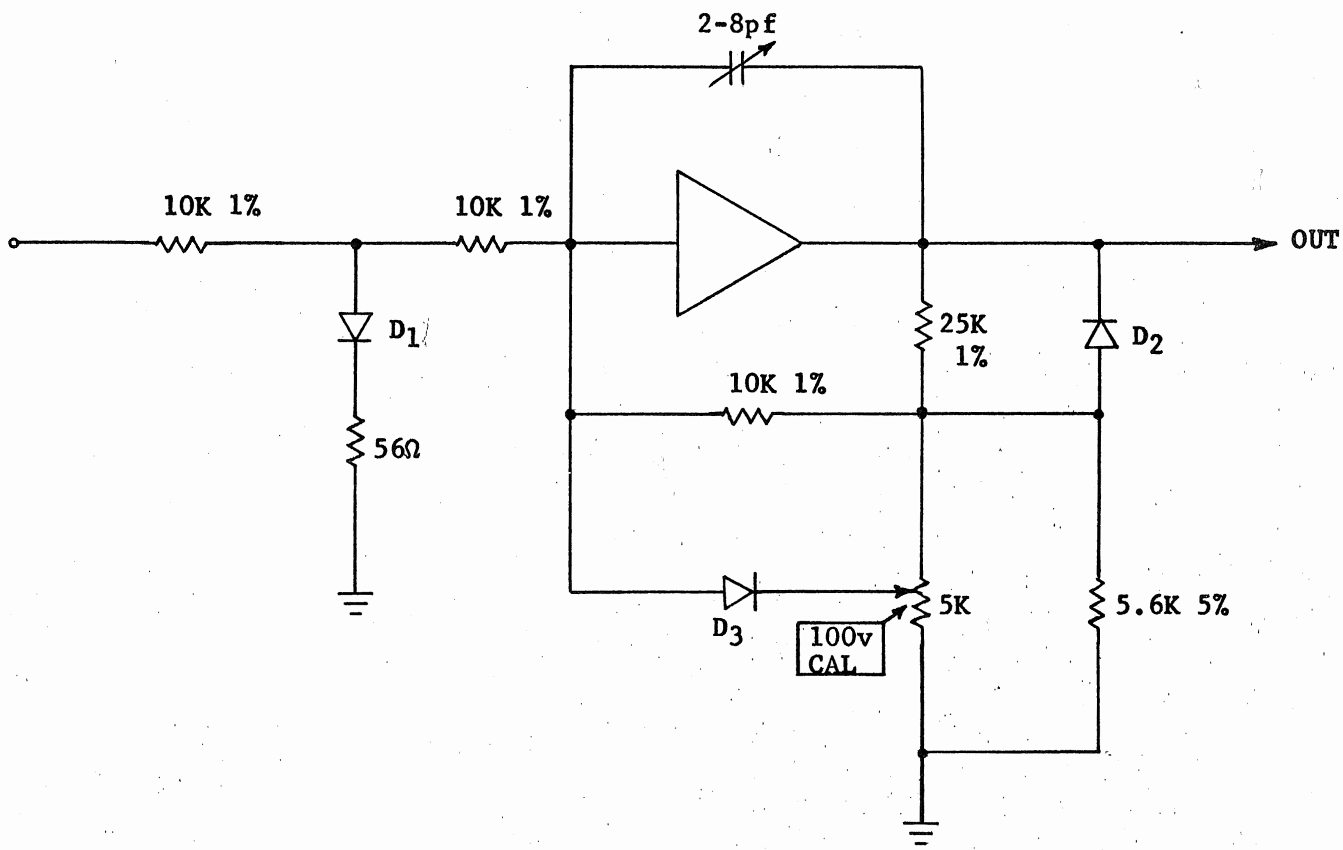
A. Circuit using penlite cell.



B. Circuit using scope mod. 270K resistor causes + gate to start at -2v.

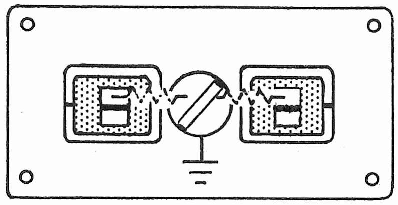
2. LOG AMPLIFIER ADAPTER

For + going signals only (0.1-100v)
Max (-) Input (Destruct Level): 15v



		(Select for (fwd)
D_1	ERIE 2007 (152-071)	0.10 - 0.13v at 10 μa 0.35 - 0.50v at 10 ma
D_2	ERIE 2007 (152-071)	0.10 - 0.13v at 10 μa 0.25 - 0.30v at 1 ma
D_3	HUGHES 5000 (152-065)	0.25 - 0.40v at 10 μa 0.35 - 0.50v at 100μa

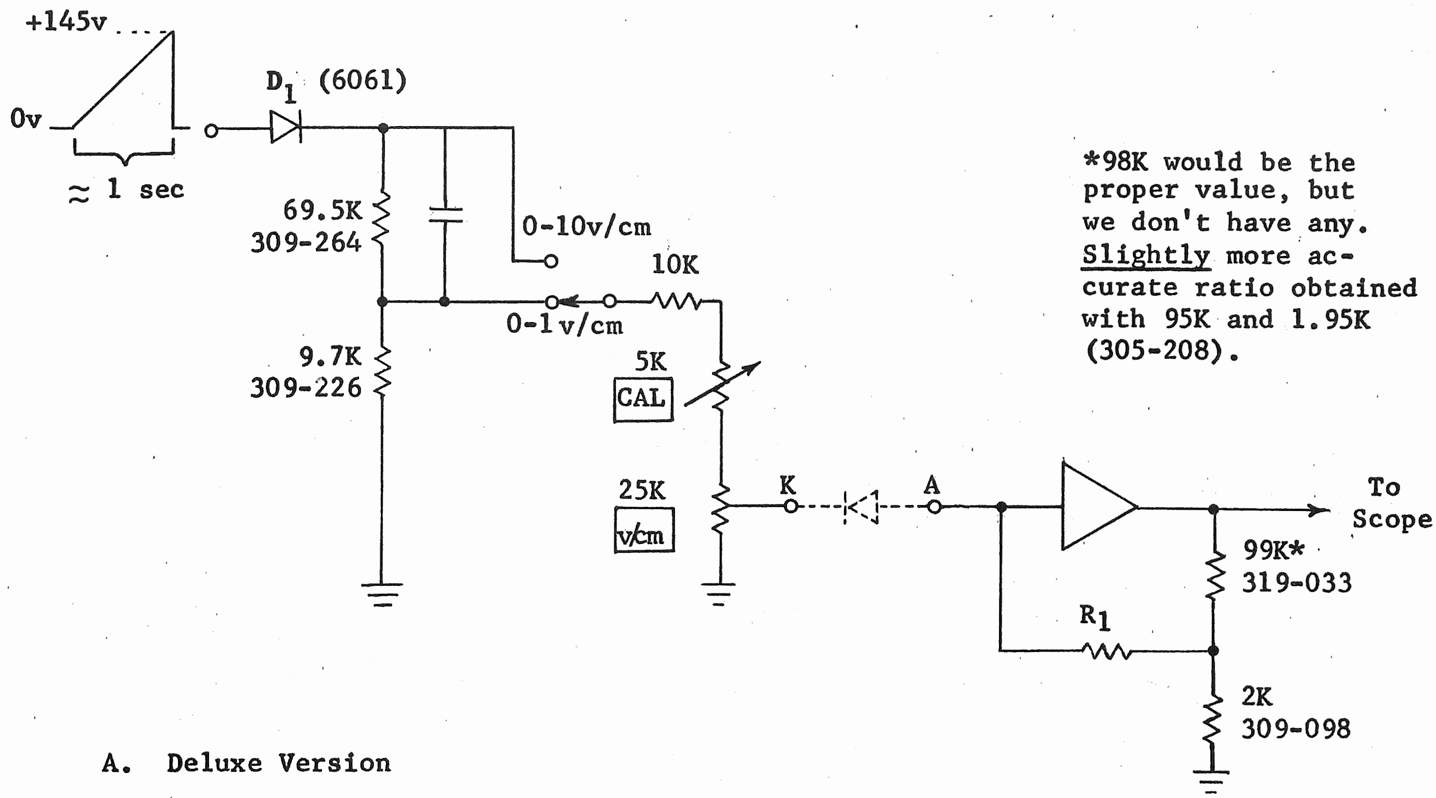
3. DIODE LEAKAGE TESTER



432-032

S-Unit Diode Holder. Mount with large binder head screw to ground all leakage paths thru plastic.

(D_1 prevents negative quiescent level of sawtooth from turning on test diode.)



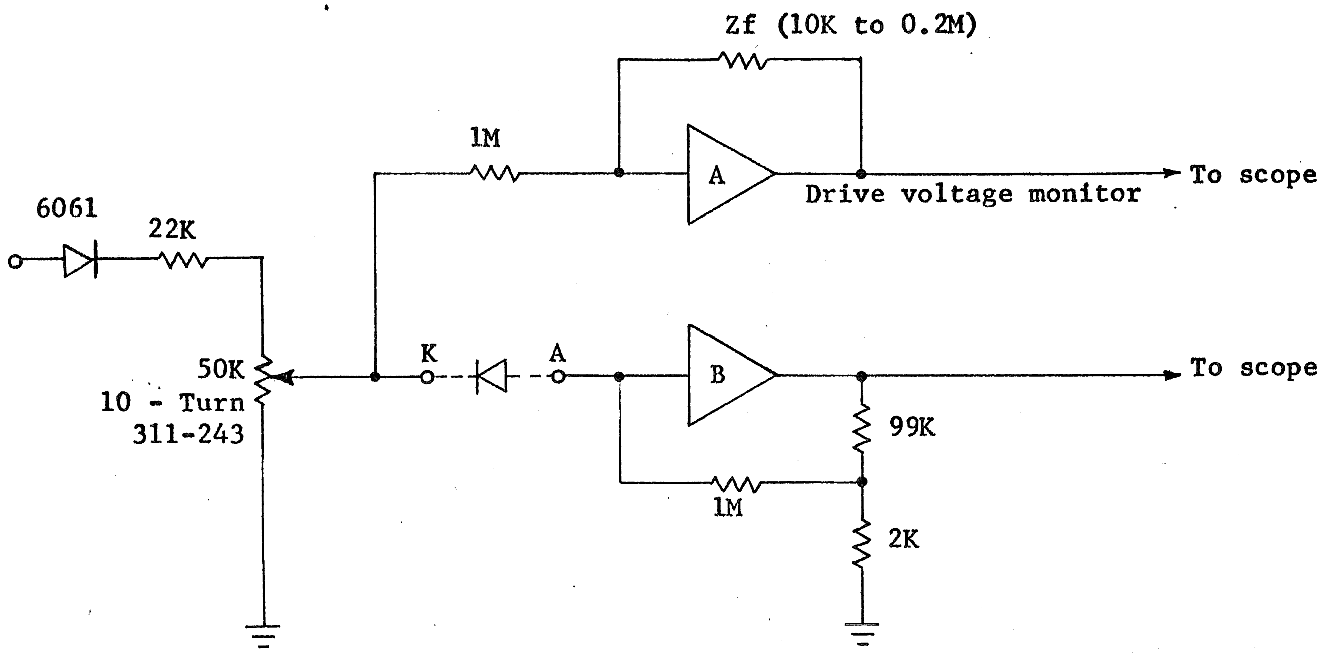
*98K would be the proper value, but we don't have any. Slightly more accurate ratio obtained with 95K and 1.95K (305-208).

A. Deluxe Version

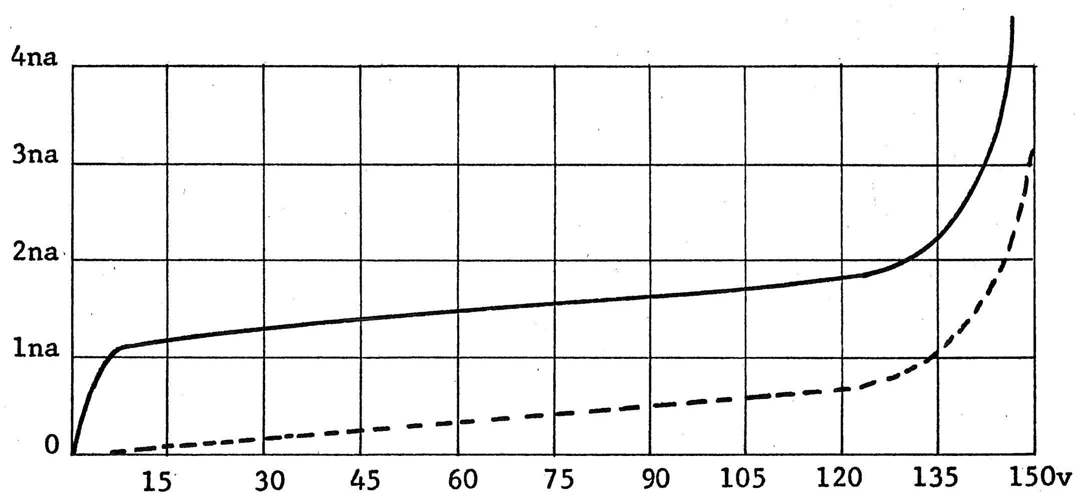
R_1 TABLE (1%)

SENSITIVITY	R_1	SCOPE
1μa/cm	100K	0.5v/cm
0.1μa/cm	100K	50mv/cm
10na/cm	1M	0.5v/cm
1na/cm	1M	50mv/cm

3. DIODE LEAKAGE TESTER (cont.)



B. Simple Version (0-10v/cm, 1na/cm at 50mv/cm)



——— Display at 20 msec/cm.

- - - - Display at 100msec/cm.

Step at start of display indicates diode capacitance. To reduce C effect, use slower Time/cm.

To measure C, set sweep Time/cm to provide display uniformly displaced from normal (dotted line) display. Vertical displacement is proportional to C.

$$C = \frac{\text{(Vertical displacement)(Time/cm)}}{\text{Volts/cm}}$$

Where vertical displacement is in amperes.

In example shown, vertical displacement = 1 nano ampere

Time/cm = 20 msec/cm

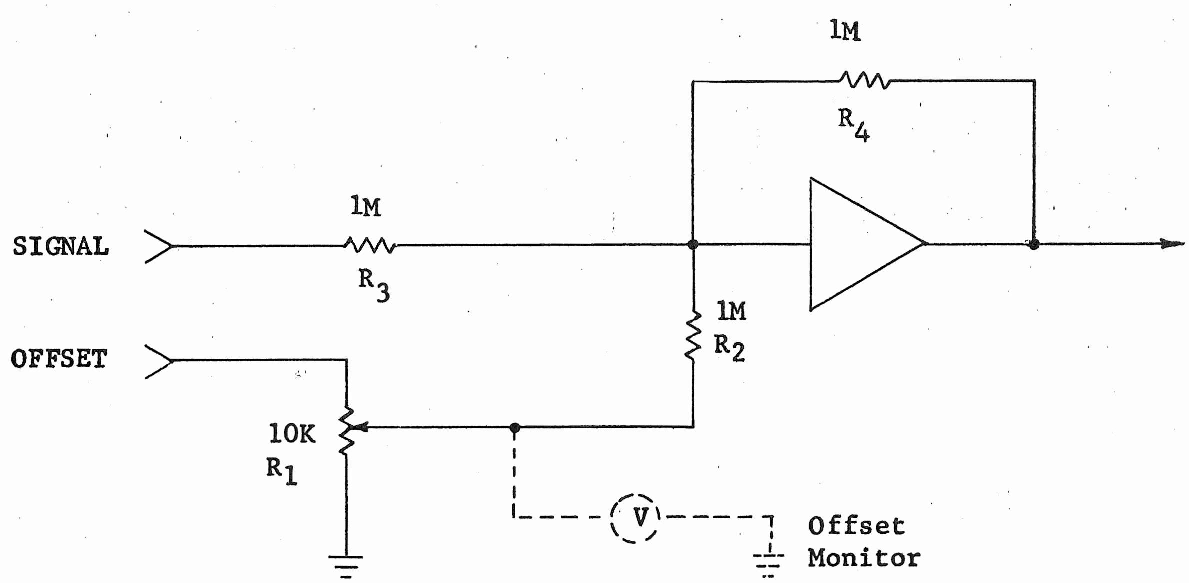
Volts/cm = 15v/cm

$$C = \frac{(10^{-9})(20\times10^{-3})}{15}$$

$$= 1.33pf.$$

NOTE: Too high a sweep repetition rate can introduce errors at start of trace. Operation in "Auto" mode recommended.

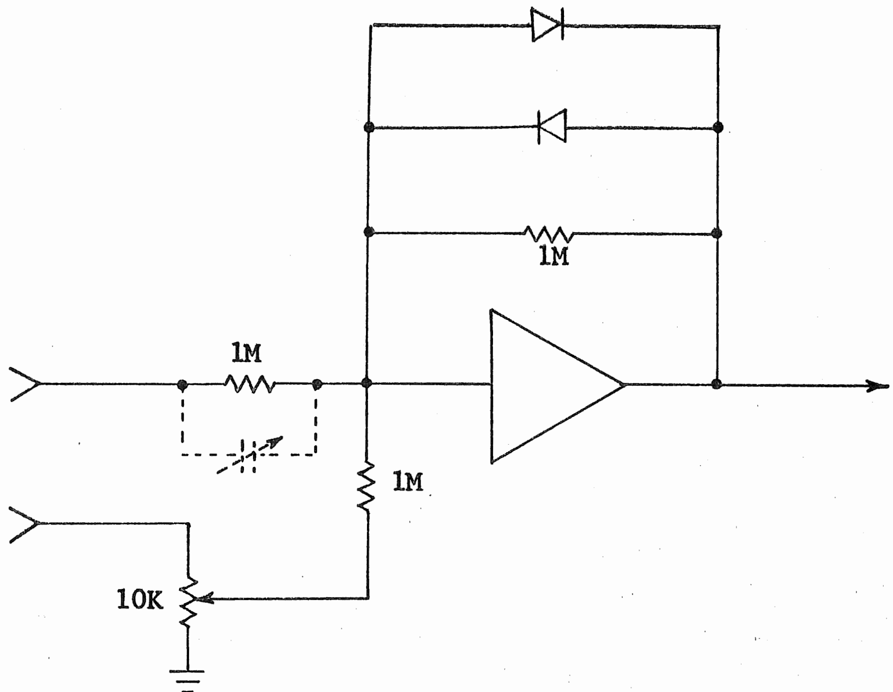
4. SLIDEBACK AMPLIFIER
with optional refinements.



A. BASIC CIRCUIT

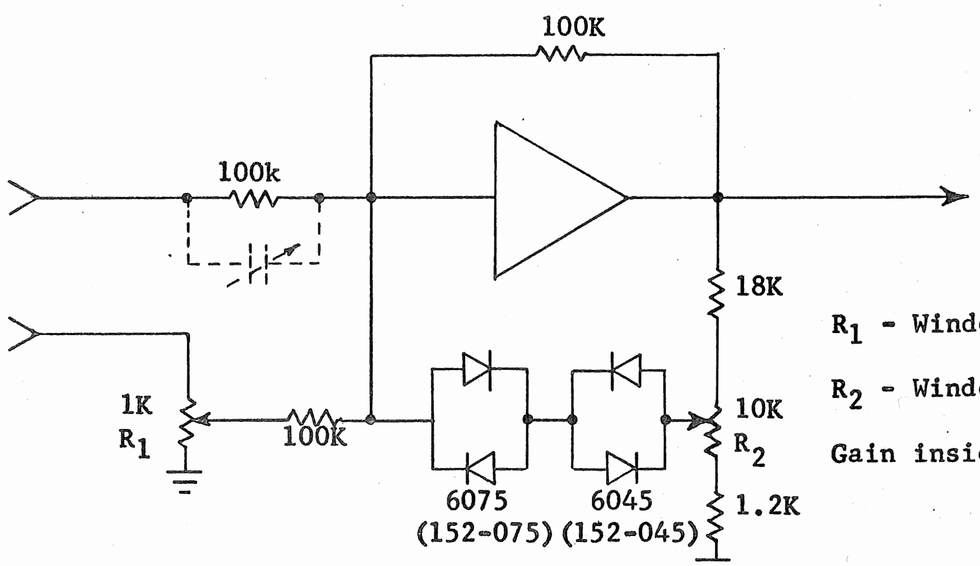
For <1% measurements, match R_2 & R_3 <1/2%. R_4 determines only <u>resolution</u>, (gain), not DC accuracy. R_1 should be low Z only if no offset monitor is used.

4. SLIDEBACK AMPLIFIER (cont)



B. WITH COMPRESSION

Diodes: T13G, 152-005, or any low-leakage germanium for smooth turnover; use silicon for sharp turnover.



R_1 - Window DC Level

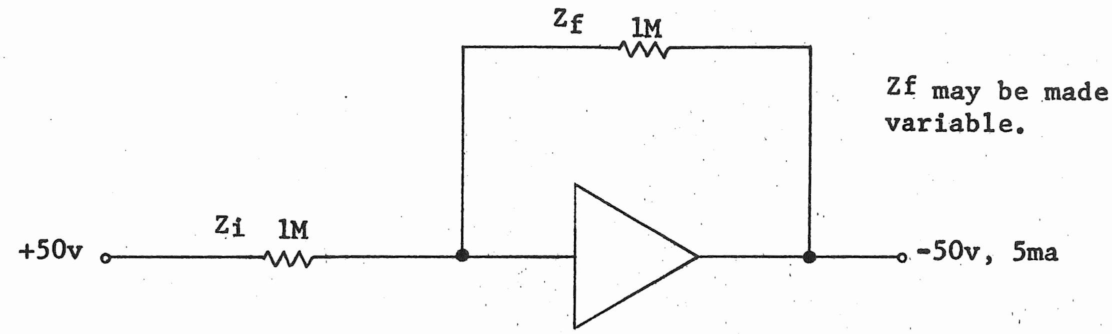
R_2 - Window Size (±1 to±10v)

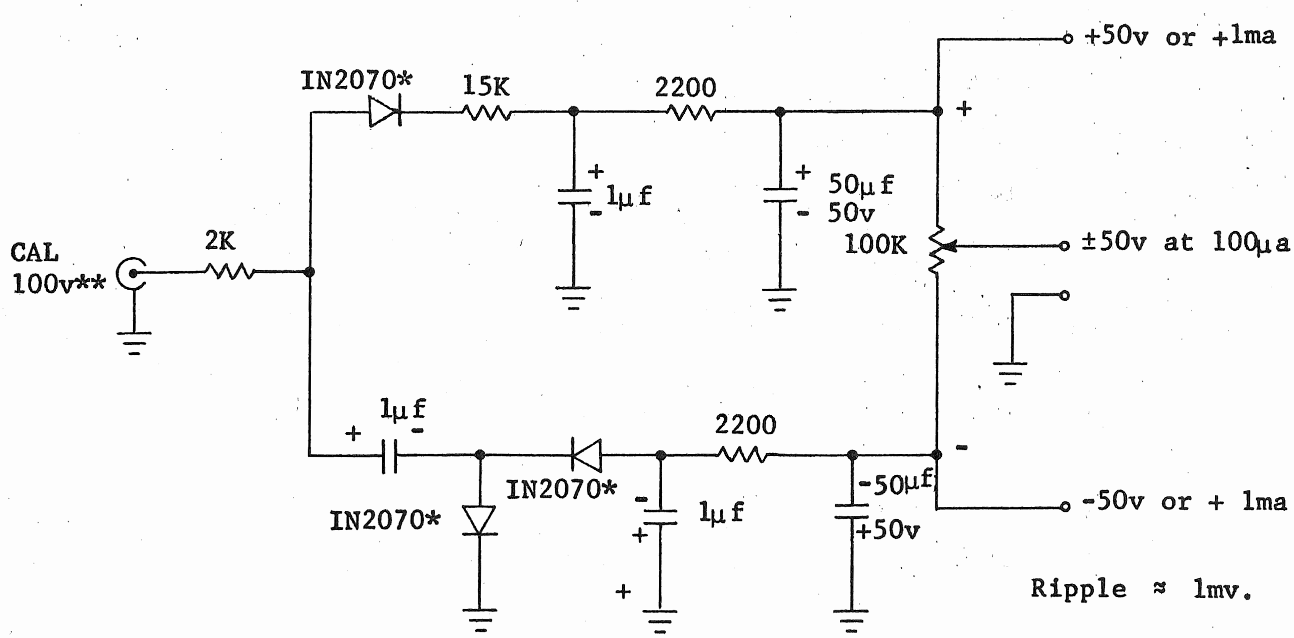
Gain inside of window is X1.

C. WITH COMPRESSION
OUTSIDE OF LINEAR "WINDOW"

CALIBRATOR RECTIFIER ADAPTER

NOTE: Where + DC voltages only are needed, it's usually easier (in 530-40A Series) to simply remove V875, and take desired voltage from Cal Out jack (5ma max). If one operational amplifier only is needed, the other may be used as 0 to -50v power supply (5ma max) by feeding + volts from calibrator to (-1) amplifier.





* or equiv.
50µf 50v 290-117
1µf 150v 290-164

**Bring calibrator up to 100v in steps to avoid heavy initial loading (50 ma!).

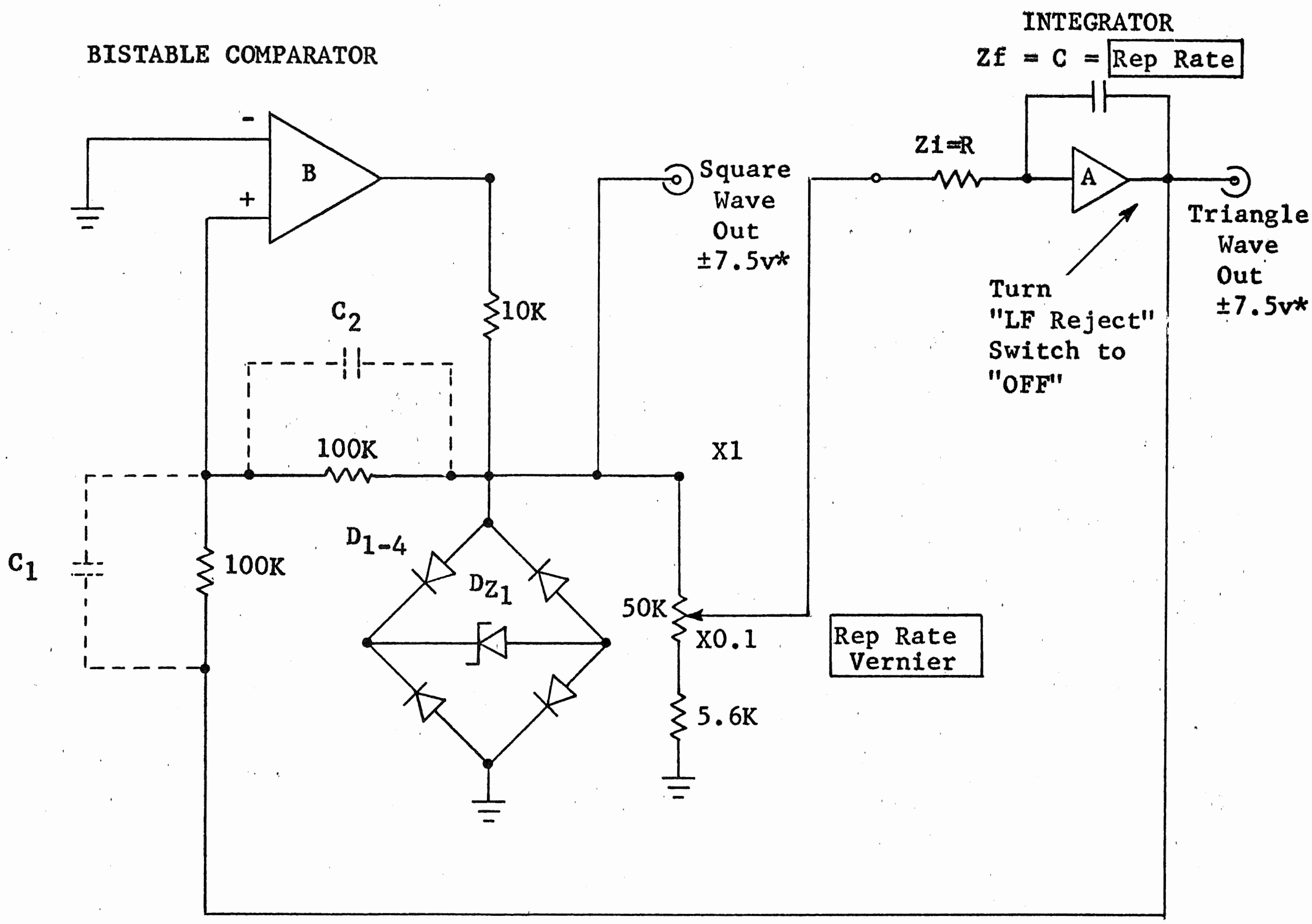
Construction Time: 20 min.

5. LOW FREQUENCY FUNCTION GENERATOR

.025 cps to 25 kc with Zi = R at 1M.

.25cps to 250 kc with Zi = R at 100K**.

(Rep-rate = $\frac{1}{4RC}$ X vernier setting)



*Square and sine wave output dependent on D_{Z_1}.

D_1 - D_4 = 6045 (152-045)

D_{Z_1} = IN707 (152-004) for 15v p-p output (±2v).
1/4 M10Z10 (152-064) for 21v p-p output (±2v).

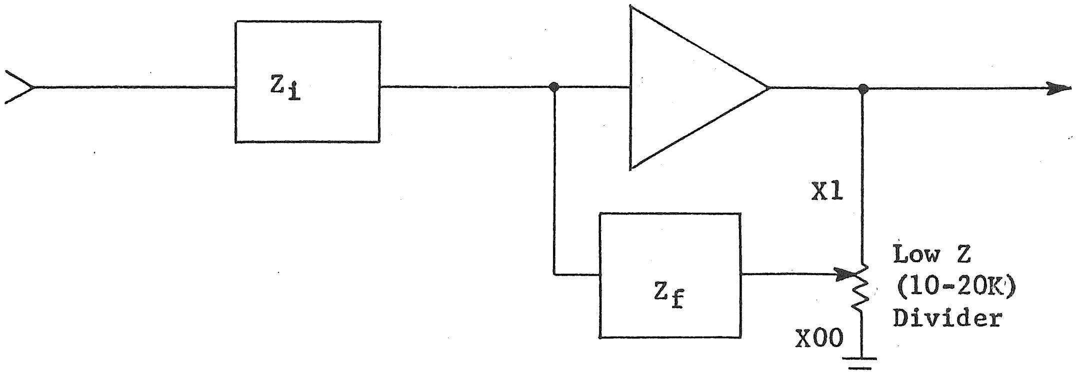
C_1 Required when fastest rep-rates used.

C_2 May be desirable for fast turn-around.

**For accurate vernier cal, keep Zi > 500K, or reduce vernier to 20K and 2.2K (lowest permissible value!).

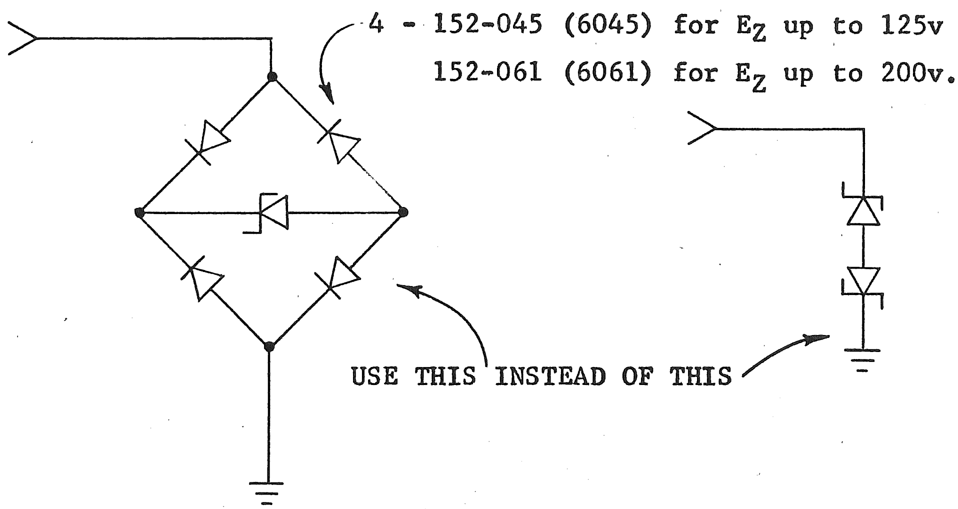
BASIC CIRCUIT TRICKS:

1. "MAGNIFYING" THE VALUE OF Zf (increasing the output excursion required to maintain a null).

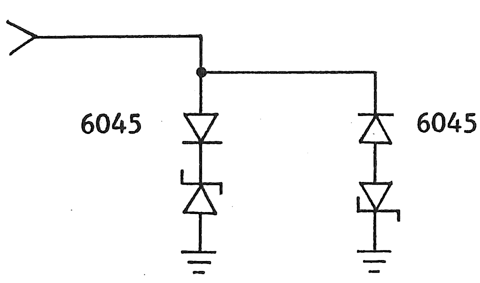


LIMITATION: Cannot provide high values of current.

2. OBTAINING A "±" ZENER WITH LOW SHUNT C: (very close symmetry)



Where specific asymmetry is required (as, +6v, -12v), use this circuit to maintain low C:



Recommended low-cost zener for fast work: (1/4w) Hoffman IN707. 7.1v ± 0.9v Tek# 152-004 Max current ≈ 30 ma.

Q and A Guide For Demos

Q: Why an open-loop gain of only 2500?

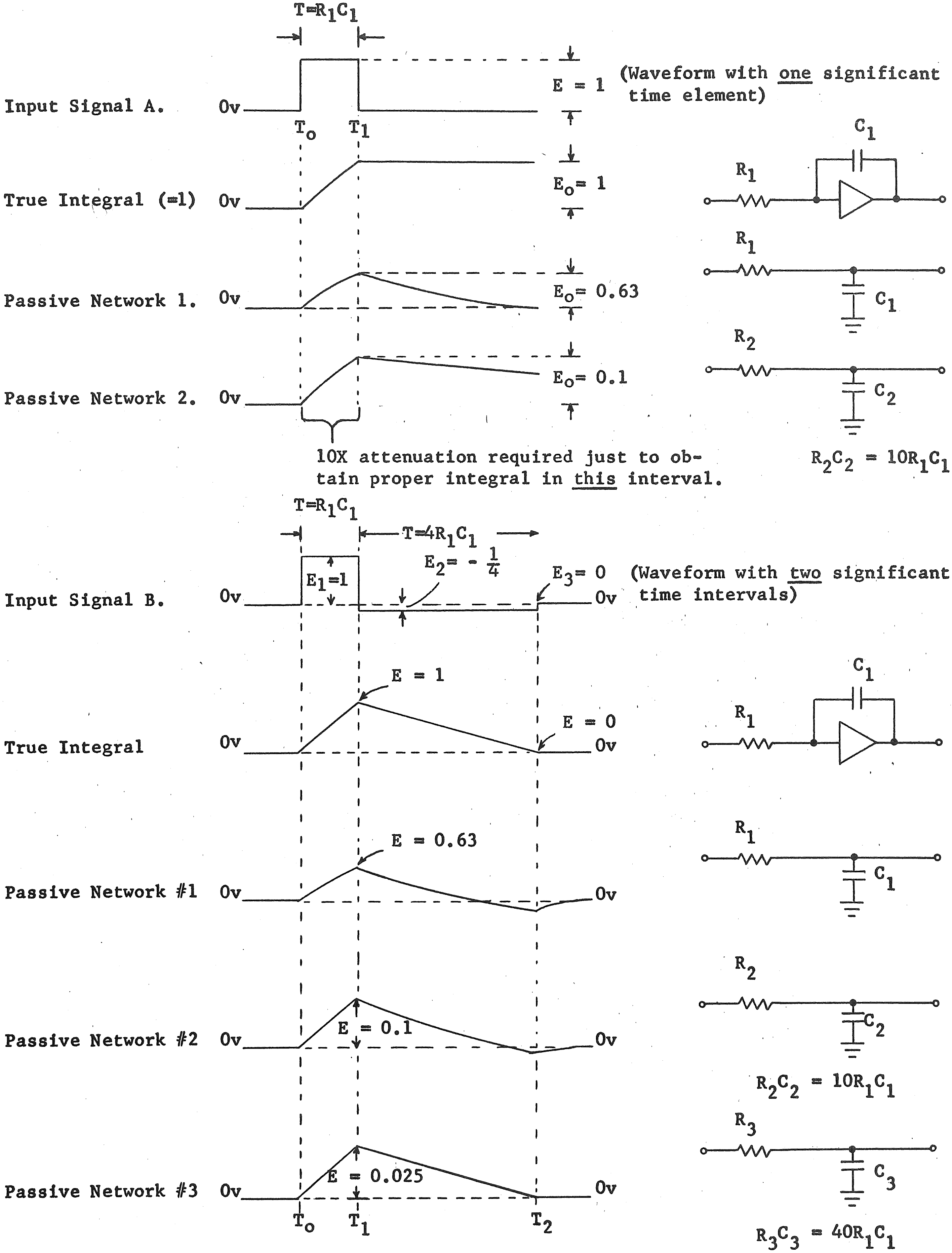
A: This is a leveled-off characteristic, and gives us a nearly constant error-factor out to about 10 KC. Most conventional operational amplifiers with open-loop DC gain of 10^4 to 10^7 are down to unity gain at 100 KC or less, and are usable as 1% devices only below about 5 KC or so for even X1 gain. By contrast, the O-Unit's amplifiers are usable at X1 gain out to about 30 KC for the same amount of error. In this respect, the O-Unit, for signals above 100 cps or so, is the equivalent of an operational amplifier with an open-loop gain of 10^5 or more. And, since the typical rolloff of super-high gain amplifiers is at a 12 db/octave rate, the O-Unit is in a class by itself beyond 100 KC.

Q: You tell me that I can use the O-Unit to integrate and differentiate. What advantage does this have over using passive components?

A: In the first place, a given passive RC time-constant is useful for integrating or differentiating only a very limited range of signals. Accurate integration or differentiation for the purpose of measurements requires first that the user be willing to accept at least a 10-1 reduction in signal level, so that only the "linear" portion of the selected time-constant curve is used, and second, that no significant time element within the waveform to be processed approaches the time-constant used any closer than a factor of ten or so.

These points are illustrated on the following pages.

Fig. 1. PASSIVE vs. ACTIVE INTEGRATION



Attenuation of 40 at T_1 required to obtain accurate measurement over the interval T_o - T_2, where $T_1 - T_2 = 4\ (T_o - T_1)$.

1. INTEGRATION. Let's suppose it's desirable to determine the "area" under a pulse (Fig. 1.) -- a measurement of volt-seconds, or, assuming a constant impedance circuit, of watt-seconds (energy). To use a passive network, the user must first determine the integrating interval, then select a time-constant at least ten times longer than this interval, then pick the components of the time-constant to be compatible with the impedance of his driving source.

The true integral of the waveform shown is illustrated below the "Input Signal A" waveform in Fig. 1. Note that the operational amplifier integral (1) provides a true integral not only for the interval T_o to T_1, but also beyond T_1, and (2) provides an answer which can be made any convenient amplitude simply by selection of R and C. The passive network, on the other hand, even in the ideal case shown, must attenuate by a factor of 10 to obtain accurate integration even over the short interval T_o-T_1. In the case of input signal B, where we are interested not only in the integral at T_1 but also again at T_2, a time constant 40 times (or more) longer than the shorter integrating interval is necessary to obtain an accurate answer. Unfortunately, this attenuates the amplitude at T_1 by a factor of 40, making accurate measurement of small signals next to impossible (the difference between 50 mv and 1.25 mv is readily appreciated by a man with a 50 mv/cm 'scope).

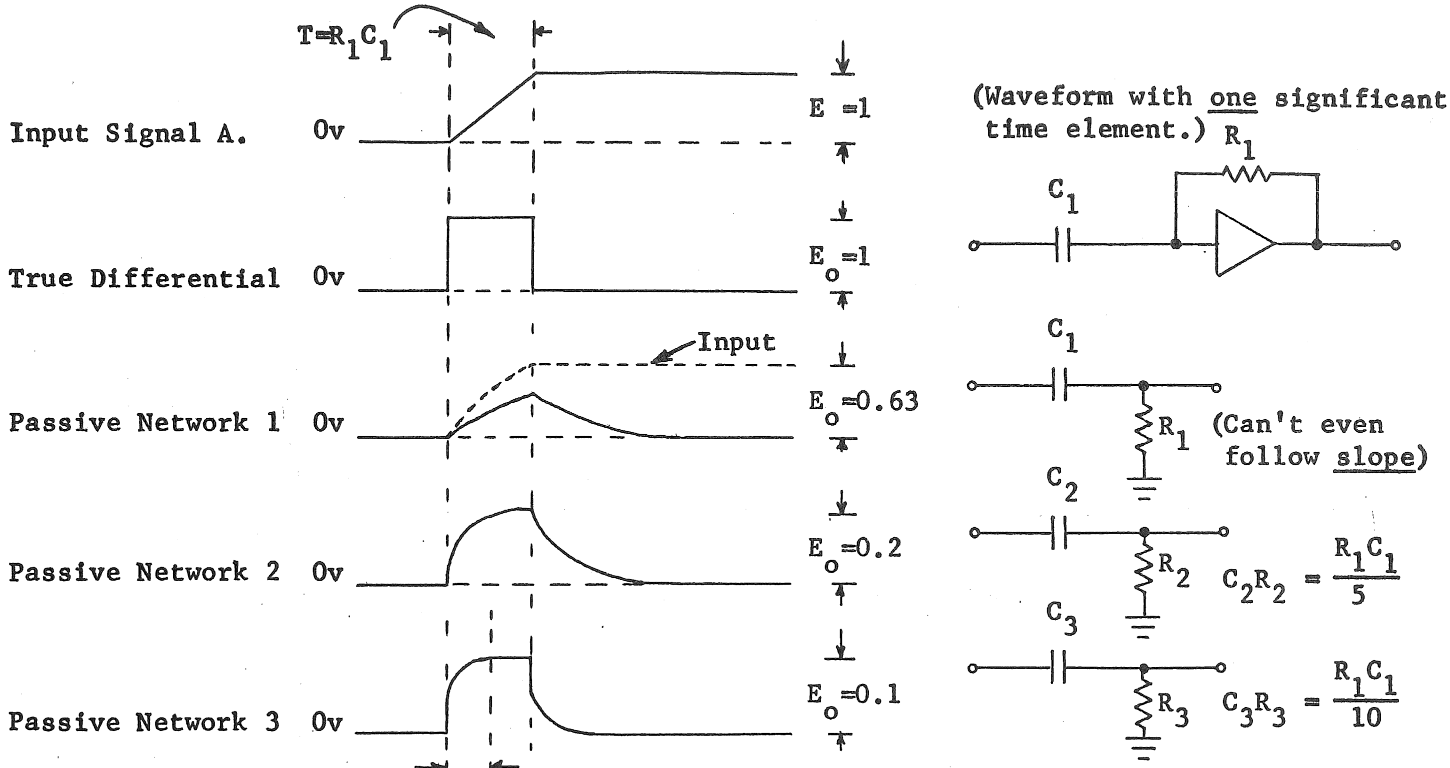
The user of an operational amplifier is not hindered particularly by these considerations. If R_1 and C_1 do not provide him with adequate amplitude for measurement -- as in the case for Input signal B, where he may want to adjust the amplitude of E_2 to a value that will bring the integral at T_2 to a value very close to zero -- he may pick smaller values of C without disturbing either the accuracy of his measurement (until he has run outside the accuracy limits of his operational amplifier), or his input impedance.

In the case of the O-Unit, the 1 second time-constant Z_i = 1 Meg, Z_f = 1μf may be used without impairment of accuracy for integrating intervals of up to 24 seconds (1% error factor due to the operational amplifier gain limitation).

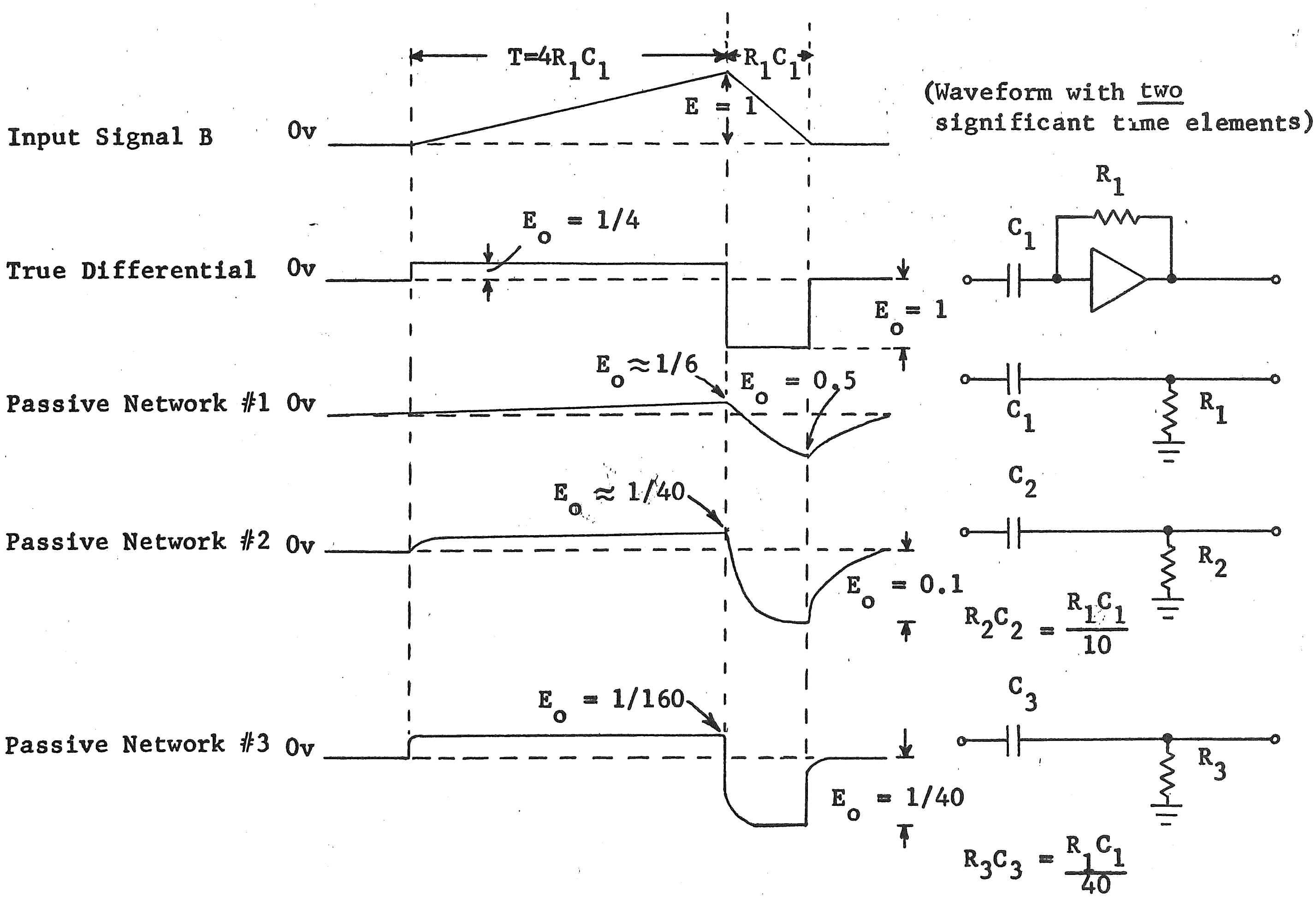
To obtain the same capability from a passive network, a 240-second time-constant would be required -- such as 10 megohms and 24 μf! Assuming a one-volt peak amplitude signal, this value would be attenuated by an absoulte minimum factor of 10 (where a 1 v level was held for 24 seconds) and probably by a factor of 50 or more, assuming a complex signal.

It must be granted, of course, that a passive network does not suffer from DC drift, etc., whereas an operational amplifier may. The O-Unit's 0.3 na grid current with a 1 μf feedback capacitor will produce 7.2 mv drift in 24 seconds.

Fig. 2. PASSIVE vs. ACTIVE DIFFERENTIATION



Note departure from true differential during first 5 RC's (R_3C_3). Rate-of-change of input signal just isn't being indicated during this time.



2. DIFFERENTIATION. The same general principles apply to passive differentiation as to passive integration. Picking an RC time-constant roughly equal to the rise-time of the signal whose rate-of-change is to be measured will result in an inaccurate answer. Only when a time constant 1/10 or less of the duration of the rate-of-change to be measured is used, can the answer approach (after 5 time-constants) the value sought. Attenuation will be at least 10X in the best possible case, which is illustrated (Input Signal A) in Fig. 2.

In attempting to view the linearity of the <u>start</u> of the ramp, a time-constant must be chosen which is less than 1/5 of the time from the start of the ramp to the first point to be examined, if a passive network is to be used. To examine the second 1% of a ramp of duration T, the RC selected must be T/500 or less. The average output level would be 1/500 of the peak ramp level (a 15 v ramp would be differentiated to a pulse 30 mv in amplitude using a passive network that would assure measurement of the rate of change over the last 99%).

Except as limited by the open-loop risetime (see page 5), the operational amplifier provides much superior performance. In the case of a 15 v, 1 msec duration ramp, the O-Unit can provide a differentiated output of 15 v, quite accurate after the second microsecond (compensation of Z_f will be required, however, to prevent ringing due to stray C at the input).

To obtain equivalent performance from a passive network, it would be necessary to follow the network (0.2 or 0.4 μsec RC) with a voltage gain of 250-500 at a bandwidth of at least 750 KC -- a pretty good amplifier.

SUMMARY: So, to summarize, an operational amplifier offers these advantages over passive integration and differentiation networks:

1. The selection of time-constants affects only output amplitude, not measurement accuracy (within the op-amp's gain limits).

2. The output amplitude from the operational amplifier for a given time-constant will be greater. The operational amplifier's output will be at least ten times greater than that of the highest-output passive network usable for comparable performance.

3. The range of operations performable with an operational amplifier and easily manipulated components (ever try to buy a 100 μf 1% mylar capacitor or a 100 meg 1% resistor?) is at least 10 times greater than with the same selection of passive components. Only at very high speeds (nanosecond region) do passive networks come into their own again.

Q: The error-factor bothers me. Do you have any rules of thumb for avoiding large measurement errors?

A: Yes. In the region where the operational amplifier gain is 2500 (DC - 1 KC), you can go for a gain of 24 with less than 1% error contributed by the amplifier (Z_i and Z_f component tolerances and signal source impedances will contribute, of course).

Out to 15 KC, an attempted gain of up to 9 won't cause more than 1% amplifier error. Above this point, compensation should be used for optimum accuracy. The absolute upper limit for less than 1% error contribution by the amplifier is 75 KC, where the maximum closed-loop gain must be held to 1 for 1% accuracy. If 10% error is tolerable, the frequency limits for X10 gain and X1 gain become 150 KC and 750 KC respectively (with compensation).

Q: That's all very nice for sine-wave amplification. Now, how about pulses and complex waveforms, and differentiation and integration?

A: This is the toughest question there is to answer in operational amplifier work. However, we can approach it by taking one element at a time.

1. PULSE WORK: If you'll look at the open loop risetime of the O-Unit, you can consider this to be the variation of open-loop gain with time after the arrival of a signal.

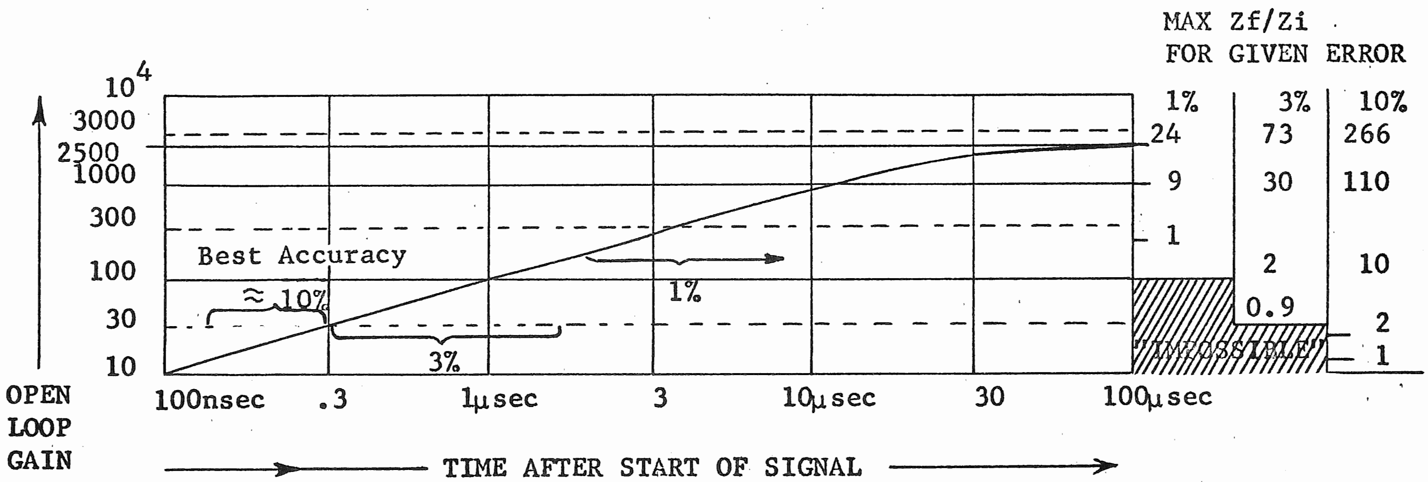


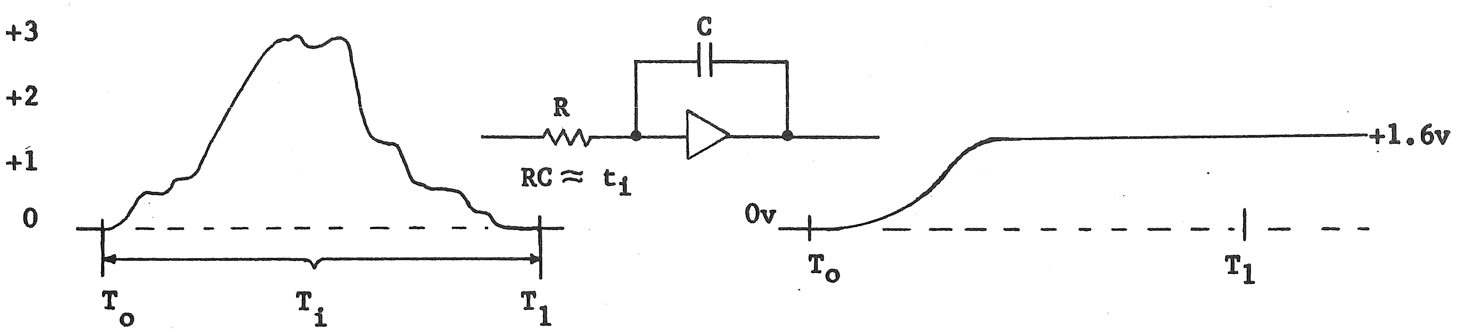
FIG. 3. Open-loop risetime viewed as variation in "A" with time after start of signal.

The open-loop gain (Fig. 3) rises to 10 after about 100 nsec. An open-loop gain of 10 will support a closed-loop gain of 1 with an error of 17%

The open-loop gain rises to 100 after about 1 μsec. This open-loop gain will support a closed-loop gain of 1 with 2% error, or a closed-loop gain of 10 with 10% error.

NOTE: These figures apply only to the "error" using exact values of external components. Because the O-Unit's 10 pf and 100 pf capacitors are adjusted IN CIRCUIT under dynamic conditions, errors will not necessarily be as shown. See page 7.

FIG. 4

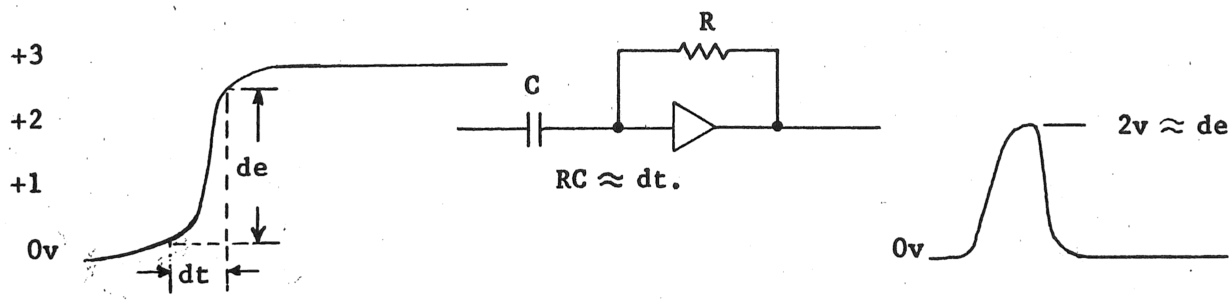


INPUT WAVEFORM

INTEGRAL (Output of Op. Amp.)

Output voltage is about equal to "eyeball average" of input voltage; virtual gain ≈ 1.

A. "VIRTUAL GAIN" in Integration



INPUT WAVEFORMS

DIFFERENTIAL

Output voltage is about equal to voltage (de) of steepest part of waveform; virtual gain ≈ 1.

B. "VIRTUAL GAIN" during differentiation.

After about 15 μsec, the open-loop gain is up to 1000, and amplifier error is negligible for a gain of 1 or 10, but 10% for a gain of 100.

NOTE: Because we allow waveform aberrations on the order of 2-5% (hook, overshoot, undershoot, etc.), however, due to imperfect R's and C's, we cannot claim "1%" measurement accuracies for high-speed work.

2. "GAIN" for differentiation or integration: "gain" for an operational amplifier is proportional to Z_f/Z_i, but during differentiation and integration, one of these impedances is time-dependent, and (except in the case of sinewaves) cannot be assigned a fixed numerical value.

However, we do have a way of assigning a "virtual gain" factor.

For integration, the virtual gain is the ratio between the integrating interval and the RC time-constant selected

$$G_v = \frac{t_i}{RC}.$$

The integrating interval, for this purpose, may be considered to be that span of time during which the integral continues to increase. With a symmetrical waveform, t_i would be considered to be the duration of one half-cycle; with an asymmetrical waveform, however, t_i would be whatever interval over which measurements continue to be taken. The larger the value of RC chosen, the lower the "virtual gain".

For ball-parking, virtual gain during integration may be taken as the ratio between the final value of the output and the "eyeball average" input voltage level (Fig. 4A).

For differentiation, virtual gain is the ratio between the RC time-constant selected and the time (t) during which a given change takes place: $G_v = \frac{RC}{t}$.

If the steepest de/dt takes place during an interval approximately equal to the RC selected, virtual gain may be taken as "1". If the selected RC is three times longer than the interval containing the steepest rate of change (de/dt), the "gain" (for that interval) is three, etc. See Fig. 4B.

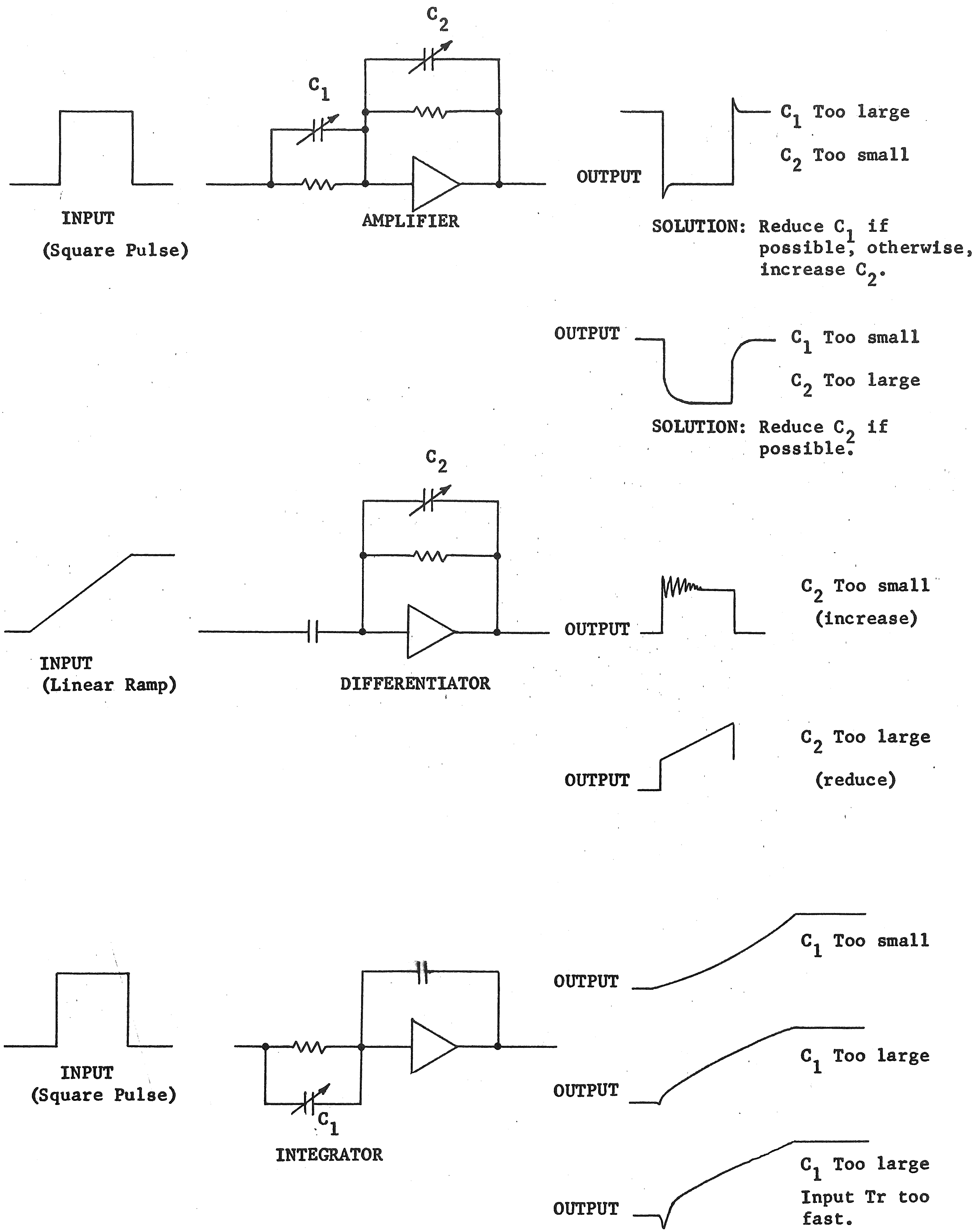
For greatest accuracy, either in differentiation or in integration, virtual gain should be kept low.

A good practice is to select RC to equal the integrating interval or the duration of the steepest portion of a signal to be differentiated.

3. Special limits:

a. Differentiation: Because open-loop gain is risetime-limited (Fig. 3), differentiation using precision external componenets cannot become very accurate until the input rate-of-change is maintained for some time. With a virtual gain of 1, the operational amplifier will require that the input rate of change be maintained for 100 nsec for the output to reach 83% of the proper level, and for 1% accuracy, for 2 μsec or longer. However, the values of internal components normally used for differentiation of fast-rising waveforms (100 pf and 10 pf) are adjusted under dynamic conditions and partially compensate for these errors. See Page 7.

FIG. 5



C2
C1
INPUT
(Square Pulse)
AMPLIFIER
OUTPUT
C1 Too large
C2 Too small
SOLUTION: Reduce C1 if possible, otherwise, increase C2.
OUTPUT
C1 Too small
C2 Too large
SOLUTION: Reduce C2 if possible.
C2
INPUT
(Linear Ramp)
DIFFERENTIATOR
OUTPUT
C2 Too small
(increase)
OUTPUT
C2 Too large
(reduce)
INPUT
(Square Pulse)
C1
INTEGRATOR
OUTPUT
C1 Too small
OUTPUT
C1 Too large
OUTPUT
C1 Too large
Input Tr too fast.

b. Integration: The shortest internally selectable time-constant is 100 nsec; shorter external time-constants should not normally be used, because of current limitations and stray-C problems.

Because of the risetime limitation of the operational amplifier, one would expect that with an integrating interval of 100 nsec for a "virtual gain" of 1, that the A-factor would never get above 10 in this interval and the error in the answer would be 17%. At the end of 1μsec, the virtual gain would be 10 and the A-factor only 100, so the error would still be 10%, and getting worse again, since after a few μsec, the virtual gain continues to rise linearly, while the A-factor starts leveling off toward 2500.

However, "all is not lost", The 10 pf and 100 pf Z_f capacitors are calibrated during integration, so that the error is much less than these figures indicate.

4. Compensation. Input C at the -grid of the O-Unit is approximately 35 pf (a little less if you switch the ±GRID switch to "+" and ground the +grid externally), which will have a definite effect on the accuracy of high-speed operations.

The closed-loop -3 db bandwidth of the operational amplifier varies from about 750 KC to about 1.5 mc for a gain of 1, depending on which pair of equal resistors you select for Z_i and Z_f. With compensation, this bandwidth may be extended to the 10-15 mc region.

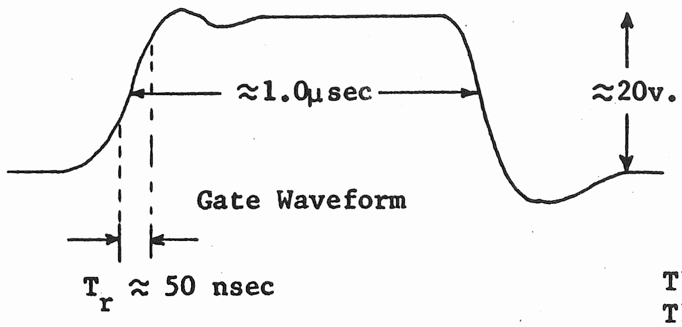
Depending on the application, Z_i, Z_f, or both may require external compensation for accurate results. Fig. 5, shows some of the results of over- and under-compensation.

5. Use of Standard Waveforms for Calibration. The obvious answer to the problem of making accurate measurements using short (external or internal) time constants is adjustment of these values to give correct answers using standard signals in the same ballpark as the signals to be measured, particularly in terms of time and virtual gain.

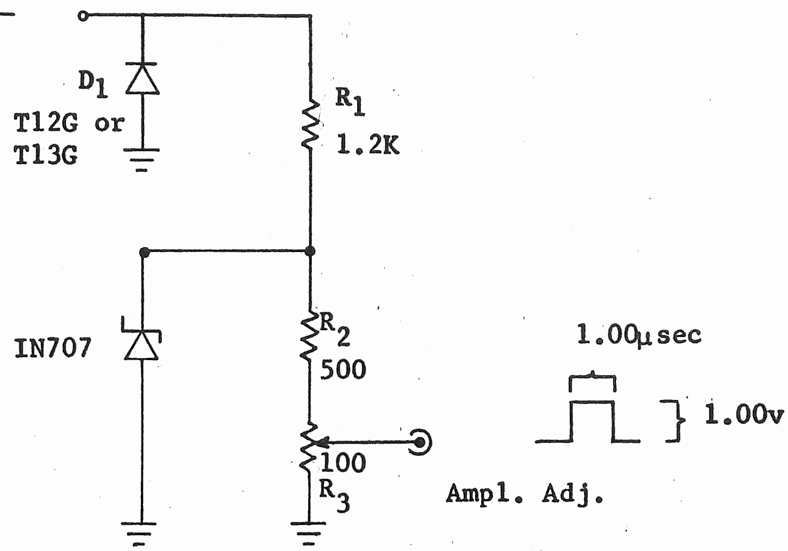
Methods of obtaining standard waveforms from a 535A/545A oscilloscope are shown in Fig. 6. The delaying-sweep scopes are preferable, since the duty-cycle of the waveform is easily controllable, to permit complete stabilization between measurements.

Fig. 6. OBTAINING "STANDARD" WAVEFORMS

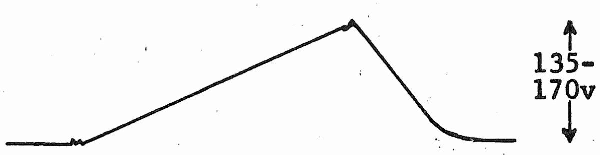
Use Z, 180A to Set Outputs



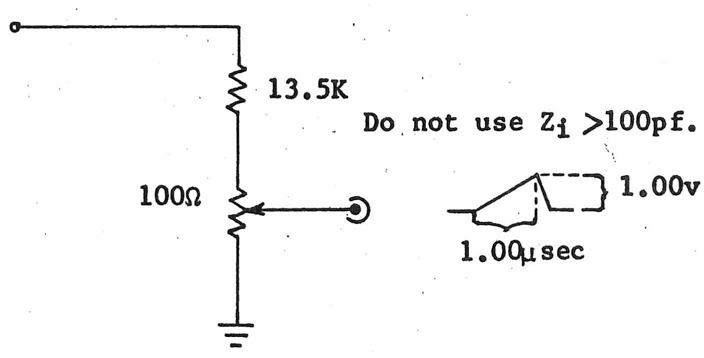
STANDARD PULSE FROM + GATE A. (ADJ swp length for accurate width at 50% points. Use 180A, Z-Unit to set up parameters.



D_1 catches trailing-edge overshoot. IN707 clips tops of pulse & improves risetime. For smaller pulses, make R_3 50Ω & select R_2 (1.2K min) - Then use 50Ω pads. Total drain on + Gate, 20 ma. (If gate > 20v, R_1 can be made larger).



STANDARD RAMP FROM SAWTOOTH A OUTPUT. DRAIN ON CF, ≈10ma.



COMPATIBILITY

PLUG-IN CYCLER

4-23-63

Bill Roberts to Geoff Gass

I have a plug-in cyler. The voltages in this unit will vary considerably depending on how many and what kind of plug-ins are installed. The -150 V will vary from 140 V to over 200 V. The other voltages vary a like amount. I would like to know whether you feel these voltages would be damaging to units such as the M unit and others which have transistorized circuitry.

I am also curious to know whether other plug-in cyclers have the same characteristics or whether I am laboring under a misaprehension.

Thank you for your assistance in this matter.

Geoff Gass to Bill Roberts

4-30-63

Bill Goard tells me that none of these plug-in cyclers have regulated supplies, and the output voltage may vary widely with line and load variations.

This says the cycler should *never* be used with a Z-Unit, which will blow transistors if the power supplies are off by more than a few percent. The tolerance of the Z to -150 v variations extends from about 140 v to 155 v at the plug-in connector, which is the range of the various direct (132-133-532) and decoupled (530/40/50 series) supplies from the scopes and power supplies if they're within tolerance.

I share your concern for the M and other new plug-ins employing semiconductors. I'd be inclined to say that use of the cycler should be strictly limited to those plug-ins which were in existence when the cycler was designed several years ago. Just for laughs, let's say it probably wouldn't hurt a: (53, 53/54) A, B, C, D, E, G, H, K, L, P (!), Q, S, T, or TU-2. It could possibly bottom out the power supplies in the R Unit and cause other problems in other more recent plug-ins. The Z should never be cycled in this box.

However, variations as great as you describe indicate that the underload relays may not be working in your instrument. There's a relay for each hole which throws a nominal load across each supply when pin 2 isn't grounded. The +350 has a fixed load on it (40 K) and isn't affected by the relay.

However, I'd be a little leery of the adequacy of this "go, no-go" kind of regulation for keeping the supplies any closer than "ballpark" for the various possible loadings. So for my money, the use of the cycler is something that can be recommended only with reservations.

I'm sure glad you brought this up, Bill -- design engineers try to make their plug-ins compatible with scopes, and most of 'em probably are not even aware of the cyclers. Even those who know about the cyclers would probably be unwilling to compromise the cost or performance of their plug-ins to assure "no problems" with the cycler.

So all in all, it looks like we have a cycler problem. I'll put this warning into the Speednote. Hope nobody's gotten into trouble.

P.S. Bill Reich points out that the 127 (only $650) makes a dandy cycler. Regulated, too, and has an underload relay for operating with only one hole filled. This would be a best bet for any boxes which look like they might be voltage-sensitive.

ADAPTERS

FEN 9-9-62

The O Unit temporary manual 070-323 jumped the gun a little bit in listing adapter boards 013-048 and terminal shields 013-049 as standard accessories for the O Unit.

These accessories were not promised to customers in the tentative spec sheets and will not be available for shipments of early production instruments. These items will be added as standard accessories as soon as they are available (tentatively by the end of April). Customers receiving early O Unit shipments may be offered the accessories as a customer accommodation, at the discretion of the field engineer. These accessories, of course are not necessary for satisfactory O Unit operation.

O UNIT GRID-CURRENT CHECKER AVAILABLE FROM FIELD MAINTENANCE SUPPORT

FEN 7-27-62

The test jig used to check grid current in the Type O Plug-in operational amplifiers is available *(internal use only)* from Field Maintenance Support. This is similar to the jig described in the Type O factory cal procedure, but with minor modifications for ease of assembly and use.

There is no Tek Number.

Order by IOC from Bill Goard, Field Maintenance Support.

ADAPTERS

Production of O Unit adapters is tentatively scheduled for mid 1963. Circuit designs were developed by Hiro Moriyasu and are being adapted for production by Bob Johnson in Lang Hedrick's accessories design group. The adapters will be made on etched circuit boards, with cast aluminum housings and photo-etched aluminum saddle-type covers.

Logarithmic amplifier: 4-decade (±.1 to 100 v) similar to the design shown on page 3-8 (circuit 13 of the O unit manual but with ac-dc coupling). At the phase A meeting, it was estimated that production might begin about March 1963. Price will probably be about $60. Don't sell 'em until they're announced in the PAL.

Low frequency function generator (circuit 24 in O Unit manual).

Gated integrator (similar to circuit 22).

Low frequency sine wave generator (circuit 25).

Amplifier compensator for X1, X10 (circuit 1, modified).

Diode leakage test jig (circuit 19, modified).

FEN 11-30-62

AC COUPLER — not for sale

A limited number of ac couplers for tweaking the 100 pf and 10 pf ranges of the O unit is available from field maintenance support. These are constructed in surplus attenuator cases and are for in-house use only -- not for sale to customers. The unit isn't absolutely necessary for calibration -- an unterminated 107 (see manual, page 7-8) is adequate for customer use.

ADAPTER HARDWARE KIT — not for sale

Because of the difficulty in obtaining finished O Unit adapters for demo use until the designs are a little further along in the phase system, we've put together an O demo kit, consisting of the following:

2	--	blank boards punched as for the 013-048 accessory
12	134-014	banana plugs with 6-32 studs (134-013 banana probe tips may also be used but they have longer shoulders)
10	210-202	solder lugs, SE-6
2	210-204	solder lugs, DE-6
12	210-407	nuts, 6-32 x 1/4"
1	--	set of suggested "operating principle" circuits, hints, kinks, etc.

You can use the hardware from a regular 013-048 adapter accessory to complete the adapter. About 150 kits are available from field training.

O UNIT ACCESSORIES SURVEY

As a result of the O Unit Accessory Adapter survey March 1, Engineering has shelved two of the projected adapters, and plans to continue with only two of the projects--the diode and transistor leakage test adapter and the amplifier compensator. The other two adapters--although estimated sales indicated possible product feasibility, the field reaction was against introducing these as products, indicating that their availability would offer little net benefit.

There is no estimate yet on availability of the two projects still in the mill--"late 63" is a good guess.

Here are the survey results:

Questionnaires sent out--99
Returns received--45 (including 4 "no bid").
Not all returns answered all questions, so totals don't add.

*E.g., O Units bought in the expectation that such an adapter would be forthcoming, or customers refusing to buy O's until adapter is available.
**Figures in parenthesis indicate probable total (domestic plus overseas) sales.

When a customer has a particular need for one of the shelved adapters and is unable to construct one of his own from the data in the O Unit Manual, negotiate with Ron Goard for a possible Custom Special.

	Predicted O Sales?*	Sell More O's?	Probable Adapters/Yr	Rank of Importance	Yes/No
Diode-Transistor Leakage Adapter	10+	Yes 7 A few 16 No 13	103+ (200)**	X1 - 16 2 - 3 3 - 6 4 - 4 5 - 3	Yes 22 No 20
Compensator Adapter	18+	Yes 6 A few 13 No 17	124+ (250)**	1 - 4 X2 - 20 3 - 4 4 - 2 5 - 1	Yes 15 No 17
Function Generator	2+	Yes 6 A few 8 No 23	85+ (175)**	1 - 6 2 - 1 X3 - 16 4 - 3 5 - 4	Yes 8 No 24
Sine-Wave Generator	2+	Yes 6 A few 7 No 24	79+ (150)**	1 - 1 2 - 4 3 - 4 X4 - 16 5 - 5	Yes 8 No 24

O UNIT GATING ADAPTER IN STOCK

FEN 6-14-63

Gating adapter 013-068 for the O Unit is available (limited quantity) from Customer Service stock. Price has been set at $75.

The manual is 070-395. Additional discussion, operating information, etc. was included in the "IEEE Instruments" book distributed in March. Corrected and updated information, and a step-by-step procedure for real-time processing of sampled data and calculation of scaling factors, is scheduled for a late summer O Unit IRB supplement. Until this is printed, manuscript copies may be obtained from Field Info if needed (specify which).

THIRD O UNIT ADAPTER IN STOCK

FEN 1-24-64

The Amplifier Compensating Adapter -- third of the four planned O-Unit accessory adapters -- is now in stock. This unit provides continuously adjustable one-knob compensation for amplifier gains from X1 to X100, and may also be used for compensating the gain <1 configuration of the O-Unit's operational amplifiers. Use of the compensator with the amplifier set for gains of X1 to X100 provides bandwidths close to the open-loop gain-bandwidth curve (See O-Unit Manual, page 2-4). With overpeaked or rolled-off settings, it may also be useful for some bandwidth-tailoring applications, though it was not especially designed for this.

The Compensating Adapter is Tek No. 013-081. The manual is 070-416, T-slotted for insertion into the O-Unit Manual. Price of the 013-081 is estimated at $35., but has not yet been set by the pricing committee. For a firm quote, contact Customer Service.

Other adapters available are:

013-067 -- Logarithmic Amplifier Adapter. The manual for this unit is 070-386; additional technical data is included in the O-Unit IRB.

013-068 -- Gating Adapter. Manual for this unit is 070-395; additional information relating to real-time processing of sampled data using the gating adapter will be included with next O-Unit IRB Supplement.

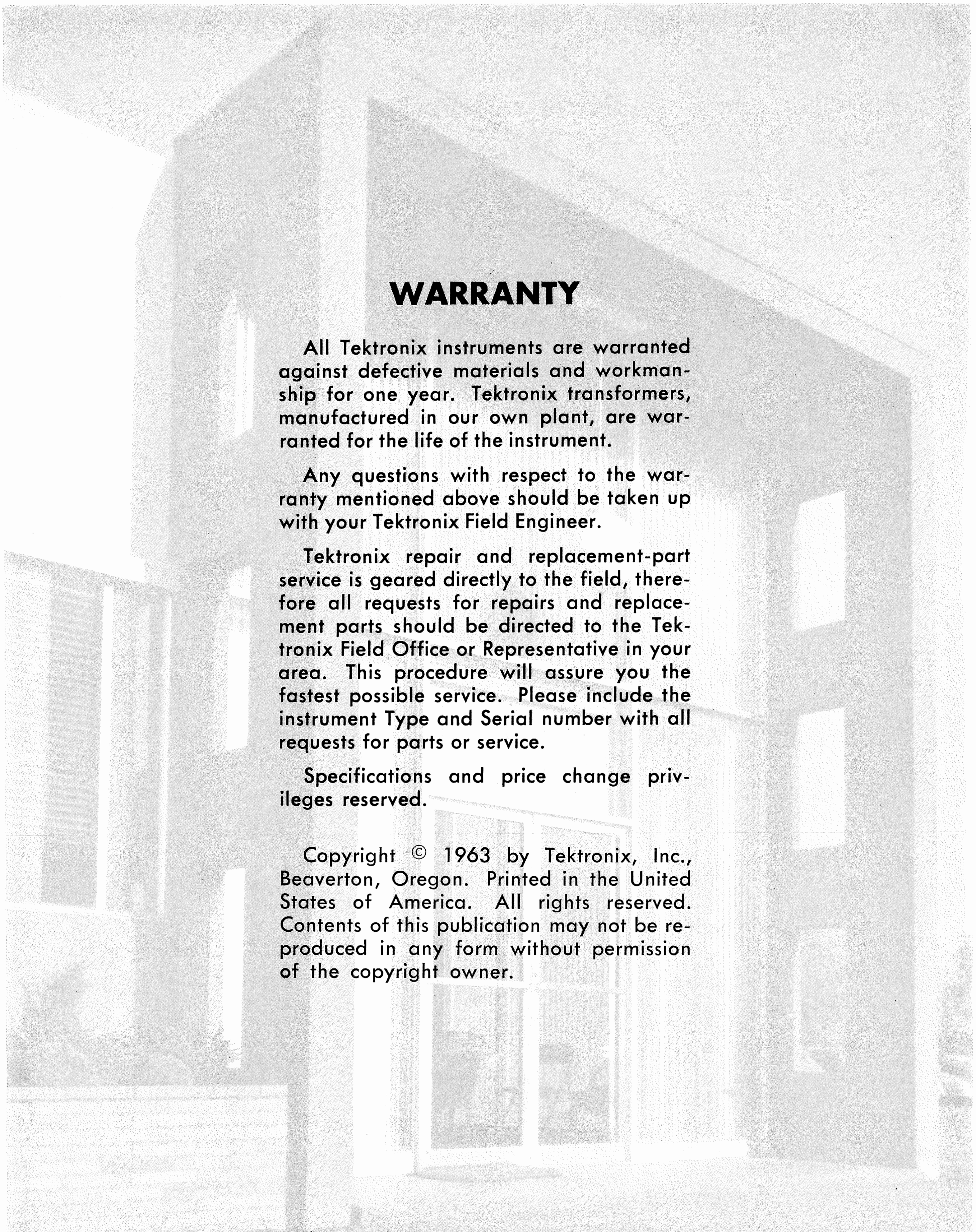
The proposed diode-transistor leakage test adapter is still in Engineering and will not be available for several months.

Gating Adapter
for
Type O Plug-In

Tektronix, Inc.
S.W. Millikan Way • P. O. Box 500 • Beaverton, Oregon • Phone MI 4-0161 • Cables: Tektronix

Tektronix International A.G.
Terrassenweg 1A • Zug, Switzerland • PH. 042-49192 • Cable: Tekintag, Zug Switzerland • Telex 53.574

070-395 563



WARRANTY

All Tektronix instruments are warranted against defective materials and workmanship for one year. Tektronix transformers, manufactured in our own plant, are warranted for the life of the instrument.

Any questions with respect to the warranty mentioned above should be taken up with your Tektronix Field Engineer.

Tektronix repair and replacement-part service is geared directly to the field, therefore all requests for repairs and replacement parts should be directed to the Tektronix Field Office or Representative in your area. This procedure will assure you the fastest possible service. Please include the instrument Type and Serial number with all requests for parts or service.

Specifications and price change privileges reserved.

Gating Adapter for the Type O Plug-In Unit

General

When the Gating Adapter is plugged into the 'B' Operational Amplifier of a Type O Plug-In Unit, an electronic switch is formed that is used to gate "on" and gate "off" the 'A' Operational Amplifier. A repetitive signal applied to amplifier 'A' will then be amplified, integrated, or differentiated only during the "on" time. There will be no signal at the 'A' output during the "off" time.

The Gating Adapter is of great value during integration operations. Except with signals which have a net integral of zero, true integration is impossible without gating since the integral of a repetitive signal would accumulate to a voltage value beyond the range of the Type O Unit. The Gating Adapter, however, permits integration for a selected time and then resets the integrator to zero. This cycle repeats with each oscilloscope sweep.

The 'A' amplifier is gated "on" by applying repetitive positive-going pulses of about 20 volts amplitude to the Gating Adapter input. The amplifier remains "on" for the duration of the pulse, and is "off" during the time between pulses.

In order for the output of the 'A' amplifier and the oscilloscope display to be coherent, both the sweep and the gating pulse must bear a fixed time relationship to the 'A' input signal. One way to accomplish this is to externally trigger the oscilloscope sweeps with the 'A' amplifier input signal and to use the plus gate output of the oscilloscope as the gating signal. Oscilloscopes such as the Tektronix Types 535A, 545A, 555, and 585, which have a delaying sweep feature, will provide greater flexibility in the gating operation. The set-up and compensation procedure at the end of this manual contains several extra steps to show how the delayed sweep gate may be used to turn on the 'A' amplifier at a selected time after the beginning of the oscilloscope sweep.

Theory of Operation

The Zener diode and capacitor at the Gating Adapter input have the effect of a battery. Repetitive positive-going pulses produce a voltage drop across the diode and maintain a charge on the capacitor. Because of this charge, the input pulses are negatively offset after passing through the diode-capacitor network.

The 'B' Operational Amplifier is connected as a unity gain, inverting amplifier. Hence, pulses of opposite polarity, but equal amplitude are applied across the diode bridge. For the duration of a pulse, all four diodes are back-biased and exhibit a resistance of about 200×10^9 ohms in parallel with the 'A' amplifier Z_f. Since this high parallel resistance does not effectively alter the Z_f value (in most practical applications), the 'A' amplifier is turned on.

When a gating pulse ends, the stored charge in the input capacitor reverses the polarity of the voltage across the diode bridge. All four diodes conduct and appear as a low resistance in parallel with the 'A' amplifier Z_f. With Z_f reduced to nearly zero ohms, there is essentially no signal at the 'A' output.

Limiting Factors

The 'A' amplifier should be turned off no more than 90% of the gating signal period or 2 seconds, whichever is the shorter time, so the charge on the Gating Adapter input capacitor is maintained. If the charge falls below a certain value, the forward bias of the diode bridge is removed and the 'A' amplifier will pass the signal during "off" time. Remember that the duty-factor of the oscilloscope plus gate output signal is a function of the sweep triggering rate as well as the ratio of the 'A' sweep duration to the 'B' sweep duration.

During the time the amplifier is gated "on", the 'A' out-'A' input signal can produce sufficient current through the 'A' Z_i component to significantly alter the diode bridge forward current. If this occurs, a small amount of signal will be passed to the 'A' output. The amount of voltage that can be applied to the 'A' input without causing signal feed-through is directly proportional to the Z_i impedance.

During the time the amplifier is gate "on", the 'A' output voltage swing must be limited to about ± 20 volts. This is to prevent variation in the Z_f value due to overriding of the diode bridge back-bias. The exact output voltage limit is directly proportional to the gating pulse amplitude.

During the transition from off to on, 'A' amplifier is unstable for a few microseconds. This generally is of little consequence except when very short gating periods are used. The effects of the transitional instability are minimized by proper adjustment of the two compensating capacitors in the Gating Adapter.

The gate period for integration is limited to about 20 times the selected integrator time constant. The limit can be raised considerably by turning on the 'A' amplifier INTEGRATOR LF REJECT, but this partially defeats the purpose of gating.

The 200 x 10^9 ohms resistance of the back-biased diode bridge parallels the 'A' Z_f during 'A' "on" time. Hence, this determines the minimum useable integrator capacitance because the bridge resistance tends to discharge the integrator capacitor. This discharge will not be noticeable unless the time constant of the bridge resistance and the integrator capacitor is a significant percentage of the integrator network time consant.

Set-Up and Compensation

The following steps give the necessary information to establish the 'A' Operational Amplifier in a Type O Plug-In unit as a gated, unity-gain amplifier. These steps should be performed regardless of the type of operation to be performed by the 'A' amplifier. Also included is the procedure for adjusting the compensating capacitors in the Gating Adapter. This procedure is intended for use with an oscilloscope having a delaying sweep feature. Other oscilloscopes may be used by substituting the appropriate control settings into the procedure for the oscilloscope used.

1. Set the oscilloscope controls and switches as follows:

'A' STABILITY	clockwise
'A' TRIGGERING MODE	AC
'A' TIME/CM	.5 mSEC
VARIABLE	CALIBRATED
HORIZONTAL DISPLAY	'B' INTENSIFIED BY 'A'
5X MAG.	OFF
'B' STABILITY	clockwise
'B' TRIGGERING MODE	AC
'B' TIME/CM	1 mSEC
'B' LENGTH	clockwise
DELAY-TIME MULTIPLIER	2.00
AMPLITUDE CALIBRATOR	1 VOLTS

2. Plug the Gating Adapter into the 'B' Operational Amplifier. The cam on the adapter housing should trip the ±GRID SEL to (—).

3. Plug the Gating Adapter leads into the 'A' Operational Amplifier —GRID and OUTPUT jacks.

4. Set the Type O Plug-In Unit controls and switches as follows:

'A' ±GRID SEL	(—)
'A' and 'B' INTEGRATOR LF REJECT	OFF
'A' Z_i and Z_f	1 MEG.
'B' Z_i and Z_f	EXT.
VOLTS/CM	.5
VARIABLE	CALIBRATED
VERTICAL DISPLAY	EXT. INPUT + DC
VERTICAL POSITION	to center trace

5. After the plug-in unit has warmed-up, check for proper adjustment of the DC BAL. and GAIN ADJ. controls, using a free running display.

6. Set the plug-in unit VERTICAL DISPLAY switch to OUTPUT —B.

7. Adjust B OUTPUT DC LEVEL as described in the Type O unit manual.

8. Set the plug-in unit VERTICAL DISPLAY to OUTPUT A.

9. Adjust A OUTPUT DC LEVEL.

NOTE

The OUTPUT DC LEVEL adjustments are very important to proper operation and should be re-checked often.

10. Connect the oscilloscope +GATE 'A' output to the Gating Adapter input.

11. Set 'A' TIME/CM to 2μSEC and 'B' TIME/CM to 5 μSEC.

12. Set the Type O Unit VOLTS/CM to 2.

13. With the cover-plate in place, adjust the Gating Adapter compensation capacitors:

    a. Adjust C_1 for minimum amplitude of the pulse at the left end of the intensified trace zone.

    b. Adjust C_2 to make the portion of the intensified trace zone following the pulse appear as straight as possible.

    c. Balance the two adjustments for the smoothest transition from the pulse to the remainder of the intensified trace zone.

NOTE

Step 13 completes the compensation and basic set-up procedure. The remaining steps are provided to illustrate a few of the basic operating procedures which can be used to obtain maximum measurement versatility.

14. Set 'A' TIME/CM to .5 mSEC and 'B' TIME/CM to 1 mSEC.

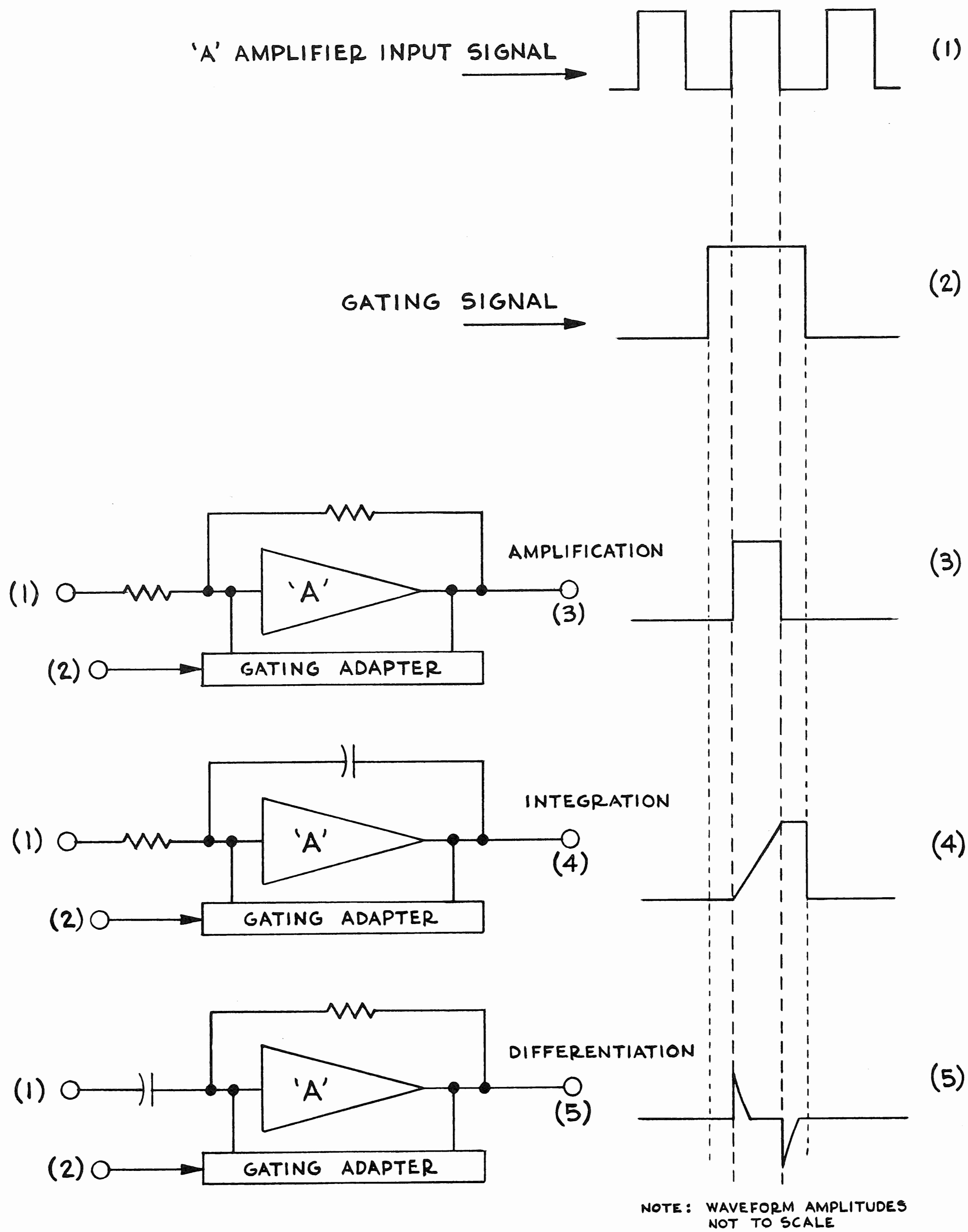
15. Connect the oscilloscope CAL. OUT to Time Base 'B' TRIGGER INPUT and to the 'A' Operational Amplifier INPUT.

16. Set the Time Base 'B' triggering controls for an externally triggered display.

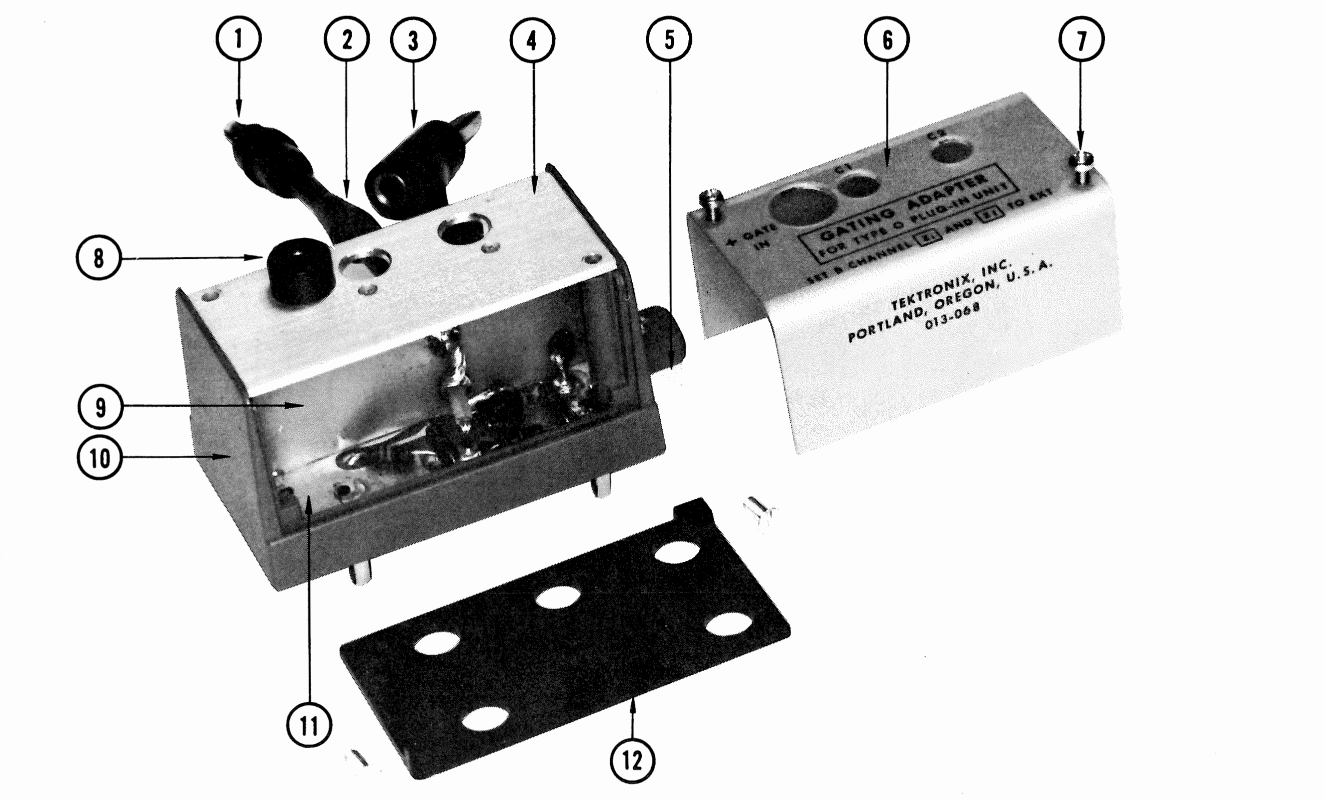
The trace should display a signal only in the intensified zone. The 'A' TIME/CM switch and the VARIABLE control settings can be changed to include the desired number of pulses in the display. The DELAY TIME MULTIPLIER control can be set so the display begins between pulses or during a pulse. The 'B' TIME/CM switch, the 'B' LENGTH control, and the 'B' sweep triggering rate determine the time seperation between the groups of pulses.

The oscilloscope controls mentioned in the foregoing paragraph are used in much the same manner when the system is operated as a gated integrator or a gated differentiator.

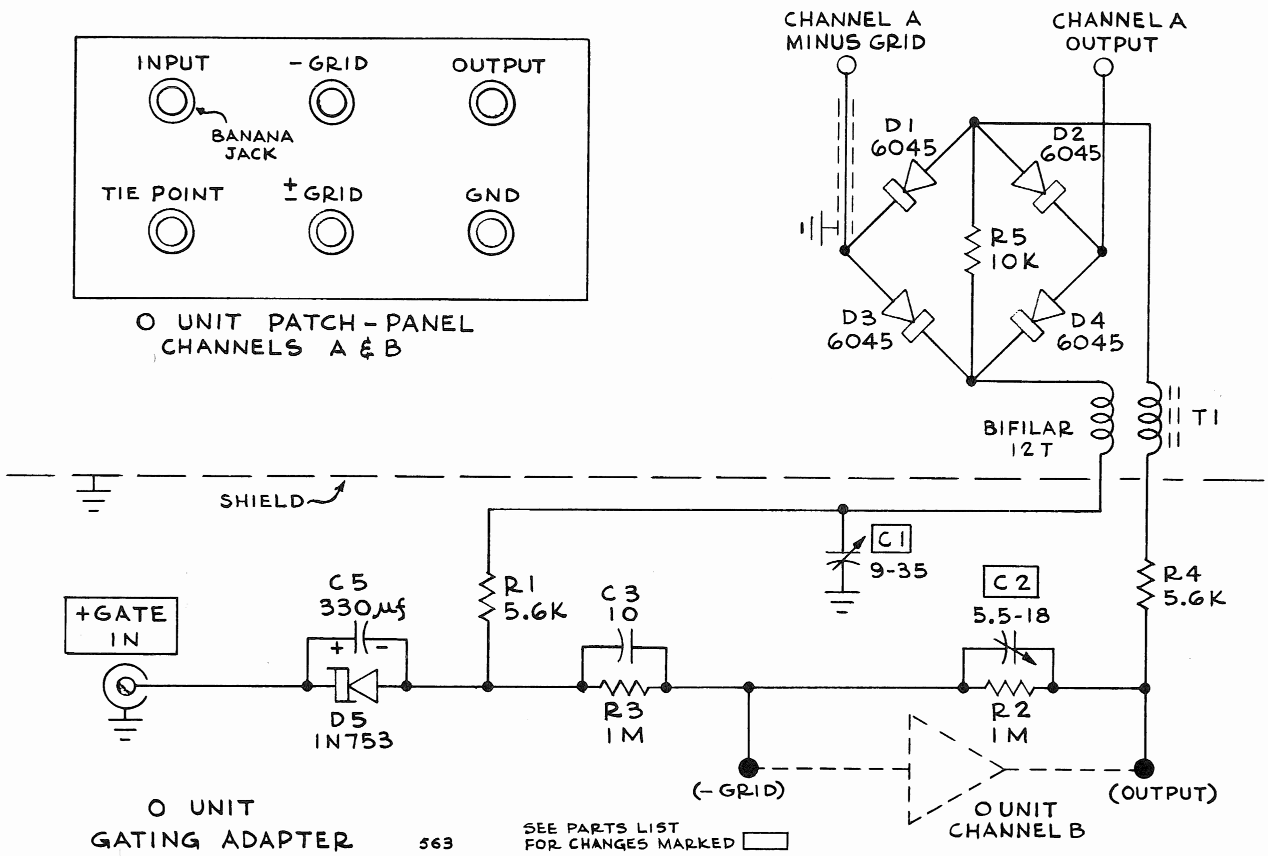
The Type O Unit, 'A' amplifier and Preamplifier controls and switches are used as described in the O unit manual, with one exception; the 'A' amplifier INTEGRATOR LF REJECT switch is normally left in the OFF position.



GATING ADAPTER 563



REF. NO.	PART NO.	SERIAL NO. EFF.	SERIAL NO. DISC.	QTY.	DESCRIPTION
1	175-294			1	CABLE ASSY., MINUS GRID (Consisting of)
	134-024			1	PLUG, Banana, Male
	200-489			1	COVER, connector
	175-026			1	CABLE, Coax. 75 Ω 5"
	358-117			1	BUSHING, outer sleeve
2	200-491			2	COVER, cable relief
3	175-295			1	CABLE ASSY., OUTPUT
4	387-799			1	PLATE, subpanel
5	401-022			1	CAM, switch actuating Mounting Hardware: (not included)
	213-055			1	SCREW, thread cutting, 2-56 x $3/16$ PHS
6	200-490			1	COVER
7	211-079			4	SCREW, 2-56 x $3/16$ Pan Head steel
8	136-138			1	SOCKET, Banana Jack Assy.
	210-223			1	LUG, solder
	210-895			1	WASHER, insulating
	210-583			1	NUT, hex, $5/16$ brass, $1/4$-32
9	337-579			1	SHIELD, gating adapter
10	204-163			1	BODY, casting
11	388-550			1	BOARD, Circuit
	131-285			2	Connector
12	392-146			1	BOARD insulating Mounting Hardware: (not included)
	211-095			2	SCREW, 2-56 x $5/16$ FHS 80°
	134-070			1	PLUG



O UNIT
GATING ADAPTER

REF. NO.	PART NO.	SERIAL NO. EFF.	SERIAL NO. DISC.	QTY.	DESCRIPTION
	120-319			1	T1 TRANSFORMER, 12T TD103
	152-034			1	D5 DIODE, zener 1N753 6.2 V 400MV 10%
	152-045			4	D1, 2, 3, 4 DIODE silicon 6045
	281-061			1	C2 CAPACITOR, 5.5-18 pf cer. var.
	281-063			1	CI CAPACITOR, 9-35 pf cer. var.
	281-504			1	C3 CAPACITOR, 10 pf cer. 500 V 10%
	290-138			1	C5 CAPACITOR, 330 μf EMT 6 V 20%
	315-562			2	R1, R4 RESISTOR, 5.6 K 1/4 W 5%
	318-004			2	R2, R3 RESISTOR, 1 meg 1/8 W prec. 1%
	315-103			1	R5 RESISTOR, 10 K 1/4 W 5%

GATING ADAPTER

Description: The Type O gating adapter makes use of one of the O-Unit's operational amplifiers to invert the +Gate of the oscilloscope for symmetrical (push-pull) drive to a diode bridge. The diode bridge gates the other operational amplifier to permit gated amplification or integration of any desired signal within the operational amplifier's capabilities. The diode gate is "open" during the time the +Gate of the oscilloscope is positive, and closed when the scope gate is down.

The primary purpose of the adapter is to facilitate integration, especially in areas where the O-Unit's LF Reject circuits contribute excessive error or inconvenience in measurements.

Gating also provides a means of eliminating unwanted or confusing portions of a waveform to be amplified, differentiated or integrated, so the operation can be performed only on the desired portion of the signal. Gated operation can thus be used to avoid severe overloads of the oscilloscope preamplifier and enhance operational accuracy.

The leads on the gating adapter are arranged for plugging the adapter itself into Operational Amplifier B, to gate operations in A via the two external leads, which connect to the -Grid and Output jacks on A.

The gate of the delayed sweep in a delaying-sweep scope or the delayed gate of the Type 532 may be used to provide gating of only a selected portion of the waveform which triggers the sweep. In other scopes, the gate is open during the entire sweep. An external gating signal may also be used. Ideal levels are +20 v for "on" and -3 v for "off".

External triggering of the scope is required for gated operation in most cases, except where synchroscope techniques can be used.

Limitations: The maximum output signal which can be handled is about ±20 v, depending on the amplitude of the scope gate output. If the output waveform exceeds this value during the gated interval, the gate will come into partial conduction, distorting the output.

The maximum input signal which can be handled is 20 v, provided that the gating duty cycle is 10% or more and the R_i resistor is 1 M. For R_i = 200 K, max input is 20 v with 10% duty cycle; for R_i = 100 K, max input is 10 v. With R_i at 10 K, duty cycle must be kept at 20% or more, max interval between gates 500 msec and input to 2 v or less for 5% feed-through during "off time" (signal current pulling open the diode gate).

The signal "pulling open" the diode gate will not affect integration accuracy if the integrator output is measured from the start of the intensified portion to the end.

The gating adapter as furnished is primarily for the gated amplification or integration of repetitive signals only, where the time during which the gate is closed is no more than 9 times the "on" time, or 2 sec, whichever is shorter. You will notice that in working from a cold start, it takes several sweeps to stabilize. Modification of the scope or the adapter is required for low rep-rate or single-shot work, to provide a negative gate voltage during "off" time. Modifications are discussed below.

There is a transient during the turn-on of the gate which can be minimized, but not completely eliminated, by compensating adjustments. The transient is usually negligible during integration, but is definitely noticeable during amplification. The transient -- its exact shape depends on diode switching characteristics, etc. -- typically consists of a short duration spike of about 1 v amplitude, followed by a smaller swing of opposite polarity and about 1 μsec duration, followed by a 1/2 v pulse of the first polarity and 30 μsec decay time.

Figs. 1 and 2 show the effect of the switching transient. Fig. 1 shows its effect on gated amplification ($R_i = R_f$ = 1 M, 5 μsec/cm, 500 mv/cm). Fig. 2, at 100 μsec/cm and 50 mv/cm shows its possible effect on integration. In the case shown, with 1 M and 100 pf (100 μsec time constant) the indicated error -- unchanged over 5 time-constants -- is 70 mv, or 7 μv sec.

In amplifying the 20 v gate, the B operational amplifier shifts the decoupled -150 v supply in the plugin by about 400 mv. This causes a level shift in A of about 10-20 mv for the duration of the gate. This level shift, unless corrected, will be integrated during gated integration, producing a fixed error of 10-20 microvolt seconds per millisecond of gate duration. In most cases this error can be significantly reduced by careful setting of the A and B dc level adjustments. Where not, the no-signal integral should be separately measured and subtracted from the measurement.

To measure the no-signal integral, be sure to ground the input end of Z_i so that all effects of zero-shift, grid-current, etc., will be included. Grid current will contribute an error up to 20 mv/msec in the output, using a 10 pf value for Z_f. Normally, the 10 pf capacitor is never used for integrating intervals in excess of 100 μsec or so.

Differential leakage through the diode gate with 20 v across it is typically on the order of 0.1 nanoampere for an effective resistance of 200 X 10^9 ohms. For this reason, it is a good idea -- where you have the choice -- to use a large value integrating capacitor to keep the leakage error to a minimum. Again, this sort of error is negligible in most cases, and is of less effect than the typical 0.3 to 0.5 nanoampere grid current.

Modifications: For single-shot work, the zener diode-capacitor network used to obtain a negative level between gates, to keep the diode bridge conducting, will not work. Two solutions are available:

Scope mod: The +Gate of the scope may be made to go negative between sweeps by adding a resistor between the cathode of the +Gate output CF and the -150 v supply. The value of this resistor should be about 60 times the value of the existing cathode resistor, for a level of about -2 v between sweeps. A lower level will be needed when Z_i = 10 K.

Hookup mod: A small penlite cell or mercury battery connected in series between the +Gate output and the gating adapter input (+ end of the battery toward the +Gate output connector) will provide a -1.5 v level between sweeps.

Adapter mod: Where it is necessary that sudden changes in sweep rep rate and duration have minimal effect on the system, one of the above modifications plus a minor mod to the adapter will be required. The adapter mod consists of installing a jumper across the zener diode. Wire from the connector to the far lead of the zener. This prevents the DC levels around the gate from shifting as the duty cycle changes.

Gated Differentiation: The switching transients discussed above might indicate at first glance that gated differentiation is impossible. Not so, fortunately. The transients are not differentiated, being impressed across Z_f only. For this reason, the "glitches", etc., shown in Fig. 1 appear about the same during differentiation as in amplification and may be accounted for in the final answer by comparison with the "no signal" waveform. For minimum interference, use a low value R_f, or (with 1 M R_f) open the gate about 30 μsec before the start of the signal to be differentiated. Fig. 3 shows gated differentiation of the trigger feed-through on the +Gate of a 535A (the gate was attenuated to 4 v so as not to overload the gating circuit). The high rate of dv/dt at the start of the gate -- about 700 x 10^6 v/sec -- would severely overload the oscilloscope preamplifier at these sensitivities, masking the smaller feed-through signal, were it not for the gating function.

Gated Slideback Amplifier: The gating adapter may also be used for gated amplification with slideback, the input signal and offset voltages being summed at the -grid and gated together, thus preventing overload of the oscilloscope preamp.

FIG. 1. Switching transient. (Worst case: $Z_i = Z_f = 1$ M, uncompensated) 5 μsec/cm, 0.5 v/cm. Transient at end of gate does not affect measurements, of course. The aberrations are materially reduced if smaller values are chosen for R_f and R_i or if R_f and R_i are compensated.

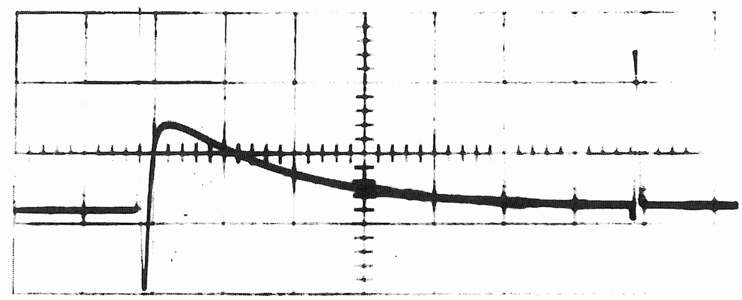


FIG. 2. Integral of switching transient. 100 μsec/cm, 0.05 v/cm, $Z_i = 1$ M, $Z_f =$ 100 pf. 70 mv error represents 7 μv-sec. Note that error is relatively constant after first 30 μsec. The error value will change with increased integrating interval because of grid-current, leakage, etc. Integrating beyond about 20 RC's not usually recommended.

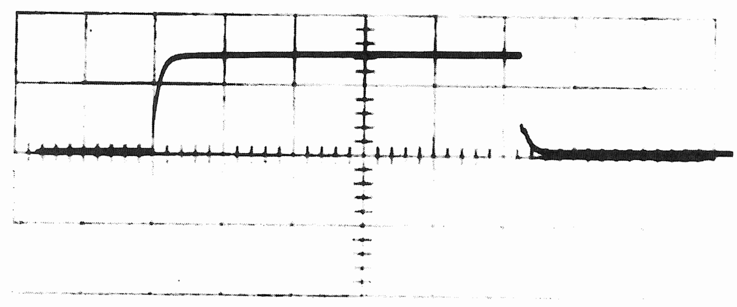
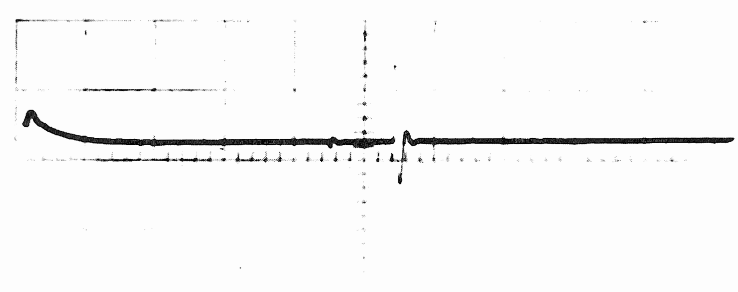


FIG. 3. Gated differentiation of trigger feed-through on 535A +Gate. 20 μsec/cm, 0.2 v/cm. $Z_i = 100$ pf, $Z_f = 200$ K. Switching transient artifact is seen at left. Gating eliminates severe scope overload (60 v at 0.2 v/cm) which would have been caused by leading edge of signal waveform.



Installation:

Using 535A or 545A Oscilloscope, O-Unit and Gating Adapter.

1. Plug gating adapter into Channel B banana jacks. Leads from gating adapter to Channel A minus grid and Channel A output.

2. Set Channel B Z_i and Z_f to external.

3. Set vertical display to B minus, adjust B dc level; set vertical to A minus, adjust A dc level.

4. Set A Channel Z_i and Z_f to 1 megohm.
Set 535A or 545A sweep to 0.2 msec/cm.
Set B sweep to 0.5 msec/cm.
Horizontal display B intensified by A.
A plus gate out to gating adapter.

5. Set volts/cm to 0.5.

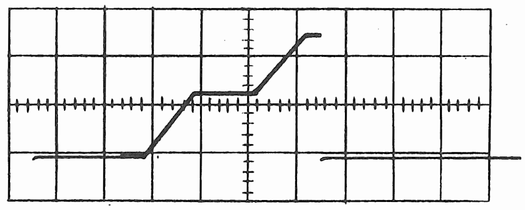
6. Adjust A Channel dc output level so that the intensified and unintensified portions of the sweeps are level.

7. Adjust C1 and C2 of gating adapter for minimum overshoot and roll-off. C1 affects the switching time of the diode gate. C2 compensates the external Z_f resistor for Channel B.

8. Connect a 1 volt calibrator signal to A input, and also to external trigger of B sweep.
A output should be 1 volt in the intensified portion of trace.
No signal on unintensified portion of trace.

9. Set A Channel Z_i to 1 megohm.
Set A Channel Z_f to 0.001 µf, LF Reject Switch "off".
Set calibrator to 20 volts.
Set volts/cm to 10 volts.
Output waveform will be as shown with no signal or unintensified portion of trace.

10. Reduce calibrator level to 1 v and volts/cm to 0.5 v/cm.
Check to see that the level portions of the waveform are quite level. If not, recheck A and B dc levels, and adjust (repeat steps 6-8-9) as necessary to obtain minimum tilt to this portion of the waveform.

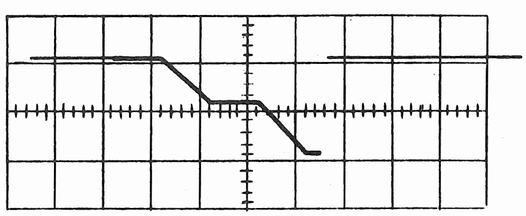


20 volt calibrator signal

≈ 40% gate cycle

Z_i 1 meg, Z_f 0.001

11. Connect signal from 105 square wave generator, 1 kc, 20 volts to A input of O Unit.
Signal will be negative going with no signal on unintensified portion.



≈ 20 volt 105 signal

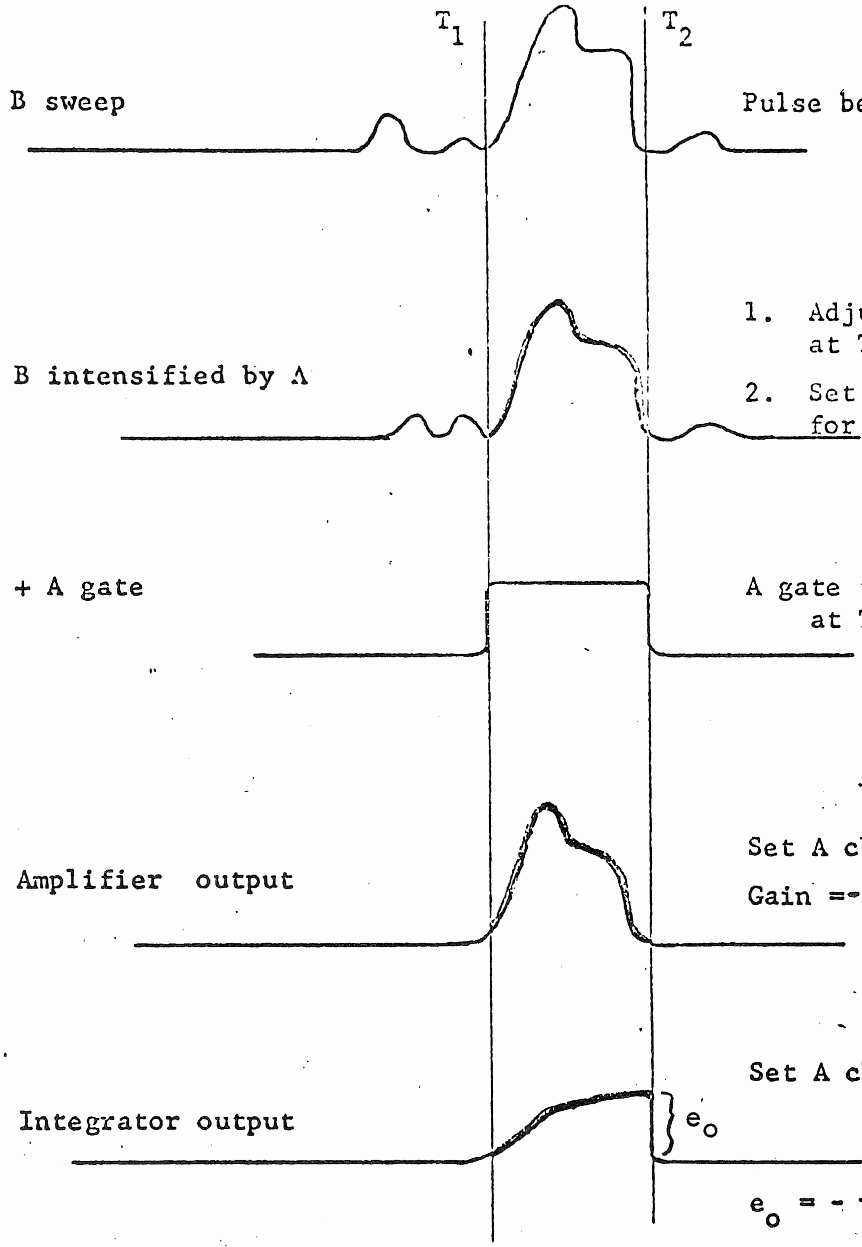
≈ 40% gate cycle

Z_i 1 meg, Z_f 0.001

12. Switch A sweep to 50 μsec (10% gate cycle).
There should be no signal appearing on the unintensified portion of the trace.
A decrease to 5% gating cycle will cause the diode gate to open during the unintensified portion.

Sample Gated Amplifier and Gated Integrator Operation

To amplify or integrate a specific pulse:



Pulse between T_1 and T_2

1. Adjust Delay to start at T_1
2. Set sweep A time/cm for T_2

A gate to start and stop at T_1 and T_2

Set A channel Z_i and Z_f to R [1]

Gain $= -Z_f/Z_i$

Set A channel Z_i to R, Z_f to C.

$$e_o = -\frac{1}{RC}\int e_i\,dt$$

$$\text{Input volt-seconds} = \int e_i\,dt = -R_i C_f\, e_o$$

Bob M. Johnson
Accessories Design

Bill Lukens
Manufacturing Staff Engineering

Future Product: A completely external gating adapter which uses a separate (AC) power supply has been designed by Chuck Edgar and may be made available later through Chuck Nolan. It would cost quite a bit more ($50 to $100 more) than the passive-component adapter, but would offer these advantages:

1. Adapter ties up only one operational amplifier, not two.

2. Two gates could be used on one O-Unit, powered from one ($60-$75) power supply, permitting gated double integration or gated amplification plus gated integration.

3. Gating is cleaner and does not shake up DC supplies in plug-ins.

4. Power supply also usable for other purposes.

Progress of the design presently awaits decisions on the power supply (just how versatile to make it).

LOGARITHMIC AMPLIFIER ADAPTER

FOR

TYPE O OPERATIONAL AMPLIFIER

The following information is based on data available during Week 49. Additional information on the Logarithmic Amplifier Adapter appears in the Type O Plug-In instruction manual.

The Logarithmic Amplifier Adapter converts the Type O Unit to a logarithmic amplifier. The adapter is built on a circuit board fitted with banana plugs to mate with the jacks on the front panel of the Type O Unit.

There is one potential problem that may be encountered with some of the earlier Type O Units when no signal is applied to the Logarithmic Adapter. Gain of the combination is maximum with little or no applied signal. Hence, some noise and/or drift may be apparent under no-signal conditions. Later Type O Units are less noisy and this is not such a problem.

Characteristics

Input Impedance: Approximately 10 kilohms.

Input Coupling: AC or DC. AC coupling time-constant ≈ 2.5 msec.

Operating Range: ±0.1 volt to ±100 volts.

Maximum Input Signal: ±100 volts, peak, ac or dc coupled.

Amplification Characteristic: The Logarithmic Amplifier Adapter in the Type O Unit closely approximates a logarithmic amplification response for input signals of ±0.1 to ±100 volts. The system is not logarithmic to signals less than about ±0.08 volt.

Deflection Characteristic: The following amounts of deflection will be obtained for the corresponding input amplitudes, (with VOLTS/CM of Type O set at .1).

INPUT SIGNAL	DEFLECTION (±0.5 mm)
±0.1 volt	1 cm
±1.0 volt	2 cm
±10.0 volts	3 cm
±100.0 volts	4 cm

Because the response is logarithmic, the graticule may be calibrated directly in db (beyond ±1 cm). The calibration, after the preamp gain is set for 1 cm per voltage decade, is 20 db per cm (2 db per mm), measured from ±1 cm, which may be taken as the "0 db" points. Since the range +100 mv to -100 mv passes through $-\infty$db, there can be no direct db calibration close to 0 volts.

Response Time (the time required to change from 100 mv to 100% of a step function): Varies with amplitude and direction of change. The adapter was originally designed for an operating range of 0.1 to 10 volts, therefore the spec we have is measured between these limits. Response time when going from 0.1 to 10 volts is approximately 0.2 μsecond; time from 10 to 0.1 volt is approximately 0.3 μsecond.

Transient Response: May be optimized for 0-10 v or 0-100 v operation.

Bandwidth: .The primary use of the logarithmic amplifier adapter will be in allowing the observation of pulse and transient waveforms differing in amplitude by up to 1000-1, on the same trace.

Because the logarithmic amplifier driven by a sinusoidal input waveform does not produce a sinusoidal output (p-p and RMS -3 db points will be quite different) and because both output amplitude and effective bandwidth are affected by both the amplitude and the DC level of the incoming signal, no clear statement of bandwidth can be made.

A few general statements regarding apparent bandwidth based on peak-to-peak output deflection may be of help in establishing the proper use-domain of the instrument, however.

1. The apparent -3 db bandwidth may be defined for our purposes as that upper frequency at which the output from the logarithmic amplifier, driven from a constant-amplitude sine-wave generator, drops to the same level as would be caused by a 30% reduction in input signal amplitude.

   When the adapter and CRT display are properly balanced, calibrated and compensated for a display of 1 cm per input voltage decade (1 cm = 100 mv; 4 cm = 100 v) the apparent -3 db bandwidth is evidenced by a decrease of 1.5 mm in peak deflection (3 mm peak-to-peak) for a signal symmetrical about zero volts.

   A decrease in peak deflection of 0.5 cm (1.0 cm p-p) represents an amplitude change of -10 db for a sinusoidal signal symmetrical with respect to 0 volts.

   In this case, typical apparent bandwidth figures are:

Signal Amplitude	Apparent Bandwidth -3 db	Apparent Bandwidth -10 db
100 mv peak (200 mv p-p)	400 KC	750 KC
1 v peak (2 v p-p)	700 KC	1.15 MC
2.5 v peak (5 v p-p)	700 KC	1.15 MC
5 v peak (10 v p-p)	1.1 MC	2.0 MC

2. Special Case: For a sine-wave with one peak clamped to exactly 0 volts, a reduction in peak-to-peak deflection of 1.5 mm represents -3 db amplitude change; a reduction of 0.5 cm represents -10 db.

   In this case, the apparent bandwidth for a signal of a given amplitude would be roughly equivalent to the bandwidth for a signal of twice that amplitude in the table above (i.e., a signal of 100 mv p-p with one peak clamped at 0 volts would provide an apparent bandwidth equivalent to that produced by 200 mv p-p symmetrical about 0).

Low Frequency Response: In the AC-coupled mode, for signals of over 500 mv peak amplitude where the effective input R is 10 K, the input time-constant for a low signal source impedance is approximately 2.5 msec, corresponding to an apparent -3 db point of about 65 cps. The time constant becomes somewhat longer for smaller signals, as the effective input R approaches 20 K.

Operation

To use the Type 0 Operational Amplifier as a logarithmic amplifier with the Logarithmic Amplifier Adapter, set the Type 0 Unit controls as follows:

Channel B Controls	
Z_i SELECTOR	EXT.
Z_f SELECTOR	EXT.
± GRID SEL	(-)
INTEGRATOR LF REJECT	OFF
Channel A Controls	Any position
Preamp Controls	
VERTICAL DISPLAY	B, + or - (dependent on signal polarity)
VOLTS/CM	.1 (for 100 volts/4 cm display)
VARIABLE VOLTS/CM	Set according to the adapter calibration procedure.
Oscilloscope Controls	Set as necessary for desired Type 0 Unit operation.

Plug the Logarithmic Amplifier Adapter into the front-panel jacks of the Channel B Operational Amplifier. Calibrate the adapter as described on page 3-9 of the Type 0 Unit instruction manual. Connect the input signal to the input connector of the adapter and select AC or DC coupling of the signal. Signals of varying amplitudes will be displayed logarithmically on the oscilloscope CRT.

Measurements and Interpretation

The convenience of having the ranges from +1 cm to +4 cm and from -1 cm to -4 cm calibrated directly in db is partly offset by the fact that the range between +100 mv and -100 mv passes through $-\infty$ db. The anomaly arises from the fact that there is no logarithmic expression for a negative number -- but since we work with both positive and negative signal voltages, the amplifier is made non-logarithmic in the 0 volt area, and treats a negative-polarity signal as $-\log |E_{in}|$, since $\log\text{-}E_{in}$ is impossible.

For this reason, when observing signals which pass through 0 volts, each peak should be treated separately, and interpretation of "peak-to-peak" measurements should be undertaken only with caution. Normally, +100 mv and -100 mv will be taken as independent "0 db" points, and separate + and - measurements referenced to these points.

DC Level Set: Because only a few mv drift in the operational amplifier output can cause a large measurement error, it is important to set this level carefully before any critical measurement. For minimum drift -- particularly where the adpater is being driven from another source which may drift slightly -- the AC coupling position is provided. O-Unit modification 6115 -- scheduled for early 1963 production -- will provide a considerable reduction in internal drift in the operational amplifiers, substituting a new circuit using type ZZ1000 gas reference tubes and transistors for the temperature-sensitive 100 v 1N3044B Zener Diodes now in use.

Theory of Operation

The Logarithmic Amplifier Adapter gets its logarithmic response by using the non-linear forward impedance characteristics of diodes. The diodes connect into the input and feedback loop of the operational amplifier. The diodes are in matched pairs and connected back-to-back to allow for signals of either polarity.

Calibration

The amplifier is essentially logarithmic from ±0.1 v to ±100 v. It is not logarithmic with signals less than about 0.08 v above or below ground.

When calibrated, the amplifier has a gain of 3 for very low level signals. At a signal level just under 0.1 v, the gain is approximately 1.4 and continues to drop logarithmically with increased signal.

Calibration is accomplished by using the oscilloscope AMPLITUDE CALIBRATOR and a Type 105 Square-Wave Generator with 93 Ω cable and termination as signal sources.

1. Let the O Unit warm up for ten to fifteen minutes. Plug the logarithmic amplifier into the B operational amplifier, AC/DC Switch at DC.

2. Free-run the sweep at about .5 msec/cm.

3. Place the VERTICAL DISPLAY switch at B -.

4. Place the Z_i and Z_f controls in the EXT. position.

5. Establish the OUTPUT DC LEVEL to match the ZERO CHECK trace position.

6. Place the O Unit VOLTS/CM switch to .1.

7. Connect a 1 v calibrator signal to the input of the logarithmic amplifier.

8. Adjust the O Unit VARIABLE control for 2 cm of signal. Re-check the OUTPUT DC LEVEL and ZERO CHECK to be sure the negative part of the display coincides with zero volts. (A little drift will alter the calibration for low level signals).

9. Raise the AMPLITUDE CALIBRATOR to 100 v. Adjust the 5 k ±100 v CAL. pot for a 4 cm display. NOTE: Most calibrator CF's in good condition will produce about 94 v into 10 K in the 100 v position (3.97 cm deflection). If in doubt, check 100 v Cal Out into 10 K, using a Z-Unit. The zero-volt part of the waveform also may not return completely to zero in the time the calibrator multivibrator output is negative. Do not adjust the vertical POSITION control for the zero volt point while adjusting the ±100 v CAL pot.

10. Re-do Steps 7, 8 and 9 several times until the amplifier functions properly. Proper operation means an 0.1 v calibratior signal will produce 1 cm deflection ±0.5 mm; a 1 volt calibrator signal will produce 2 cm ±0.5 mm; the 10 volt calibrator signal will produce 2.95 cm ±0.5 mm (only 8.5 v actual cal output into 10 K); and 100 v will produce 4 cm ±0.5 mm -- provided the Cal Out CF is in good condition.

    Failure of the calibrator to reach its negative peak will be particularly noticeable in AC-coupled operation with 100 v calibrator input, but this is normal -- due to the high impedance of the Cal Out CF to -going signals. In the AC coupled mode, the effect will appear as a sizeable apparent error in p-p deflection.

11. Substitute a 25 KC square wave from a Type 105. Use 93 Ω cable and 93 Ω termination and adjust the HF Comp capacitor for minimum spike at the leading edge of the square wave at about a 10 volt signal level. This adjustment will change with maximum signal amplitude. If you intend using the amplifier up to 100 volts, set the HF Comp capacitor at maximum capacitance.

    With 10 v input from the Type 105, response time should be about 0.2 μsec going from 100 mv to 10 v, and about 0.3 μsec from 10 v to 100 mv.

12. The amplifier is now calibrated. Always re-check the OUTPUT DC LEVEL against the ZERO CHECK trace position just prior to any measurement.

LOG ADAPTER FOR TEKTRONIX TYPE O PLUG-IN UNIT

The Log Adapter is a logarithmic feedback network that converts the A or B operational amplifier in a Tektronix Type O Plug-In Unit from a linear to an essentially logarithmic amplifier. It is fitted with banana plugs so it can be plugged directly into the jacks provided on the front panel of the Type O Unit. The circuit configuration and operation of the device is similar to that given in the Applications Section (Application 13) of the Type O Unit instruction manual.

CHARACTERISTICS

Input Resistance

Approximately 10 kilohms

Operating Range

±0.1 volt to ±100 volts

Maximum Input Signal

±100 volts peak, ac- or dc-coupled

Input Voltage vs Oscilloscope Vertical Deflection

An essentially logarithmic relationship for signals between ±0.1 volt and ±100 volts. System is not logarithmic for signals between about —0.05 and +0.05 volt.

Any convenient vertical deflection factor per input voltage decade can be used. (One cm/decade is used in the following example of input and display relationship.)

Dc-Coupled Input Signal (dc plus peak ac)	Deflection (from zero reference)
+0.1 volt	+1 cm, ±0.5 mm
—0.1 volt	—1 cm, ±0.5 mm
+1.0 volt	+2 cm, ±0.5 mm
—1.0 volt	—2 cm, ±0.5 mm
+10.0 volts	+3 cm, ±1.0 mm
—10.0 volts	—3 cm, ±1.0 mm
+100.0 volts	+4 cm, ±1.0 mm
—100.0 volts	—4 cm, ±1.0 mm

In the foregoing example, a 1-centimeter change in display peak deflection equals approximately a 20-db change in input peak voltage (2 db/mm).

NOTE

Amplitude must be measured with respect to the zero-volt dc input trace position. When one peak of a dc-coupled signal is at zero volts, the peak-to-peak deflection amplitude can be used to calculate the peak-to-peak signal voltage. However, when neither display peak is at the zero-volt trace position, the peak voltages must be calculated separately. (This will usually be the case when ac coupling is used.) The sum or difference of the peak voltages will then indicate the peak-to-peak voltage of the ac portion of an input signal. See Fig. 1.

Response Time

Depends upon the direction and the amount of change. For a 10-volt step input (either + or —), the time required for the amplifier output to rise from the 0.1-volt level to the 10-volt level is typically 0.2 μsecond, and to fall from the 10-volt level to the 0.1-volt level it is typically 0.3 μsecond. Response times for this and other peak voltages depend upon proper adjustment of the HF ADJ control for the particular peak signal voltage involved.

Low-Frequency Response (Ac-coupled input)

3-db down at about 70 cps for signals over 500 mv peak amplitude with a signal source impedance of 50 ohms. (Low-frequency response is measured in a manner similar to that described in "Apparent Bandwidth".)

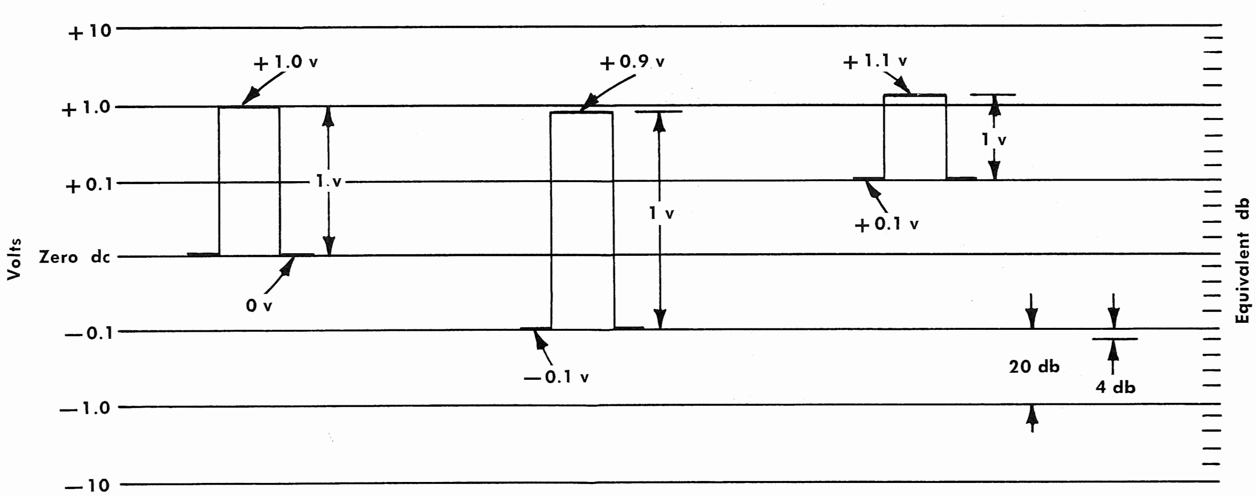


Fig. 1. Display amplitudes resulting from equal peak-to-peak signal voltages, but different dc levels.

Input to Output Voltage Relationship

$e_o = k_1 + k_2 \log_{10} e_i$

where:

e_o = operational amplifier output voltage

e_i = Log Adapter input voltage

$k_1 = e_o$ when $e_i = 1$ volt (typically, $k_1 = 0.3$ to 0.4 volt)

$k_2 = \dfrac{e_o - k_1}{2}$ when $e_i = 100$ volts (typically, $k_2 =$ 0.15 to 0.2). k_1 and k_2 must be determined experimentally for a particular Log Adapter.

Apparent Bandwidth

Sinewave bandwidth measurements generally are not directly applicable to devices having a nonlinear output since the rms and peak bandwidths differ. However, the apparent bandwidth data given here does provide a means of verifying proper performance of the Log Adapter/Type O Unit system.

Apparent bandwidth is measured using a constant-amplitude sinewave generator with low output impedance and signal that is symmetrical about zero volts dc. Under these conditions, the apparent -3 db bandwidth is defined as that approximate upper frequency at which the output from the logarithmic amplifier drops to the same peak voltage as would be obtained with a 30% reduction in input signal amplitude at lower frequencies.

Typical apparent bandwidth figures are:

Signal Amplitude	Apparent Bandwidth —3 db	Apparent Bandwidth —10 db
100 mv peak (200 mv p-p)	400 kc	750 kc
1.0 v peak (2.0 v p-p)	700 kc	1.1 mc
2.5 v peak (5.0 v p-p)	700 kc	1.1 mc
5.0 v peak (10.0 v p-p)	1.1 mc	2.0 mc

SET-UP AND CALIBRATION

The set-up and calibration of the Log Adapter are described in the Applications Section, Application 13, of the Type O Unit instruction manual.

For accurate calibration of the Log Adapter, the applied voltage must be accurately known. The Amplitude Calibrator in your Tektronix Oscilloscope will provide a reasonably accurate voltage at the 1.0-volt and 100-volt levels. However, when the Amplitude Calibrator is connected directly to the Log Adapter, the calibrator will deliver less than the indicated voltage in the 10-volt and adjacent switch positions. This occurs because the Amplitude Calibrator impedance varies with the switch position. To avoid this, isolate the Amplitude Calibrator and Log Adapter impedances by using the second operational amplifier as a unity gain, high input-impedance amplifier (maximum input ±50 volts; see Application 5 in the Type O manual).

As an alternate calibration voltage source, use a regulated variable dc power source, monitored with an accurate meter.

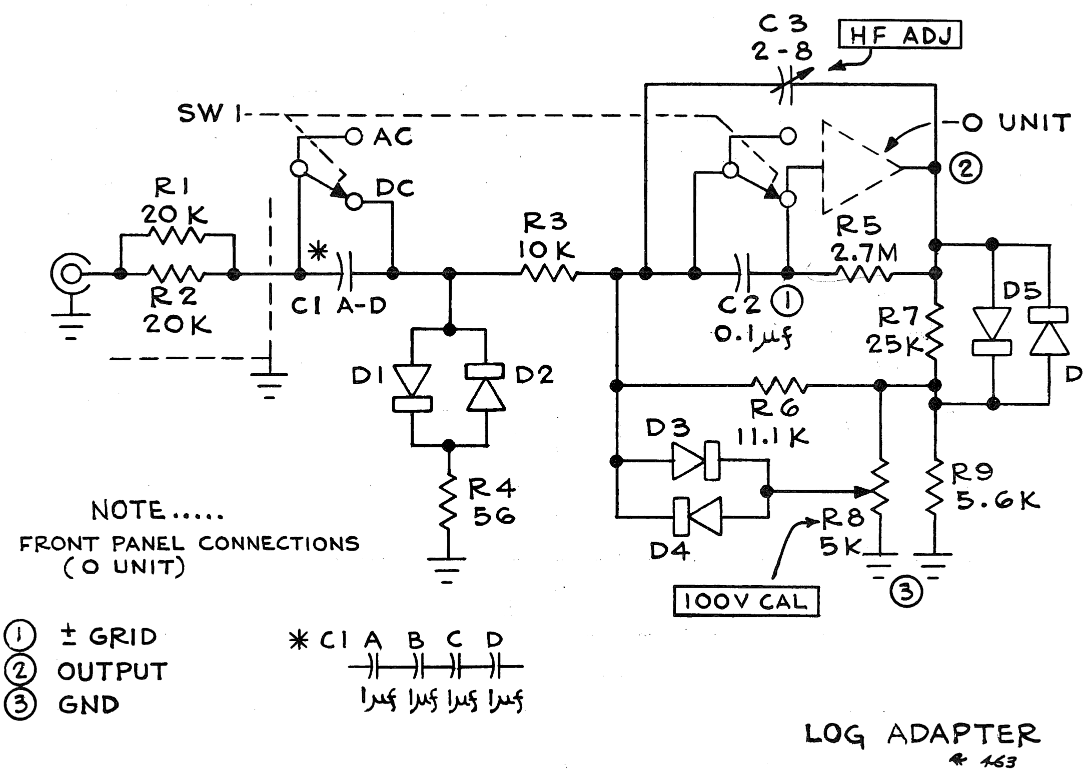
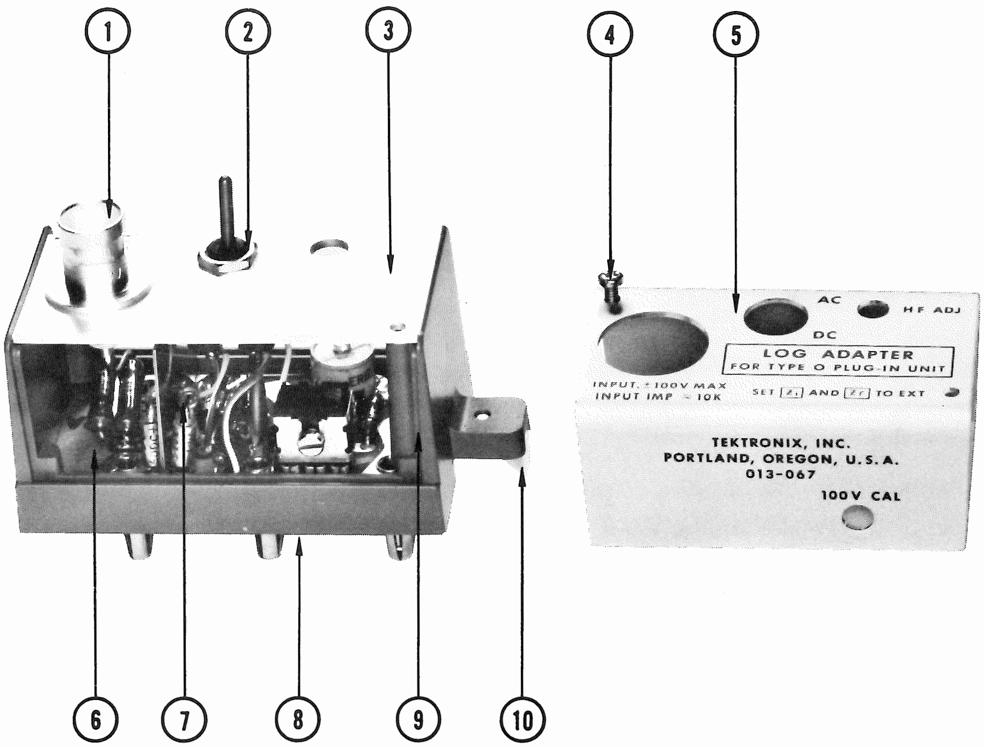


Fig. 2. Log Adapter Schematic

LOG ADAPTER PARTS LIST



REF. NO.	PART NO.	SERIAL NO. EFF.	SERIAL NO. DISC.	QTY.	DESCRIPTION
1.	131-126			1	CONNECTOR
2.	260-511			1	SWITCH
	210-562			2	NUT, hex $\frac{1}{4}$-40 x $\frac{5}{16}$
3.	387-739			1	PLATE
4.	211-079			4	SCREW, 2-56 x $\frac{3}{16}$ PHS slotted
5.	200-457			1	COVER
6.	388-545			1	BOARD, etched circuit
7.	337-565			1	SHIELD
8.	392-146			1	BOARD (not shown)
	211-095			2	SCREW, 2-56 x $\frac{5}{16}$ FHS slotted, 80° (not shown)
	134-070			1	PLUG (not shown)
9.	204-163			1	BODY
10.	401-022			1	CAM
	309-392			2	R1-R2 RESISTOR, 20 k, $\frac{1}{2}$ w, prec, 1%
	318-084			1	R3 RESISTOR, 10 k, $\frac{1}{8}$ w, prec, 1%
	315-560			1	R4 RESISTOR, 56 Ω $\frac{1}{4}$ w, comp, 5%
	316-275			1	R5 RESISTOR, 2.7 meg, $\frac{1}{4}$ w, comp, 10%
	318-003			1	R6 RESISTOR, 11.1 k, $\frac{1}{8}$ w, prec, 1%
	318-012			1	R7 RESISTOR, 25 k, $\frac{1}{8}$ w, prec, 1%
	311-359			1	R8 POTENTIOMETER, 5 k, var. 20%
	315-562			1	R9 RESISTOR, 5.6 k, $\frac{1}{4}$ w, comp, 5%
	290-177			4	C1A-D CAPACITOR, 1 μf, 50 v, EMT, 20%
	283-023			1	C2 CAPACITOR, .1 μf, 10 v disc type
	281-060			1	C3 CAPACITOR, 2-8 pf, cer, var.
	152-110			1	D1-D2 DIODE, matched pair coded 6110
	152-109			1	D3-D4 DIODE, matched pair coded 6109
	152-111			1	D5-D6 DIODE, matched pair coded 6111

COMPENSATING ADAPTER FOR TYPE O PLUG-IN UNIT

Tektronix Part No. 013-081

The Compensating Adapter is an accessory for the Tektronix Type O Operational Amplifier Plug-In Unit. It extends the high-frequency performance of either operational amplifier when the internal Z_i and Z_f resistors are used in any combination for either gain or attenuation. (See "Closed-Loop Gain-Bandwidth Characteristics of Operational Amplifiers" in Section 2 of the Type O instruction manual.)

Without the Compenstating Adapter, the stray capacitance associated with the internal Z_i and Z_f resistors limit the operational amplifier high-frequency performance. With the adapter, the Z_i and Z_f resistors are paralleled by the HF ADJ differential capacitor (see Fig. 1) to supplement their stray capacitance. HF ADJ may then be set to equalize the Z_i and Z_f time constants so that optimum performance is obtained.

Characteristics

The following characteristics apply to the Compensating Adapter and Type O combination:

Frequency Response—See Fig. 2.

Input Resistance—0.01 to 1 megohm; determined by Z_i SELECTOR position.

Input Capacitance—About 40 to 450 pf; dependent on Z_i and Z_f SELECTOR positions (maximum at X100 gain).

Maximum Input Voltage—400 v dc or 150 v rms. Derating required above 1 mc; see Fig. 3.

Maximum Output Voltage—±50 v peak.

Maximum Temperature—+55° C.

Operating Information

Plug the Compensating Adapter into the front-panel jacks of either operational amplifier of the Type O plug-in unit. The cam on the adapter body insures that the ± GRID SEL switch will be set to (—). Set the Z_i and Z_f SELECTOR switches to the desired values and check the DC BAL., GAIN, and OUTPUT DC LEVEL adjustments as described in Section 2 of the Type O instruction manual. Set the switch on the adapter to the appropriate position. Apply a square-wave signal (50 nsec or less risetime) to the operational amplifier and set the adapter HF ADJ control for optimum square wave display. It will be necessary to repeat the HF ADJ adjustment for each combination of Z_i and Z_f resistors used.

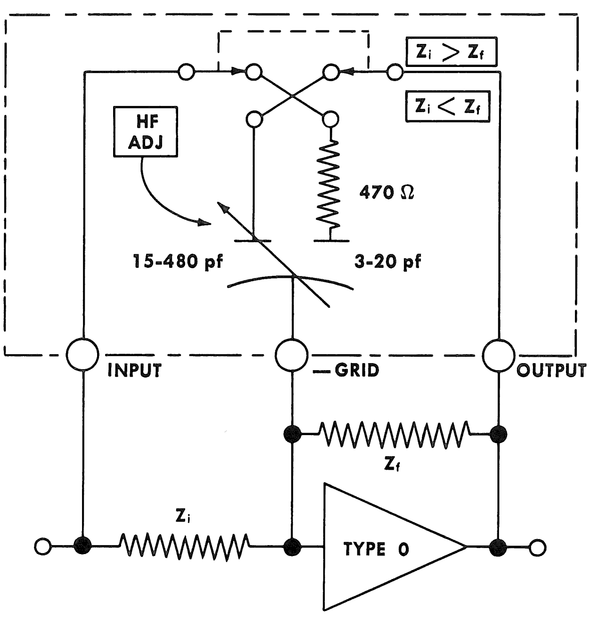


Fig. 1. Compensating Adapter with Type O.

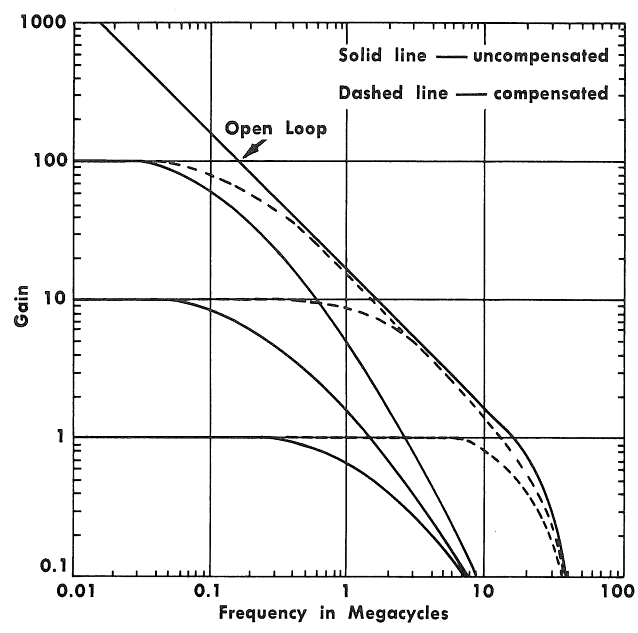


Fig. 2. Typical frequency response of a Type O with and without compensation.

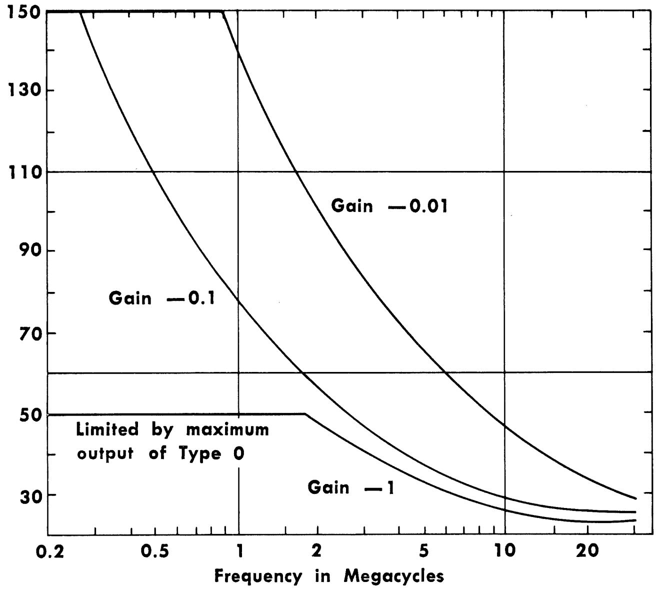
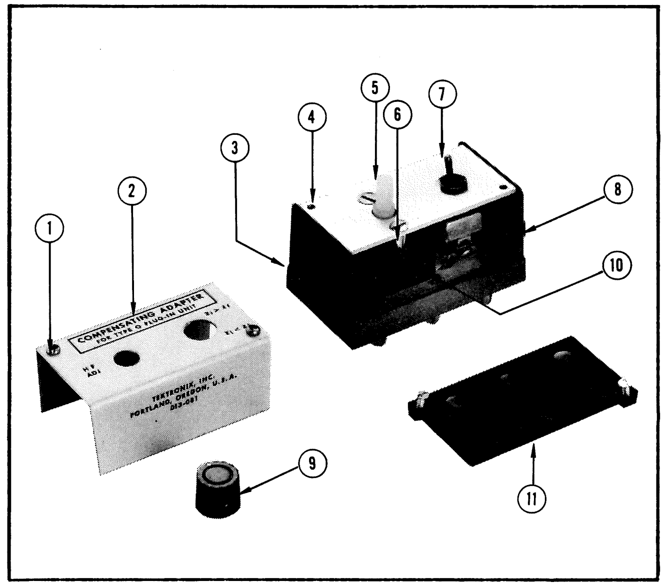


Fig. 3. Input voltage derating vs frequency.



Parts List Reference

PARTS LIST

REF. NO.	PART NO.	SERIAL NO. EFF.	SERIAL NO. DISC.	QTY.	DESCRIPTION
1	211-079			2	SCREW, 2-56 x 3/16 inch, PHS phillips
2	200-546			1	COVER
3	204-163			1	BODY
4	387-869			1	PLATE, sub panel
5	384-298			1	ROD, shaft extension, nylon, .656 inch long
	214-110			1	PIN, roll, 1/16 x 1/4 inch
	354-220			1	RING, external retaining
	210-803			1	WASHER, 6L x 3/8 inch
6	166-025			2	TUBE, spacer, 1/4 inch long
	- - - -			-	Mounting Hardware: (not included)
	211-103			2	SCREW, 4-48 x 7/16 inch FHS
7	260-398			1	SWITCH, miniature
	- - - -			-	Mounting Hardware: (not included)
	210-562			1	NUT, hex, 1/4-40 x 5/16 inch
	210-046			1	LOCKWASHER, internal, .400 OD x .261 inch ID
	210-854			1	WASHER, 17/64 x 1/2 inch OD
8	401-022			1	CAM, switch actuating
	- - - -			-	Mounting Hardware: (not included)
	213-055			1	SCREW, thread cutting, 2-56 x 3/16 inch PHS
9	366-210			1	KNOB, charcoal
	- - - -			-	Includes:
	213-020			1	SCREW, set, 6-32 x 1/8 inch HSS
10	388-571			1	BOARD, etched circuit
	- - - -			-	Mounting Hardware: (not included)
	211-079			2	SCREW, 2-56 x 3/16 inch, PHS phillips
11	392-146			1	BOARD, insulating
	- - - -			-	Mounting Hardware: (not included)
	211-095			2	SCREW, 2-56 x 5/16 inch FHS
					ELECTRICAL COMPONENTS
	304-471			1	RESISTOR, 470 Ω, 1 w, 10%
	281-090			1	CAPACITOR, var., differential, 3-20 and 15-480 pf.

Application Notes

Tektronix No. 013-086

Semiconductor Leakage Current Adapter For Type O-Unit
(Rev. 10-1-64)

General Description:

The 013-086 Diode Leakage Test Adapter is a plug-on assembly for the Type O operational amplifier plug-in unit for the 530-40-50-80* series oscilloscopes, facilitating the measurement of semiconductor leakage currents over the range of 50 picoamperes** to 50 microamperes, for reverse-bias voltages of 200 mv to 100 v.

The oscilloscope sawtooth provides the bias source for the junction under study, to display the entire leakage characteristic (leakage versus bias) in a single presentation.

The adapter may also be used to plot junction capacitance as a function of bias voltage, for junctions whose leakage-to-capacitance ratio does not exceed about .01 microampere per picofarad at the voltages of interest. Techniques are discussed below for measuring capacitance in the presence of leakages up to 0.1 μa per pf of capacitance.

The leakage test adapter contains a step attenuator for the input sawtooth, an adjustment to standardize the sawtooth amplitude applied to the attenuator, a carefully designed and shielded test jig having negligible end-to-end leakage and capacitance, and a precision feedback network to provide the high sensitivity required for present-day leakage measurements.

Junction bias ranges of 0-10, 0-20, 0-50 and 0-100 percent may be calibrated directly in volts when the adapter is used with any 530-40-50-80 series oscilloscope except Types 544-546-547, where the available sawtooth amplitude is usually about 9 to 9.5 v per cm. With these latter instruments, the user has the option of standardizing the bias ranges to 9-18-45-90 v or to 5-10-25-50 v. The 544-6-7 may be modified to provide a larger sawtooth output amplitude. However, because in a large number of cases the 544-6-7 replace former 530-40 series instruments, the user may find it most economical to make the replaced instrument available for leakage tests, since the measurements do not require critical trigger adjustments, sweep rates faster than 0.1 msec/cm or bandwidth and transient response characteristics beyond the kilocycle-microsecond range. Only the mid- and slow-range sweep rate accuracy and the stability of the power supplies are critical to the application.

* Type 81 adapter required to use the O-unit in 580-series oscilloscopes.

** For production testing by unskilled operators, the device should be considered limited to leakage tests at preamp sensitivities near 1 v/cm (1 μA or 1 nA/cm). An operator sufficiently skilled to check and set the output DC level of the O-unit at intervals may extend the measurement sensitivity to 50 pA/cm or better, using techniques discussed in the application notes below.

Familiarization -- Connections and Controls

Sawtooth Input. The banana jack on the front panel accepts a patch cord from the oscilloscope Sawtooth Out connector.

Horiz Volts (X% of Preset Sawtooth) Per Centimeter. The rotary switch on the front panel of the adapter selects how much of the available sawtooth amplitude is applied to the junction under test. The available sawtooth amplitude is set by the:

Preset Adj., which is accessible through the lower lip of the panel. If the oscilloscope provides a front-panel sawtooth output of more than 10 v/cm (about 105 v for 10.5 cm sweep length), the sawtooth adjustment on the adapter may be set to provide display calibration and bias drive of:

%/Cm Setting	Horiz Calib. in v/cm	Peak drive (approx)
10	10	105 v
5	5	52.5 v
2	2	21 v
1	1	10.5 v

The Sensitivity Switch provides direct calibration of the O-Unit's volts/cm switch in .05 to 20 nanoamperes per centimeter or microamperes per centimeter. For measurements of greater than 50 nanoamperes, the μA/cm position must be used.

The Test Jig contains facilities for both transistor and diode measurements. The shield hinges downward and contains the diode carrier (simply drop the diode into the notches with its cathode at the left, and close the shield). For transistors, a divided and internally shielded 4-hole socket allows measurement of small PNP or NPN transistors without having to re-bend the leads. The transistor is rotated for insertion according to its type: collector to the left for NPN, collector to the right for PNP. The base and emitter connections are tied together inside the adapter for the measurement (I_{ces}).

Size Limits for the device under test must be observed for high-sensitivity measurements. To be effective in suppressing coupling to and between diode leads during leakage or capacitance measurements on high-performance glass-cased signal diodes, the diode-holder shield is purposely made close-fitting.

As a consequence, the device is not well suited for making high-sensitivity leakage and capacitance measurements on power or zener diodes with large metal cases, stud-mounts, etc., or on transistors in cases larger than TO-18 or TO-5. It will not usually be possible to close the shield on a transistor, but with most types observed so far (PNP with collector tied to case is the critical type) the noise level does not prove objectionable if the operator keeps his body grounded. Noise suppression and other techniques for specific measurements are discussed below.

Familiarization -- First Time Operation

1. Calibration

Install the adapter in operational amplifier A. Set the operational amplifier controls as follows:

Z_i	EXT
Z_f	EXT
LF Reject	OFF.

Patch from oscilloscope Sawtooth Output to adapter Sawtooth Input.

Set the oscilloscope sweep controls (Sweep A) as follows:

Triggering:	Automatic, +Line.
Time/Cm:	10 msec/cm.

Install a X10 probe on the O-unit preamplifier input; set the preamplifier input selector to +DC, and V/Cm to 1 v/cm. Apply probe tip to CAL OUT and set calibrator for 100 v output; adjust variable v/cm for exactly 4 cm (540, 540A, 550 or 580 series) or 6 cm (530, 530A, 540B-series or 544-6-7). See that the top and bottom of the display coincide with the graticule lines across the entire 10 cm.

Now, apply probe tip to one of the two left-hand holes in the adapter transistor socket. Set Horiz Volts/Cm switch to 10 v/cm (CW). Adjust the Preset Adj. control on the underside of the adapter for a diagonal trace starting in the lower left corner of the graticule and crossing the top graticule line just at the 10 cm line.

Important: The trace should be horizontally positioned so that the level portion of the trace to the left of the sawtooth-start point is positioned to the left of the first graticule line, and the start of the sawtooth occurs exactly at the corner of the graticule. The flat portion is the vertical system zero reference, which occupies about 1% of the total sawtooth duration.

Special Note for 544-6-7: Sawtooth amplitude in these instruments will in most cases be insufficient to obtain 10 v/cm drive, as intended the above procedure. In these instruments, the Preset Adj. should be set to provide some other convenient value, such as 9 v/cm (sawtooth crossing the right hand graticule line 6 mm from the top), or 5 v/cm (sawtooth crossing the right hand graticule line at the center).

Use of the 10X probe and the 100 v position of the calibrator is important in this step to avoid loading down the Preset Adj. divider in the adapter (1 M loading will introduce a few percent calibration error) and to avoid possible accumulated divider errors.

Once the Preset Adj. has been set, the adapter is calibrated for use in either operational amplifier of any O-unit plugged into the same indicator, i.e., the adapter is matched to the indicator. It must be recalibrated if the indicator is changed, but may be freely interchanged between operational amplifier A and B, or between different O-units in the same indicator.

Return variable V/Cm to the calibrated position after setting the sawtooth adjustment.

2. Operation

> Note: In the discussions below, use of an oscilloscope providing sufficient sawtooth for direct calibration of horizontal drive (10%/cm = 10 v/cm) is assumed.

A. List of suggested equipment for demonstration:

(1) O-Unit, Leakage Test Adapter, 530-40 series oscilloscope.
(2) Red patch lead.
(3) Diodes (other representative types may be used).

Type T13G (25 v low leakage germanium)
Type 6045 (125 v low leakage silicon)
Type 1N999 (75 v low leakage, low capacitance silicon)
Type 1N707 (7 v Zener diode)

(4) Transistors (other representative types may be used).

Type 2N711A (15 v, Ge, PNP) 151-092
Type 151-096 (125 v Si, NPN)

*(5) 1 pf capacitor 281-538
*(6) BNC-BNC Cable 012-076 (20") or 012-057 (42")

If O-unit has UHF connectors, use UHF-UHF cable, such as 012-019 (14").

*(7) O-Unit Compensating Adapter 013-081
*(8) Short patch cord (Cal out to alligator clip) and 1 M resistor

*For high-sensitivity capacitance measurements only.

B. Demonstration No. 1 (25 v germanium diode leakage)

Preset controls as follows:

Sweep Trigger:		+Line, Auto
Sweep Time/Cm:		50 msec/cm
Preamp	Input	-B
	V/Cm	.05 v/cm

Operational Amplifier B -- Install Leakage Test Adapter.

Z_i	EXT
Z_f	EXT
LF Reject	OFF

Leakage Test Adapter

Horiz v/cm	1
Sens.	1 μA/v

(1) Adjust DC level Out of Operational Amplifier B for minimum shift from preamp "Zero Check".

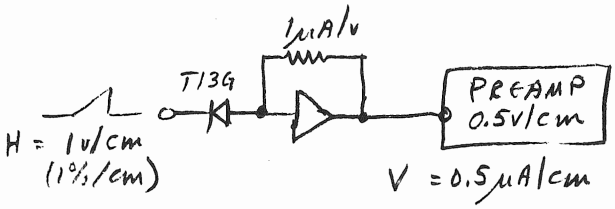
(2) Position baseline to bottom graticule line.

(3) Patch from Sawtooth Out to adapter Sawtooth Input.

(4) Set preamp v/cm to 0.5 (vertical calibration is now 0.5µA/cm) and install T13G diode in adapter shield (cathode to the left) and close shield.

(5) Position the step at the left of the sweep to the left-hand graticule line.

(6) Step Time/Cm to 5 msec/cm, then to 500 msec/cm. Note that display characteristics do not change. This indicates that the indication is of leakage alone, and is not influenced by diode capacitance. With a T13G, the display should be similar to that of Fig. 1.



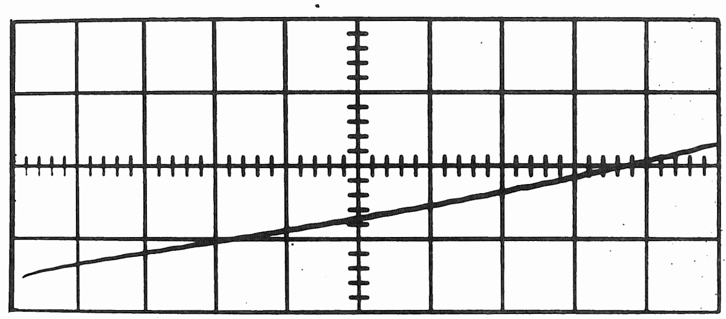


Fig. 1. T13G Leakage, 0-10v

(7) Change preamp v/cm to 1 (= 1 μA/cm), increase horizontal drive to 2%/cm (2 v/cm). The display should now be similar to Fig. 2.

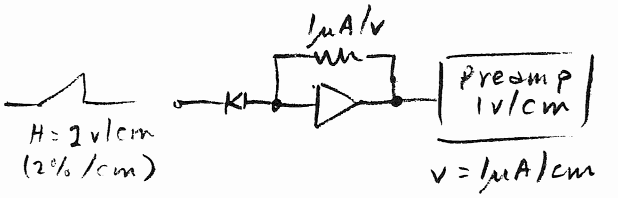
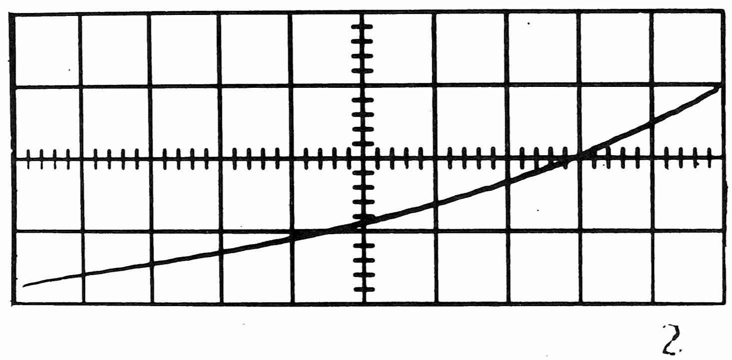


Fig. 2. T13G Leakage, 0-20 v



C. Demonstration No. 2 (125 v low leakage Silicon diode).

Preset controls as follows:

Sweep Trigger:	+Line, Auto
Sweep Time/Cm:	50 msec/cm
Preamp	
Input	-B
V/Cm	0.5 v/cm

Operational Amplifier B (with leakage test adapter installed).

Z_i	EXT
Z_f	EXT
LF Reject	OFF

Leakage Test Adapter

Horiz v/cm	10
Sens.	1 nA/v

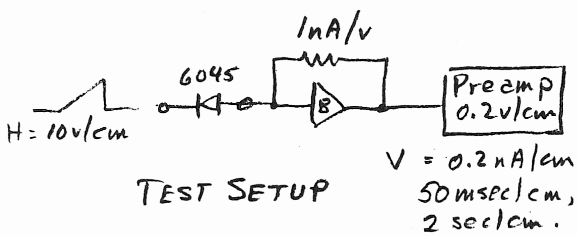
(1) With diode holder empty, check and set output DC level with switch in "Adj" position. Baseline should not shift more than 6 mm (300 pA grid current) when the switch is returned to its normal position. Position baseline to bottom graticule line, <u>using the preamp POSITION control</u>.

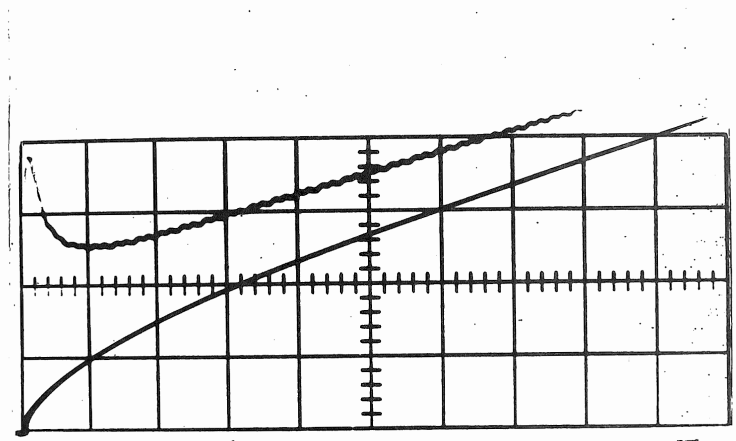
(2) Install Type 6045 diode in shield, close shield.

CAUTION: The diode lead extending to the left reaches a peak voltage of about 105v during this test. Avoid touching it with the fingers when the shield is closed. The diode is disconnected from the drive circuit when the shield is open, allowing safe handling.

(3) Note that the display starts with a spike of 1.5 to 3 cm (upper trace of Fig. 3 is at sensitivity of 0.2 nA/cm. Spike at 0.5 nA/cm should be only 40% as high). The 50 msec/cm sweep rate is too fast for a high-capacitance, low leakage diode, and the trace indicates leakage <u>plus</u> capacitive displacement current. Step sweep rates to slower and slower sweeps until stepping the sweep time/cm has negligible effect on the vertical amplitude indication. For a typical 6045 (5 to 8 pf at 0 v), it will be necessary to set sweep time/cm to 2 sec/cm to obtain a vertical display of leakage alone. A good rule of thumb is to set the sweep at least 10X slower than the range which dropped the amplitude by about 1 cm.

Figure 3 (lower trace) was taken at 2 sec/cm, preamp at 0.2 v/cm for 0.2 nA/cm. The diode is shown to have a leakage of 0.8 na at 90 v bias (1.1 x 10^5 MegΩ equivalent).





3

Fig. 3. Type 6045 diode.

V: 0.2 nA/cm.
H: 10 v/cm.

Upper trace, 50 msec/cm (Leakage, plus capacitive displacement).
Lower trace, 2 sec/cm (Leakage).

Capacitive displacement current is (C) x (v/sec). Vertical trace shift is

$$\text{pf/cm} = \frac{\text{nA/cm} \times \text{msec/cm}}{\text{Horiz. v/cm}}$$

D. Demonstration No. 3 (Capacitance of low leakage, high capacitance Silicon Diode)

Note: For the purposes of this demonstration, capacitance is displayed using the 1 nA/v sensitivity position of the leakage test adapter. In normal use, capacitance would be measured by an alternative technique discussed below, making use of the second operational amplifier with a compensating adapter to obtain the sensitivity required without the bandwidth sacrifice involved in the 1 nA/v position of the adapter sensitivity switch.

Proceed as follows:

(1) Display the leakage characteristic as in Demonstration No. 2, noting the value (cm) at some bias voltage of interest. (It will not be necessary to convert this to nA).

(2) Step the sweep time/cm faster until there is a significant (>1 cm) increase in vertical amplitude indication at the voltage of interest. Note the displacement. The capacitance sensitivity at this sweep rate is:

$$\text{pf/cm} = \frac{\text{Vertical nA/cm x Sweep Msec/cm}}{\text{Horiz Volts/cm}}$$

The number of picofarads capacitance will be proportional to the number of centimeters difference between the leakage current and the leakage-plus-displacement current noted at the faster sweep.

CAUTION: Do not change the vertical sensitivity between noting the leakage value and observing the displacement due to capacitance -- only change the sweep rate. If sensitivity is changed, both zero level and leakage value must be rechecked to obtain a proper difference reading.

In Fig. 3, taken at a sensitivity of 0.2 nA/cm, the sweep was changed from 2 sec/cm to 50 msec/cm, for a capacitance sensitivity of

$$\frac{0.2 \times 50}{10} = 1 \text{ pf/cm}.$$

At the 50 v bias point (graticule centerline) the displacement difference was 9 mm or 0.9 pf.

Note that at a much lower bias point the difference -- and hence capacitance -- is much greater. Though the capacitance indicated is about 3.5 pf at perhaps 1 v bias the measured value at 100 mv was actually about 6 pf. The risetime of the 1 nA/v circuit is insufficient to show the value below about 1 v in this 10 v/cm display.

In the interest of accuracy, it should be noted that the "leakage only" curve is somewhat distorted by capacitive indication. The capacitance sensitivity at 2 sec/cm was

$$pf/cm = \frac{0.2 \times 2000}{10} = 40\,pf/cm.$$

About 1 mm of that initial step in the leakage curve at 0 v, then, is due to capacitance. At the 50 v mark, where capacitance is only 1 pf, the error (~1/4 mm) is negligible. Where leakage at 0 v is critical, the capacitance effect can be minimized by going to 1 v/cm horizontal drive at the same sweep rate, rather than delay measurements by going to an even slower sweep rate.

E. Demonstration No. 4. (Preferred Capacitance Method)

Demonstration No. 3, using the 1 nA/cm sensitivity of the leakage adapter is adequate for cases where the capacitance is relatively high, the leakage low, and measurements very close to 0 v are not required.

Using the second operational amplifier as a X1 to X100 compensated amplifier (with compensator 013-081 or equivalent), sweep rates up to 0.1 msec/cm may be used to evaluate small values of capacitance, and -- with special techniques, moderate values of capacitance in the presence of considerable leakage. With the second operational amplifier in place, offset, tailored noise reduction and other techniques are also available to facilitate measurements.

Set up as follows:

(1) Install compensating adapter in operational amplifier A.

(2) Carefully check DC output level of operational amplifiers A and B (open the shield to disconnect the diode in B and set sensitivity to 1 µA/v before checking level, then connect the output of B to the input of operational amplifier A, using coaxial (shielded) cable. Set the preamp to 1 v/cm, +A and operational amplifier A to $Z_i = 10k$, $Z_f = 1\,M$. Touch up DC balance of B (using Adj switch) as the preamp sensitivity is increased to 0.1 v/cm for an overall sensitivity of 1 nA/cm. Position trace to bottom graticule line.

(3) Set horizontal drive for 1%/cm (1 v/cm).

(4) Install diode and close shield. Step sweep rate clockwise until a rate is found which provides a substantial increase (>1 cm) in vertical deflection <u>but not faster than 0.2 msec/cm</u>. If the sweep rate is 200, 100, 20, 10, 2, 1 or 0.2 msec/cm, calculate capacitance sensitivity at this rate by the formula

$$pf/cm = \frac{nA/cm \times msec/cm}{Horiz\ v/cm}$$

(When horiz v/cm is 1 and nA/cm is 1, vertical capacitance sensitivity in pf is equal to the msec/cm setting of the time/cm control).

If the sweep rate is 500, 50 or 5 msec/cm, set the preamp to 0.2 v/cm (2 nA/cm), step up to the next faster sweep (200, 20 or 2 msec/cm) and calculate the capacitance sensitivity at the faster rate (400, 40 or 4 pf/cm). If the sweep rate is 0.5 msec/cm, step back to 1 msec/cm and set the preamp to .05 v/cm (0.5 nA/cm). In this case the capacitance sensitivity is now 0.5 pf/cm.

(5) Adjust compensation of operational amplifier A for best risetime with minimum noise, having the compensator selector switch at $Z_i < Z_f$. At highest sensitivity, it may be an advantage to set Z_f of operational amplifier B to 10 pf (make sure that LF Reject is off, as it will otherwise degrade measurement accuracy). Do not set Z_f as high as 100 pf, however, as risetime will be considerably degraded.

(6) Step sweep to a rate 1 step faster than that for which capacitive sensitivity was calculated (i.e., double the sweep speed), and note the difference in vertical displacement. This difference, multiplied by the calculated capacitance sensitivity for the slower sweep, is the actual capacitance.

Explanation: The vertical displacement in the first setting was due to capatance plus leakage. By doubling the sweep speed, the capacitance sensitivity was doubled, giving effectively a reading of 2C+L if calculated for the slower rate. The difference between the readings is (2C+L) - (C+L), or C.

The same technique may be used for very low capacitances, using a 10X change in sweep rate, the difference then becoming (10C+L) - (C+L), or 9C. The indicated capacitance in this case is divided by 9 to obtain the true value. The 2X change, of course, is easier to handle. The trick is to select a slower rate which is a factor of 2 slower than the next faster step. When the rate change is 2-1/2 to 1, as it is when the slower rate is 500, 50, 5 or 0.5, it's usually easier to change sensitivity combinations than to divide the answer by 1.5. For a demonstration of the 10X sweep, 9X capacitance technique, see section H (4) and Fig. 10.

F. Demonstration No. 5. (Capacitance near 0 v).

The point at which diode capacitance is highest is in the region of zero bias, and it's often most important to obtain a reading as close to zero as possible, but without driving the diode into conduction, since recovery characteristics (diode and/or amplifier) after conduction will mask the capacitive effects.

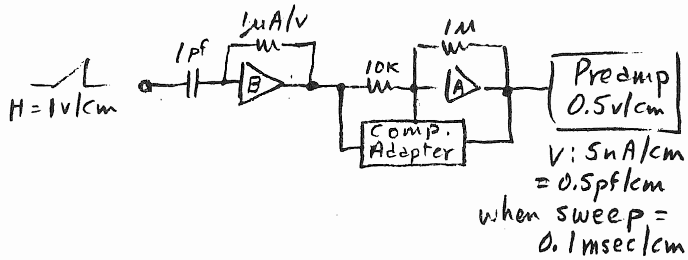
For measurements near zero volts, it is necessary to obtain as fast a risetime as possible in the measuring system, consistent with noise level and sensitivity requirements.

The smallest value of drive obtainable in the leakage test adapter is 1%/cm, or normally, 1 v/cm.

By turning on the magnifier, however, the CRT horizontal display resolution may be improved without affecting the drive to the device under test, since the oscilloscope magnifier affects only the CRT display, not the generated sawtooth which appears at the front panel.

With a X5 magnifier, the drive to the device under test in the "1%/cm" position remains 1 v per centimeter of the unexpanded sweep, but the CRT display is expanded so the display is 200 mv/cm.

It is possible (Fig. 4, lower trace) to obtain a sensitivity of 0.5 pf/cm with the system 0-98% risetime occupying no more than the first 50 to 100 millivolts of the diode characteristic.



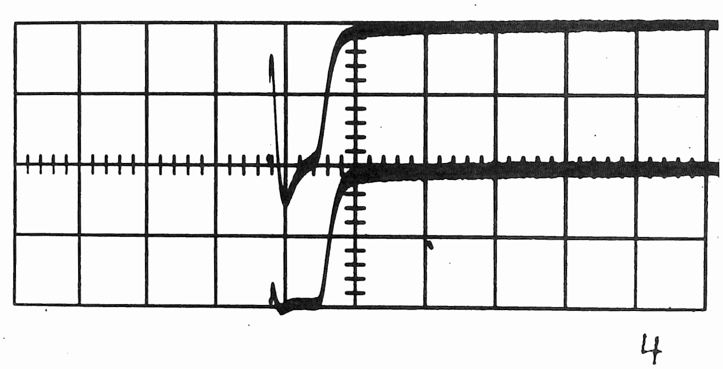


Fig. 4. Capacitance Measurement.

Use of compensating adapter provides fast (<15 µsec) overall system 0-98% risetime. Unmagnified sweep rate, 0.1 msec/cm; X5 Mag on for better resolution.

Upper trace: interference from +Gate.
Lower trace: 470 pf across sawtooth output, allows 0.5 pf/cm sensitivity.

The device used for figure 4 is a 1 pf capacitor having a slight positive voltage coefficient of capacitance.

The upper trace of Fig. 4 shows an anomoly -- loss of adequate zero reference at the start of the trace when operating at 0.1 msec/cm. The cause is a small amount of the leading edge of the +Gate waveform capacitively coupled into the sawtooth waveform, momentarily turning on the coupling diode in the leakage test adapter. This spike can be materially reduced and zero reference restored by installing a 470 pf capacitor (.00047 µf) between the sawtooth output terminal and the adjacent ground terminal, or between the +Gate output terminal and ground. It is not recommended that this modification be made permanent, however, as it will cause material degradation of the sweep or gate waveform at rates faster than those used in this application, and may cause internal sweep waveform aberrations at the fastest sweep rates.

Some other precautions to be observed in attempting noise reduction are noted below under "noise reduction techniques" (Section I below).

To measure input capacitance near 0 v, then, proceed as follows:

(1) With the preamp set to 0.5 v/cm and sweep time/cm at 1 msec, +Internal triggering, move input cable of operational amplifier A from B to the Cal Out connector and apply 2 mv of calibrator waveform. Adjust compensating adapter for optimum squarewave compensation (minimum risetime with minimum overshoot).

(2) Return cable to the output of operational amplifier B, and reset triggering controls for line triggering. Set Time/cm to 0.2 msec/cm. If Z_f of operational amplifier B was set to 10 pf in the previous demonstration, return it to "EXT".

(3) Install 470 pf capacitor from Sawtooth Out terminal to ground (stop the sweep generator for this operation).

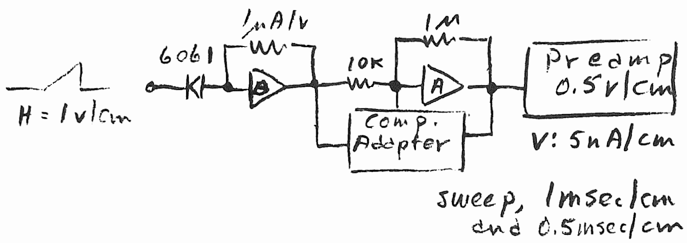
(4) Install 1 pf capacitor in diode holder, set horizontal drive (%/cm) for 1 v/cm and sensitivity to 1 µA/cm.

(5) Close adapter shield. Display should be a step 1 cm high.

(6) Position the step horizontally to graticule center and turn on the magnifier. Note that there is no increase in display amplitude. Now increase sweep speed by a factor of two to 0.1 msec/cm. The display should resemble the lower waveform of Fig. 4, with the system risetime no greater than 1/2 cm horizontally (100 mv).

Having confirmed the system's measurement capability, remove the 1 pf capacitor and substitute a diode, proceeding with the two-step capacitance measurement technique outlined in demonstration No. 4. Figure 5 shows the capacitance of a Type 6061 diode at 100 mv (0 v is 1/2 cm to the left of graticule center), at sweep rates differing by a factor of two (1 msec/cm and 0.5 msec/cm). The calculated sensitivity at the slower rate is 5 pf/cm. The difference in deflection being 1.8 cm, the actual capacitance of this diode is shown to be 9 pf (out of spec, incidentally, as this diode is supposed to be 8 pf max at 0 v).

Current sensitivity of the system set up for Fig. 5 is 5 nA/cm; at this point we might backtrack to find the leakage. If 1.8 cm of the initial 2 cm step was capacitance, leakage must be 0.2 cm or 1 nA. The actual measured value was 1.5 nA at 100 mv. This backtracking method is not recommended as a leakage measurement except where speed is essential. The technique is most useful in estimating leakage measurement error due to capacitive displacement current at the slow sweeps used for leakage tests (see pages 8 and 9).



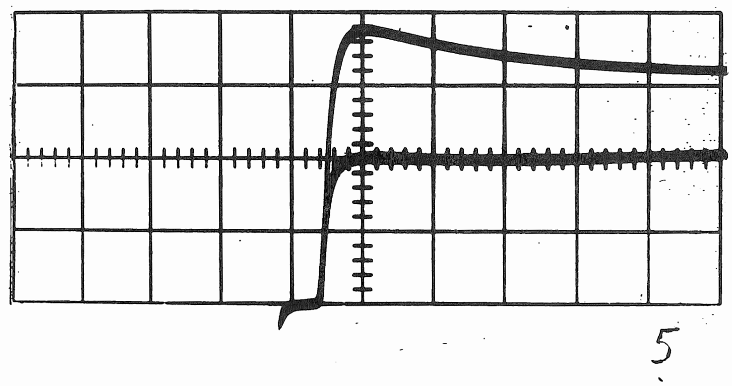


Fig. 5. Capacitance of 6061.

Two-step sweep method (1.0 and 0.5 msec/cm) shows ΔI of 1.8 cm. At 5 pf/cm, C = 9 pf at centerline, representing 100 mv bias (X5 Mag on to improve display resolution).

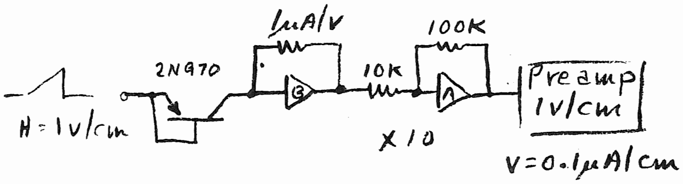
G. Demonstration No. 6: Capacitance Measurement in the Presence of Substantial Leakage.

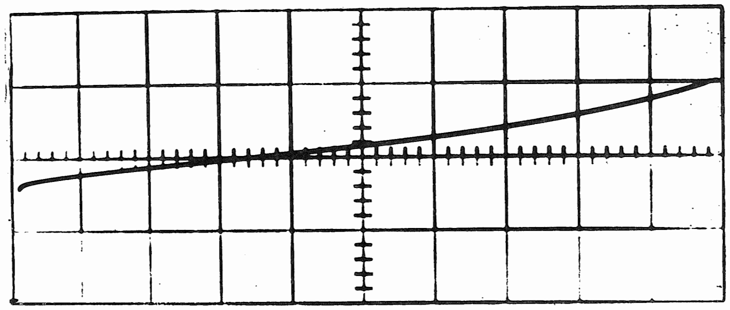
The current sensitivity required to measure small amounts of capacitance in the presence of substantial leakage currents presents a problem in having the trace onscreen (i.e., within range of the position control) in order to observe the capacitive displacement current.

To solve the problem, use of the second operational amplifier is again required, this time as a slideback amplifier, to put the signal of interest on-screen.

Rather than go to the bother of constructing a DC offset voltage supply, we suggest using the oscilloscope's square-wave calibrator, which provides effectively a two-trace display, one trace offset by the calibrator amplitude.

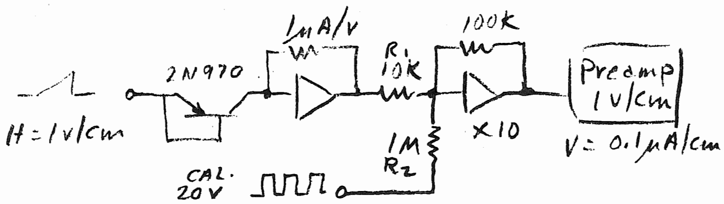
The problem and solution are illustrated in Fig. 6. In (a) we see the leakage characteristic (I_{ces}) of a Type 2N970 transistor, displayed at a suitably slow sweep rate. (The step at the start is characteristic of many semiconductors and is not due to capacitance. However, it can be a serious source of measurement error -- see Section K below).

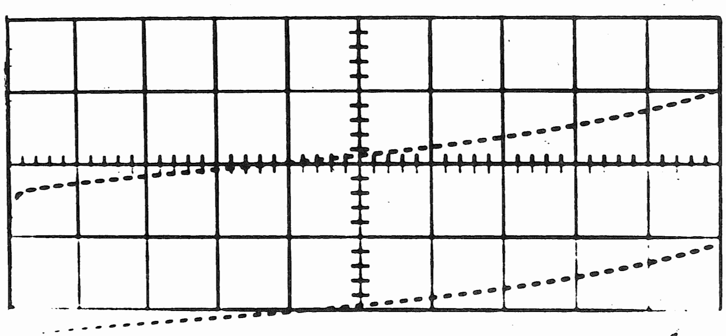




6a

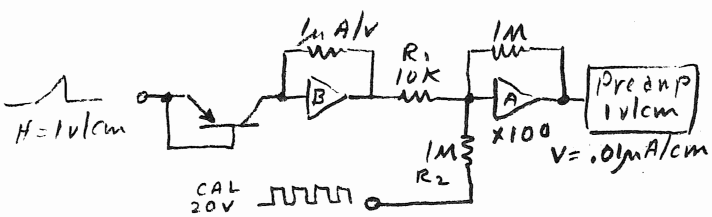
(a) Leakage, 2N970 (I_{ces}) 0–10 v, 2 msec/cm. Step is due to high leakage near 0 v, not primarily due to C. Leakage at 5 v is > 200 nA.

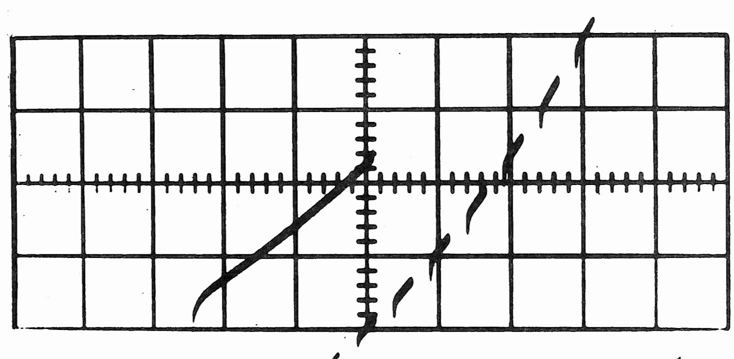




6b

(b) Calibrator used to offset leakage to near 0 v level.





6c

(c) Using offset to allow increasing gain for two-step sweep rate method of measuring C. Lower trace, 2 msec/cm; upper trace, 0.2 msec/cm. ΔI = 2.25 cm ("45 pf") ÷ 9 = 5 pf at 5 v. See text for details.

FIG. 6. USING CALIBRATOR SIGNAL FOR OFFSET.

Display was triggered from calibrator for this illustration; normally, asynchronous sweep would be used to obtain continuous trace.

The schematic and waveform (b) show how the injection of 20 v of the calibrator waveform via a 1 M resistor ($Z_f/Z_i = 0.1$ for this signal) produces 2 v chopped offset to the preamp, approximately equal to the amplifier A output for the leakage at the point ($E_c = 5$ v) we want to examine.

The voltage steps available from the calibrator will usually be sufficient to offset the leakage signal to within the range of the O-Unit position control.

If it is desired to make the offset exactly equal to the leakage, in order to retain zero reference and to avoid adjustment of the position control, a 470 k resistor in series with a small 500 k pot (such as 311-361) is recommended, installed in series with the calibrator signal. The resistor and pot should be located as close to the -grid input of the operational amplifier as possible, to avoid noise pickup. Once the offset (Cal signal via R_2) matches the input via R_1, Z_f may be varied at will, and only the gain will change, not the position.

The waveforms in Fig. 6 (c) show the result (lower waveform) of increasing gain 10X -- which may be done either by changing operational amplifier A Z_f to 1 M or by stepping the preamp sensitivity to 100 mv/cm -- and of increasing sweep speed by 10X, from 2 msec/cm to 0.2 msec/cm rate (upper trace). The capacitance sensitivity at 2 msec/cm and 10 nA/cm with 1 v/cm horizontal drive is 20 pf/cm. Increasing sweep speed X10 gives us 2.25 cm increase in vertical deflection, or a "45 pf" indication at the centerline (representing $E_c = 5$ v). Actual capacitance is 1/9 of this amount (see page 10 for calculation) or 5 pf.

The leakage/capacitance ratio for this transistor is 40 nA/pf. The technique may be extended by a factor of better than two before sweep-rate, sensitivity, noise and risetime limitations come seriously into play to limit measurement accuracy.

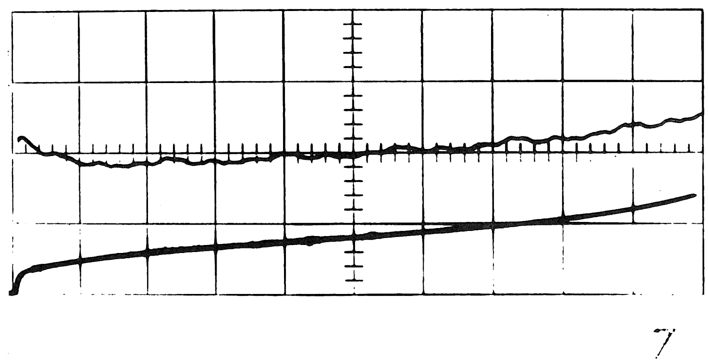
One limitation not mentioned is that the O-Unit compensating adapter was not installed when the offset technique was used, as the compensating adapter provides no convenient external access to the -grid. The lack of compensation makes for a longer risetime in the system, and so limits the capacility of making these slideback measurements near zero volts. The compensating adapter may be modified, or another adapter may be constructed in the O-Unit blank terminal adapter chassis 013-048 to provide both the compensation and summing circuitry if desired. See also H (2) below.

H. Miscellaneous Demonstrations:

(1) Low-leakage, low-capacitance diode Type 1N999. Figure 7 illustrates the problem of measuring capacitance and leakage in the same display when the leakage is very low. The sensitivity required for the leakage measurement (0.5 nA/cm or better) is at the limit for short-term stability of the cascaded amplifier arrangement required for capacitance measurement (1 sec/cm or slower sweep required for leakage test to avoid capacitive effects). The

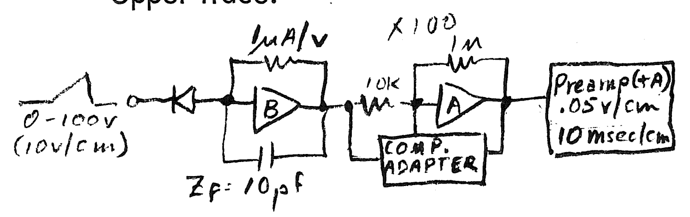
sensitivity of the system (overall nA/cm) must be the same for both measurements if leakage is to be subtracted from the capacitance indication.

Fig. 7. Leakage (lower trace) and capacitance (upper trace) of 1N999, 0.5 nA/cm vertically, 10 v/cm horizontally. Capacitance sensitivity (difference) is 0.5 pf/cm.

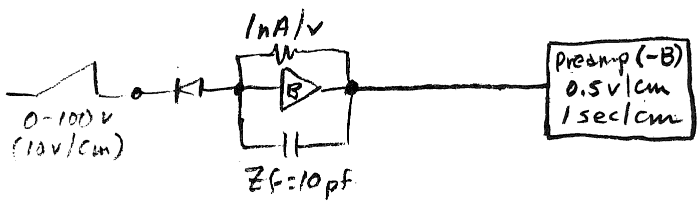


Test setup:

Upper trace:



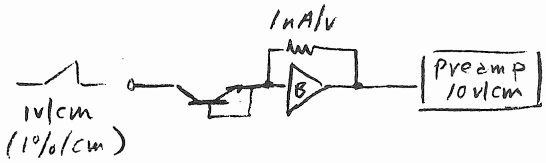
Lower trace:

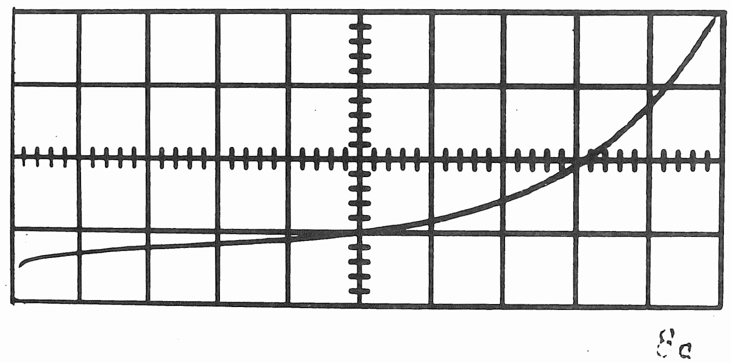


The measurement may be made using photography or a storage indicator, changing measurement systems between shots as shown in Fig. 7. The preferable solution is the two-sweep-rate method of capacitance measurement, obviating the need for observation of leakage during the capacitance measurement (see (2) and Fig. 8 below).

(2) Separate Leakage and Capacitance Measurement (2N706B: NPN Silicon, 25 v transistor). With the operational amplifiers cascaded for high sensitivity for the 1 µA/v adapter setting, the DC output level control of the <u>second</u> operational amplifier may be used for a limited amount of offset to bring the capacitance measurement (2-sweep-rate method) on screen in the presence of substantial leakage. Figure 8 (a) is the leakage measurement (10 nA/cm) showing leakage of ~5 nA at low voltage and 10 nA at 5 v. Waveform (b) shows the capacitance measurement taken at higher sensitivity (1 nA/cm), stepping from 2 msec/cm (2 pf/cm) to 1 msec/cm, for a capacitance measurement of 2.8 pf at 5 v. Waveform (c) shows the capacitance at 0.5 v by the same method (4 pf), using the DC output level control of operational amplifier A (used as X50 amplifier) to bring the waveform within position control range.

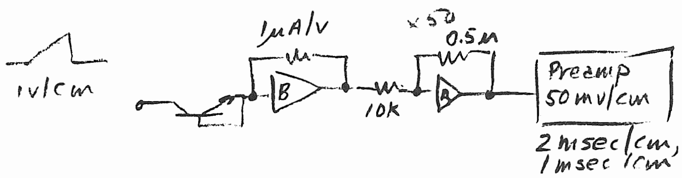
CAUTION: The DC output level of the operational amplifier used with the leakage test adapter MUST BE SET TO ZERO at all times, and particularly with germanium devices. See <u>Precautions and Limitations</u> below.

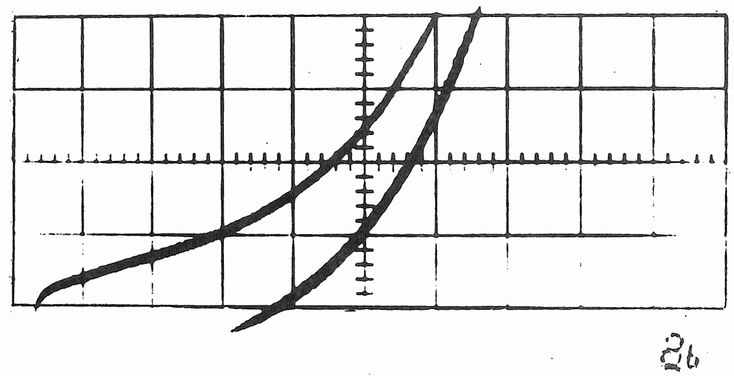




(a) Overall calibration

V : 10 nA/cm
H : 1 v/cm
T : 1 sec/cm

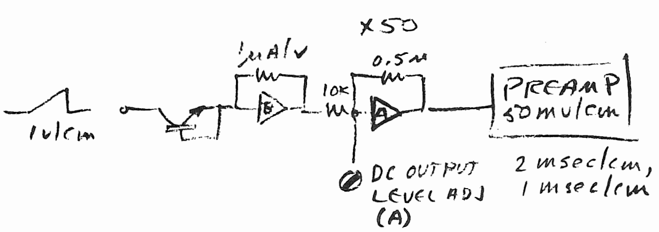


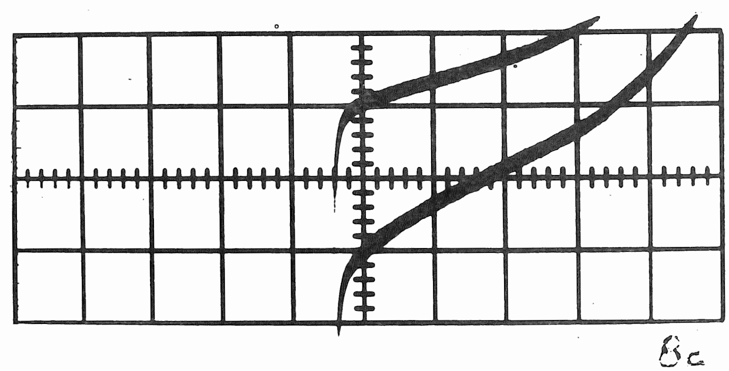


(b) Overall calibration

V : 1 nA/cm ⟶ 2 pf/cm
H : 1 v/cm, zero at 0 cm
T : 2 msec and 1 msec/cm

C at 5 v, 2.8 pf





(c) Overall calibration

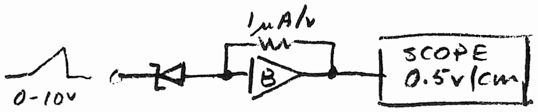
V : 1 nA/cm ⟶ 2 pf/cm
H : 1 v/cm, zero at +4.5 cm
T : 2 msec and 1 msec/cm

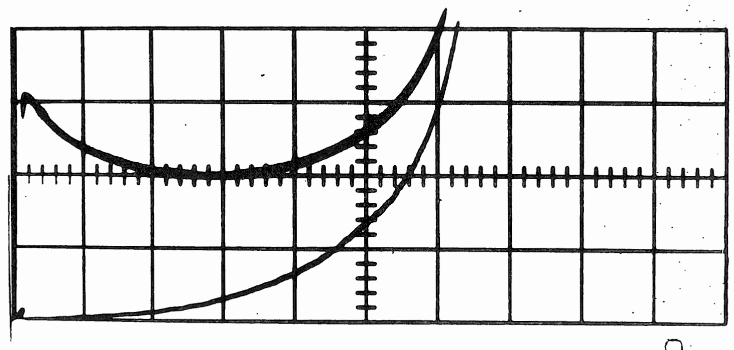
C at 0.5 v, 4 pf

FIG. 8. Capacitance measurement, 2N706B, by 2-step sweep method, using separate sensitivity settings for (a) leakage (10 nA/cm) and (b, c) capacitance (1 nA/cm) measurements. See Section H (2).

Fig. 9a. Leakage and capacitance of 1N707 Zener diode (7 v Zener).

Horizontal	1 v/cm
Vertical	0.5 μA/cm
Upper trace	0.2 msec/cm (100 pf/cm)
Lower trace	50 msec/cm



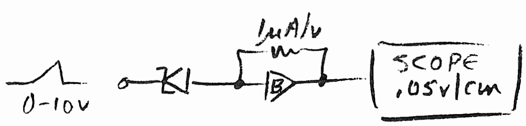


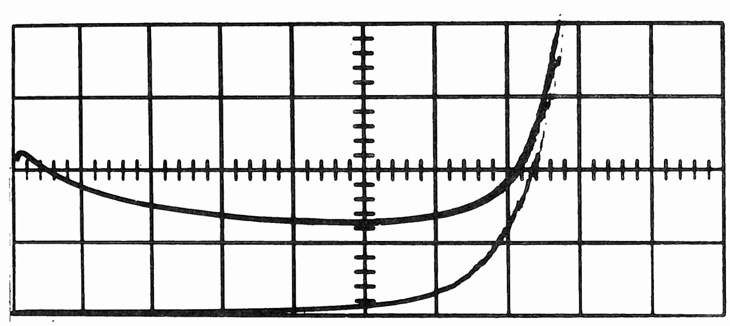
9a

Capacitance at 0.2 v, 310 pf.

Fig. 9b. Leakage and capacitance of 1/4 M10Z10 (10 volt Zener).

Horizontal	1 v/cm
Vertical	.05 μA/cm
Upper trace	10 msec/cm (500 pf/cm)
Lower trace	2 sec/cm

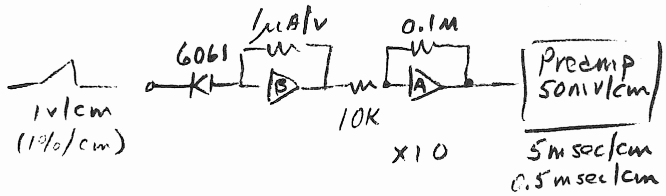




9b

Capacitance at 0.2 v, ~1150 pf.

(3) Zener Diodes. Capacitance of even the small 250 mw zener diodes is quite large. Figure 9 shows the test setup for two popular types, (a) the Hoffman 1N707 and (b) the Motorola 1/4 M10Z10. Figure 9b reveals a noise characteristic typical of the region just below the Zener "knee". At the point where the noise becomes evident (~6.5 v), the diode appears as 260 MegΩ paralleled by 450 pf.



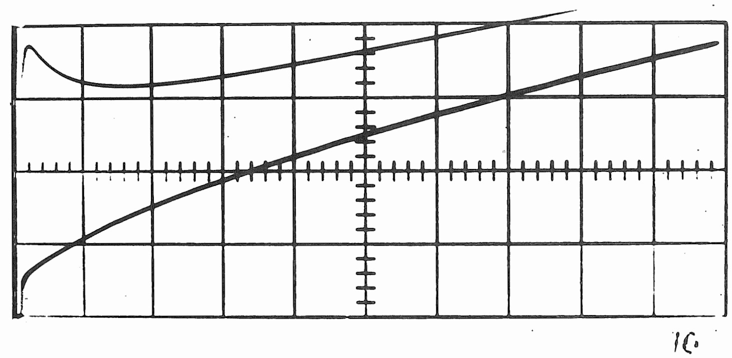


FIG. 10.

Capacitance measurement with 10X sweep step. Indication is 9X capacitance sensitivity of slower sweep (5 nA/cm at 5 msec/cm w/1 v/cm drive ⟶ 25 pf/cm).
Difference in indication at 0.5 msec/cm is

$\frac{25}{9}$ pf/cm or 2.8 pf/cm.

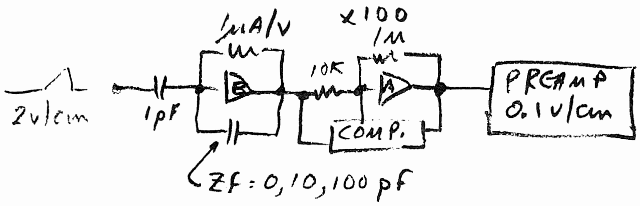
(4) 10X Sweep Technique (Type 6061 diode). Fig. 10 shows a method of obtaining maximum capacitive indication for a given nA/cm sensitivity -- in this case 5 nA/cm. The lower trace is 5 msec/cm (slowest sweep showing noticeable capacitive displacement), the upper trace is 0.5 msec/cm. Lower trace capacitance sensitivity is 25 pf/cm. Displacement of 1.1 cm at 5 v bias is an indication of "27.5" pf; divided by 9 yields 3 pf as the actual capacitance at this voltage. At about 200 mv, actual capacitance is 8.5 pf. As we have seen (Fig. 5) use of a compensated amplifier and the magnifier would reveal an actual capacitance of 9 pf at 100 mv on this particular diode.

I. <u>Noise Reduction Techniques</u>.

Particularly when working with the second operational amplifier to provide extra bandwidth at high sensitivity, noise can get to be a problem as a limit to resolution.

(1) <u>Using Zf</u>. Under most circumstances -- provided the leakage test adapter sensitivity is set to 1 µA/v -- Zf of its operational amplifier may be set to 10 pf to reduce noise without materially affecting risetime. A higher value (100 pf) may be used if risetime is not material -- i.e. if capacitance value near 0 v is not to be measured. A setting of .001 µf may cause oscillations and is not recommended. The 1 nA/v sensitivity position bandwidth is already swamped by the existing strays, and further trimming will have considerable effect on accuracy at the start of the trace.

The LF Reject circuit must be OFF for all operations; it will introduce large measurement errors if switched in when Z_f is set to C.



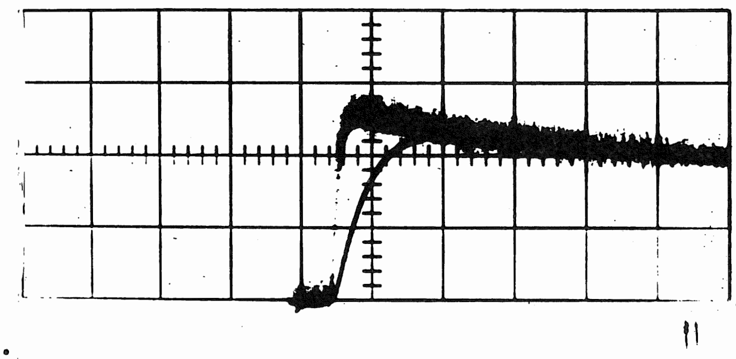


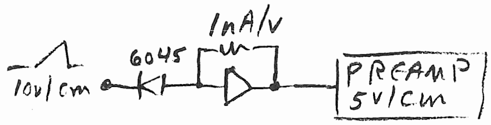
FIG. 11.

Using Z_f for noise reduction (triple exposure). Traces for $Z_f = 0$ or 10 pf are superimposed; 100 pf degrades risetime. Horizontal drive 2 v/cm, 1 msec/cm. Display: 0.2 msec/cm, 0.4 v/cm (X5 Mag.); V = 1 nA/cm (=0.5 pf/cm). (Compensating adapter was set for overcompensation, to emphasize any variations.)

Figure 11 shows the effect of adding 0 pf, 10 pf and 100 pf to a noisy trace at 1 nA/cm obtained by X100 gain (overcompensated) and 0.1 v/cm preamp sensitivity. Note that the 0 pf and 10 pf traces are superimposed, but the degradation of risetime is quite noticeable with 100 pf. The waveform is of a 1 pf capacitor with 2 v/cm horizontal drive at 1 msec/cm for a vertical sensitivity of 0.5 pf/cm. With the X5 magnifier on, the horizontal display is 0.4 v/cm and 0.2 msec/cm. The capacitance indication with 100 pf Z_f is substantially in error for the first half volt and more due to the ~300 µsec risetime of the degraded circuit.

(2) Using the Compensating Adapter. Frequently it's desirable to suppress high-frequency amplifier noise for a short time by more than can be accomplished with the Z_f control of the operational amplifier carrying the leakage test adapter (setting Z_f to .001 µf can frequently cause oscillations). When the amplifier compensating adapter 013-081 is installed on the operational amplifier serving as a X50 or X100 amplifier, however, simply flipping the compensating adapter mode switch from $Z_i < Z_f$ to $Z_f > Z_i$ will reverse the differential capacitor connections and provide a marked reduction in bandwidth, and hence HF noise sensitivity.

Use of this switch position is particularly handy in photographing or storing leakage characteristics at 1 sec/cm or 2 sec/cm. It must be kept in mind, however, that noise characteristics of the semiconductor under test will also be suppressed, along with system noise.



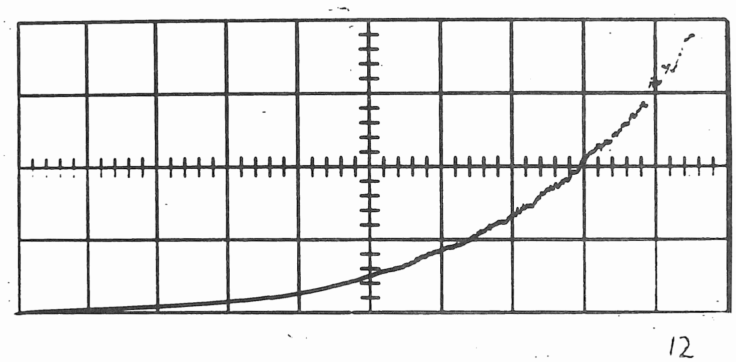
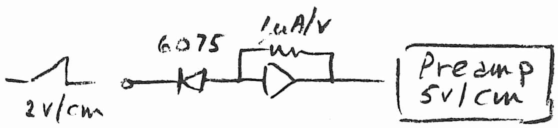


FIG. 12

Type 6045 diode showing some noise above 50 v.

H = 10 v/cm; V = 5 nA/cm.



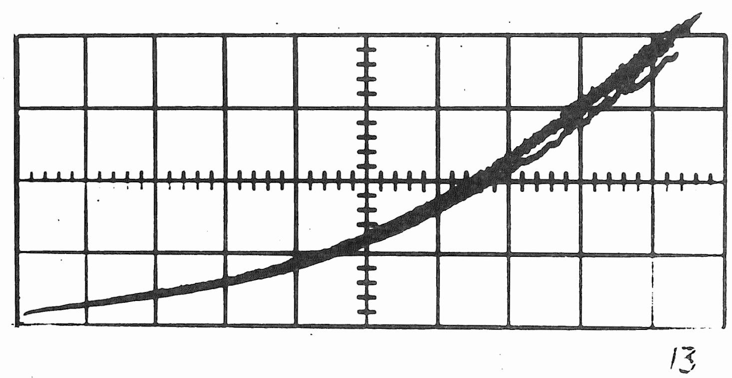


FIG. 13

Noisy Type 6075 diode (5 consecutive traces, 50 msec/cm).

H = 2 v/cm; V = 5 μA/cm.

(3) Non-System Noise. The noise showing on the trace in some cases is not due to the O-Unit but to the device under test. Figures 12 and 13 show noise which is characteristic of the particular semiconductors under test, and not due to the measurement system. Figure 12 is a Type 6045 diode at V = 5 nA/cm, H = 10 v/cm (single shot); Fig. 13 shows 5 consecutive traces (at 50 msec/cm) of a noisy Type 6075 at V = 5 μa/cm, H = 2 v/cm. The noise is not necessarily typical of these diode types; other samples of the same types show negligible noise under the same test conditions. Indicated noise amplitudes should only be taken as a relative noise index, not absolute, since the O-Unit system bandwidth is quite limited under the conditions of measurement. Noise is generally proportional to leakage.

J. Precautions and Limitations.

Serious errors can be introduced into measurements taken without some basic precautions. The most critical are:

(1) (Leakage tests). DC Level Output must be as close to 0 v as the final trace position must be to the true leakage figure, particularly for germanium devices. This is discussed further below (K).

(2) (All tests). Always operate at a low sweep rep-rate (60 cps or slower). For very sensitive measurements, as much as 2 sec between sweeps may have to be allowed.

(3) (All tests). Z_i must be set to EXT in the operational amplifier carrying the leakage test adapter.

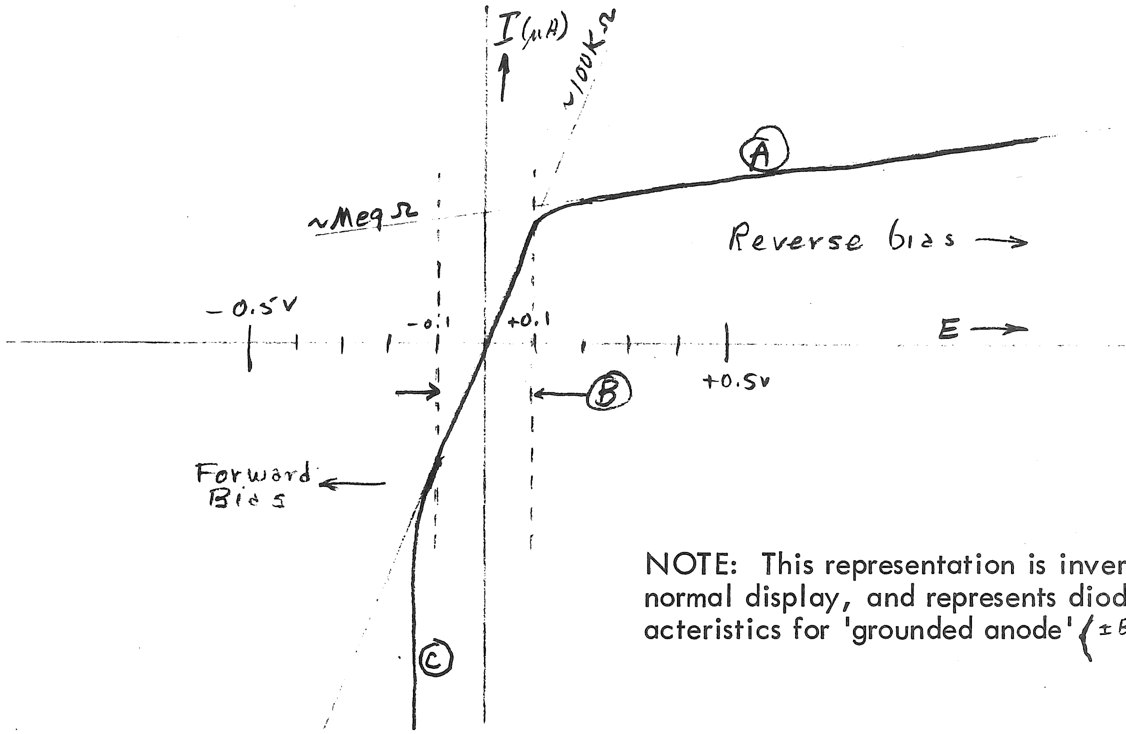
(4) (All tests). Z_f must be set to EXT (or to 10 pf or 100 pf if noise reduction is desirable and risetime degradation is not critical).

(5) (All tests). LF Reject should be OFF. If inadvertently left ON when Z_f is set to C for noise reduction, the resistive feedback will introduce serious (>50%) error.

(6) (Capacitance Tests). Capacitance and leakage must be measured at the same vertical sensitivity if the difference in readings (C) is to be meaningful. It's easy to forget this when using storage or photography. An exception to this rule applies when measuring capacitance by the two-sweep-rate technique; the deflection due to leakage is automatically subtracted in the calculation.

K. Special Note on DC Level. The "foot" at the start of the sweep is a convenient zero reference for positioning the trace, but can be a trap if the operational amplifier Output DC Level is not set accurately. The problem is this: the cathode of the diode under test (we can look at the collector of a transistor simply as a cathode) is always returned to 0 v between sweeps by the adapter's input divider resistor and by the silicon disconnect diode at the adapter input. We assure freedom from circuit recovery (charge and discharge of stray capacitance by picoampere currents) by specifying that the sweep be operated at a slow rate -- 60 cps for most measurements, but with a holdoff up to 2 sec if this proves necessary.

However, because the operational amplifier input is connected to the output by a DC path (Z_f), changing the output DC level (which we do actually by moving the ground reference for the -grid) causes the input grid to move away from ground.



NOTE: This representation is inverted from normal display, and represents diode characteristics for 'grounded anode' (±Bias).

FIG. 14.

Three of the four conductance regions of a germanium diode (the fourth is avalanche -- reverse breakdown -- to the right of the region illustrated).

Note relatively abrupt transitions between (A) - (B) and (B)- (C). The same phenomena may be noted (with different proportions) in silicon.

(A) Reverse leakage at low voltage. Slope is of resistance much higher than for region (B).

(B) Low resistance region: resistance is almost constant for region ~±100 mv from 0. Slope and extent vary with type.

(C) (Forward) Logarithmic region. Above ~100 mv, conductance increases logarithmically with bias.

Failure to start leakage measurement at 0 v bias may produce leakage measurement error if drive-start point is assumed to be 0 v point. Error can be as great as the current difference between region (A) and region (C) before being noticeable, and hence leakage measurement error can be 50% or more, for a low-leakage diode. See text (K).

In the 1 µa/v position of the leakage test adapter the relation between output and -grid voltages is 1:1; moving the output level over its range of approximately ±150 mv moves the -grid by the same amount if the input Z is $>>$1 MegΩ. With a diode installed, mis-setting the DC level will start the leakage curve at bias points which may be as much as ±150 mv from 0 v. With many low-leakage types of diodes, this error in assignment of the 0 v point on the diode curve (see Fig. 1, T13G, and Fig. 14, representing diode characteristics near zero volts) may cause a leakage measurement error of 50% or more, because of the steep characteristic of the leakage near zero volts (imagine moving the zero volts point back and forth between the dotted lines of range B , in order to visualize the current level where the sweep start "foot" would appear on the display.

Here is the one saving feature:

<u>Regardless of how the output DC level may be mis-set, the zero current reference obtained by removing the diode under test is correct within 300 picoamperes if the O-Unit is properly calibrated</u>.

Thus, regardless of the location of the "foot" of the display, the leakage <u>as measured from zero current reference</u> is correct, though the zero voltage point may be off by ±150 mv from actual 0 at the left of the display.

In the 1 nA/v position, the situation is a little different. The -grid moves only 1/20 the amount that the output DC level moves, so an output DC level setting error of 200 mv will make the zero volt reference incorrect by only 10 mv.

However, in the case of very low leakage silicon diodes, we find a region of relatively high conductivity similar to that seen in germanium, and failure to start the drive at exactly 0 v potential difference between the cathode of the diode and the -grid of the leakage test adapter can produce an error in leakage measurement, even though the possible zero-setting error is only a few millivolts.

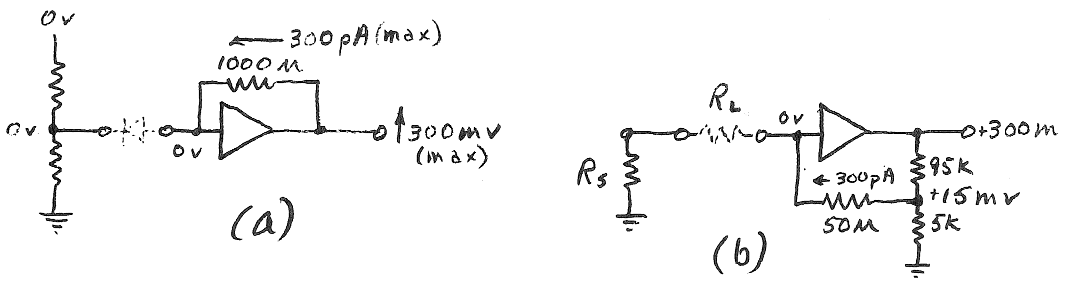


FIG. 15. EFFECT OF O-UNIT GRID CURRENT AT 1 nA/v.

(a) Equivalent circuit for 1 nA/v sensitivity. With Output DC Level correctly set, 300 pA grid current will produce 300 mv output, but -grid will remain at 0 v, assuring zero bias on diode under test at start of display.

(b) Actual circuit used. If output DC level is changed, only 1/20 of this voltage change will appear at input. If DC level set is used to buck out grid current, however, -grid will shift by 1/20 of the output voltage change and may affect leakage reading.

Effective input R of -grid in both cases is 400 k. If output DC level is correctly set, shunting -grid to ground with $R_L + R_S = 400$ k should return output just halfway to 0 v.

In the 1 nA/v sensitivity position, the zeroing problem is complicated by the fact that the current indication for zero current from the diode will differ from the trace position representing optimum setting of the DC level adjustment, by the amount of the grid current (<300 pA if the O-Unit is properly adjusted). This difference should not be bucked out with the DC Level adjustment. If the DC level adjustment is properly made using the DC Level Adj. switch, the -grid of the operational amplifier will be kept at 0 v when there is no external current applied via the test adapter, and the output only will move to supply the grid current (see Fig. 15 a). The output reading under these circumstances, then, will be displaced from the level-set point by the amount of the grid-current, but will be the correct "no-diode-current" reference.

For setting up high sensitivity measurements then, always proceed as follows:

(1) Using the front-panel switch, set the Output DC Level adjustment to the preamp "zero check" level, using 0.2 v/cm or better preamp sensitivity.

(2) With the adapter empty and set to 1 nA/v, use the position control to put the "no current" reference trace to a convenient point on the graticule.

(3) If the "foot" of the display with a diode installed is displaced from the zero reference, try increasing holdoff between sweeps (e.g., go to single sweep, manually triggered sweeps, or delayed sweep), and double-check steps 1 and 2.

(4) If the foot of the display remains offset from the zero-current reference, there are two possible causes:

(a) If the foot is offset in the direction away from the zero-check point on the preamp, try removing the cable from the sawtooth out connector, leaving the diode in place. Change of zero level probably indicates that the sawtooth divider tap connected to the cathode of the diode under test is not at zero volts at the start of the sweep. (Check for proper operation of the sweep and sawtooth output circuits, or for a defective disconnect diode in the adapter).

(b) If the foot is offset in the direction toward the preamp zero-check, the above may be true, or the device under test may have an extremely high conductivity region near zero volts.

A resistance of 400 k would cause the output DC level to move halfway from its quiescent (zero current input) level to the zero check point (zero ohms would move it to the check point). Such a conductivity as that of a 400 k resistance would produce a leakage current of 2.5 nA per millivolt drive. A resistance of 4 M (250 pA per mv drive) would move the zero 10% of the way to the zero-check point. It's improbable that conventional semiconductors having this type of characteristic near 0 v would require nanoampere sensitivities for leakage observations at 1 v to 10 v bias.

--o--

GG/cmh

Specification 112

September 18, 1964

ENGINEERING INSTRUMENT SPECIFICATION

LEAKAGE CURRENT ADAPTER

FOR TYPE O UNIT

FOR INTERNAL USE ONLY
TEKTRONIX, INC.

REVISED APRIL 6, 1965

Please make the following addition to your copy (or copies) of the Instrument Specification No. 112 for the Leakage Current Adapter.

On page 4:

Characteristic	Performance Requirement
Maximum Internal Leakage	50 picoamps at 100 volts

In Section 3, Electrical Test Methods

3.3 Maximum Internal Leakage

Set the Type O unit Volts/cm switch to .05 and the oscilloscope Time/cm switch to .1 sec. Set the Adapter Horiz Volts/cm switch to 10, and the µA/V-nA/V switch to nA/V. Any deviation from a straight horizontal trace indicates leakage within the Adapter. This deviation must not exceed 1 cm (50 picoamps) at 100 volts.

Specification 112

September 18, 1964

ENGINEERING

INSTRUMENT SPECIFICATION

LEAKAGE CURRENT ADAPTER

FOR TYPE O UNIT

Prepared By:

Technical Writing Staff

Preproduction Engineering

Approved By:

Project Engineer Robert M Johnson Rob't. M. Johnson

Project Engineer Duane Romine Duane Romine

Evaluation Engineer Gordon Mery Gordon Mery

Evaluation Engineer Richard Wilson Rick Wilson

FOR INTERNAL USE ONLY

TEKTRONIX, INC.

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INTRODUCTION

This is the Instrument Specification for the Leakage Current Adapter (013-086) for the Type O Unit, and is the reference document for all company activity concerning performance requirements.

This Instrument Specification is for INTERNAL USE only.

General Information

The Leakage Current Adapter is a Type O Unit accessory for measuring reverse leakage in semiconductor diodes. In addition to accepting regular axial-lead diodes, the adapter will also accept small-signal transistors for measuring reverse leakage in a diode-connected configuration. The diode clips and the transistor socket are completely shielded to minimize leakage capacitance.

Diodes are easily loaded by placing them in a notched retainer in the swing-down cover. When closed, the cover places the diode leads against spring clips. The swing-down cover must be left in the open position when measuring leakage in transistors. Tweezers may be required to load long-lead transistors in the socket and to remove short-lead transistors from the socket.

The four-pin transistor socket accepts NPN and PNP small-signal transistors. The C-B-E pins in the socket are marked in black for NPN transistors and in red for PNP. The emitter and base leads of the socket are tied together. If all three transistor leads are inserted in the socket, the crt display will indicate reverse leakage from collector to base + emitter. Collector to base only or collector to emitter only leakage will be displayed if the undesired transistor lead is not inserted in the socket.

A positive-going sawtooth drive voltage is required at the IN jack on the adapter. The Sawtooth Out (or Sweep Out) jack on most oscilloscopes that will accept the Type O Unit is a convenient source of this voltage. A 100-volt sawtooth is required at the diode cathode clip to establish the horizontal deflection factor and obtain a direct calibration from the HORIZ VOLTS/CM switch. (The diode cathode clip is the left-side clip, and is indicated as Test Point on the schematic.) Sawtooth voltages in excess of 100 volts (200 volts ± 20 volts* maximum) may be set to 100 volts at the Test Point with the PRESET adjustment. Sawtooth voltages less than 100 volts may be set to some desired percentage of 100 volts; the HORIZ VOLTS/CM switch setting must then be multiplied by this percentage to obtain the proper horizontal deflection factor.

* PRESET is a 20% pot.

The setting of the Type O Unit VOLTS/CM switch is directly translated to the same number of µamps/cm or namps/cm, depending on the setting of the µA/V-nA/V switch on the adapter. The resultant crt display is therefore a plot of leakage current vertically against reverse bias voltage horizontally. See Fig. 1.

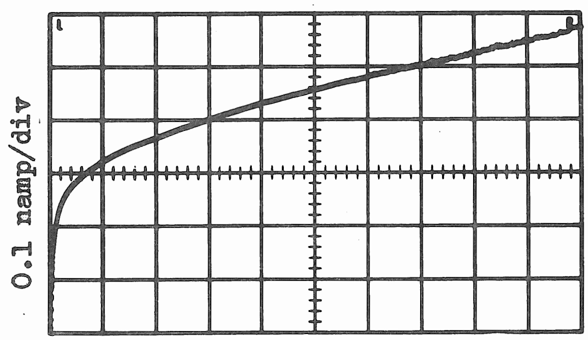

Fig. 1. Reverse leakage in 1N999 diode. At 25 volts reverse bias, for example, the leakage is ≈ 0.455 namps.

The adapter may also be used to provide a crt display of resistance values as high as 10^{12} ohms. Figs. 2 and 3 illustrate the use of the adapter to calculate unknown resistance values.

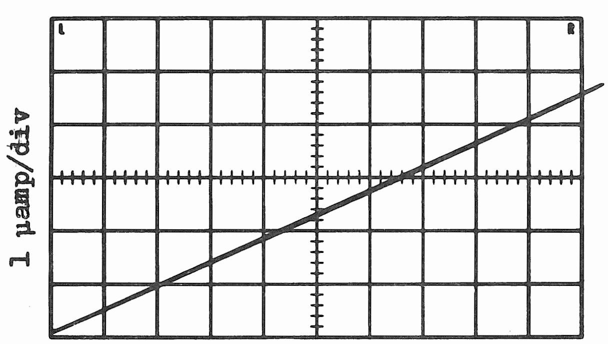

Fig. 2. Current vs. voltage in unknown resistance. Current ≈ 4.6 µa @ 50 volts, corresponding to ≈ 10.87×10^6 ohms.

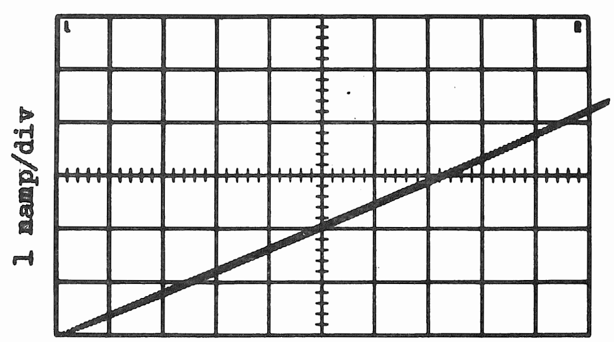

Fig. 3. Current vs. voltage in unknown resistance. Current ≈ 4.2 na @ 50 volts, corresponding to ≈ 11.9×10^9 ohms.

Characteristics Summary (Performance figures are typical)

Vertical Deflection Factors	1 µamp/volt and 1 namp/volt
Horizontal Deflection Factors	1, 2, 5, and 10 volts/cm for 100-volt sawtooth at TEST POINT
Maximum Sweep Rate	1 msec/cm at µA/V 0.1 sec/cm at nA/V
Maximum Input Sawtooth Voltage	200 volts

SECTION 1

1.0 Performance Requirements

1.1 Electrical Characteristics

The following electrical characteristics are valid throughout the environment specified in Section 1.2. Tests and measurements are performed according to Sections 3 and 4.

Characteristic	Performance Requirement	Supplemental Information
Vertical Deflection Factors	1 µamp/volt or 1 namp/volt, ±8%	
Horizontal Deflection Factors		
For 100-volt sawtooth at Test Point	1, 2, 5, or 10 volts/cm, ±2%	
For < 100-volt sawtooth at Test Point	1, 2, 5, or 10 volts/cm × % of 100-volt sawtooth at Test Point	PRESET adjusted for desired % of 100 volts at Test Point
Maximum Sweep Rate		Faster sweeps less accurate due to junction capacitance of device under test
At µA/V	1 msec/cm	
At nA/V	0.1 sec/cm	
Maximum Input Drive Voltage	200 volts	

1.2 Environmental Characteristics

The Leakage Current Adapter is a laboratory instrument. Only the following environmental limits are applicable.

1.2.1 Storage

No visible damage or electrical malfunction after storage at -40°C to +65°C, as described in Section 4.1.

1.2.2 Humidity

The Adapter will perform to limits indicated in Section 1.1 following humidity tests described in Section 4.2.

1.2.3 Vibration

The Adapter will perform to limits indicated in Section 1.1 following vibration tests described in Section 4.3.

1.2.4 Shock

The Adapter will perform to limits indicated in Section 1.1 following shock tests described in Section 4.4.

1.2.5 Transportation

The Adapter will be packaged so that it will meet the requirements of the National Safe Transit specifications described in Section 4.5.

Section 2

2.0 Miscellaneous Information

2.1 Finish

Casting is Zamak #5 finished in a blue vinyl paint; panel is anodized aluminum.

2.2 Overall Dimensions

See Fig. 4.

2.3 Weight

5 oz. maximum.

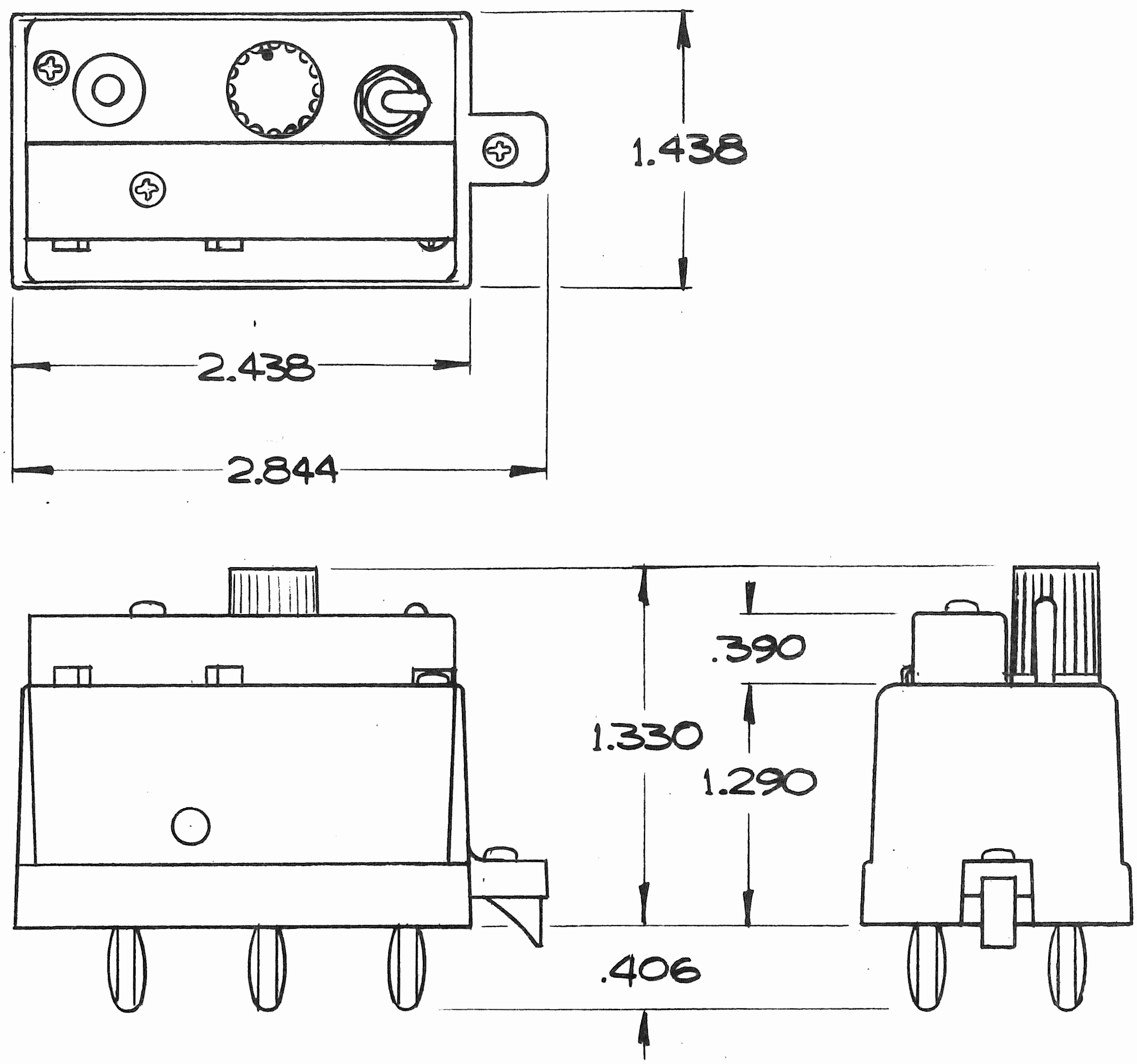

Fig. 4. Nominal overall dimensions of Leakage Current Adapter.

2.4 Connectors

_/_ IN connector is banana jack; all connectors to Type O Unit are banana plugs.

2.5 Accessories

1	Instruction Manual	070-461
1	Banana-Plug Patch Cord, 18"	012-031

Section 3

3.0 Electrical Test Methods

NOTE: Calibration accuracy of the Leakage Current Adapter is dependent on the vertical gain accuracy of the Type O Unit. To check the O Unit gain, set the VOLTS/CM switch to 20, VARIABLE to CALIBRATED, and VERTICAL DISPLAY to DC, +. Connect a patch cord between the oscilloscope CAL OUT connector and the O Unit EXT. INPUT connector.

If the crt has a 4-cm (vertical) graticule, set the Calibrator for a 50-volt pulse and check for exactly 2½ cm of vertical deflection. If the crt has a 6-cm graticule, set the Calibrator for a 100-volt pulse and check for exactly 5 cm of vertical deflection. If the vertical deflection is not that indicated, adjust the O Unit GAIN ADJ. control for the correct deflection.

3.1 Horizontal Deflection Factor

Be sure the Type O Unit gain has been checked as indicated in the note above. Set the O Unit VOLTS/CM control to 2 (leave all other controls as indicated in the above note) and remove the patch cord between the EXT. INPUT and CAL OUT connectors. Install the Leakage Current Adapter in the desired O Unit channel (A or B), and set the HORIZ VOLTS/CM switch to 10. Connect the long patch cord (with banana plugs) between the oscilloscope Sawtooth Out (or Sweep Out) connector and the _/_ IN connector on the adapter. Then connect a 10X probe to the O Unit EXT. INPUT connector, and touch the probe tip to the left-side diode clip (diode cathode clip, labeled Test Point on schematic). CAUTION: Do not clamp the probe tip to the diode clip; the diode clip cannot support the probe weight. Adjust the Vertical and Horizontal positioning controls so that the trace starts exactly in the lower left corner of the graticule.

If the oscilloscope has a 4-cm graticule, set the adapter _/_ PRESET adjustment so that the vertical deflection is exactly 4 cm* at the 8th horizontal graticule mark, as shown in Fig. 5. If the oscilloscope has a 6-cm graticule, set the _/_ PRESET adjustment so that the vertical deflection is exactly 5 cm* at the 10th horizontal graticule mark, as shown in Fig. 6.

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* See NOTE on page 8.

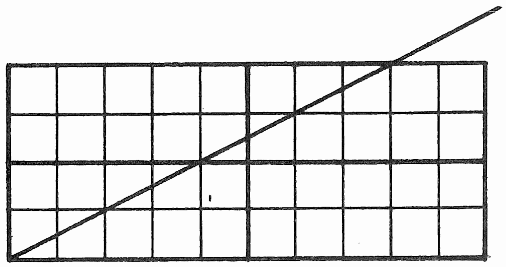

Fig. 5. Correct setting of _/_ PRESET adjustment for 4-cm graticule.

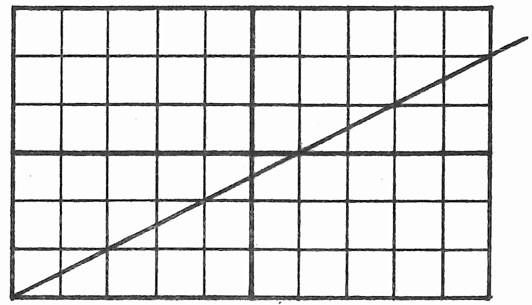

Fig. 6. Correct setting of _/_ PRESET adjustment for 6-cm graticule.

After setting the horizontal deflection factor at 10 volts/cm, switch the HORIZ VOLTS/CM switch to 5, 2, and 1. Check the attenuation at each step for 5% accuracy (2% for the adapter, and 3% for the O Unit).

3.2 Vertical Deflection Factors

Vertical deflection factors are checked by measuring the current through resistances of known values (bridged) at an exact voltage. For example, if the Q Unit VOLTS/CM switch is set to 10, VARIABLE to CALIBRATED, and the adapter µA/V-nA/V switch set to µA/V, a 10-megohm resistance should produce exactly 1 cm of vertical deflection at 100 volts (current = 10 µa). With the adapter switch at nA/V, a 10-kilomegohm resistance should produce exactly 1 cm of vertical deflection at 100 volts (current = 10 na). The tolerance of the resistance values must be added to the nominal 5% accuracy specified for the adapter/O Unit combination.

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* NOTE: If the amplitude of the oscilloscope sawtooth is less than 100 volts, it will not be possible to obtain the required 4 or 5 cm of vertical deflection at the proper horizontal graticule mark. In this event, set the _/_ PRESET adjustment so that the vertical deflection at the indicated horizontal graticule mark is some exact percentage of 4 or 5 cm. The HORIZ VOLTS/CM switch setting must then be multiplied by this percentage to obtain the correct horizontal deflection factor.

Section 4

4.0 Environmental Test Methods

4.1 Temperature

4.1.1 Storage

Store for 4 hours at -40°C and 4 hours at +65°C, one cycle only.

4.1.2 Failure Criteria

The Adapter must meet performance requirements before and after storage. Cracking, warping, and significant color discoloration which interferes with normal function will not be permitted.

4.2 Humidity

4.2.1 Nonoperating

Perform two cycles (48 hours) of Mil-Std-202B, Method 160A. Delete freezing and vibration requirements. Humidity 90-98% at +25°C to +65°C. The Adapter will then be permitted to dry for 24 hours at room ambient before conducting electrical checks.

4.2.2 Failure Criteria

The Adapter must meet performance requirements before and after humidity tests. There shall be no significant deterioration of components, materials, or finishes. Deformation which interferes with normal mechanical function will not be permitted.

4.3 Vibration

4.3.1 Nonoperating

Perform resonant searches along all 3 axes at .015" ±.003" total displacement from 10 to 55 c/s (2.3g at 55 c/s). All major resonances should be above 55 c/s. The swing-down cover shall be taped down during this test.

4.3.2 Failure Criteria

The Adapter must meet performance requirements before and after vibration tests. There shall be no broken leads, chassis, or other components, loose parts, excessive wear, or component fatigue. Deformation which interferes with normal mechanical function will not be permitted.

4.4 Shock

4.4.1 Nonoperating

Subject the Adapter to a "guillotine" type shock of 30g, ½ sine, 11 msec nominal duration. One shock to be applied in each direction along each major axis for a total of 6 shocks.

4.4.2 Failure Criteria

The Adapter must meet performance requirements before and after shock tests. Cracked or broken chassis, components, or leads, or deformed chassis or components on the order of 0.100", or deformation which interferes with normal mechanical function, will not be permitted.

4.5 Transportation

The Adapter when packaged must meet the National Safe Transit type of tests.

4.5.1 Package Shake Test

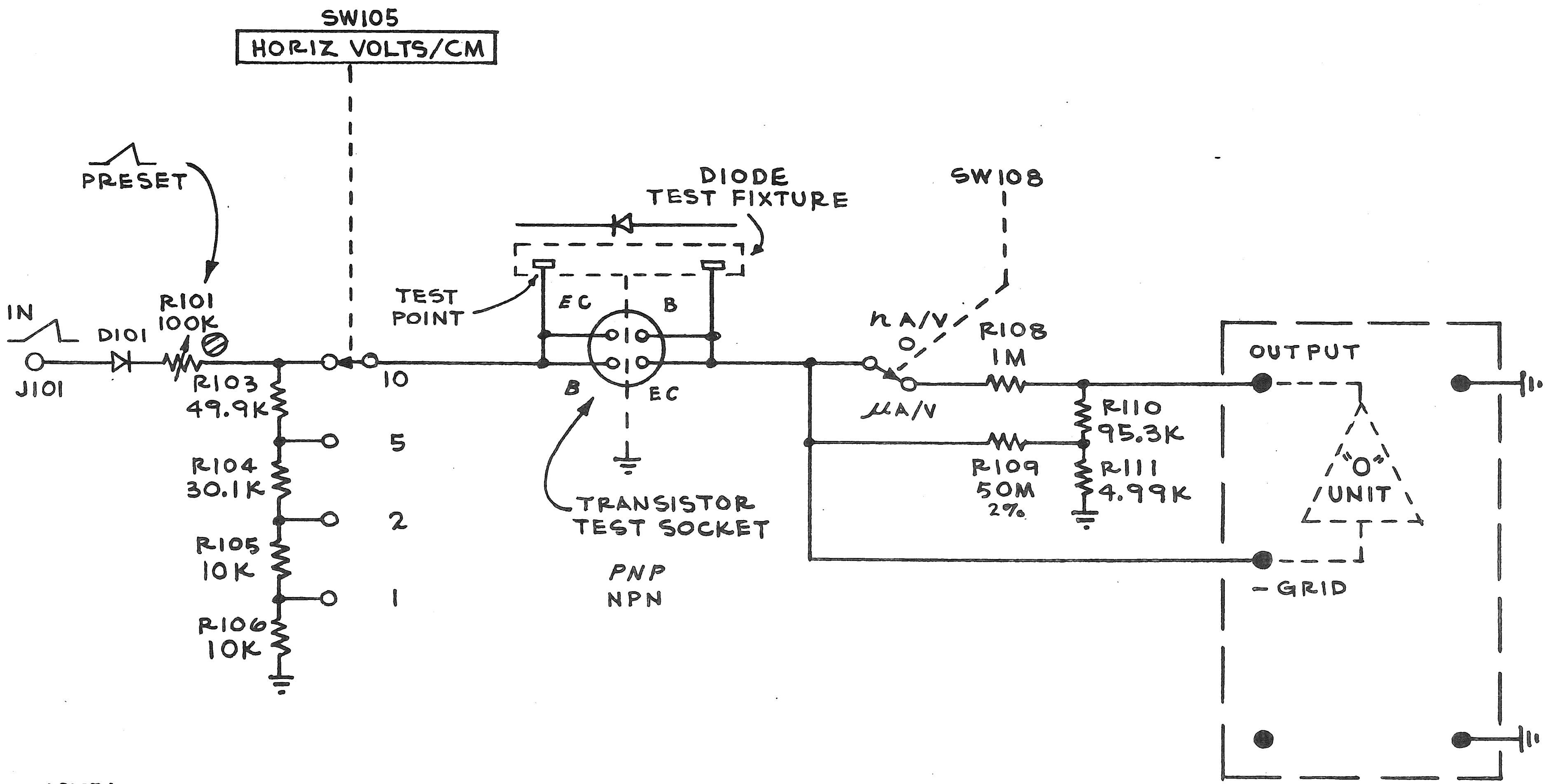
One hour on the vibration table with an amplitude slightly in excess of 1g which just causes the package to leave the vibration platform.

4.5.2 Package Drop Test

Drop from a height of 30" on all corners, edges, and flat surfaces.

4.5.3 Failure Criteria

The Adapter must meet performance requirements before and after the transportation tests. There must be no serious damage such as broken leads, chassis, or components. Deformation which interferes with normal operation will not be permitted.



NOTE:
ALL RESISTORS 1%
EXCEPT AS NOTED.

GAB
8-12-64

LEAKAGE CURRENT ADAPTER

ENGINEERING

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SPECIFICATIONS

GRID CURRENT

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Specs = .5 namp max; adjustable to .3 (-), .15 (+)

We've changed final specs for O unit grid current to "less than .5 namp; internally adjustable to less than .3 namp at the -GRID and less than .15 namp at the +GRID" for two reasons:

1. Grid current tends to vary slightly (±100 pamp in some cases) over 50 to 100 hours operation, and also may vary when the O unit is plugged into different scopes or when it is shipped. A little leeway between test specs and advertised specs means less chance of customer rejects based on picoamp fly specking.

2. There isn't much possibility that the test spec (.3 namp and .5 namp) could be tightened and still provide a reasonable tube yield.

Tube brands

Recent evaluation of tubes of various brands aged 100 hours indicates we can get a good yield of 12AU6's to meet grid current specs from the following brands:

Westinghouse (95%)
Raytheon
GE (passive cathode)
Nippon
Brimar (British)

Evaluation of Telefunken and RCA 12AU6's showed poor yield, even after aging. We're setting up a 157 selected tube number for matched pairs meeting grid current requirements. In the meantime, the plant is selecting Westinghouse 157-050's.

Grid current adjustment mod at sn155

Mod 5728-O, effective sn155, adds a grid current adjustment pot and aids considerably in getting a good yield of tubes. It also helps the sharp-type customer adjust grid current to extremely low (but less stable) values for critical low level applications.

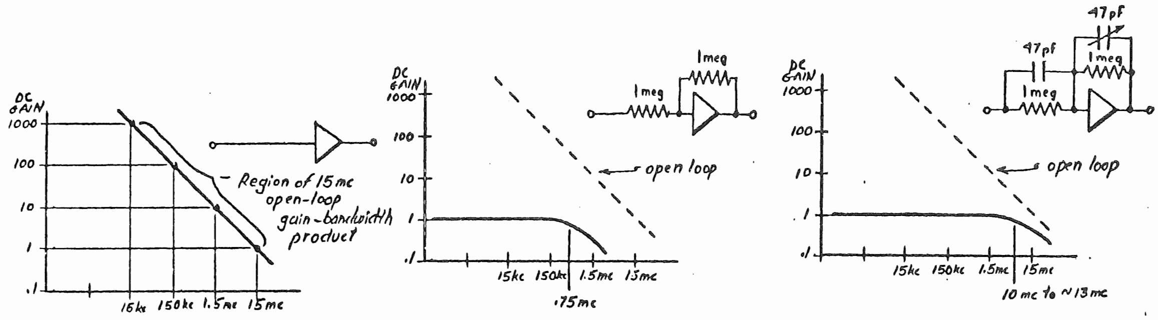
PERFORMANCE

BANDWIDTH LIMITATIONS

FEN 1-12-62

The specifications of a "15 Mc gain-bandwidth pro-duce" for the O Unit operational amplifiers has apparently been quite confusing. This specification is for *open-loop gain-bandwidth product* -- the frequency at which the nominal DC open-loop gain of 2500 drops to unity....and is *not* an indication of conventional 3 db bandwidth.

Using internal switched components, we obtain a "unity gain" 3 db bandwidth greater than 750 kc. The actual value varies somewhat depending on what pairs of equal resistors are used. Using two 10 k resistors the 3 db unity gain bandwidth may be as high as 1.5 Mc.



For those customers requiring maximum gain-band width, we hope to make a capacitive compensation adapter board. This should allow us to extend the closed-loop unity gain to above 10 Mc. With internal switched components alone, this wide a band is not possible due to switching difficulties.

Customers should not be led to expect many-mega-cycle miracles from the instrument without external compensation. While the O Unit's operational ampli-fiers are a significant improvement over those generally available and are certainly valuable tools for many, many applications, we shouldn't allow the customer to oversell himself by misinterpretation of bandwidth specs.

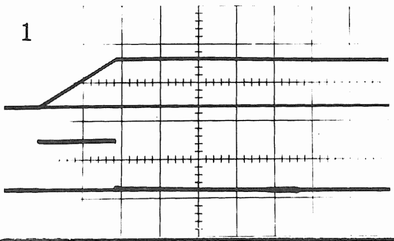
O UNIT INTEGRATING PROBLEMS

Jim Griffin, Geoff Gass

Jim Griffin to Geoff Gass, 11-18-64

The attached series of photos and data is self-explanatory. We noticed a definite ringing when using one time constant with one RC network; but if we keep the same time constant and change the RC values, the ringing disappears. Can you explain?

1

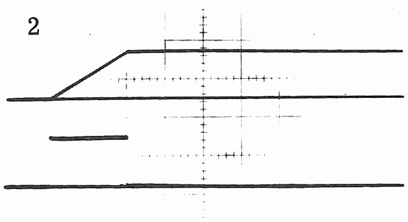


Time/cm 5μsec/cm
Z_i 10k
Z_f .001

RC = 10μsec

Integral 2v/cm
Pulse 2v/cm

2

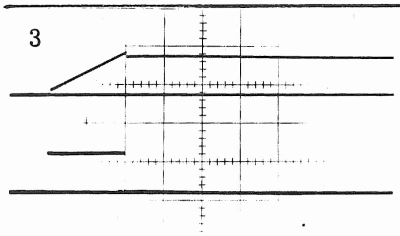


Time/cm 5μsec/cm
Z_i 100k
Z_f .001

RC = 100μsec

Integral 0.2v/cm
Pulse 2v/cm

3

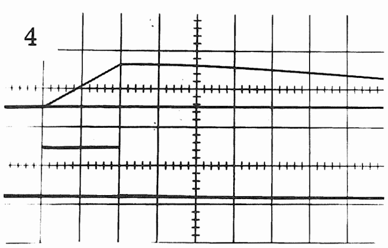


Time/cm 5μsec/cm
Z_i 1M
Z_f .001

RC = 1msec

Integral .05v/cm
Pulse 5v/cm

4

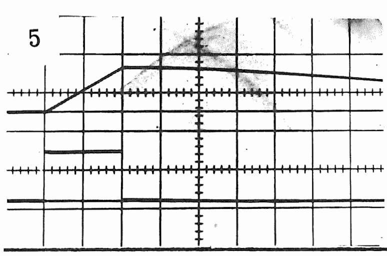


Time/cm 5μsec/cm
Z_i 10k
Z_f 100pf

RC = 1μsec

Integral 5v/cm
Pulse 2v/cm

5

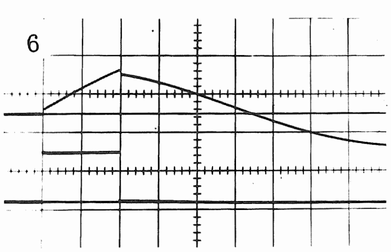


Time/cm 5μsec/cm
Z_i 100k
Z_f 100pf

RC = 10μsec

Integral 2v/cm
Pulse 2v/cm

6

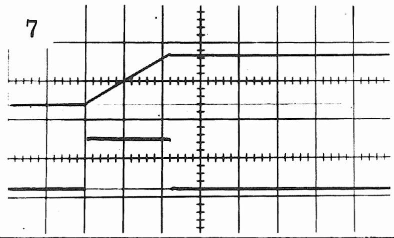


Time/cm 5μsec/cm
Z_i 1M
Z_f 10pf

RC = 10μsec

Integral 2v/cm
Pulse 2v/cm

7

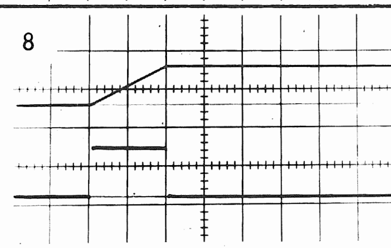


Time/cm 0.5μsec/cm
Z_i 10k
Z_f 100pf

RC = 1μsec

Integral 2v/cm
Pulse 2v/cm

8

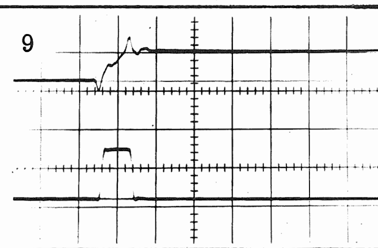


Time/cm 0.5μsec/cm
Z_i 10k
Z_f 10pf

RC = 0.1μsec

Integral 20v/cm
Pulse 2v/cm

9

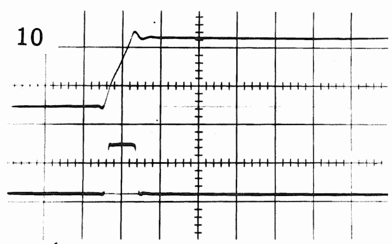


Time/cm 0.2μsec/cm
Z_i 10k
Z_f .001

RC = 10μsec

Integral .05v/cm
Pulse 2v/cm

10

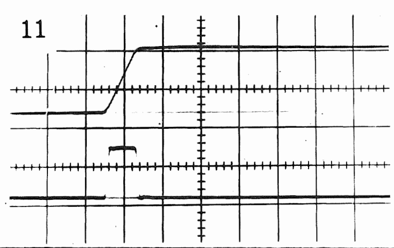


Time/cm 0.2μsec/cm
Z_i 10k
Z_f 100pf

RC = 1μsec

Integral 0.2v/cm
Pulse 2v/cm

11



Time/cm 2μsec/cm
Z_i 10k
Z_f 10pf

RC = 0.1μsec

Integral 2v/cm
Pulse 2v/cm

Inter-Office Communication

BEAVERTON

To: Jim Griffin,
Union

Date: December 2, 1964

From: Geoff Gass,
Product Information Liaison

Subject: O-Unit Integrating Problems
Your IOC 11-18

Dear Jim:

Thanks much for the photos and complete problem report. Helps lots in pinning things down.

On your No. 1 problem (Photo No. 9) -- a ratty looking integral of a 2.6 v x 0.16 µsec pulse (0.416 µv-sec) -- this is about 1% feed-through of the leading and trailing edges of the input waveform to the output. In photo No. 9, your time-constants have you shooting for a virtual gain (Integrating interval ÷ RC) of 0.016 -- that is, you're counting on the integrator to attenuate the average voltage amplitude of your particular signal by a factor of 60. When the answer is only 1.6% of the input, 1% feedthrough looks pretty horrible. The solution is to use a shorter time-constant and get your virtual gain back closer to 1 where 1% feed-through of many-megacycle information is not such a large part of the answer.

The dimensions of your pulse for photo No. 9 came out 416 nanovolt-seconds. Divide this by 10 µsec (the time constant used) and you get 41.6 millivolts as the answer you should have gotten. The answer you actually have there (after the pulse quits bashing things around) seems to be 40 millivolts, and that's certainly reasonable, considering the eyeball measurements I took on your original pulse.

As noted in the Feb.-April '63 Service Scope article (we'll refer to this again later -- might as well dig it out) the O-Unit is pre-compensated for these little difficulties, so even though things may be lumpy while you're taking a fast integral, as soon as you quit kicking the poor little fella, he does hand you the right answer, although during the kicking, his responses may be somewhat garbled.

Ultimately, Jim, the problem is this: The O-Unit has a finite transit time, which says the output does not develop an <u>immediate</u> response to any input, but one delayed in time by a few nanoseconds. The O-Unit also has a non-zero risetime, so even after it <u>starts</u> responding to an input, there's some time before all of its muscles come into play. (Maybe I've said the same thing twice here). The ultimate result is, that the input drives the output via the series combination of Z_i and Z_f for a little bit before the amplifier starts driving the output <u>in the opposite direction</u>. So you see about 1% preshoot when you slam the integrator with a fast-rise pulse.

In sine-wave terms, we say the gain-bandwidth product of the op-amp is 15 mc, but this does not mean that we maintain an attenuation of precisely 1000 at 15 Gc. As a matter of fact, the open-loop gain-of-one point may not make 15 mc. The gain-bandwidth product is primarily useful back below 1/3 the frequency at which the gain is 1.

Now that takes care of the cases where there is a "pre-shoot" (wrong direction) sort of step at the start of an integral.

How about the case where you get a "right-direction" step at the start of the integral (photos Nos. 3 and 6)? This is end-to-end capacitance and/or wiring strays through and around the input resistor. In other words, you never really have a "pure" resistor there. A Welwyn or TI 1 M resistor has maybe 0.6 pf end-to-end capacitance.

If the current through this capacitance (due to dE_{in}/dt) is any significant fraction of the current that flows through the resistor (due to E_{in}), it will come out as error in the answer during the time that the capacitive current flows.

Let's take the specifics of photo No. 3. The input pulse is 5 v x 10 µsec = 50 µv-sec. It jumps those 5 v in maybe 20 nsec, for a rate-of-change 250×10^6 v/sec (aren't you glad it doesn't keep rising for a second?). The current that 5 v should be pumping through 1 Meg is 5 µa. The current that 250×10^6 v/sec should pump through 6×10^{-13} farads ($Q = It = CE$ and all that) is $6 \times 10^{-13} \times 250 \times 10^6 = 1500 \times 10^{-7}$ or 150 µa during the risetime of the pulse. About 3000% error for a little while, there. However, since the pulse being integrated is about 500 risetimes long, the output capacitor is pumped up much farther by the 10 µsec duration of the 5 µa resistor current than by the ~20 nsec duration of the ~150 µa current from the stray capacitance. The 1000 pf Z_f capacitor should be pumped up about 3 mv by the hypothetical 150 µa x 20 nsec = 3 picoamp-second displacement current.

In photo No. 3 we can see that the actual step is about 1.4 mm corresponding to an error of 7 picoampere-seconds in 1000 pf. This means the actual strays involved in the switches and wiring, and between the input and -grid connectors, and also to the -grid connectors from external radiation sources amounts to an effective value of about 1.4 pf. (An error in guessing the pulse risetime doesn't matter here, 'cause we used it twice and effectively canceled it out.)

It's easier to calculate than by this roundabout means, though. Provided that the dE_{in}/dt is great enough so that during a "step" type transition the displacement current through the strays is going to be much, much greater than the current running where it's supposed to through the resistor, the voltage output as the result of the step will be the voltage of the step, times C_i/C_f, where C_i is the effective input stray capacitance existing between the signal source and the -grid (capacitance to ground doesn't count). To turn it around, then, the input stray capacitance is $C_i = E_oC_f/E_{in}$. In this case, you put in an input step of 5 v and got out 7 mv, so $C_i = 7 \times 10^{-3} \times 10^{-9} \div 5 = 1.4$ pf (effective). If you started out with a 50 v step having the same risetime, the value of E_o would have been easier to measure and C_i would be more accurately calculated in case anybody really cares.

The step in the opposite direction at the end of the ramp, caused by the dE_{in}/dt of the fall of the input step, corrects the final answer, so all is not lost. But if the customer wanted his integral at sometime other than when the input was resting quietly at zero volts, his answer would be in error by 7 microvolt-seconds. Not so hot, especially when you're working with waveforms other than neat 5v x 10 µsec rectangular pulses (which can be eyeball-integrated anyway and don't need an O-Unit).

What to do?

Well, the O-Unit manual says, stick around virtual gains of 1. In this case (photo No. 3) the setup was aiming for a virtual gain (integrating interval ÷ RC) of .01. However, in moving to photo No. 6, holding the megohm and changing the C from 10^{-9} to 10^{-11}, and thus shooting for a virtual gain of 1, the step looks much worse in absolute value (up from 7 mv to around 200 mv even though the input has been halved) and is not much improved in terms of percentage of the final answer.

The answer to this specific problem is not in shuffling the feedback <u>capacitor</u> to change the virtual gain, but in this case, reducing the value of the input <u>resistor</u>. Since we're stuck with a certain amount of stray capacitance coupling a certain amount of current from the input connector to the -grid every time we have a fast transition in the input signal, about the only thing we have left to do is increase the "right-answer" current (through the resistor) so that the "error" current through the strays won't be so large in proportion. Increasing the size of the signal won't help, because the error currents will increase just as fast as the current through the resistor.

The only practical solution is to reduce the size of the resistor.

If the best answers are to be obtained around a gain of one, then for a 10 µsec integrating interval, we should use an RC of 10 µsec.

If we have 1.4 pf C_i that we're stuck with, and we have a 5 v input step, a 1.4 pf Z_f will give us a 5 v output error (~100% error if the right answer is 5 v). A 14 pf Z_f will give us an 0.5 v output step (~10%). A Z_f of 140 pf will give us an output step of 50 mv (~1%). That sounds better. A 100 pf feedback cap will give us an output step of 70 mv, or 1.4% error if the right answer is 5 v. So let's go with 100 pf and 10 µsec -- which just naturally leads us to 100 k for the input resistor. Try it out and -- voilà! -- photo No. 5 shows that we get the right answer during the entire integrating interval and for another 5 µsec to boot. No step.

You can approach this logically another way by just saying, to increase the "right answer" current through Z_i for a given voltage input, we want to reduce the resistance of R_i, so instead of having 350 µa x 20 nsec error from the capacitance lousing up 5 µa x 10 µsec through the resistor, let it try to louse up 50 µa x 10 µsec. Well, however you like to angle it, the answer comes out the same. No step.

This is mentioned -- though not in as verbose a fashion -- in the Service Scope article, Part 2, April '63 issue. You <u>can</u> sometimes reduce the step effect by using an external Z_i, switching the ±grid switch to + and grounding the ±grid connector externally, thus eliminating one front-panel "antenna" connection to the -grid.

But hold on, you say. Photograph No. 4 may have solved the <u>step</u> problem of photo No. 3, but 4 and 5 are showing up a new problem, and that problem is getting really rotten in photo No. 6 which has the step, to boot!

Jim Griffin,
Union

December 2, 1964

Well, that brings up the third factor that gets into the picture, and that's the LF Reject circuit. (Please do not smite your forehead like that, Jim, you could hurt yourself. The O-Unit does have ingredients other than limitations, problems, errors, diddle-factors and funnies. It's just unfortunate you managed to overturn so many stones at once here.)

If you scrabble about a bit in the Engineering section of the O-Unit IRB, you'll find a rambling 19-page monograph we put together last year on the mechanics and limitations of LF Reject circuits in general (why-have-'em and what-they-do), and of the two ("1 cps" and "1 kc") LF Reject circuits provided in the O-Unit in particular, and instructions on how to build your own if you can't get the answers you want with the ones provided.

Actually, to get "universal" reject circuits, the O-Unit would need, instead of the little toggle switch, something more like the 533 sweep and mag system.....a lot of different time-constants, and several different ways of getting each one.

The two that are provided in the O-Unit are of general-purpose utility, and are optimum for a lot of applications. But page 12 of this writeup shows what kind of errors are introduced by the circuits if you try to hang onto answers too long, or to keep on integrating over too long an interval.

Let's take photo No. 5 as an example of an integral that's started showing a "droop" problem (which will turn into damped sinusoidal ringing at around 3.4kc, falling off to an amplitude of about 1 mm in another 500 μsec -- didn't it?).

The integrating interval of 10 μsec using 100 k and 100 pf is $\sim$0.1 of the "lag" (lousy terminology by operational amplifier standards -- "lag" in the feedback, i.e., delayed DC feedback, is known as "lead" in the trade, but I understand it better as lagging DC feedback) time constant provided by the 1 kc LF Reject circuit which you were using. So from Table 2, page 12, we get the basic error factor B $\approx$ .002. Plugging that into the formula above, with R_f = 200 k and C_f = 100 pf and a T_L value of 100 μsec we come up with an error at the end of 10 μsec of $\sim$1%. Not bad.

Let's run over to the 9 cm mark, where we've stretched the integrating interval out to 40 μsec (8 cm x 5 μsec/cm).

The interval/lag ratio is now about 0.4 and the error should fall between 3.5 and 20% (component tolerances) with a center value of about 12%.

The actual droop in photo No. 5 is 25%, or around twice the predicted value.

The reason for this is not well explained in the LF Reject writeup. At the top of page 11 the fact is mentioned that integrals of waveforms which start out slowly and then increase rapidly to their final value will be more accurate than those which start out quickly and then slowly sneak up to their final value. But it's not explained that the table on page 12 applies only to the "middle" case -- where the integral increases at a fairly constant rate during the entire integrating interval (looking time, time before the value is measured).

Jim Griffin, Union

December 2, 1964

In the case of photo No. 5, we have a worst-case condition, where the integral increased for 10 μsec and then did substantially nothing for the next 30 μsec. The source of the error signal (i.e., output voltage) that will be coming back to the input via the LF Reject circuit is there in the first 10 μsec. For the remainder of the 30 μsec, there's nowhere to go but wrong.

If the integral had continued to increase for the full 40 μsec, the output would have been 2.4 v (1.2 cm) per time-constant for a final value of 9.6 v, minus the predicted error of 12-13% (assuming the LF Reject components were "right on" center values). The total error of, say 1.2 volts would have been compounded of 0 from the zero-time point, 600 mv from the +10 μsec point, maybe 400 mv from the +20 μsec area and maybe 200 mv from the +30 μsec area (actually, this is a continuous process -- I'm only lumping them up to convey an idea). The point is that after 40 μsec, the 2.4 v output that existed at +10 μsec will have pumped a total of 6 nanocoulombs back to the input from the output -- that is, it will have pulled that much charge out of the integrating capacitor -- regardless of whether thereafter the integral increased or not.

So that's how what the table says should be a 12-13% error actually is about 25% if the integral reaches its ultimate value in a quarter of the total integrating interval. Conversely, if there was practically nothing going on in the first 20 μsec, and the big contribution came in the last 20 μsec, then the error at 40 μsec would have been 25% of what was there at 10 μsec (like, hardly anything) plus maybe 7% of what was there at 20 μsec plus 2% of what was there at 30 μsec, and the overall total would be much smaller than the predicted 12-13%.

Well, now. On to the ringing. Page 13 of the writeup explains why the ringing exists and what the frequency will be for LF Reject circuits in general; Table 3, page 14, gets specific for the combinations in the O-Unit. The ringing and the errors due to ringing will be aggravated whenever the period of the signal being integrated falls within the decay time (column 5) for the setup. If the period is $\ll$ ring start, you get "AC" integration.

There are several things you can do when you just can't get answers using the LF Reject circuits of the O-Unit.

(a) Go to a "partial integrator" circuit, simply using feedback resistance across the integrating capacitor and omitting the LF Reject shunt capacitor, which is the phase-shifting cause of the ringing but is also the component that lets you get maximum use out of the circuit. Alternatively, use the DC Reject circuit of Figure 9, page 17.

(b) Build an LF Reject circuit externally that's better tailored to the application. Instructions on page 15.

(c) Use the reed-relay trick on page 16, which dumps the shunt capacitor right after it's done all the good it can and before it can start doing harm.

(d) Use the gating adapter, which we also happen to sell. Tek No. 013-068. But see too the gating adapter limitations described in the O-Unit IRB and the "error" photos on page 6 of this writeup.

Jim Griffin,
Union

December 2, 1964

Two last notes.

(1) A not-very-quickie formula is given on pages 18-19 of the LF Reject writeup for picking R's and C's and integrating intervals to stay out of LF Reject errors. A universal rule for picking R_i and C_f for a given application is pretty hard to state. The RC should be picked to give a virtual gain around 1 -- say X10 to ÷ 10 -- the R should be the lowest value that doesn't louse up the signal source; the C should be one that's most compatible with the LF Reject circuit used.

(2) What are the limits when the "O" is used as an integrator? That's another tough one, because there are various limits. A single pulse which at X10 virtual gain provides the smallest usable deflection on the scope using 10 k and 10 pf would seem to be one limit. How much deflection is "usable"? A centimeter? A pulse of 1 μsec duration and 5 mv amplitude will produce 50 mv out (1 cm deflection) using the 100 nsec time-constant. That's 5 nanovolt-seconds. You'd get the same answer (assuming your O is tweaked correctly) from a 50 mv x 100 nsec pulse, or a 500 mv x 10 nsec pulse. A 5 v x 1 nsec pulse would show up the feedthrough that bugged you in photo No. 9, but the ultimate answer would be close to correct. But we're pushing another limit there -- distortion and error in the total at any time where the input isn't sitting quietly at zero.

If we try stretching it the other way, pushing for X100 virtual gain with a 500 μv x 10 μsec pulse, we start running into other limits caused by open loop gain-bandwidth limitations.

However, by using the compensating adapter and amplifying the output, I did once get a single-shot measurement of a 50 mv x 10 nsec pulse from a 109, which a 545A/L cannot reproduce at full amplitude. The integrator plus amplifier shipped it upstairs correctly, however, displaying the integral of 500 picovolt seconds one centimeter high. What actually was pumped into the O-Unit was 310,000 electrons -- i.e., 500 picovolt-seconds at the g'zinta end of 10 k. (Of course, several hundred million electrons were being pumped through a 50 Ω termination at the same time, but 310,000 were all that the O-Unit used.) The single-sweep picture was out of focus and hence barely visible; otherwise I'd send you a copy. For low frequency work, hum, noise, grid current, drift, LF Reject limits and open loop gain are the chief limitations, each coming into play in a different way.

Well, lucky you, Jim. You sure bought yourself a book on this one. Hope some of it proves useful.

Best regards,

Geoff

GG/cmh

CC - Hiro Moriyasu

CALL REPORT EXTRACTS

PERFORMANCE

USE AS AMPLIFIER

CUSTOMER
CBS Laboratories

FIELD ENGINEER
Jim Johnson

CITY AND STATE
Stamford, Connecticut

AUG 1 2 1964

MONTH-DAY-YEAR 7/28/64

CALL # 03

GROUP
M & I Group

GROUP FUNCTION
Design and Evaluation of Military Products

NAMES
Marty Kay, Engr.

PROBLEM

0 Unit discussion. 10 mv per hour drift ~~too much~~ for their application. Marty was using both channels of the 0 Unit. The second channel was used as an additional stage of amplification so that the unit could more correctly integrate the signal. Integrator alone is more accurate than with amp. ahead of it. Down below 5 nanovolt-seconds, there are some problem though. O-Unit 10pf integrator needs about 3.1 million electrons to provide a usable output, and for any kind of accuracy, they must be delivered at a rate of better than 40,000 electrons per μsec (grid current is ~2000 electrons/μsec) to the − grid. If the op-amp had ~1500mc gain bandwidth product and a 0 nsec transit time, we'd have those measurements whipped. As it is, we daren't operate the 100 nsec integrator for >1μsec (gain of 10), and the customer with the 100pV × 50 μsec or the 500μv by 10 nsec pulse (3,100 electrons in 10K) gonna hafta use a different "looker" box. In any case, best rule is integrate first, then amplify, if the 0 unit is to be used for the amplification.

agh

—CG

CUSTOMER Stelma, Inc.
FIELD ENGINEER Jim Johnson
CITY AND STATE Stamford, Connecticut
AUG 12 1964
MONTH-DAY-YEAR 7/30/64
CALL # 03
GROUP Engineering
GROUP FUNCTION Filter Design
O Engineering
NAMES Eric Aliot, Engr.

PROBLEM

Problem was checking a 90 degree phase shift (3 kc) network. They had accomplished this by integrating the output signal ~~of~~ with a Type O. I explained that since the gain of the O Unit was not infinite, true integration could not be easily performed and felt that by differentiation a more accurate 90 degrees could be obtained. On checking this we found a 3 cps variance between the integrated and differentiated results. This was far greater than Eric could stand on his filter.

Open loop gain is still 2500 at 3KC. Betcha your difference in readings was due to input Z ($=Z_i$) of op-amp loading down customer's circuit differently when $Z_i = R$ and $Z_i = C$. Phase error of O-Unit at 3KC calculates to $\ll 0.3°$ when $Z_i = .001$, $Z_F = 100K$. But if customer's signal source Z was 2.5KΩ, that would account for a 3° phase shift into .001μf that wouldn't be there into 100K or 1M. — GG

agh

4

INTRODUCTION
TO
OPERATIONAL AMPLIFIERS

SERVICE SCOPE, February and April 1963

INTRODUCTION TO OPERATIONAL AMPLIFIERS

Prepared by
Tektronix Field Information Department

Part 1.

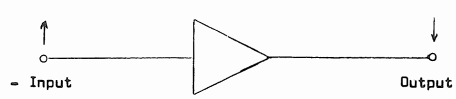
Functionally speaking, an operational amplifier is a device which, by means of negative feedback, is capable of processing a signal with a high degree of accuracy limited primarily only by the tolerances in the values of the passive elements used in the input and feedback networks.

Electronically, an operational amplifier is simply a high-gain amplifier designed to remain stable with large amounts of negative feedback from output to input.

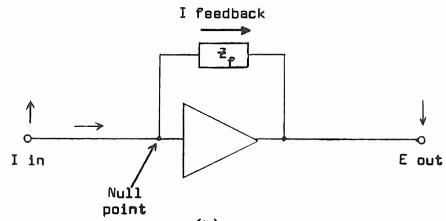
General-purpose operational amplifiers, useful for linear amplifications with precise values of gain, and for accurate integration and differentiation operations, have low output impedance and are DC-coupled, with the output DC level at ground potential.

The primary functions of the operational amplifier are achieved by means of negative feedback from the output to the input. This requires that the output be inverted (180° out of phase) with respect to the input. The conventional symbol for the operational amplifier is the triangle shown in Figure 1-a. The output is the apex of the triangle; the input is the side opposite the output. Negative feedback, through a resistor, capacitor, inductor, network or nonlinear impedance, designated "Z_f" is applied from the output to the input as shown in Figure 1-b. The input to which negative feedback is applied is generally termed "-input"* or "grid" (in the case of vacuum-tube operational amplifiers).

* The operational amplifiers of the Tektronix Type O Operational Amplifier also provide access to a non-inverting input. Uses of this "+input" or "+grid" are discussed later.



(a)



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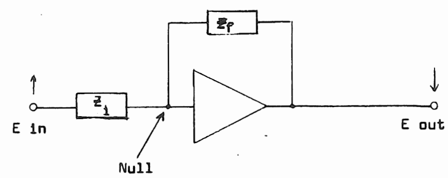


(c)

Figure 1. Conventional Operational Amplifier Symbols.

(a) The input is to the base of the triangular symbol, the output is from the apex opposite. The —input and output are out-of-phase (arrows).

(b) Feedback element Z_f provides the negative feedback to permit high-accuracy operations. The amplifier seeks a null at the input by providing feedback current through Z_f equal and opposite to the input current I_{in}. Output voltage is whatever is necessary to provide required balancing current through Z_f.

(c) Input element Z_i converts a voltage signal (E_{in}) to current, which is balanced by current through Z_f.

Operational Amplifier Seeks Voltage Null at —Input

An operational amplifier, using negative feedback, functions in the manner of a self-balancing bridge, providing through the feedback element whatever current is necessary to hold the —input at null (ground potential). See Figure 1-b. The output signal is a function of this current and the impedance of the feedback element.

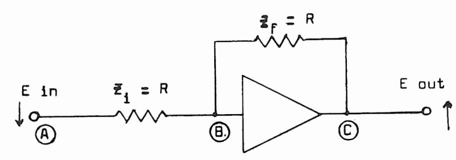
The —input, held to ground potential by the feedback current, appears as a very low impedance to any signal source. Using resistive feedback, for instance, the input appears to be the resistance of the feedback element, divided by the open-circuit gain of the operational amplifier.

If current is applied to the —input, it would tend to develop voltage across the impedance of feedback element, and move the —input away from ground potential. The output, however, swings in the opposite direction, providing current to balance the input current and hold the —input at ground. If the impedance of the feedback element is high, the output voltage must become quite high to provide enough current to balance even a small input current.

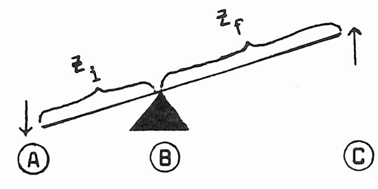
Input Element Z_i Converts Input Signal to Current

Since we more often have to deal with voltage rather than current signals, an additional element is used in most operational amplifier applications, designated "Z_i" (input impedance). This is an impedance placed in series with the —input, converting into *current* that parameter of the input signal which we want to appear as voltage at the output (Figure 1-c).

If Z_i and Z_f are both resistors (Figure 2), the operational amplifier becomes a simple voltage amplifier, the gain of which is $-Z_f/Z_i$.



(a)



(b)

Figure 2

(a) Operational amplifier using resistors for both Z_i and Z_f becomes fixed-gain linear amplifier. Gain is $\frac{-Z_f}{Z_i}$.

(b) "See-Saw" operation of operational amplifier. System appears to pivot about a fulcrum (the null point B) whose "location" is determined by Z_f/Z_i.

Let's examine the mechanism by which this works. Referring again to Figure 2, we apply a voltage to point A, causing current to flow through Z_i. Were it not for the operational amplifier, this current would also flow through Z_f and to ground through the low impedance at point C, making Z_i and Z_f a voltage divider, and raising the voltage at point B. However, the operational amplifier operates to hold the voltage at point B (the —input) at ground potential. To do this, it must supply at point C a voltage which will cause a current to flow through Z_f which will just balance the current flowing through Z_i. When point B is thus held at ground potential, the voltage across Z_i is obviously equal to the applied voltage at A.

Output Voltage is Input Current X Impedance of Z_f

The current through Z_i is equal to the applied voltage at A divided by the impedance (in this case, resistance) of Z_i, or E_{in}/Z_i. This same value of current must flow through Z_f in order to keep point B at ground. The voltage at point C, then, must be E_{in}/Z_i (which is the value of the current in Z_f) multiplied by Z_f. The output is inverted (of opposite polarity) from the input, so we say that $E_{out} = (-E_{in})\left(\frac{Z_f}{Z_i}\right)$, and the voltage gain of this amplifier configuration is seen to be $\frac{-Z_f}{Z_i}$.

See-Saw Operation

As indicated in Figure 2-b, the operational amplifier with resistive input and feedback elements acts in see-saw fashion, the amplifier moving the output end of the see-saw in response to any motion of the input end, causing the system to pivot about an imaginary fulcrum, which is the "sensing point" (—input). The distance from the near end to the sensing point or fulcrum corresponds to the Z_i or input resistor, and the distance from the fulcrum to the far end corresponds to Z_f. The motion of the far end depends on the motion of the near end and the ratio of the two distances. This analogy suggests that the operational amplifier may be used to solve dynamic problems in mechanical engineering, and so it can. One of the principal uses of operational amplifiers has been in the rapid solution of complex mechanical or hydraulic problems by means of electronic analogs of mechanical or hydraulic systems: operational amplifiers are the basic components of an analog computer.

As may be expected, simple linear voltage amplification by precise gain factors is, though useful, not by any means the limit of the operational amplifier's capabilities.

Capacitor as Z_i Senses Rate-of-Change

Remembering that an operational amplifier with a resistor as a feedback element responds with an output voltage equal to the product of the input current and the feedback resistance, let's consider what happens if a capacitor is used instead of a resistor as Z_i (Figure 3).

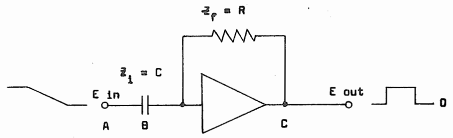


Figure 3.
Operational Amplifier as Differentiator. Output is proportional to *rate-of-change* of input voltage. $E_{out} = \frac{-dE_{in}}{dt}$ X RC.

The current through a capacitor is proportional to the *rate-of-change* of the voltage across the capacitor. A steady state DC voltage across a capacitor (assuming an "ideal" capacitor) passes no current through the capacitor, so no balancing current need be furnished by the output to hold the —input of the operational amplifier at ground. The output voltage then, is zero.

If the voltage at the input is changed, however, the *change* causes a current to flow through capacitor Z_i. The *amount* of current that flows is directly proportional to the capacitance of Z_i times the *rate of change* of the input voltage.

Let's assume that the potential at point A is +100 v DC, and that we change it smoothly to +95 v DC in five seconds. This represents a rate of change of one volt per second, the change taking place over a period of five seconds. If the value of Z_i is 1 μf, then, a current of —1 microampere will flow through Z_i for those 5 seconds.

The operational amplifier will cause an equal and opposite current to flow in Z_f. If we select a value of 1 megohm for Z_f, the one microampere current necessary to balance the circuit will require +1 v to appear at the output of the operational amplifier, during the time that 1 μa current flows through the capacitor.

This operation is *differentiation*: sensing the *rate-of-change* of an input voltage, and providing an output voltage proportional to that rate of change.

The actual relationship of output to input is this: $E_{out} = -\left(\frac{dE_{in}}{dt}\right)$ (RC), where the expression $\frac{dE_{in}}{dt}$ indicates the rate of change (in volts per second) of the input signal at any given instant, and R and C are Z_f and Z_i respectively.

In our example, we used a constant rate of change, and obtained a constant voltage level out. Had the rate been less even, the output signal would have demonstrated this dramatically with wide variations in amplitude. The differentiator senses both the rate and direction of change, and is very useful in detecting small variations of slope or discontinuities in waveforms.

Differentiator Has Rising Sine Wave Response Characteristic

In responding to sine-waves, the differentiator has a rising characteristic directly proportional to frequency, within its own bandwidth limitations (see Figure 7). The output voltage is equal to (E_{in}) (2πfRC), and the output waveform is shifted in phase by —90° from the input (the phase shift across the capacitor is actually +90°, but the output is inverted, shifting it another 180°).

Capacitor as Z_f Senses Input Amplitude and Duration

If we interchange the resistor and capacitor used for differentiation, and use a resistor for Z_i and a capacitor for Z_f (Figure 4) we obtain, as might be expected, the exact opposite characteristics from those obtained above. While in differentiation we obtained an output voltage proportional to the rate of change of the input, by swapping the resistor and capacitor, the output signal becomes a rate of change which is proportional to the input voltage.

This characteristic allows us to use the operational amplifier for integration, since the instantaneous value of output voltage at any time is a measure of both the amplitude and duration (up to that time) of the input signal — to be exact, a sum of all the amplitudes, multiplied by their durations, of the input waveform since the start of the measurement.

Here's how integration works: Let's assume the conditions of Figure 4 ($Z_i = 1$ meg, $Z_f = 1\ \mu$f), and an input signal level of zero volts. No current flows through Z_i, so the operational amplifier needs to supply no balancing current through Z_f. Suppose now we apply a DC voltage of —1 v to Z_i. This will cause a current of —1 μa to flow in Z_i, and the operational amplifier will seek to provide a balancing current through Z_f. To obtain a steady current of 1 μa through 1 μf, the operational amplifier will have to provide a continually rising voltage at the output, the rate of rise required being 1 volt per second. It will continue to provide this rate of rise until the input voltage is changed or the amplifier reaches its swing limit ("bottoms out"), or approaches its open-loop gain.

Now, this rate-of-rise, though helpful in understanding the mechanism by which the operational amplifier performs integration, is not the "answer" we seek from an integrator. The significant characteristics is the exact voltage level at a certain time, or after a certain interval.

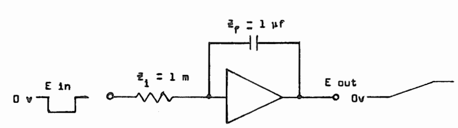


Figure 4.

Operational Amplifier as Integrator. Output *rate of change* is proportional to input voltage. $\frac{dE_{out}}{dt} = \frac{-E_{in}}{RC}$, or $E_{out} = \frac{-1}{RC}$ $\int E_{in}$ dt. RC in the example here is 1 second. Output, then, is 1 volt per second per volt input, and—most important—the output level at anytime is *one volt per volt-second input.*

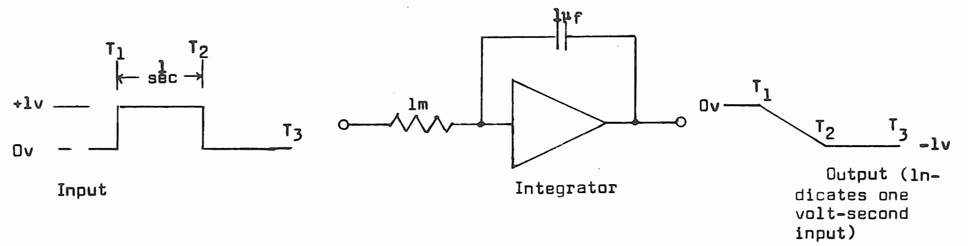


Figure 5.
Simple case of integrating 1-volt-second pulse. Integrator does not improve measurement accuracy in so simple a case.

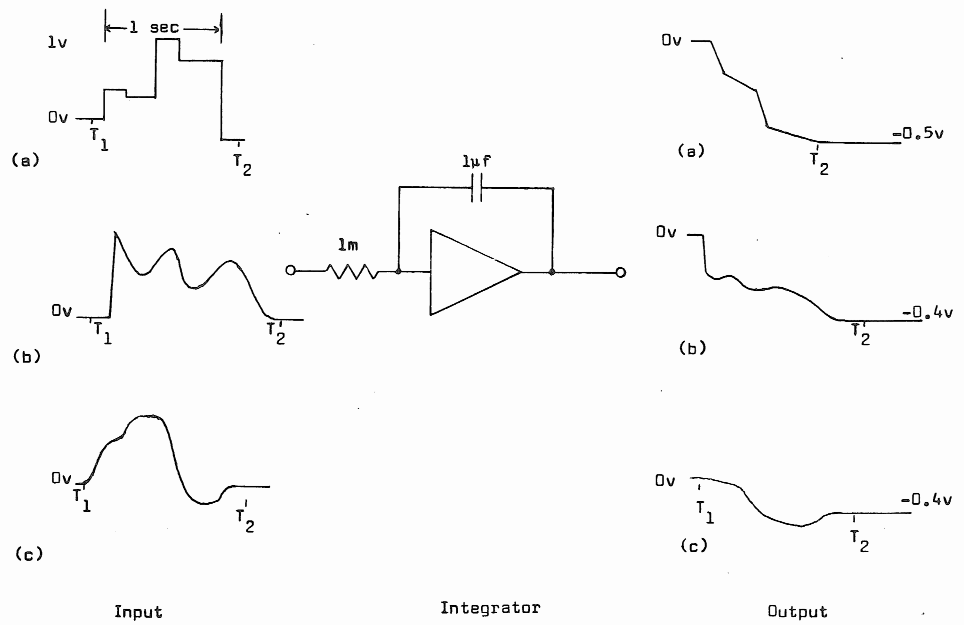


Figure 6.
Integrating more complex waveforms to determine "area under the curve" between T_1 and T_2. Note that in (c) the negative portion of the input waveform *reduces* the net integral.

Integrator Holds Final Level Until Reset

Before the amplifier reaches its output limit, suppose we remove the input voltage to Z_i. The output does not return to ground, but remains at the level it reached just before the signal was removed. The rate of rise has stopped because the necessity for providing $+1\,\mu a$ through Z_f to maintain the null at the —input has been removed. With an ideal capacitor and amplifier, the output voltage would remain at the last level reached indefinitely, until an input signal of the opposite polarity were applied to Z_i, and a negative-going rate of change at the output were required to maintain the null at the —input.

If the positive input signal is greater than our original —1 volt, it will take less time for the output voltage to reach zero than it originally took to rise. If the positive signal is smaller, it will take more time.

The absolute output level of the integrator at the end of some interval is the sum of the products of all the voltages applied to Z_i since the output was at zero, times the durations of these voltages, that sum divided by —RC.

Interpreting Answers Obtained From Integrator

The mathematical expression for the output level reached in a given interval of time $(T_2 - T_1)$ is as follows:

$$E_{out} = \left(\frac{-1}{RC}\right) \int_{T_1}^{T_2} E_{in}\, dt$$

The integral sign indicates that the value to be used is the *sum* of all of the products (E_{in} X dt) shown, between the limits (T_1, T_2) noted. The expression "dt" indicates infinitely small increments of time.

It is not necessary, however, to understand and be able to manipulate expressions in integral calculus to understand and make use of an operational amplifier integrator.

The integrator provides a voltage output proportional to the net number of volt-seconds applied to the input. If the total volt-seconds of one polarity is equalled by those of the opposite polarity, the output level at the end of the selected interval will be zero. Let's look at some examples.

Simple Example of Data From Integrator

First, we'll assume the signal we want to integrate is a simple one-volt positive pulse of one second duration (Figure 5). The sum of all voltages times durations between T_1 and T_2 is one volt-second. Using 1 megohm and 1 microfarad for Z_i and Z_f, the operational amplifier output will fall at the rate of one volt per second $\left(\frac{-E_{in}}{RC}\right)$ for one second, reaching —1 v when the pulse ends, and remaining at that level.

In reading this output level at T_2 we know that the input signal has amounted to 1 volt-second during the interval T_1 to T_2. Note also that a later observation, at T_3, gives the same answer, since E_{in} has been 0 between T_2 and T_3.

More Complex Cases

Now, take the more complicated case of the waveform in Figure 6-a. Its four voltage levels, of different duration, cause the integrator output to fall at four different rates, reaching a final level representing the total number of volt-seconds contained in the waveform. It should be apparent now that the integrator can measure the total volt-seconds contained in even the very complex waveform of Figure 6-b — something that would be difficult to measure by direct observation of the waveform. This type of operation is often referred to as "taking the area under the curve," since the area underneath a waveform plotted against time (i.e., the area bounded by T_1, T_2, the waveform and the line representing 0 volts) is the number of volt-seconds involved. Note, too, that we needn't wait for T_2 to obtain a reading: the instantaneous value of E_{out} at *any time* is proportional to the input volt-seconds up to that time.

Using Different Values of R and C

In the cases we've used for illustration, RC was 1 ($10^6 \times 10^{-6}$), and the numerical value of the output voltage at the end of the integrating interval was the number of volt-seconds in the input waveform. Using other values of R and C requires some additional calculation. To find the actual input volt-seconds, multiply the output voltage by (—RC). Example: R is 200 k, C is .01 μf and the output voltage after the selected interval is —2.5 volts. Multiplying —2.5 by $(-2 \times 10^5 \times 1 \times 10^{-8})$ gives us 5×10^{-3}, or 5 millivolt-seconds, positive polarity. Note that because of the polarity-reversal in the amplifier, we multiply by (—RC), to obtain the proper sign in the answer.

Measuring Ampere-Seconds (Coulombs)

To measure ampere-seconds, Z_i is omitted, and the current source is applied directly to the —input. The output level reached in a given time $(T_2 - T_1)$ is $\frac{-1}{C}\int_{T_1}^{T_2} I_{in}\, dt$.

Integrator Response to + and — Signals

If a waveform to be integrated contains both positive and negative polarity portions during the integrating interval, the output will be proportional to the *difference* between the volt-seconds of each polarity, the integrator being an averaging device. If it's desired to add the two polarities instead of allowing them to be subtracted, it is necessary to precede the integrator with an "absolute-value amplifier" (full wave rectifier) which inverts one of the polarities.

Necessity to "Reset" Integrator After T_2

The "integrating interval" (T_1 to T_2) has been mentioned several times. Because we frequently deal with repetitive signals; and continued integration of a waveform which is not perfectly symmetrical with respect to zero volts will eventually drive the operational amplifier to its output voltage limit, it's desirable to have some way of returning the output to zero at or after T_2, the end of the desired interval.

For slow work, a pushbutton which can be used to discharge Z_f manually is usually sufficient. Other circuits which may be used to perform this function automatically are shown in the applications section of the Type O-unit manual. Where the integrating interval is quite short, RC networks may be placed around Z_f to return the output level to 0 v through a time constant much longer (e.g., 100X) than the integrating interval.

In the Type O Unit, the "Integrator LF Reject" switch-positions perform this function whenever Z_f is set to a capacitive value.

Since the LF Reject circuit operates continually to return the integrator output to zero, it is necessary not only to keep the integrating interval short with respect to the LF Reject time-constant, but also to measure E_o before it has had a chance to decay, whenever these circuits are used. The value of resistors used in the circuit will also limit the maximum output obtainable for any given amplitude input (Max $\frac{E_{out}}{E_{in}} = \frac{R_r}{Z_i}$, where R_r is the resistance of the LF Reject circuit.

Reset or LF Reject Imperative When $Z_f = C$ is Small

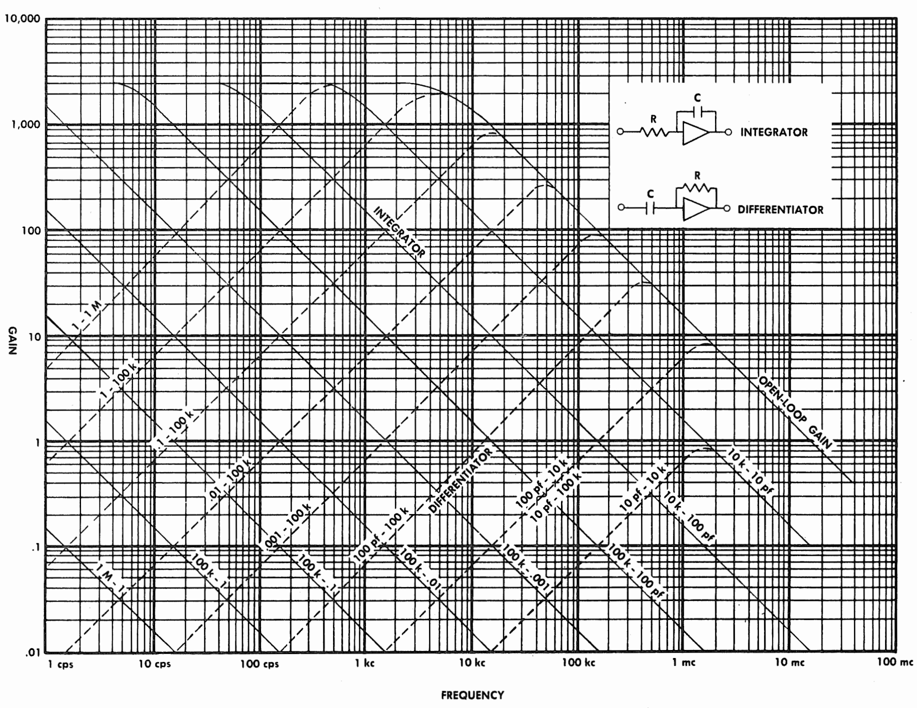
Use of resetting or LF reject circuits is usually imperative when small values of C are used for Z_f, since the small amount of grid current which flows in the —input grid even in the absence of an input signal is sufficient to cause a relatively rapid rise in output voltage as the operational amplifier tries to hold the —input null with balancing current through Z_f.

Response of Integrator to Sine Waves

For sine waves, the gain of the integrator varies inversely with frequency, the actual gain being $\frac{-1}{2\pi\, fRC}$, except as limited by the open-loop gain (at low frequencies) and the open-loop gain-bandwidth product at high frequencies (see Figure 7). At low frequencies, the gain becomes less than the formula would indicate, the effect becoming noticeable at the point where the formula indicates a gain of approximately 1/3 the open loop gain. At high frequencies, the error becomes significant above approximately 1/10 of the open-loop gain-bandwidth product. Except as limited above, the integrator shifts the phase of the input sine wave by +90°.

Editor's note: The second (and concluding) part of this article will appear in the April '63 issue of SERVICE SCOPE. This second part will discourse on the + input, a feature of some operational amplifiers. It will also discuss limitations of operational amplifiers, chief of which are:

1. *Open-loop Gain*
2. *Gain-bandwidth product.*
3. *Grid current (primarily of concern during integration).*
4. *Output-current and voltage capability.*
5. *Signal-source impedance.*



Average Gain-Frequency characteristics for integration and differentiation.

Figure 7.

INTRODUCTION TO OPERATIONAL AMPLIFIERS

Prepared by
Tektronix Field Information Department

Part 2

Use of The +Input

Many operational amplifiers (including those in the Tektronix Type O unit) provide access to a non-inverting input, referred to as the +grid or +input. A positive-going signal injected at this point produces a positive-going signal at the output. Conventional identification of + and — inputs is shown in Figure 8.

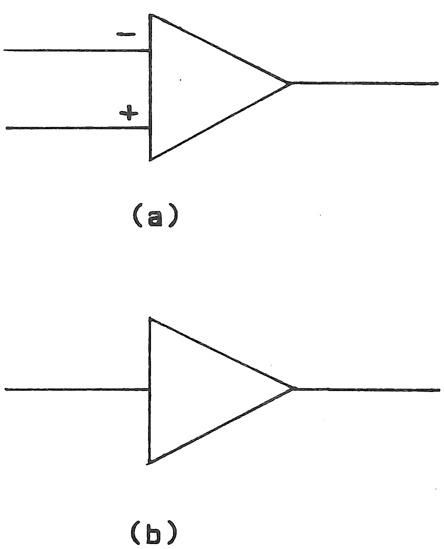


Figure 8

Identification (a) of + and — inputs of an operational amplifier. If only one input is shown (b), it is always assumed to be the —input.

If the output is connected directly to the —input, the operational amplifier becomes a non-inverting gain-of-one voltage amplifier for a signal applied to the +grid, with very high input impedance and very low output impedance.

Non-Inverting Amplifier With Gain > 1

With less than 100% negative feedback (Figure 9), obtained by putting the —input on a voltage divider between the output and ground, gains of greater than one may be realized, the actual gain being $\frac{R_i + R_f}{R_i}$ or 2 where $R_i = R_f$.

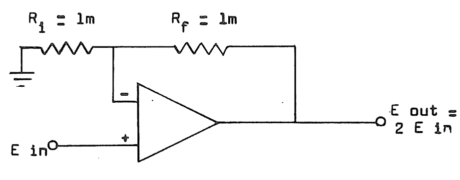


Figure 9

Gain of Two Using +Input. Very high input resistance ($>10^9\ \Omega$). for signals on the order of 1 v amplitude is possible. Other values of gain may be obtained using different ratios of R_i and R_f.

Feedback applied to the +input from the output is positive feedback, which tends to raise the input impedance of the +input toward infinity as the amplitude of the feedback approaches the amplitude of the input signal. If the loop gain (feedback amplitude compared to signal amplitude) exceeds 1 for any frequency, the amplifier becomes unstable (negative input resistance) and will oscillate at that frequency. If the loop gain exceeds 1 at DC, the amplifier will swing to its output voltage limit and stay there. The +input is useful for applications combining positive and negative feedback, and for use of the operational amplifier as an oscillator, waveform generator or multivibrator. The +input may also be used to provide a balanced or differential input, in which the operational amplifier responds only to the instantaneous *difference* between the signals applied to the + and — inputs. Other uses are suggested in the applications section of the Tektronix Type O Operational-Amplifier instruction manual.

Operational Amplifier Limitations

In performing linear operations with an operational amplifier, it is necessary to recognize and allow for the limitations of the amplifier and technique used, to obtain accurate results. The chief limitations are:

1. Open-loop gain.
2. Gain-bandwidth product.
3. Grid current (chiefly of concern during integration).
4. Output current and voltage capability.
5. Signal source impedance.

1. *Open Loop Gain*

The accuracy of all operations is ultimately limited by the open-loop gain of the amplifier, which determines how closely the amplifier is capable of holding the —input null. An amplifier with infinite gain would provide a null of exactly 0 volts, and the impedance at the —input (using feedback) would be exactly 0 ohms.

With finite gain, the —input does not quite null, and does not appear as 0 ohms. With an open-loop gain of A*, the —input moves $\frac{1}{A}$ times the output voltage swing, and appears as an impedance which is $\frac{Z_f}{1 - A}$. If this voltage swing of $\frac{E_{out}}{A}$ is a significant fraction of the input signal E_{in}, or if the impedance $\frac{Z_f}{1 - A}$ is a significant fraction of Z_i, there will be a definite output signal error in addition to the error introduced by the tolerances of

*Common usage in the analog computer field assigns a *negative number* to the open-loop gain between the —input and output (and a positive number to the gain from the +input). Therefore, in calculating values from formulas involving A and the —input, it is necessary to keep in mind that *A is a negative number,* and the expression "1 — A" for instance, when A is —2500, equals +2501, not —2499.

One simplification has been made in this article. Closed-loop gain, commonly expressed

as $\frac{-Z_f}{Z_i}\left[\frac{1}{1 - \frac{1}{A}\left(1 + \frac{Z_f}{Z_i}\right)}\right]$

has been reduced to:

$\frac{-Z_f}{Z_i}\left[\frac{A}{A - 1 - \frac{Z_f}{Z_i}}\right]$. It may also be

written $\frac{-Z_f}{Z_i}\left[\frac{1}{1 - \frac{1 + Z_f/Z_i}{A}}\right]$, if

this seems to indicate the effect of A on accuracy more clearly.

Z_i and Z_f. The exact value of this error is $1 - \dfrac{A}{A - 1 - \dfrac{Z_f}{Z_i}}$. So long as $\dfrac{Z_f}{Z_i}$ is small and A is large, the error is not serious. For instance, using the O-Unit's operational amplifiers (at low frequencies where $A = -2500$) in the simple fixed-gain amplifier mode with resistors for Z_i and Z_f, we see that the error for the gain of 1 ($Z_i = Z_f$) is only $1 - \dfrac{2500}{2502}$, or less than 0.1%. For a gain of 100, however, the error becomes $1 - \dfrac{2500}{2601}$, or almost 4%. (A gain-correcting resistor is automatically shunted across Z_i in the O-Unit when the internal Z_i resistor is set to 10 k and Z_f to 0.5 or 1.0 meg.

Using external components, similar precautions should be observed when high gain is required).

Approximate Error Calculation Using C for Z_i or Z_f.

Since it is not easy to assign a single impedance value in the error formula for Z_i or Z_f when one of them is a capacitor, it is convenient to use the ratio E_{out}/E_{in}, representing the actually obtained voltage gain, to compute the approximate error. The error ε is found by this formula:

$$\varepsilon = 1 - \left[\frac{\dfrac{E_{out}}{E_{in}} - A}{1 - A} \right] \text{, or, more simply, } \varepsilon = \frac{1 - \dfrac{E_{out}}{E_{in}}}{1 - A}$$

where A, as before, is the open-loop gain, and E_{out}/E_{in} is the actually obtained voltage gain (Don't forget — both A and E_{out}/E_{in} are *negative numbers*). Example: where A is -1000 and the observed E_{out}/E_{in} is -50, the error has been 51/1001 or 5.095%. The output "-50" represents, then, 94.905% of the correct value, and the correct value is $-50/0.94905$, or almost -52.7.

For convenience, you may want to rearrange the terms as shown below, to determine how large an output signal to allow, for a given input and an arbitrarily selected maximum error:

$$\frac{\text{Max } E_{out}}{E_{in}} = 1 - \varepsilon\ (1 - A)$$

Using the Tektronix Type O Operational Amplifier for integration, for instance, to keep error due to amplifier gain below 1%, the output voltage during or at the end of the integrating interval should not exceed the average value of the signal being integrated by more than a factor of $1 - (.01 \times 2501)$, or -24, for low frequencies. The same limitation should be observed during differentiation.

The minimum open loop gain required by an operational amplifier to operate within a given error even at "zero" Z_f/Z_i is

$$A = \frac{(\varepsilon - 1)}{\varepsilon}$$

, where ε is the error expressed as a decimal fraction ($.01 = 1\%$, $0.1 = 10\%$, etc.).

Where Z_f/Z_i is a finite number, the minimum open-loop gain required for a given maximum error is:

$$A = \frac{(\varepsilon - 1)\ (1 + Z_f/Z_i)}{\varepsilon}$$

The application of these formulas will be most useful in observing gain-bandwidth limitations, discussed below.

2A. Gain-Bandwidth Product:

The gain-factor A varies with frequency, and it's important to know what the effective value of A is for the frequencies or signal frequency components being used. In the Type O, the gain factor A is constant (-2500) only to about 1 kc, dropping off to -1000 at about 15 kc, and reaching a value of -1 at approximately 15 Mc.

The error introduced by the gain factor, then, becomes greater with frequency, and for accurate measurements the allowable ratio of E_{out} to E_{in} must be reduced as higher-frequency information is processed.

Although the drop in gain at high frequencies in the open-loop bandwidth characteristic follows the same pattern as that of an integrator, it must be remembered that this response is obtained *without* input and feedback elements. The effect of this rolloff will *add* to the effect of the integrating components, altering their effect.

At a frequency approximately 1/10 of the open-loop gain-bandwidth product, the open-loop gain will be insufficient (on the order of 10 or so) to provide accuracy better than 9% even at "zero" closed loop gain, or 16.7% when Z_f/Z_i is 1, (i.e., $E_{out} \approx E_{in}$). Above 1/10 of the open-loop gain-bandwidth product, answers will be only approximate, although the data will be useful for frequencies as high as 1/3 of the open-loop gain-bandwidth product. For high-frequency work, then, the nominal values of Z_f and Z_i are usually trimmed to compensate for gain-factor error and improve functional accuracy.

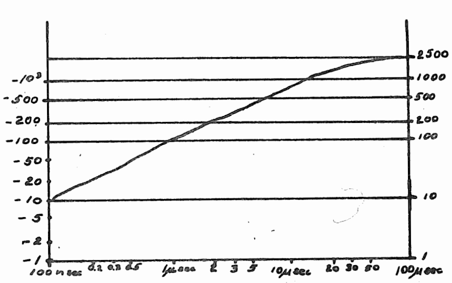


Figure 10 (a)

Variation in open-loop gain after application of signal, for O-Unit.

2B. Gain-versus-Time Factor — Complex Waveshapes:

In working with pulse and complex waveforms, open-loop gain in terms of frequency is not too useful. Instead, the open-loop risetime characteristic, Figure 10 (a), may be used to determine the time after the start of a signal at which the A-factor has reached a sufficiently high level to permit the desired accuracy.

Figure 10 (b) shows the A-factor required to support a given accuracy at a given attempted or "virtual" gain (Z_f/Z_i).

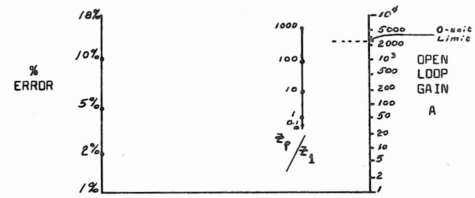


Figure 10 (b)

Nomograph for determining A-FACTOR, ERROR and Z_f/Z_i. Given any two factors, the third may be found. (Lay straightedge across chart.)

"Virtual gain" (roughly, E_{out}/E_{in}) in the case of integration or differentiation is the ratio between the RC time constant chosen and the time interval involved in the operation.

In the case of integration, virtual gain G_v will be $G_v = \dfrac{-t}{RC}$, where t is the integrating interval — i.e., that span of time during which the integral continues to increase. The larger the values of integrating components, the smaller the virtual gain.

In the case of differentiation, the virtual gain will be: $G_v = \dfrac{-RC}{t}$, where t is that span of time during which the input signal has its steepest slope. The larger the values of differentiating components, the higher the virtual gain.

As can be seen from Figure 10 (b), holding virtual gain to a value of one or so is a good general rule of thumb for accurate measurements.

NOTE: It should be kept in mind that the values of the internal 10 pf and 100 pf Z_f and Z_i components of the O-Unit have been adjusted under *dynamic conditions*, to compensate partially for the time-dependent errors indicated in Figure 10 (a). For greatest measurement accuracy, standard waveforms involving a similar time interval and virtual gain as the signal to be measured should be used to determine the probable measurement error, or to trim the values of external components to provide direct readings for the particular waveform to be measured (comparison method). However, correction of this sort can be optimized for only a limited range of waveforms, and cannot extend the operating range of the system indefinitely.

3. *Grid Current*:

During integration, any grid-current flowing in the —input will be integrated along with the current through E_{in}, except when this current is bucked out through a DC path from output to input (in the Type O Unit, "Integrator LF Reject" circuits are provided for this purpose).

The amount of grid-current flowing in the —input circuit may be determined by switching out any "LF Reject" circuit and measuring the length of time it takes the output signal to rise or fall 1 v with a capacitor as Z_f (no signal input). The grid current I_g is found by the formula $I_g = \frac{C}{t}$, where t is the time (in seconds) required for the output to move one volt. A grid current (electron current) at the —input of 300 picoamp is normal.

It is not usually practical to try to adjust a wide band operational amplifier for "zero" input current, since this condition is not as stable as is a fixed value of grid-current appropriate to the input tube type and amplifier design. In low-frequency operational amplifiers using electrometer tubes as input elements, extremely low values of input grid current can be obtained with good stability. In wide-band units, higher values must be tolerated.

Once the grid current has been set to a known value, its effect on a given integrating operation can be computed. So long as the value of I_g is very small compared to the average value of the current through Z_i during the integrating interval, the effect of I_g can be largely ignored.

4. *Output Current and Voltage Limits*:

Any operational amplifier is limited in the amount of current and voltage it can deliver to its feedback network and any external load with good linearity. If these limits are exceeded during any part of an operation, the accuracy of that part of the operation, at least, will be impaired.

In the case of the Type O Unit, maximum output is ± 50 v and ± 5 ma. At high speeds, the maximum rate-of-change at the output will be limited by the available current, and should not exceed 20 v per µsec, when loaded by the O-Unit's oscilloscope preamplifier (47 pf) and 10 pf of other loading (e.g., Z_f).

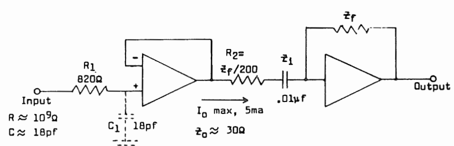


Figure 11

Operational amplifier connected as gain-of-one, non-inverting amplifier to drive low-input-impedance differentiator from high impedance signal source. If output current capability is 5 ma (as in the O-Unit), driver amplifier will reproduce faithfully an input rate-of-change as high as 0.5 v/µsec into 0.01 µf, several orders of magnitude in excess of the amount necessary to obtain usable output from the differentiator. Component R_1 combines with the input C of the first operational amplifier to compensate its response in this mode. R_2 limits the current to the second operational amplifier to prevent overdriving, and reduces noise components possibly introduced by first amplifier.

5. *Input Signal Source Impedance*:

A part of Z_i, the input element of the operational amplifier circuit, is the source impedance of the signal being processed. Linear operations using precision input and feedback components will be accurate only if the source impedance of the signal is very small compared to the impedance of the input component, or the value of the input or feedback component is trimmed to allow for the impedance of the signal source.

Where trimming of components is not practical, or the signal source impedance is not resistive and linear, the usual practice is to process the signal first through a gain-of-one, high input-impedance, low-output-impedance amplifier, such as that shown in Figure 11, to obtain a low-impedance signal source for the desired operation. In the case shown, the output impedance of the first amplifier is *too* low, making it capable of overdriving the second. A current-limiting resistor helps keep down noise as well as prevent overdriving.

Shunt Impedance Across —Input

Though we tend to think of the —input or —grid as a "virtual ground", and that impedances between this point and ground will have a negligible effect on the performance of the operational amplifier, this is only partially true. The true impedance of this point is $Z_f/1 - A$, and that instead of holding a perfect voltage null (as would be the case if A were infinite), its voltage excursions actually amount to E_{out}/A.

So long as A is large and Z_f has a fairly low value, an impedance across the —input which is large compared to Z_f or Z_i will have little effect on performance. However, in high-frequency work, where the effective value of A is low, more and more care must be exercised to assure that shunt impedances — particularly capacitive reactance, which becomes lower with increasing frequency — do not interfere with the operation (Figure 12).

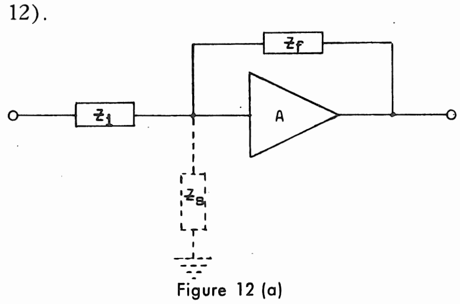


Figure 12 (a)

Shunt Impedance across —input. Where Z_s is large compared to Z_i and Z_f, and open-loop gain A is high, effect of Z_s is negligible.

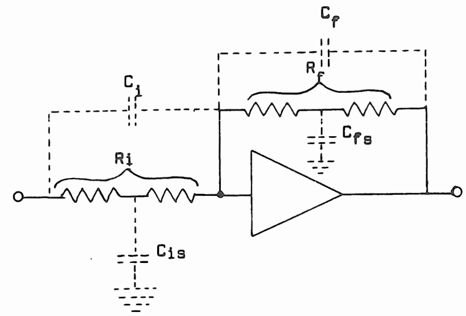


Figure 12 (b)

Where Z_i or Z_f is a resistor, and particularly if a large (>100 k) value, more serious errors may be caused by capacitance from the resistor body (highest impedance point) to ground, and, in the case of R_i during integration, end-to-end capacitance of R_i. Time constants involved in shunt capacitance C_{fs} and C_{is} are approximately RC/4.

The general expression for the closed loop gain of an operational amplifier

$$\frac{E_{out}}{E_{in}} = \frac{-Z_f}{Z_i}\left[\frac{A}{A - 1 - \frac{Z_f}{Z_i}}\right]$$

may be modified as follows to show the effect of shunt impedance Z_s across the —grid:

$$\frac{E_{out}}{E_{in}} = \frac{-Z_f}{Z_i}\left[\frac{A}{A - 1 - \frac{Z_f}{Z_i} - \frac{Z_f}{Z_s}}\right]$$

keeping in mind that A is a negative number. As you can see, unless Z_s is very high compared to Z_f, its effect on accuracy may become comparable to that of Z_f/Z_i.

The terms in the above equation can be rearranged to show the effect of Z_s as related to Z_i:

$$\frac{E_{out}}{E_{in}} = \frac{-Z_f}{Z_i}\left[\frac{A}{A - 1 - \frac{Z_f}{Z_i}\left(\frac{Z_s + Z_i}{Z_s}\right)}\right]$$

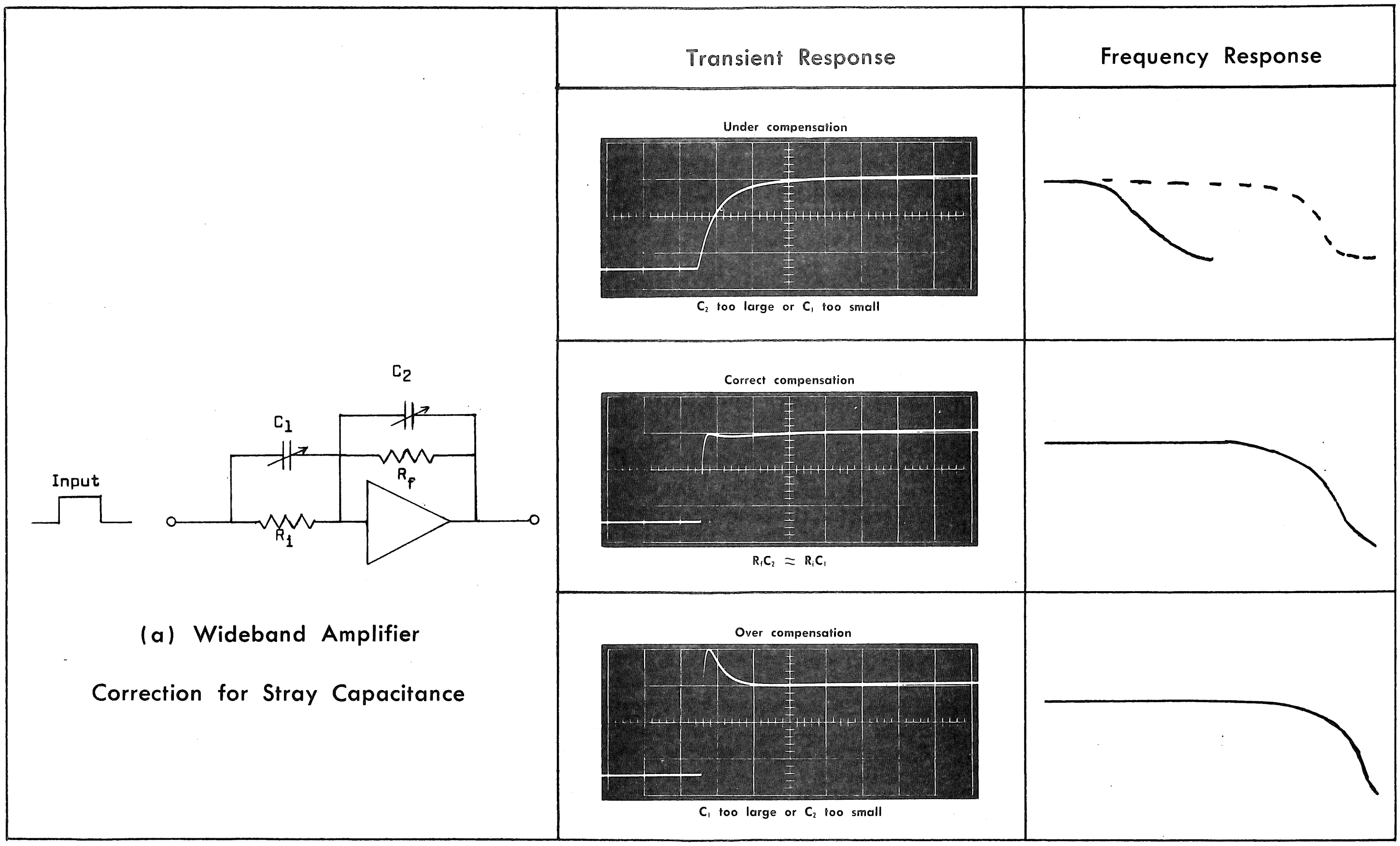
Correcting For The Effects of Stray C

In high-speed work, the accuracy of operations will be affected by Cs during the start of an operation when the effective value of A is low, and also by the end-to-end and distributed capacitance to ground of the resistors used for Z_i and Z_f (Figure 12b).

To correct for strays and the variation in A, the 100 pf and 10 pf values of Z_i and Z_f in the Type O operational amplifiers are factory adjusted under dynamic conditions, and no external compensation of these components is generally required. If it is intended to use values in this range externally, they should be padded or trimmed as necessary under conditions similar to those of the contemplated measurements.

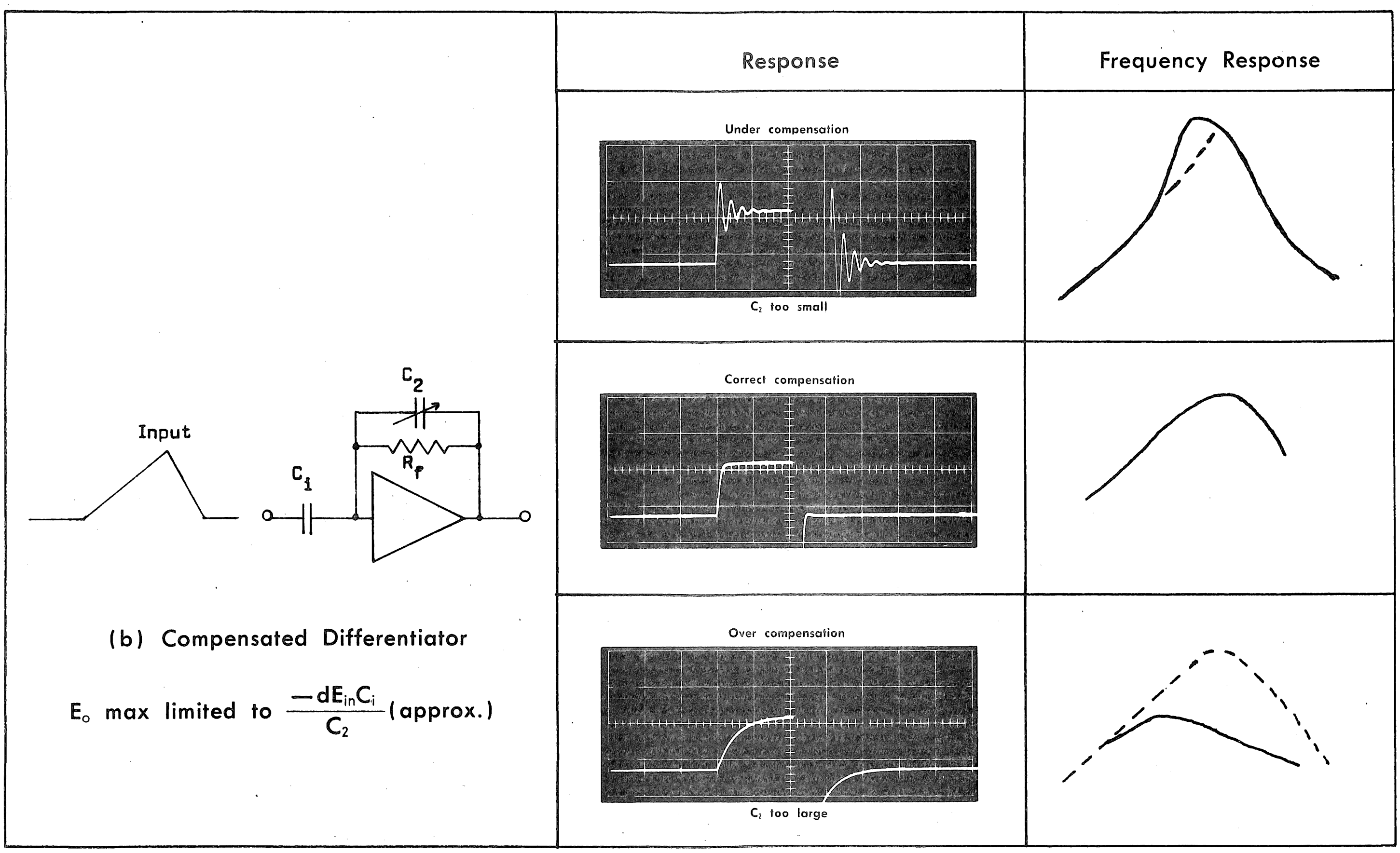
The resistors used as Z_i and Z_f, however, can be given only partial compensation internally, since the optimum value of compensation varies with the application. For this reason, it is usually necessary in dealing with short-duration or high-frequency

Figure 13



(a) Wideband Amplifier

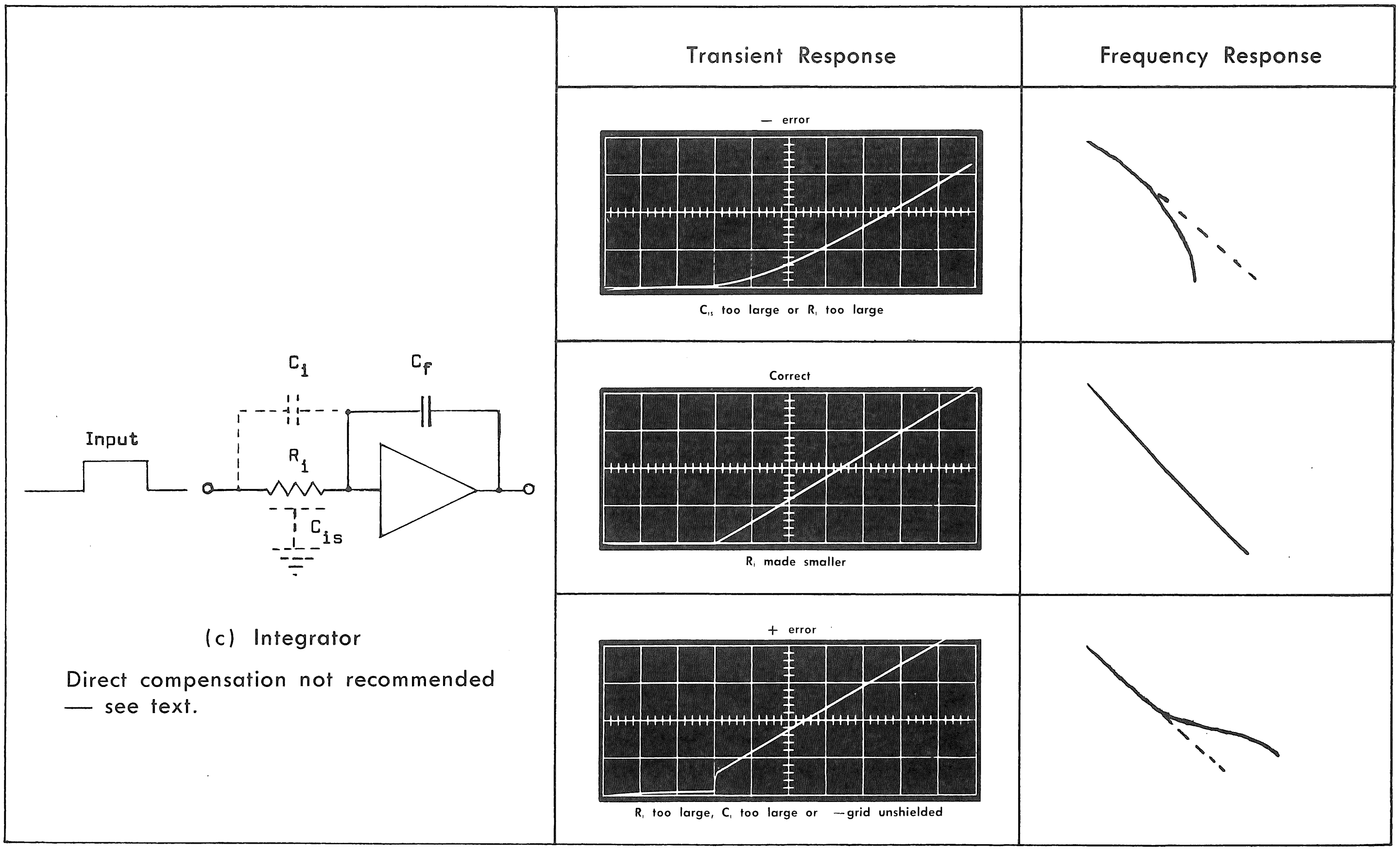
Correction for Stray Capacitance



(b) Compensated Differentiator

E_o max limited to $\frac{-dE_{in}C_i}{C_2}$ (approx.)

 11-25-64 TEK 0 IRB



(c) Integrator

Direct compensation not recommended — see text.

signals to add external compensation to R_f or R_i when these components are used in amplification and differentiation.

Figure 13 illustrates the corrections necessary to improve operational accuracy for each of the three basic operations.

Note, however, that except in the case of straight amplification (Figure 13a), the compensation itself introduces possible errors which must be recognized and allowed for in interpretation of results.

Compensated Amplifier

In the case of amplification, selecting small values of capacitance (on the order of 2-25 pf) for C_1 and C_2, the closed-loop risetime can be made to approach the slope of the open-loop risetime (Figure 9), providing a gain-bandwidth product about equal to the open-loop gain-bandwidth product. Without compensation, the amplifier may typically achieve only 1/20 of this figure.

Compensated Differentiator

Without compensation, the differentiator (Figure 13b) may respond to a sudden change in dE_{in}/dt by overshoot, followed by sinusoidal ringing, due to the fact that excess output voltage must be developed to charge via R_f the input capacitance and the distributed stray capacitance of R_f itself, as well as provide the current needed to obtain a null at the —input. As soon as the strays are charged, however, the excess current through R_f upsets the null, and the output must swing in the opposite direction to re-establish the null and discharge the capacitance associated with R_f — hence the ringing. A small capacitance across R_f provides the current needed to establish the null at the start of the waveform without having to develop excess voltage across R_f.

Differentiator Compensation Limits Initial Accuracy

The presence of this capacitance, however, limits the output voltage maximum to approximately $\frac{-dE_{in}\,C_i}{C_2}$. After an abrupt change in the input waveform, then, when dE_{in} is small, but $\frac{dE_{in}}{dt}$ X RC may be quite large, the output voltage limitation of $\frac{-dE_{in}\,C_i}{C_2}$ may result in a significant error. The solution in this case is to select a larger value of C_i and smaller values for R_f and C_2 (keeping the $R_f\,C_i$ time constant the same) to minimize the error, and keep its duration as short as possible.

Integrator Compensation Rarely Needed

Failure of the integrator to start integrating at the proper rate at the beginning of a fast-rise pulse or after a sharp step in the input waveform is usually due largely to the distributed stray capacitance to ground in R_i. This is infrequent; more commonly the error is in the opposite direction because of excessive capacitance coupling of the input waveform around R_i into the —grid directly, producing a step of approximately $\frac{-dE_{in}\,C_{in}}{C_f}$. The first (—error) waveform in Figure 13c was obtained by deliberately putting a ground plane near the center of a 9 megohm R_i and carefully shielding the —grid. Removing the ground plane and shield produced the third (+error) waveform, using the same input signal (a rectangular pulse) and components.

Normally, the "undercompensated" effect would only occur when R_i is composed of several resistors in series, or when a high-value potentiometer is used as, or in series with, R_i.

The solution usually is to select a smaller value for R_i and a larger one for C_f, to maintain the same time-constant. Normally, if a signal source is capable of driving a large value of R_i with capacitive compensation, it is also capable of driving a smaller value of R_i without compensation.

Theoretically — as when a potentiometer is used in conjunction with R_i — it is possible to compensate the RC losses in R_i by shunting R_i with a series RC network of the proper time-constant, or by using a small value of R in series with C_f. In practice, these added components usually add nearly as many stray-C problems as they cure, and "compensation" of this sort is not recommended. Compensating with simple capacitance across R_i produces a "step" error at any abrupt transition, and usually an error of greater magnitude than the one to be corrected.

If R_i is a single component, an environmental "guard" driven by the input signal (e.g., a short piece of wire soldered to the input end of R_i and dressed near the body of the resistor) can make some correction, but its use requires more complete shielding of the —grid and the —grid end of R_i.

Using "Standard" Waveforms For Comparison

The use of standard waveforms (pulses and ramps) with known parameters, is of considerable help in adjusting compensation and assuring best accuracy for critical measurements near the limits of the instrument's capabilities. For many purposes, such "standard" waveforms may be obtained by attenuation of the oscilloscope gate and sawtooth output waveforms. Selection of time and amplitude parameters close to those of waveforms to be measured will give best assurance against possible system errors.

Editor's note: This concludes the article "Introduction to Operational Amplifiers". If you missed Part 1, which appeared in the February 1963 issue of Service Scope, you can obtain a copy of that issue by contacting your local Tektronix Field Office or Field Engineer.

O-UNIT LF REJECT CIRCUITS

GEOFF GASS

7-3-63

Additional copies of this material are not immediately available. The conclusions are empirical, sufficient for the area under discussion, but rigorous proofs are not provided because additional effort cannot be spared.

O-Unit LF Reject Circuits

The "Integrator LF Reject" circuits in the O-Unit are provided to extend the usefulness of the integrating operation as much as possible without the expense and complexity of keyed reset circuits (such as the gating adapter).

The LF Reject Circuits have certain limitations, however, and some care must be exercised to recognize these, particularly when operating close to them, or large errors may result.

The following discussion covers these topics:

1. Purposes of the LF Reject Circuits:

   (a) Automatic Reset.
   (b) Limiting DC and LF Response.
   (c) Limiting grid-current effects.

2. LF Reject Circuitry and Theory.

   (a) DC Feedback.
   (b) Lag.

3. Specific limitations of 100 msec and 100 µsec circuits.

   (a) Factors to consider -- R_i and DC gain, Lag Time Constant, Integrating interval, virtual gain.
   (b) Specific Limits.
   (c) Ringing, oscillation, and rep-rate effects.
   (d) O-Units S/N 814 up -- Switch must be "off" when using +grid.
   (e) About the 1 µf 3 v capacitor 283-017.

4. Selecting components for external LF Reject Circuits.

   (a) Selecting resistance value.
   (b) Selecting lag time-constant.
   (c) Special circuits; use of input blocking capacitors, etc.

5. Gating Adapter (or equivalent) -- Where Required.

6. Addendum:

   (a) Verification of LF Reject circuit performance.
   (b) Alternate rule for avoiding integration errors with LF Reject circuit.

1. Purposes of the LF Reject Circuit.

(a) Automatic Reset. One of the purposes of the LF Reject circuit is in the integration of repeating waveforms, where the integral over only a limited span of time is required. An integrator is like an adding machine without a mechanism to clear the register. Every total you obtain includes the totals of all previous measurements you've taken since the last time the register was cleared. The reject circuit operates continuously to clear the register, but slowly enough that if you run up a fast total, the error in that small amount of time can be negligible.

(b) Limiting DC and LF Response. The O-Unit is capable of handling quite short-duration events with accuracy. However, a perfect integrator (Fig. 1A) has a frequency response inversely proportional to frequency, rising at 6 db per octave as you go <u>down</u> in frequency, reaching infinite gain at DC. In time-domain terms, this calls for infinite gain with infinitely long risetime. Neither is achievable, of course. All real integrators level off in frequency response at some low frequency above DC, and, in the time-domain, level off at some DC gain short of infinity and before the expiration of infinite time.

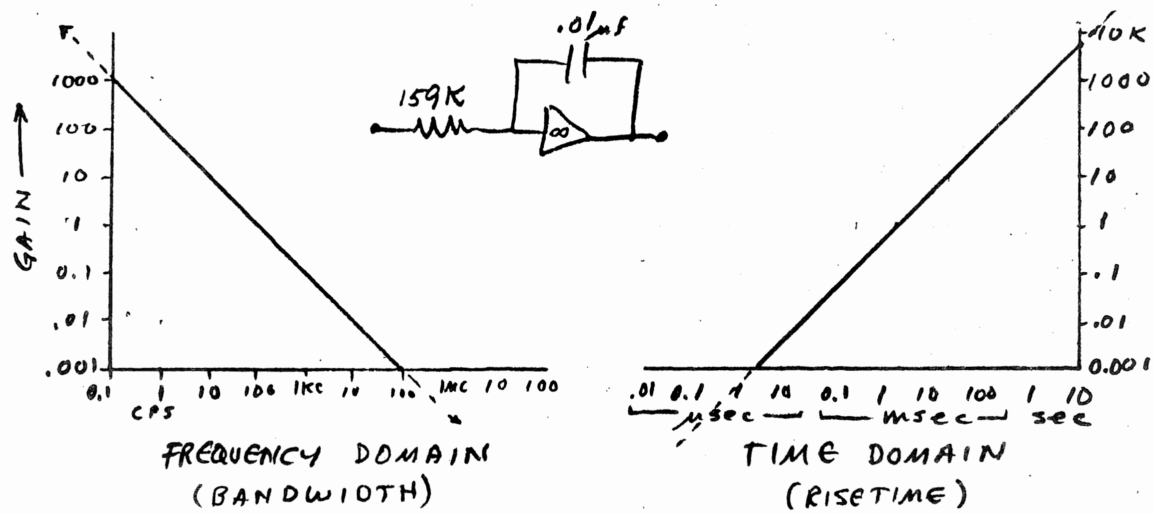


Fig. 1-A IDEAL INTEGRATOR R and C values only shift the location of these curves, not their shape or slope (O-Unit levels off at 2500 x gain).

These finite but still quite large values of low-frequency gain can be troublesome when working with high-frequency, short-duration phenomena.

Example: Using the integrator with 10k and 10 pf for a 100 nsec time-constant, integrating a pulse of about 100 nsec duration and 1 volt peak amplitude. The integrator response for this R_i C_f setting (see O-Unit Manual, page 2-6), rises at 6 db per octave down to about 1 kc, then flattens out at a gain of 2500 from about 500 cycles to DC. The integrator output for a 100 nsec wide pulse will be 1 volt per 0.1 µv/sec input, or probably about a half a volt (triangular pulse) to a volt (rectangular pulse) output for the pulse described.

Let's assume that along with our signal pulse, we have an incidental DC level of 1 millivolt and also 1 millivolt p-p of 60 cycle ripple.

The output of the 60 cycle ripple will be 2.5v p-p, and the 1 millivolt DC level will drive the trace offscreen at the rate of 10 v/msec until the operational amplifier reaches its swing limit.

Obviously, measuring a 1/2 to 1 volt integral under these conditions would be quite difficult, if not impossible.

The LF Reject circuit limits the DC response to the value R_f/R_i, where R_f is the resistance of the reject circuit and R_i is the integrating resistor. Low frequencies may also be attenuated to this level or a somewhat higher level, depending on the exact circuit employed.

(c) Limiting grid-current effects. Grid current flowing in the input grid circuit will act the same as a DC level in the input signal -- and tend to drive the trace off-screen. The O-Unit's 300 pico ampere grid current will cause a 1 µf capacitor as Z_f to charge at the rate of 300 µv/sec, or a 10 pf capacitor to charge at the rate of 30 v/sec. The actual amount of grid current flowing is not always 300 pA; this is only a maximum value; the amount may vary widely with DC Balance and output voltage level (which is a reflection of the input tube plate voltage and hence affects grid current).

The LF Reject circuit will limit the output DC level change due to grid current to I_gR_f, where R_f is the reject circuit resistance. With a typical value of 200k to 1.2 M, the offset is only 60 to 400 µv at the output for a 300 pA grid current.

2. LF Reject Circuitry and Theory.

(a) DC Feedback. The ideal integrator has no DC feedback, and so has infinite DC gain. However, since DC is "forever", it's obvious that actual DC response is never needed in an integrator (or a scope either, for that matter). Substantially uniform performance <u>over the span of the integrating interval</u> is all that's required of an integrator. Whenever the integrating interval is substantially less than the available linear slope on the integrator "risetime" characteristic, a resistive DC feedback circuit (Fig. 1B) can be used to limit the long-term response which is not needed. Fig. 1B shows the effects in both the time and frequency domains.

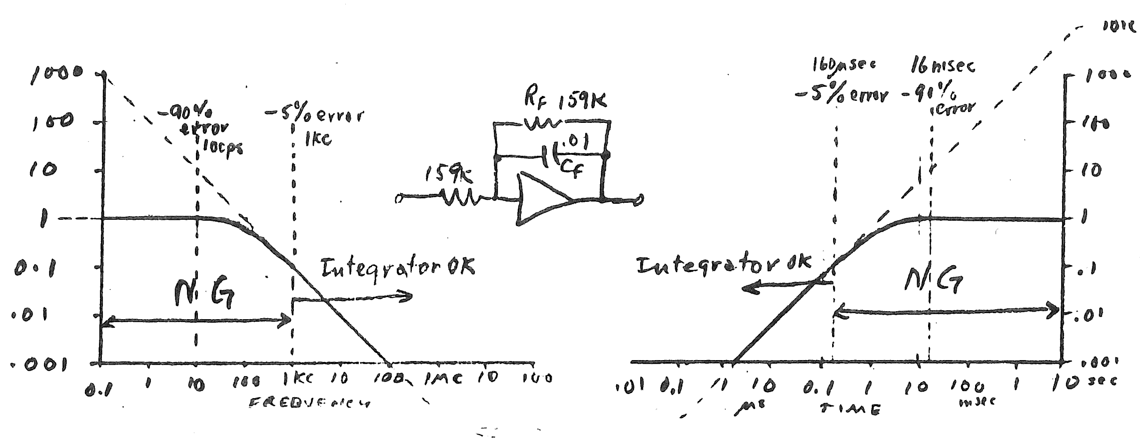


Fig. 1-B LOW FREQUENCY RESPONSE LIMITING. Simple resistor across C_f serves to limit LF response and DC gain, and resets integrator by providing DC feedback path, but limits accuracy near R_fC_f turnover or time constant. For extended range, larger value resistor may be used.

You will notice, however, that a simple gain-limiting circuit affects accuracy out to 10X the R_fC_f turnover point, or back to a tenth of the R_fC_f time-constant. This is not too desirable, and is the reason why this type of reject circuit is not used in the Type O.

(b) Lag.* The technique providing the greatest amount of integrator usability with minimum circuit complexity is the LF Reject circuit used in the O-Unit, employing lag, or delayed DC Feedback (akin to the delayed AGC of a radio receiver). The lag time-constant (see Fig. 1C) prevents the feedback resistance from providing any significant amount of current to the -grid until some time after the start of the signal to be integrated.

The values shown in Figure 1C are not those used in the O-Unit, but are shown only as examples. Also, the accuracy limitations indicated are typical only of the values shown; the actual limitations using O-Unit circuits are taken up later. The overshoot and ringing -- for instance -- occur only with certain combinations of integrating and reject circuit values.

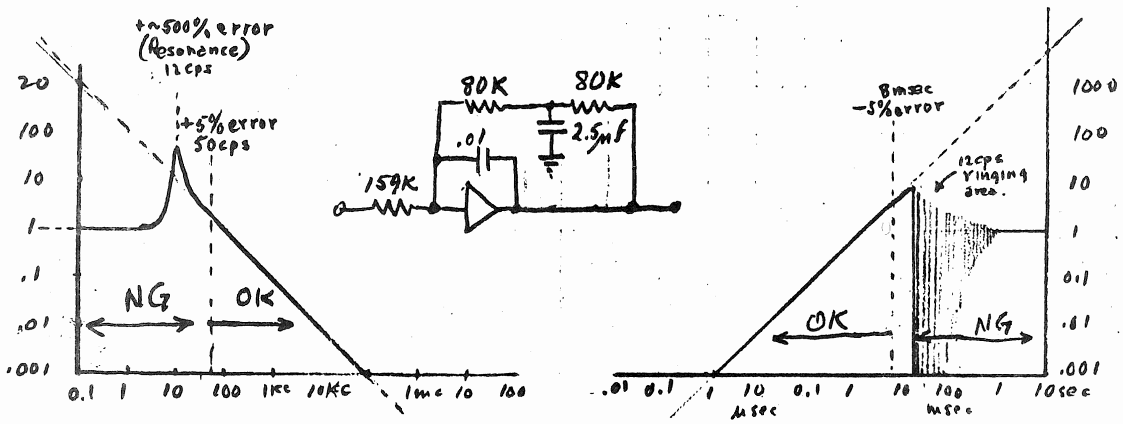


Fig. 1-C REJECT CIRCUIT WITH LAG IN DC FEEDBACK. Lag time constant of 100 ms extends usable range of integrator without increasing DC response. Larger values of R and longer time constants provide greater improvement.

The action of the lag circuit in the feedback loop does not yield accurately to simple algebraic analysis. However, a study of the action of a passive RC circuit in responding to the integral of a step waveform and a step-by-step approximation of the effects in the feedback loop will indicate the basic principles and limitations of the lag circuit.

We'll assume that we want to integrate a one-volt pulse over some number of milliseconds, using a 100 msec integrating time-constant of 0.1 M Z_i and 1 μf Z_f, and that the O-Unit's 1 cps LF Reject circuit (Fig. 2A) is to be used to minimize drift.

*Hiro Moriyasu points out that in operational amplifier practice, it is usual to refer to all functions with reference to their effect on E_o as a function of E_{in}. From this point of view, all low-pass circuitry in the feedback is "lead" (differentiation) circuitry, since its net effect on the overall operation is to enhance high-frequency gain. However, for simplicity's sake, we refer here only to its effect on the feedback signal, and call it "lag" (integration). This is just for in-house use; some care will have to be exercised in talking with customers on this subject.

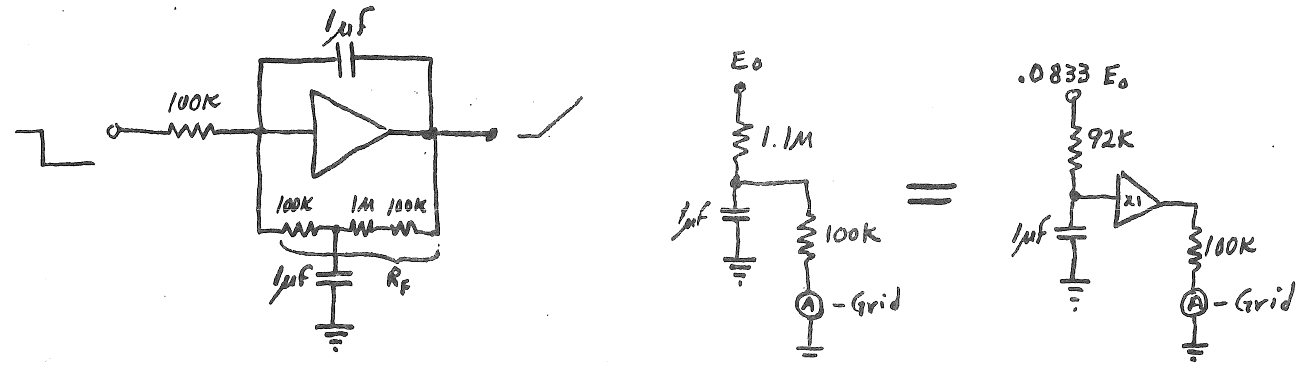


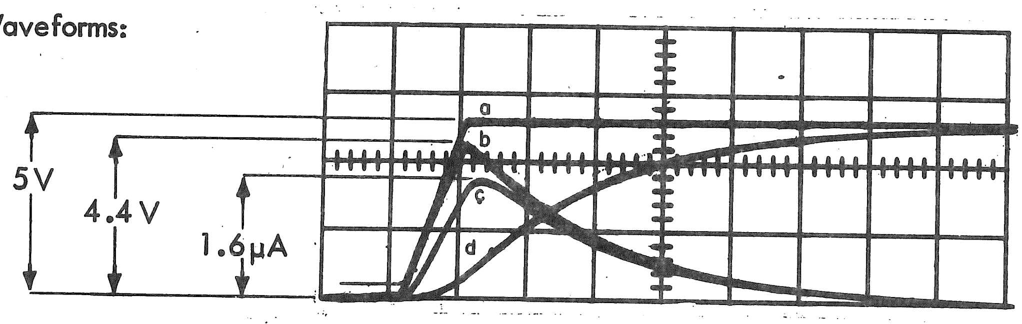
Fig. 2-A 100 ms integrator with 90 ms, "1 cps", LF Reject circuit and equivalent networks for figuring time constants.

To analyze the operation of the circuit, we first find equivalent networks for the LF Reject circuit.

First, looking at the dc path only (imagine disconnecting the lag capacitor) we see that the total resistive path from output to -grid is 1.2 M, making the dc gain limit 12 (R_f/R_i).

Second, we determine the effective AC network for the feedback path (not including C_f). The lag time-constant consists of the lag capacitor and the parallel combination of the resistance from the lag capacitor to the -grid and the resistance from the output to the lag capacitor. This comes out 92 k, which, with 1 µf, gives us about a 90 msec time constant. The voltage to which the capacitor charges is $\frac{100\,k}{100\,k + 1.1\,M}$ times the integrator output voltage, or .0833 E_o. Looking on this capacitor as a zero-impedance voltage source, the equivalent network shows it as supplying current to the -grid (which should be considered as a zero impedance point) through a 100 k resistance. So for every millivolt of charge on the capacitor, 10 nanoamperes flows to the -grid.

Fig. 2-B Waveforms:



a. True integral of 1 V x 500 ms pulse. (2 V per cm)
b. Output of integrator using 1 cps LF Reject. Over first 500 ms result is about 10% low.(2 V per cm)
c. LF-reject circuit current to negative grid. Note lag of about 90 ms. (1 µA per cm)
d. Accumulated error. Negligible for about 200 ms; 10% at 500 ms; Oscillogram should not be scaled because of parallax and calibration error.

Any current flowing to the -grid null point via any path except R_i or the C_f integrating capacitance, represents integration error, at least so far as the signal into R_i is concerned. The value of current represent the rate of error accumulation; the coulombs or ampere-seconds delivered represent the total error at the end of the operation. This becaomes evident when you remember that input current always equals feedback current, and the -grid is always at (or very close to) ground potential. If any part of the input current from R_i is balanced by current not flowing through the integrating capacitance C_f, then the voltage across C_f will be low at the end of the integrating interval.

The actual percentage rate of error accumulation at any time is the ratio of the current reaching the -grid via circuits other than C_f, to the total input current through R_i at that time. The accumulated error is the average error rate times the length of time it was allowed to continue.

For this reason, current reaching the -grid by means of the LF Reject circuit will introduce error; the object of our lag circuit is to delay the arrival of this error current as long as possible.

The next step in understanding the lag circuit is to examine the response of a passive network of series R and Shunt C to a ramp of voltage such as the first part of the integral of a rectangular pulse (before the reject circuit has had a chance to introduce any significant error).

The response of an RC circuit to a step function is quite familiar, of course. The voltage across the capacitor rises in an exponential fashion toward the value E of the step function, and its value at any time T after the step is $1-e^{-t/RC}$, or 1-1/e exp (t/RC). At the end of 5 time-constants, the value has become $1-e^{-5}$, which for many purposes can be considered as equal to 1.

The response of the same network to a continuing ramp of voltage is related. Instead of starting from an initial slope of E/RC and approaching the amplitude of the driving signal according to the exponential function, in this case the voltage across the capacitor starts with an initial slope of zero and approaches the slope of the driving ramp signal according to the exponential function. In doing so, it also approaches an instantaneous amplitude difference from the ramp which difference is the value reached by the ramp in RC seconds, and also approaches a time difference (for the same amplitude) from the ramp of RC seconds -- that is, if the driving ramp reaches 1 volt at X seconds, the charge on the capacitor would reach 1 volt at X + RC seconds, provided a few RC time-constants had elapsed between the start of the ramp and X. The expression for the amplitude at time t of the output voltage of this network is t/RC-1 + e exp (-t/RC). The relationships are drawn to scale in Fig. 3.

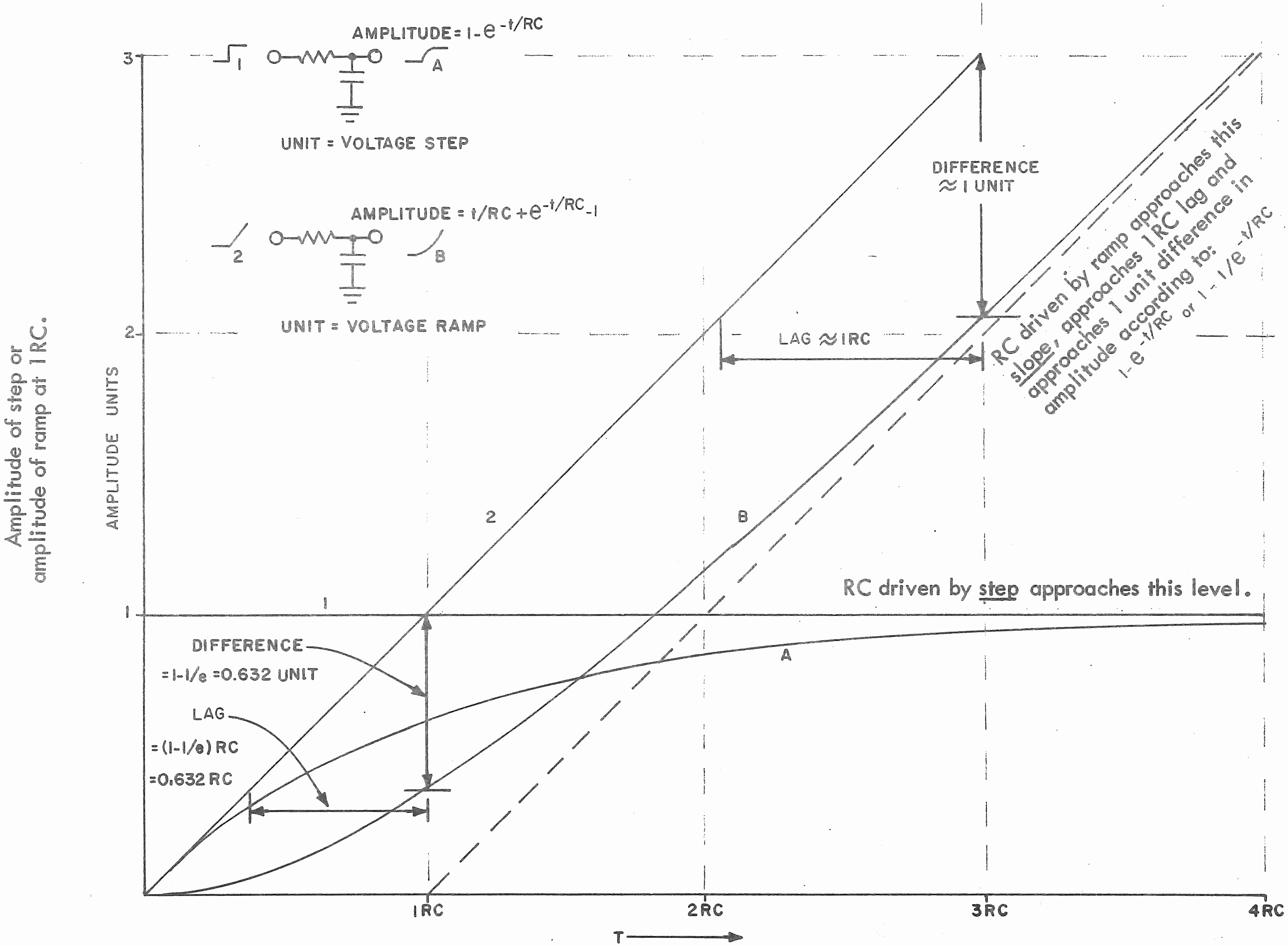


Fig. 3 Lag

3-25-64

TEK O IRB

Getting back to our integrator circuit (Fig. 2), it might be well to note that because of the 12-1 ratio between the LF Reject circuit resistance and R_i, we would be able to integrate fairly accurately over the first 100 msec integrating time-constant even if the "lag" capacitor were open. At the end of 100 msec, the output voltage would be 1.0 volt, and the feedback current to the -grid only 0.83 µa compared to the 10 µa current through R_i. The error rate after 100 msec would be 8.3%, but the accumulated error only about half that amount.

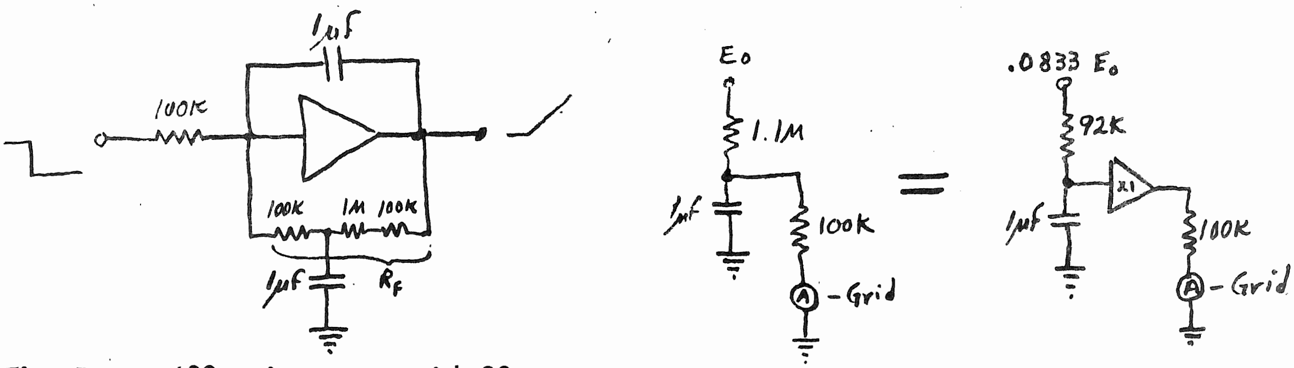


Fig. 2-A 100 ms integrator with 90 ms etc.

So we may safely assume that the LF Reject circuit will be driven by a fairly linear ramp at least over the first 100 msec. Since our chart is based on lag-circuit time-constants, we'll work in terms of these:

In 90 msec, our one-volt pulse will produce an output integral of 0.9 volts ($E_{in} \times T/RC$).

Using the equivalent network of Fig. 2A, we determine that the lag capacitor will see this as a ramp rising 0.083 x 0.9 v or 75 mv, in 90 msec. From Fig. 3, we find that the capacitor will charge to 36.8% of this value, or about 27-1/2 mv in the 90 msec, and be delivering error current to the -grid at a rate of 27.5 x 10^{-3}v/100 k, or 0.275 µa, for an error rate of 2.75%. However, this rate has not been maintained for the entire 90 msec; we can see from Fig. 3 that the average rate has been less than half this much.

The actual accumulated error is the integral of the current since the start of the waveform, and simply by counting squares on the chart, we can see that it is about 13% of 0.75 µa (0.083 x 0.9, or 75 mv corresponds to the amplitude unit of Fig. 3; 75 mv/100 k = 0.75 µa) times 90 msec, or 13% of 67.5 nanoampere seconds -- slightly under 9 na-sec, or less than 1% of the 900 na-sec delivered by R_i during this time.

You can see the benefit of the lag circuit. Without the capacitor, accumulated error after 90 msec would have been about 4% and the rate entering the second 90 msec about 8%; using the shunt capacitor in the reject network, the accumulated error is less than 1% and the error rate at 90 msec only 2-3/4%.

The table below shows the effects during the second, third, fourth and fifth time-constants. Note that after the third time-constant, the linearity of the ramp is sufficiently impaired that corrections must be made to allow for the effects of this curvature on the LF Reject circuit. Beyond about 5 lag time-constants, this method of approximation ceases to yield useful answers and direct analog computation (using the O-Unit, of course!) or more rigorous mathematical analysis is required.

TABLE 1

Lag Time Constant	1ST	2ND	3RD	4TH	5TH
Duration	90 msec	90 msec	90 msec	90 msec	90 msec
Ideal integrator output	900 mv	900 mv	900 mv	900 mv	900 mv
Seen by equiv. Reject ckt	75 mv	75 mv	75 mv	75 mv	75 mv
Error current to -grid (at end of TC)	0.275 μa	0.845 μa	1.54 μa	2.04 μa	3 μa
THIS TC:					
Unit charge (90 msec x 75 mv)	67.5 na-sec	67.5 na-sec	67.5 na-sec	67.5 na-sec	67.5 na-sec
Delivered to -grid (From Fig. 3)	13%	73%	158%	251%	350%
Charge to -grid (uncorrected)	8.8 na-sec	49 na-sec	106 na-sec	170 na-sec	235 na-sec
Input charge ($\int \frac{E_{in}}{R_i} dt$)	900 na-sec	900 na-sec	900 na-sec	900 na-sec	900 na-sec
Average error, this TC	1%	5-1/2%	12%	19%	26%
*Correction for output error	--	--	--	-1%	-2%
Corrected output error	1%	5-1/2%	12%	18%	24%
TOTALS:					
Total input charge since T_o	900 na-sec	1800 na-sec	2700 na-sec	3600 na-sec	4500 na-sec
Total error charge since T_o	8.8 na-sec	58 na-sec	165 na-sec	333 na-sec	548 na-sec
Accumulated error since T_o	1%	3-1/4%	6%	9%	12%
Accumulated error if lag removed	3.7%	7-1/2%	11-1/2%	14%	17%

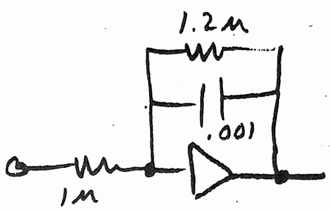
*Ballpark correction to allow for accumulated slope error in output.

SUMMARY:

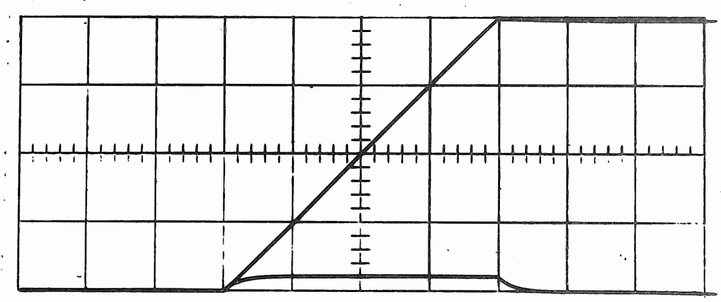
The Integrator LF Reject circuit aids integration by providing a dc feedback path in the integrator to minimize the effects of dc levels and grid-current. Because of the lag time-constant, the dc feedback is delayed, to postpone the arrival of error-producing current, allowing integration over longer periods using a given integrating time-constant and dc gain-limit with less accumulated error.

The effects were not exciting in the example used -- an accuracy improvement of 3-5% net over one to 5 lag time-constants. Normally, the LF Reject circuit would not be used beyond one lag time-constant; the example was chosen primarily to illustrate the lag principle.

Fig. 4 shows the more dramatic improvement in the case of a 1 millisecond time-constant and a dc gain of 1.2. The 90 msec lag (Fig. 4B) permits accuracy within 15% after 10 msec, while simple resistive dc feedback (Fig. 4A) causes an error of 80% in half that interval.

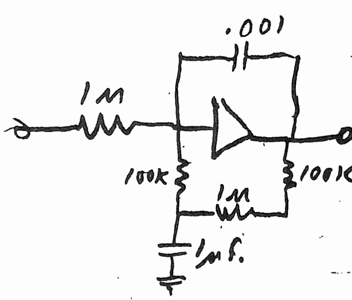


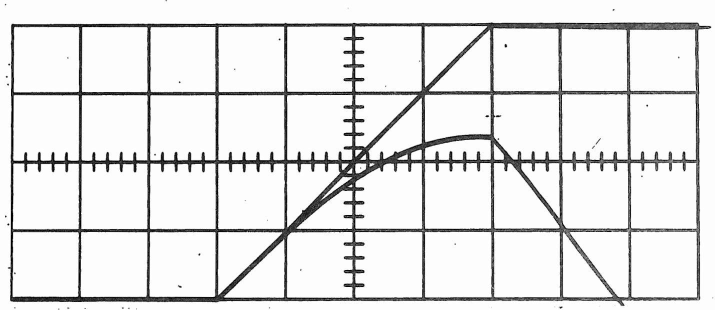
Integrating time constant = 1 ms integral of 1 V x 20 ms pulse vertical gain of 20



5 V/cm; 5 ms/cm

Fig. 4-A 1.2 M resistor used to limit DC gain to x 1.2 Upper trace: True integral (resistor removed). Lower trace: Output of "integrator". Error is 95% after 20 ms.





5 V/cm; 5 ms/cm

Fig. 4-B 1 cps LF Reject circuit. Upper waveform: True integral. Lower trace: Output of "integrator". Error is 42% after 20 ms, but negligible (<2%) below 5 ms.

It should be kept in mind that the integration of rectangular pulses represents a special (though typical) case, and the performance of the lag circuit will vary with different types of waveforms to be integrated.

In general, best performance will be obtained integrating waveforms which build up slowly to a maximum value and decay rapidly; poorest performance for waveforms which reach a peak rapidly and then decay slowly, particularly if the integrating interval represents a significant part of the lag time-constant.

It should also be kept in mind that our discussion has had primarily to do with <u>single shot</u> phenomena, or, if repetitive, of such low duty-cycle that the LF Reject circuits have had sufficient time (i.e., 5 to 10 R_fC_f time-constants) between repetitions to reach zero-signal equilibrium ("initial conditions") again. Special problems arising from certain signal repetition-rates are discussed below.

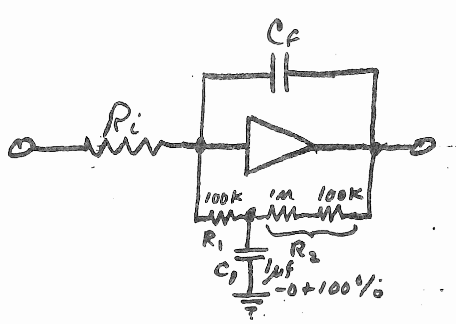
3. Specific Limitations of 100 msec and 100 µsec Circuits (O-Unit).

(a) Factors to Consider: R_i and DC Gain, Lag Time-Constant, Integrating Interval, Virtual Gain.

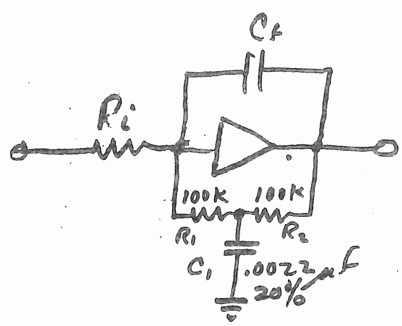
The four factors listed above all affect the performance of an integrator used with an LF Reject circuit. The integrating time-constant is implicit in these terms, it being the integrating interval divided by virtual gain.

Figure 5 shows the configuration of the two LF Reject circuits used in the O-Unit; they may generally be referred to as "100 msec" and "100 µsec" time-constants, the tolerance on the 1 µf capacitor being -0, +100% and the other components 10% (Resistors) and 20% (the .0022 µf capacitor).

In general, greatest integration accuracy is obtained by keeping virtual gain low compared to DC gain, and the integrating interval short compared with the lag time-constant.



(a) 1 cps LF Reject
$R_f = R_1 + R_2 = 1.2M$ Lag
Time constant $= R_1R_2C_1/(R_1+R_2)$
$=92$ ms + 110%, -10%
DC gain: 1.2 M/R



(b) 1 kc LF Reject
$R_f = R_1 + R_2 = 200k$
Lag time minus constant
$R_1 R_2 C_1/(R_1 + R_2) = 111\,\mu s \pm 30\%$
DC gain: 200k/R

Fig. 5 O-Unit LF Reject Circuits

(b) Specific Limits. The integrating interval may be allowed to approach the lag time-constant if the virtual gain (integrating interval $\div R_iC_f$) is kept less than the DC gain (R_f/R_i).

When the virtual gain is equal to the DC gain and the integrating interval equal to the lag time-constant, the output error will be in excess of 10% and will increase rapidly with any extension of the integrating interval.

For a given LF Reject circuit, integrating interval and selection of integrating components, it is possible to compute a ballpark error figure. Since the error function is not linear, though, it's easiest to approximate the probable error for the typical case by this formula:

$\frac{B\,T_L}{R_fC_f} \times 100\%$, where B is taken from Table 2 below, T_L is the lag time-constant, and R_f is the resistance of the LF Reject circuit.

For the O-Unit's 1 cps (100 msec) circuit, R_f is 1.2 M; for the 1 kc (100 µsec) circuit, R_f is 200 k.

TABLE 2

Integrating Interval / Lag Time-Constant	B (Basic error factor)
0.1 TC	.002
0.2	.0075
0.3	.016
0.4	.026
0.5	.037
0.6	.052
0.7	.07
0.8	.09
0.9	.11
1.0	.1325
1.1	.157
1.5	.375
2.0	.43
3.0	.84
4.0	1.25
5.0	1.7
6.0	2.15
7.0	2.65
8.0	3.1
9.0	3.6
10.0	4.1
N (>10)	4.1 + 0.5 (N - 10)

Since the formula does not take into account the fact that the reduced output due to the error tends to reduce the actual error, it becomes progressively inaccurate for indicated errors of over 20% or so.

The limitations covered above will be in addition to the errors due to R_i and C_f component tolerances, stray coupling, and the effect of the A-factor, which last will introduce 1% error at a virtual gain of 24 unless corrected for.

(c) Ringing and Rep-Rate Effects. The delayed low-frequency feedback, which enhanced operational accuracy during the start of the integration, was only delayed, not destroyed.

As the integrating interval approaches the value $\frac{\pi}{2}\sqrt{T_L T_R}^{*}$ (for the integral of a step), the error current through the LF Reject circuit will just equal input signal current, and the integral will stop increasing and start to decrease. The error current, because of the lag, is still increasing, however, and becomes positive feedback momentarily. Depending on the reset and lag time-constants, this positive feedback may cause the reset to overshoot, and then overshoot again in recovering from the initial overshoot (ringing).

The ringing frequency (which will be noted as a resonance in using sinewaves) will be approximately $\frac{1}{2\pi\sqrt{T_L T_R}}^{*}$. The frequency is actually "exactly" this amount, but component tolerances are so great that considerable deviation will be noted in actual units.

The Q of this ringing will be affected by both the ratio between T_L and T_R and by the value of R_i used. Table 3 shows typical figures for O-Unit LF Reject ringing characteristics. Note that when T_R approaches or exceeds the value of T_L, ringing disappears, since the amount of charge in the lag capacitor becomes much less than that in the integrating capacitor, and cannot drive the integrating capacitor very far in the way of overshoot.

The ringing is always damped when the internal LF Reject and integrating components are used. However, continuous ringing (oscillation) will be sustained whenever the input signal slope is in phase with the slope of a ring cycle from a previous signal. This creates not only annoying instabilities in the waveform level (vertical jitter) but may cause severe inaccuracies in the reading of any integral taken during the ringing period. Fig. 6 (AC integration of cal waveform following integration of a step) shows the effect. While it's obvious in the 10 msec/cm picture what's happening, it would be far from obvious at a faster sweep and with unknown, irregular waveforms (it's not often that the user is taking the integral of anything so simple and predictable as a cal waveform).

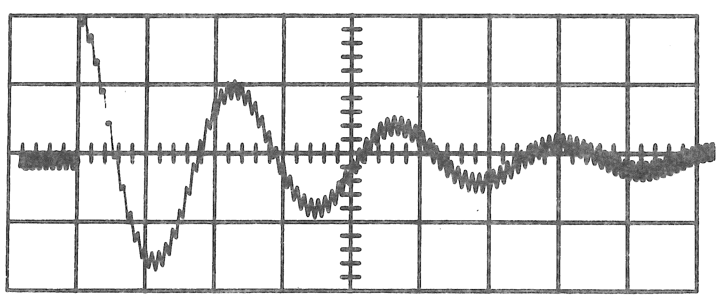


Fig. 6 LF-Reject ringing effect on later integrals. This is typical of an error that could be introduced by a large pulse preceding the waveform to be integrated. 10 ms/cm, .0001 µf, 1 cps reject.

The same sort of trap may be encountered when inadvertently using the LF Reject circuit and the gating adapter at the same time. If the "off" time of the gate is not long enough to discharge the lag capacitor fully, the remaining charge can introduce error into a subsequent "gated" integral. It is rare that use of a reject circuit is necessary when using the gating adapter.

*Where T_L is the lag time-constant and T_R is the reset time-constant -- the resistance of the LF Reject circuit and the integrating capacitor C_f.

TABLE 3

LF REJECT RINGING DATA

Z_f	Z_i	Ring Starts After	Frequency (Nominal)	Decay Time (to 10%)	Notes
1 CPS LF REJECT					
1 µf	Any	--	--	--	(No ringing)
0.1 µf	--	--	--	400 msec	(Reset overshoot only)
.01 µf	Any	55 msec	4.6 cps	700 msec	
.001 µf	1 M	17.5 msec	14.5 cps	500 msec	
	10 k	17.5 msec	14.5 cps	300 msec	
.0001 µf	1 M	5.5 msec	46 cps	160 msec	
	10 k	5.5 msec	46 cps	30 msec	Very 60 cps sensitive
10 pf	1 M	1.75 msec	145 cps	20 msec	Very 60 cps sensitive
	100 k-10 k	1.75 msec	145 cps	10 msec	
1 KC LF REJECT (No ringing 1 µf to .001 µf)					
.0001 µf	Any	75 µsec	3.4 kc	500 µsec	
10 pf	1 M	20 µsec	11 kc	400 µsec	
	100 k-10 k	20 µsec	11 kc	300 µsec	

(d) O-Units S/N 814 up -- LF Reject Switch Must Be "Off" When Using the +Grid. In order to maintain exactly the same capacitance between the output and the -grid in all positions of the LF Reject switch (a 3 pf change produced up to 30% integration error in the "off" position using C_f = 10 pf in S/N's 101-813), it was necessary to modify the O-Unit to leave half the "LF Reject" circuit hanging across the -grid in all the "R", "C" and "Ext" positions of the Z_f selector switch, except when the LF Reject switch is set to "off".

Since the LF Reject capacitors are isolated from the -grid by 100 k, and the -grid appears in most operations as a very low impedance, the R and C hanging across the grid do not have any appreciable effect.

However, when the +grid is used, the -grid frequently is required to move as much as ±10 v, and R and C shunting this grid will have a definite effect on operations. So whenever the +grid is used, be sure the LF Reject switch is set to "off" (exception: when the output is strapped directly to the -grid for a X1 non-inverting amplifier, this precaution is unnecessary).

(e) About the 1 µf 3 v capacitor 283-017? It may seem strange that a capacitor rated at 3 v is used in a circuit which theoretically can swing ±50 v. However, this is an unusual capacitor, and the actual voltage across it never exceeds about 2.85 v.

The reason is leakage. The 50 v that might be impressed across this capacitor is isolated by 1.1 M. The specification for this capacitor is 25k (min) leakage at 1.5 v, and 4k at 3 v DC. The unusual dielectric used in this capacitor increases leakage so rapidly with increasing voltage that no amount of voltage which the O-Unit output can impress across it via 1.1 Meg can generate more than 2.85 v drop. Typical leakage is on the order of 200-500 k at 1.5 v and 50 k at 2-1/2 v. The shunt leakage will cause an increase in DC gain for large excursions, since much of the DC feedback is shunted to ground.

The capacitor also has an unusual voltage coefficient, typically -10% per volt. As more voltage is impressed across it, its capacitance drops radically. For this reason, the behavior of the 1 cps reject circuit will vary somewhat with signal and DC voltage levels, due not only to the increased leakage (which tends to shorten its time-constant) but lower lag capacitance as well.

4. Selecting Components for External LF Reject Circuits.

The values provided by the O-Unit's two reject circuits may not be optimum for some applications; some suggestions follow for the selection of alternate values.

(a) Selecting Resistance Values:

(1) Total Resistance: The total resistance from the output to the -grid sets the available DC gain and also the reset time-constant. For greatest single-shot accuracy, the resistance should be large compared to R_i. For fast reset, for handling repetitive signals, the value must be made smaller.

(2) Isolation: To prevent the lag capacitor from loading down the -grid or the output, at least 10 k should isolate the capacitor from the output, and, if possible, 100 k isolate the capacitor from the -grid. A smaller isolation from the -grid is permissible if the R_i resistor is small ($<$100 k).

(b) Selecting the Lag Time-Constant. Using the table and formula on page 8, a lag time-constant can be selected to provide the desired degree of accuracy for a given value of C_f, R_f and integrating interval. The lag time constant is

$\frac{R_{f_1} R_{f_2} C_L}{R_{f_1} + R_{f_2}}$, where R_{f_1} is the resistance to the -grid from the lag capacitor and R_{f_2} is the resistance to the lag capacitor from the output. $R_{f_1} + R_{f_2}$ add up to the total R_f resistance selected in (a) (1) and (2) above.

The ringing frequency and ringing-start point can be determined from the formulae on page 9, though if the lag time-constant is close to the value of the reset time-constant, ringing will probably be negligible. If a proper compromise cannot be reached, see (c) next page.

(c) Special Circuits; Use of Input Blocking Capacitors, etc.

(1) Relay Circuit. Relays generally cannot be made to operate fast enough to be used as reset devices over a wide range, as can be the gating adapter. However, an LF Reject circuit operating in conjunction with a disabling relay can sometimes be used to good advantage. Such a circuit, using a fast-acting reed relay, is shown in Fig. 7. The reset circuit is operative only during the period between sweeps; the lag is only operative during the interval between the start of the sweep and the time the relay finally closes. The closure of the relay disables the entire reject circuit and discharges the lag capacitor, allowing the integrating interval to be as long as desired without error-signal interference, since none of the output current reaches the -grid except via C_f. The reed relay can be operated at repetition rates as high as several kc, covering most mid-range sweep rates.

(2) Input Blocking Capacitor: A capacitor may be placed ahead of R_i to block any DC signal component. Where this is done, swamping resistors to ground should be used (Fig. 8) to nullify any capacitor leakage, and the capacitor value should be much larger than C_f, since the R_iC_i time-constant will have the same effect on low-frequency accuracy as the R_fC_f time-constant encountered in shunting C_f to limit DC gain (Fig. 1B). Use of an input blocking capacitor alone is not generally recommended, since it does not limit drift due to grid-current.

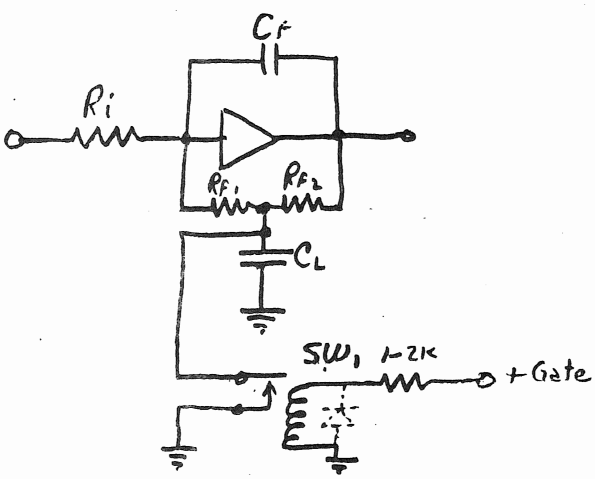


Fig. 7 Reed Relay (Sw-1) used to disable LF Reject circuit after start of sweep. R_{f1} and R_{f2} should be of low enough value to discharge C_f between sweep. Lag time-constant should hold error to desired minimum between sweep start and relay closure.

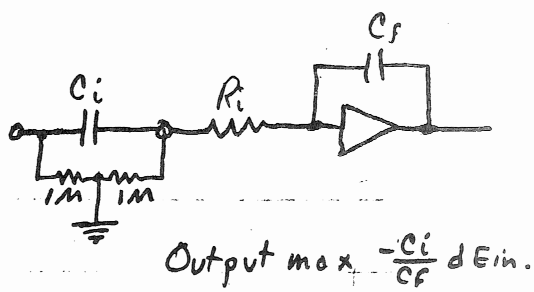


Fig. 8 Use of input blocking capacitor to block signal DC level. C_i should be much larger than C_f. Resistors to ground prevent capacitor leakage (>>1 MΩ) from affecting answer.

(3) AC Integration Using Blocking Capacitor Inside the Feedback Loop. An interesting alternative to the delayed DC feedback used in the LF Reject circuits is the circuit of Fig. 9. The blocking capacitor is put inside the feedback loop and a large value resistor used for grid-current rejection. DC response of the circuit is now zero. The circuit behaves very similarly to the LF Reject circuit, and offers no unique advantage beyond its greater DC rejection.

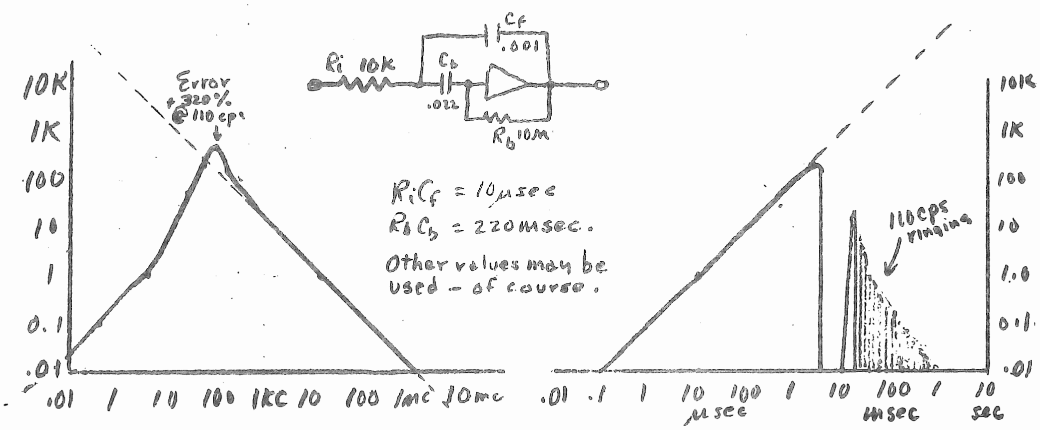


Fig. 9 Alternate scheme for AC integration. Capacitor C_b blocks DC signal components. Resistor R_b prevents integration of grid current. $R_b C_b >> R_i C_f$. Ring frequency is $1/(2\pi\sqrt{T_i T_b})$, and starts at $\pi/(2\sqrt{T_i T_b})$ for step. T_i is $R_i C_f$, T_b is $R_b C_b$. Circuit has advantages of zero gain at DC.

5. Gating Adapter.

Although this was not intended as a pitch for gating adapters (Tek No. 013-068), these -- or their equivalent -- are the appropriate answer whenever the following problems are beyond the means of LF Reject circuits to cope with:

(a) Very long integrating intervals with high values of R_i.

(b) Signal rep-rates likely to interfere with the reset-recovery characteristics of the lag circuits.

(c) Very rapid (μsec) reset.

(d) Integration of only a small part of a waveform.

(e) Integration of a waveform preceded by a transient which would "ring" the lag circuit.

6. Addendum

(a) Verification of LF Reject Circuit Performance. Neither the factory cal procedure nor the manual cal procedure at the present writing actually checks out the proper performance of the O-Unit's LF Reject circuits.

This can be done by checking to see that the circuits ring at approximately the right frequency under specified conditions, or by following the procedure below, which fits in better with the equipment on hand during regular calibration.

Equipment required:

Type 105
50Ω 10-1 T
50Ω Termination
O-Unit AC Coupler (Available from FMS, or see FCP for details).
UHF-BNC Adapters (depending on S/N's and hardware).

Initial Setup:

O-Unit Preamp: 100 mv/cm, A — or B —.
Operational Amplifier A or B: Z_i 1 M; Z_f 1 M; LF Reject 1 cps.

Apply 50 cps (exactly) from 105 to operational amplifier input via 50Ω T and termination and O-Unit AC-coupler.

Adjust 105 for just 400 mv (4 cm) display; adjust for best symmetry.

Set O-Unit preamp for 0.5 v/cm.

Switch Z_f to .001 μf. You should obtain just 2.0 v p-p of triangle waves, with no appreciable curvature. You may obtain just slightly more. If you obtain substantially less than 4 cm, the LF Reject circuit is defective. If you obtain about 0.5 cm, the 1 μf lag capacitor is open.

Return Z_f to 1 M and set 105 to 10.0 kc. Recheck symmetry and amplitude (400 mv) at 100 mv/cm.

Switch LF Reject to 1 kc and Z_f to .0001 μf. Set preamp to .05 v/cm. You should obtain 100 mv (2.0 cm) of triangle waves, again with little if any visible curvature. There will be a small amount of spike feedthrough, making the p-p slightly more than 2.0 cm. If the output is as low as 40 mv, the .0022 μf lag capacitor is open.

(b) Alternate Rule for Avoiding LF Reject Errors

A fairly good formula to use in holding errors due to the LF Reject circuit action to a minimum is related to the ringing period. As you noticed, in the integration of a step, the integrator output stops rising at $\frac{\pi}{2}\sqrt{T_L T_R}$, and error at this point is typically about 40%.

The expression $\sqrt{T_L T_R}$, when restated in terms of components, reduces to $\sqrt{R_{f_1} R_{f_2} C_L C_f}$, where R_{f_1} and R_{f_2} are the two components of the total feedback resistance, C_L is the lag capacitor, and C_f is the integrating capacitor.

For an integrating interval of $\sqrt{R_{f_1} R_{f_2} C_L C_f}$ seconds, error is about 20%. For $0.6\sqrt{R_{f_1} R_{f_2} C_L C_f}$, it's about 10%; for $0.3\sqrt{R_{f_1} R_{f_2} C_L C_f}$, 3% or so.

So keeping the integrating interval below $0.3\sqrt{R_{f_1} R_{f_2} C_L C_f}$ is a good rule of thumb for avoiding errors due to the LF Reject circuit.

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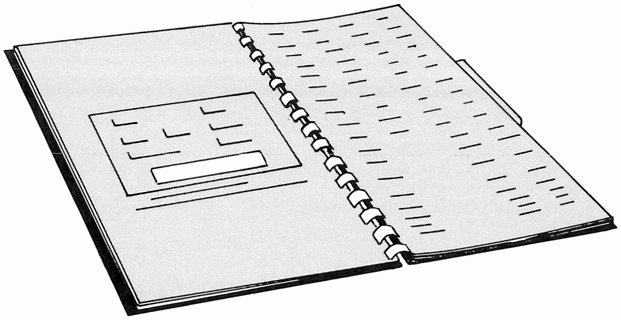


SECTION 2
OPERATING INSTRUCTIONS

General Information

The Type O Plug-In Unit is a very versatile device. The number of applications for the unit is limited only by the imagination of the user. To realize the full potentialities of the unit, it is important for you to understand the operation and function of each control. Much of this understanding will come only as the result of actual use. This section provides the basic information you will require.

Typical application examples are included in this section of the manual to aid you in becoming acquainted with the unit. These examples provide set-up instructions for integration, differentiation, and amplification.

Connecting The Plug-In Unit To The Oscilloscope

The Type O Unit can be used with any of the Tektronix Type 530-, 540-, 550-, or 580-Series Oscilloscopes. In the Type 530-, 540-, or 550-Series Oscilloscopes, the plug-in unit need only be inserted in the plug-in compartment of the Series Oscilloscope is used, the Type 81 or Type 81A Plug-In Adapter must be inserted into the plug-in compartment of the oscilloscope ahead of the plug-in unit. The plug-in unit can then be inserted into the compartment of the Type 81 or Type 81A. The plug-in fastener knob should be turned until tight to insure that the plug-in unit makes good connection with the oscilloscope.

PREAMPLIFIER

The Type O Unit contains a vertical preamplifier which is used in much the same manner as the preamplifiers in other vertical plug-in units. The preamplifier can be used with or without the operational amplifiers. When the preamplifier is used alone, input signals are connected to the EXT. INPUT connector on the front panel.

The VERTICAL DISPLAY switch selects the input signal used by the preamplifier. Possible selections are (1) external signals applied to the EXT. INPUT connector, and (2) output of either of the operational amplifiers. The + or — sign at each position of the VERTICAL DISPLAY switch indicates whether the oscilloscope display is normal or inverted. In the — positions of the VERTICAL DISPLAY switch the input signal is inverted before being displayed.

The preamplifier vertical deflection factor is controlled by the VOLTS/CM switch. Nine calibrated deflection factors from 0.05 to 20 volts per centimeter are provided. The VARIABLE control allows uncalibrated deflection factors between ranges and gives a continuous range of deflection factors between 0.05 and approximately 50 volts per centimeter. With the desired signal displayed on the crt, the VOLTS/CM controls are adjusted to give a convenient amount of vertical deflection.

External input signals to the preamplifier may be either ac or dc coupled depending on the position of the VERTICAL DISPLAY switch. In many cases only the ac component of the input signal is of interest. In such cases, use of ac coupling allows the display of signal information while blocking the dc component. Ac coupling also permits observation of ac information at high sensitivities without dc components deflecting the display off the crt.

Dc coupling must be used to observe very low-frequency signals (the ac response is 3-db down at approximately 2 cps). Dc coupling must also be used when measuring the dc component of the input signal or making measurements which include the dc component.

Input signals may be displayed with or without inversion when using either ac or dc coupling.

Output signals from the two operational amplifiers are dc coupled through the VERTICAL DISPLAY switch to the preamplifier. To ac couple the output of an operational amplifier, connect a short coaxial lead from the OUTPUT of the operational amplifier to the EXT. INPUT connector. Either of the AC positions of the VERTICAL DISPLAY switch can then be used.

First Time Operation

Initial operation of the Type O Unit requires that certain adjustments must be checked. Set both the A and B Operational Amplifier Z_i and Z_f controls at 1 MEG. Set the VERTICAL DISPLAY switch to +DC.

After inserting the plug-in unit in the oscilloscope and switching on the power, wait a few minutes for the instrument to warm up. Adjust the oscilloscope for a free-running sweep and, using the POSITION controls, position the trace on the crt. Set the intensity of the trace at a convenient level and adjust the oscilloscope FOCUS and ASTIGMATISM controls for a sharply focused trace.

OPERATIONAL AMPLIFIERS

With no input signal to the operational amplifiers, the output dc level of the amplifiers should be zero. To insure this condition, a check on the dc level should occasionally be made. The procedure is outlined in the following paragraph using the OUTPUT DC LEVEL switch in each amplifier. The OUTPUT DC LEVEL ADJ. control is used to set the output to zero.

In order to set or check the output dc level of the operational amplifiers, it is first necessary to determine the zero input dc level of the preamplifier. To do this, set the VOLTS/CM switch to .5 and press the ZERO CHECK switch. (The ZERO CHECK switch disconnects all signals to the preamplifier and permits only its dc level to be displayed on the crt.) Then use the POSITION control to position the trace to a convenient horizontal graticule line. The zero input dc level of the preamplifier then corresponds to this graticule line.

When the zero input dc level line of the preamplifier has been determined, set the VERTICAL DISPLAY switch to A+. This connects the output of the A operational amplifier to the input of the preamplifier. Place the OUTPUT DC LEVEL switch in the ADJ. position and hold it there. Set the ADJ. control to position the trace on the crt to the zero input dc level line previously determined. This sets the output level of the A operational amplifier at zero volts. When the OUTPUT DC LEVEL switch is pressed to the ADJ. position, the external circuit is disconnected, the input end of the Z_i component is grounded, and a gain of 100 is automatically provided in the operational amplifier. The 100× gain permits a more precise adjustment to be made. It is important, however, to recognize that any large amount of drift and noise is due to the extra gain.

Now set the VERTICAL DISPLAY switch to B+ and adjust the dc level at the output of the B operational amplifier in the same manner as for the A amplifier.

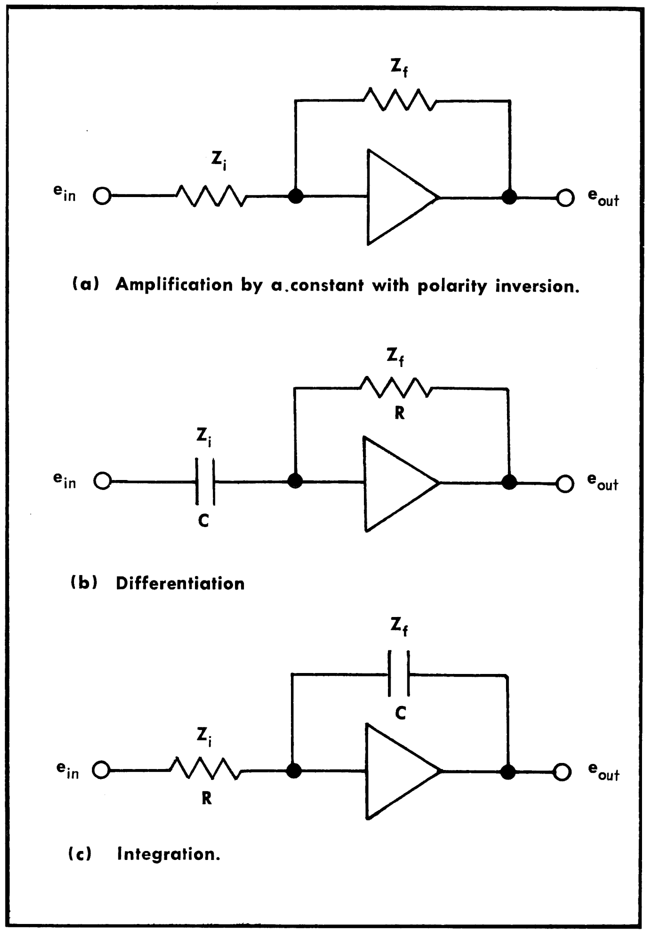


Fig. 2-1. The three most basic uses of an operational amplifier.

Feedback Controls

The most basic functions of an operational amplifier are those of amplification by a constant, integration, and differentiation. In amplification, both the input and feedback impedances are normally resistors (although capacitors or inductors can be used). The ratio of the feedback resistor to the input resistor detemines the gain of the feedback amplifier. In integration, the feedback impedance is a capacitor while the input impedance is a resistor. In differentiation, the input impedance is a capacitor while the feedback impedance is a resistor. In both integration and differentiation, the time constant of the feedback network determines the characteristics of the amplifier. The basic circuits for these three types of operations are shown in Fig. 2-1.

The front-panel controls labeled Z_i and Z_f select the input and feedback impedances from several possible internal values. External positions for these controls also permit connection of desired external values to the jacks on the front panel of the unit. Paralleling internal components with external components is practical; however, it is possible to perform any of the following basic operations using the internal values supplied with the unit.

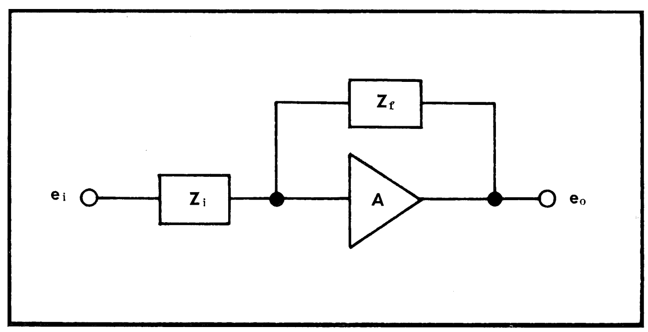


Fig. 2-2. Major terms used in general gain expression for an operational amplifier.

Operational Amplifier Gain

The gain of the operational amplifier with resistance input and feedback elements is the ratio of the feedback resistance to the input resistance: Gain $= -R_f/R_i$. This ratio is selected by means of the Z_i and Z_f SELECTOR controls or by externally mounted components. (See Fig. 2-2)

If the operational amplifier had infinite gain, the accuracy of the input and feedback resistors (to establish a given gain) would determine the accuracy of the gain ($-R_f/R_i$). However, the O Unit operational amplifier open-loop gain is 2500, not infinite, so a small error is introduced at higher closed-loop gain. When using internal input and feedback resistors, this error is internally compensated at ×50 and ×100 gain settings.

Error Factor

The gain of a feedback (operational) amplifier with finite gain may be expressed by the formula:

$$\frac{e_o}{e_i} = -\frac{Z_f}{Z_i} \quad \frac{1}{1 - \frac{1}{A}\left(1 + \frac{Z_f}{Z_i}\right)}$$

where e_o = output voltage, e_i = input voltage, Z_f = impedance of feedback component, Z_i = impedance of input component, and A = amplifier open-loop gain.

Since the first part of the formula ($-Z_f/Z_i$) corresponds to the expression for the closed-loop gain of an operational amplifier with infinite open-loop gain, the remainder of the formula is the Error Factor.

An example of the need for gain correction due to the Error Factor follows: If Z_i is set at .01 MEG and Z_f at 1 MEG, the gain would be —100 if the operational amplifier open-loop gain were infinite. Since the open-loop gain is 2500, the error for this example will be approximately 100/2500 or 4%.

A second example is with Z_i set at .01 MEG, and Z_f set at .5 MEG. The gain should be —50, but the error will be approximately 50/2500 or 2%.

To keep the feedback gain within the O Unit (using internally selected input and feedback resistances) to a tolerance of 3%, special precautions were taken with the Z_i and Z_f SELECTOR switches. The 3% gain tolerance is made up of a 1% limit due to the Error Factor and a 2% limit due to the 1% tolerance of the resistors. Therefore, the Z_i and Z_f switches have been wired to place a shunting resistor across the .01 MEG Z_i resistor when the Z_f control chooses a resistor that will give a gain error over 1%.

The schematic diagrams of the Operational Amplifiers show the .01 MEG Z_i resistors shunted with 240 k when the Z_f resistor is 1 MEG, and shunted with 510 k when the Z_f resistor is .5 MEG. The Z_i—Z_f switches do not shunt the .01 MEG Z_i resistor at any other time.

When using external input and feedback components, it may be important to use the Error Factor to correct for gain errors.

Closed-Loop Gain-Bandwidth Characteristics of Operational Amplifiers

Open- and closed-loop gain-frequency characteristics of the Type O Unit Operational Amplifiers are shown in Chart 2-1. As mentioned in the Characteristics (Section 1), the Type O Operational Amplifiers have an open-loop gain-bandwidth product of approximately 15 mc. The open-loop gain-frequency characteristics set a maximum theoretical limit for the closed-loop characteristics. When using internal Z_i and Z_f resistors, the closed-loop bandwidth is always somewhat less that the theoretical limit due to wiring and stray capacitances.

Average closed-loop gain frequency characteristics in various closed-loop gain settings are shown in Chart 2-1. Since the individual Z_i and Z_f stray capacitances change with switch settings, the data presented in the chart may not fit all switch combinations for the same gain settings.

In a critical application, external capacitive compensation is recommended to reduce the effect of internal stray capacity and extend the closed-loop bandwidth.

Capacitive compensation is accomplished by placing external variable capacitors across the input and output resistors. With the oscilloscope Amplitude Calibrator signal applied to the input, the capacitors may be adjusted for optimum flat top response while observing the output waveform on the crt. Optimum response will be obtained when the time constant of the input circuit is equal to the time constant of the feedback circuit. Typical compensated unity-gain amplifier response is shown by the dotted line in Chart 2-1.

Sample Amplification Setup

1. Connect the output of the oscilloscope Amplitude Calibrator to the A INPUT connector on the front of the O Unit. Adjust the calibrator for a 1-volt signal.
2. Connect the output of the A operational amplifier to the preamplifier by setting the VERTICAL DISPLAY switch to A+.
3. Set the ±GRID SEL switch to (—).
4. Adjust the A operational amplifier to give a gain of —1 by setting both the Z_i and Z_f controls to 1 MEG.
5. Set the VOLTS/CM switch to .5, VARIABLE control to CALIBRATED.

You should now have 2 centimeters of vertical deflection. Thus with 1 volt in, —1 volt out is obtained, resulting in a gain of —1. Note that whenever the Z_i and Z_f SELECTOR controls are both set to the same value of resistance, a gain of —1 is obtained.

Next set the VOLTS/CM switch to 5, Z_i to .1 MEG, and Z_f to 1 MEG. The ratio of Z_f to Z_i is 1/0.1 = 10, so a gain of —10 should be obtained. There should now be 2 centimeters of vertical deflection. This corresponds to an output of 10 volts. The gain is —10 as calculated. Try other settings of the Z_i and Z_f controls where the ratio of Z_f and Z_i is 10. You will see that in each case the gain of the amplifier is —10.

Differentiation

In differentiation, a capacitor is used as the input component while the feedback component is a resistor. It is similar in some respects to a simple RC differentiation circuit except that the high gain of the amplifier allows differentiation to be obtained without loss of signal level.

In differentiation, the output voltage from the operational amplifier is given approximately by the equation

$$e_o = -R_f C_i \frac{d}{dt}(e_i)$$

where e_o is the output voltage, R_f is the feedback resistance, C_i is the input capacitance, e_i is the input voltage. The output voltage varies directly with the time constant and the rate at which the input voltage changes with respect to time.

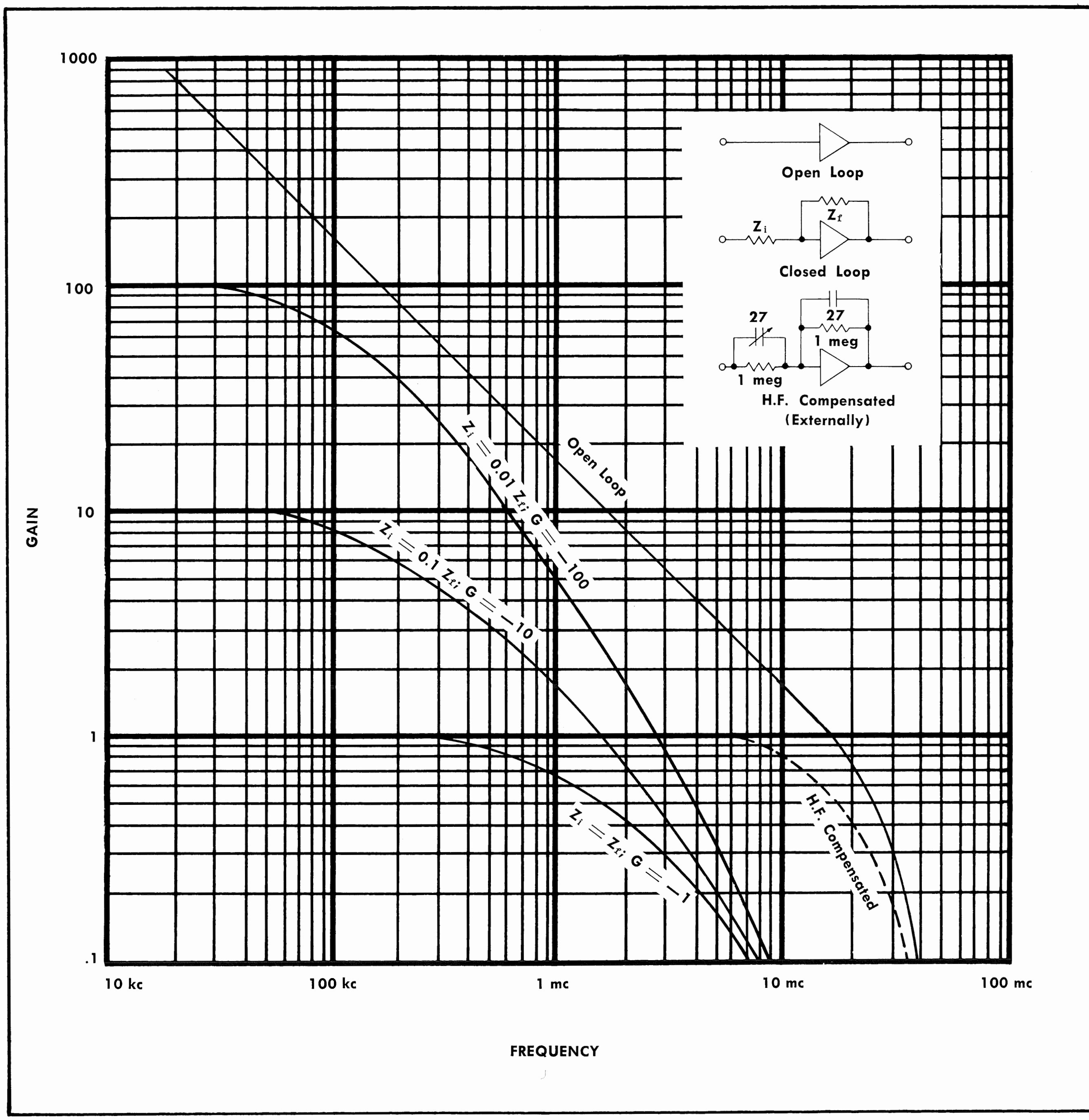


Chart 2-1. Average closed-loop and open-loop Gain-Frequency characteristics for the operational amplifiers. The dashed line indicates an example of external capacitance compensation extending the high-frequency performance.

A good starting point is to choose a time constant equal to the fastest risetime of the signal you are attempting to differentiate. The oscilloscope calibrator has approximately a 1-μsec risetime. Thus with the calibrator signal, an RC time of 1 μsec should be selected as a starting value. The VOLTS/CM switch can then be set to permit the display to be viewed. If necessary, the time constant can be changed to permit a better display. (Normally, the smaller values of capacitance will produce better results.)

Sample Differentiation Setup

1. Connect the oscilloscope Amplitude Calibrator to the A INPUT connector and adjust the calibrator for a 1-volt output.
2. Set the Z_i control to 10 pf and the Z_f control to 0.1 MEG. These values produce a time constant of 1 μsec.
3. Set the ±GRID SEL switch to (—).

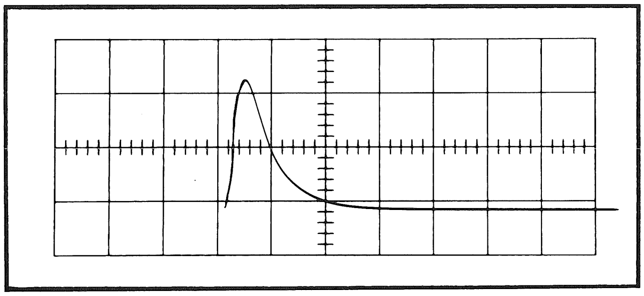


Fig. 2-3. Differentiated oscilloscope-calibrator signal; 1-μsec/cm sweep rate.

4. Set the VOLTS/CM switch to .5 and the Oscilloscope TIME/CM switch to 1 μSEC.

The display should be a differentiated waveform, as shown in Fig. 2-3.

Observe the effects of other RC combinations. It is important to use the lower capacitance values in order to minimize circuit loading. There will be minor differences in the waveforms obtained with various RC combinations. This is true even for simple RC differentiators that produce the same time constant.

Integration

In integration, the input component is a resistor while the feedback component is a capacitor. This is analogous to the simple RC circuit. When set up for integration, the output (e_o) of the operational amplifier is approximately

$$e_o = -\frac{1}{R_i C_f}\int e_i \, dt.$$

The output voltage is inversely proportional to the time constant and directly proportional to the integral of the input voltage.

Sample Integration Setup

1. Connect the oscilloscope Amplitude Calibrator to the A INPUT connector and adjust the calibrator for a 1-volt output.

2. Set the Z_i control to 1 MEG, Z_f to .001 μf, and the INTEGRATOR LF REJECT switch to 1 CPS.

3. Set the $\pm$GRID SEL switch to (—).

4. Set the VOLTS/CM switch to .2 and the oscilloscope TIME/CM switch to .5 mSEC.

The display should be an integrated calibrator signal, as shown in Fig. 2-4.

The time constant of the values chosen is 1 millisecond, the period of the Amplitude Calibrator waveform. Try other values of R and C which produce the same time constant. Then try other time constants to see the effect of changing the time constant.

With a good integrated calibrator waveform displayed, set the INTEGRATOR LF REJECT switch to OFF. The oscilloscope trace will probably deflect off the screen. If this happens, it is because of dc components in the input signal and/or inherent drift in the integrator. The 1 CPS position of the INTEGRATOR LF REJECT switch reduces dc offset due to integration of low frequency signals. This allows normal ac components to be integrated while permitting the trace to remain on the screen. For signals with a significant dc component, use ac coupling. A 1 KC position of the INTEGRATOR LF REJECT switch is also provided. The 1 KC position permits elimination of low-frequency noise and power line hum pickup from integration in the medium- to high-frequency range.

Gain-Frequency Characteristics of Differentiator and Integrator

As stated in the Characteristics section, the Type O Operational Amplifiers have an open-loop gain-bandwidth product of 15 mc. This means the open-loop gain will drop from 2500 at low frequencies to unity at 15 mc.

It is also important to know the gain-frequency characteristics for both integration and differentiation. Chart 2-2 illustrates the gain-frequency characteristics for most Z_i and Z_f control settings for both integration and differentiation. This information can be used by the operator to avoid inaccurate measurements from erroneous waveforms due to gain-bandwidth limitations for each mode of operation.

Output Connections

Each of the three basic operational amplifier applications just described were employed by the oscilloscope crt. The Type O Unit front panel has two OUTPUT connectors, one of which is suited for coaxially connecting any application output signal to an external device.

The OUTPUT terminal located to the left of the Z_i-Z_f SELECTOR switches is available for connection of external feedback components that can augment the internal values of the Z_f selector switch. This OUTPUT connector is in parallel with the coaxial OUTPUT connector located to the right of the Z_i-Z_f SELECTOR switches. Either of these OUTPUT terminals can be used for external connection of one operational amplifier to the other operational amplifier INPUT. Or, the coaxial OUTPUT connector can be conveniently used to drive a 50 Ω coaxial cable to a fairly remote system requiring the use of one of the features of an O Unit operational amplifier. The output impedance is low enough to drive a 50 Ω coax with a reasonably good im-

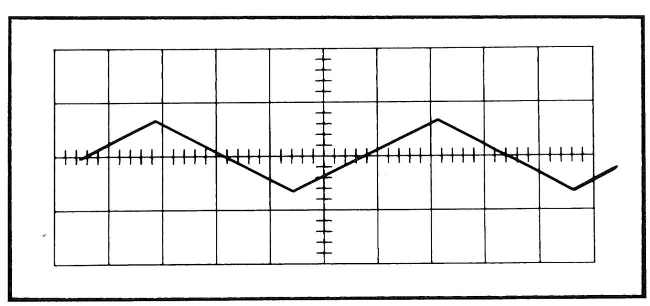


Fig. 2-4. Integrated oscilloscope-calibrator signal; 0.5-msec/cm sweep rate.

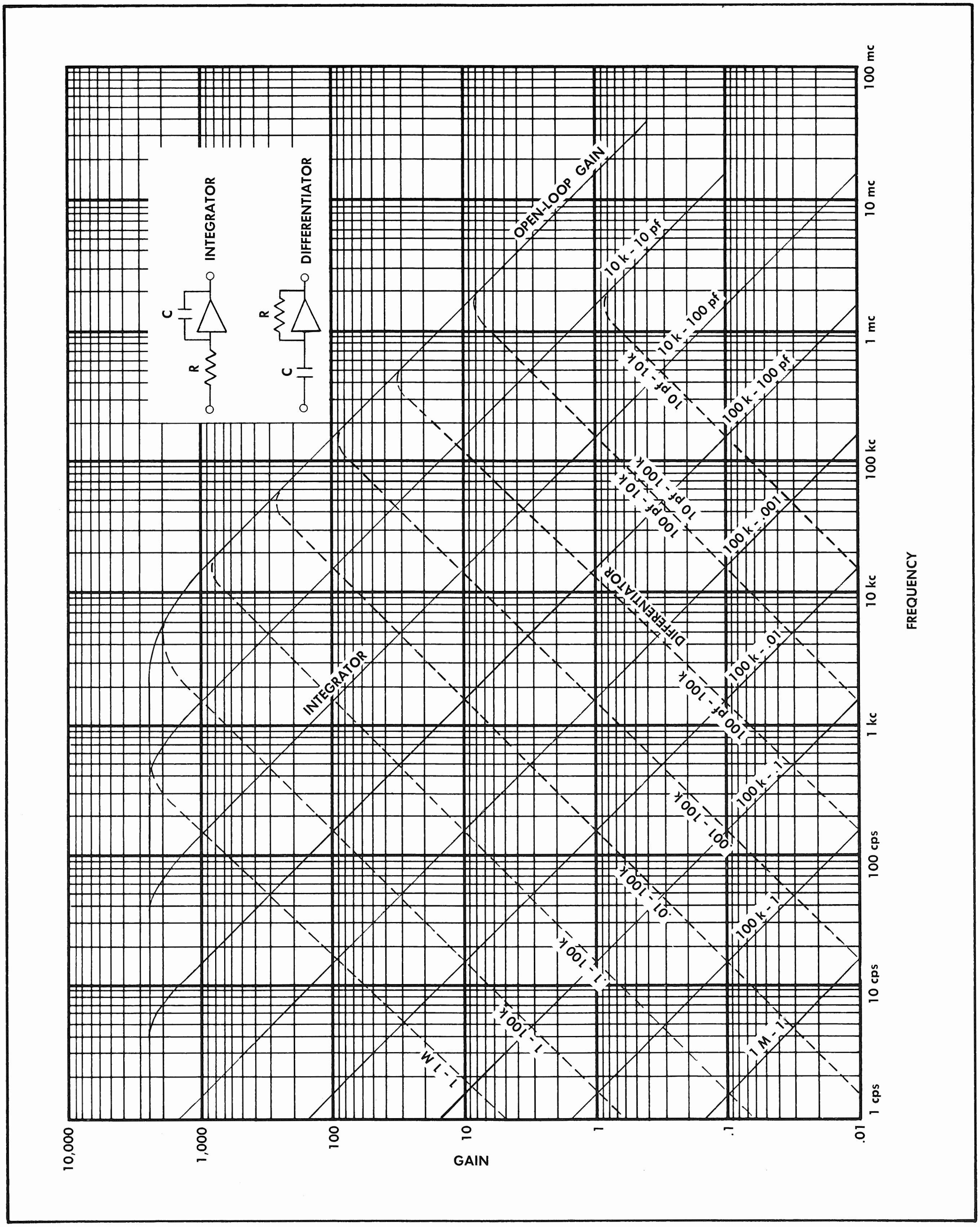


Chart 2-2. Average Gain-Frequency characteristics for integration and differentiation.

pedance match throughout the frequency response limits of the operational amplifier.

Signal Connection Precautions

Certain precautions should be taken in connecting signals to the input of the operational amplifiers to assure good results. First, when dealing with low level signals it may frequently be necessary to use shielded leads in order to minimize stray pickup. This is particularly important when the unit is used for differentiation. High-frequency noise is particularly troublesome with differentiation since the output of the differentiator is proportional to frequency. Whether shielded leads are required for a particular application can usually be determined from the resulting oscilloscope display.

Precautions for connecting signals to the preamplifier are similar to those which must be observed for the operational amplifiers. When using only the preamplifier, avoid errors in readings by guarding against stray electric or magnetic coupling between circuits, particularly in the input signal leads. In general, unshielded leads of appreciable length are unsuited for this use. This is true even in the audio-frequency range. Shielded input cables are recommended for signal measurements when signals are obtained from a high-impedance source, or when leads are long. When shielded leads are used, the shields should be securely grounded to the chassis of both the signal source and the oscilloscope.

In broadband applications, it may be necessary to terminate signal cables to prevent ringing and standing waves in the cable. The termination is generally placed at the oscilloscope end of the cable, although some sources also require a termination at the source end.

As nearly as possible, simulate actual operating conditions in the equipment being tested by permitting it to work into a normal load.

Consider the effect of loading upon the signal source due to the input circuit of the preamplifier. The input impedance of the preamplifier is 1 megohm to ground, paralleled by 47 pf. With a few feet of shielded cable, the capacitance may well be 100 pf. Where the effects of these resistive and capacitive loads are not negligible, it may be necessary to use a probe to lessen their effect.

Use of Probes with the Type O Unit

Standard Tektronix probes can be used with the preamplifier of the Type O Unit. When used, the probes must be connected to the EXT. INPUT connector.

When probes with 10× or more attenuation are used, they must first be properly compensated for high-frequencies. This compensation is most easily accomplished using the oscilloscope Amplitude Calibrator signal.

To compensate the probe, first obtain a display of the calibrator signal on the crt. Then adjust the probe compensation control for the squarest possible corners on the displayed waveform. This condition results when the undershoot or overshoot has been reduced to a minimum.

The attenuation factor of the probe must be considered when measurements are made. The vertical deflection factor of the O Unit is effectively increased by the attenuation factor of the probe. When a 10× probe is used with a VOLTS/CM switch setting of 5, the actual deflection factor is 50 volts per centimeter.

Probes reduce both capacitive and resistive loading on the signal source. They also permit larger signals to be displayed than would otherwise be possible.

Attenuator probes are not normally used with the operational amplifiers, because of the variable input characteristics. A special case, however, is illustrated in Fig. 2-5. This application is used when it is necessary to obtain the signal from a high-impedance circuit and deliver it to a low-impedance load without attenuation. To permit the use of a 10× probe, the operational amplifier input must look like the input to the preamplifier; that is, 1 megohm paralleled by 47 pf.

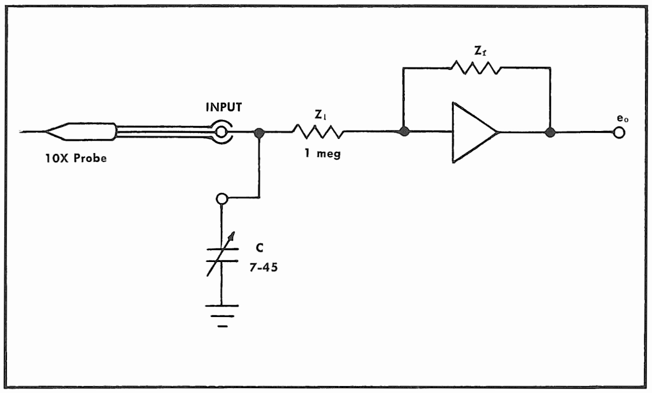


Fig. 2-5. Special case of a 10X probe used with an operational amplifier.

Fig. 2-5. shows an internally selected Z_i of 1 megohm, with an externally connected variable capacitor. By using an external feedback resistor (Z_f) of 10 megohms, the amplifier gain will be 10, although the probe attenuation of 10 makes the overall gain unity.

Using the oscilloscope Amplitude Calibrator, the external capacitor can be adjusted to compensate the system for optimum response (as explained earlier).

This system now appears as a dc-coupled 10-megohm probe, without attenuation, having a very low output impedance.

The OUTPUT connector can be coaxially connected to an external system requiring the low output impedance probe just described.

Connection of Signals to the ± GRID Jack

The ± GRID jack can be used to connect signals to either input grid of the operational amplifiers. When the ± GRID SEL switch is in the (—) position, signals connected to the ±GRID jack are applied to the —grid; when the ±GRID SEL switch is in the (+) position, signals connected to the ±GRID jack are applied to the +grid. (Signals connected to the —GRID jack are always applied to the —grid, regardless of the ±GRID SEL switch setting.) Signals applied to the —grid are inverted at the output of the amplifier; signals applied to the +grid are not inverted.

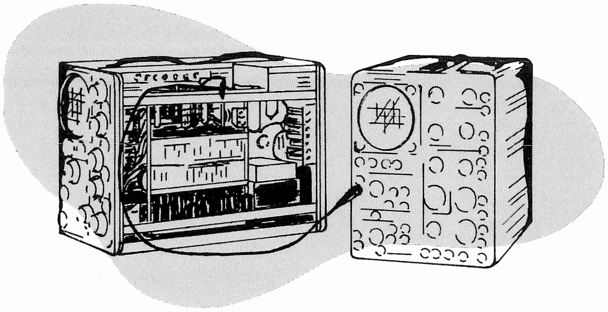
When using the +grid, both grids can be active, as in a differential amplifier. Three ways the —grid can be connected (when using the +grid) are: (a) 100% feedback from the output to the —grid; (b) —grid grounded; and (c) a feedback element connected from the output to the —grid. Each of the three —grid circuits affects the operational amplifier characteristics when driving the +grid. It is safe to say that when the ±GRID SEL switch is at (+), both the — and + grid circuits must be planned and properly connected.

Input signals applied to either grid through either the —GRID or ±GRID jacks bypass the internal input impedances and are therefore not affected by settings of the Z_i control.

Access to the +grid increases the versatility of the Type O Unit over conventional operational amplifiers and permits its use in certain applications where it could not otherwise be used. Thus external components can be used to provide positive rather than negative feedback to the operational amplifier. Some applications for using the + grid input to the operational amplifiers are described in the Applications section of this manual.

Connecting External Feedback Components

It will occasionally be necessary to mount external feedback components on the O Unit. This is necessary because of the limited number of components which may be mounted internally. A convenient means for mounting these parts is through the use of the adapter board received with your Type O Unit. The parts can easily be installed on the adapter board and the connectors can then be inserted in the jacks on the front panel of the O Unit. A shield should be placed over the adapter board to prevent stray signal pickup. The standard 3/4" spacing of the jacks on the front panel of the O Unit also permits use of double banana plug connectors, such as General Radio Type 274-B. See the end of the Applications Section of this manual for additional component mounting devices.



SECTION 3

APPLICATIONS

Operational Amplifiers

An operational amplifier is a very high-gain, direct-coupled amplifier having out-of-phase input-output characteristics (180° phase shift) which permits negative feedback. Since the open-loop gain of the amplifier is very high, closed-loop characteristics can be controlled by feedback components within the frequency and gain limits of the amplifier. (See Error Factor discussion on pages 2-3, and 4-1.)

The output level of the operational amplifiers is normally at dc ground so that two or more may be cascaded to perform successive operations.

Normally, resistors and capacitors are used as input and feedback components. By selecting proper feedback networks, many operations including linear amplification by a constant, summation of two or more signals, and integration and differentiation of voltage waveforms with respect to time, can be performed.

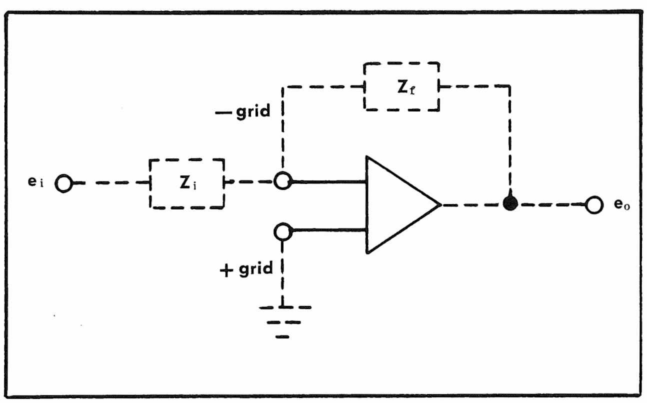


Fig. 3-1. Operational amplifier symbols.

A symbol frequently used for an amplifier is a triangle with the vertex pointing in the direction of signal flow (see Fig. 3-1). This symbol is used throughout the manual to represent the operational amplifiers. Note that Fig. 3-1 is drawn with both a + grid and a — grid. Signals applied to the — grid are inverted at the output, while signals applied to the + grid arrive at the output with the same polarity. The — grid is normally used in an operational amplifier because the inverted output permits the use of negative feedback through the Z_f feedback components. When not used, the + grid is grounded and will be omitted from diagrams.

The most basic form of an operational amplifier, illustrated in Fig. 3-2, is a time insensitive circuit that includes both negative and positive feedback. While many versions of the basic form can be used in time insensitive circuits, the four most common forms are: (1) an input voltage generates an output voltage; (2) an input voltage generates an output current; (3) an input current generates an output voltage; and (4) an input current generates an output current. These four basic Operational Amplifiers are illustrated in Fig. 3-3, and various forms appear in the applications that follow.

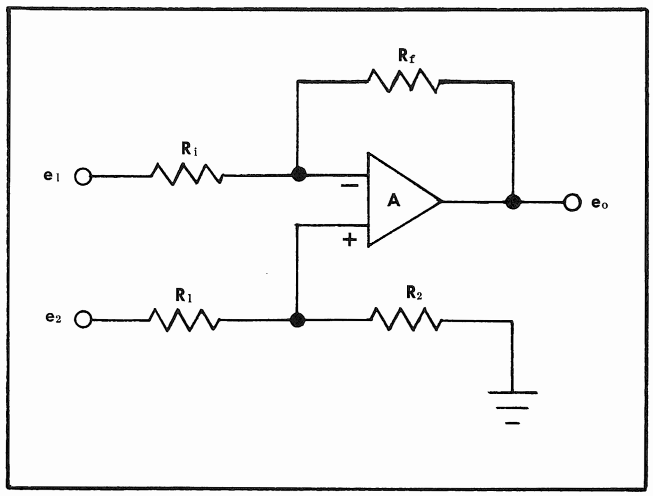


Fig. 3-2. Basic form of operational amplifier with both negative and positive active grids.

Virtual Ground

The dc level at the — grid input of an operational amplifier is very close to ground. When an input signal is applied, the signal tends to move the grid away from ground. However, the negative feedback from the output of the amplifier resists this tendency. The amount that the — grid voltage varies with a signal is dependent on the open-loop gain of the amplifier; the higher the gain, the less the — grid voltage varies. With the high open-loop gain normally encountered in an operational amplifier, the — grid voltage varies only slightly under closed-loop conditions. Therefore, it is convenient to assume that for all practical purposes the — grid voltage does not change with signal.

Since the — grid voltage remains essentially constant with input signal changes, it appears as though the — grid

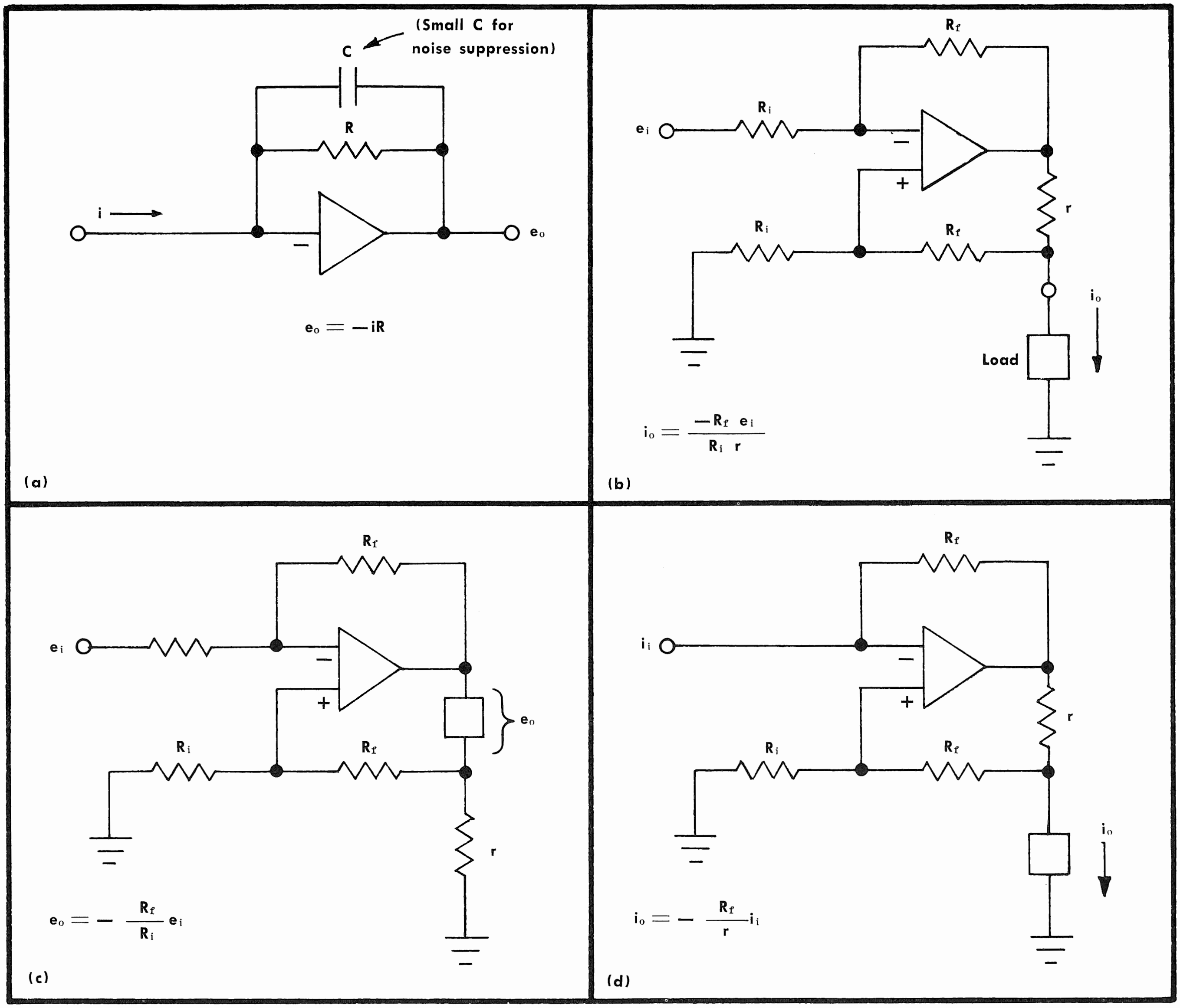


Fig. 3-3. Operational amplifiers in Voltage-Current conversions. (a) $i_{in}/-e_{out}$; (b) $e_{in}/-i_{out}$; (c) $e_{in}/-e_{out}$; (d) $i_{in}/-i_{out}$.

is grounded. Thus, a virtual ground can be considered to exist at the — grid input. The word "virtual" is used to indicate that although the input of the amplifier appears to be grounded, it actually is not. Many equations for the functions performed by an operational amplifier can be most easily derived by use of the concept of a virtual ground.

It should also be noted that since a virtual ground exists at the — grid, the input impedance of the amplifier is essentially determined by the value of the Z_i component.

Formulas and their derivation for finding the true input impedance at the — grid are presented in Section 4 of this manual.

TYPICAL APPLICATIONS

The remainder of this section explains typical applications for the O Unit operational amplifiers. The circuit for each application is shown, along with a brief discussion of the circuit and method of use. Most circuits are typical only, and not necessarily the only configurations that can be used. Modifications of these basic circuits may be employed to adapt the O Unit for individual needs.

The derivation of the equations relating to these applications will be found in Section 4. Equations indicating the operation of a particular circuit were derived assuming infinite open-loop amplifier gain. In most cases this assumption is quite good and calculations will be very close to conditions actually obtained. In practical applications it may be necessary to consider several factors such as open-loop and gain-frequency characteristics, stray capacity, and the output impedance of the signal source. The input impedance of the operational amplifier is really a combination of the signal source impedance and the impedance selected by the Z_i control. If the output impedance of the signal source is low, it can usually be ignored.

Applications Index

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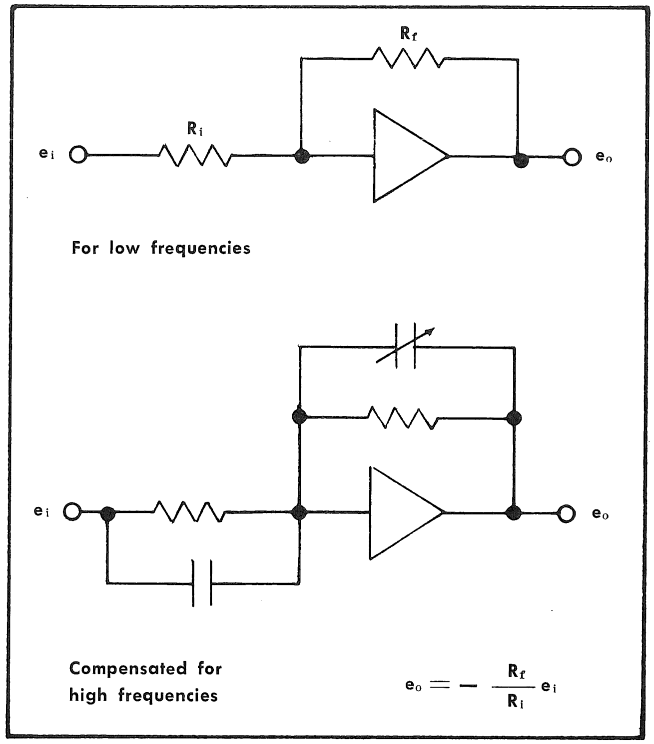
Description and Additional Information

1. Amplification by a Constant

This circuit is useful for providing amplification by a desired constant. The closed-loop gain is determined by the feedback components within the frequency and gain limits of the amplifier. Either internal or external feedback components may be used to provide the desired constant gain.

The output of the amplifier in this configuration is inverted. This makes the amplifier useful as a sign changer, with or without amplification. Applications 5 and 6 show circuits which can be used if inversion of the input signal is not desired.



1. Amplification by a Constant

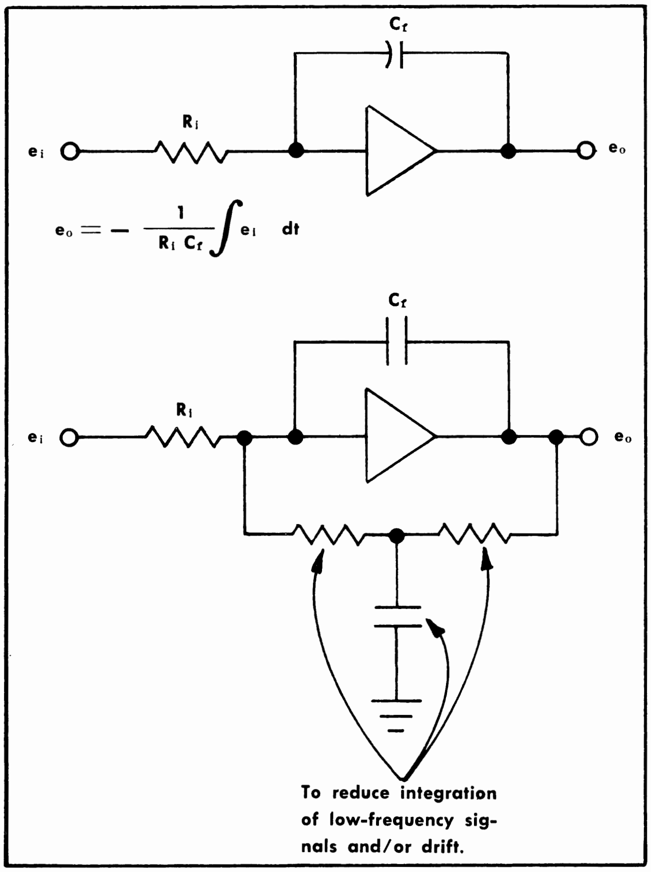
The circuit can be made to provide variable gain by replacing the feedback resistor with a potentiometer. As the potentiometer setting is changed, the gain of the amplifier will also change. Replacing R_f with a potentiometer will allow the input impedance of the amplifier to remain essentially constant while permitting the gain to be varied.

The fixed resistors may also be replaced by thermistors, photoresistors, or other variable-resistance elements. The gain will then be a function of the temperature, light level, or other variable.

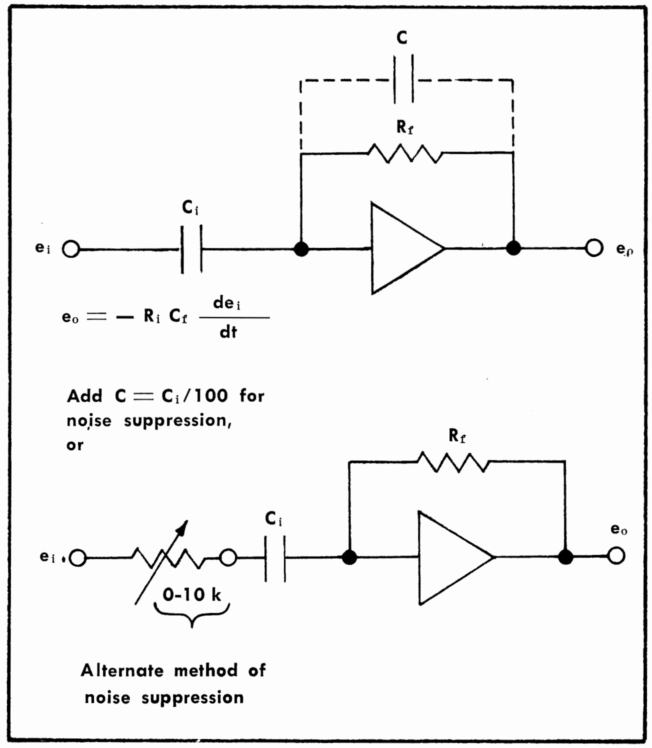
2. Integration

Integration of various signals can be accomplished by means of the Application 2 circuit. Feedback elements shown in the diagram may be selected from internal values or connected externally. The output voltage from the integrator is inversely proportional to the time constant of the feedback network and directly proportional to the integral of the input voltage. A good starting point in any integration application is to make the R_iC_f time constant approximately equal to the period of the signal to be integrated.

This basic circuit integrates not only the ac components of signals, but the dc components and drift as well. Integration of dc components or amplifier drift will cause the trace to gradually drift off the crt. To prevent this condition when integrating repetitive signals, an integration low-frequency rejection circuit is incorporated in each operational amplifier of the Type O Unit. This circuit prevents the integration of component signals with frequencies lower than approximately 1 cps or 1 kc, depending on the setting of the INTEGRATOR LF REJECT switch.



2. Integration



3. Differentiation

The integrator circuit can be used as a precision 90° phase shifter within the frequency limits of the amplifier. Sine waves applied to the input of the integrator are shifted in phase by exactly +90°.

In some applications it may be desirable to start and stop the integration at a known time interval. This type of integrator is discussed as the Gated Integrator, Application 22.

3. Differentiation

The output voltage of the differentiation circuit is inversely proportional to the feedback time constant and is directly proportional to the time rate of change of the input voltage. In practical applications, the R_fC_i time constant will have to be chosen somewhat on a trial and error basis to obtain a reasonable output level. A good starting point, however, is to make the time constant approximately equal to the risetime of the signal to be differentiated.

Differentiation permits slight changes in input slope to produce very significant changes in the output. An example of the usefulness of this feature would be in determining the linearity of a sweep-sawtooth waveform. Nonlinearity results from changes in the slope of the waveform. Therefore, if nonlinearity is present, the differentiated waveform indicates the points of nonlinearity quite clearly. Repetitive waveforms with a rise and fall of differing slopes can show erroneous waveforms. Under such conditions it is best to view the waveform using the oscilloscope SINGLE SWEEP mode.

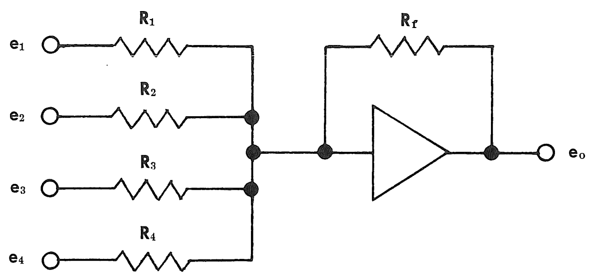
The output voltage of the differentiator is directly proportional to the frequency of the input signal (within the frequency limits of the circuit). This permits the differentiator to be used as a frequency to voltage converter. The frequency of an unknown signal can be determined by comparing the amplitude of the output voltage to that obtained using a known input frequency. The oscilloscope graticule can be calibrated, if desired, for frequency per centimeter. A constant-amplitude input signal must be used in this application, to prevent changes in amplitude from disturbing the measurement.

Differentiation of a sine wave results in a 90° phase shift at the output of the differentiator. The circuit can thus be used as a precision 90° phase shifter within its frequency limits. The output of the differentiator is shifted by —90° as opposed to +90° for the integrator circuit.

In general, differentiation accentuates high-frequency noise; if this is objectional, a noise suppression capacitor may be placed in the feedback circuit, or a resistor in the input circuit, to limit the high-frequency response above the signal frequency.

Conversely, in some applications a differentiator may be advantageously used to detect the presence of distortion or high-frequency noise in the signal. A differentiator can often detect hidden information that could not be detected in the original signal.

4. Summation



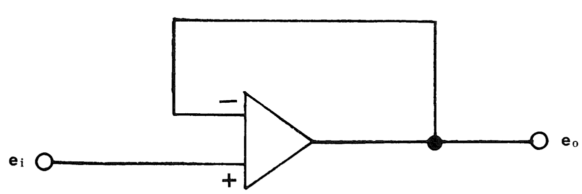
$$e_o = -\left(\frac{R_f}{R_1} e_1 + \frac{R_f}{R_2} e_2 + \frac{R_f}{R_3} e_3 + \frac{R_f}{R_4} e_4\right)$$

When $R_1 = R_2 = R_3 = R_4 = R_f$,
then $e_o = -(e_1 + e_2 + e_3 + e_4)$

4. Summation

Summation of a number of voltages can be accomplished with the Application 4 circuit. The feedback resistor R_f can be selected from internal values, however it will be necessary to provide external input resistors. When the values of all resistors are the same, the output of the amplifier is the inverted sum of all of the input voltages. By proper resistor selection, many input voltages can be amplified at the output, limited only by the ability of the output to swing either ±50 volts or ±5 ma.

5. Unity-Gain Amplifier with High Input Impedance



$e_o = e_i$

5. Unity-Gain Amplifier with High Input Impedance

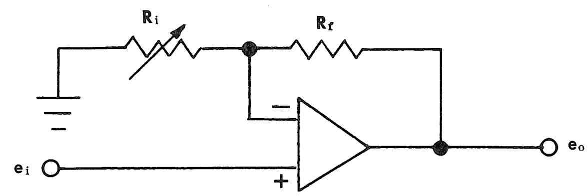
NOTE

The following method of determining input impedance signifies a concept only, and is not necessarily valid for all circuit configurations.

In this circuit, a gain of +1 is obtained. The input signal is applied directly to the +grid with feedback applied from the output to the −grid. Since the signal is applied directly to the +grid, and since there are no resistance elements between it and ground, the input impedance of the circuit is determined primarily by the grid current. This current is on the order of 0.3 nanoampere (see Characteristics Section) for signals of less than about ±10 volts; thus the input impedance is on the order of 10,000 megohms. For signals of ±1 v, the input impedance is on the order of 1000 megohms. This is, of course, the dc impedance. For ac signals, the shunt capacitance brings the impedance down to a much lower level. The input impedance varies with the input voltage (assuming $R_i = e_i/i_g$) since the grid current (i_g) remains relatively constant..

The feedback to the −grid insures a gain of 1 at the amplifier output.

6. High Input Impedance Amplifier



$$e_o = \frac{R_i + R_f}{R_i} e_i$$

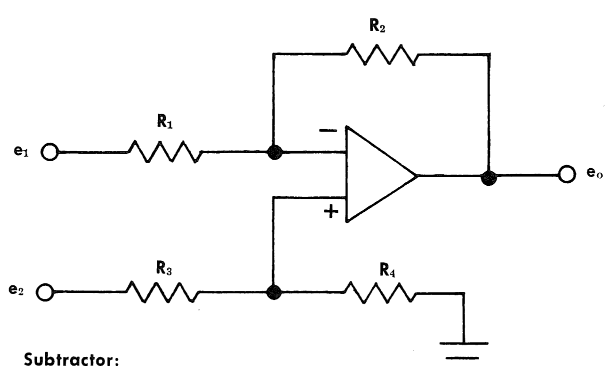
6. High Input Impedance Amplifier

The high input impedance feature of the previous circuit is combined with gain in this circuit. The gain is positive and is determined by the formula shown with the diagram.

In this circuit, the signal is again applied directly to the +grid with feedback from the output to the −grid. The amount of feedback is controlled by the values of R_i and R_f. Note that it is not possible to obtain a gain of less than one with this circuit.

Both R_i and R_f can be selected from internal values. If internal values are used, it will be necessary to externally ground the A INPUT connector in order to ground one end of R_i.

7. Subtractor and/or Difference Amplifier



Subtractor:

$$e_o = -\frac{R_2}{R_1} e_1 + \left[\frac{R_4}{R_3 + R_4}\right]\left[\frac{R_1 + R_2}{R_1}\right] e_2$$

Difference (special case)
When $R_1 = R_2 = R_3 = R_4$,
then $e_o = e_2 - e_1$

7. Subtractor and/or Difference Amplifier

One signal voltage can be subtracted from another through simultaneous application of signals to both grids of the amplifier, as shown in the preceding diagram. The signal applied to the — grid is subtracted from the signal applied to the +grid.

If the values of the resistors are not all the same, the gain of the amplifier for signals applied to the +grid is

$$\frac{R_1 + R_2}{R_1} \times \frac{R_4}{R_3 + R_4}.$$

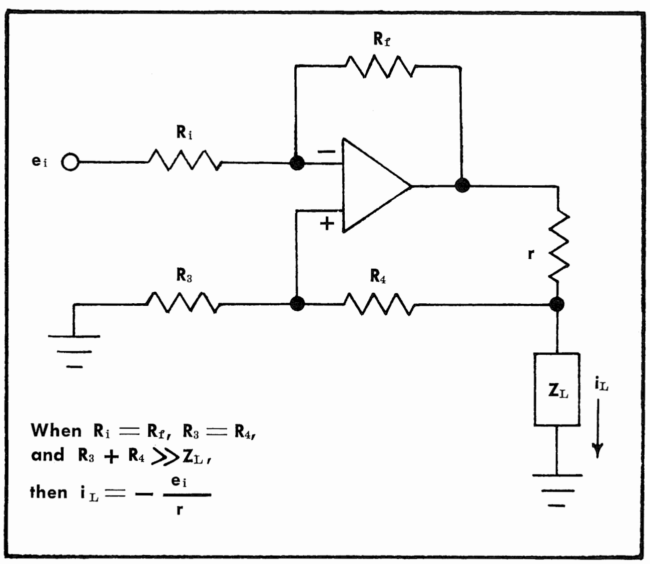
Gain of the amplifier for signals applied to the — grid is $-R_2/R_1$. Through use of these two equations, a desired amplification can be combined with the subtractive process shown by the generalized expression,

$$e_o = -\frac{R_2}{R_1} e_1 + \frac{R_1 + R_2}{R_1} \frac{R_4}{R_3 + R_4} e_2.$$

In this application, the amplifier is used essentially as a difference amplifier.

A difference amplifier may be operated with compensated frequency response by adding small variable capacitors across R_3 and R_4 of the preceding circuit. This permits balancing the time constants, extending the usable bandpass of the difference amplifier. In cases where all resistors are not equal, compensation for high frequencies may be accomplished by making all time constants equal.

8. Voltage to Current Converter (Transadmittance Amplifier)



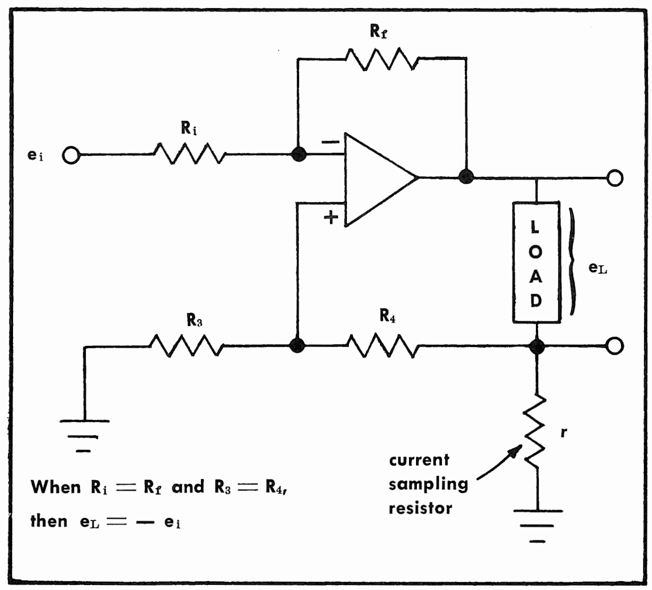
8. Voltage to Current Converter

This circuit can be used to supply a current to a load which is proportional to the voltage applied to the input of the amplifier. The current supplied to the load is relatively independent of the load characteristics. This circuit is essentially a current feedback amplifier. Load current is limited to ±5 ma.

A current sampling resistor is used to provide the feedback to the +grid. When $R_i = R_f = R_3 = R_4$, the feedback maintains the voltage across r at a value $-R_f/R_i \times e_i$ regardless of the load. If a constant input voltage is applied to the input of the amplifier, the voltage across r also remains constant regardless of the load. If the voltage across r remains constant, the current through r must also remain constant. With R_3 and R_4 normally much higher than the load impedance, the current through the load must remain nearly constant regardless of its impedance.

The values of R_i, R_f, R_3, and R_4 should normally be the same (1 megohm for each is satisfactory). The current sampling resistor is then selected for the desired load currents. The current through the load $i_L = e_i/r$ milliamperes per volt of signal when r is expressed in kilohms. The value of r should be selected to limit the maximum current drawn from the operational amplifier to less than 5 ma.

9. Voltage to Voltage Amplifier (Voltage Gain Amplifier)



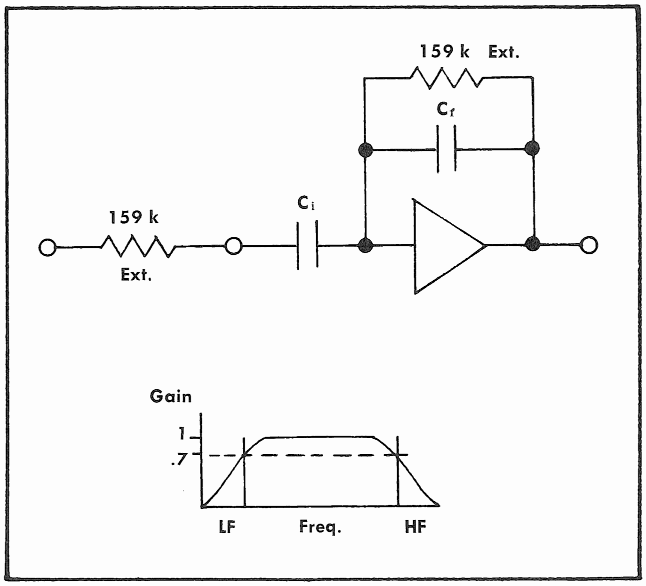
9. Voltage to Voltage Amplifier

This circuit is similar to the voltage to current converter described previously, except that the load is now placed where the current sampling resistor was in the previous circuit. With $R_i = R_f = R_3 = R_4$, the feedback to the + grid maintains the voltage across the load equal to minus the input voltage regardless of load (within the current limitations of the amplifier). Operation of the circuit is essentially the same as that of the voltage to current converter.

Since the voltage across the load is equal to the input voltage, sweeping the input voltage results in the voltage across the load also being swept. A voltage proportional to the current through the load can be obtained across r. The combination of the voltage across the load and the current through r can then be used to display the characteristic curves of devices such as tunnel diodes.

When using this circuit, care should be taken in adjusting the input voltage and/or load impedance so that the ±5 ma rating of the operational amplifier is not exceeded.

10. Bandpass Amplifier



10. Bandpass Amplifier

An operational amplifier can be applied as a bandpass amplifier with proper input and feedback elements. The adjacent circuit illustrates an example with only the addition of two external resistors. The C_i and C_f capacitors are internally chosen. The bandpass curve shown with the circuit illustrates an RC rolloff at each end of the bandpass with the signal amplitude dropping 6 db per octave. To make such a system guassian, with a rolloff of 12 db per octave after the 3 db down point, two bandpass amplifiers can be cascaded.

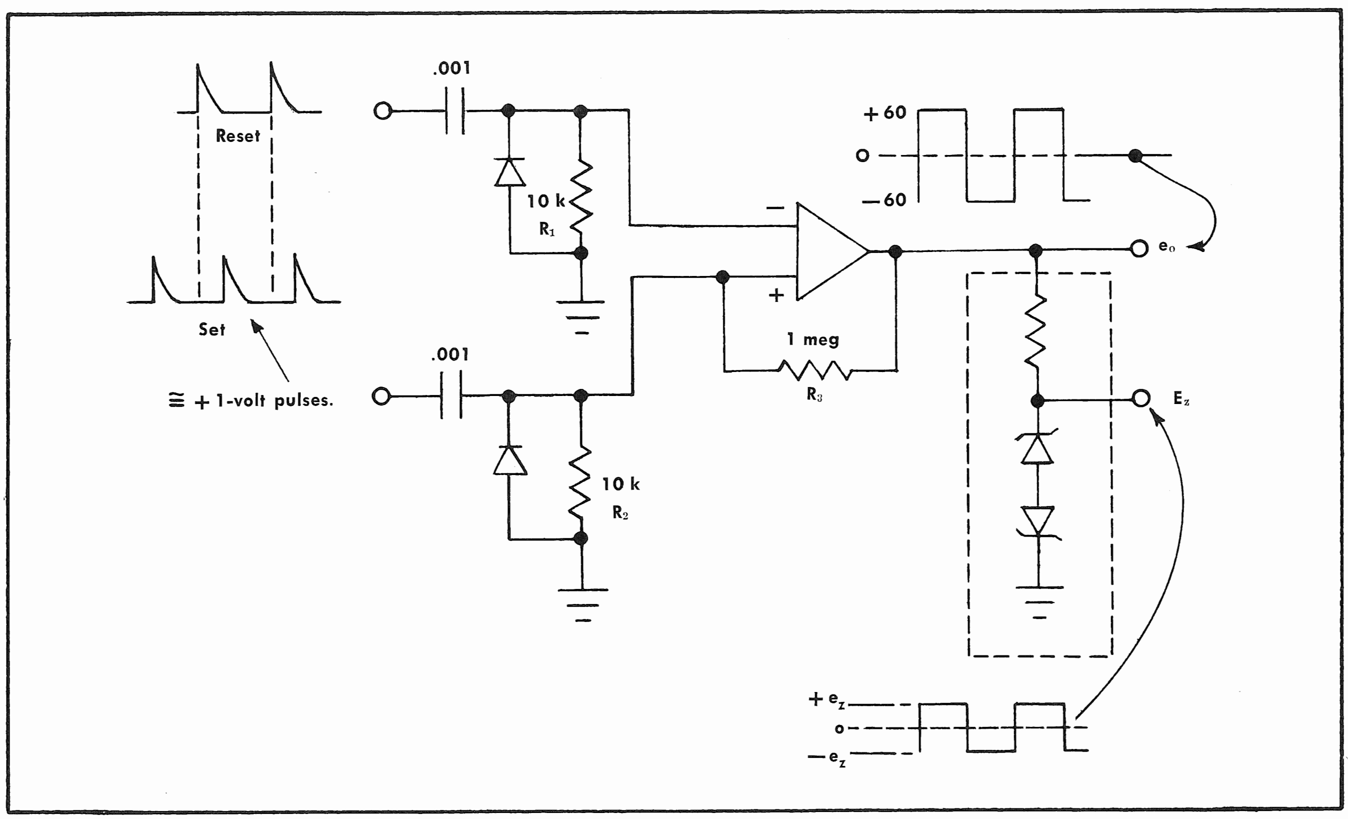
The principle of operation of the bandpass amplifier is that the input series R and C attenuate low frequencies, and the feedback parallel R and C attenuate high frequencies. The maximum gain, in the flat area of the frequency response curve, is unity with the following suggested values:

C_i and C_f, for —3-db frequency values, when external resistors are 159 k.

L. F.	C_i	*H. F.*	C_f
1 c	1 μf	100 kc	10 pf
10 c	.1 μf	10 kc	.0001 μf
100 c	.01 μf	1 kc	.001 μf
1 kc	.001 μf	100 c	.01 μf
10 kc	.0001 μf	10 c	.1 μf
100 kc	10 pf	1 c	1 μf

11. Flip-Flop Multivibrator

In the flip-flop multivibrator circuit, +1-volt or greater pulses are applied to both the + and — grids. When



11. Flip-Flop Multivibrator

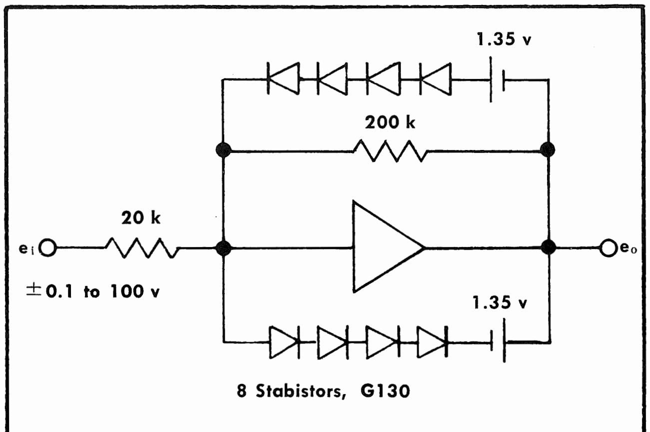
the multivibrator is flipped to one state of conduction by application of a pulse to one of the grids, it remains in that state of conduction until a positive pulse is applied to the other grid.

The output of the multivibrator is a series of positive and negative pulses approximately 50 volts in amplitude. This circuit is useful to approximately 10 kc. The sensitivity of the circuit can be changed to satisfy application requirements. The circuit can be made more sensitive by increasing the ratio of R_3 to R_2.

Waveforms shown on the circuit drawing as e_o are idealized. To obtain flat tops and bottoms on the output waveforms use a double Zener output circuit, as indicated by the dotted area below the e_o terminal.

12. Logarithmic Amplifier

In many applications it is desirable to have a device whose output is proportional to the logarithm of the input. However, a practical amplifier cannot give a true logarithmic response because of two primary limitations: (1) the logarithm of zero is —infinity, thus a true logarithmic amplifier would have an output of —infinity with zero input; (2) the logarithm of a negative number is not defined, therefore a true logarithmic amplifier could not accept negative input signals.



$e_o = -[0.7 + (0.35 \log_{10} e_i)]$

12. Logarithmic Amplifier

An approximate logarithmic output can be obtained from the indicated circuit; the relation between input and output is shown in the equation under the diagram.

The stabistors shown provide feedback which produces the logarithmic output response for input voltages between ±0.1 and ±100 volts.

13. Logarithmic Amplifier (fast response)

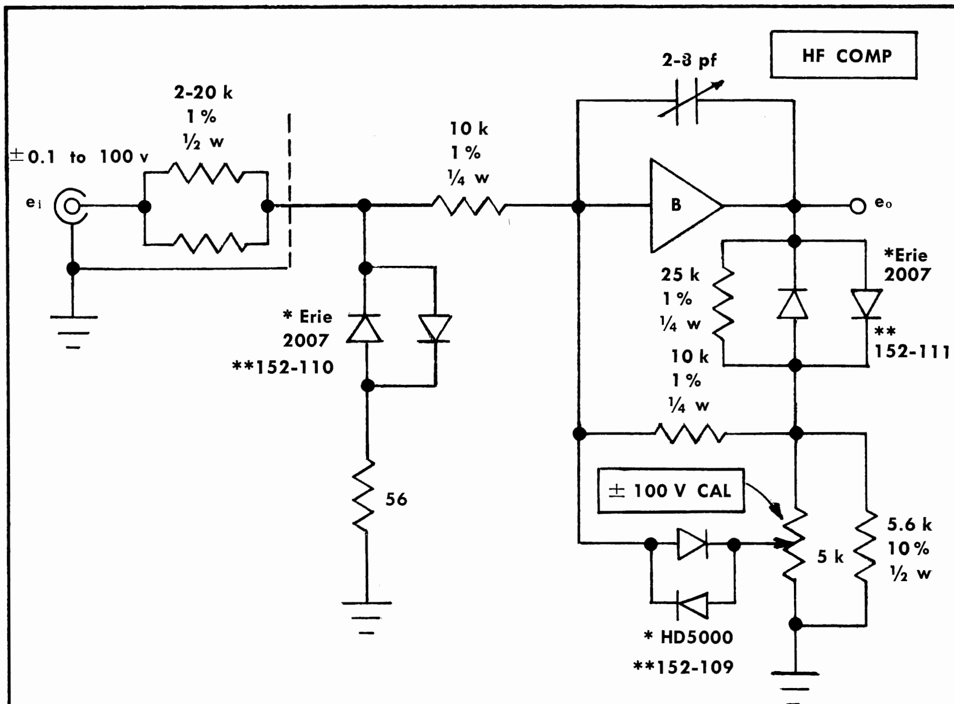
This circuit performs with a much faster logarithmic response than circuit 12. It operates to a higher frequency, employing high-frequency components and circuitry.

The amplifier is essentially logarithmic from ±0.1 volt to ±100 volts. It is not logarithmic with signals less than about ±0.08 volt.

When calibrated, the amplifier has a gain of 3 for very low-level signals. At a signal level just under 0.1 volt, the gain is approximately 1.4, and continues to drop logarithmically with increased signal.

Calibration is accomplished by using the oscilloscope AMPLITUDE CALIBRATOR and a Type 105 Square-Wave Generator with a 93 Ω cable and termination as signal sources.

1. Let the O Unit warm up for 10 to 15 minutes. Set the ±GRID SEL to —. Plug the logarithmic amplifier into the B operational amplifier.



* Matched pairs
**Tektronix Part Number

13. Logarithmic Amplifier (fast response)

2. Free run the sweep at about .5 mSEC/CM.
3. Set the VERTICAL DISPLAY switch to B—.
4. Place the Z_i and Z_f controls in the EXT. position.
5. Adjust the OUTPUT DC LEVEL to match the ZERO CHECK trace position.
6. Set the VOLTS/CM switch to .1
7. Connect a 1-volt calibrator signal to the input of the logarithmic amplifier.
8. Adjust the VARIABLE VOLTS/CM control for two centimeters of crt display. Recheck step 5 to be sure the zero-volt portion of the display coincides with the ZERO CHECK trace position. (A little drift will alter the calibration for low level signals.)
9. Raise the calibrator signal to 100 volts. Adjust the 5-k potentiometer, ±100 v CAL, so the 100 volt level of the display is 4 cm from the ZERO CHECK trace position. The zero-volt part of the display may not return completely to the ZERO CHECK trace position in the time the calibrator signal remains at zero volts. Do not adjust the vertical POSITION to offset this while adjusting the ±100 v CAL potentiometer.
10. Repeat steps 7, 8 and 9 as necessary until the amplifier functions properly. Proper operation means a 0.1-volt calibrator signal will produce 1 centimeter of deflection; 1 volt will produce 2 centimeters, and 100 volts will produce 4 centimeters. A tolerance of ±1 mm can be expected.

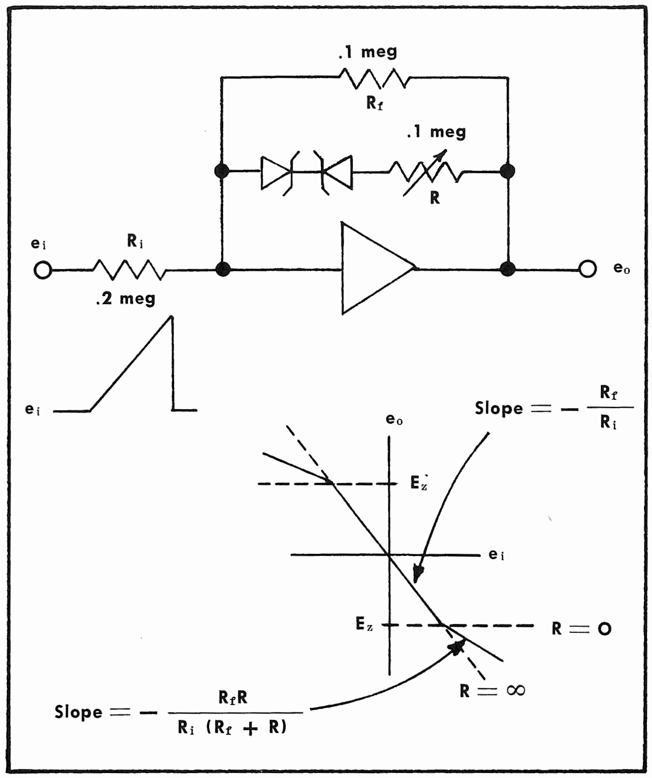
    To check for proper operation throughout the operating range, another voltage source must be used. A regulated, variable dc power source, monitored with an accurate meter will provide the accuracy needed.
11. Disconnect the calibrator signal and connect a 25-kc square wave from the Type 105 Square-Wave Generator to the input. Use a 93 Ω cable and a 93 Ω Termination Resistor and adjust the HF COMP capacitor for minimum spike at the leading edge of the square wave at about a 10-volt level. This adjustment will change with maximum signal amplitude, so it should be made at the maximum value of voltage the amplifier is expected to handle.
12. The amplifier is now calibrated. Always recheck the OUTPUT DC LEVEL against the ZERO CHECK trace position prior to any measurement.

Response-time of the logarithmic amplifier (the time to change 100% of a step signal amplitude) varies with the direction of change. The amplifier was originally designed to operate between 0.1 to 10 volts, therefore the response-time performance has been measured within these limits. Response-time when going from 0.1 volt to 10 volts is approximately 0.2 μsecond. Response-time when going from 10 volts to 0.1 volt is approximately 0.3 μsecond.

If your instrument serial number is below 813, we suggest that you modify the circuitry to agree with that found in instruments numbered 814 and above. This should reduce thermal drift problems when using the Logarithmic Amplifier.

14. Limiter Amplifier

The limiter amplifier operates as a normal amplifier until the output reaches the Zener breakdown voltage of the diodes. When this occurs, the Zener diode places R in parallel with R_f, thereby increasing the negative feedback and decreasing the gain of the amplifier. By changing the setting of R, the gain of the amplifier after Zener breakdown can be controlled. The curve illustrates the input-output characteristics of the circuit.



14. Limiter Amplifier

The input and output voltages of an amplifier are related by the expression $e_o = Ge_i$, where e_o is the output voltage, G is the gain, and e_i is the input voltage. From this expression it can be seen that G is the slope of the curve. For voltages between the two Zener breakdown points, the gain (and the slope of the curve) is the usual expression $-R_f/R_i$. After the Zener breakdown, the effective feedback resistance is that of R_f and R in parallel. The gain (and slope) of the limited curve is

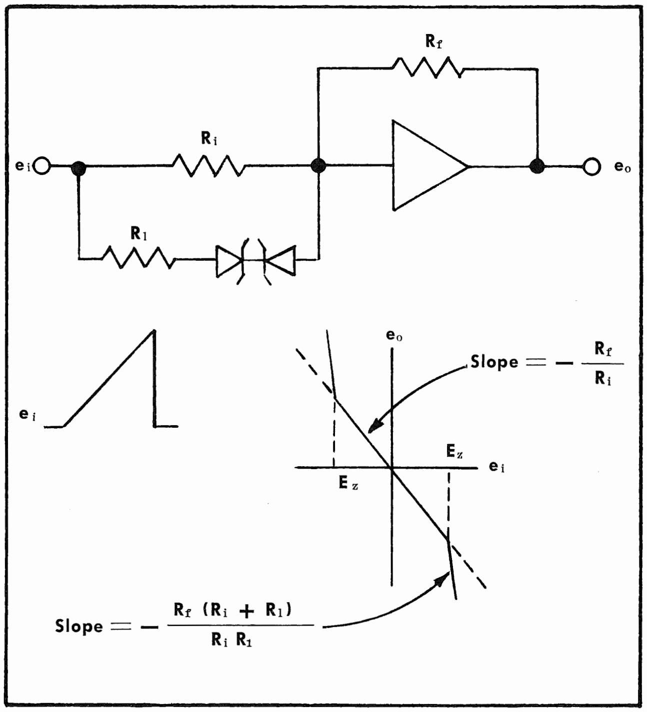
$$G = -\frac{R_f R}{R_i (R_f + R)} .$$

The slope of the limited curve can be varied between zero and $-R_f/R_i$ by varying the value of R between zero and infinity.

The purpose of the circuit is to limit the gain of the amplifier for output voltages greater than the Zener breakdown voltage. If R is reduced to zero, the output voltage is limited to the Zener breakdown voltage.

15. Expansion Amplifier

The expansion amplifier is similar in operation to the limiter amplifier of application 14. The primary difference in the two circuits is in the location of the back-to-back Zener diodes.



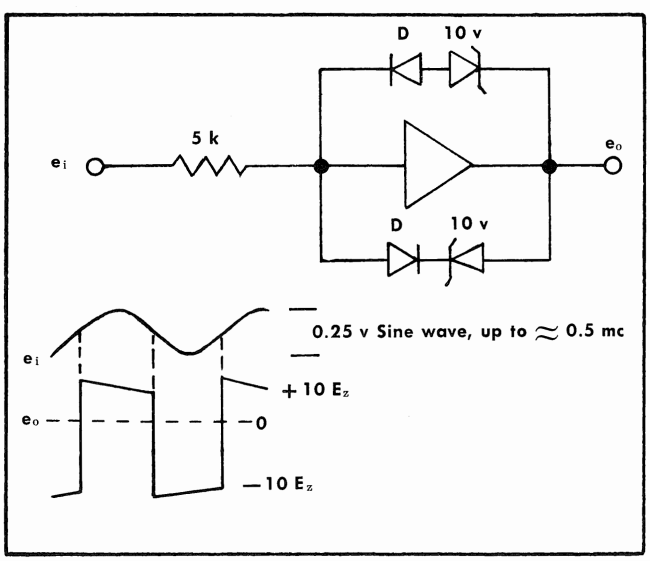
15. Expansion Amplifier

In this circuit, the gain of the amplifier is the usual $-R_f/R_i$ until the input voltage reaches the Zener breakdown level. When this occurs R_1 is placed in parallel with R_i thereby increasing the gain of the amplifier for input voltages above this level.

As mentioned previously, the gain of the amplifier (and thus the slope of the input-output curve) is $-R_f/R_i$ for input voltages below Zener breakdown. Above Zener breakdown the gain (and slope of the limited curve) is

$$G = -\frac{R_f(R_i + R_1)}{R_i R_1} .$$

Input signals above the Zener breakdown voltage are effectively expanded by the amplifier. Signals below the Zener breakdown voltage are amplified in the normal manner.



16. Clipper Circuit

16. Clipper Circuit

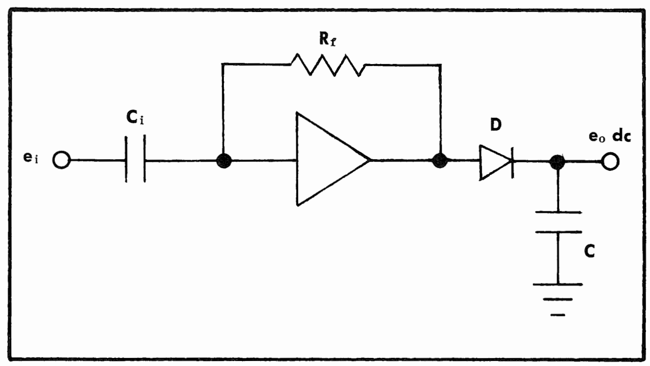
The clipper circuit is similar to the circuit shown in application 14 except that no feedback resistor is used. This means that the open-loop gain of the amplifier is obtained until the output reaches the Zener breakdown voltage. The output is then limited to the Zener breakdown voltage. The low-capacity signal diodes *D* disconnect each Zener during its forward voltage excursion to keep leakage feedback to a minimum.

The combination of the very high gain and clipping at the Zener breakdown voltage permits small amplitude sine waves to produce large amplitude square waves.

The example shows a gain of about 80 with clipping at the Zener level on both halves of each cycle.

17. Frequency to Voltage Converter

The frequency to voltage converter is essentially a differentiating circuit. The output of the differentiator in this circuit is rectified and used to charge a capacitor. Since the output of a differentiator is proportional to frequency, the capacitor charge is also proportional to frequency. A dc voltage is thus obtained which is proportional to frequency.



17. Frequency to Voltage Converter

It is important that the proper values of R_f and C_i are used in order to obtain best results from this circuit. Refer to Chart 2-2 for the proper values to use in the frequency range you wish to cover. Practical limits of performance are from approximately 1 cps to 1.5 mc, shown as several ranges in Chart 2-2.

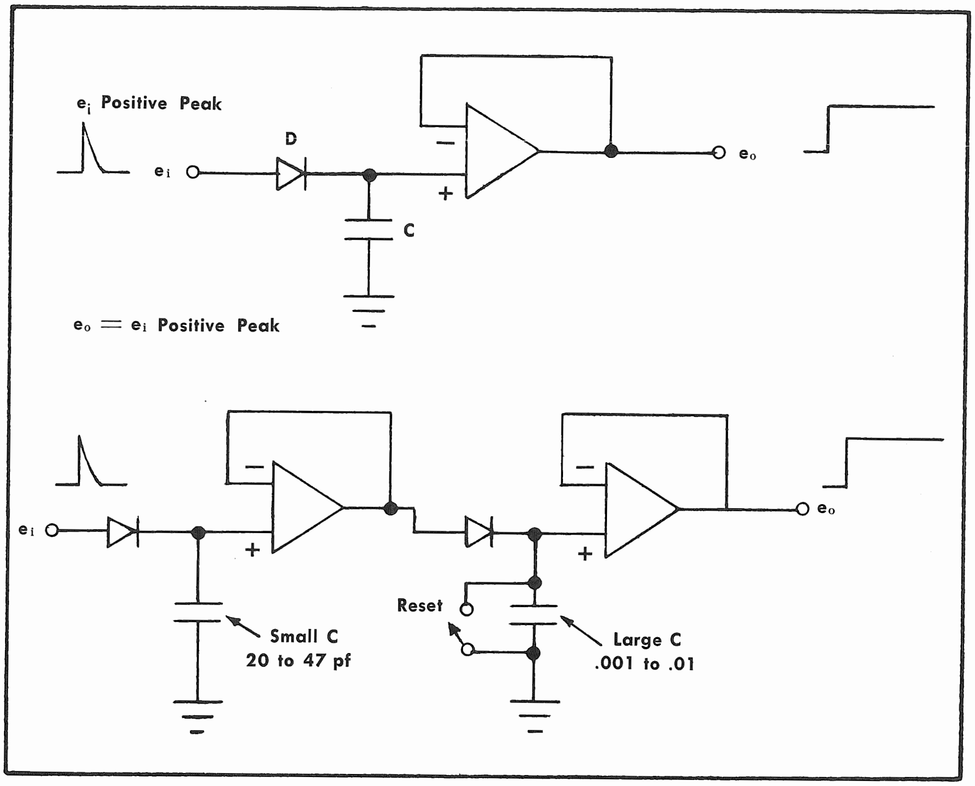
The diode *D* and capacitor *C* must both be selected to have high leakage resistance to permit maintaining the proper dc output voltage during the part of each cycle that *D* is not conducting in the forward direction.

18. Peak Reading Amplifier

In this circuit, advantage is taken of the high input impedance feature of the + grid. When a positive pulse is applied, the diode conducts, charging the capacitor to the peak voltage. Because of the high input impedance, the capacitor charge is retained for a relatively long period. The gain of the amplifier is unity under these conditions, so the output is equal to the peak voltage of the input pulse.

In order for the circuit to operate properly, the time constant of the source impedance and the capacitor to be charged must be short enough so that the capacitor can charge to the peak voltage in the time that the pulse remains at the peak. For this reason, the value of the capacitor should be as small as possible. The capacitor cannot be too small, however, or it will discharge too rapidly. A capacitor with very low leakage should be selected to prevent rapid loss of the charge. Also, the diode reverse current should be very low to prevent the capacitor charge from being lost too rapidly. The forward drop across some silicon diodes is great enough to prevent the capacitor from charging to the peak voltage.

In practice, two peak reading amplifiers can be cascaded to minimize input loading and extend peak memory time.



18. Peak-Reading Amplifier

19. Very Low Current Measurements

The output voltage of this circuit is proportional to the input current. The circuit can be used to measure very low currents, such as reverse current of semiconductor diodes, or the leakage current of capacitors.

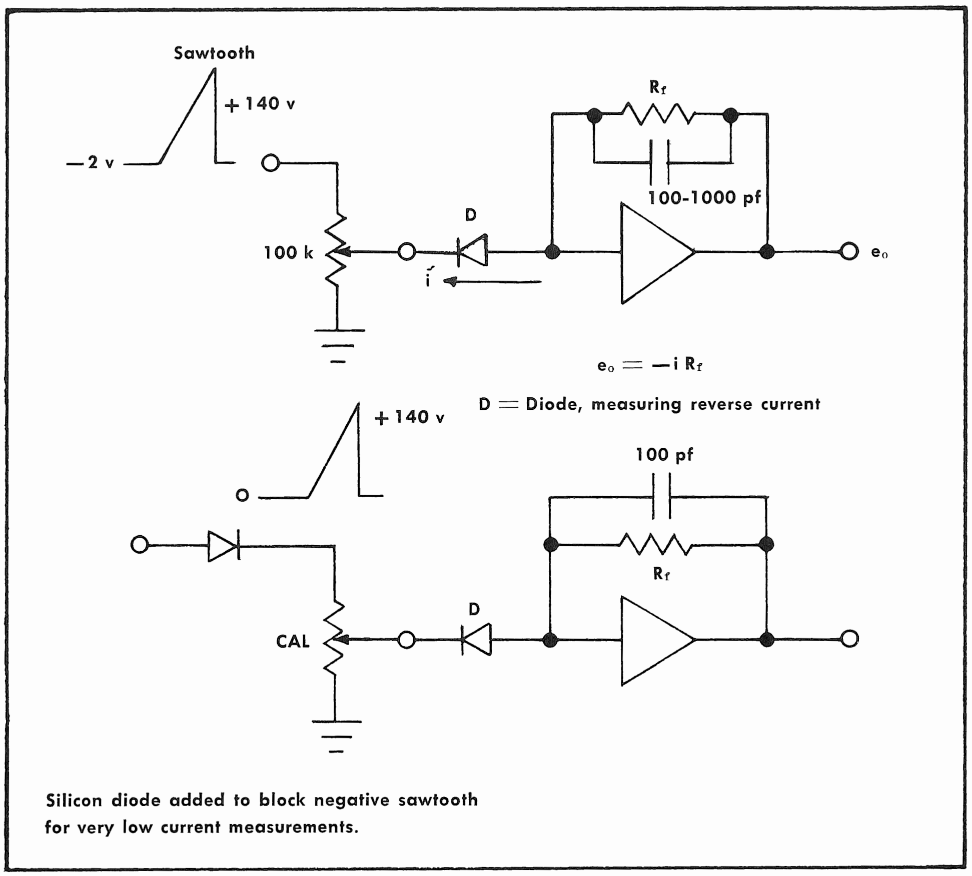
If R_f is 1 megohm, the output will change 1 volt per microampere input.

If R_f is 50 megohms (external resistor), the output will change 50 millivolts per nanoampere input.

The capacitor across R_f may be required to reduce output noise when the gain of the overall system is high.

A convenient voltage source for measuring diode reverse current is the oscilloscope SAWTOOTH OUT waveform.

The peak value of the sawtooth can be varied to suit the particular diode by means of a 100-k potentiometer. With



19. Very Low Current Measurements

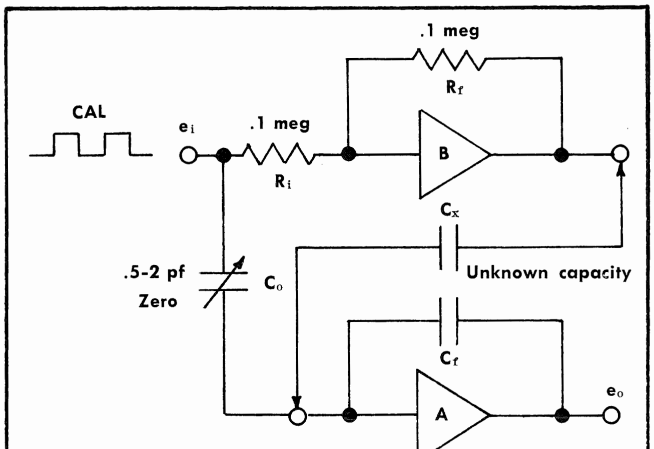
such a system, the horizontal display can be calibrated in volts per centimeter. Thus the exact voltage at which the diode reverse current exceeds a given value can be seen dynamically on the crt.

For very high sensitivities (the measurement of very low current) it may be necessary to shield the input and output circuits to eliminate hum pickup. The noise of the system can be reduced by placing a capacitor across the feedback resistor.

The oscilloscope sawtooth-out voltage usually contains a small negative offset when the crt spot is at rest. To prevent the diode being tested from going into forward conduction and saturating the amplifier, a silicon diode may be inserted between the sawtooth-out voltage and the potentiometer to block the negative offset voltage.

Capacitance of the added diode can affect the display at high sweep rates. To minimize the capacitance effect, slow sweep rates should be chosen for nanoampere current measurements. The oscilloscope single sweep feature may also be used to improve the display.

20. Capacitance Measurements



20. Capacitance Measurements

The above circuit can be used to make accurate capacitance measurements by comparing the value of an unknown capacitor against one of the internal standard capacitors. The fixed internal capacitors are within 1% of the front-panel values. The two lowest value capacitors are adjusted during calibration. In the diagram, C_x is the unknown capacitance and C_f is the selected internal capacitor.

Although resistors are normally used as the input and feedback components when an operational amplifier is used to provide amplification by a constant, capacitors may be used also. The gain of the amplifier with capacitors as the input and feedback elements is the ratio of the capacitive reactance of the feedback capacitor to the capacitive reactance of the input capacitor. The gain is therefore equal to $-C_x/C_f$. If a known input signal from the oscilloscope calibrator is applied to the amplifier, the gain can be determined by comparing the output displayed on the oscilloscope to the input signal. With the gain and C_f known, C_x can be calculated. This is the general method used.

Operational amplifier B is used as a unity gain driver for operational amplifier A. The low output impedance of amplifier B is necessary to obtain accurate measurements on large capacitors. The low output impedance permits charging C_x in less than one half the period of the calibrator waveform.

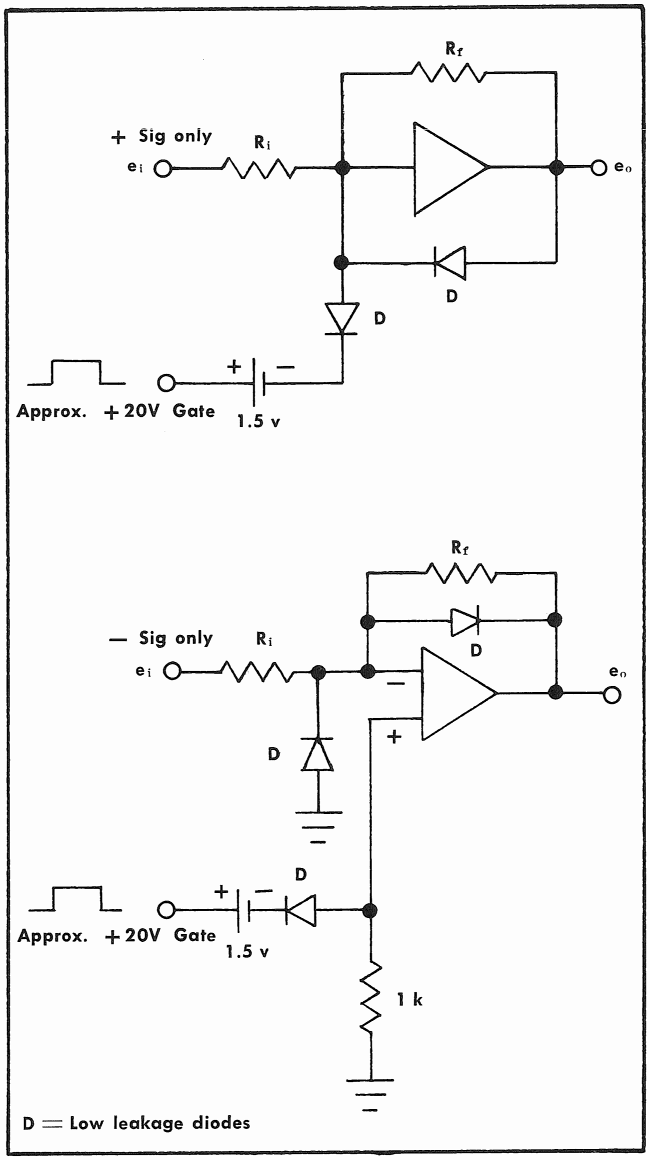
To make a capacitance measurement, proceed as follows:

1. Connect the circuit as shown in the diagram but do not connect C_x. Set the A Operational Amplifier Z_i switch to EXT.
2. Connect the output of the oscilloscope AMPLITUDE CALIBRATOR to the input of Operational Amplifier B and adjust the calibrator output voltage according to the following chart for the range of capacitance to be measured.
3. Set the VERTICAL DISPLAY switch to A+. Set the VOLTS/CM switch to .05. Connect any device which will be used to hold the unknown capacitor to the A INPUT and A — GRID connectors. If the capacitor to be measured is 10 pf or less, adjust C_o, the neutralizing capacitor, for the least amount of signal displayed on the crt. This neutralizes the stray capacitance between the input and A — GRID connectors.
4. Connect the amount of signal indicated below for the range of capacitance expected. Connect the capacitor to be measured and read the output of the amplifier displayed on the crt. Calculate the gain and the value of C_x. If C_x is equal to the value of the C_f switch setting, the display amplitude will be equal to the calibrator peak-to-peak voltage. If C_x is one half the value of the C_f switch setting, the display amplitude will then be one half the calibrator peak-to-peak voltage.

RANGE	CALIBRATOR SIGNAL	C_f SWITCH SETTING
0— 10 μf	10 mv	.1
0— 1 μf	.1 v	.1
0— .1 μf	.1 v	.01
0— .01 μf	.1 v	.001
0—.001 μf	1 v	.001
0— 100 pf	10 v	.001
0— 10 pf	10 v	.0001

21. Gated Amplifier

It is possible to gate on and off an operational amplifier. The simplest form of this type of circuit would be to use a relay with contacts that short across the feedback component when the amplifier is to be turned off, but the speed of response is limited by the response of the relay used.



21. Gated Amplifier

The simple relay may be replaced by a diode gate that is turned on and off to gate off and on the operational amplifier.

The ≅ 20 volt gate indicated can be obtained from either the A or B sweep of any Tektronix dual-sweep oscilloscope in which the Type O unit is used, such as the 535A or 545A. The 1.5-volt battery biases diode D to conduction when there is no + GATE signal. With the arrival of the + GATE, diode D opens, permitting the amplifiier to function.

The back resistance of the diode gate must be very high to prevent altering the amplifier performance. Also, its forward resistance must be low to permit firm clamping of the amplifier.

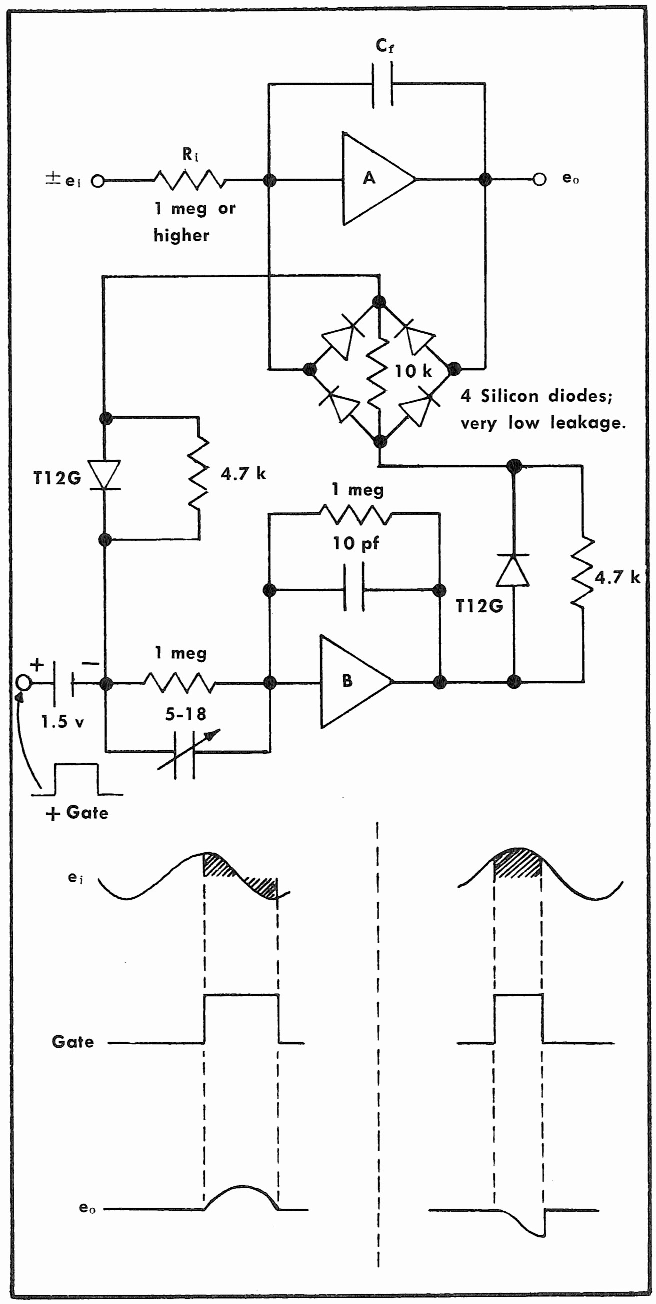
The gated amplifier illustrated here can accept only one polarity of signal. If the feedback capacitor of application 22 is replaced by a feedback resistor, a signal containing both positive and negative polarity can be gated.

22. Gated Integrator

This circuit is a modification of application 21.

The gated integrator can be used to integrate a signal over a specific time interval such as a portion of a sine wave or a small pulse that follows a large pulse. If a large pulse preceding a small pulse were to be integrated too, the area under the small pulse might not be easily resolved from the display.

By using a two-sweep oscilloscope, such as a Type 535A or Type 545A, the oscilloscope HORIZONTAL DISPLAY should first be set to 'B' INTENSIFIED BY 'A'. Adjust the two sweep rates and the DELAY-TIME MULTIPLIER dial so



22. Gated Integrator

the desired portion of the display is viewed by the 'A' sweep. Then, switching the HORIZONTAL DISPLAY switch to 'A' DEL'D BY 'B', the crt presentation will be that portion of the signal to be integrated. By using the + GATE A to turn on the integrator, the full crt display will then be the integral of the desired part of the original display.

If the waveform is of a recurrent nature, it may be necessary to use external triggering to the Time-Base B external TRIGGER INPUT. If the waveform is a single-shot type, the system described above will function correctly. In either case, it may be advantageous to read the operating instructions about the 'A' DEL'D BY 'B' mode of operation in the instruction manual for the oscilloscope used.

Integration of a repetitive waveform requires the use of the Type O Unit INTEGRATOR LF REJECT switch to keep the integrated waveform on the crt. By using a gated integrator, it is possible to integrate a signal plus any reasonable dc component associated with the signal. Under such conditions the LF REJECT switch would be OFF. By virtue of the shorted feedback element when there is no turn-on gate, the gated integrator is not subject to dc drift between gating periods. Therefore, it is possible to integrate essentially single-shot waveforms with a dc component included in the integral waveform.

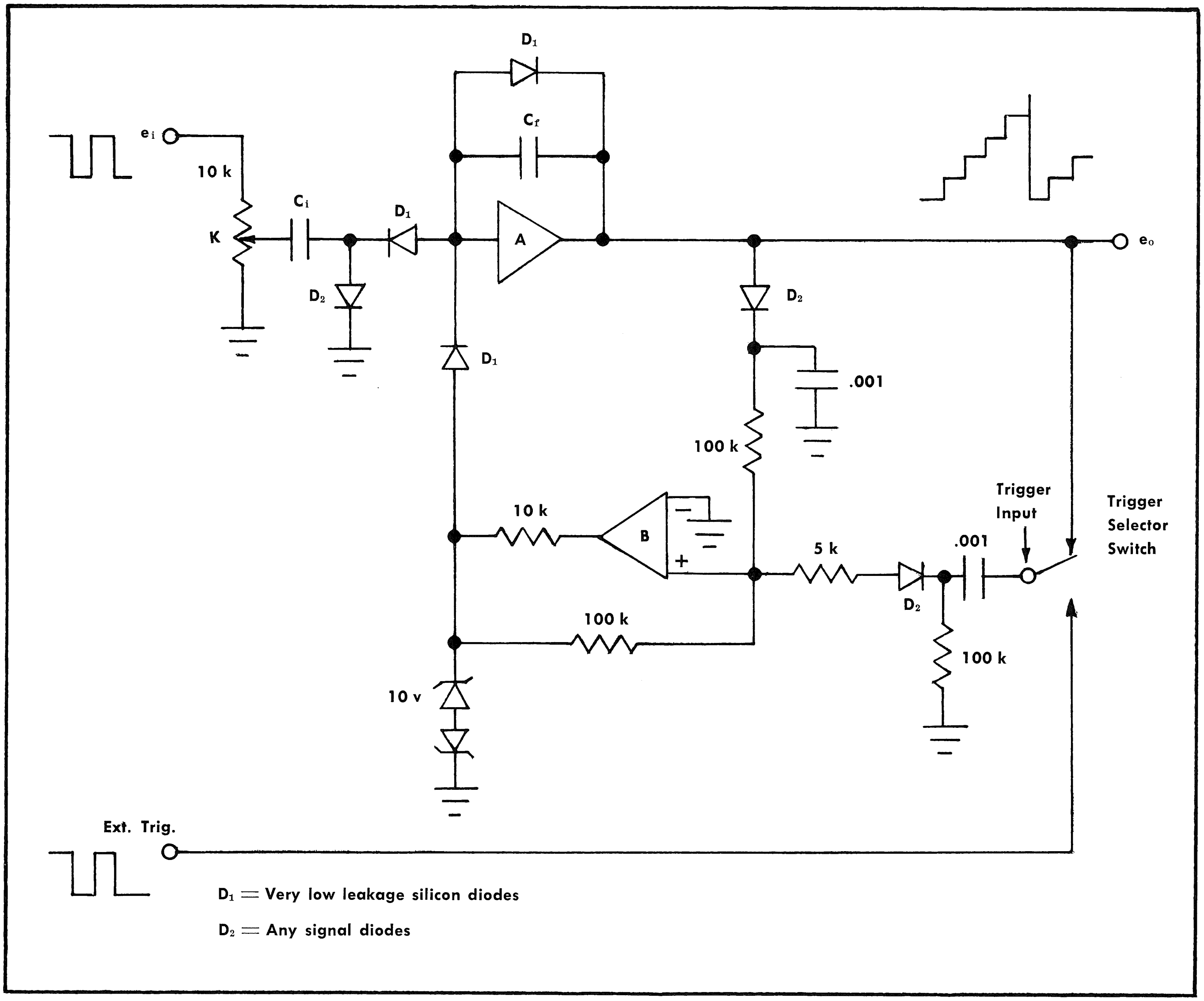
In any use of the gated integrator, refer to Chart 2-2 so that errors will not be made due to the gain-frequency limits of the operational amplifier integrator.

23. Stair-Step Generator

Another form of gated system is shown here with the B amplifier used as a multivibrator to gate an operational integrator. The integrator is fed a train of equal amplitude pulses via a 'Bucket' capacitor circuit to form a stair-step output waveform.

With e_i equal to the amplitude of the pulse feeding the operational integrator, the output amplitude per step is

$$e_o \text{ per step} = \frac{K C_i}{C_f} e_i,$$



23. Stair-step Generator

where K is the potentiometer setting (as a fraction of the pulse amplitude), C_i is the input capacitor, and C_f the feedback capacitor.

Diode coupling from the output stair-step to the + GRID of the B amplifier multivibrator assures the system will revert to zero output at the voltage of the back-to-back Zener diodes.

Once reverted to essentially zero output, a trigger pulse is required at the multivibrator + grid to open the gate again and permit another stair-step cycle. Thus the circuit can either produce a continuous series of stair-steps, or only one at a time, on the arrival of a single trigger. To make it repetitive, connect the output stair-step to the trigger input.

The maximum time per step (with good fidelity) is determined by the value of C_i. Two examples of maximum time limit per step, when C_f is a .01-μf capacitor, are: if C_i = .001 μf, the maximum time per step is about 10 μseconds; if C_i = 47 pf, the maximum time per step is about 1 μsecond. The minimum time per step is limited by the impedance of the driving source at C_i, and by the bandwidth of the A amplifier.

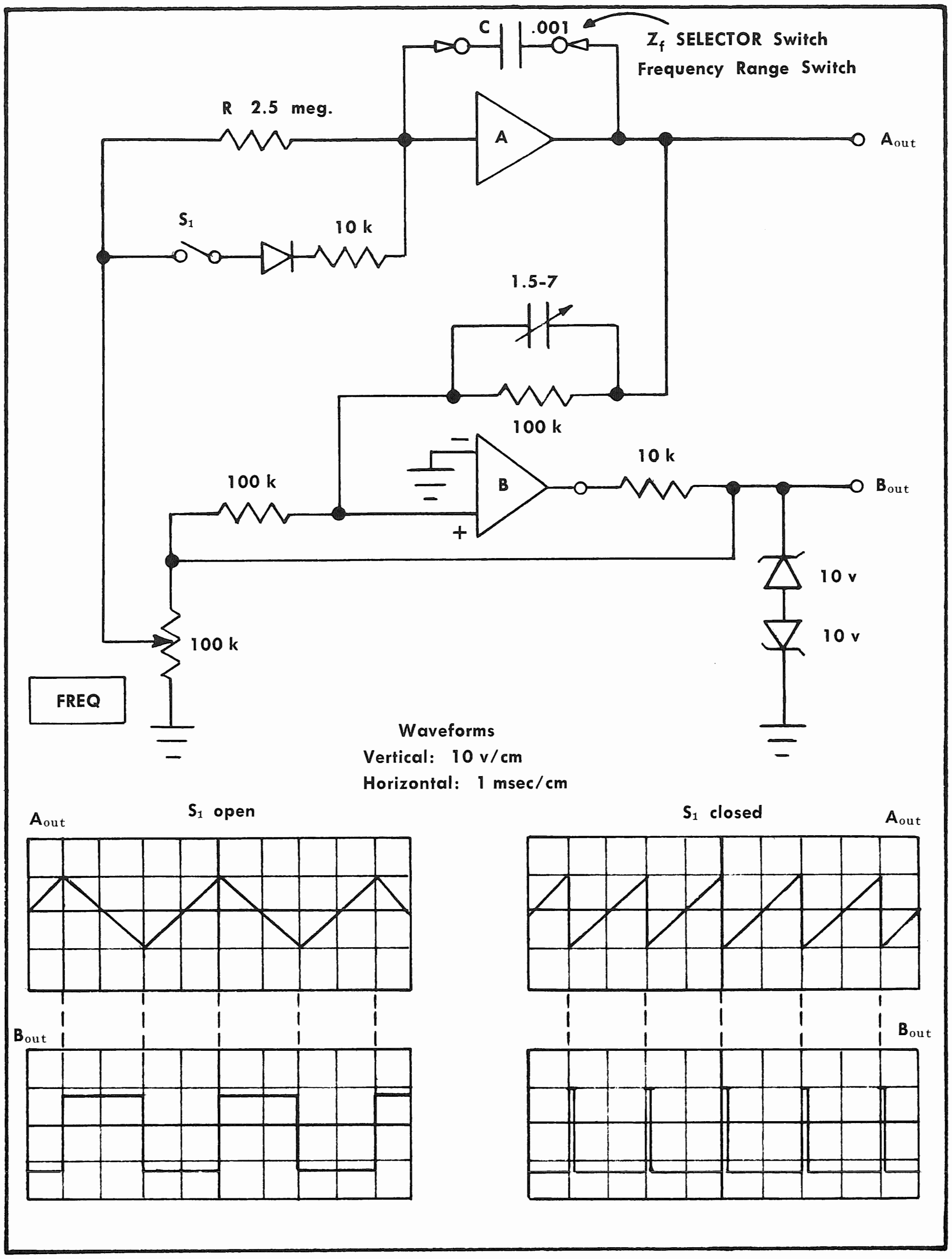
24. Function Generator

The function generator makes simultaneous use of both operational amplifiers. The A amplifier functions as an integrator while the B amplifier functions as a flip-flop multivibrator.

When the unit is switched on, the dc level at the output of the B amplifier is integrated by the A amplifier to form a linear ramp as shown by the waveform from the A output. When the ramp voltage reaches a certain level, it causes amplifier B to switch into its other state. The change in voltage at the output of B is integrated by A to form the remainder of the sawtooth from A. The output from A is again applied to B and causes B to switch to its original state after the output of A reaches the required level. This completes one cycle of operation. The cycle then repeats. The operation of the circuit can be compared in many respects to a simple relaxation oscillator.

The frequency of the output is determined by the time required for the voltage at the output of amplifier A to rise to the required switching level of B. This is determined by the value of R and C, and the setting of the *FREQ.* control. A nominal upper limit of 60 kc is possible without seriously changing the waveforms from those shown.

The output waveforms are modified by closing switch *S1*. This switch permits a diode to reduce the charging time for C in one direction only, changing the symmetrical ramp waveform to an unsymmetrical one as shown. This also affects the switching time of amplifier B and its duty cycle.



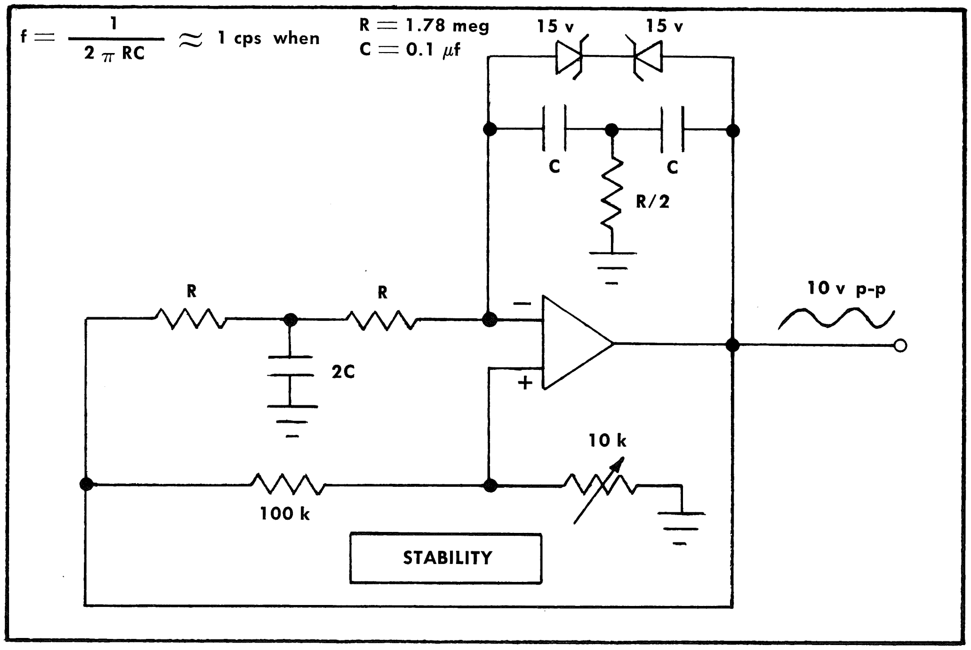
24. Function Generator

The back-to-back Zener diodes at the output of amplifier B square off the tops and bottoms of the waveforms and limit their amplitude, providing a square-wave output.

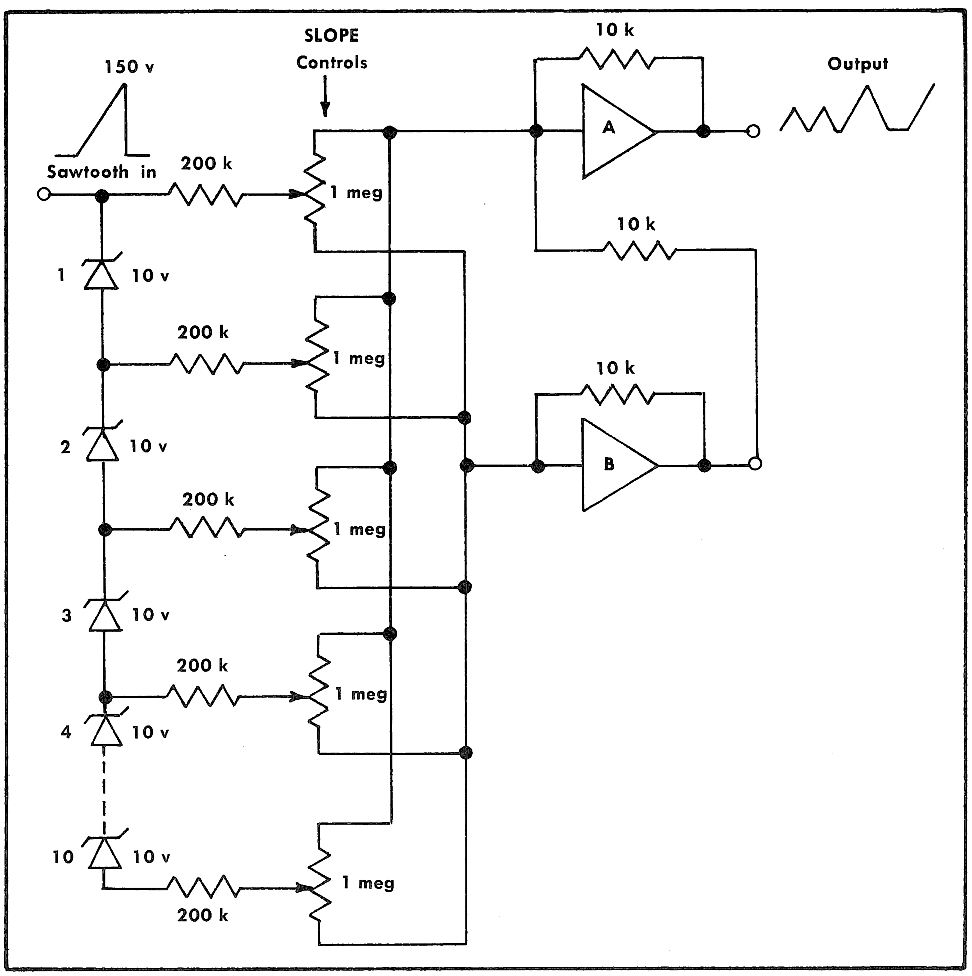
25. Low-Frequency Sine-Wave Generator

This circuit is a parallel-T oscillator. Feedback to the — grid becomes positive at the frequency indicated in the formula, while positive feedback is applied to the + grid at all frequencies. The amount of positive feedback to the + grid is sufficient to cause the amplifier to oscillate. In combination with the feedback to the — grid, feedback to the + grid can be used to stabilize the amplitude of oscillations.

The output of the oscillator is a stable sine wave with very low harmonic content. Output voltage is approximately 10 volts, peak-to-peak. The nonlinear resistance of the back-to-back Zener diodes is used to limit the output amplitude and maintain good linearity.



25. Low-Frequency Sine Wave Generator



26. Segments Function Generator

26. Segments Function Generator

The simple segments function generator shown employs Zener diodes to determine the starting voltage of each segment, and 200-k potentiometers to control the slope of the segment after each Zener conducts.

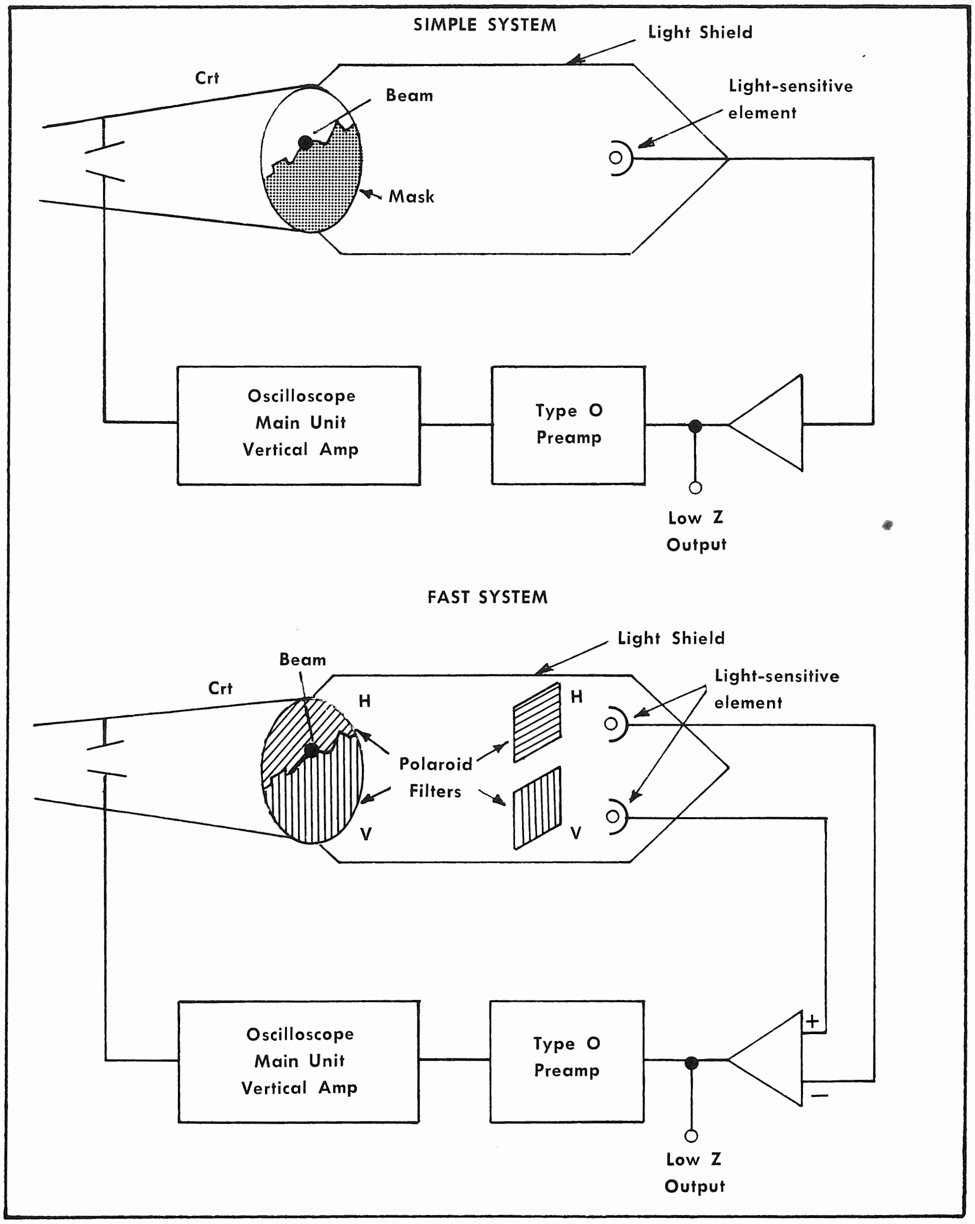
Many Tektronix oscilloscopes can provide the input signal from the SAWTOOTH OUT connector. The oscilloscope sweep generator therefore controls the waveform rate, either on a repetitive basis or on a manually- or externally-triggered single-shot basis.

27. CRT Function Generator

The top illustration shows the circuit and physical arrangement for a simple crt arbitrary function generator. A mask is placed over the face of the oscilloscope crt in the shape of the waveform to be generated. A light-sensitive element is then placed in front of the crt. The output of the high-gain operational amplifier is applied to the preamplifier to control the vertical position of the crt spot. The light-sensitive element is connected so that a bright display moves the spot down.

In operation, the crt spot is positioned (with the Positioning controls) to be just above the highest edge of the mask. If the spot then moves higher, the additional light reaching the phototube moves the spot down. If the spot moves down too far, the reduced light moves the spot back up. The net result is that the spot stays at the edge of the mask. As the spot is swept horizontally by the oscilloscope sweep circuit, it follows the edge of the mask, tracing out the pattern. The dc-coupled voltage from the operational amplifier OUTPUT connector is nearly an exact replica of the mask. Any type of arbitrary function waveform can be generated by cutting out the proper mask and placing it over the oscilloscope crt.

The simple crt function generator has upper rate limitations for the crt spot movement. These limitations are due to the response time of the light-sensing element, the bandwidth of the operational amplifier, and the type of phosphor and its light output. If a fast sweep rate is required, it may be necessary to undercut the mask at sharp corners. Experimental masks will lead to the correct shape to give the desired waveform output.



27. CRT Arbitrary Function Generator

A more sophisticated version of the crt function generator is shown in the lower illustration. This system uses two light-sensing elements and two sets of polaroid filters for differential operation. The response time of this system is greater than the simple system, with a limitation now including the oscilloscope delay line.

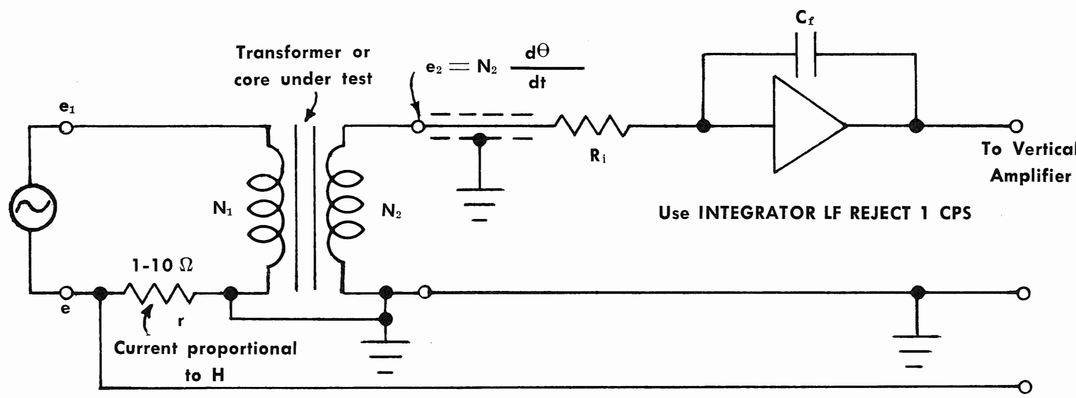
28. Displaying B-H Curves for Magnetic Materials

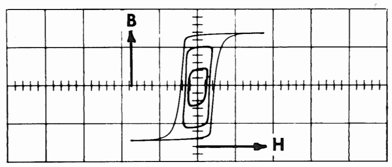
B-H curves for magnetic materials can be displayed using this circuit. A transformer is constructed using a core of the test material and the transformer is excited from the output of a variable autotransformer. The magnetic intensity *H* in the core is proportional to the current through the primary winding. The voltage across a current sampling resistor in the primary circuit is applied to the horizontal deflection system of the oscilloscope. Horizontal deflection on the oscilloscope is thus proportional to *H*.

The output voltage obtained from the secondary winding is proportional to the time rate of change of the flux. The transformer secondary voltage is applied to an integrator circuit which gives an output voltage proportional to the flux. The output of the integrator is applied to the preamplifier where it produces vertical deflection of the crt beam. Since the flux density *B* is equal to ϕ/A (where *A* is the cross sectional area of the core), the oscilloscope vertical deflection is also proportional to *B*.

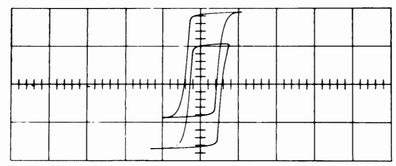
The net result of the signals applied to the horizontal and vertical deflection systems is to produce patterns similar to those indicated. Vertical deflection is proportional to *B* and horizontal deflection is proportional to *H*.

If it is desired to determine qualitative measurements of a transformer core from the oscilloscope display, the proportionality constants relating the horizontal and vertical deflections to *H* and *B* must be determined (see page 4-4).

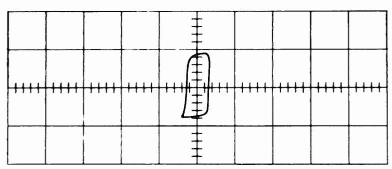




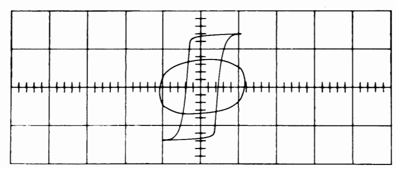
(a) Three levels of supply voltage to an inductor.



(c) Top: Normal B-H curve.
Bottom: Small dc offset bias effect.



(b) Same inductor, indicating residual magnetism.

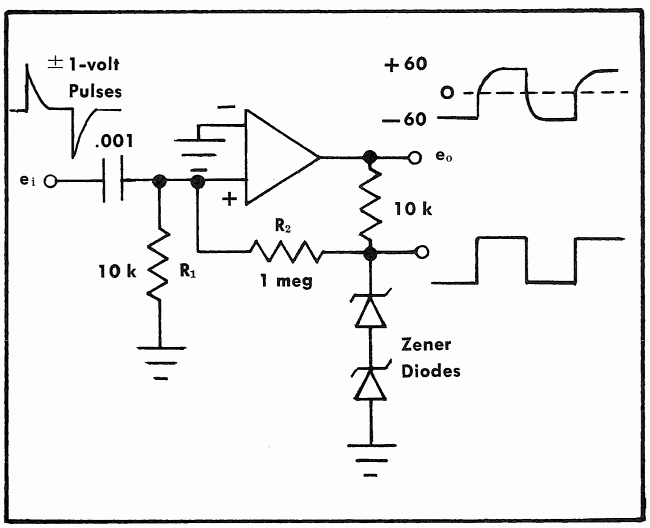


(d) Normal B-H curve compared with case of one shorted turn.

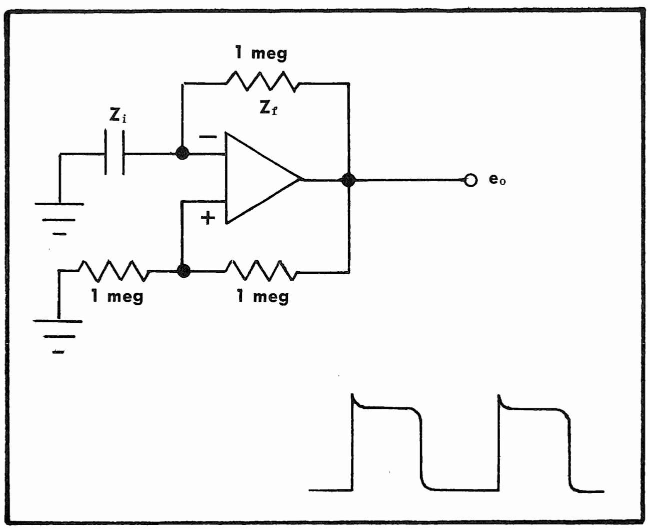
28. Displaying B-H Curves for Magnetic Materials

ADDITIONAL APPLICATIONS

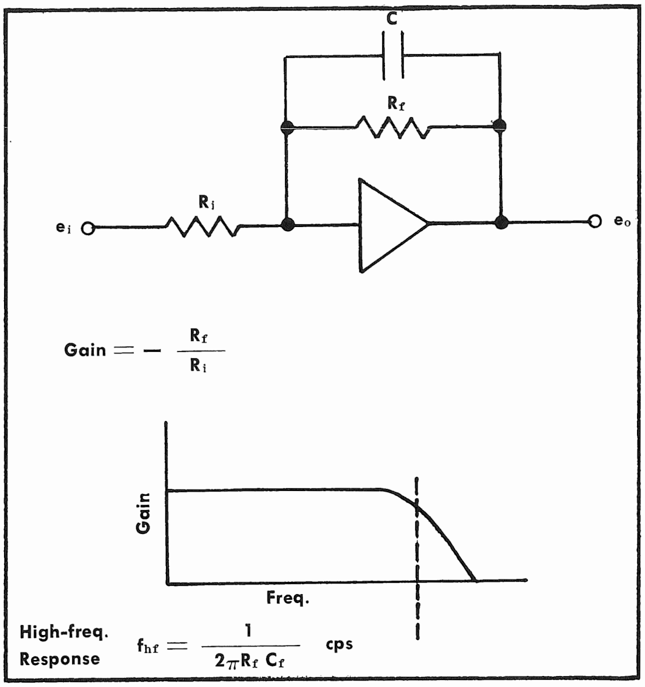
The following applications are in schematic form only, offering suggestions for other uses of the Type O Unit.



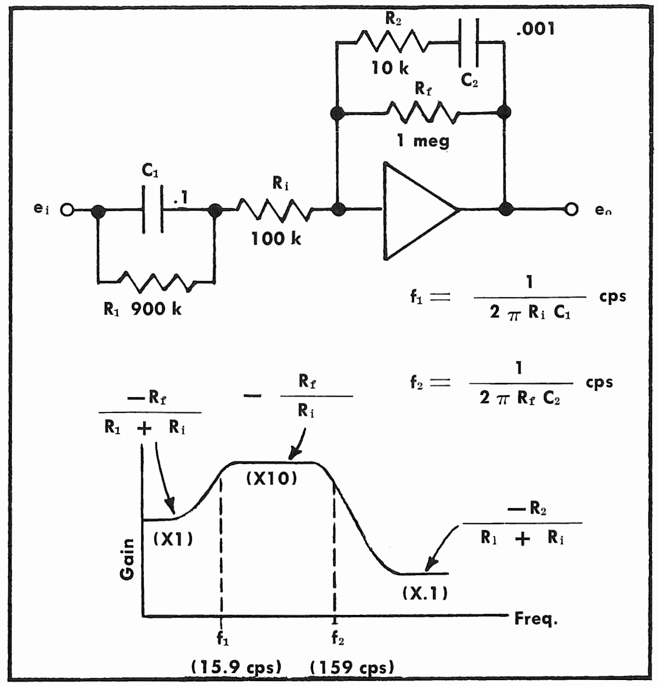
A. Bistable Multivibrator



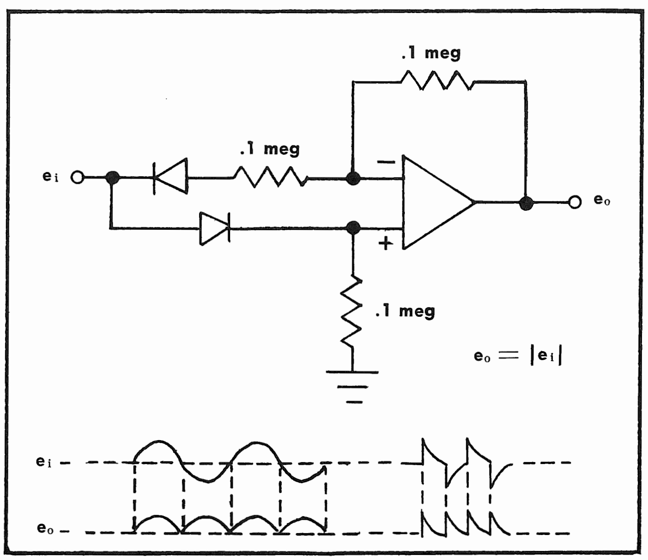
B. Free-running Multivibrator



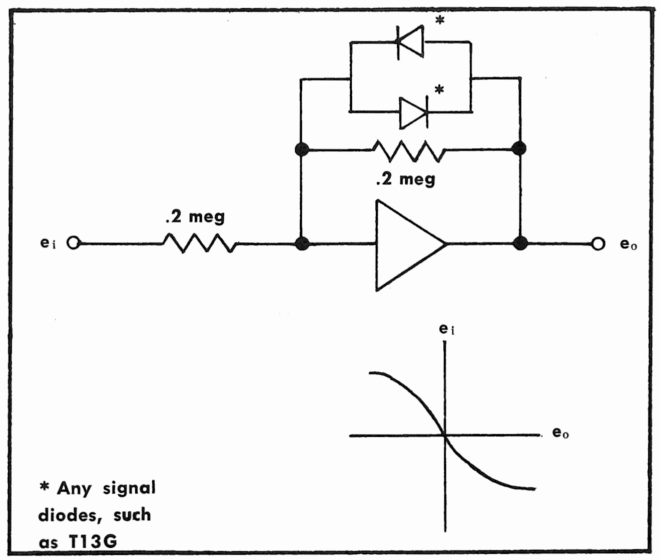
C. Bandpass amplifier



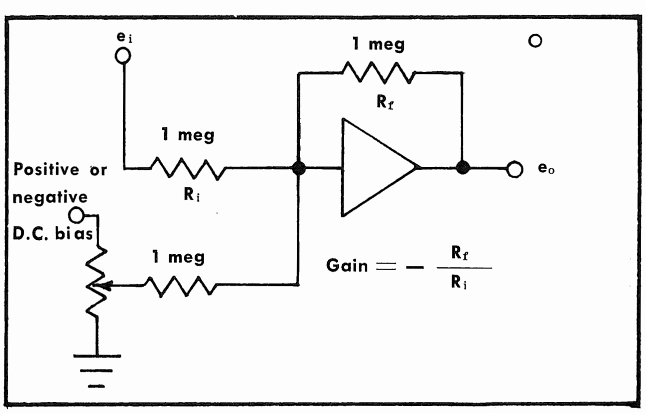
D. Special Bandpass amplifier



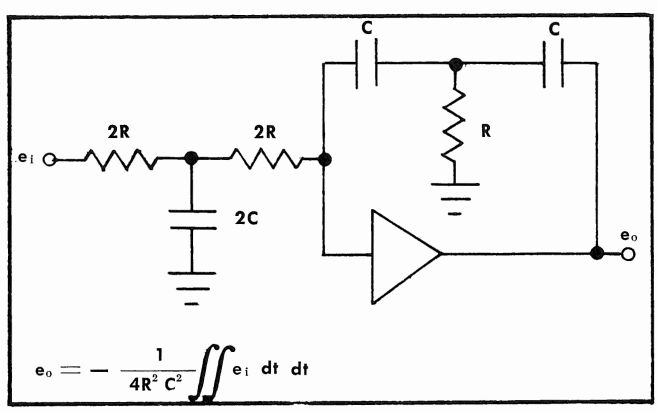
E. Full-wave rectifier



F. Simple non-linear amplifier



G. DC Slideback amplifier



H. Double integrator

Additional equipment that might be used for various applications of the Type O Unit is shown in the pictures following. Most of this equipment can be purchased locally.

NOTE

To use some of the equipment shown it will be necessary to use a BNC to UHF adapter.

Several of the applications in this section have been constructed into front-panel plug-on adapters. The adapters have circuitry similar to the numbered application which follows the adapter name.

Log Adapter (13) — Tektronix Part Number 013-067

Gating Adapter (22) — Tektronix Part Number 013-068

Compensating Adapter — Tektronix Part Number 013-081

Leakage Current Adapter (19) — Tektronix Part Number 013-068

To provide independent operation of Type O Plug-In Units, the Type 132 or 133 Plug-In Unit Power Supply may be used. This will allow for a more versatile system. The Type 133 has a high current output.

For additional information contact your local Tektronix Representative.

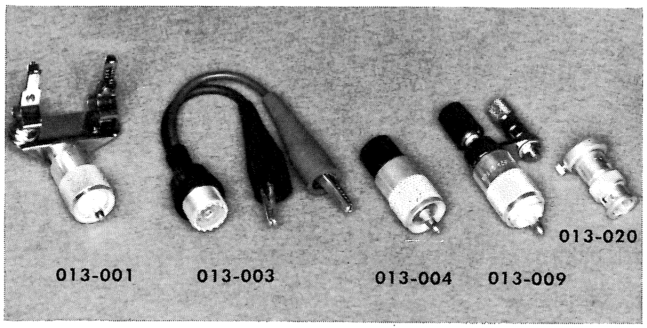


Fig. 3-4.

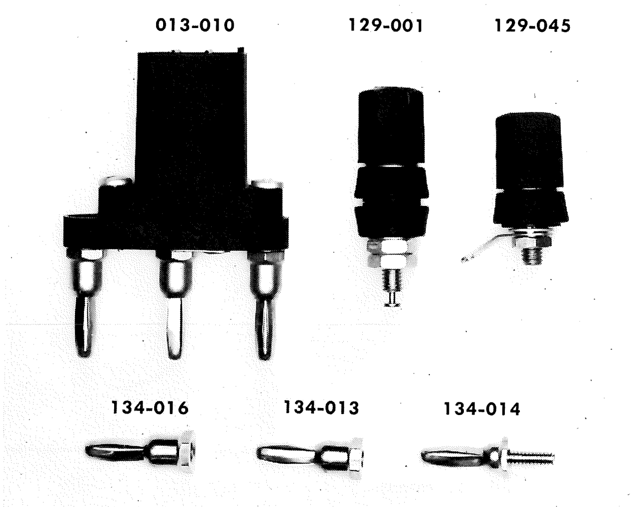


Fig. 3-5

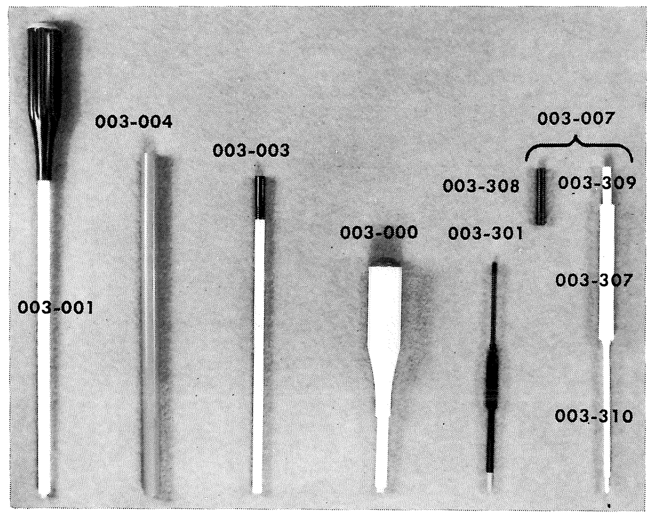


Fig. 3-6.

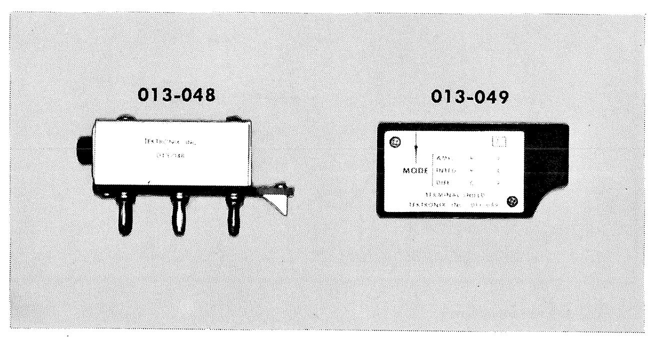


Fig. 3-7.

SELF-RESETTING STAIRSTEP GENERATOR
FOR USE WITH TEKTRONIX TYPE O PLUG-IN UNIT

For stepping the position of a waveform displayed on a Tektronix Type 564 storage oscilloscope, a self-resetting stairstep generator may be constructed for use with the Type O plug-in or other operational amplifier. The stairstep voltage available from the output may be either mixed directly with the incoming signal, or (preferred) introduced into the negative input of a differential amplifier. The stepping signal may also be introduced into the DC Balance or Position circuit of a non-differential input plug-in, if necessary. Since the Type O plug-in cannot be operated from the Type 564 power supplies, the use of a Type 132 or 133 power supply is suggested. The stepping command required may be obtained from the internal gating signal in a 2B67, 3B1 or 3B3 Time Base plug-in. The reset pulse capacitively coupled from the stairstep generator may be used as the "erase" command for the Type RM564.

The stair-step generator may also be used as a pulse counter or incremental ramp generator.

Operational Amplifier Requirements: Nearly any operational amplifier may be used which provides for simultaneous access to both plus and minus grids. For a given operational amplifier, the limiting factors will be available voltage and current output swing, and grid-current. The latter will affect drift and hence the maximum useful duration of any one step.

Signals: Suggested component values provide a ramp of 15 v peak, with the number of steps continuously variable from 2 to 200 or more, using the O-Unit's internal Z_f capacitors. The required step command is a 20 to 25 volt gate waveform from a source impedance of 5k or less for its negative-going portion. A higher source impedance may be used, but the time required to complete the step transition may become excessively long. Component selection allows the use of nearly any step-command waveform greater than about 1.5 volt. For uniform steps, the step-command must be of uniform amplitude and the duration of its negative portion at least $(3 \times 10^3)(C_1)$ seconds.

Construction Details (Refer to schematic diagram and parts list for components discussed): The stairstep generator may be constructed in a Tektronix 013-048 terminal adapter assembly for use on the Type O plug-in or any other operational amplifier with the same terminal arrangement. The cam at the right of the adapter assembly should be turned so that the "±Grid" switch (Type O) is forced to the "+Grid" position when the adapter is installed. Use of an AB Type G or similar (miniature) potentiometer for the 10-1 variable steps-per-ramp control allows mounting in one of the connector holes in the adapter. Wiring may be done point-to-point. It is helpful to run a heavy wire ground-bus from right to left in the assembly, just under the cover, to provide the seven ground connections needed, plus convenient component mounting (see sketch).

Selection of C_1, C_f and R_5:

If the Type O plug-in is used, C_f may be selected by the O-Unit's Z_f control for any decade value from 10 pf to 1 µf. For operational amplifiers without internal feedback components, 0.1 µf is a good starting value. The value of Cf, grid-current, and leakage of D_2, D_3 and D_4 will determine the drift rate between steps.

It is recommended that R_5 be selected for a ramp amplitude in the range of 10 to 25 volts. A smaller voltage range will show proportionally greater drift; a larger range requires a more expensive transistor. The output can be attenuated as desired externally. To select R_5, divide the desired peak output voltage of the ramp by 2.2×10^{-3}.

The minimum number of steps per ramp is determined by the voltage amplitude of the input pulse or gate divided into the peak output voltage, that ratio multiplied by C_f/C_1:

$$\text{Steps per ramp} \approx \left(\frac{E_p}{Ein}\right)\left(\frac{C_f}{C_1}\right)$$

Approximately 10 times this number of steps per ramp will be secured when R_1 is set to minimum. If the input pulse is 25v, the output ramp is to be 15v, C_f is to be 0.1 µf, and a minimum of 5 steps per ramp is required,

$$C_1 = \frac{(E_p)(C_f)}{(Ein)(Steps)}$$

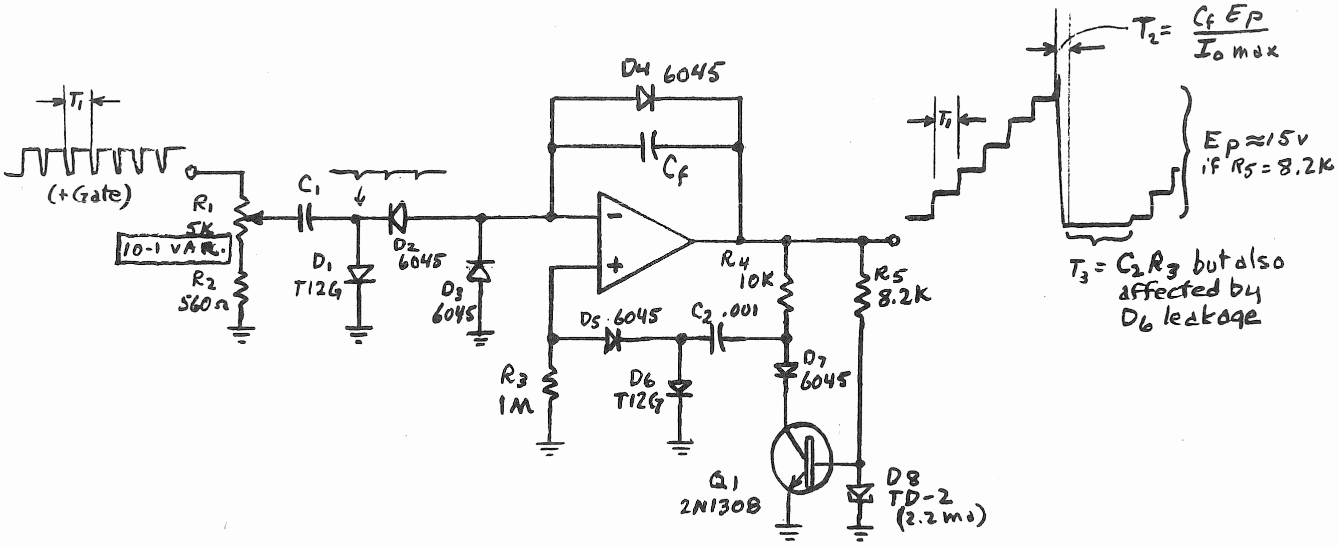
$$= \frac{(15)(0.1)}{(25)(5)}$$

$$= .012\,\mu f.$$

The value of C_1 required will be slightly higher because of the forward drop (~400 mv) in D_1 during charging of C_1 and the forward drop (~700 mv) in D_2 when transferring the charge to C_f. The suggested value is .033 µf, which provides 2-200 steps using 0.1 µf and 1.0 µf values for C_f if Ein is 25v.

The time required for the operational amplifier to reset the ramp is determined by the peak voltage, the C_f value selected and the maximum current the operational amplifier can supply in the negative-going direction. In the case of the O-unit, this is somewhat greater than the 5 ma available for steady-state operational accuracy, and amounts to about 12.5 ma. Operational amplifiers having a PNP emitter-follower output should have a current-limiting resistor between the output and C_f.

The same circuit can be made to provide a negative-going staircase (reverse all diodes and use a PNP transistor) but some means should be employed to limit output current during reset.

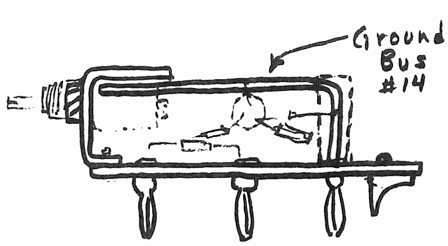


SELF-RESETTING STAIRSTEP GENERATOR

PARTS LIST

C_1	.033 µf (See construction notes)			R1	5k pot, miniature	311-310
C_2	.001 µf discap 200 v 10%	283-067		R2	560Ω 1/2 w 10%	302-561
C_f	Selected by O-unit Z_f control			R3	1 M 1/2 w 10%	302-105
				R4	10k 1/2 w 10%	302-103
D_1	T12G General purpose, Ge	152-008		R5*	8.2k 1/2 w 10%	302-822
D_2	6045 Low leakage, Si	152-045				
D_3	6045*** " " "	"		Q_1	2N1308**	151-072
D_4	" " " "	"				
D_5	" " " "	"			Enclosure. O-Unit	
D_6	T12G	152-008			Terminal Adapter Assy	013-048
D_7	6045	152-045				
D_8	TD-2 2.2 ma Tunnel Diode	152-081			Knob (if desired)	366-085

Notes:
* For 15 v p-p stairstep. Select larger value for higher E_p.
** If stairstep $E_p > 25$ v, use germanium transistor with higher V_{cbo}.
*** T13G, 152-005, may be used if $C_f \geq 0.1$ µf.



CONSTRUCTION SUGGESTION:

Use long 'overhead' ground bus for component mounting.

GG/PI 7-63 3-64

Theory: A negative charge is applied to D_2, which charge is transferred to C_f by the operational amplifier to provide a voltage step at the output. The charge in the circuit shown is the charge in C_1 developed during the "on" time of the plus-gate of the oscilloscope with which the stairstep is used. The charge is transferred to C_f during the interval between sweeps. The amount of voltage developed across C_f for each step is dependent on the +gate amplitude, the setting of R_1 and the ratio C_i/C_f.

The positive going ramp provides both collector voltage for Q_1 and current for the tunnel diode D_8 which is normally in the quiescent state. When the ramp reaches the amplitude at which the current through R_5 and D_8 reaches 2.2 ma, the tunnel diode switches to its high voltage state (+500 mv approx), turning on Q_1. The collector of Q_1 snaps sharply negative. This negative-going step is coupled to the +Grid of the operational amplifier via C_2 and D_5, causing the output to fall negative. The negative transistion at the output is also coupled via C_2 and D_5 to the positive grid, completing a regenerative loop and forcing the output to continue falling even when D_8 has reverted to its low voltage state and Q_1 is turned off. D_8 and Q_1, then, only trigger the reset; the "work" is done by the operational amplifier.

Diode D_3 prevents the falling output from driving the -grid negative and canceling the regenerative action; D_4 catches the output when it reaches ground potential. Diode D_7 prevents the collector of Q_1 from being driven negative.

The negative transition drives the left hand end of C_2 to approximately -15 volts. It then recovers to ground potential via the C_2R_3 time-constant. Until the +grid recovers to 0 volts and D_5 disconnects, further input pulses to the -grid are shunted to ground via D_3 and do not charge the stairstep. This provides a natural holdoff action. The circuit comes out of holdoff with a sharp regenerated step to 0 v.

Geoff Gass, 2-15-65

CALL REPORT EXTRACTS
TECHNIQUES

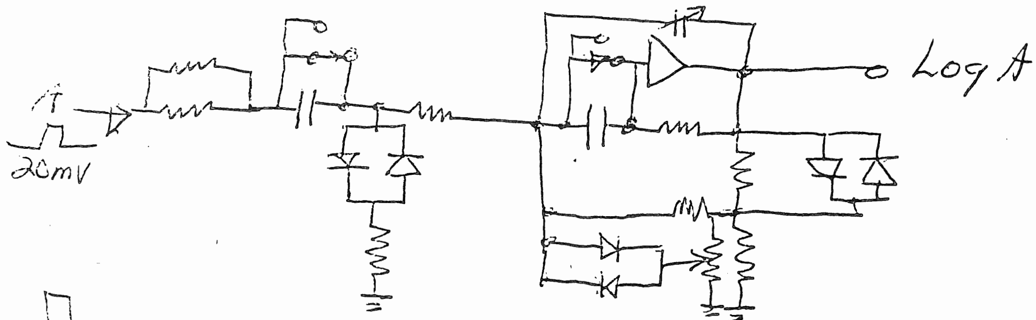
PHOTO CATHODE RESPONSE MEASUREMENT

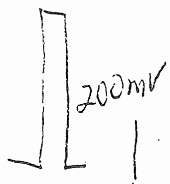
ISTOMER U. S. AIR FORCE	FIELD ENGINEER CHUCK BENDER
CITY AND STATE ALBUQUERQUE, NEW MEXICO	SEP 23 1964 · MONTH-DAY-YEAR 9/1/64 · CALL #
GROUP RADIATION EFFECTS LABORATORY	GROUP FUNCTION PHYSICS DEPARTMENT/PHOTO EMISSION DEVICES RESEARCH
NAMES GG LT. BRUCE	

Lt. Bruce is trying to run some normalized responses from the output of a string of photo cathodes. His pulse outputs are two general levels, approximately 20 MV and 200 MV. He would like to display the ratio of these pulses on a CRT. I suggested that he take the log of each of these pulses, subtract, and then take the anti-log and display this. This will in effect give him division by a variable.

Idea works ok except for the accuracies attainable. Each log amp can have up to ±1mm error = ±2db = ±20% in the 100 mv – 100 v range. Antilog circuit would add one more tolerance. Since log amp is not logarithmic between 0v and ±50 mv, 20 mv signal will have to be amplified. Direct ÷x unit needed for decent accuracy — and we just don't have one. — GG.

O UNIT #1



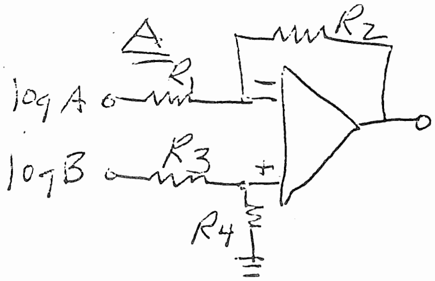


B - Same as A

o Log B

O UNIT #2 Con't from preceding page

A



log B − Log A

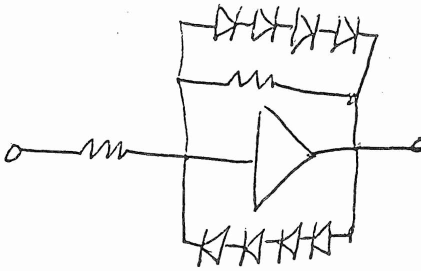
$$e_o = -\frac{R_2}{R_1}\log A + \left[\frac{R_4}{R_3+R_4}\right]\left[\frac{R_1+R_2}{R_1}\right]\log B$$

$$e_o = \log B - \log A$$

when $R_1 = R_2 = R_3 = R_4$

B

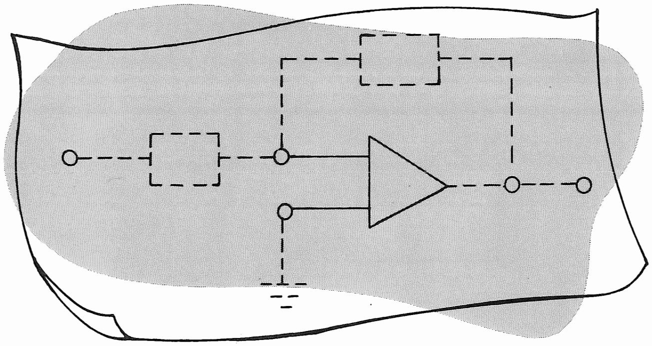
Log B − log A



$$e_o \approx -[1.4 + (.35\ \text{antilog}\ e_{in})]$$

$$e_o = \text{Antilog } B - A$$

$$e_o = \text{linear}$$



SECTION 4
THEORY OF OPERATIONAL AMPLIFIERS

General

Information in this section of the manual is intended to provide some understanding of the theory of operational amplifiers. In addition, derivations presented in this section support some of the formulas that appear throughout other sections of this manual.

Operational Amplifier

An operational amplifier is basically a high-gain dc amplifier with feedback. The feedback elements selected permit the amplifier to perform various operations such as amplification by a constant factor, integration, differentiation, summation, etc. In conventional operational amplifiers, the feedback is negative. However, the operational amplifiers in the Type O Unit also permit positive feedback, thereby increasing the number of possible applications.

Generalized Feedback Arrangement

In the feedback operational amplifier shown in Fig 4-1, the summation of input and feedback currents is equal to zero at the input grid (assuming the input grid current is negligible), or

$$i_1 + i_2 = 0. \qquad (1)$$

From Fig. 4-1 we find that

$$i_1 = \frac{e_i - e_g}{Z_i} \qquad (2)$$

and $$i_2 = \frac{e_o - e_g}{Z_f}. \qquad (3)$$

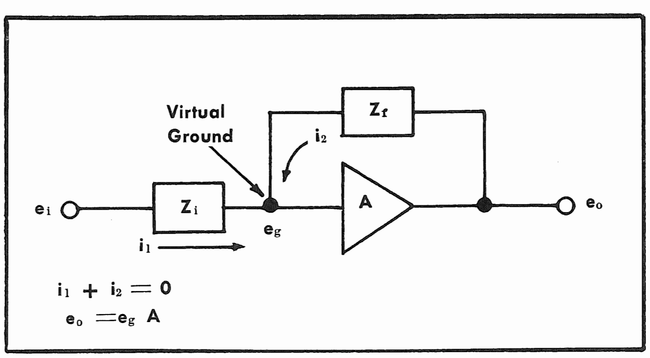


Fig. 4-1. Generalized feedback circuit.

Substituting these values for i_1 and i_2 into equation (1), we have

$$\frac{e_i - e_g}{Z_i} + \frac{e_o - e_g}{Z_f} = 0. \qquad (4)$$

Since $e_g = e_o/A$ (where A is the open-loop gain, a negative number), equation (4) becomes

$$\frac{e_i - \frac{e_o}{A}}{Z_i} + \frac{e_o - \frac{e_o}{A}}{Z_f} = 0. \qquad (5)$$

Rewriting equation (5):

$$\frac{e_i}{Z_i} - \frac{e_o}{Z_i\,A} + \frac{e_o}{Z_f} - \frac{e_o}{Z_f\,A} = 0. \qquad (5)$$

Then

$$e_o\left(\frac{1}{Z_f} - \frac{1}{Z_f\,A} - \frac{1}{Z_i\,A}\right) = -\frac{e_i}{Z_i}. \qquad (6)$$

Solving for e_o:

$$e_o = -\frac{Z_f}{Z_i}\left[\frac{e_i}{1 - \frac{1}{A}\left(1 + \frac{Z_f}{Z_i}\right)}\right] \qquad (7)$$

The gain of a closed-loop amplifier is

$$G = \frac{e_o}{e_i} = -\frac{Z_f}{Z_i}\underbrace{\left[\frac{1}{1 - \frac{1}{A}\left(1 + \frac{Z_f}{Z_i}\right)}\right]}_{\text{Error Factor term.}} \qquad (8)$$

Virtual Ground at the — Grid

(The concept of a virtual ground applies only when the +grid is grounded.)

An operational amplifier with negative feedback tends to maintain a very small voltage change at the — grid terminal. Insofar as circuit behavior is concerned, it will appear as if a very low impedance is placed between the — grid and ground; it is considered a "virtual ground".

To find the equivalent impedance at the — grid (Z_g) refer to Fig. 4-1 again, where $e_g = e_o/A$.

Assuming the — grid current to be negligible, the current entering the virtual ground terminal is equal to the current through Z_f:

$$i_2 = -\frac{e_g - e_o}{Z_f} = -\frac{\frac{e_o}{A} - e_o}{Z_f} \qquad (9)$$

and

$$i_1 = -i_2 = \frac{\frac{e_o}{A} - e_o}{Z_f}. \qquad (10)$$

The equivalent impedance looking into the — grid is obtained by dividing e_g by i_1, or

$$Z_g = \frac{e_g}{i_1} = \frac{\frac{e_o}{A}}{\frac{\frac{e_o}{A} - e_o}{Z_f}} = \frac{Z_f}{1 - A}. \qquad (11)$$

Fig. 4-1 can now be redrawn as Fig. 4-2, with the feedback resistor Z_f replaced by the equivalent resistance it creates, Z_g, as shown at the — grid terminal.

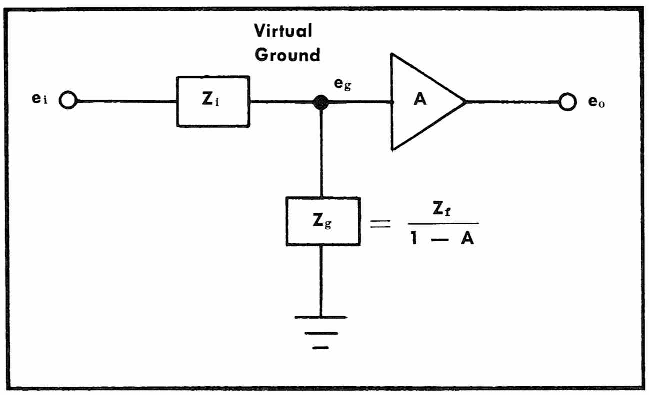


Fig. 4-2. Equivalent virtual ground impedance at the — grid.

Grid Current Calculations

With no external resistor between the grid and ground, and with a capacitor in the Z_f position, the rate, polarity, and amount of change in amplifier output voltage determines the grid current. Fig. 4-3 illustrates a typical grid current measurement setup; the principle is valid for either the — grid or the + grid.

Grid current is measured by essentially measuring the charge stored by capacitor C in a given time after S_1 is opened.

$q = Ce$; then, with C constant,

$$i = C\frac{de}{dt}, \text{ or} \qquad (12)$$

$$i = C\frac{\Delta e}{\Delta t}. \qquad (13)$$

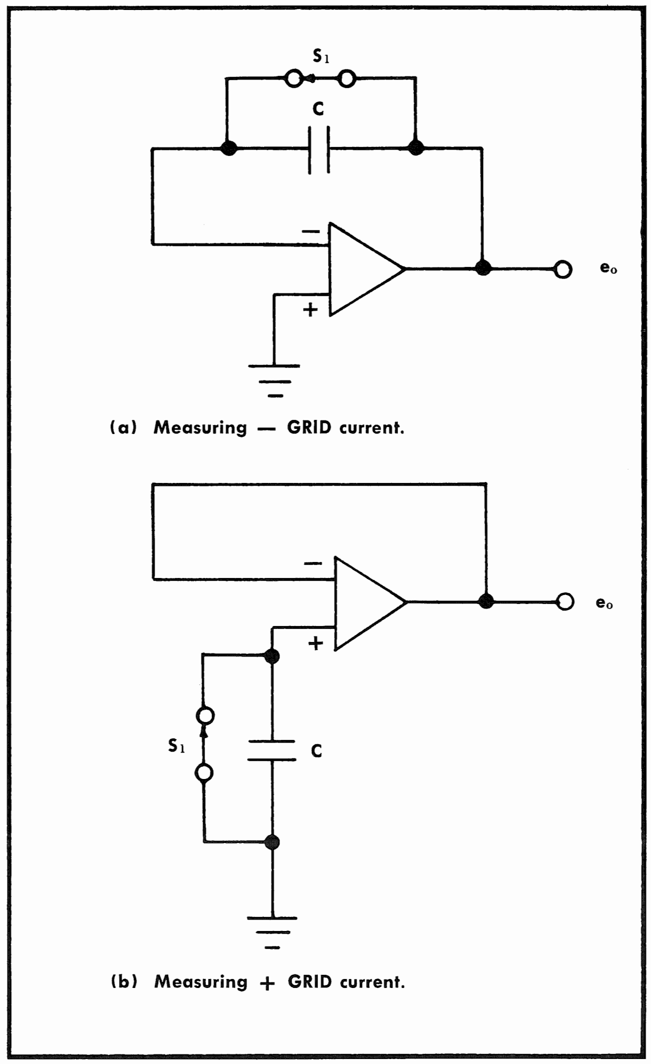


Fig. 4-3. Typical grid current measurement circuits.

If C in Fig. 4-3 is .001 μf, then

$$i = 0.001 \times 10^{-6}\frac{\Delta e}{\Delta t} = \frac{\Delta e}{\Delta t} \text{ nanoamperes.}$$

If the capacitor voltage (after opening S_1) changes 2.5 volts in 5 seconds, the current is 0.5 nanoampere.

Combined Operations

The two operational amplifiers in the Type O Unit permit combined operations to be performed. For example, a signal can be amplified by a desired constant using one amplifier, and its output can be applied to the other for integration or differentiation.

Both operational amplifiers have zero input and output no-signal dc levels, permitting the amplifiers to be stacked. Stacking permits their operations to be combined, greatly increasing their applications.

FORMULA DERIVATION

The remainder of this section is used for the derivation of formulas employed in the Typical Applications portion of Section 3.

Amplification by a Constant

Application 1 shows an amplifier circuit. Both the input and feedback components are resistors. The concept of a virtual ground at the amplifier input is used in the following expression for the circuit gain.

The total current entering and leaving the junction at the — grid must be equal to zero. If we consider the grid current to be negligible (a good approximation of the actual case), then all current entering the junction through R_i must leave through R_f. The current through R_i must therefore be e_i/R_i. Because of the virtual ground at the — grid, the output from the amplifier is the voltage across R_f. The output voltage

$$e_o = -\frac{e_i}{R_i} R_f,$$

where the — sign results from the direction of the current through R_f. The gain of the amplifier is then

$$G = \frac{e_o}{e_i} = -\frac{R_f}{R_i}.$$

Thus, the gain depends only on the input and feedback resistances. The gain can therefore be controlled by adjusting the ratio of R_f to R_i. In a practical application, use the desired gain to find the ratio of R_f to R_i and then select appropriate values to give this ratio. In many cases you can obtain the required ratio by means of the internal Z_i and Z_f resistors. In other cases it will be necessary to use external resistors.

Integration

Application 2 shows an operational amplifier used for integration. The circuit is the same as for amplification except that R_f is replaced by a capacitor. Using the concept of the virtual ground, the current through R_i is found to be e_i/R_i. Assuming the grid current to be negligible, the current through R_i charges the feedback capacitor. The output of the amplifier is essentially the voltage across the capacitor, again because of the virtual ground at the grid of the amplifier. The voltage across the capacitor (and also the output voltage) is

$$e_o = -\frac{1}{C_f}\int i \, dt,$$

but since $i = e_i/R_i$, it can be written as

$$e_o = -\frac{1}{R_i C_f}\int e_i \, dt.$$

Thus, the output voltage of the amplifier is proportional to the integral of the input voltage. The proportionality constant $1/R_iC_f$ gives the scaling of the output voltage. In a practical application, the values of R_i and C_f must be chosen to provide a useful output. If the values of either or both R_i and C_f are decreased, the output of the integrator will increase. The value of R_i includes any internal resistance of the signal source and the virtual ground resistance at the input grid.

A good starting point in a practical problem is to let the R_iC_f product equal (approximately) the period of the waveform to be integrated. This choice generally results in a satisfactory output voltage.

Integration tends to eliminate the high-frequency components in a waveform. This is because the output voltage from the integrator is inversely proportional to the frequency of the input. This can be seen if we assume a sinusoidal input waveform of the form $e_i = E \sin \omega t$. Substituting for e_i in the preceding equation:

$$e_o = -\frac{1}{R_i C_f}\int E \sin \omega t \, dt$$

$$e_o = \frac{E}{R_i C_f \omega} \cos \omega t$$

where E is the peak amplitude, ω is the angular frequency, and t is time.

This equation shows that the output voltage is inversely proportional to the frequency of the input signal.

It should be noted that the output voltage is leading the input by exactly 90°. Thus the integrator can be used as a 90° phase shifter for sinusoidal input signals.

The gain of an integrator circuit as a function of frequency is determined by

$$G = \frac{e_o}{e_i} = \frac{\frac{E}{R_i C_f \omega}}{E} = \frac{1}{R_i C_f \omega} = \frac{1}{2\pi R_i C_f f}.$$

Differentiation

The circuit for differentiation is shown in application 3. The input component is a capacitor and the feedback component is a resistor. Since the — grid is at virtual ground, the current through the capacitor at the input is

$$i = C_i \frac{de_i}{dt}.$$

The output voltage is the voltage across the feedback resistor:

$$e_o = -iR_f.$$

Substituting for i in the above equation:

$$e_o = -R_f C_i \frac{de_i}{dt}.$$

The output voltage of the differentiation circuit shown in application 3 is directly proportional to frequency. This is apparent if we assume an input signal of the form $e_i = E \sin \omega t$. Substituting in the preceding equation for e_i:

$$e_o = -R_f C_i \frac{d}{dt}(E \sin \omega t)$$

$$e_o = -R_f C_i E \omega \cos \omega t.$$

The output voltage is thus proportional to frequency ($\omega = 2\pi f$) and is shifted 90° with respect to the input signal. This permits use of the differentiator as a 90° phase shifter and as a frequency-to-voltage converter as described in application 17.

Summation

Application 4 shows the circuit for a summing amplifier. Because of the virtual ground at the — grid, the current in the total R_i network is

$$i = \frac{e_1}{R_1} + \frac{e_2}{R_2} + \frac{e_3}{R_3} + \frac{e_4}{R_4} \ldots + \frac{e_n}{R_n}.$$

The output voltage of the summing amplifier is obtained by multiplying R_f by the current through it.

$$e_o = -i R_f = -\left(\frac{R_f}{R_1} e_1 + \frac{R_f}{R_2} e_2 + \frac{R_f}{R_3} e_3 + \frac{R_f}{R_4} e_4 \ldots + \frac{R_f}{R_n} e_n\right).$$

Any of the input voltages may be amplified by a constant before the summation is made. In the special case where

$$R_1 = R_2 = R_3 = R_4 = R_n = R_f$$

the equation reduces to

$$e_o = -(e_1 + e_2 + e_3 + e_4 \ldots + e_n).$$

Displaying B-H Curves for Magnetic Materials

The circuit for displaying B-H curves for magnetic materials is illustrated as application 28. The magnetizing force H and the flux density B of a magnetic sample can be calculated from the following.

H: Magnetizing Force

The relation between H and the current in a magnetic core is given by

$$H = \frac{N_1 i}{l}$$

where H is the magnetizing force in ampere-turns per meter, N_1 is the number of turns in the primary winding, i is the primary current in amperes, and l is the mean length of the magnetic path in meters.

The voltage across the current sampling resistor is due to the primary current through it. This current is equal to e/r, where e is the voltage across r applied to the horizontal deflection system of the oscilloscope. Substituting for i in the equation:

$$H = \frac{N_1 e}{l r}.$$

The voltage e can be determined from the calibrated horizontal deflection factors of the oscilloscope. Since all other factors are known, H can then be determined.

ϕ: Flux

The secondary voltage of the transformer e_2 is given by

$$e_2 = N_2 \frac{d\phi}{dt}$$

where N_2 is the number of secondary turns in the transformer, and $d\phi/dt$ is the time rate of change of flux. Since

$$d\phi = \frac{e_2}{N_2} dt,$$

$$\phi = \int \frac{e_2}{N_2} dt.$$

If the secondary voltage of the transformer is applied to the O Unit integrator circuit, the output voltage from the integrator is

$$e_o = -\frac{1}{R_i C_f} \int e_2 \, dt.$$

It can be resolved from the preceding equations that

$$e_o = -\frac{N_2}{R_i C_f} \phi \text{ and}$$

$$\phi = -\frac{R_i C_f}{N_2} e_o.$$

The output voltage from the integrator circuit e_o can be measured from the oscilloscope display by means of the calibrated vertical deflection factors. Since all other factors are known, the flux can be calculated.

B: Flux Density

The flux density B is the total flux ϕ divided by the cross sectional area A, or $B = \phi/A$. The cross sectional area of the core can be determined from its physical dimensions. Substituting for ϕ in the preceding equation and solving for B:

$$B = -\frac{R_i C_f}{N_2 A} e_o.$$

It is therefore possible to determine both H and B from the oscilloscope display and the constants of the particular configuration.

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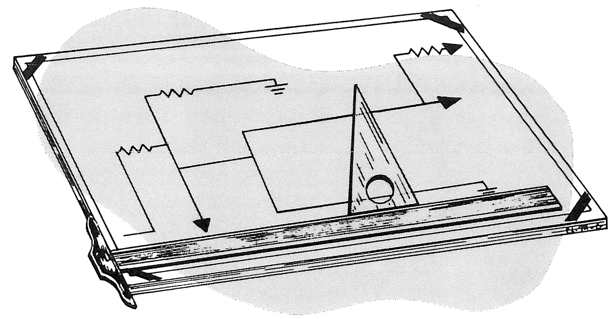

SECTION 5

CIRCUIT DESCRIPTION

General

This section of the manual describes the Type O Unit circuitry with reference to the block diagram and schematics in Section 9.

Block Diagram

The block diagram shows that the output of the two operational amplifiers and the external input signals are individually applied to the VERTICAL DISPLAY switch. The switch selects the desired input and applies it to a step attenuator. The attenuator reduces the signal to the required level before it is applied to the amplifier stages. The output of the amplifiers is then applied to the vertical amplifier of the oscilloscope.

Preamplifier

Input signals to the preamplifier are selected by the VERTICAL DISPLAY switch. Signals are then applied through the ZERO CHECK switch to the VOLTS/CM switch and its attenuator. The ZERO CHECK switch allows input signals to be quickly disconnected so that the zero-signal dc level of the preamplifier can be determined.

The VOLTS/CM switch attenuators are frequency-compensated voltage dividers that reduce the input signal amplitude to a level suitable for driving the grids of the input stage. Each attenuator presents an input impedance of 1 megohm paralleled by 47 pf, regardless of signal frequency. The frequency compensation is accomplished by a variable capacitor across the input resistor of each divider. The input capacitance is adjusted with a variable capacitor in parallel with both resistors of the divider. In the $\times 2$ attenuator, for example, these capacitors are C6508C and C6508B, respectively.

No attenuator is used in the .05 position of the VOLTS/CM switch. The preamplifier input capacitance in this position is adjusted to 47 pf by means of either C6521 or C6541, depending on whether the VERTICAL DISPLAY switch is set to a + or — position.

Input signals from the attenuator are applied to the grid of either V6524 or V6544 depending on the setting of the VERTICAL DISPLAY switch. Positive-going signals applied to the grid of V6524 appear on the oscilloscope crt in the normal position. Positive-going signals applied to the grid of V6544 are inverted. Resistor R6518 prevents excessive input tube grid current if a large positive signal is applied to the input without attenuation. C6518 prevents R6518 from affecting the high-frequency response of the amplifier.

The input stage of the preamplifier, V6524 and V6544, is a cathode-coupled paraphase amplifier. The signal is applied to one grid, and the other grid is at ac ground at all times. Cathode coupling produces equal and opposite (push-pull) output signals. The VARIABLE control R6530 (concentric with the VOLTS/CM switch), varies the cathode degeneration and hence the stage gain. With the VARIABLE control at its detent position (minimum resistance), the GAIN ADJ. control R6536 is used to set the preamplifier maximum gain by varying the total cathode current.

The DC BAL. control R6533 permits an exact balance in dc cathode currents of V6524 and V6544. This prevents any change in dc voltage at the cathodes of V6524 and V6544 when the VARIABLE control is rotated. When the DC BAL. control is adjusted correctly, the crt trace will not shift vertically when the VARIABLE control is rotated.

The push-pull output of the paraphase amplifier is applied directly to the bases of output amplifier transistors, Q6564 and Q6574. The output amplifier is a difference amplifier with emitter coupling. The emitter-coupling network employs a time constant equal and opposite to the transistor thermal time constant, providing low-frequency gain stabilization. The collector circuits contain networks to improve the high-frequency response of the amplifier. L6564 and L6574 peaking coils are adjustable, permitting proper collector circuit compensation for the capacitance of the interconnecting plug and the oscilloscope main-unit vertical amplifier.

The preamplifier input capacitance is stabilized against changes in the setting of the VARIABLE VOLTS/CM control by neutralization from the collectors of Q6564 and Q6574 to the grids of V6524 and V6544. When a signal is applied to the grid of V6524, for example, the collector voltage of Q6564 changes in the same direction as the cathode of V6524. When the VARIABLE control is rotated, increasing its resistance, the signal voltage appearing at the cathode of V6524 will increase while the signal voltage appearing at the cathode of V6544 will decrease. When neutralizing capacitor C6574 is properly adjusted, the change in effective capacitance between the grid and cathode of V6524 will be offset by the opposite change in effective capacitance between the grid of V6524 and the collector of Q6564. The net result is that the effective input capacitance remains constant. The same effect is produced on the other side of the amplifier when the signal is applied to the grid of V6544.

Operational Amplifiers

Since the two operational amplifiers are identical, only the Amplifier A will be discussed.

Input signals to the operational amplifier are normally applied through the A INPUT connector on the front panel of the unit. The signals are then connected through the input impedance selected by the Z_i SELECTOR switch to the grid of V5524. Or, the input signals may be applied directly to the grid through the — GRID connector (or the ± GRID connector, providing the ± GRID SEL switch is in the (—) position). Signals connected directly to the —GRID connector bypass the Z_i SELECTOR switch.

Tubes V5524 and V5534 comprise an input difference amplifier stage. Signals applied to the grid of V5524 (the —grid) are inverted at the A OUTPUT connector while signals applied to the grid of V5534 (the + grid) are not inverted. Signals applied to either grid are amplified and applied from the plate of V5524 to the bootstrap cathode follower V5543A. The polarity of the signal at the grid of V5543A depends upon the polarity of the input and on whether the signal is applied to the + or — grid.

Type O plug-ins from serial number 814 and up have the Zener diodes replaced by more temperature-stable components.

Zener diode D5528 has been replaced by the combination of Q5523 and V5539. In the later circuit the voltage regulator tube V5539 sets the base-to-collector voltage of Q5523. V5539 is maintained in a conducting state by current supplied through R5539 from the —150 volt supply. The emitter voltage of Q5523 will follow the base voltage so that from emitter to collector, the transistor will have a voltage across it of about 82 volts.

With a constant voltage established across the transistor the current through the transistor will be constant. The percent of drift per degree centigrade is less for both the voltage regulator tube and the transistor than it is for the Zener diode. This circuit will therefore provide extremely good stability with temperature changes.

At serial number 814 Zener diode D5529 was replaced by the series combination of D5529 (a smaller voltage Zener diode) and V5529. V5529 is a voltage regulator tube which has a very small percentage of drift per degree. D5529 has been installed in series with V5529 to provide the proper voltage.

The temperature stability of a Zener diode improves as the voltage rating of the Zener diode approaches five or six volts, hence this circuit will have very good stability with a temperature change.

The plate load resistors of V5524 are part of a bootstrap circuit driven by the cathode of V5543A through Q5523 (Zener diode D5528) and capacitor C5528. The bootstrap circuit makes the plate load resistors of V5524 appear larger than they actually are, and the supply voltage (+350 v) much higher than it actually is. This permits the plate voltage of V5524 to rise and fall with almost no change in plate current.

V5543A also drives the grid of the output cathode follower V5543B through V5529, D5529 (Zener diode D5529) and capacitor C5529. Any voltage change at the cathode of V5543A is coupled to the grid of V5543B without attenuation. D5529 assures that the A OUTPUT connector voltage can be set at ground.

The open-loop gain of this system is less than 2500. To increase the gain, a small amount of positive feedback is applied from the A OUTPUT terminal to the cathodes of the input stage through R5547 and the OPEN LOOP GAIN A control R5548.

To assure that the A OUTPUT connector voltage can be set at ground, the screen voltage in the input stage is adjustable. The DC LEVEL RANGE A control (internal adjustment) and the front-panel OUTPUT DC LEVEL control permit the screen voltage of one of the input tubes to be raised while the screen voltage of the other tube is lowered. Thus the conduction of V5524 can be changed to suit the required ground level output voltage when there is no input signal.

The front-panel OUTPUT DC LEVEL ADJ. switch, SW5520, allows the output dc level of the operational amplifier to be adjusted to ground even when input circuits and signals are connected. SW5520 disconnects all input circuitry to the amplifier and switches in a feedback network that makes the operational amplifier gain 100. The 100× gain permits the operator to make the OUTPUT DC LEVEL adjustments with good display resolution.

A second form of screen voltage adjustment of the input tubes permits the selection of the amplifier operating conditions that are required for minimum grid current. Minimum grid current occurs when the operating bias of the input tubes is adjusted to the correct value. A limited adjustment range for the screen voltage of the input amplifier (an adjustment that changes both screen voltages equally), assures that the proper operating conditions for minimum grid current can be obtained. By adjustment of the GRID CURRENT A control, R5535, the input stage operating bias can be changed to the optimum value without appreciably affecting the system gain. Since the V5524 plate circuit is a bootstrap system, a change in the bias caused by a small change in the screen voltage will not significantly disturb the output dc level.

When an operational amplifier is connected for integration, normal amplifier drift and small dc components of the signal will also be integrated. The result is that the display will be slowly forced off the crt. To prevent this, a special feedback network can be switched in series with the integrator Z_f component. The special network is composed of R5514, R5515, C5514, and C5515. With the front-panel INTEGRATOR LF REJECT switch in either the 1 CPS or 1 KC position, any dc or low-frequency changes at the A OUTPUT connector are applied through the feedback network to the grid of V5524, restoring the trace position on the crt. The high-frequency components of the integrated signal are not fed back to the grid of V5524.

By means of the INTEGRATOR LF REJECT switch, the time constant of the feedback network can be selected. In the 1 CPS position, the time constant of the feedback network is about 1 second. Signals much above 1 cps are not fed back to the grid of V5524. In the 1 KC position, the time constant is about 1 millisecond. In this case signals much above 1 kc are not fed back to the grid of V5524. The 1 KC position can be used to prevent the integration of line-frequency hum pickup or other noises, while still permitting the integration of desired signals above 1 kc.

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UNDERSTANDING OPERATIONAL AMPLIFIERS

Operational amplifiers are important in measurements and R&D work. Capable of simulating a variety of circuits, they can be used to add, subtract, differentiate, integrate or amplify either linearly or with controlled nonlinear coefficients, under signal conditions. Here is a chance to learn how they work.

By GEOFFREY GASS
Staff Field Engineer
Tektronix, Inc.
Beaverton, Oregon

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UNDERSTANDING OPERATIONAL AMPLIFIERS

OPERATIONAL AMPLIFIERS are devices which make use of negative feedback to process signals with a high degree of accuracy. This accuracy is limited in the ideal case only by tolerances in the values of the passive elements in the input and feedback networks.

An operational amplifier is a high-gain amplifier designed to remain stable with large amounts of negative feedback from output to input.

General-purpose types are useful for linear amplification with precise values of gain, and for accurate integration and differentiation operations. These have low output impedance and are dc-coupled, with the output dc level at ground potential.

* * *

To obtain negative feedback from the output to the input, the output must be inverted with respect to the input. Negative feedback (Z_f, Fig. 1b)—through a resistor, capacitor, inductor, network or nonlinear impedance—is applied from the output to the input. The input to which negative feedback is applied is generally termed $-$input or $-$grid.

An operational amplifier, using negative feedback, operates like a self-balancing bridge. It provides whatever current is needed through the feedback element to hold the $-$input at null (ground potential). The output signal is a function of this current and the impedance of the feedback element.

The $-$input, held to ground potential by the feedback current, appears as a low impedance to any signal source. Using resistive feedback, for instance, the input appears to be the resistance of the feedback element, divided by the open-circuit gain of the amplifier.

If current is applied to the $-$input, it would tend to develop voltage across the impedance of the feedback element, and move the $-$input away from ground. The output, however, swings in the opposite direction, providing current to balance the input current and hold the $-$input at ground. If the feedback element impedance is high, the output voltage must become quite high to provide enough current to balance even a small input current.

Since, usually, we are concerned with voltage rather than current signals, an additional element is used in most applications; input impedance Z_i. This impedance, in series with the $-$input, converts into *current* that parameter of the input signal appearing as voltage at the output; Fig. 1c.

Voltage Amplifier

If Z_i and Z_f are both resistors (Fig. 2), the operational amplifier becomes a simple voltage amplifier, the gain of which is $-Z_f/Z_i$.

When a voltage is applied to A in Fig. 2, current flows through Z_i. Were it not for the amplifier, this current would also flow through Z_f and to ground through the low impedance at C; Z_i and Z_f would then operate as a voltage divider, raising the voltage at B. But, the amplifier serves to hold the voltage at B ($-$input) at ground potential. To insure such operation, the amplifier must supply a voltage at C so that current through Z_f exactly equals the current flowing through Z_i. When B is thus held at ground potential, the voltage across Z_i is obviously equal to the applied voltage at A.

The current through Z_i is equal to the applied voltage at A divided by the impedance (in this case, resistance) of Z_i, or E_{in}/Z_i. This identical value of current must flow through Z_f to keep B at ground. The voltage at C, then, must be E_{in}/Z_i (the value of current in Z_f) multiplied by Z_f. The output is inverted from the input; thus $E_{out} = (-E_{in})\left(\frac{Z_f}{Z_i}\right)$, and the voltage gain of this amplifier setup is $-Z_f/Z_i$.

Differentiation

Earlier, it was noted that an operational amplifier with a resistor as a feedback element responds with an output voltage equal to the product of the input current and the feedback resistance. Now let us see what happens if a capacitor is used instead of resistor Z_i; Fig. 3.

The current through a capacitor is proportional to the *rate-of-change* of the voltage across the capacitor. A steady state dc voltage across a capacitor (assuming an *ideal* capacitor) passes no current through the capacitor, and thus no balancing current need be furnished by the output to hold the −input of the amplifier at ground. Output voltage then is zero.

If the voltage at the input is changed, however, the *change* causes a current to flow through the capacitor. Amount of current that flows is directly proportional to the capacitance times the *rate-of-change* of the input voltage.

To show this, let us assume that the potential at the input is +100 vdc, and it is changed smoothly to +95 vdc in 5 sec. This represents a rate-of-change of 1 v/sec, the change taking place over a period of 5 sec. If the value of Z_i is 1 µf, then, a current of −1 µa will flow through Z_i for that 5-sec period.

The amplifier will cause an equal and opposite current to flow in Z_f. If a value of 1 megohm were selected for Z_f, the 1 µa necessary to balance the circuit will require +1 v to appear at the output during the time 1 µa flows through the capacitor.

This operation is differentiation: sensing the rate-of-change of an input voltage, and providing an output voltage proportional to that rate-of-change.

The actual relationship of output to input is: $E_{out} = -\left(\frac{dE_{in}}{dt}\right)(RC)$, where the expression dE_{in}/dt indicates the rate-of-change (in volts-per-second) of the input signal at any given instant, and R and C are Z_f and Z_i, respectively.

In our example, a constant rate of change was used and a constant voltage level out was obtained. Had the rate been less even, the output signal would have shown this dramatically with wide variations in amplitude. The differentiator senses both the rate and direction of change. It is very useful in detecting small variations of slope or discontinuities in waveforms.

Integration

If the resistor and capacitor used for differentiation were interchanged, a resistor used for Z_i and a capacitor for Z_f (Fig. 4), the characteristics would be exactly opposite. While in differentiation an output voltage proportional to the rate-of-change of the input was obtained, by interchanging the resistor and capacitor, the output signal becomes a rate-of-change proportional to the input voltage.

This characteristic permits the use of the operational amplifier for integration. This is because the instantaneous value of the output voltage at any time is a measure of both amplitude and duration (up to that instant) of the input signal.

In the operational amplifier, integration operates in the following manner: Let us assume the conditions where $Z_i = 1$ meg, and $Z_f = 1$ µf, and the input signal level is 0 v. No current flows through Z_i; thus, no balancing current is needed through Z_f. Suppose now a dc voltage of −1 v is applied to Z_i. This will cause a current of −1 µa to flow in Z_i, and the amplifier will have to provide a balancing current through Z_f. To obtain a steady current of 1 µa through 1 µf, a continually rising voltage will be needed at the output, the rate of rise being 1 v/sec. This rate of rise will continue until the input voltage is changed, or the amplifier reaches its swing limit or approaches its open-loop gain.

This rate-of-rise, though helpful in understanding the mechanism by which the amplifier performs integration, does not provide the answer sought from an integrator. The significant characteristic is the exact voltage level at a certain time, or after a certain interval.

Before the amplifier reaches its output limit, suppose the input voltage to Z_i is removed. The output does not return to ground, but remains at the level it reached just before the signal was removed. Rate of rise has stopped, because the necessity for providing 1 µa through Z_f to maintain the null at the −input has been removed. With an ideal capacitor and amplifier, the output voltage would remain indefinitely at the last level reached, until an input signal of opposite polarity were applied to Z_i, and a negative-going rate-of-change at the output were needed to maintain the null at the −input.

Absolute integrator output level at the end of some interval is the sum of the products of each voltage applied to Z_{io} since the output was at 0, times the duration of each voltage, that sum divided by −RC.

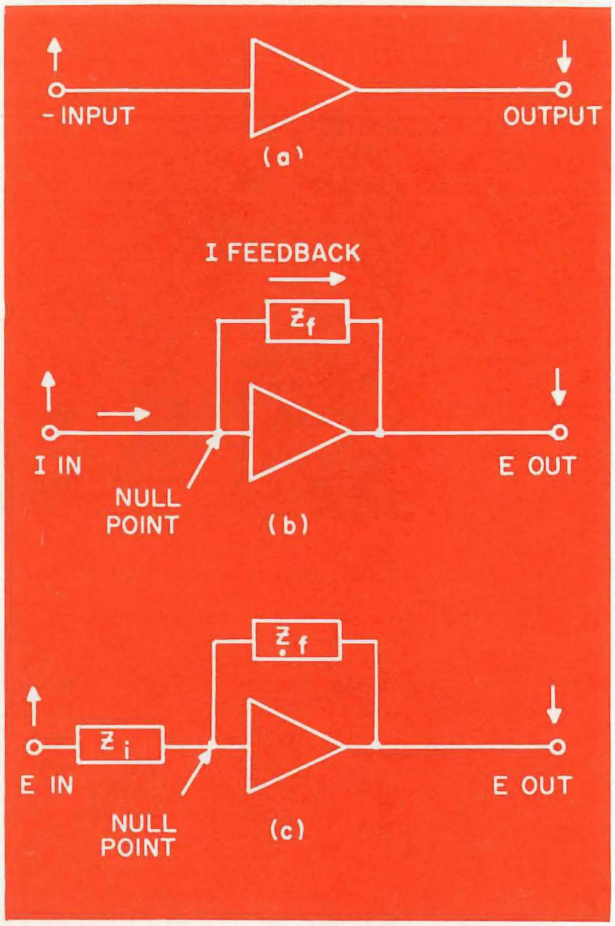
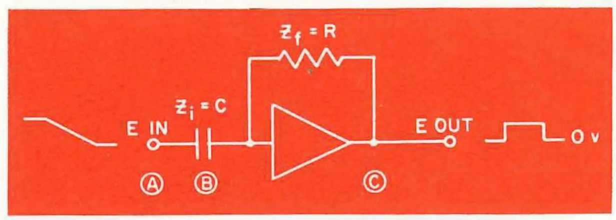


Fig. 1: Normal operational amplifier symbols are shown in (a.) The —input and output are out-of-phase (arrows). Feedback element Z_f is shown in (b). In (c) the input element Z_i converts a voltage signal (E_{in}) to current, balanced by current through Z_f.

Fig. 2: Operational amplifier, using resistors for both Z_i and Z_f, performing as a voltage amplifier.

Fig. 3: Operational amplifier as a differentiator. Output here is proportional to the rate of change of the input voltage.



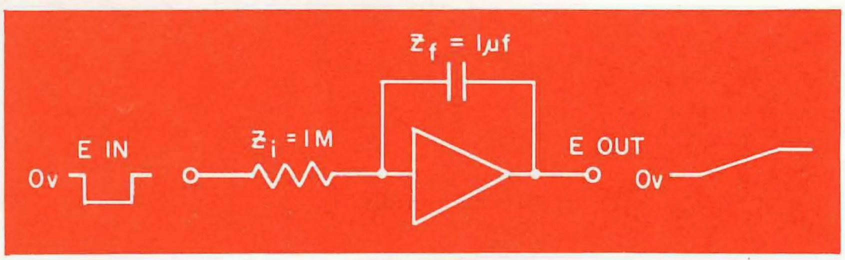
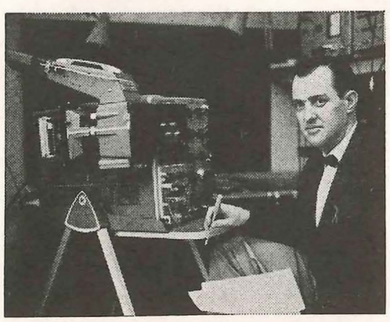
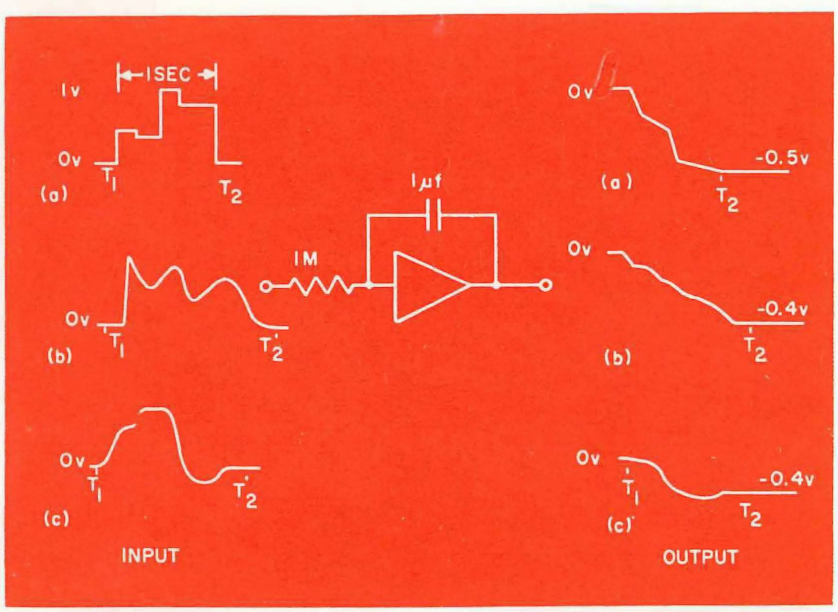


Fig. 4: Operational amplifier as an integrator (left). Output rate of change in this instance is proportional to the input voltage.

Fig. 5: Series of waveforms (right) integrated to determine "area under the curve" between T_1 and T_2. Note that in (c) the negative portion of the input waveform reduces the net integral.



By GEOFFREY GASS
Staff Field Engineer
Tektronix, Inc.
Beaverton, Oregon



The mathematical expression for the output level reached in a given interval of time (T_2-T_1) is:

$$E_{out} = \left(\frac{-1}{RC}\right) \int_{T_1}^{T_2} E_{in}\, dt$$

The integral sign indicates that the value to be used is the sum of all of the products ($E_{in} \times dt$) shown, between the limits (T_1, T_2) noted. The expression *dt* shows infinitely small increments of time.

It is not necessary to understand and be able to manipulate expressions in integral calculus to make use of an operational amplifier integrator.

The integrator provides a voltage output proportional to the net number of volt-seconds applied to the input. If the total volt-seconds of one polarity is equalled by those of the opposite polarity, the output level at the end of the selected interval will be zero.

Typical integrator waveforms are shown in Fig. 5. An interesting example is shown in (c). Its 4 voltage levels, of different duration, cause the integrator output to fall at 4 different rates, reaching a final level representing the total number of volt-seconds contained in the waveform. It should be apparent that the integrator can measure the total volt-seconds contained in even the complex waveform of (b). Such a measurement would be difficult to make by direct waveform observation. This type of operation is often referred to as "taking the area under the curve," since the area underneath a waveform plotted against time (i.e., the area bounded by T_1, T_2, the waveform and the 0 v line) is the number of volt-seconds involved. Also, the instantaneous value of E_{out} at any time is proportional to the input volt-seconds up to that time.

In the cases used as examples, RC was 1 ($10^6 \times 10^{-6}$), and the numerical value of the output voltage at the end of the integrating interval was the number of volt-seconds in the input waveform. Using other values of R and C does require some added calculation. To find actual input volt-seconds, the output voltage must be multiplied by ($-RC$). To illustrate, let us suppose R is 200,000, C is 0.01 μf and the output voltage, after the selected interval, is -2.5 v. Multiplying -2.5 by ($-2 \times 10^5 \times 1 \times 10^{-8}$) provides 5×10^{-3}, or 5 mv/sec positive polarity. Because of the polarity-reversal in the amplifier, it is necessary to multiply by ($-RC$) to obtain the proper sign in the answer.

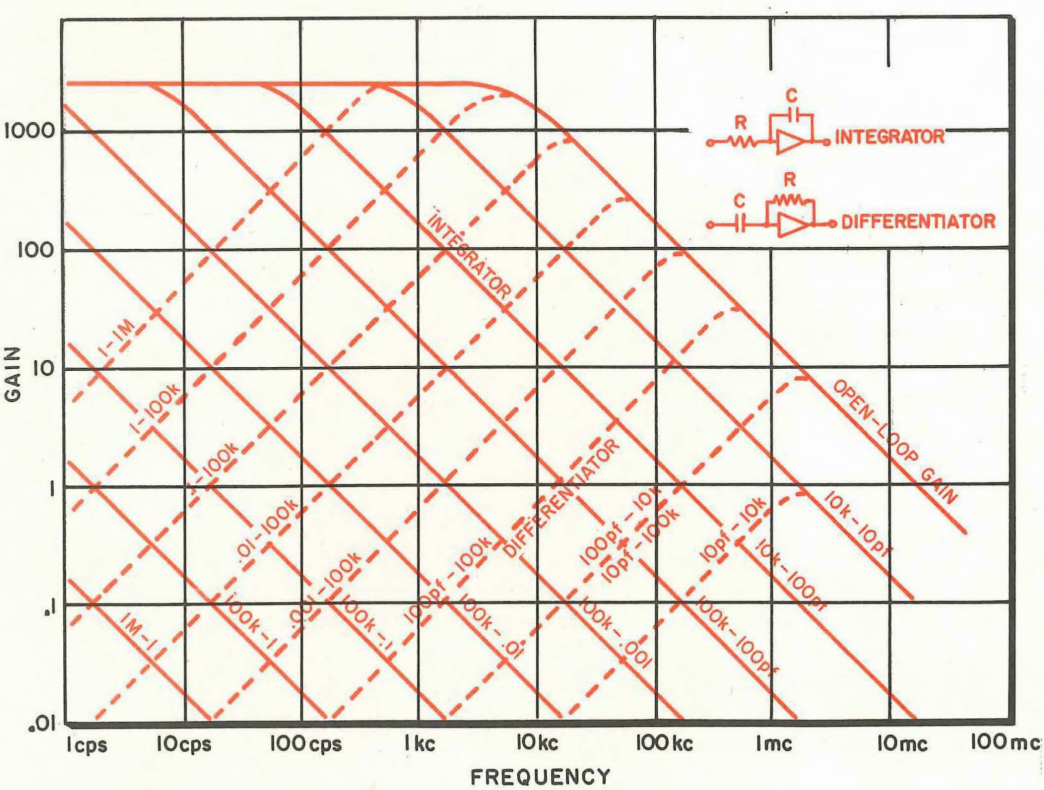


Fig. 6: Average gain characteristics for integration and differentiation.

Fig. 7: Operational amplifier driving tunnel diode with very low impedance (r). Diode, stabilized at high frequencies by rf-terminated jig, can be driven by a very slow ramp voltage. Linear drive of the tunnel diode allows differentiation of the current to obtain the di/de curve (see Fig. 8).

OPERATIONAL AMPLIFERS (Concluded)

If a waveform to be integrated contains both + and − polarity portions during the integrating interval (d) in Fig. 5, the output will be proportional to the *difference* between the volt-seconds of each polarity, the integrator being an averaging device. To add the 2 polarities, instead of allowing them to be subtracted, precede the integrator with an absolute-value amplifier (full-wave rectifier) which inverts one of the polarities.

The operational amplifier is now being used not only as an electrical and electronic tool, but in control engineering and for mathematical studies.

As an electronic device, the amplifier can be used to make capacitance, resistance and impedance measurements. In solid-state fields, the amplifier has been used to study tunnel-diode performance through plots of E-I and di/de curves. Also, diode reverse leakage current can be measured and B-H curves plotted. Many sampled data studies can also be made: sampling and sampled-pulse integrations and differentiation.

Using semiconductors and tubes as external active elements, the amplifiers are also being used for a number of nonlinear functions: compression, expansion, root and power function, limiting, clipping and fast-response log-arithmetic amplification.

The amplifier can also be used to provide rate-intensification of oscilloscope traces for photography of transients. Other uses include dc-coupled current measurement and high-input impedance amplification.

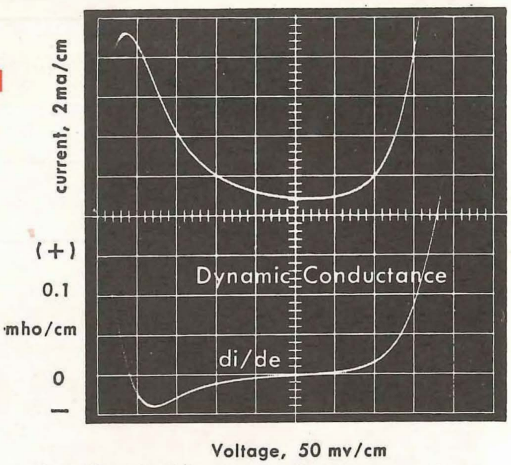
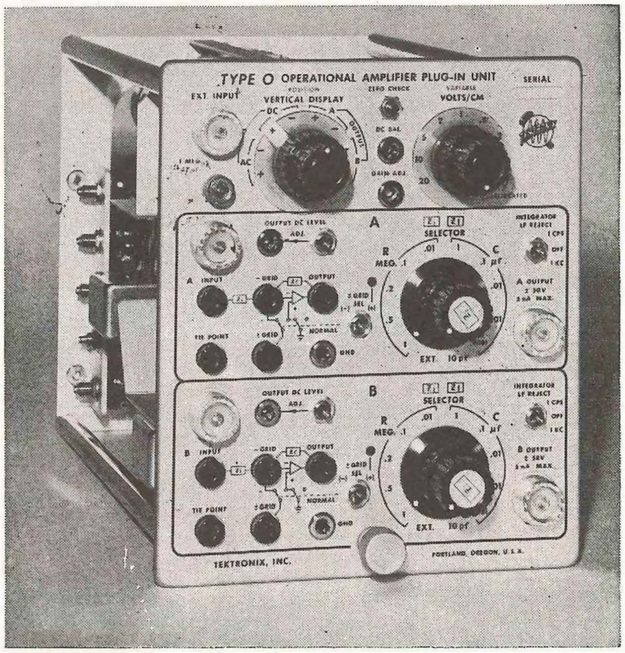
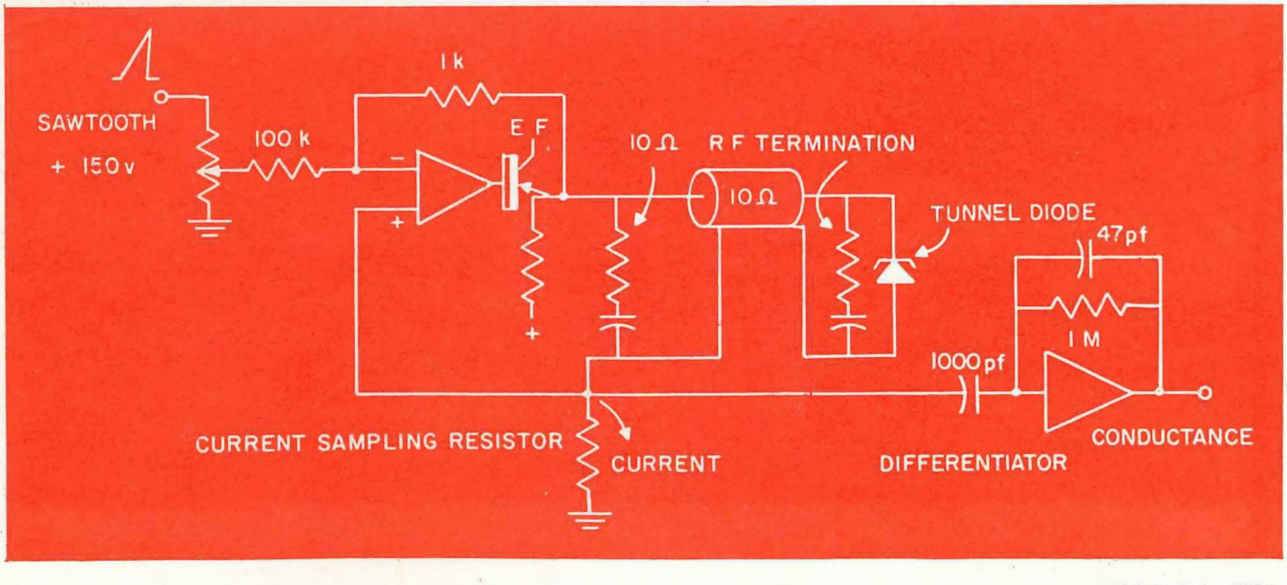


Fig. 8: Dynamic Conductance waveform.

Fig. 9: Tektronix Type O Operational Amplifier Plug-In Unit.





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USING THE O UNIT FOR "CONSTANT INTENSITY" DISPLAYS

FEN 2-23-62

The Type O Plug-in can be used to provide differential beam-brightening for fast segments of a displayed waveform at slow sweep speeds, to facilitate viewing and trace photography. (At faster sweep speeds, this feature is not usually required, because the vertical and horizontal spot velocities are more nearly alike.)

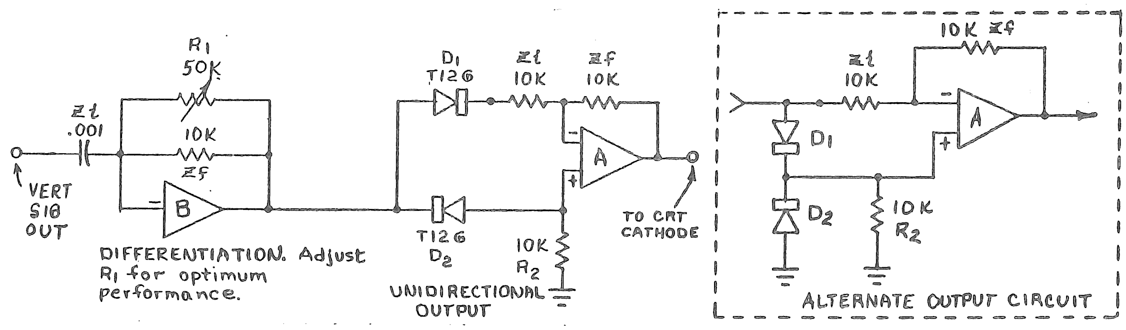


Fig. 4. O Unit circuit for "constant intensity" feature. All Components internal to O Unit except R_1, R_2, D_1, and D_2.

Figure 4 shows the O Unit settings and connections to provide (1) differentiation of the vertical signal and (2) "absolute value" operation (negative-going signal regardless of input signal polarity).

Use of the circuit allows reducing the intensity control setting at slow sweeps, and making the "flat tops" and base line more nearly equal in intensity to the rising and falling portions of the waveforms. Do attempt to make rising and falling portions as bright as base lines and flat tops at "normal" intensity -- it's not only impossible, but leads to defocussing and geometry aberrations.

There is a certain amount of transit-time delay in the intensification circuitry--apparently a good part of a microsecond. For this reason, even the presence of the delay line and CRT transit time is not sufficient to give 100% brightening of a fast transistion. However, as can be seen in Figures 5 and 6, the circuit can provide a vast improvement in many applications.

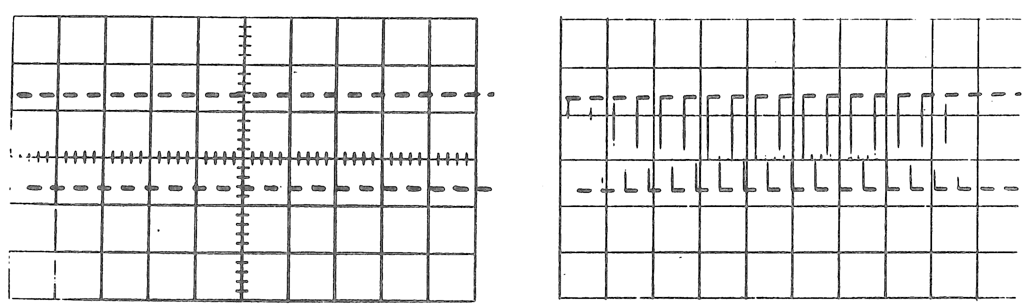


Fig. 5 Waveform Improvement; Calibrator -- 2 msec/cm

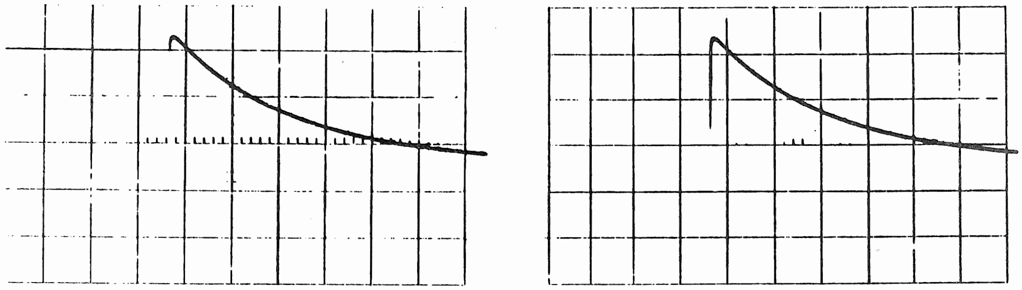


Fig. 6 Waveform Improvement; 10 msec time mark -- 20 μsec/cm

DIFFERENTIAL BEAM BRIGHTENING USING THE O UNIT -- ADDENDUM FEN 3-9-62

The circuit shown in the last issue of the FEN (2-23-62) is not the only -- or even the best-- circuit to use for differential beam brightening. A more economical circuit using only a single operational amplifier is shown on page 25 of Hiro Moriyasu's "Notes on Operational Amplifier Applications," copies of which were sent last week to all field engineers and repair centers. Variations on any of these circuits may be worked to fit specific applications problems.

To obtain measurements in real time from operations on sampled data (e.g., integration of diode stored charge from 661/4S1/291/TF1), it is essential that the scaling factor for time conversion be known and held constant--i.e., that the relation between equivalent time (horizontal deflection on the sampling scope display) and real time (the time by the clock it takes the sampling trace to cross the screen) be known at all times, since the O-Unit operates in real time, and the sampling display data is in equivalent time.

The relationship between real and equivalent time is determined by the sampling rate and the samples/cm setting unless this relationship is *forced*. The sampling rate is governed by the triggering signal or vertical signal rep-rate up to about 100 KC, then by the sampling rate limit.

If the sampling oscilloscope sampling rate cannot be fixed at a known, stable value below 100 KC, it becomes necessary to *force* the real/equivalent time relationship by driving the sampling scope horizontal with the real-time saw-tooth from the O-Unit's scope. (It is *not* necessary to trigger the real-time scope.)

To obtain a sufficient dot density for accurate integration, it may be necessary to run the real-time scope sweep quite slowly. A density of 10-100 samples per cm is generally desirable. Very high density is needed for differentiation, of course.

When the horizontal drive to the sampling oscilloscope is correctly adjusted so that the (non-integrated) displays on the sampling scope and the real-time scope occupy an equal number of centimeters horizontally, the time conversion factor is the ratio of the Time/Cm settings of the real-time and the sampling oscilloscope. An external, adjustable divider will be required to attenuate the 150 v saw-tooth to the horizontal input of a type 661; this is not required with the 3T77.

The vertical conversion factor may be calculated step-by-step (sampling input volts to centimeters, to Vert-Sig-Out volts, through the operational amplifier, to volts and finally to centimeters again), but the errors, loading effects and tolerances involved make this calculation quite uncertain without a lot of extra work.

The most accurate method (eliminating any sampling scope display or output calibration error, loading effects, etc.) is to apply the sampling oscilloscope calibrator to the sampling input, set the operational amplifier for X1 amplification using the same value of Z_i that will be used in the intended measurement, and adjust the O-Unit Volts/Cm and Variable Volts/Cm for calibrated deflection on the real-time scope. For a vertical scaling factor of 1:1, set the O-Unit to display the same number of centimeters as the sampling scope *should* be displaying.

The remaining errors will be due to (a) sampling scope calibrator error, (b) tolerance buildup in the operational amplifier from setting up with one value of Z_f and measuring with another, and (c) error due to measurement technique.

Particular care must be taken to avoid *technique* errors, particularly in integration. Since DC levels are critical in integration, the level representing the reference from which the measurement is to be made must--by adjustment of the DC offset control in the sampling oscilloscope--be made to be 0 volts DC at the input to the integrator. Any error in judgement of the reference level may be magnified considerably in the answer. The vertical signal output of the 3S series sampling systems comes out at a DC level of about +10 v. See addendum for special considerations required for use of these instruments.

Because integration is a bit tricky, a detailed procedure and example follows, assuming use of a 535A or 545A to provide positionable gating.

1. Set up real-time scope for "B delayed A" operation, so that Sawtooth A may be used to drive the sampling oscilloscope (this special setup is not necessary on a 555).
   (a) Horiz display, "A". Free-run Sweep A at 20 msec/cm.
   (b) Patch DELAYED TRIG to B EXT TRIG IN. Adjust B Trig controls for triggering from the delayed trigger.
   (c) Patch +Gate B to the gating adapter.
   (d) Set Sweep B to 20 msec/cm, with the delay time multiplier set to 0.05 and Sweep Length B to about 3 o'clock.

2. Set up operational amplifier A for gated X1 amplification, using the same value R_i as will be used in integration (set R_f to the same value). Check DC levels and balance and confirm proper triggering of Sweep B, with input to R_i grounded.
3. If a 661 is used, attenuate Sawtooth A to 5 v/cm or less (a 20 K pot equipped with banana plugs is adequate for slow sweeps) and apply to the sampling scope external horizontal input. If the sampling unit is a 3T77, the full 150 v sawtooth may be used and attenuated internally by the 3T77 Ext Atten pot.
4. Patch from the Vert Sig Out on the sampling oscilloscope to the gated amplifier. Adjust v/cm of the real-time oscilloscope and the saw tooth attenuator until the display on the real-time scope matches "exactly" (e.g., to the desired accuracy) the display on the sampling oscilloscope. For maximum accuracy, use the amplitude/time calibrator (if available) for simultaneous setting of the vertical and horizontal scaling factors.

   If 1 cm on the real-time scope vertical now represents 1 cm on the sampling scope vertical, the amplitude conversion factor is 1.0. Whatever is represented by 1 cm vertical deflection on the sampling scope will be represented by 1 cm on the real-time scope--volts, amps, psi or whatever.

   Return the Time/Cm controls on the real-time scope to a slow setting (e.g., 0.1 sec/cm) which permits a good sampling density. The sampling scope time per cm divided by the real time/cm gives you the horizontal scaling factor, something like 10^{-7} or 10^{-8}.
5. Adjust Delay Time Multiplier and B Time/Cm and Sweep Length so that the gated portion of the display corresponds to the part of the waveform to be integrated.
6. Adjust DC offset in sampling oscilloscope so that the start of the segment to be integrated rests at 0 volts on the real-time scope (good idea to recheck operational amplifier DC output level again, too). Note: In the 4S-sampling series, only the DC offset affects the Vert Sig Out level. In the 3S-series, *both offset and position controls* affect this level, as well as whatever circuitry is used to bring the center-screen output level to 0 volts.
7. Turn Z_f to an appropriate value of C to obtain a few cm of deflection. Record integral in *centimeters* of deflection reached just before the gate is closed.
8. Ground the input to R_i and record integrating error, if any. Subtract from figure obtained in step 7. This gives you the integral in terms of centimeters.
9. Compute integral. The answer you are looking for ($\int N dt$, where N is the quantity represented on the sampling scope by vertical deflection) is equal to: (N/cm) X(Cm)X(R_iC_f)X(K), where N/cm is the *sampling* scope sensitivity, Cm is the number of centimeters net integral (answer minus error, from step 8) and K is the horizontal scaling factor. If N is volts, your answer comes out in volt-seconds. If the vertical scaling factor is other than 1, multiply by this factor also.

Example (Using 661, 4S1, 5T1, O-Unit and 545A):

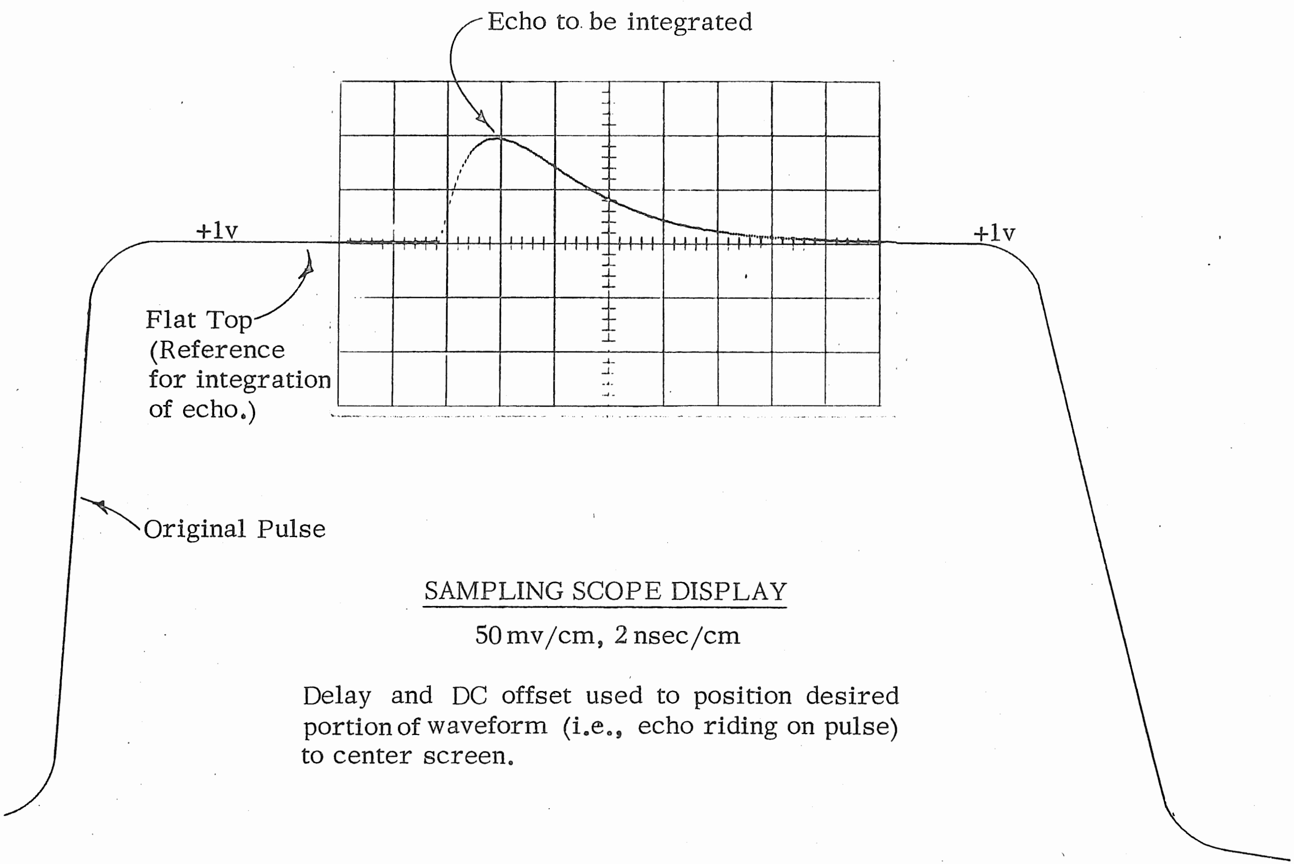
Determine the energy in a reflected pulse from a transmission line. The reflection occurs during the flat top of the original +going pulse, which is 1 volt in amplitude.

On the sampling scope at 2 nsec/cm and 50 mv/cm, the reflection appears as a positive pulse of 2 cm peak amplitude, with a fairly steep leading edge, the trailing edge occupying 4-5 cm before tailing out completely.

Procedure:

1. *Adjust Vertical Scaling Factor:* Using the 661 calibrator at 10 μsec/cycle and 100 mv for best accuracy (at the 50 mv/cm vertical setting), slow the sampling sweep to 10 μsec/cm and, with the Vert Sig Out connected to the operational amplifier input (X1 amplifier using 0.2 meg R_f and R_i) adjust the O-Unit vertical preamp for just 2 cm vertical display. (It is not necessary to correct the sampling oscilloscope v/cm under these circumstances even if it is out of calibration.) The vertical scaling factor is now 1:1--and with a direct relationship between input voltage to the sampler and deflection on the real-time scope, the intermediate calibrations cancel out.

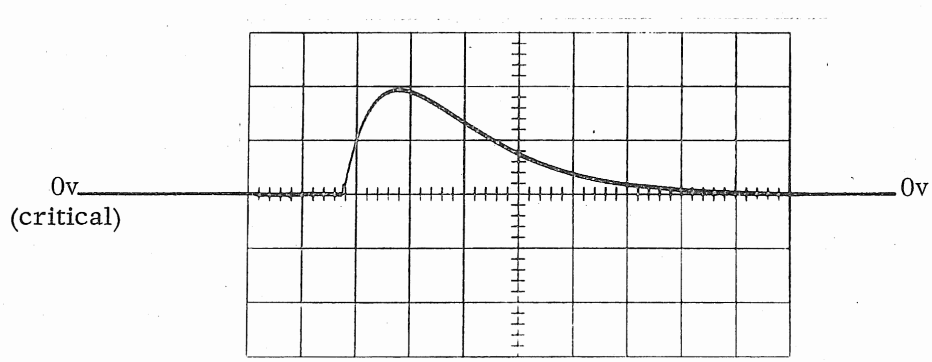
   The scaling factor must be rechecked if R_i is changed, however.
2. *Adjust Horizontal Scaling Factor:* Set the 661 Time Calibrator to 10 nsec/cycle and the sampling sweep to 2 nsec/cm, and adjust the real-time sawtooth attenuator so that two full cycles occupy 10.0 cm on the real-time oscilloscope screen. (The lead from the gating adapter to the -grid of operational amplifier A may be removed momentarily to permit a full 10.5 cm display.) Then restore the waveform to be measured, and set A and B Time/cm to a slow enough rate to obtain a sampling density of at least 10 samples/cm. We'll assume that 100 msec/cm is used. The horizontal scaling factor becomes $2X10^{-9}/10^{-1} = 2X10^{-8}$.



<u>SAMPLING SCOPE DISPLAY</u>

50 mv/cm, 2 nsec/cm

Delay and DC offset used to position desired portion of waveform (i.e., echo riding on pulse) to center screen.



<u>REAL TIME SCOPE DISPLAY</u>

Sampling scope offset must be carefully set so that reference level for integration of echo is at 0 rolts at input to operational amplifier. (Step 3- Z_f set to R = R_i)

3. *Set DC Level:* After checking the DC level setting of operational amplifier A, adjust the DC offset in the sampling scope so that the flat top of the original pulse is represented on the real-time scope by a level of 0 volts. This is extremely critical--any error here will be magnified in the answer. See fig. 1.
4. *Set Integrating Interval:* Use the Delay Time Multiplier and Sweep B length to set the gating interval to extend from just before the pulse reflection to just following its "tail out".
5. *Integrate:* Set Z_f to C (be sure the LF reject switch is "off") and select a value of C which will provide a few centimeters of output. With Z_i set at 0.2 meg and real time/cm at 100 msec/cm, the proper value of C_f turns out to be (say,) 1 μf. The output level reaches 2.2 cm at the end of the integrating interval. Adjustment of B sweep length to a little longer value does not materially increase the value, so we can feel we have "all" or the pulse.
6. *Check Error:* Grounding the input to R_i shows no climbing or falling of the baseline during the gated interval, so our first answer does not need correction.

7. *Calculate:* $\int Edt$ = (Sampling Scope v/cm setting) X (Cm of Integrator Output on Real Time Scope) X (R_i C_f) X (K). Plugging in our numbers, we have (.05 v/cm) X (2.2 cm) X (0.2 X 10^6 X 1.0 X 10^{-6}) X (2 X 10^{-8}) = 4.4 X 10^{-10} = 440 pico-volt-seconds. In a 50 Ω system, this represents 19.5 X $10^{-20}/50$ watt-secmnfs, or 3.9 nanopico-joules of energy, and each reflection contains 44/50 X 10^{-11} = 8.8 pico coulombs, or 55 million electrons.
8. *Probable Accuracy Check:* Step 6 checked only the DC-level errors in the gated integrator system itself (drift, leakage, zero shift, etc.). It did not check the most probable cause of error, the DC level of the incoming signal from the sampling scope.

   Either inaccurate setting of the DC offset or subsequent drift of the output level can cause a significant error in the integrator output.

   To determine the probable accuracy with which your measurement was made, repeat step 3, this time observing the extremes of offset adjustment which still appear "right". Now, repeat steps 4 and 5, adjusting the DC offset level between the two extremes noted. The two integrator output levels you obtain will indicate the possible error range of any single measurement.
9. *Direct Calibration* Once the vertical scaling factor has been set to 1 and the values of v/cm, R_i, C_f and the horizontal scaling factor have been set, the real-time scope graticule may be considered as directly calibrated in N X seconds, where N is the parameter displayed vertically on the sampling oscilloscope. In the example above, the direct calibration was 0.2 nanovolt-seconds per cm. Note, however, that if watt-seconds or joules is the required parameter, either the input to the sampling system must be in watts (per cm), or else the real-time calibration must be considered as proportional only to $\sqrt{\text{joules}}$ per centimeter, and the value obtained must be squared (and then multiplied by an appropriate factor) to obtain the numerical value for energy.

Limitations: In addition to the problems discussed above, there are some basic limitations which should be recognized in attempting real-time operations on sampled data.

1. DC Drift (Integration). A small amount of drift in the DC offset voltage in a 4S-series sampling instrument (or offset and positioning voltage drift plus drift of any external level-setting circuit in the 3S-series) is not bothersome when the sampling instrument is operated at low sensitivity, with relatively long integrating time constants and small horizontal scaling factors. But a combination of worst-case conditions of operation can produce serious measurement errors.

   Worst-case Conditions:

   (a) Low sampling rate. To obtain high dot-density, the real-time sawtooth scan must be run very slowly if the sampling rate is low. A given drift-rate then will have more time to be effective. Also, the tendency of the sampling instrument to "slash" (memory discharge) will be greater, and the integrating error due to grid current, etc., will also increase at very slow sampling rates.

   (b) High sampling scope sensitivity setting. A shift of 1 mv (0.1%) in a 1 v DC offset level is of minor consequence at 100 mv/cm. At 2 mv/cm, it may be quite serious, especially if the real-time involved in the integration is long. At 2 mv/cm, and with a one-second time-constant in the O-Unit (largest internal value), a shift of 1 mv in a 1 v offset will produce an error of 10 mv integrator output per 100 msec real time.

   Cures: The higher the sensitivity required, the faster the minimum sampling rate must be to avoid errors due to low dot-density or real-time errors due to drift. Conversely, if the sampling rate is low, the signal level must be larger to avoid the effects of small internal level changes.

2. Noise (Differentiation). Because a sampling trace is not a continuous function, differentiation is bound to produce a "noisy" output, reflecting the vertical "steps" representing amplitude changes. The lower the sampling density the greater the possible amplitude step per sample--and hence noise spike in the differentiator output.

   With a moderate sampling rate, the 10 K output impedence of the vertical signal output helps in reducing sampling noise, but in many cases, an additional capacitor across R_f will be needed to further reduce high-frequency gain and limit the response of the differentiator to the significant real time analog information from the high-speed signal being sampled.

   As the sampling scope sensitivity is increased toward maximum, the inherent noise in the sampler will be of significant effect, and maximum use of the smoothing control and use of resistance in series with C_i and capacitance across R_f will be needed to obtain meaningful answers. Operation at the maximum possible horizontal scaling factor will be necessary in some cases, to effect as great as possible a frequency-domain difference between the sampling noise and the real time analog of the signal under observation, in order that the two can be separated by filtering.

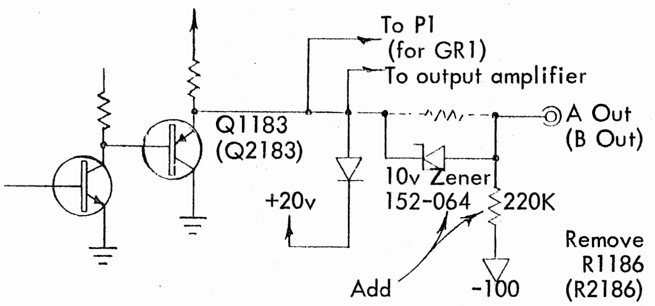
Addendum: Special considerations for using the 3S-series sampling systems.

All 3S-series sampling vertical outputs, to accomodate the special requirements of the 6R1, come out at an amplitude of 1 v/cm, at a dc level of approximately +10 v (center screen) and an impedance of 10 K. For useful integration, it is necessary to drop this DC level to 0 v, or at least perform the equivalent of this operation at the operational amplifier. Whatever method is employed, the *stability* requirement of the level-setting circuit is fairly critical (see notes on drift limitations). The 3S-series output signal is large, however, and under most circumstances fairly simple circuitry may be employed to give quite usable answers.

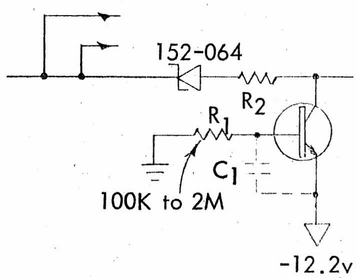
Four possible techniques are open:

1. Put a 10 v battery in series with the output signal. Since our real-time signal is relatively low-frequency information, the stray capacitance of the battery does not seriously affect accuracy.
2. Make the 10 K series output resistor in the plug-in into a dc level-setting divider by running a 12.2 K resistor (or thereabouts--say, 12.4 K 1/2 w 309-350) from the output end of this resistor to the -12.2 v supply. The output signal will now be 0.525 to 0.565 v/cm at a dc level close to ground, and the plug-in calibration will be made about 1% low.
3. Add a zener diode and 220 K constant-current resistor to the vertical signal output circuit instead of the 10 K resistor. This trick will translate the dc level to near 0 v without loss of signal amplitude. (See Fig. 1a). Select the zener diode to match the actual center-screen dc level if possible. A transistor may also be used as a constant-current source (Fig. 1b), for better-going current capability, which may be needed in some cases for differentiation.
4. Add slideback to the operational amplifier itself, to buck out the dc level from the sampling scope. This is done by injecting current to the -grid equal and opposite to the current representing the dc level of the signal to be integrated ($1 \approx 10\,v/R_i$). The source impedance of this slideback signal should be kept high, and of course it must be stable. Whenever R_i is changed, the voltage or the impedance of the slideback circuit must also be changed, so that the *currents* at the -grid balance exactly. If R_i is kept at 1 meg, a 22-1/2 battery, a 2 meg resistor and 500 K pot all in series will do the job nicely (Fig. 2). If this last method is used, grounding R_i to determine integrating error (step 8 in procedure, step 6 in example) will *not* be possible.

G. Gass



(a) Using Zener and resistor.max
output current, 400+μa for-signals.



(b) Using Zener and trancistor
for more output current.

Fig. 1. Modifying 3S76 for 0 v DC level output from front panel connectors. Replace R1186 with 10 v zener diode, provide constant-current source for zener. 3S3,etc. may be treated similarly. If transistor is used, select R_1 for $\approx$1 ma collector current. With zener slightly less than 10 v, R2 (100 Ω to 1 k) and collector current may be selected for exact 0 v center-screen DC level.

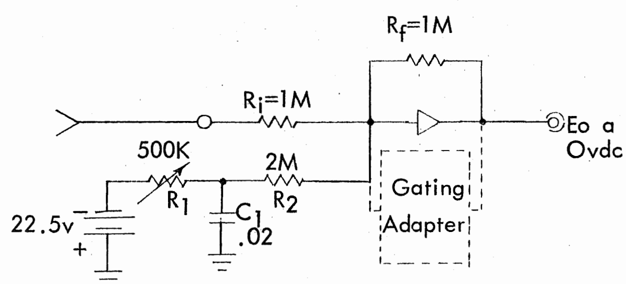


Fig. 2. Slideback added to operational amplifier to handle signal at elevated level. R_2 must be changed whenever R_i is changed. After setting R_1 for 0 v DC output level, Z_f may be changed to C.

TYPE O APPLICATIONS
Page D-15

This cover sheet identifies unnumbered page.

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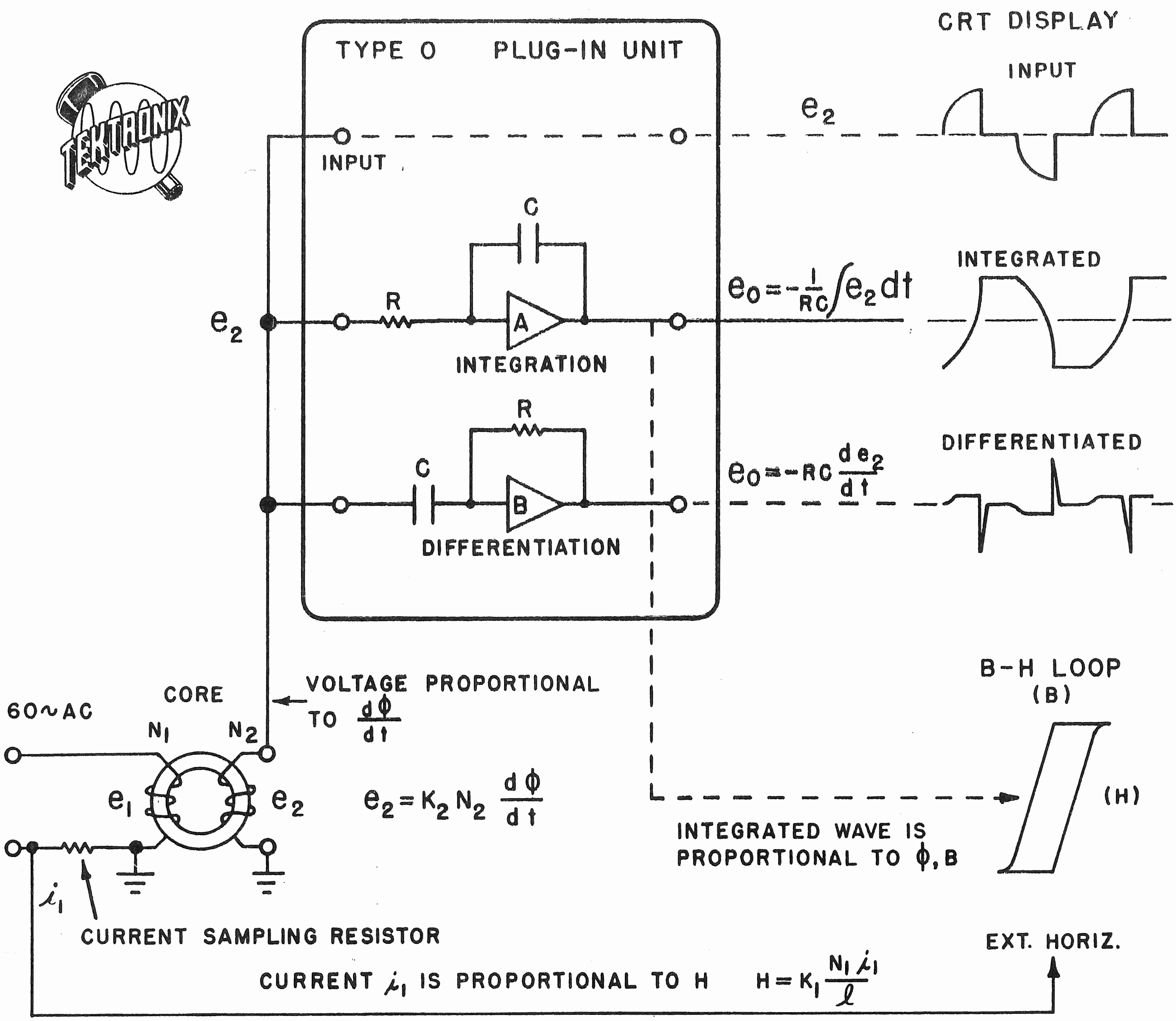
OPERATIONAL AMPLIFIER PLUG-IN UNIT

FOR AMPLIFICATION, INTEGRATION, DIFFERENTIATION, AND NONLINEAR – FEEDBACK APPLICATIONS



MAINTENANCE

CONTENTS-MAINTENANCE

MODIFICATION SUMMARY

0

PREAMPLIFIER FAST RISE RESPONSE IMPROVED

See SQB

M5510

Effective Prod 109
w/exceptions 105, 107

Usable in field instruments SN 101-108

DESCRIPTION:

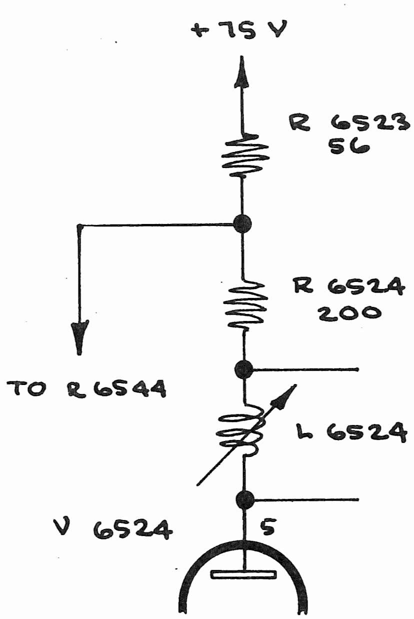
Components in V6524 plate circuit are changed/added to improve the fast rise waveform response. (See schematics below.) Also see M9564.

Parts Removed:

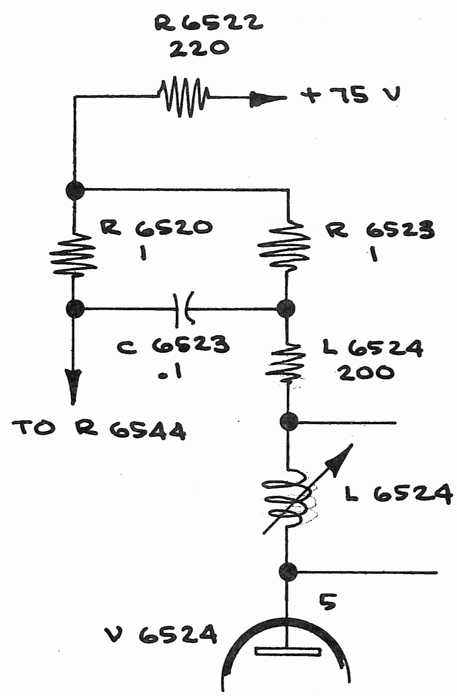
R6523	56Ω 1/4w 10%	316-0560-00

Parts Added:

C6523	0.1 μf 10 v	283-0023-00
R6522	220Ω 1/4w 10%	316-0221-00
R6520, R6523	1Ω 1/2w 5%	308-0141-00



BEFORE



AFTER

Parts Required for Field Installation:

See 'Parts Added'.

INSTALLATION INSTRUCTIONS:

Rewire the V6524 plate circuit as indicated in the schematics above.

A AND B OPERATIONAL AMPLIFIER GRID CURRENT POTENTIOMETERS ADDED

See SQB — M5728

Effective Prod s/n 155 — Usable in field instruments s/n 101-154

DESCRIPTION:

Potentiometers R5535 and R5585 are added to provide adjustment for the input grid current in the operational amplifiers. This necessitates replacing R6557 with a physically smaller potentiometer and changing the pot bracket to one accommodating 5 potentiometers. At the same time, the silk-screening for the Open Loop Gain DC Level range pots was changed by adding an "A".

Parts Removed:

R6557	100 k ±20% AB	311-0301-00
Bracket, pot		406-0796-00

Parts Added:

R6557	100 k 0.2 w mini	311-0088-00
R5534, R5584	12 k 2 w 5%	305-0123-00
R5535, R5585	5 k 10%	311-0171-00
Bracket, pot		406-0848-00

Parts Required for Field Installation:

See 'Parts Added'.

INSTALLATION INSTRUCTIONS:

a) Replace the potentiometer mounting bracket on the rear plate with the larger bracket.

b) Replace R6557 with the smaller potentiometer.

c) Mount the 5 k potentiometers in the R5535 and R5585 positions.

d) Unsolder the white-brown-black-brown wires from the center terminal of R5532 (DC Level Range A). Solder these wires together and cover the splice with a piece of tubing.

e) Move the white-brown-black-brown wire from the center terminal of R5582 (DC Level Range B) to the counter-clockwise (ccw) terminal of R5585 (Grid Current B).

f) Solder a white-brown-black-brown wire between the ccw terminals of R5585 and R5535 (Grid Current A).

g) Solder a 12 k resistor between R5535 center terminal and the rear cw terminal of the POSITION control.

h) Solder a 12 k resistor between R5585 center terminal and the front ccw terminal of the POSITION control.

i) Solder a white-red wire between the center terminals of R5582 and R5585.

k) Solder a white-red wire between the center terminals of R5532 and R5535.

A AND B OPERATIONAL AMPLIFIER
RESISTOR WATTAGE INCREASED

See SQB

M5636

Effective Prod s/n 160

Usable in field instruments s/n 101-159 w/exceptions

DESCRIPTION:

Resistors R5529 and R5579, in the A and B Operational Amplifier, were changed from 1w to 2w to relieve over-dissipation under certain conditions.

Parts Removed:		Parts Added:	
R5529, R5579 47k 1w 10%	304-0473-00	R5529, R5579 47k 2w 10%	306-0473-00

Parts Required for Field Installation:

See 'Parts Added'.

INSTALLATION INSTRUCTIONS:

Replace R5529 and R5579, located on the ceramic strips near V5524 and V5574 respectively, with 2w resistors.

A AND B OPERATIONAL AMPLIFIER
OUTPUT AMPLITUDE ACCURACY
AND CROSS-TALK RATIO IMPROVED

See SQB

M5956

Effective Prod s/n 319

Usable in field instruments s/n 101-318

DESCRIPTION:

Capacitors C5535 and C5585 are added, and circuitry is changed to increase the accuracy of the output amplifier when Zf is set at 10pf and the INTEGRATOR LF REJECT is OFF, and to improve the cross-talk ratio.

This modification is included in Field Modification Kit 040-0301-01.

Parts Removed:		Parts Added:	
		C5535, C5585 0.01 µf 150v	283-0003-00

Parts Required for Field Installation:

See 'Parts Added'.

INSTALLATION INSTRUCTIONS:

See page 6.

continued

INSTALLATION INSTRUCTIONS:

NOTE: The following method is used to identify the SELECTOR switch terminals:

The wafers are numbered from the front to the rear.
The contact positions are numbered 1 through 12 relative to the index key, as shown in Fig. 1.
The contacts have an 'F' or 'R' suffix which denotes that they are on the front or the rear of the wafer.

Example: W2-7R (denoted by * on Fig. 1) is contact 7 of the rear of wafer 2.

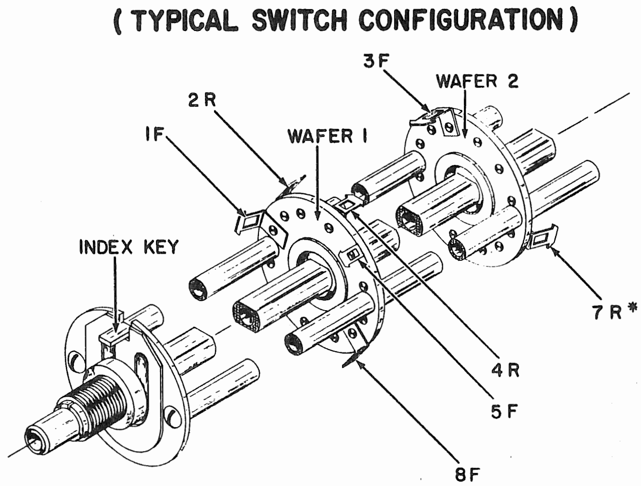


Fig. 1

a) Unsolder the following components and wires from the Amplifier 'A' INTEGRATOR LF REJECT switch:
   100k 1/4w resistor (save for re-use)
   white-yellow wire
   1 meg 1/4w resistor between switch terminals (save for re-use)

b) Unsolder the end of the 0.0022 µf capacitor from the switch and resolder to the terminals indicated in Fig. 2. Clean the excess solder from the switch terminals.

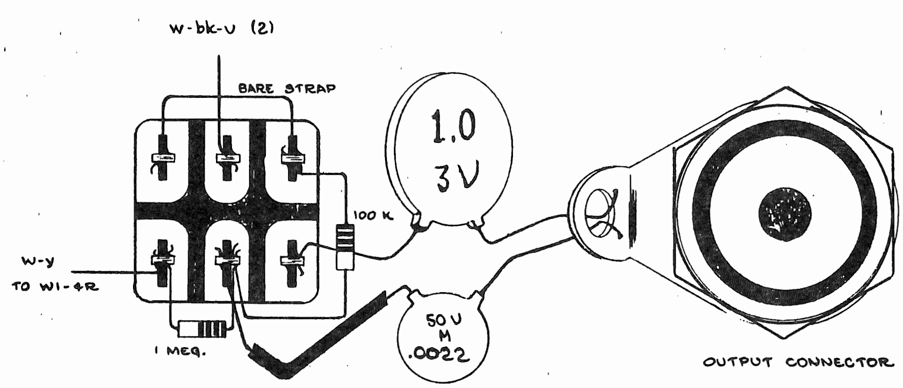


Fig. 2

Installation Instructions: (con'd)

c) Unsolder the white-black-violet wire from 'A' SELECTOR switch W2-8R.

d) Remove the nylon spacing clip between the white-black-violet and white-yellow wires.

e) Locate the 100k 1/4w resistor on the 'A' SELECTOR switch, soldered between W1-5R and W2-8R. Unsolder the end from W2-8R and resolder to W1-8F.

f) Dress the white-black-violet wire, unsoldered in step c, to the 'A' LF REJECT switch. Cut off the excess wire and solder, along with another 6-1/4 in. length of white-black-violet wire, to the switch terminal indicated in Fig. 2.

g) Resolder the remainder of the components and wires, as indicated in Fig. 2, using a length of bare wire.

h) Dress the other end of the white-black-violet wire (step f) through the grommet in the 'A' SELECTOR switch bracket and solder to W2-8R.

i) Locate the 100k 1/4w resistor on the 'B' SELECTOR switch, soldered between W1-5R and W2-8R. Unsolder the end from W2-8R and resolder to W1-8F.

k) Unsolder the following components and wires from the 'B' INTEGRATOR LF REJECT switch:
100k 1/4w resistor (save for re-use)
1 meg 1/4w resistor between switch terminals (save for re-use)
bare wire between switch terminals.

m) Unsolder and remove the bare wire between the 'B' INTEGRATOR LF REJECT switch and the 'B' SELECTOR switch W1-4R.

n) Unsolder the end of the 0.0022 µf capacitor from the 'B' INTEGRATOR LF REJECT switch. Resolder it to the terminal indicated in Fig. 3. Clean the excess solder from the switch terminals.

p) Solder a 2-1/2 in. length of white-yellow wire between the 'B' SELECTOR switch W2-8R and the switch terminal indicated in Fig. 3.

q) Solder a 1-1/2 in. legnth of bare wire to 'B' SELECTOR switch W1-4R. Place a 1 in. length of plastic tubing over the wire and dress the wire to the switch terminal indicated in Fig. 3. DO NOT SOLDER until next step.

r) Resolder the components and wires to the switch as indicated in Fig. 3, using length of bare wire.

s) Solder a 0.01 µf ceramic capacitor between the two wired terminals of the Grid Current 'A' potentiometer (R5535) on the potentiometer bracket.

t) Similarly, solder a 0.01 µf capacitor between the two wired terminals of the Grid Current 'B' potentiometer (R5585).

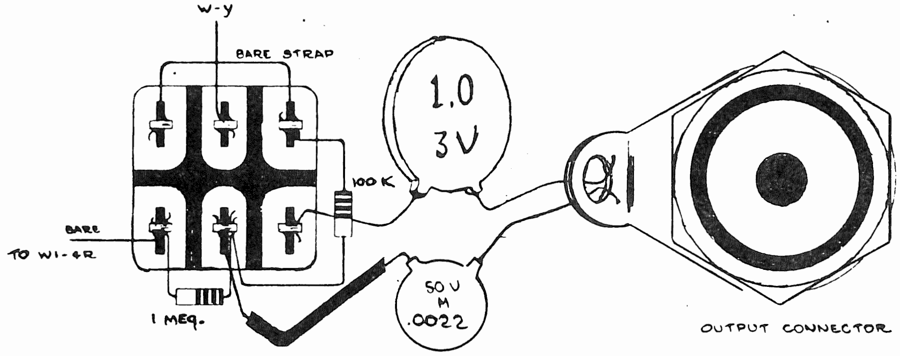


Fig. 3

A AND B OPERATIONAL AMPLIFIER
TUBES REPLACED WITH CHECKED PAIR

See SQB — M5731

Effective Prod s/n 490 — Usable in field instruments s/n 101-489

DESCRIPTION:

To more easily meet specifications for V5524, V5524, V5574, and V5584, new 12AU6 tubes are set up which are checked for microphonics, noise, and grid current.

Parts Removed:

V5524, V5534, V5574, V5584	12AU6 (2 pr)	157-0050-00

Parts Added:

V5524, V5534, V5574, V5584	12AU6, ck'd (2 pr)	157-0071-00

Parts Required for Field Installation:

See 'Parts Added'.

INSTALLATION INSTRUCTIONS:

Replace the tubes in pairs, as indicated above.

A AND B OPERATIONAL AMPLIFIER
GAS REGULATORS ADDED TO REDUCE DC THERMAL DRIFT

See SQB — M6115

Effective Prod s/n 814 — Usable in field instruments s/n 101-813

DESCRIPTION:

Gas Regulator tubes and associated circuitry are added to the A and B Operational Amplifiers to improve the temperature-sensitive drift characteristics of the Operational Amplifiers.

Parts Removed:

R5523, R5573	33 k 1 w 1%	310-0070-00
R5525, R5575	45 k 1/2 w 1%	309-0354-00
R5526, R5527	22 k 2 w 5%	305-0223-00
D5528, D5529, D5578, D5579	1N3044B	152-0087-00

Parts Added:

R5523, R5573	39 k 2 w 5%	305-0393-00
R5525, R5575	40.2 k 1/2 w 1%	309-0437-00
R5526, R5576	22 k 7 w 3%	308-0241-00
R5539, R5589	390 k 1/2 w 10%	302-0394-00
R5529, R5539, R5579, R5589	ZZ1000	154-0370-00
D5529, D5579	11 v 1/4 w 5%	152-0055-00
Q5523, Q5573	NPN, Tek Spec	151-0096-00

Parts Required for Field Installation:

Modification Kit — 040-0301-02

INSTALLATION INSTRUCTIONS:

Refer to kit instructions.

PREAMPLIFIER INPUT TUBES REPLACED | See SQB | M6490

Effective Prod s/n 850 | Usable in field instruments s/n 101-849

DESCRIPTION:

V6524 and V6544 are changed to a pair selected for grid current, grid bias/heater current, and low microphonics. This eliminates an additional grid current check.

Parts Removed:			Parts Added:		
V6524, V6544	12AU6 (1 pr)	157-0050-00	V6524, V6544	12AU6 (1 pr)	157-0077-00

Parts Required for Field Installation:

See 'Parts Added'.

INSTALLATION INSTRUCTIONS:

Replace V6524 and V6544 in pairs.

PREAMPLIFIER GAIN ADJ RANGE INCREASED | See SQB | M6470

Effective Prod s/n 1150 | Usable in field instruments s/n 101-1149

DESCRIPTION:

GAIN ADJ potentiometer R6536 is changed from 5k to 10k to increase the range of adjustment.

Parts Removed:			Parts Added:		
R6536	5k 2w	311-0300-00	R6536	10k 2w	311-0392-00

Parts Required for Field Installation:

See 'Parts Added'.

INSTALLATION INSTRUCTIONS:

Replace the GAIN ADJ potentiometer with the 10k pot.

UHF CONNECTORS REPLACED WITH BNC CONNECTORS

INFORMATION ONLY M6860

Effective Prod s/n 1270

DESCRIPTION:

The UHF connectors are replaced with BNC connectors, to match the military and manufacturing trend toward the BNC type. The BNC type has a constant 50Ω impedance and a lower input capacitance. It also requires less front panel space.

See M8313.

Parts Removed:		Parts Added:	
Connector, UHF, female (5)	131-081	Connector, BNC, female (2)	131-126
Adapter, probe, BNC to UHF (2)	103-015	Adapter, BNC to binding post	103-033

ACCESSORIES CHANGED TO PERMIT PATCHING WITHOUT ADAPTERS

INFORMATION ONLY M8313

Effective date 2-26-65

DESCRIPTION:

To permit patching from BNC to BNC connectors or from BNC to UHF (or banana jack) connectors without the use of adapters, the present patch cords and/or adapters are changed/added as indicated below.

Also, these patch cords are set up as optional accessories:

6 in. red BNC to BNC	012-085
6 in. red BNC to banana plug	012-089
6 in. black BNC to BNC	012-084
6 in. black BNC to banana plug	012-088
18 in. black BNC to BNC	012-086
18 in. black BNC to banana plug	012-090

See M6860.

Parts Removed:		Parts Added:	
Adapter, BNC to binding post	103-033	Patch cord, 18 in. BNC to BNC	012-087

PREAMPLIFIER INPUT CAPACITORS RANGE INCREASED — See SQB — M7437

Effective Prod s/n 1800 — Usable in field instruments s/n 101-1799

DESCRIPTION:

Capacitors C6521 and C6541 are changed to provide a larger range necessary because of the lower input capacity of some preamplifier tubes. This change necessitated a change in the Vertical Display switch part numbers.

Parts Removed:			Parts Added:		
C6521,C6541	1.5-7 pf	281-0034-00	*C6521,C6541	3-12 pf	281-0036-00
SW6500	VERTICAL DISPLAY	262-0423-00	SW6500	VERTICAL DISPLAY	262-0634-00

Parts Required for Field Installation:

See 'Parts Added' with asterisk.

INSTALLATION INSTRUCTIONS:

Replace C6521 and C6541, located on the third wafer of the VERTICAL DISPLAY switch, with the 3-12 pf capacitors.

PLUG-IN SPACER RODS STANDARDIZED — INFORMATION ONLY — M8087

Effective Prod s/n 2230

DESCRIPTION:

Install new plug-in spacer rods which have hex shape near one end to allow better tighening. This will insure more positive grounding of the plug-in to the indicator (via the spacer rods).

Parts Removed:			Parts Added:		
Rod, spacer	(4)	384-0508-00	Rod, spacer	(4)	384-0631-00

VERTICAL AMPLIFIER OSCILLATIONS PREVENTED	See SQB	M8983
Effective Prod s/n 2429	Usable in field instruments s/n 101-2428 w/exceptions	

DESCRIPTION:

Ferramic Suppressor beads are added to the vertical Output Amplifier leads to prevent oscillations at approximately 200mc when the unit is used with Type 547, 546 and 544 instruments. See M9564 which removes L6565 and L6575.

Parts Removed:

Parts Added:

L6565, L6575	0.1 μH	276-0528-00

Parts Required for Field Installation:

See 'Parts Added'.

INSTALLATION INSTRUCTIONS:

Add the cores to the end of the white-brown wire at the top of L6574, and the end of the white-blue wire at the top of L6564. Both wire ends must be stripped back 3/4 in. L6564 and L6574 are the Q6564 and Q6574 Collector peaking coils.

TRANSISTOR SOCKETS CHANGED	INFORMATION ONLY	M8208
Effective Prod s/n 2470		

DESCRIPTION:

Provides a better and more economical way to mount transistor sockets, by replacing sockets with new snap-in type.

Parts Removed:

Socket, 4-pin transistor (2)	136-0095-00

Parts Added:

Socket, 3-pin transistor (2)	136-0181-00
Ring, transistor socket (2)	365-0234-00

A AND B OPERATIONAL AMPLIFIER INPUT LEAD RESISTOR REPLACED WITH FERRITE BEAD TO ELIMINATE TEMPERATURE EFFECTS ON GRID CURRENT — See SQB — M9434

Effective Prod s/n	2651				Usable in field instruments s/n 101-2650				
w/exceptions	2261	2424	2442	2479	2523	2558	2576	2603	2644-5
	2335	2427	2460	2489	2543	2562	2578	2635	
	2401	2429	2474	2509	2556	2570	2585	2639	

FRONT PANEL SYMPTOM: Trace drift with temperature change.

PROBLEM: Humidity and temperature change the ceramic strip leakage resistance, causing the grid current to vary.

PRODUCTION CHANGE: The grid leads were removed from the ceramic strips where they connected to the grids through 47Ω resistors (R5520, R5530, R5570, R5580) and connected directly to the tube socket pin connections (pin 1). Ferrite beads were added over the leads to take the place of the 47Ω resistors (see Figs. 2 and 3).

This modification is included in Modification Kit 040-0301-01.

Parts Removed:

R5520, R5530, R5570, R5580	Resistor, comp, 47Ω 1/4W 10%	316-0470-00

Parts Added:

L5520, L5530, L5570, L5580	Core, type 101	276-0532-00

INSTALLATION INSTRUCTIONS:

Parts Required: See 'Parts Added'.

Installation Procedure:

Refer to Fig. 1 for ceramic strip locations.

a) Remove 47Ω resistor (R5520) between CSF-16 and pin 1 of V5524.

b) Remove 47Ω resistor (R5530) between CSF-19 and pin 1 of V5534.

Refer to Fig. 2 when replacing the following wires:

c) Remove white-black-violet wire between Channel A OUTPUT DC LEVEL switch and CSF-16.

Solder a 3 in. white-black-violet wire between the same terminal of the Channel A OUTPUT DC LEVEL switch and pin 1 of V5524. Place a ferrite bead over the tube socket end of the wire.

d) Remove white-black-blue wire between the same terminal of the Channel A OUTPUT DC LEVEL switch and pin 1 of V5534. Place a ferrite bead over the tube socket end of the wire.

e) Remove 47Ω resistor (R5570) between CSE-16 and pin 1 of V5574.

f) Remove 47Ω resistor (R5580) between CSE-19 and pin 1 of V5584.

continued

Installation Procedure:

Refer to Fig. 3 when replacing the following wires:

g) Remove white-black-violet wire between Channel B OUTPUT DC LEVEL switch and CSE-16.

Solder a 7in. white-black-violet wire between the same terminal of the Channel B OUTPUT DC LEVEL switch and pin 1 of V5574. Place a ferrite bead over the tube socket end of the wire.

h) Remove the white-black-blue wire between Channel B OUTPUT DC LEVEL switch and pin 1 of V5584. Place a ferrite bead over the tube socket end of the wire.

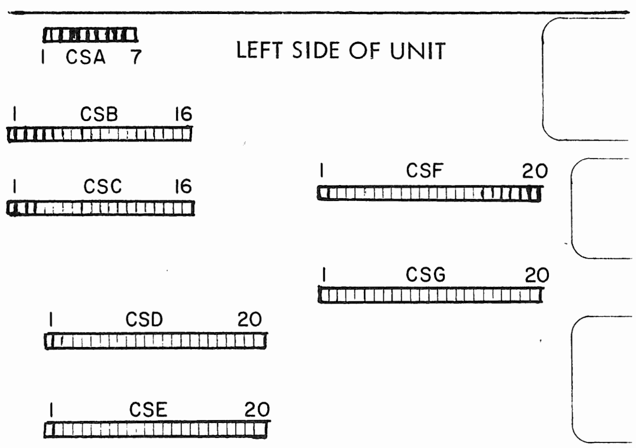


FIG. 1

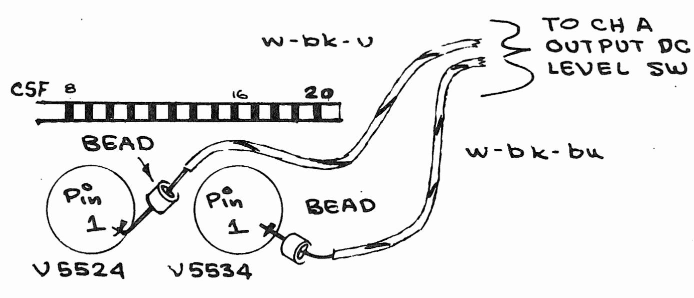


FIG. 2

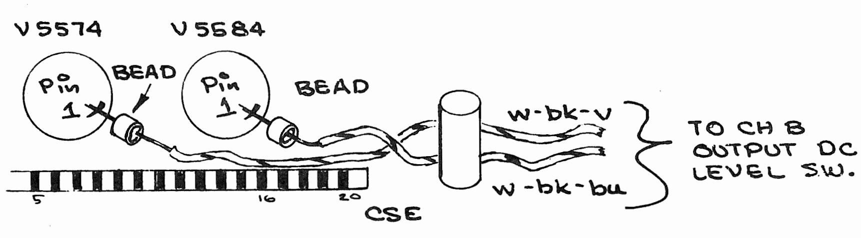


FIG. 3

FRONT PANEL KNOB COLOR CHANGED TO CHARCOAL FOR COMPATIBILITY WITH NEW INSTRUMENTS | INFORMATION ONLY | M9172

Effective Prod s/n 2940

DESCRIPTION:

To standardize indicator and plug-in knob colors -- all knobs, switch buttons, binding posts, etc, on older instruments are changed to the charcoal colored ones used on new instruments.

Parts Removed:			Parts Added:		
Knob assembly, black	(2)	366-0029-00	Knob assembly, gray	(2)	366-0142-00
Knob assembly, black	(2)	366-0087-00	Knob assembly, gray	(2)	366-0230-00
Knob assembly, black	(2)	366-0132-00	Knob assembly, gray	(2)	366-0331-00
Jack, banana	(10)	136-0138-00	Jack, banan	(10)	136-0140-00

OUTPUT AMPLIFIER CHANGES REDUCE OVERSHOOT WHEN USED WITH 540B SERIES OSCILLOSCOPES — See SQB — M9564

Effective Prod s/n 2950 — Usable in field instruments s/n 101-2949

FRONT PANEL SYMPTOM: 3-5% overshoot in 540B instruments (only) which is not correctable by peaking adjustments.

PROBLEM: Present amplifier output impedance of 200 Ω is not low enough.

PRODUCTION CHANGE: The output stage transistors Q6564 and Q6574 are replaced and associated components changed (see Before and After schematics) to lower the output impedance to 100 Ω.

Overshoot is reduced to about 1-1/2% in 540B oscilloscopes after calibration of the 'O' unit in the 544, 546 and 547. Transient Response remains excellent in 540A, 544, 546 and 547 instruments.

Parts Removed:

R6520, R6523	Resistor, WW, 1 Ω 1/2 W 5%		308-0141-00
R6522	Resistor, comp, 220 Ω 1/4 W 10%		316-0221-00
R6563, R6573	Resistor, comp, 3.9 k 1/2 W 10%		302-0392-00
R6569, R6579	Resistor, comp, 9.1 k 1/2 W 5%		301-0912-00
R6568	Resistor, prec, 119 Ω 1/4 W 1%		319-0050-00
R6564, R6574	Resistor, prec, 200 Ω 1/8 W 1%		318-0083-00
R6565, R6575	Resistor, comp, 4.7 Ω 1/2 W 10%		307-0023-00
R6576	Resistor, comp, 3.9 k 2 W 10%		306-0392-00
C6523	Capacitor, disc, 0.1 µF 10 V		283-0023-00
C6565	Capacitor, disc, 0.0022 µF 50 V		283-0028-00
C6579	Capacitor, disc, 0.02 µF 150 V		283-0004-00
D6576	Diode, Zener, 6.3 V		152-0016-00
L6565, L6575	Core, ferramic suppressor		276-0528-00
L6524, L6544	Coil, variable, 0.2 - 0.325 µH		114-0149-00
L6564, L6574	Coil, variable, 0.5 - 1.0 µH		114-0043-00
Q6564, Q6574	Transistor, 2N1143		151-0067-00
	Socket, transistor, 3-pin	(2)	136-0181-00
	Screw, thread-forming, 4 x 1/4	(2)	213-0088-00
	Cable, chassis		179-0712-00

Parts Added:

R6522	Resistor, comp, 330 Ω 1/2 W 5%	301-0331-00
R6568	Resistor, prec, 40.2 Ω 1/8 W 1%	321-0059-00
R6560, R6570	Resistor, comp, 10 k 1 W 5%	303-0103-00
R6567, R6577	Resistor, prec, 100 Ω 1/8 W 1%	321-0097-00
R6562, R6572	Resistor, comp, 220 Ω 1/4 W 5%	315-0221-00
R6578	Resistor, comp, 8.2 k 1/4 W 5%	315-0822-00
R6546	Potentiometer, comp, 20 k	311-0337-00
C6562, C6572	Capacitor, cer, 0.001 µF 100 V	283-0065-00
C6525, C6545	Capacitor, cer, 15 pF 500 V	281-0509-00
C6546	Capacitor, variable, cer, 9-35 pF	281-0063-00
D6576	Diode, Zener, 3 V	152-0076-00
L6524, L6544	Coil, variable, 0.5 - 1.0 µH	114-0043-00
Q6564, Q6574	Transistor, (similar 2N2475)	151-0120-00
	* Socket, transistor, 4-pin	136-0182-00
	* Cable, chassis	179-0712-00

continued

INSTALLATION INSTRUCTIONS:

Parts Required: See Parts Added, except cable and socket marked with asterisks.

Installation Procedure:

Refer to Figs. 1 and 2 for ceramic strip locations while performing the following steps.

a) Unsolder and remove the following components and wires (some components may be re-used):

white-brown wire from top terminal of L6574
white-blue wire from top terminal of L6564
bare wire between the top terminals of L6574 and C6564
bare wire between the top terminals of L6564 and C6574
R6520, 1Ω resistor between CSB-16 and CSC-16
R6523, 1Ω resistor between CSB-14 and CSC-14
R6522, 220Ω resistor between CSB-15 and CSC-15
C6523, 0.1μF capacitor between CSB-14 and CSB-16
bare wire between CSB-15 and CSC-16
R6524, 200Ω resistor between CSB-14 and L6524
R6551, 100k resistor between CSB-13 and L6524
R6556, 100k resistor between CSB-12 and L6524
D6576, Zener diode between CSB-11 and CSC-11
C6576, 0.01μF capacitor between CSB-10 and CSC-10
bare wire between CSC-10 and CSC-11
bare wire between CSB-10 and ground
bare wire between CSB-11 and CSB-15
bare wire between CSC-14 and CSC-16
R6565, 4.7Ω resistor between CSB-9 and CSC-9
R6564, 200Ω resistor between CSB-9 and L6564
R6574, 200Ω resistor between CSB-7 and L6574
R6575, 4.7Ω resistor between CSB-7 and CSC-7
bare wire between CSC-7 and CSC-9
C6565, 0.0022μF capacitor between CSB-7 and CSB-9
R6569, 9.1k resistor between CSB-5 and CSC-5
R6563, 3.9k resistor between CSB-4 and CSC-4
R6573, 3.9k resistor between CSB-2 and CSC-2
R6579, 9.1k resistor between CSB-1 and CSC-1
C6579, 0.02μF capacitor between CSC-1 and CSC-5
bare wire between CSC-2, CSC-3 and CSC-4
bare wire between CSB-1 and CSB-2
bare wire between CSB-4 and CSB-5
bare wire between CSB-5 and Q6564 emitter
bare wire between CSB-2 and Q6574 emitter
R6568, 119Ω resistor between Q 6574 and Q6564 emitters
L6574, coil located on chassis
L6564, coil located on chassis

b) Unsolder the white-red wire from CSC-10 and clip it off where it enters the cable (the other end will be removed later).

c) Move the white-green wire from CSB-16 to CSB-15.

d) Move the two white-violet wires from CSC-15 to CSC-16.

continued

Installation Procedure:

e) Move the white-orange wire from CSC-13 to CSC-14.

f) Move the white-green wire from CSC-12 to CSC-13.

g) Move the white-brown-black-brown wires from CSC-2 and CSC-3 to CSC-4.

h) Trim and solder the white-brown wire (unsoldered from L6564 in step a -- other end to rear connector pin 3), to CSC-6. Do not re-install the toroid core (L6575) if present.

i) Trim and solder the white-blue wire (unsoldered from L6564 in step a -- other end to rear connector pin 1), to CSC-9. Do not re-install the toroid core (L6565) if present.

k) Check Step -- The following wiring should now be on CSB and CSC:
white-green wire at CSB-15
three white-brown-black-brown wires at CSC-4
white-brown wire at CSC-6
white-blue wire at CSC-9
bare wire between CSC-9 and CSC-10
bare wire between CSB-12 and CSC-12
bare wire between CSB-13 and CSC-13
white-green wire at CSC-13
white-orange wire at CSC-14
two white-violet wires at CSC-16

m) Replace L6544 and L6524 with 0.5-1 μH coils

n) Install the following wires and components:
bare wire between CSB-6 and CSB-10 (inside)
bare wire between CSB-10 and CSB-11 (outside)
bare wire between CSB-11 and CSB-16 (inside)
bare wire between CSC-1 and CSC-3 (outside)
bare wire between CSC-3 and CSC-11 (inside)
bare wire between CSC-11 and CSC-12 (outside)
bare wire between CSB-12 and ground on V6524 socket
bare wire between CSC-6 and CSC-7
bare wire between CSB-16 and CSC-15
bare wire between CSB-14 and CSC-14
bare wire between CSA-1 and Q6574 base
bare wire between CSA-4 and Q6564 base
bare wire between CSB-1 and Q6574 emitter
bare wire between CSB-3 and Q6564 emitter
bare wire between CSB-7 and Q6574 collector
bare wire between CSB-9 and Q6564 collector
bare wire between Q6574 collector and C6564 top terminal
bare wire between Q6564 collector and C6574 top terminal
C6545, 15 pF capacitor between pin 5 of V6544 and ground on socket
C6525, 15 pF capacitor between pin 5 of V6524 and ground on socket
R6568, 40.2 Ω resistor between Q6574 emitter and Q6564 emitter
C6546, 9-35 pF capacitor between CSA-1 and CSA-3
R6546, 20 k potentiometer between CSA-3 and CSA-4
R6570, 10 k resistor between CSB-1 and CSC-1
R6560, 10 k resistor between CSB-3 and CSC-3

continued

Installation Procedure:

n) continued

R6577, 100Ω resistor between CSB-6 and CSC-6
R6572/C6572, 220Ω-0.001 μF resistor-capacitor combination between CSB-7 and CSC-7
R6562/C6562, 220Ω-0.001 μF resistor-capacitor combination between CSB-9 and CSC-9
R6578, 8.2k resistor between CSB-7 and CSB-9
R6567, 100Ω resistor between CSB-10 and CSC-10
** C6576, 0.01 μF capacitor between CSB-11 and CSC-11
D6576, 3 V Zener diode, between CSB-15 and CSC-15 (cathode)
R6522, 330Ω resistor between CSB-16 and CSC-16
** R6524, 200Ω resistor between CSB-15 and L6524 top terminal
** R6551, 100k resistor between CSB-14 and L6524 top terminal
** R6556, 100k resistor between CSB-13 and L6524 top terminal

** Removed in step a

p) Unsolder and remove the 3.9k resistor (R6576) located between CSH-5 and CSJ-5.

q) Remove the bare wire between CSH-4 and CSH-5.

r) Remove the bare wire between CSJ-5 and CSK-1.

s) Unsolder the white-red wire from CSK-1 and clip it off where it enters the cable (other end clipped in step b).

t) Remove the "L6564" and "L6574" silkscreening from the chassis with lacquer thinner or similar mineral solvent.

u) Replace Q6564 and Q6574 with the new 151-0120-00 transistors.

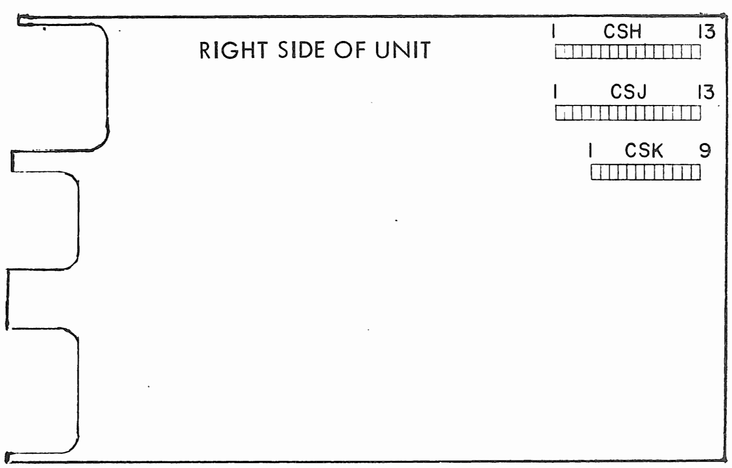


Fig. 1

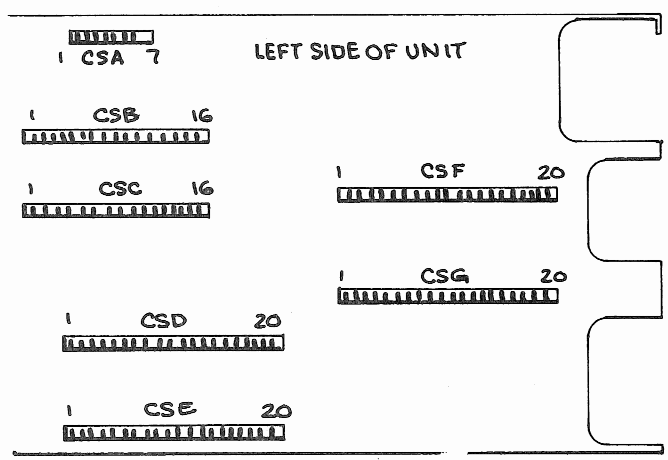
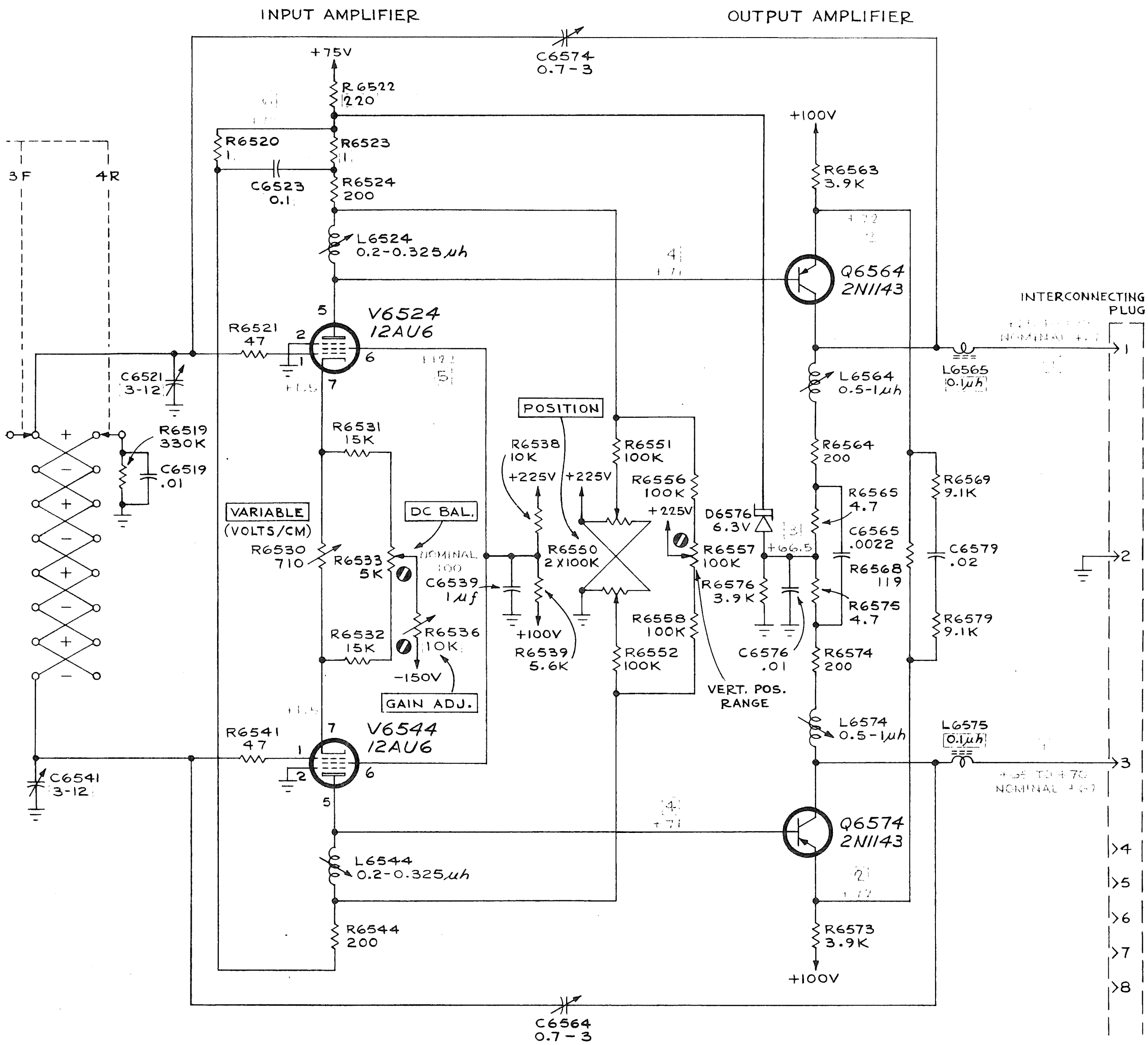


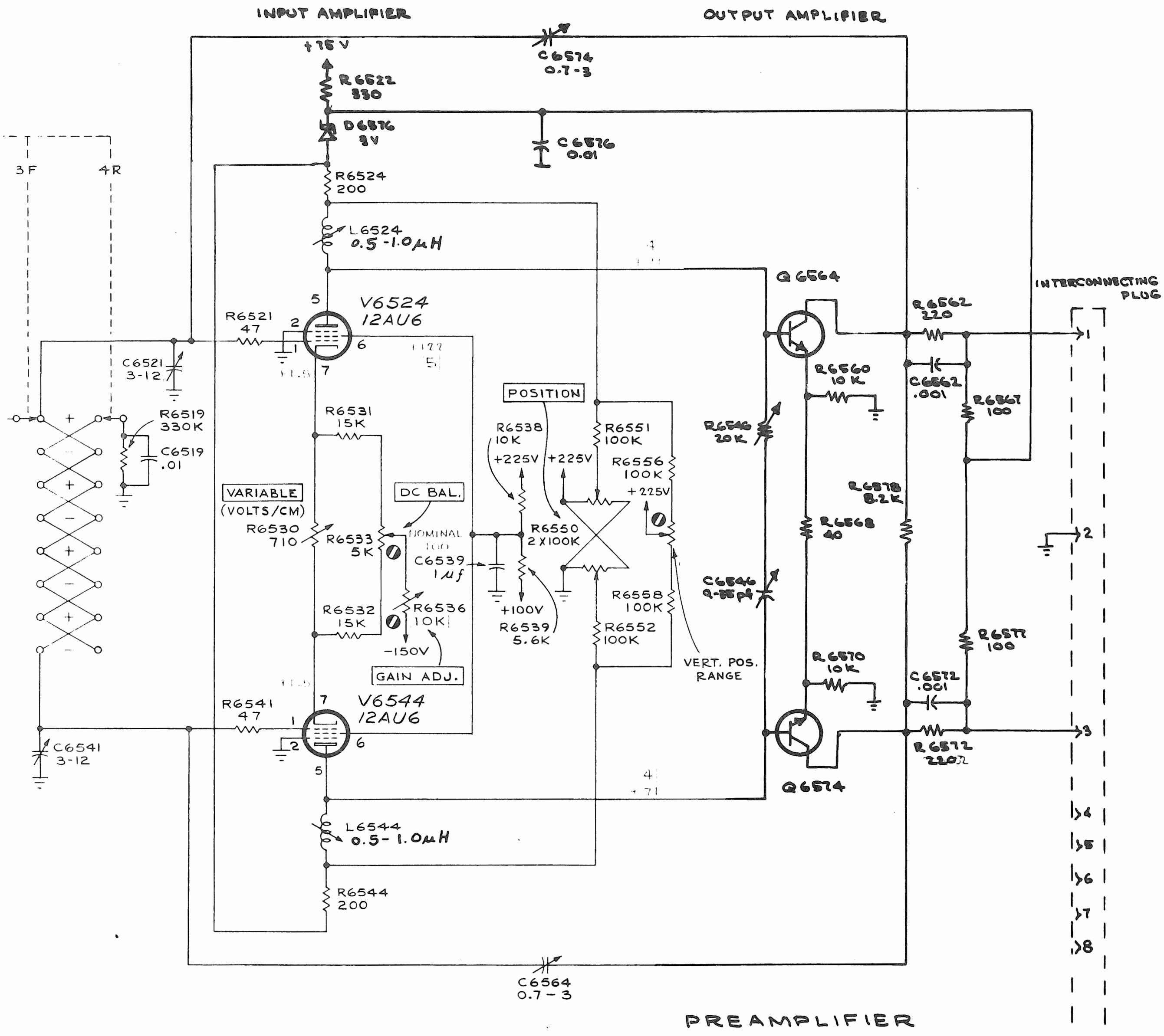
Fig. 2

continued



BEFORE

continued


INPUT AMPLIFIER
OUTPUT AMPLIFIER
+15 V
R6522
330
D6576
C6574
0.7-3
C6576
0.01
3F
4R
R6524
200
L6524
0.5-1.0μH
Q6564
INTERCONNECTING
PLUG
R6521
47
V6524
12AU6
R6562
220
C6521
3-12
C6562
.001
R6560
10K
R6519
330K
C6519
.01
POSITION
R6538
10K
+225V
R6551
100K
R6556
100K
R6546
20K
R6567
100
R6578
8.2K
VARIABLE
(VOLTS/CM)
R6530
710
R6531
15K
DC BAL.
R6533
5K
NOMINAL
C6539
1μf
R6550
2X100K
R6568
40
C6546
9-35pf
R6558
100K
R6532
15K
R6536
10K
-150V
GAIN ADJ.
+100V
R6539
5.6K
R6552
100K
VERT. POS.
RANGE
R6570
10K
R6577
100
C6572
.001
R6541
47
V6544
12AU6
R6572
220Ω
C6541
3-12
Q6574
L6544
0.5-1.0μH
R6544
200
C6564
0.7-3
PREAMPLIFIER

AFTER

EXTERNAL INPUT COUPLING CAPACITOR REPLACED WITH IMPROVED MORE RELIABLE TYPE

INFORMATION ONLY — M9957

Effective Prod s/n 3020

FRONT PANEL SYMPTOM: None.

PROBLEM: Impregnated polycarbonate film is more reliable and has a better dielectric absorption characteristic than mylar film.

PRODUCTION CHANGE: C6501 was changed from the type with mylar dielectric to a type with impregnated polycarbonate film dielectric. The value and voltage rating remained unchanged.

Parts Removed:

C6501	Capacitor, PTM, 0.1 μF 600 V	285-0556-00

Parts Added:

C6501	Capacitor, PTM, 0.1 μF 600 V	285-0672-00

6AU6 & 12AU6 TUBES REPLACED BY PREMIUM 8425 & 8426 TUBES TO REDUCE MICROPHONICS, INTERFACE AND GRID CURRENT PROBLEMS

INFORMATION ONLY — M10548

Effective Prod SN none given

FRONT PANEL SYMPTOM: Microphonics, interface, and grid current.

PROBLEM: Usage of 6AU6 and 12AU6 tubes results in high reject rate and length of selection time because of microphonics, interface, and grid current.

PRODUCTION CHANGE: Type 6AU6 and 12AU6 tubes were replaced by 8425 and 8426 premium tubes.

Parts Removed:

V5524, V5534, V5574, V5584	Tube, raw 12AU6(154-0040-00) Subpart of	157-0071-00
V6524, V6544	Tube, raw, 12AU6(154-0040-00) Subpart of	157-0077-00

Parts Added:

V5524, V5534, V5574, V5584	Tube, raw 8426/12AU6(154-0040-05) Subpart of	157-0071-00
V6524, V6544	Tube, raw 8426/12AU6(154-0040-05) Subpart of	157-0077-00

OPERATIONAL AMPLIFIER
ZZ1000 CURRENT INCREASED FOR MORE STABLE OPEN-LOOP GAIN — See SQB — M10869

Effective Prod SN 3240 — Usable in field instruments SN 813-3239*

* Or SN 101-813 with Mod Kit 040-0301-00 or 040-0301-01 installed

FRONT PANEL SYMPTOM: Erratic closed-loop gain at very high (much larger than 100) gain.

PROBLEM: It was hard to set open-loop gain accurately during calibration due to insufficient current through V5539 and V5589. This was especially a problem at very high (much larger than 100) gain settings.

PROD CHANGE: The current through V5539 and V5589 was increased by changing the values of the biasing resistors. This mod is incorporated into Mod Kit 040-0301-02.

Parts Removed:

R5523, R5573	Resistor, comp,	39k	2W	5%	305-0393-00
R5539, R5589	Resistor, comp,	180k	1W	10%	304-0184-00

Parts Added:

R5523, R5573	Resistor, comp,	27k	2W	5%	305-0273-00
R5539, R5589	Resistor, comp,	82k	2W	5%	305-0823-00

INSTALLATION INSTRUCTIONS:

Parts Required:

See 'Parts Added'.

Installation Procedure:

Replace the resistors as listed above, located on the ceramic strips near the gas tubes (V5539 and V5589).

NEON INDICATING LAMPS AND HOLDERS REPLACED WITH IMPROVED TYPE — INFORMATION ONLY — M8002

Effective Prod SN 3280

FRONT PANEL SYMPTOM: None.

PROBLEM: None.

PRODUCTION CHANGE: The indicating neon holders were replaced with a type which increased wide-angle visibility and is neater in apperance. The new holders, being slightly shorter, require a type NE-2V neon bulb and a shorter mounting screw.

Parts Removed:

B5517, B5567	Bulb, neon, NE-23	150-0027-00
	Holder, neon, single	352-0008-00
	Screw, 4-40 x 1 FHS	211-0031-00

Parts Added:

B5517, B5567	Bulb, neon, NE-2V	150-0030-00
	Holder, neon, single	352-0067-00
	Filter, lens, neon indicator	378-0541-00
	Screw, 4-40 x 7/8 FHS	211-0109-00

FERRITE CORE REPLACED WITH MORE READILY AVAILABLE TYPE AND TO REDUCE COST — INFORMATION ONLY — M10297

Effective Prod SN 3430

FRONT PANEL SYMPTOM: None.

PROBLEM: The 0.7 μH ferrite core is not available in sufficient quantity and is replaceable with a 0.6 μH ferrite core at a cost saving.

PRODUCTION CHANGE: The 0.7 μH ferrite cores, L5520, L5530, L5570 and L5580, were replaced by a 0.6 μH ferrite core.

Parts Removed:

L5520, L5530, L5570, L5580	0.7 μH ferrite core	276-0532-00

Parts Added:

L5520, L5530, L5570, L5580	0.6 μH ferrite core	276-0507-00

<u>10% AND 20% ZENER DIODE</u>
<u>CHANGED TO STANDARD 5%</u> UNIT

INFORMATION ONLY

M11191

Effective Prod SN not given

FRONT PANEL SYMPTOM: None.

PROBLEM: Zener diode values are at present widely scattered in both voltage and tolerance. The proposed modifications will standardize all 400 mW, 1 W, 1.5 W and 10 W Zeners now listed as 10 and 20%, to 5% tolerance; and change the majority of non-standard parts to standard JEDEC units. One of these changes is to minimize the number of active part numbers. There will be no increase in cost for the 5% Zeners.

PRODUCTION CHANGE: Voltage tolerance for 10% and 20% Zener diodes was changed to 5% for all uses. At the same time, all 250 mW Zener diodes were changed to 400 mW. Refer to parts removed and added list for details.

Parts Removed:

D6576	Diode 1N4372 3 V ±10%	152-0076-00

Parts Added:

D6576	Diode 1N4372A 3 V ±5%	152-0278-00

OPERATIONAL AMPLIFIERS
Z_i Z_f RESISTORS CHANGED TO 1/2% METAL FILM TO ELIMINATE SELECTION

See SQB

M11478

Effective Prod SN 3520

Usable in field instruments SN 101-3519

FRONT PANEL SYMPTOM: None.

PROBLEM: Z_i Z_f resistors often must be selected to meet A & B SELECTOR switches 1% gain matching specification.

PRODUCTION CHANGE: The tolerance of the Z_i Z_f resistors was changed from 1% to 1/2%. R5509B, R5509C, R5511B, and R5511C will be changed when the parts become available.

Parts Removed:

R5511A, R5509A	Resistor, prec 1 M 1/2 W 1%	309-0148-00
R5511D, R5509D	Resistor, prec 100 k 1/2 W 1%	309-0260-00
R5511E, R5509E	Resistor, prec 10 k 1/2 W 1%	309-0100-00
R5561A, R5559A	Resistor, prec 1 M 1/2 W 1%	309-0148-00
R5561D, R5559D	Resistor, prec 100 k 1/2 W 1%	309-0260-00
R5561E, R5559E	Resistor, prec 10 k 1/2 W 1%	309-0100-00

Parts Added:

R5511A, R5509A	Resistor, prec MF 1 M 1/2 W 1/2%	323-0481-01
R5511D, R5509D	Resistor, prec MF 100 k 1/2 W 1/2 %	323-0385-01
R5511E, R5509E	Resistor, prec MF 10 k 1/2 W 1/2%	323-0289-01
R5561A, R5559A	Resistor, prec MF 1 M 1/2 W 1/2%	323-0481-01
R5561D, R5559D	Resistor, prec MF 100 k 1/2 W 1/2%	323-0385-01
R5561E, R5559E	Resistor, prec 10 k 1/2 W 1/2%	323-0289-01

INSTALLATION INSTRUCTIONS:

Parts Required: See 'Parts Added'.

Installation Procedure:

a) Replace the following resistors, located on the Channel A Selector switch, as required:

R5511A (1 M)	located between W1-9F and W2-9R.
R5511D (100 k)	located between W1-12F and W2-12R.
R5511E (10 k)	located between W1-1F and W2-1R.
R5509A (1 M)	located between W4-7R and W3-7F.
R5509D (100 k)	located between W4-10R and W3-10F.
R5509E (10 k)	located between W4-11R and W3-11F.

b) Replace the following resistors, located on the Channel B Selector switch as required:

R5561A (1 M)	located between W1-9F and W2-9R.
R5561D (100 k)	located between W1-12F and W2-12R.
R5561E (10 k)	located between W1- 1F and W2-1R.
R5559A (1 M)	located between W3-9F and W4-9R.
R5559D (100 k)	located between W3-12F and W4-12R.
R5559E (10 k)	located between W3-1F and W4-1R.

continued

NOTE: The following method is used to identify the Channel A or B Selector switch terminals:

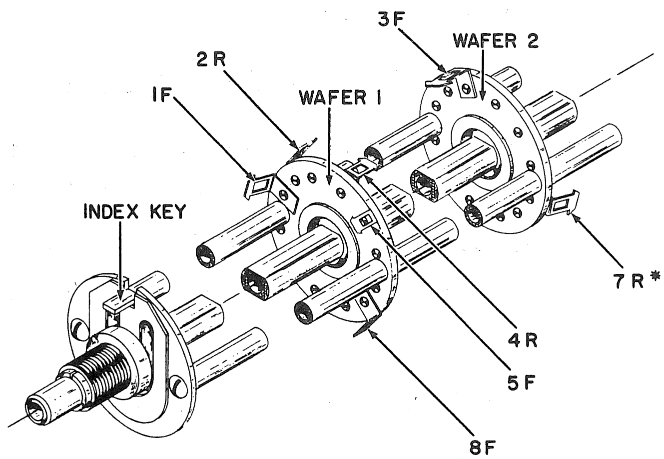
The wafers are numbered from front to the rear.

The contact positions are numbered 1 through 12 relative to the index key as shown in drawing.

The contacts have an "F" or "R" suffix which denotes that they are on the front or the rear of the wafer.

Example: W2-7R (denoted by * on drawing) is contact #7 on the rear of wafer 2.

(TYPICAL SWITCH CONFIGURATION)



Z_i AND Z_f RESISTOR TOLERANCE CHANGED TO REDUCE EFFECT OF TOLERANCE ACCUMULATION

INFORMATION ONLY

M11605

Effective Prod SN 3800

FRONT PANEL SYMPTOM: None.

PROBLEM: Parts tolerance accumulations was causing test to have to select components to meet instrument specifications.

PRODUCTION CHANGE: Tolerance of 1/2W resistors was changed from 1% to 1/2%.

Parts Removed:

R5509B, R5511B, R5559B, R5561B	Resistor, precision, 500k 1/2W 1%	309-0140-00
R5509C, R5511C, R5559C, R5561C	Resistor, precision, 200k 1/2W 1%	309-0444-00

Parts Added:

R5509B, R5511B, R5559B, R5561B	Resistor, precision, 500k 1/2W 1/2%	323-0740-01
R5509C, R5511C, R5559C, R5561C	Resistor, precision, 200k 1/2W 1/2%	323-0414-01

OPERATION AMPLIFIERS
TRANSISTOR TYPE CHANGED
TO ELIMINATE OSCILLATION

INFORMATION ONLY

M11567

Effective Prod SN 3880

FRONT PANEL SYMPTOM: Operational amplifiers oscillate.

PROBLEM: It is possible to operate Q5523 and Q5573 above the BV_{ceo} specifications for the 151-0096-00 type transistors used there.

PRODUCTION CHANGE: The 151-0096-00 transistors were changed to 151-0150-00.

This modification has been incorporated into Field Modification Kit 040-0301-02.

Parts Removed:

Q5523,Q5573	Transistor, silicon, 2N1893	151-0096-00

Parts Added:

Q5523,Q5573	Transistor, silicon, 2N3440	151-0150-00

BE:fb

MODIFICATION KIT

OPERATIONAL AMPLIFIER IMPROVEMENTS

040-301

For Tektronix Type 'O' Plug-in Units
Serial numbers 101-813*

DESCRIPTION

This modification incorporates several refinements in the operational amplifiers of the Type 'O' unit.

Section A improves the cross-talk characteristics by relocating several wires and changing the decoupling arrangement.

Sections B and C improve the temperature-sensitive drift characteristics of the 'A' and 'B' amplifiers.

Section D increases the accuracy of the output amplifier when Z_f is set at 10pf and the INTEGRATOR LF REJECT is OFF.*

The instructions are divided so that any part of the modification may be performed separately if desired.

*Section D applies to s/n 101-318 (with the exception of a few instruments in this range already modified at the factory).

Publication:
Instructions for 040-301
November 1963

Supersedes:
February 1963

PARTS LIST

Quantity	Description				Part Number
2 ea.	Transistor, NPN, TEK Special				151-096
2 ea.	Diode, Zener,	11 v	1/4 w	5%	152-055
4 ea.	Tube, gas diode, ZZ1000				154-370
2 ea.	Capacitor, cer,	0.01 pf	150 v	Hi-Kap	283-003
2 ea.	Capacitor, cer,	0.02 pf	150 v	Hi-Kap	283-004
2 ea.	Resistor, comp,	180 k	1 w	10%	304-184
2 ea.	Resistor, comp,	39 k	2 w	5%	305-393
2 ea.	Resistor, WW,	22 k	7 w	1%	308-241
4 ea.	Resistor, prec,	45 k	1/2 w	1%	309-354
2 ea.	Resistor, comp,	1 k	1/4 w	5%	315-102
2 ea.	Resistor, prec,	40.2 k	1/2 w	1%	323-347
4 ea.	Clamp, neon bulb, no.20 wire, bare				343-043
4 ea.	Tag, MODIFIED INSTRUMENT, gummed back				(001-910)
1 ea.	Tubing, plastic, no.20 black		6 in.		(162-504)
1 ea.	Wire, no.22 solid,		12 in.	black-brown-green-brown	(175-514)
1 ea.	Wire, no.22 solid,		4 in.	black-brown-green-brown	(175-514)
1 ea.	Wire, no.22 solid, twisted pair,		8-1/2 in.	white-yellow/white-brown	(175-522)
1 ea.	Wire, no.22 solid, twisted pair,		6 in.	white-green/white-brown	(175-522)
1 ea.	Wire, no.22 solid,		6-1/4 in.	white-black-violet	(175-522)
1 ea.	Wire, no.22 solid,		2-1/2 in.	white-yellow	(175-522)
1 ea.	Wire, no.22 solid,		12 in.	bare	(176-005)
2 ea.	Wire, no.22 solid, pre-bent for 5 small notches				(176-139)
1 ea.	Wire, solder, silver-bearing		24 in.		

INSTRUCTIONS

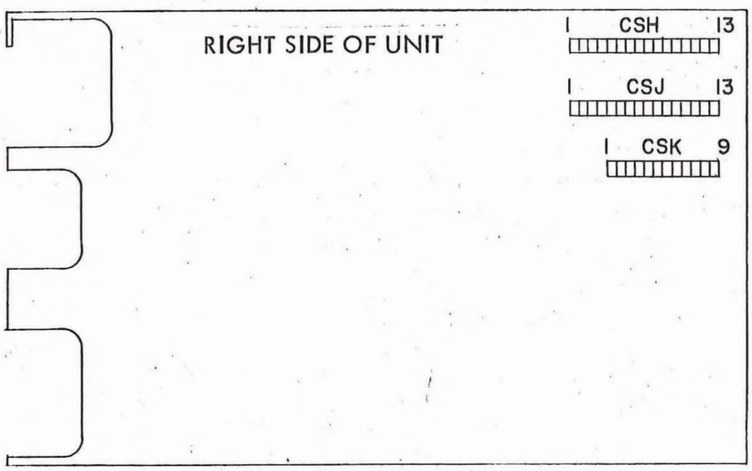


Fig. 1

IMPORTANT: When soldering to the ceramic strips, use the silver-bearing solder supplied with this kit.

A. TO IMPROVE THE CROSS TALK CHARACTERISTICS:

( ) 1. Unsolder the white-green and white-brown wires from CSG-17 and CSG-18.

( ) Cut off these leads at the point where they enter the wiring harness.

( ) 2. Unsolder the white-green and white-brown wires from the amplifier 'A' OUTPUT DC LEVEL ADJ potentiometer.

( ) Solder both of these wires to the ground terminal on the potentiometer.

3. Dress and solder the white-green/white-brown twisted pair (from kit) as follows:

( ) white-green to CSG-17 and lower (cw) terminal of 'A' OUTPUT DC LEVEL ADJ potentiometer.

( ) white-brown to CSG-18 and upper (ccw) terminal of 'A' OUTPUT DC LEVEL ADJ potentiometer.

( ) 4. Unsolder the white-yellow and white-brown wires from CSE-17 and CSE-18.

( ) Cut off these leads at the point where they enter the wiring harness.

( ) 5. Unsolder the white-yellow and white-brown wires from the Amplifier 'B' OUTPUT DC LEVEL ADJ potentiometer.

( ) Solder both of these wires to the ground terminal on the potentiometer.

( ) 6. Dress the white-yellow/white-brown twisted pair (from kit) under the shield and along the wiring harness. Solder as indicated:

( ) white-yellow to CSE-17 and lower (cw) terminal of 'B' OUTPUT DC LEVEL ADJ potentiometer.

Step 6 (con'd)

( ) white-brown to CSE-18 and upper (ccw) terminal of 'B' OUTPUT DC LEVEL ADJ potentiometer.

ALL WIRES REFERRED TO IN STEPS 7 THROUGH 13 ARE COLOR-CODED BLACK-BROWN-GREEN-BROWN

( ) 7. Unsolder all three wires from CSE-20.

( ) 8. Unsolder the wire which goes through the grommet from R5596 (12k, 10 w resistor, mounted on other side of chassis).

( ) With an ohmmeter, locate the other end of this wire (unsoldered in step 15) and clip off both ends, or remove it from the cable.

( ) 9. Determine which of the wires, unsoldered in step 7, goes to CSD-7.

( ) Dress this wire through the grommet and solder to R5596.

( ) 10. Solder the remaining wire (unsoldered in step 7) and one end of the 12 in. wire (from kit) to CSE-20.

( ) Dress the other end of this wire through the grommet, across the chassis, and solder it to CSH-13 (see Fig. 1).

( ) 11. Unsolder both wires from CSG-20.

( ) 12. With an ohmmeter, determine which of these wires goes to CSF-7.

( ) Unsolder the wire from CSF-7 and clip both ends where they enter the cable.

( ) Resolder the remaining wire to CSG-20.

( ) 13. Solder one end of the 4 in. wire (from kit) to CSF-7.

( ) Dress the wire through the hole near R5532 and solder it to R5546 (12k, 10 w) at the terminal nearest the chassis.

REFER TO FIG. 2 FOR CERAMIC STRIP LOCATIONS AND NUMBERING

B. TO IMPROVE THE AMPLIFIER 'A' DRIFT CHARACTERISTICS:

14. Remove the following components and wires. (DO NOT discard any components until the modification is completed):

NOTE: The shield below CSG may be temporarily removed if desired.

( ) 100k (R5540) 1/4 w between CSF-4 and CSG-4

( ) 1N3044B (D5529) between CSF-5 and CSG-5

( ) 0.1 μf (C5528) 200 v between CSF-8 and CSG-8

( ) 45k (R5525) 1% between CSF-9 and CSG-9

( ) 1N3044B (D5528) between CSF-12 and CSG-12

Step 14 (con'd)

( ) 33k (R5523) 1% between CSF-14 and CSG-14

( ) 47k (R5521) 1/2 w between CSF-17 and CSG-17

( ) 47k (R5531) 1/2 w between CSF-18 and CSG-18

( ) 22k (R5526) 2 w between CSF-20 and CSG-20

( ) bare wire between CSF-3 and CSF-4

( ) bare wire between CSF-5 and CSF-8

( ) bare wire between CSF-10 and CSF-12

( ) bare wire between CSF-12 and CSF-14

( ) bare wire between CSG-5 and CSG-7

( ) bare wire between CSF-5 and pin 3 of V5543

( ) 15. Move the white-orange-green-brown wire from CSG-14 to CSG-15.

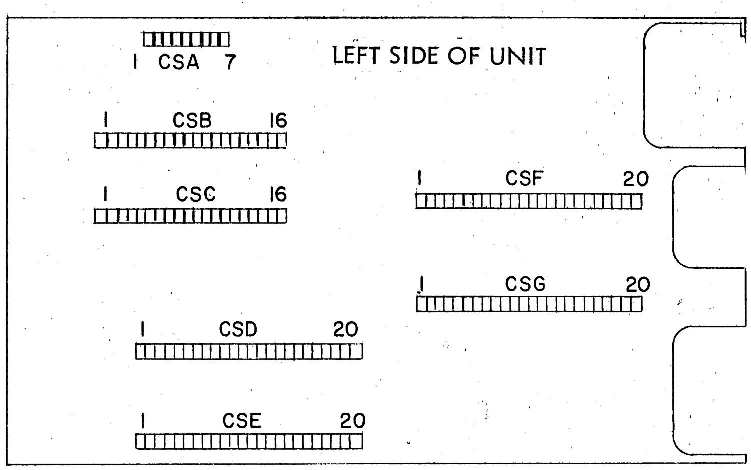


Fig. 2

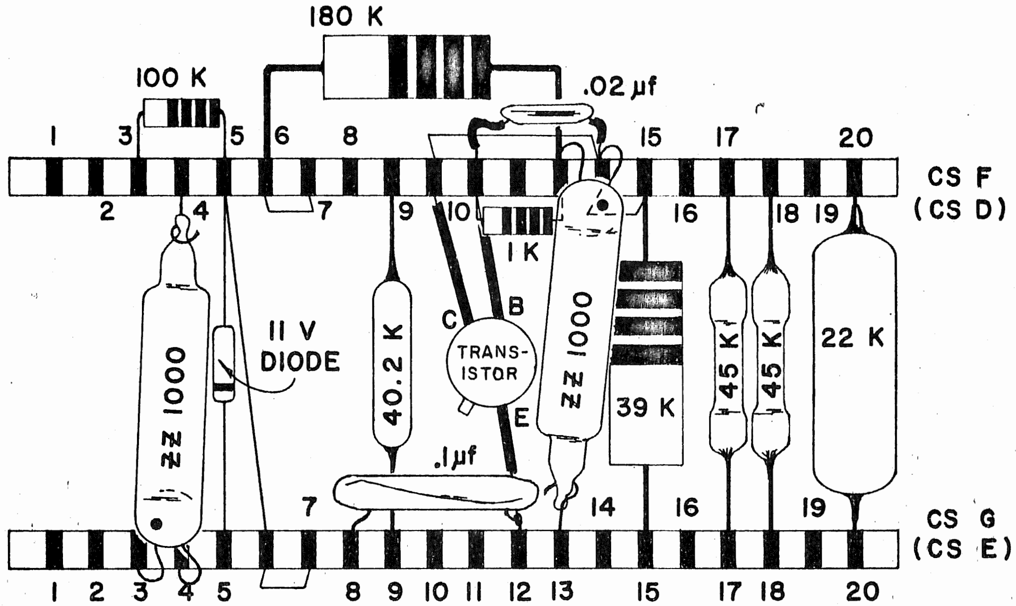


Fig. 3

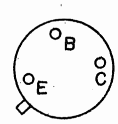


Fig. 4

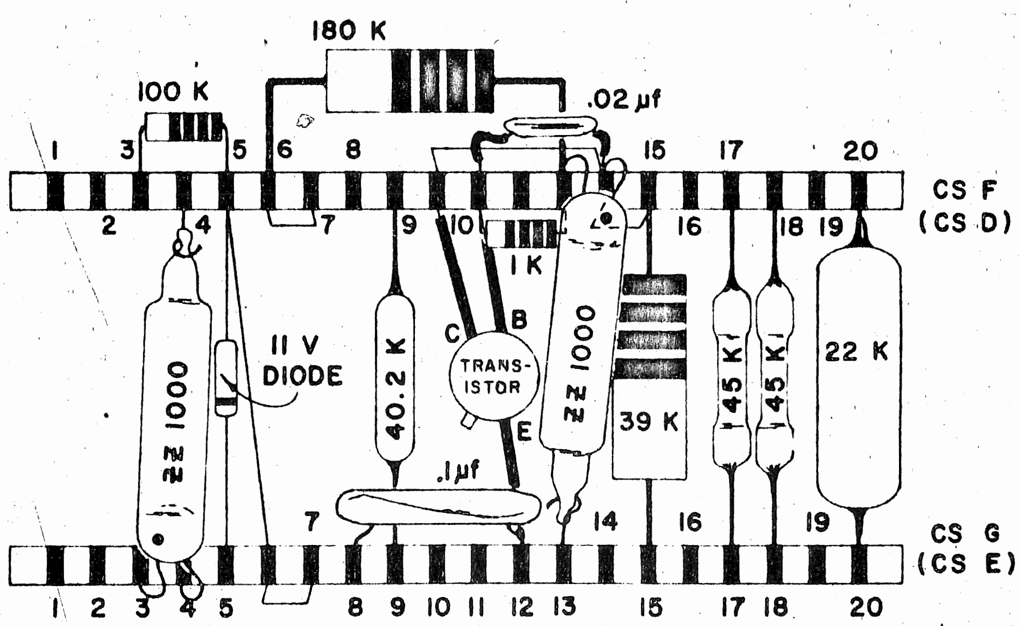


Fig. 5

16. Solder the following wires and components to the points indicated in Fig. 3. (Parts from kit except as noted):

( ) bare wire from CSF-5 to CSG-6

( ) bare wire from CSF-6 to CSF-7

( ) bare wire from CSG-6 to CSG-7

( ) pre-bent wire from CSF-10 to CSF-14

( ) bare wire from CSF-14 to CSF-15

( ) 100k 1/4w (removed in step 14) from CSF-3 to CSF-5

( ) ZZ1000 {dimpled lead to CSG-3; unmarked lead to CSG-4}

( ) neon bulb clamp to CSF-4

( ) 11v Zener {cathode (banded) end to CSG-5; unmarked end to CSF-5}

( ) 0.1 μf 200 v (removed in step 14) from CSG-8 to CSG-12

( ) 180k 1 w from CSF-6 to CSF-13

( ) 40.2k 1% from CSF-9 to CSG-9

( ) 1k 1/4 w from CSF-11 to CSF-13

( ) 0.02 μf 150 v from CSF-11 to CSF-14

NOTE: Place a 1/4 in. length of plastic tubing (from kit) on each capacitor lead.

( ) 39k 2 w from CSF-15 to CSG-15

( ) ZZ1000 {dimpled lead to CSF-14; unmarked lead to CSF-13}

( ) neon bulb clamp to CSG-13

( ) 45k 1% from CSF-17 to CSG-17

( ) 45k 1% from CSF-18 to CSG-18

( ) 22k 7 w from CSF-20 to CSG-20

( ) 17. Place a 1/2 in. length of plastic tubing (from kit) on each lead of the special transistor from the kit.

Step 17 (con'd)
Solder the transistor leads as follows (see Fig. 5):

NOTE: Transistor base diagram is shown in Fig. 4.

( ) emitter to CSG-12

( ) base to CSF-11

( ) collector to CSF-10

( ) Replace the shield, if removed in step 14.

C. TO IMPROVE THE AMPLIFIER 'B' DRIFT CHARACTERISTICS:

18. Remove the following components and wires. (DO NOT discard any components until the modification is completed):

( ) 100k (R5590) 1/4 w between CSD-4 and CSE-4

( ) 1N3044B (D5579) between CSD-5 and CSE-5

( ) 0.1 μf (C5578) 200 v between CSD-8 and CSE-8

( ) 45k (R5575) 1% between CSD-9 and CSE-9

( ) 1N3044B (D5578) between CSD-12 and CSE-12

( ) 33k (R5573) 1% between CSD-14 and CSE-14

( ) 47k (R5571) 1/2 w between CSD-17 and CSE-17

( ) 47k (R5581) 1/2 w between CSD-18 and CSE-18

( ) 22k (R5576) 2 w between CSD-20 and CSE-20

( ) bare wire between CSD-3 and CSD-4

( ) bare wire between CSD-5 and CSD-8

( ) bare wire between CSD-10 and CSD-12

( ) bare wire between CSD-12 and CSD-14

( ) bare wire between CSE-5 and CSE-7

( ) bare wire between CSD-5 and pin 3 of V5593

( ) 19. Move the two white-orange-green-brown wires from CSE-14 to CSE-15.

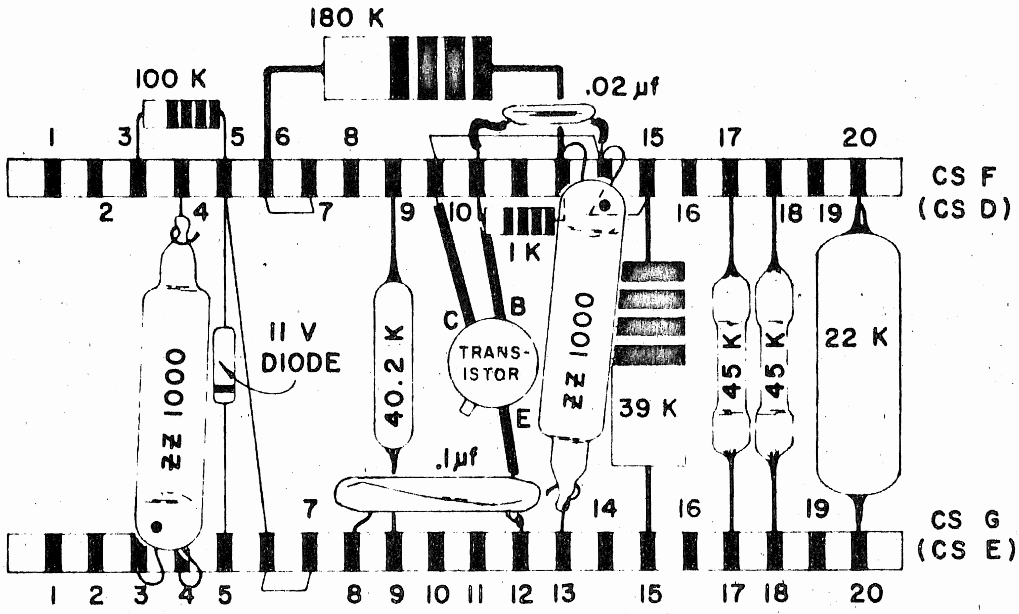

180 K
100 K
.02 µf
1
3
5
6
8
15
17
20
CS F
(CS D)
2
4
7
9
10
16
18
19
1 K
ZZ 1000
11 V
DIODE
40.2 K
C
B
TRANS-
ISTOR
E
ZZ 1000
39 K
45 K
45 K
22 K
.1µf
7
14
16
19
CS G
(CS E)
1
2
3
4
5
8
9
10
11
12
13
15
17
18
20

Fig. 6

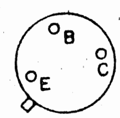

B
C
E

Fig. 7

20. Solder the following wires and components to the points indicated in Fig. 6. (Parts from kit except as noted):

( ) bare wire from CSD-5 to CSE-6

( ) bare wire from CSD-6 to CSD-7

( ) pre-bent wire from CSD-10 to CSD-14

( ) bare wire from CSD-14 to CSD-15

( ) bare wire from CSE-6 to CSE-7

( ) 100k, 1/4 w (removed in step 18) from CSD-3 to CSD-5

( ) ZZ1000 {dimpled lead to CSE-3
unmarked lead to CSE-4

( ) neon bulb clamp to CSD-4

( ) 11 v Zener {cathode (banded) end to CSE-5
unmarked end to CSD-5

( ) 0.1 μf, 200 v (removed in step 18) from CSE-8 to CSE-12

( ) 180k, 1 w from CSD-6 to CSD-13

( ) 40.2k, 1% from CSD-9 to CSE-9

( ) 1k, 1/4 w from CSD-11 to CSD-13

Step 20 (con'd)

( ) 0.02 μf, 150 v from CSD-11 to CSD-14

NOTE: Place a 1/4 in. length of plastic tubing (from kit) on each capacitor lead.

( ) 39k, 2 w from CSD-15 to CSE-15

( ) ZZ1000 {dimpled lead to CSD-14
unmarked lead to CSD-13

( ) neon bulb clamp to CSE-13

( ) 45k, 1% from CSD-17 to CSE-17

( ) 45k, 1% from CSD-18 to CSE-18

( ) 22k, 7 w from CSD-20 to CSE-20

( ) 21. Place a 1/2 in. length of plastic tubing (from kit) on each lead of the special transistor from the kit.

Solder the transistor leads as follows (see Fig. 6):

NOTE: Transistor base diagram is shown in Fig. 7.

( ) emitter to CSE-12
( ) base to CSD-11
( ) collector to CSD-10

(TYPICAL SWITCH CONFIGURATION)

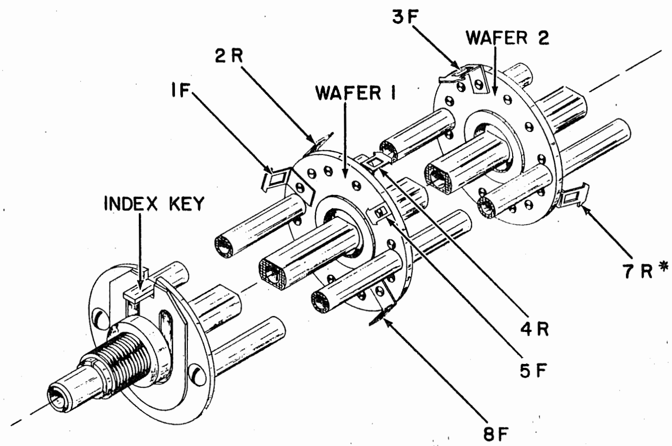


Fig. 8

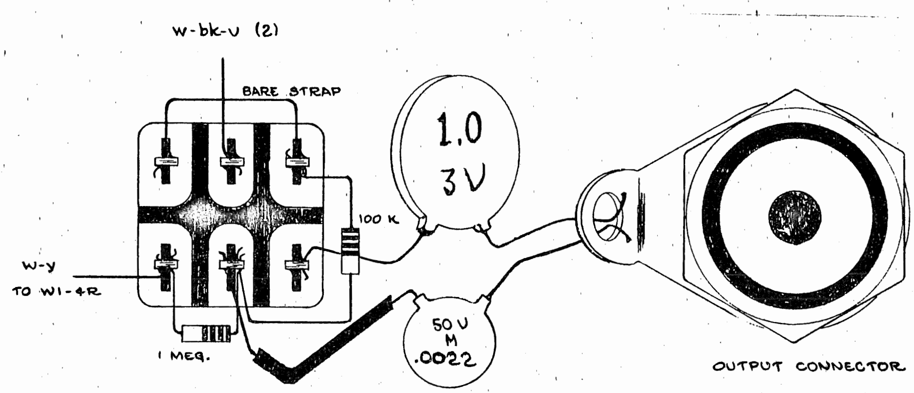


Fig. 9

D. TO INCREASE THE OUTPUT AMPLITUDE ACCURACY:
(Applies only to s/n 101-318, with exceptions. These exceptions may be determined by comparing the wiring of either INTEGRATION LF REJECT switch with Fig. 9. If they are the same, omit steps 22 through 40.)

NOTE: The following method is used to identify the SELECTOR switch terminals:

The wafers are numbered from the front to the rear.

The contact positions are numbered 1 through 12 relative to the index key, as shown in Fig. 8.

The contacts have an "F" or "R" suffix which denotes that they are on the front or the rear of the wafer.

EXAMPLE: W2-7R (denoted by * on Fig. 8) is contact no.7 on the rear of wafer 2.

22. Unsolder the following components and wires from the Amplifier 'A' INTEGRATOR LF REJECT switch:

( ) 100 k, 1/4 w resistor (save for re-use)

( ) white-yellow wire

( ) 1 meg, 1/4 w resistor between switch terminals (save for re-use)

( ) bare wire between switch terminals

( ) 23. Unsolder the end of the 0.0022 μf capacitor from the switch.

( ) Resolder it to the terminal indicated in Fig. 9.

( ) 24. Clean the excess solder from the switch terminals.

( ) 25. Unsolder the white-black-violet wire from 'A' SELECTOR switch W2-8R.

( ) 26. Remove the nylon spacing clip between the white-black-violet and white-yellow wires.

( ) 27. Locate the 100k, 1/4 w resistor on the 'A' SELECTOR switch, soldered between W1-5R and W2-8R.

( ) Unsolder the end from W2-8R.

( ) Resolder it to W1-8F.

( ) 28. Dress the white-black-violet wire, unsoldered in step 25, to the 'A' LF REJECT switch.

( ) Cut off the excess wire and solder, along with the white-black-violet wire (from kit), to the switch terminal indicated in Fig. 9.

( ) 29. Resolder the remainder of the components and wires, as indicated in Fig. 9, using the bare wire from the kit.

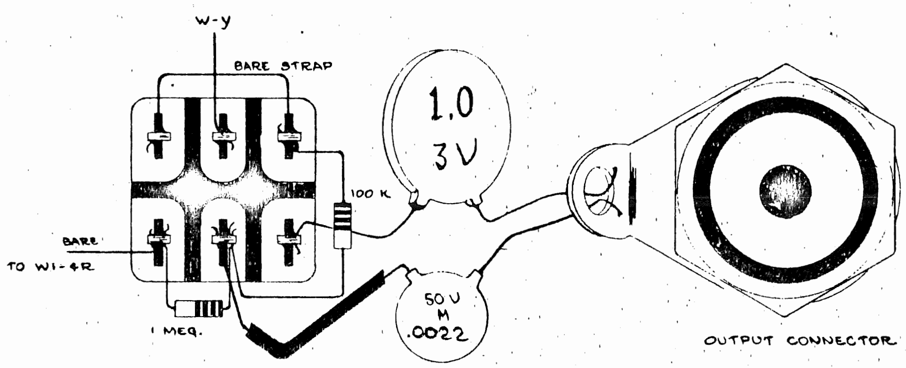

W-Y
BARE STRAP
1.0
3V
100 K
BARE
TO W1-4R
1 MEG.
50 V
M
.0022
OUTPUT CONNECTOR

Fig. 10

( ) 30. Dress the other end of the white-black-violet wire (step 28) through the grommet in the 'A' SELECTOR switch bracket.

( ) Solder this wire to the 'A' SELECTOR switch W2-8R.

( ) 31. Locate the 100k, 1/4w resistor on the 'B' SELECTOR switch, soldered between W1-5R and W2-8R.

( ) Unsolder the end from W2-8R.

( ) Resolder it to W1-8F.

32. Unsolder the following components and wires from the 'B' INTEGRATOR LF REJECT switch:

( ) 100k, 1/4w resistor (save for re-use)

( ) 1 meg, 1/4w resistor between switch terminals (save for re-use)

( ) bare wire between switch terminals

( ) 33. Unsolder and remove the bare wire between the 'B' INTEGRATOR LF REJECT switch and the 'B' SELECTOR switch W1-4R.

( ) 34. Unsolder the end of the 0.0022 μf capacitor from the 'B' INTEGRATOR LF REJECT switch.

( ) Resolder it to the terminal indicated in Fig. 10.

( ) 35. Clean the excess solder from the switch terminals.

( ) 36. Solder the 2-1/2 in. length of white-yellow wire (from kit) between the 'B' SELECTOR switch W2-8R, and the switch terminal indicated in Fig. 10.

( ) 37. Solder a 1-1/2 in. length of bare wire (from kit) to 'B' SELECTOR switch W1-4R.

( ) Place a 1 in. length of plastic tubing (from kit) over the wire and dress the wire to the switch terminal indicated in Fig. 10. DO NOT SOLDER until the next step.

( ) 38. Resolder the components and wires to the switch as indicated in Fig. 10, using the bare wire from the kit.

( ) 39. Solder a 0.01 μf ceramic capacitor (from kit) between the two wired terminals of the Grid Current 'A' potentiometer (R5535) on the potentiometer bracket.

( ) 40. Similarly, solder a 0.01 μf capacitor (from kit) between the two wired terminals of the Grid Current 'B' potentiometer (R5585).

THIS COMPLETES THE INSTALLATION

( ) Check wiring for accuracy.

( ) Adjust the Variable Z_i and Variable Z_f capacitors as described in your Instruction Manual.

( ) Fasten the insert page in your Instruction Manual.

( ) Moisten the back of the MODIFIED INSTRUMENT tags (from kit) and place them on the Manual schematic pages affected by this modification.

GG:cc

MODIFICATION KIT

OPERATIONAL AMPLIFIER IMPROVEMENTS

For Tektronix Type O Plug-in Units
Serial numbers 155-813*

DESCRIPTION

This modification incorporates several refinements in the operational amplifiers of the Type O unit.

Section A improves the crosstalk characteristics by relocating several wires and changing the decoupling arrangement.

Sections B and C improve the temperature-sensitive drift characteristics of the 'A' and 'B' amplifiers.

Section D increases the accuracy of the output amplifier when Z_f is set at 10pF and the INTEGRATOR LF REJECT is OFF.*

The instructions are divided so that any part of the modification may be performed separately if desired.

This kit replaces 040-0301-01.

*Section D applies to SN 155-318 (with the exception of a few instruments in this range already modified at the factory).

040-0301-02

Publication:
Instructions for 040-0301-02
October 1966

Supersedes:
August 1966

PARTS LIST

Quantity	Part Number	Description			
2 ea	151-0096-00	Transistor, NPN, Tek Special			
2 ea	152-0055-00	Diode, Zener	11V	1/4W	5%
4 ea	154-0370-00	Tube, gas diode, ZZ1000			
1 ea	214-0210-00	Spool, w/3 ft. silver-bearing solder			
4 ea	276-0507-00	Core, ferramic suppressor 255, 0.6 μH			
2 ea	283-0003-00	Capacitor, cer,	0.01 pF	150V	Hi-Kap
2 ea	283-0004-00	Capacitor, cer,	0.02 pF	150V	Hi-Kap
2 ea	305-0823-00	Resistor, comp,	82 k	2W	5%
2 ea	305-0273-00	Resistor, comp,	27 k	2W	5%
2 ea	308-0241-00	Resistor, WW,	22 k	7W	1%
4 ea	309-0354-00	Resistor, prec,	45 k	1/2W	1%
2 ea	315-0102-00	Resistor, comp,	1 k	1/4W	5%
2 ea	323-0347-00	Resistor, prec,	40.2 k	1/2W	1%
4 ea	343-0043-00	Clamp, neon bulb, #20 wire, bare			
1 ea	(162-0504-00)	Tubing, plastic, #20		6 in.	black
1 ea	(175-0514-00)	Wire, #22 solid,		12 in.	black-brown-green-brown
1 ea	(175-0514-00)	Wire, #22 solid,		4 in.	black-brown-green-brown
1 ea	(175-0522-00)	Wire, #22 solid, twisted pair		8-1/2 in.	white-yellow/white-brown
1 ea	(175-0522-00)	Wire, #22 solid, twisted pair		6 in.	white-green/white-brown
1 ea	(175-0522-00)	Wire, #22 solid,		6-1/4 in.	white-black-violet
1 ea	(175-0522-00)	Wire, #22 solid,		2-1/2 in.	white-yellow
1 ea	(175-0522-00)	Wire, #22 solid,		2-1/2 in.	white-black-blue
1 ea	(175-0522-00)	Wire, #22 solid,		3 in.	white-black-violet
1 ea	(175-0522-00)	Wire, #22 solid,		6 in.	white-black-blue
1 ea	(175-0522-00)	Wire, #22 solid,		7 in.	white-black-violet
1 ea	(176-0005-00)	Wire, #22 solid,		12 in.	bare
2 ea	(176-0139-00)	Wire, #22 solid, pre-bent for 5 small notches			
4 ea	1-910D	Tag, MODIFIED INSTRUMENT, gummed back.			

INSTRUCTIONS

IMPORTANT: When soldering to the ceramic strips, use the silver-bearing solder supplied with this kit.

A. TO IMPROVE THE CROSSTALK CHARACTERISTICS: (Refer to Fig. 2 for ceramic strip locations except as noted.

( ) 1. Unsolder the white-green and white-brown wires from CSG-17 and CSG-18.

( ) Cut off these leads at the point where they enter the wiring harness.

( ) 2. Unsolder the white-green and white-brown wires from the amplifier 'A' OUTPUT DC LEVEL ADJ potentiometer.

( ) Solder both of these wires to the ground terminal on the potentiometer.

3. Dress and solder the white-green/white-brown twisted pair (from kit) as follows:

( ) white-green to CSG-17 and lower (cw) terminal of 'A' OUTPUT DC LEVEL ADJ pot.

( ) white-brown to CSG-18 and upper (ccw) terminal of 'A' OUTPUT DC LEVEL ADJ pot.

( ) 4. Unsolder the white-yellow and white-brown wires from CSE-17 and CSE-18.

( ) Cut off these leads at the point where they enter the wiring harness.

( ) 5. Unsolder the white-yellow and white-brown wires from the amplifier 'B' OUTPUT DC LEVEL ADJ potentiometer.

( ) Solder both of these wires to the ground terminal on the potentiometer.

( ) 6. Dress the white-yellow/white-brown twisted pair (from kit) under the shield and along the wiring harness. Solder as indicated:

( ) white-yellow to CSE-17 and lower (cw) terminal of 'B' OUTPUT DC LEVEL ADJ pot.

( ) white-brown to CSE-18 and upper (ccw) terminal of 'B' OUTPUT DC LEVEL ADJ pot.

ALL WIRES REFERRED TO IN STEPS A-7 THROUGH A-13 ARE COLOR-CODED BLACK-BROWN-GREEN-BROWN

( ) 7. Unsolder all three wires from CSE-20.

( ) 8. Unsolder the wire which goes through the grommet from R5596 (12k 10W resistor, mounted on other side of chassis).

( ) With an ohmmeter, locate the other end of this wire (unsoldered in step A-7) and clip off both ends, or remove it from the cable.

( ) 9. Determine which of the wires, unsoldered in step A-7, goes to CSD-7.

( ) Dress this wire through the grommet and solder to R5596.

( ) 10. Solder remaining wire (unsoldered in step A-7) and one end of the 12 in. wire (from kit) to CSE-20.

( ) Dress other end of this wire through the grommet, across the chassis, and solder it to CSH-13 (see Fig. 1).

INSTRUCTIONS (cont)

A. (cont)

( ) 11. Unsolder both wires from CSG-20.

( ) 12. With an ohmmeter, determine which of these wires goes to CSF-7.

( ) Unsolder the wire from CSF-7 and clip both ends where they enter the cable.

( ) Resolder the remaining wire to CSG-20.

( ) 13. Solder one end of the 4 in. wire (from kit) to CSF-7.

( ) Dress the wire through the hole near R5532 and solder it to R5546 (12k 10W) at the terminal nearest the chassis.

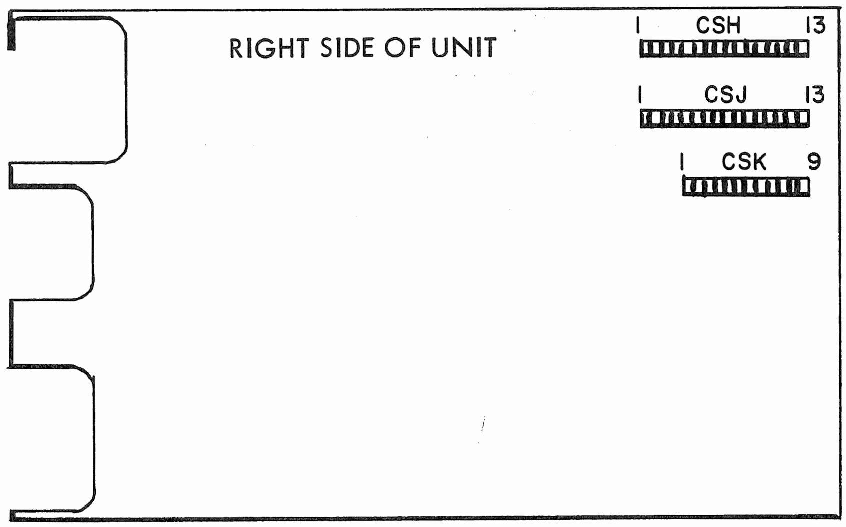


Fig. 1

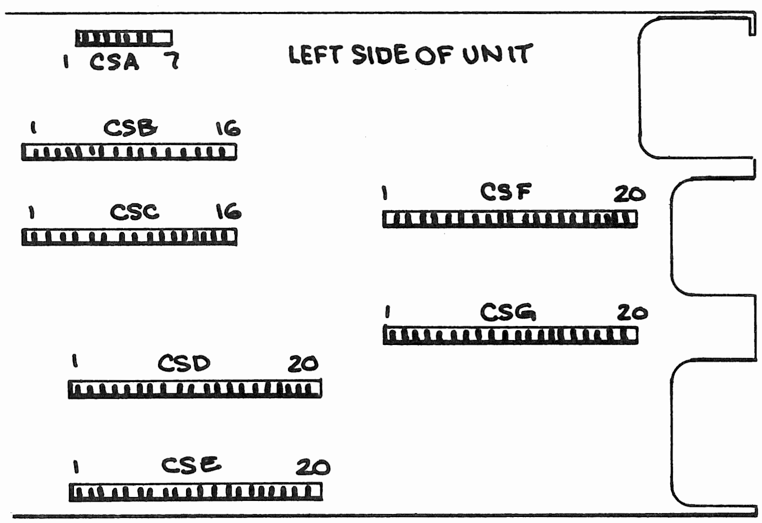


Fig. 2

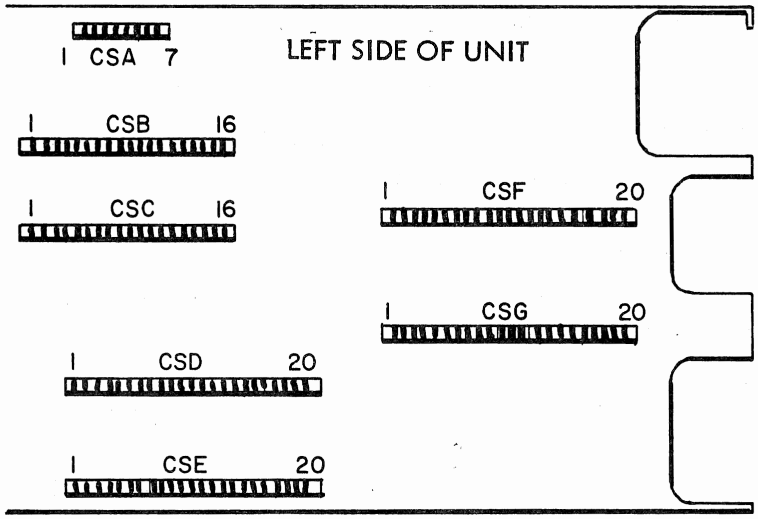

1 CSA 7
LEFT SIDE OF UNIT
1 CSB 16
1 CSC 16
1 CSF 20
1 CSG 20
1 CSD 20
1 CSE 20

Fig. 2

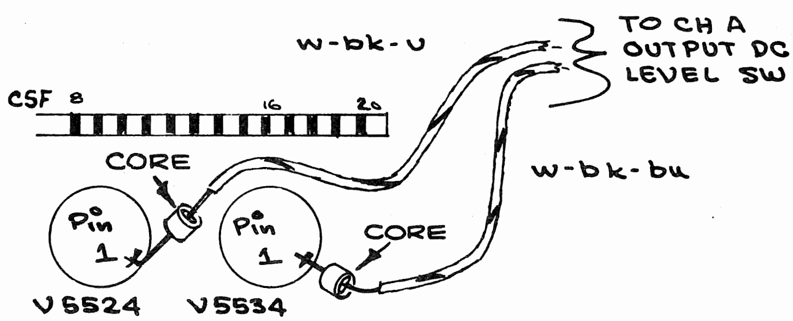

w-bk-v
TO CH A
OUTPUT DC
LEVEL SW
CSF 8
16
20
CORE
w-bk-bu
Pin
1
CORE
V5524
V5534

Fig. 3

INSTRUCTIONS (cont)

B. TO IMPROVE THE AMPLIFIER 'A' DRIFT CHARACTERISTICS: (Refer to Fig. 2 for ceramic strip locations.)

1. Remove the following components and wires. (DO NOT discard any components until the modification is completed):

NOTE: The shield below CSG may be temporarily removed if desired.

( ) 100 k 1/4W (R5540) between CSF-4 and CSG-4

( ) 1N3044B (D5529) between CSF-5 and CSG-5

( ) 0.1 μF 200V (C5528) between CSF-8 and CSG-8

( ) 45 k 1% (R5525) between CSF-9 and CSG-9

( ) 1N3044B (D5528) between CSF-12 and CSG-12

( ) 33 k 1% (R5523) between CSF-14 and CSG-14

( ) 47 k 1/2W (R5521) between CSF-17 and CSG-17

( ) 47 k 1/2W (R5531) between CSF-18 and CSG-18

( ) 22 k 2W (R5526) between CSF-20 and CSG-20

( ) bare wire between CSF-3 and CSF-4

( ) bare wire between CSF-5 and CSF-8

( ) bare wire between CSF-10 and CSF-12

( ) bare wire between CSF-12 and CSF-14

( ) bare wire between CSG-5 and CSG-7

( ) bare wire between CSF-5 and pin 3 of V5543

( ) 47 Ω (R5520) between CSF-16 and pin 1 of V5524

( ) 47 Ω (R5530) between CSF-19 and pin 1 of V5534

( ) 2. Move the white-orange-green-brown wire from CSG-14 to CSG-15.

Refer to Fig. 3 when replacing the following wires:

( ) 3. Remove the white-black-violet wire between Channel A OUTPUT DC LEVEL switch and CSF-16.

( ) Solder a 3 in. white-black-violet wire (from kit) between the same terminal of the Channel A OUTPUT DC LEVEL switch and pin 1 of V5524. Place a ferrite core (from kit) over the tube socket end of the wire.

( ) 4. Remove the white-black-blue wire between Channel A OUTPUT DC LEVEL switch and CSF-19

( ) Solder a 2-1/2 in. white-black-blue wire (from kit) between the same terminal of the Channel A OUTPUT DC LEVEL switch and pin 1 of V5534. Place a ferrite core (from kit) over the tube socket end of the wire.

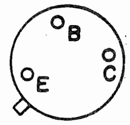

B
C
E

Fig. 4

##

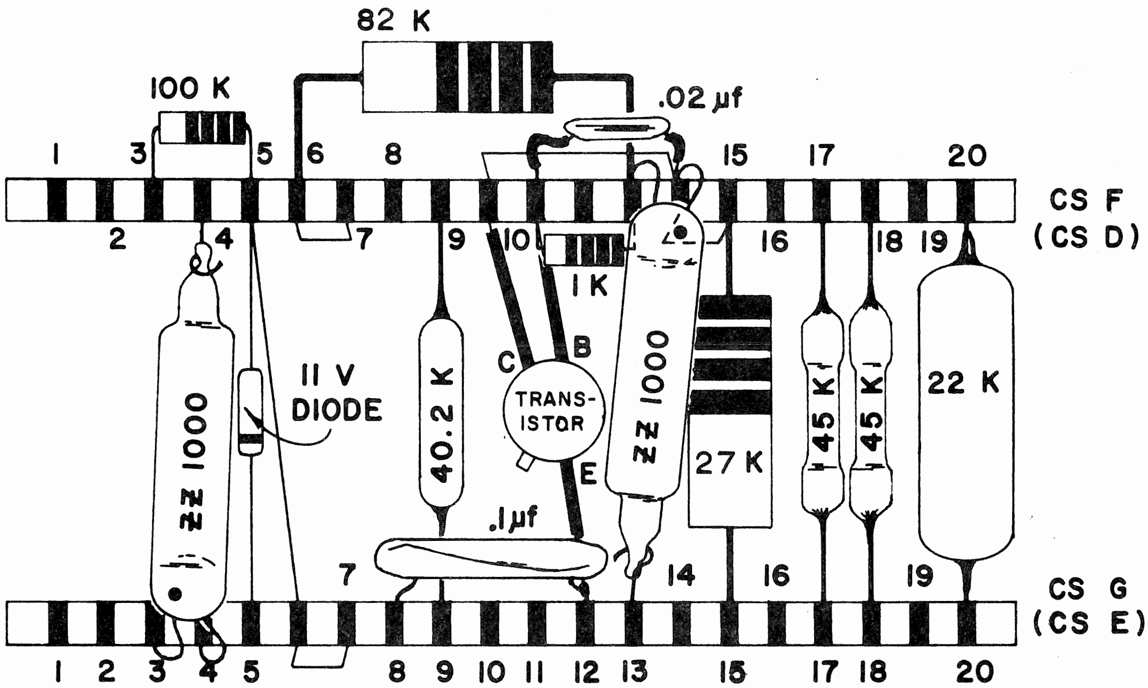

82 K
100 K
.02 µf
1 K
11 V
DIODE
ZZ 1000
40.2 K
C
B
TRANS-
ISTOR
E
.1µf
27 K
45 K
45 K
22 K
CS F
(CS D)
CS G
(CS E)
1
2
3
4
5
6
7
8
9
10
11
12
13
14
15
16
17
18
19
20

Fig. 5

B. 5. Solder the following wires and components to the points indicated in Fig. 5. (Parts from kit except as noted.

( ) bare wire from CSF-5 to CSG-6

( ) bare wire from CSF-6 to CSF-7

( ) bare wire from CSG-6 to CSG-7

( ) pre-bent wire from CSF-10 to CSF-14

( ) bare wire from CSF-14 to CSF-15

( ) 100 k 1/4W (removed in step B-1) from CSF-3 to CSF-5

( ) ZZ1000, dimpled lead to CSG-3, unmarked lead to CSG-4

( ) neon bulb clamp to CSF-4

( ) 11V Zener, cathode (banded) end to CSG-5, unmarked end to CSF-5

( ) 0.1 μF 200V (removed in step B-1) from CSG-8 to CSG-12

( ) 82 k 2W from CSF-6 to CSF-13

( ) 40.2 k 1% from CSF-9 to CSG-9

( ) 1 k 1/4W from CSF-11 to CSF-13

( ) 0.02 μF 150V from CSF-11 to CSF-14

NOTE: Place a 1/4 in. length of plastic tubing (from kit) on each capacitor lead.

( ) 27 k 2W from CSF-15 to CSG-15

( ) ZZ1000, dimpled lead to CSF-14,. unmarked lead to CSF-13

( ) neon bulb clamp to CSG-13

( ) 45 k 1% from CSF-17 to CSG-17

( ) 45 k 1% from CSF-18 to CSG-18

( ) 22 k 7W from CSF-20 to CSG-20

( ) 6. Place a 1/2 in. length of plastic tubing (from kit) on each lead of the special transistor from the kit. Solder transistor leads as follows (see Fig. 5). NOTE: Transistor base diagram is shown in Fig. 4.

( ) emitter to CSG-12

( ) base to CSF-11

( ) collector to CSF-10

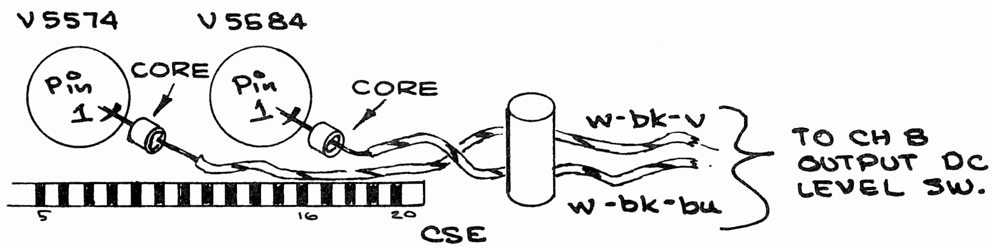

V5574
V5584
Pin
1
CORE
Pin
1
CORE
w-bk-v
w-bk-bu
TO CH B
OUTPUT DC
LEVEL SW.
5
16
20
CSE

Fig. 6

C. TO IMPROVE THE AMPLIFIER 'B' DRIFT CHARACTERISTICS:

1. Remove the following components and wires. (DO NOT discard any components until the modification is completed):

( ) 100 k 1/4W (R5590) between CSD-4 and CSE-4

( ) 1N3044B (D5579) between CSD-5 and CSE-5

( ) 0.1 μF 200V (C5578) between CSD-8 and CSE-8

( ) 45 k 1% (R5575) between CSD-9 and CSE-9

( ) 1N3044B (D5578) between CSD-12 and CSE-12

( ) 33 k 1% (R5573) between CSD-14 and CSE-14

( ) 47 k 1/2W (R5571) between CSD-17 and CSE-17

( ) 47 k 1/2W (R5581) between CSD-18 and CSE-18

( ) 22 k 2W (R5576) between CSD-20 and CSE-20

( ) bare wire between CSD-3 and CSD-4

( ) bare wire between CSD-5 and CSD-8

( ) bare wire between CSD-10 and CSD-12

( ) bare wire between CSD-12 and CSD-14

( ) bare wire between CSE-5 and CSE-7

( ) bare wire between CSD-5 and pin 3 of V5593

( ) 47 Ω (R5570) between CSE-16 and pin 1 of V5574

( ) 47 Ω (R5580) between CSE-19 and pin 1 of V5584

( ) 2. Move the two white-orange-green-brown wires from CSE-14 to CSE-15.

REFER TO FIG. 6 WHEN REPLACING THE FOLLOWING WIRES

( ) 3. Remove the white-black-violet wire between Channel B OUTPUT DC LEVEL switch and CSE-16.

( ) Solder a 7 in. white-black-violet wire (from kit) between the same terminal of the Channel B OUTPUT DC LEVEL switch and pin 1 of V5574. Place a ferrite core (from kit) over the tube socket end of the wire.

( ) 4. Remove the white-black-blue wire between Channel B OUTPUT DC LEVEL switch and CSE-19.

( ) Solder a 6 in. white-black-blue wire (from kit) between the same terminal of the Channel B OUTPUT DC LEVEL switch and pin 1 of V5584. Place a ferrite core (from kit) over the tube socket end of the wire.

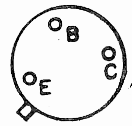

B
C
E

Fig. 7

##

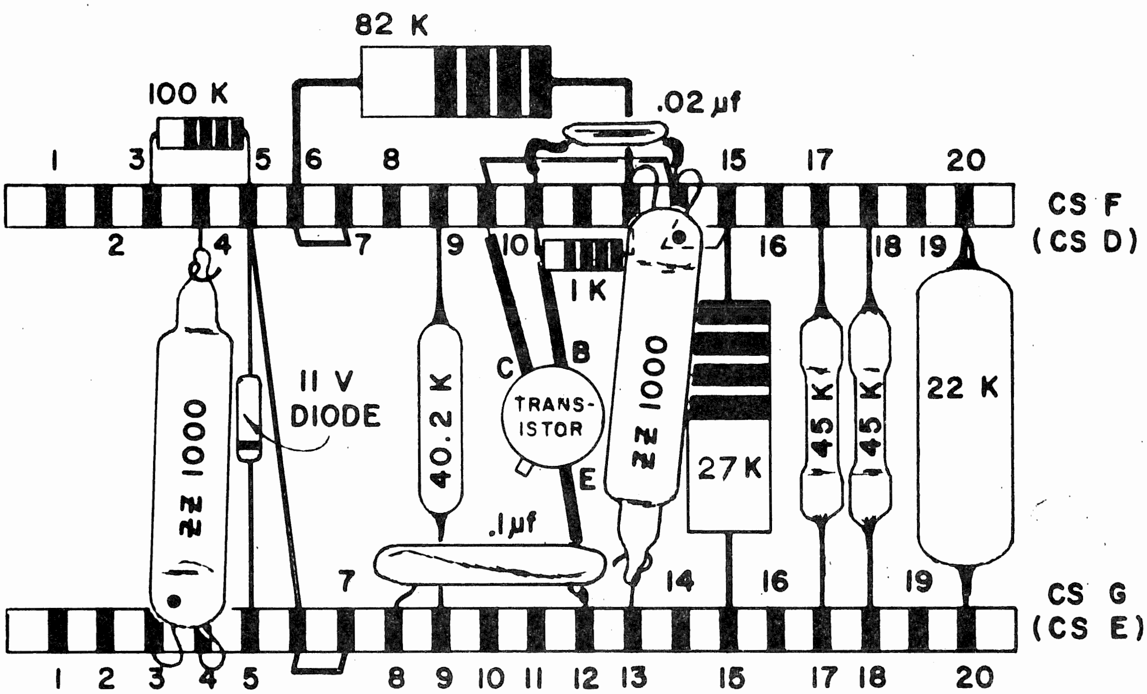

82 K
100 K
.02 µf
1 3 5 6 8 15 17 20
CS F
(CS D)
2 4 7 9 10 16 18 19
1 K
ZZ 1000
11 V
DIODE
40.2 K
C
B
TRANS-
ISTOR
E
ZZ 1000
27 K
45 K
45 K
22 K
.1 µf
7 14 16 19
CS G
(CS E)
1 2 3 4 5 8 9 10 11 12 13 15 17 18 20

Fig. 8

C. 5. Solder the following wires and components to the points indicated in Fig. 8. (Parts from kit except as noted):

( ) bare wire from CSD-5 to CSE-6

( ) bare wire from CSD-6 to CSD-7

( ) pre-bent wire from CSD-10 to CSD-14

( ) bare wire from CSD-14 to CSD-15

( ) bare wire from CSE-6 to CSE-7

( ) 100k 1/4W (removed in step C-1) from CSD-3 to CSD-5

( ) ZZ1000, dimpled lead to CSE-3, unmarked lead to CSE-4.

( ) neon bulb clamp to CSD-4

( ) 11V Zener, cathode (banded) end to CSE-5, unmarked end to CSD-5

( ) 0.1 μF 200V (removed in step C-1) from CSE-8 to CSE-12

( ) 82k 2W from CSD-6 to CSD-13

( ) 40.2k 1% from CSD-9 to CSE-9

( ) 1k 1/4W from CSD-11 to CSD-13

( ) 0.02 μF 150V from CSD-11 to CSD-14

NOTE: Place a 1/4 in. length of plastic tubing (from kit) on each capacitor lead.

( ) 27k 2W from CSD-15 to CSE-15

( ) ZZ1000, dimpled lead to CSD-14, unmarked lead to CSD-13

( ) neon bulb clamp to CSE-13

( ) 45k 1% from CSD-17 to CSE-17

( ) 45k 1% from CSD-18 to CSE-18

( ) 22k 7W from CSD-20 to CSE-20

( ) 6. Place a 1/2 in. length of plastic tubing (from kit) on each lead of the special transistor from the kit.

Solder transistor leads as follows (see Fig. 8). NOTE: Transistor base diagram is shown in Fig. 7.

( ) emitter to CSE-12

( ) base to CSD-11

( ) collector to CSD-10

( ) 7. Replace the shield below CSG, if removed in step B-1.

D. TO INCREASE THE OUTPUT AMPLITUDE ACCURACY:

(Applies only to SN 155-318, with exceptions. These exceptions may be determined by comparing the wiring of either INTEGRATOR LF REJECT switch with Fig. 10. If they are the same, omit steps D-1 through D-19.)

NOTE: The following method is used to identify the SELECTOR switch terminals:

The wafers are numbered from front to rear.

The contact mounting holes are numbered 1 through 12 relative to the index key, as shown in Fig. 9.

The contacts have an 'F' or 'R' suffix which denotes that they are on the front or rear of the wafer.

Example: W2-7R (denoted by * on Fig. 9) is contact #7 on the rear of wafer 2.

(TYPICAL SWITCH CONFIGURATION)

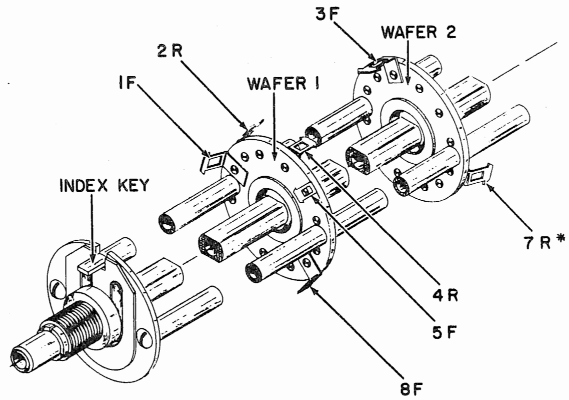


Fig. 9

1. Unsolder the following components and wires from the Amplifier 'A' INTEGRATOR LF REJECT switch:

( ) 100k 1/4W resistor (save for re-use)

( ) white-yellow wire

( ) 1M 1/4W resistor between switch terminals (save for re-use)

( ) bare wire between switch terminals

( ) 2. Unsolder the end of the 0.0022 μF capacitor from the switch.

( )) Resolder it to the terminal indicated in Fig. 10.

( ) 3. Clean the excess solder from the switch terminals.

( ) 4. Unsolder the white-black-violet wire from 'A' SELECTOR switch W2-8R.

( ) 5. Remove the nylon spacing clip between the white-black-violet and white-yellow wires.

##

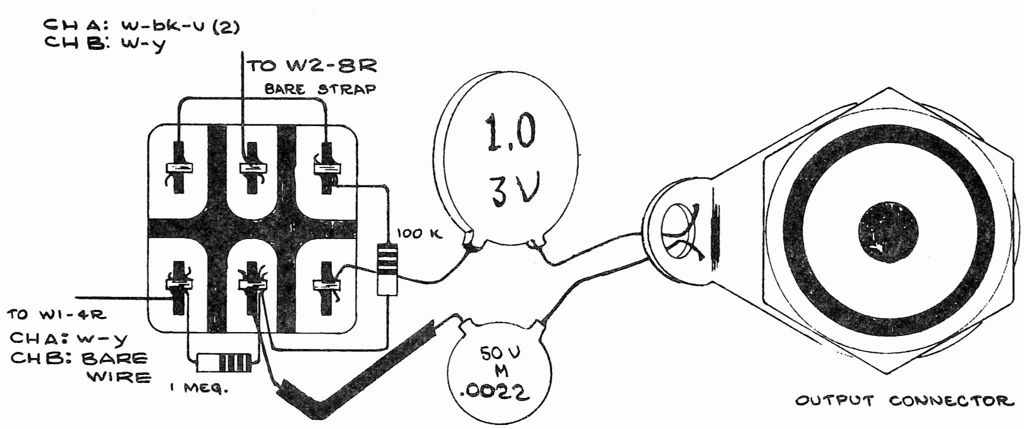


Fig. 10

( ) 6. Locate the 100k 1/4W resistor on the 'A' SELECTOR switch, soldered between W1-5R and W2-8R.

( ) Unsolder the end from W2-8R and resolder it to W1-8F.

( ) 7. Dress the white-black-violet wire, unsoldered in step D-4, to 'A' LF REJECT switch.

( ) Cut off the excess wire and solder, along with the white-black-violet wire (from kit), to the switch terminal indicated in Fig. 10.

( ) 8. Resolder the remainder of the components and wires as indicated in Fig. 10, using the bare wire from the kit.

( ) 9. Dress the other end of the white-black-violet wire (step D-7) through the grommet in the 'A' SELECTOR switch bracket.

( ) Solder this wire to the 'A' SELECTOR switch W2-8R.

( ) 10. Locate 100k 1/4W resistor on 'B' SELECTOR switch, soldered between W1-5R and W2-8R.

( ) Unsolder the end from W2-8R and resolder it to W1-8F.

11. Unsolder the following components and wires from 'B' INTEGRATOR LF REJECT switch:

( ) 100k 1/4W resistor (save for re-use)

( ) 1 M 1/4W resistor between switch terminals (save for re-use)

( ) bare wire between switch terminals

( ) 12. Unsolder and remove the bare wire between the 'B' INTEGRATOR LF REJECT switch and the 'B' SELECTOR switch W1-4R.

( ) 13. Unsolder the end of the 0.0022 μF capacitor from 'B' INTEGRATOR LF REJECT switch.

( ) Resolder it to the terminal indicated in Fig. 10.

( ) 14. Clean the excess solder from the switch terminals.

( ) 15. Solder the 2-1/2 in. length of white-yellow wire (from kit) between the 'B' SELECTOR switch W2-8R and the switch terminal indicated in Fig. 10.

( ) 16. Solder a 1-1/2 in. length of bare wire (from kit) to 'B' SELECTOR switch W1-4R.

( ) Place a 1 in. length of plastic tubing (from kit) over the wire and dress the wire to the switch terminal indicated in Fig. 10. DO NOT SOLDER until the next step.

( ) 17. Resolder the components and wires to the switch as indicated in Fig. 10, using the bare wire from the kit.

( ) 18. Solder an 0.01 μF ceramic capacitor (from kit) between the two wired terminals of the Grid Current 'A' potentiometer (R5535) on the potentiometer bracket.

( ) 19. Similarly, solder an 0.01 μF capacitor (from kit) between the two wired terminals of the Grid Current 'B' potentiometer (R5585).

INSTRUCTIONS (cont)

THIS COMPLETES THE INSTALLATION

( ) Check wiring for accuracy

( ) Adjust the Variable Z_i and Variable Z_f capacitors as described in your Instruction Manual.

( ) Fasten the insert page in your Instruction Manual.

( ) Moisten the back of the MODIFIED INSTRUMENT tags (from kit) and place them on the Manual Schematic pages affected by this modification.

JB:cet

OPERATIONAL AMPLIFIER IMPROVEMENTS

Type 'O' Plug-in Units -- SN 155-813*

GENERAL INFORMATION

This modification incorporates several refinements in the operational amplifiers of the Type O unit.

Section A improves the crosstalk characteristics by relocating several wires and changing the decoupling arrangement.

Sections B and C improve the temperature-sensitive drift characteristics of the 'A' and 'B' amplifiers.

Section D increases the accuracy of the output amplifier when Z_f is set at 10 pF and the INTEGRATOR LF REJECT is OFF.*

The instructions are divided so that any part of the modification may be performed separately if desired.

*Section D applies to SN 155-318 (with the exception of a few instruments in this range already modified at the factory).

The following parts list assumes all parts of the modification have been completed.

ELECTRICAL PARTS LIST

Values fixed unless marked variable. Only new parts listed.

Ckt. No.	Part Number	Description			
		CAPACITORS			
Tolerance ±20% unless otherwise indicated.					
C5535	283-0003-00	0.01 µF	Cer	150V	
C5538	283-0004-00	0.02 µF	Cer	150V	
C5585	283-0003-00	0.01 µF	Cer	150V	
C5588	283-0004-00	0.02 µF	Cer	150V	
		DIODES			
D5529	152-0055-00	Zener	11V	1/4W	5%
D5579	152-0055-00	Zener	11V	1/4W	5%
		INDUCTORS			
L5520	276-0507-00	Core, ferramic suppressor 255, 0.6 µH			
L5530	276-0507-00	Core, ferramic suppressor 255, 0.6 µH			
L5570	276-0507-00	Core, ferramic suppressor 255, 0.6 µH			
L5580	276-0507-00	Core, ferramic suppressor 255, 0.6 µH			

Ckt. No.	Part Number	Description			
		RESISTORS			

Resistors are 10% composition unless otherwise indicated.

R5520	Delete				
R5530	Delete				
R5570	Delete				
R5580	Delete				
R5521	309-0354-00	45 k	1/2W	prec	1%
R5523	305-0273-00	27 k	2W		5%
R5525	323-0347-00	40.2 k	1/2W	prec	1%
R5526	308-0241-00	22 k	7W	WW	1%
R5531	309-0354-00	45 k	1/2W	prec	1%
R5538	315-0102-00	1 k	1/4W		5%
R5539	305-0823-00	82 k	2W		5%
R5571	309-0354-00	45 k	1/2W	prec	1%
R5573	305-0273-00	27 k	2W		5%
R5575	323-0347-00	40.2 k	1/2W	prec	1%
R5576	308-0241-00	22 k	7W	WW	1%
R5581	309-0354-00	45 k	1/2W	prec	1%
R5588	315-0102-00	1 k	1/4W		5%
R5589	305-0823-00	82 k	2W		5%

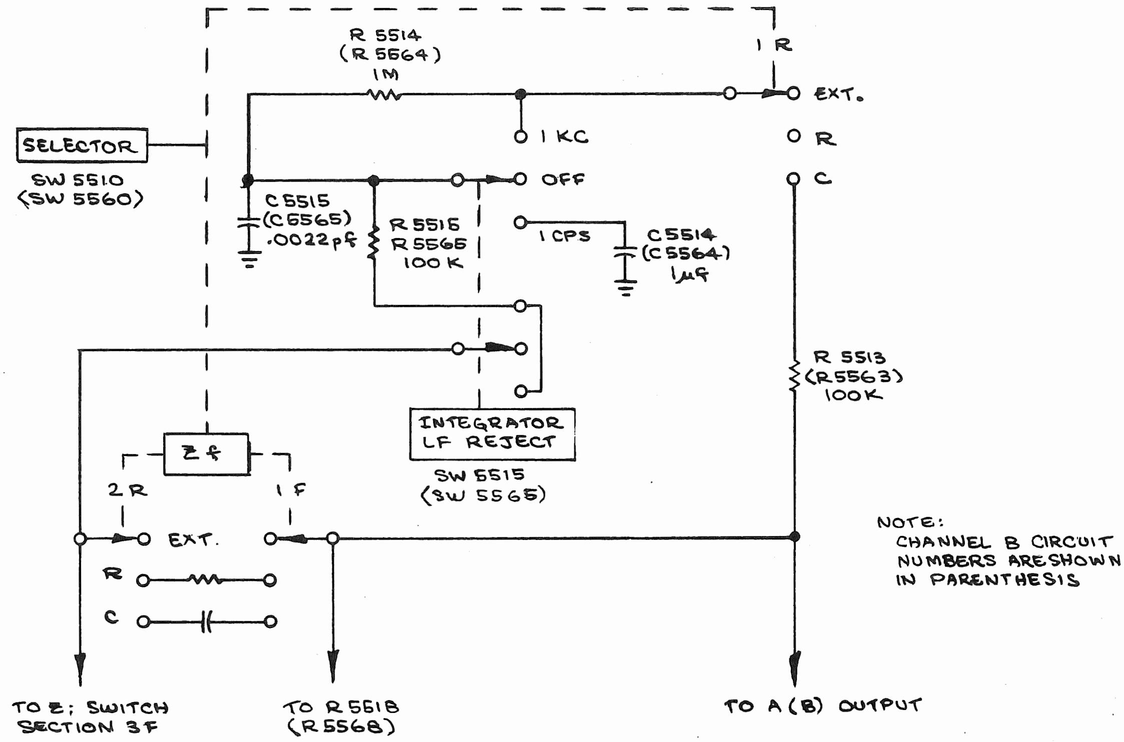
TRANSISTORS

Q5523	151-0096-00	NPN, Planar, TEK Special
Q5573	151-0096-00	NPN, Planar, TEK Special

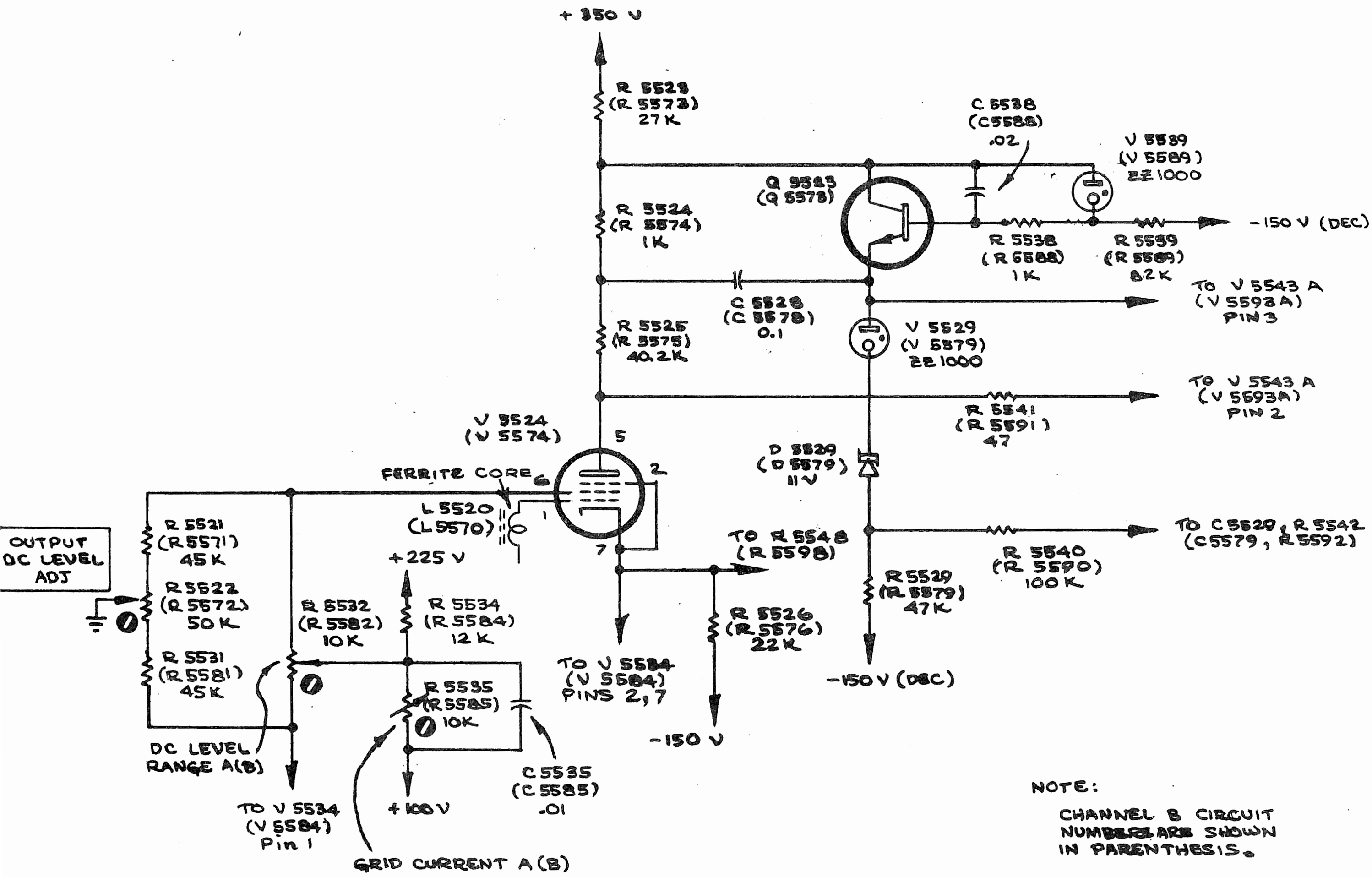
TUBES

V5529	154-0370-00	ZZ1000	Gas diode
V5539	154-0370-00	ZZ1000	Gas diode
V5579	154-0370-00	ZZ1000	Gas diode
V5589	154-0370-00	ZZ1000	Gas diode

##



Channels A and B OPERATION AMPLIFIERS
(Partial Diagram)

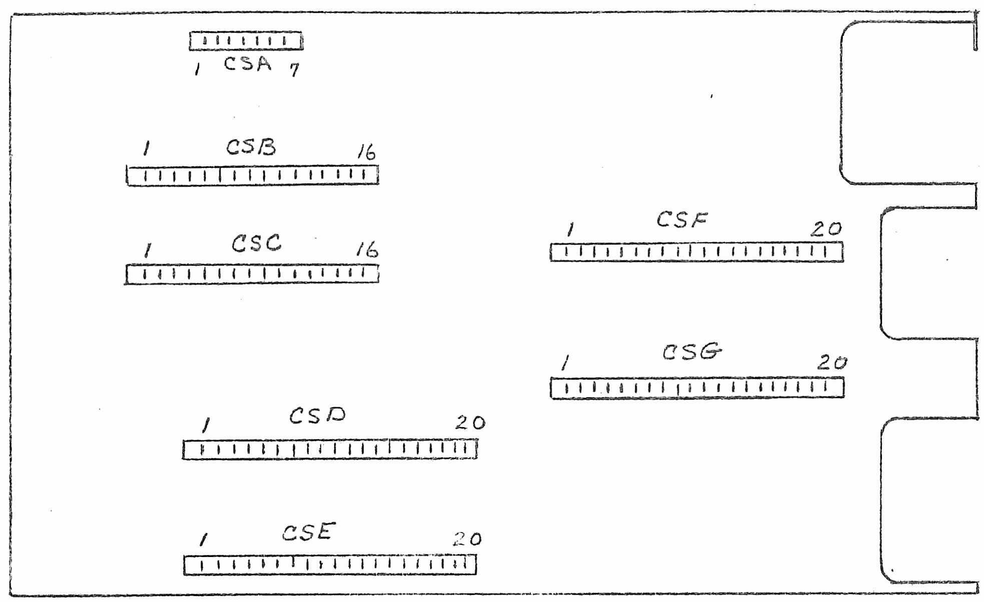


Channel A and B OPERATIONAL AMPLIFIERS
(Partial Diagram)

Also change R5530 (R5580) from 47 Ω resistor to ferrite core, L5530 (L5580).

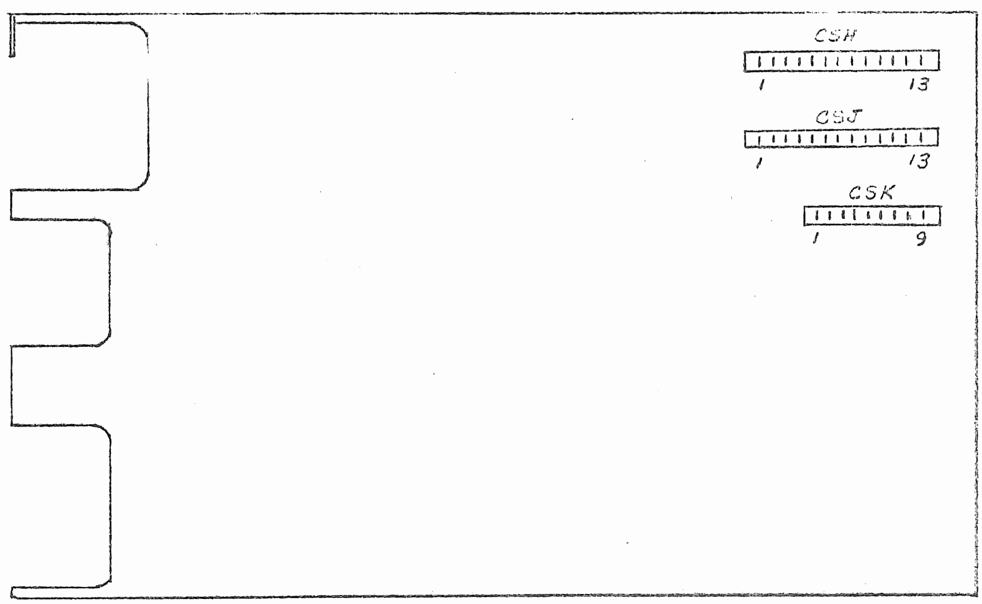
STRIP LAYOUT – 0

FRONT VIEW



CHASSIS – 610-051

BACK VIEW



MOD #	
DATE	
KIT #	

2-6-62

MAINTENANCE NOTES

GRID CURRENT

Specs = .5 namp max; adjustable to .3 (-), .15 (+)

We've changed final specs for O unit grid current to "less than .5 namp; internally adjustable to less than .3 namp at the -GRID and less than .15 namp at the +GRID" for two reasons:

1. Grid current tends to vary slightly (±100 pamp in some cases) over 50 to 100 hours operation, and also may vary when the O unit is plugged into different scopes or when it is shipped. A little leeway between test specs and advertised specs means less chance of customer rejects based on picoamp fly specking.

2. There isn't much possibility that the test spec (.3 namp and .15 namp) could be tightened and still provide a reasonable tube yield.

Tube brands

Recent evaluation of tubes of various brands aged 100 hours indicates we can get a good yield of 12AU6's to meet grid current specs from the following brands:

Westinghouse (95%)
Raytheon
GE (passive cathode)
Nippon
Brimar (British)

Evaluation of Telefunken and RCA 12AU6's showed poor yield, even after aging. We're setting up a 157 selected tube number for matched pairs meeting grid current requirements. In the meantime, the plant is selecting Westinghouse 157-050's.

Grid current adjustment mod at sn155

Mod 5728-O, effective sn155, adds a grid current adjustment pot and aids considerably in getting a good yield of tubes. It also helps the sharp-type customer adjust grid current to extremely low (but less stable) values for critical low level applications.

RELIABILITY -- C6589 .1 μf FAILURE

March 12, 1963

Problem - C6589 failures

C6589 (.1 μf 100 v disc type 283-012) may fail prematurely or excessively. We're not sure of the reason but it may be because C6589 is tied to +75 v, only 25 v away from its rating or because this particular type of disc capacitor has a poor overall reliability record in general--we had a lot of trouble with the 283-057's, both are Erie.

Temporary solution--use 283-012's

About the best thing you can do now is replace the 283-012's with 283-057 (Sprague, only) 200 volters. We hope the GE's flat, square Mylar caps work out, however, at the moment we seem to be doing GE's evaluation engineering for them--their samples look like they're still learning how to build 'em.

INPUT CAPACITOR RANGE

Greater range of the input capacitors (C-6541, C-6521) is needed.

SOLUTION: Mod 7437 will install 3-12 pf capacitors (281-036) in order to utilize wider variations of the input amplifier tubes.

VARIABLE VOLTS/CM POTENTIOMETER

FEN 5-11-62

Modification 3862 converts O Units, sn 101 up, to the new Tek-made "stopless" potentiometers. These new pots have two advantages:

1. Provide a new, more realistic design-center resistance value and tolerance.
2. Remove the stop, for continuous rotation (but with a detent for the "calibrated" position).

This should eliminate many shaft and pot problems, and provide greater operator convenience.

R6530 resistor, 710 Ω ±15% Tek-made var 311-259

All of the new "stopless" pots will have a distinctive grey delrin cover, with the Tek number, value, and (±15%) tolerance molded in.

POSITION POT DEFECTS, SN 1340-1689

FEN 5-8-64

Some 5000 Clarostat 311-028 2 x 100k pots from batches code-dated 6313 and 6314 have been shipped in instruments, prior to discovery that a sizeable percentage of these pots may be defective. The defect is a mechanical misalignment in some cases sufficient to allow the rotor in some positions to contact the element in two places, thus shorting out about 2/3 of the element. In its typical VERTICAL POSITION control applications, the pot is wired directly to the power supply buss, and may be damaged by excessive dissipation.

The instrument types which may contain defective pots were all shipped in summer and early fall 1963.

Defective 311-028 pots should be returned to Roger Ady.

OSCILLATION/COMPATIBILITY PROBLEM

GS 11-25-64

The Type O plug-in/547 combination may oscillate at approximately 200 mc.

Production mod 8983, starting sn 2426 (with 64 exceptions) will:

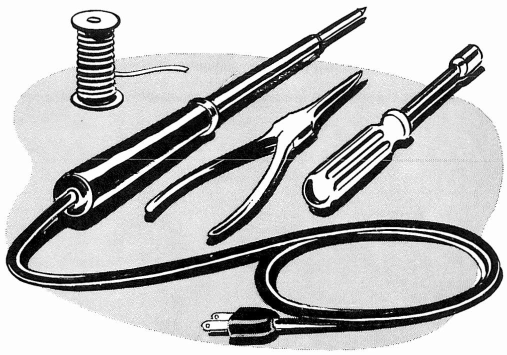
Add L6575, core 276-528, to end of 9-1 wire at top lead of L6574.
Add L6565, core 276-528, to end of 9-6 wire at top lead of L6564.
Both wire ends must be stripped back to accept the beads.

CERAMIC STRIP LEAKAGE VS GRID CURRENT

FEN, 11-20-64
Revised, 11-65

The range of the grid-current adjustment pots in the O-Unit operational amplifiers in s/n 101-2651 was predicated on a certain small amount of leakage across the ceramic strips from the +350 V supply (A) or from the anode of the ZZ1000 (B) to the -grid, and from the -150 V supply to the +grid. If the surface of the strips becomes contaminated from operation in adverse atmospheric environments or from soldering, the leakage may become excessive (10^6 megohms will provide leakage equal to the actual grid current) and prompt the rejection of 12AU6's for "too much grid current".

The problem became particularly noticeable in the installation of kit 040-0301-00 in sn 101-813. The layout used in production sn 814-2651 was based on the strip leakage after walnut-blast; without the same degree of flux-removal in the field, the leakage became excessive. Complete removal of the grid connections from the strips threw the grid current adjustment pots out of range in the opposite direction in some sample instruments, leading to the belief that this most obvious solution could not be used. Later, however, excessive variation of apparent grid current with temperature even in walnut-blasted production units prompted a more thorough investigation, and it was determined that the leakage was not necessary to keep the grid-current pots in range. Production mod 9434 was implemented on a #1 rework basis in March 1965 to get the grid connections off the strips and directly to the tube sockets, replacing the 47 Ω "stopper" resistors with ferrite beads. Effective s/n of the mod was 2651, with 27 lower s/n's modded out of sequence. Kit 040-0301-01 incorporates the layout change, which is now recommended for all early s/n's. Previous FEN and Green Sheet suggestions for ultra-thorough strip cleaning, SC-87 dri-film, etc., may be considered obsolete.



SECTION 6

MAINTENANCE

PREVENTIVE MAINTENANCE

Calibration

The Type O Plug-In Unit will not require frequent calibration. However, to insure that the unit is operating properly at all times we suggest that you check the calibration after each 500-hour period of operation (or every six months if the unit is used intermittently). A complete step-by-step procedure for calibrating the unit and checking its operation is given in the Calibration Section of this manual.

The accuracy of measurements made with the O Unit depends not only on the accuracy of calibration, but also on the calibration of the associated oscilloscopes. It is important for the oscilloscope to be maintained in proper calibration.

Visual Inspection

Troubles can sometimes be found by a visual inspection of the unit. For this reason, you should perform a complete visual check every time the instrument is calibrated or repaired. Look for such defects as loose or broken connections, damaged connectors, improperly seated tubes, scorched or burned parts, broken terminal strips, etc. The remedy for these troubles is apparent, except for heat-damaged parts. Heat damage is often the result of other, less apparent trouble. It is essential for you to determine the cause of overheating before replacing damaged parts.

Tube Checks

Tube-tester checks on the tubes used in the Type O Unit are not recommended. Tube testers sometimes indicate a tube to be defective when that tube is operating satisfactorily in a circuit, or they may fail to indicate tube defects which affect the performance of the circuits. The criterion for useability of a tube is whether or not it works properly in the circuit. If it does not, then it should be replaced. Unnecessary replacement of tubes is not only expensive but may also result in needless recalibration of the instrument.

To obtain maximum reliability and performance, we check some of the tubes used in our instruments for such characteristics as G_m, microphonics, balance, etc. We age other tubes to stabilize their characteristics. The checked tubes are labeled and identified with a part number beginning with 157___. Raw-stock . . . that is, unchecked tubes . . . are unlabeled tubes assigned the part number 154___.

COMPONENT REPLACEMENT

General

The procedures for replacing most parts in the O Unit are easy. Detailed instructions for their removal are therefore not required. In some cases, however, additional information may help you. This information is contained in the following paragraphs. Because of the circuit configuration, it will be necessary to recalibrate portions of the circuit when certain parts are replaced. Refer to the Calibration Section of this manual.

Switches

Procedures for the removal of defective switches are, for the most part, obvious and only a normal amount of care is required. If a switch is removed, careful notation of the leads to the switch should be made to facilitate connecting the new switch.

Single wafers are not normally replaced on the switches used in the O Unit. If one wafer is defective, the entire switch should be replaced. Switches may be ordered from Tektronix either unwired or with parts wired in place.

Soldering and Ceramic Strips

Many of the components in your Tektronix instrument are mounted on ceramic terminal strips. The notches in these strips are lined with a silver alloy. Repeated use of excessive heat, or use of ordinary tin-lead solder will break down the silver-to-ceramic bond. Occasional use of tin-lead solder will not break the bond if excessive heat is not applied.

If you are responsible for the maintenance of a large number of Tektronix instruments, or if you contemplate frequent parts changes, we recommend that you keep on hand a stock of solder containing about 3% silver. This type of solder is used frequently in printed circuitry and should be readily available from radio-supply houses. If you prefer, you can order the solder directly from Tektronix in one-pound rolls. Order by Tektronix part number 251-514.

Because of the shape of the terminals on the ceramic strips it is advisable to use a wedge-shaped tip on your soldering iron when installing or removing parts from the strips. Fig. 6-1 will show you the correct shape for the tip of the soldering iron. Be sure to file smooth all surfaces of the iron which will be tinned. This prevents solder from building up on rough spots where it will quickly oxidize.

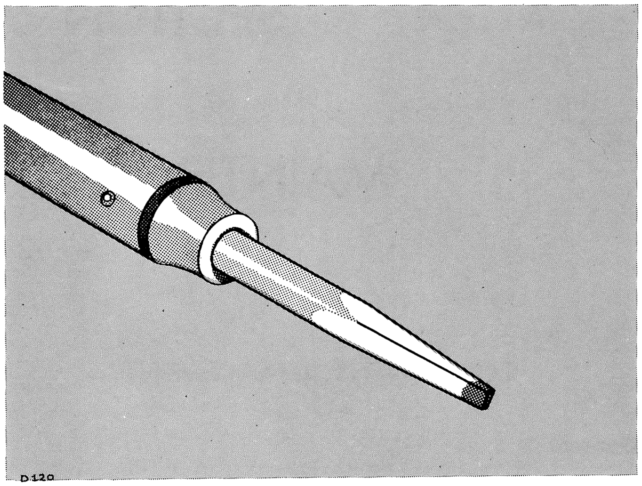


Fig. 6-1. Soldering iron tip properly shaped and tinned.

When removing or replacing components mounted on the ceramic strips you will find that satisfactory results are obtained if you proceed in the manner outlined below.

1. Use a soldering iron of about 75-watt rating.
2. Prepare the tip of the iron as shown in Fig. 6-1.
3. Tin only the first $1/16$ to $1/8$ inch of the tip. For soldering to ceramic terminal strips tin the iron with solder containing about 3% silver.
4. Apply one side of the tip to the notch where you wish to solder (see Fig. 6-2).

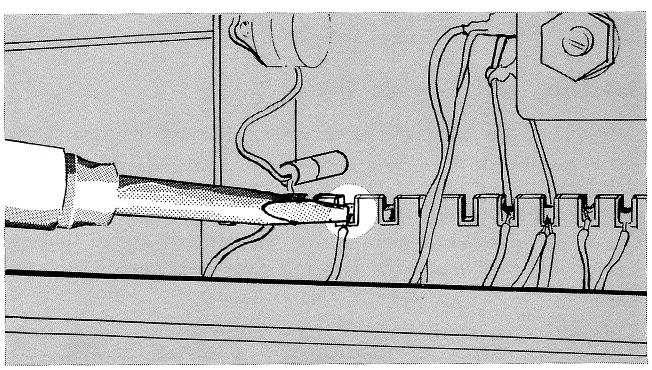


Fig. 6-2. Correct method of applying heat in soldering to a ceramic strip.

5. Apply only enough heat to make the solder flow freely.
6. Do not attempt to fill the notch on the strip with solder; instead, apply only enough solder to cover the wires adequately, and to form a slight fillet on the wire as shown in Fig. 6-3.

In soldering to metal terminals (for example, pins on a tube socket) a slightly different technique should be employed. Prepare the iron as outlined above, but tin with ordinary tin-lead solder. Apply the iron to the part to be

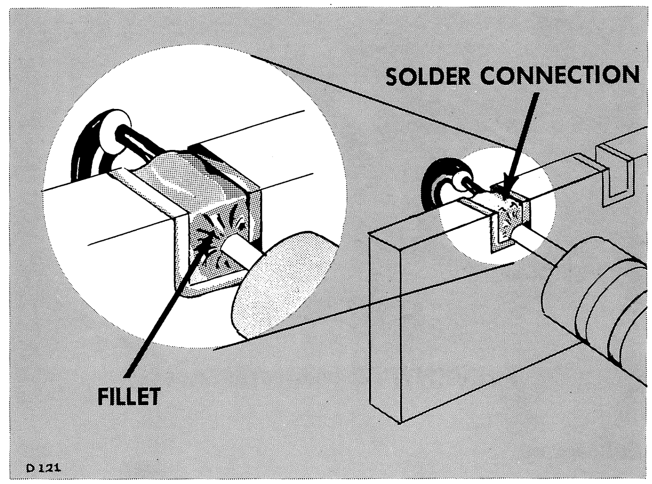


Fig. 6-3. A slight fillet of solder is formed around the wire when heat is applied correctly.

soldered as shown in Fig. 6-4. Use only enough heat to allow the solder to flow freely along the wire so that a slight fillet will be formed as shown in Fig. 6-4.

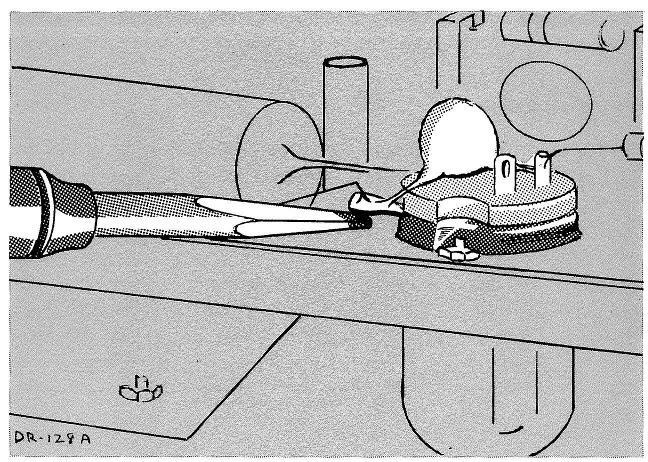


Fig. 6-4. Soldering to a terminal. Note the slight fillet of solder —exaggerated for clarity—formed around the wire.

General Soldering Considerations

When replacing wires in terminal slots clip the ends neatly as close to the solder joint as possible. In clipping ends of wires take care the end removed does not fly across the room as it is clipped.

Occasionally you will wish to hold a bare wire in place as it is being soldered. A handy device for this purpose is a short length of wooden dowel, with one end shaped as shown in Fig. 6-5. In soldering to terminal pins mounted in plastic rods it is necessary to use some form of "heat sink" to avoid melting the plastic. A pair of long-nosed pliers (see Fig. 6-6) makes a convenient tool for this purpose.

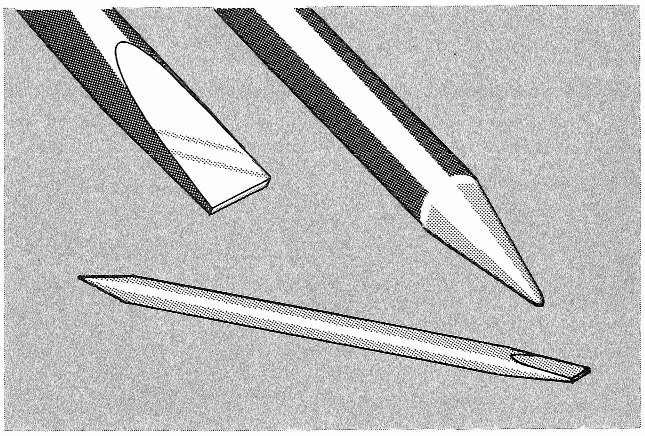


Fig. 6-5. A soldering aid constructed from a 1/4 inch wooden dowel.

Ceramic Strips

To replace strips which mount with snap-in plastic fittings, first remove the original fittings from the chassis. Assemble the mounting post on the ceramic strip. Insert the nylon collar into the mounting holes in the chassis. Carefully force the mounting post into the nylon collars. Snip off the portion of the mounting post which protrudes below the nylon collar on the reverse side of the chassis.

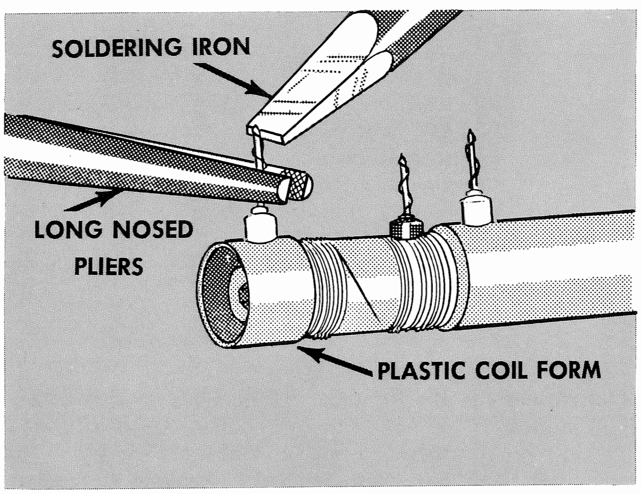


Fig. 6-6. Soldering to a terminal mounted in plastic. Note the use of the long-nosed pliers between the iron and the coil form to absorb the heat.

NOTE

Considerable force may be necessary to push the mounting posts into the nylon collars. Be sure that you apply this force to that area of the ceramic strip directly above the mounting posts.

TROUBLESHOOTING

General Troubleshooting Information

This portion of the manual is intended to help you troubleshoot the Type O Unit.

Since the Type O Unit derives all of its operating voltages from the oscilloscope, and depends on the oscilloscope for its display, you must be sure that the oscilloscope is not the cause of trouble. Trouble can usually be isolated to either the oscilloscope or plug-in unit by substituting another plug-in for the suspected one and checking for proper operation. Or you can insert the suspected O Unit in another oscilloscope and check for proper operation.

If trouble occurs in the Type O Unit, try to isolate it by quick operational and visual checks. First check the settings of all controls. Then operate the controls to see what effect, if any, they have on the trouble. The normal or abnormal operation of each control may help you to establish the trouble symptoms. (The cause of trouble which occurs only in certain positions of a control can usually be determined immediately from the trouble symptoms.)

After the trouble symptoms are established, look first for simple causes of trouble. Check to see that the pilot light of the oscilloscope is on, feel for any irregularities in the operation of the controls, listen for any unusual sound, see that the tube filaments are lit, and visually check the entire instrument. The type of trouble will generally indicate the checks to make.

In general, a troubleshooting procedure consists of two parts: circuit isolation and circuit troubleshooting. Since the Type O Unit is a relatively simple unit, divided into three independent circuits, it will be apparent which of the three is defective. After isolating the circuit, you can then troubleshoot in the circuit to find the cause of the trouble.

Table 6-1 lists troubles which can occur in the Type O Unit, the probable causes, and checks to make. The table is divided into two sections, the Preamplifier and the Operational Amplifiers. If trouble occurs, determine which section of Table 6-1 to use. Then try to identify the trouble with one of the steps in the table.

Most troubles will be caused by tube or semiconductor failures. Therefore, when trouble has been isolated to a circuit, the tubes and semiconductors in that circuit should be checked (by substitution). Be sure to return tubes and transistors found to be good to their original socket.

Switch wafers shown with the circuit diagrams are coded to indicate the position of the wafer on the switches. The number portion of the code refers to the wafer number of the switch assembly. Wafers are numbered from the front of the switch to the rear. The letters F and R indicate whether the front or the rear of the wafer is used to perform the particular switching function.

Test Points

Major test points in the unit are shown on the schematics and in Fig. 7-1. A test point is indicated by a number with a line indicating the location of the test point in the circuit. Test points are used as an aid in troubleshooting and calibrating the unit, and reference to these points is made in Table 6-1 and in the Calibration Procedure (Section 7).

Voltage measurements, and the conditions under which they were obtained, are shown on the schematics.

Test points on the schematics are numbered consecutively starting with the diagram for the Preamplifier. Numbers increase from right to left across the page. The Operational Amplifiers have identical test points and voltages.

TABLE 6-1

TROUBLE	PROBABLE CAUSE	CHECKS TO MAKE
	PREAMPLIFIER	
1. Gain of all signals low.	1. Front-panel GAIN ADJ. needs adjustment. 2. Oscilloscope vertical amplifier gain needs adjustment.	1. Check gain with the calibrator signal. 2. If unable to get enough gain, check the oscilloscope amplifier by using a Gain-Set Adapter between the plug-in and the oscilloscope.
2. Trace shifts vertically as VARIABLE control is rotated.	DC BAL. control needs adjustment.	Adjust the DC BAL. control until the trace does not move as the VARIABLE control is rotated.
3. Severe loss of gain and the vertical POSITION control works backward.	D6576 shorted.	D6576 should have 6.3 volts across it. If shorted, test points (1) and (3) will show higher than normal voltage.
4. When the unit is in a dual-sweep scope such as a 535A or 545A, noise is observed on trace when the B Sweep is running at 10 μSEC/CM.	C6539 open. C6576, C6582, C6584, C6586, C6588 or C6589 open.	Check C6539. Check all power-supply by-pass capacitors.
5. Noise on trace when both operational amplifiers are used as multivibrators at full 5-ma output.	Same as step 4.	Same as step 4.
6. Oscillations appear with signal near the top or bottom of the crt, but not in the center two centimeters.	C6564 or C6574 need adjustment.	See adjustment 10 in Calibration Procedure.
7. Tube heaters cold, oscilloscope +100-volt power supply out of regulation.	C6589 shorted. C6576 shorted.	If either C6589 or C6576 are shorted, look for damaged D6576, Q6564, Q6574, and tube heaters or resistors in oscilloscope that are in series with interconnecting plug terminal No. 15. Also look for burned resistors associated with Q6564 and Q6574.
	OPERATIONAL AMPLIFIERS SN 814-up	
1. Constant gain about half normal, and output voltage at about +25 volts.	Q5523 (A) or Q5573 (B) shorted. Q5529 (A) or C5579 (B) shorted.	The transistor should have approximately 85 volts across it. If shorted, replace, and check to see if R5523 (A) or R5573 (B) has been damaged. Replace if there is doubt. Check C5529 (A) or C5579 (B).
2. Overshoot of calibrator signal at unity gain with Z_i and Z_f both at 1 MEG. Output voltage about 3 or 4 volts negative.	D5529 (A) or D5579 (B) shorted. C5528 (A) or C5578 (B) shorted.	The Zener diode should have approximately 11 volts across it. If shorted replace. Check C5528 (A) or C5578 (B) for short. Check R5523 (A) or R5573 (B) for damage.
	SN 101-813	
1. Constant gain about half normal, and output voltage at about +25 volts.	D5529 (A) or D5579 (B) shorted. C5529 (A) or C5579 (B) shorted.	The Zener diode should have from 95 to 105 volts across it. If shorted, replace, and check to see if R5523 (A) or R5573 (B) has been damaged. Replace if there is doubt. Check C5529 (A) or C5579 (B).

TROUBLE	PROBABLE CAUSE	CHECKS TO MAKE
All serial ranges		
3. Gain of operational amplifier low for high resistance values in constant-gain circuit. Gain is correct for low resistance values.	With internal feedback resistors: R5509, F and G (A), or R5559, F and G (B), open. With external feedback resistors: use Error Factor formula, page 4-1.	Check R5509, F and G (A). Check R5559, F and G (B).
4. Trace is shifted off crt from normal position. Can be returned to crt with vertical POSITION control. Operational Amplifier is very sensitive to external line-frequency interference.	±GRID SEL switch is in + position, and the + grid circuit is open.	Place ±GRID SEL switch in (—) position, or ground + grid.

NOTES

PARTS REPLACEMENT KIT

SELECTOR "A" SWITCH

For Tektronix Type "O" Plug-in Units
Serial numbers 101-318, with exceptions

050-0074-00

DESCRIPTION

New operational amplifier "A" SELECTOR switch 262-0518-00 replaces 262-0424-00 previously used.

The new switch is changed to accommodate a modification which improves the accuracy of the output amplitude with a Z_f of 10 pF and the INTEGRATOR LF REJECT off.

NOTE: If the serial number of your instrument is above those listed, or if this kit has been installed, disregard the instructions as P/N 262-0518-00 is a direct replacement.

Publication:
Instructions for 050-0074-00
March 1966

Supersedes:
September 1963

PARTS LIST

Qty.	Part Number	Description			
(1 ea)		Assembly, SELECTOR switch (modified 262-0518-00), consisting of:			
1 ea	210-0012-00	Lockwasher, int. 3/8 x 1/2			
2 ea	210-0406-00	Nut, hex, 4-40 x 3/16			
1 ea	210-0413-00	Nut, hex, 3/8-32 x 1/2			
1 ea	210-0840-00	Washer, steel, 0.390 x 9/16 x 0.020			
1 ea	260-0431-00	Switch, raw			
2 ea	281-0031-00	Capacitor, cer, var,	3-12 pF		
2 ea	281-0048-00	Capacitor, cer, var,	5-25 pF		
1 ea	281-0504-00	Capacitor, cer,	10 pF	500V	±1 pF
2 ea	281-0528-00	Capacitor, cer,	82 pF	500V	±8.2 pF
2 ea	281-0537-00	Capacitor, cer,	0.68 pF	500V	±0.136 pF
3 ea	281-0538-00	Capacitor, cer,	1 pF	500V	±0.2 pF
1 ea	301-0244-00	Resistor, comp,	240 k	1/2W	5%
1 ea	301-0514-00	Resistor, comp,	510 k	1/2W	5%
2 ea	309-0100-00	Resistor, prec,	10 k	1/2W	1%
2 ea	309-0140-00	Resistor, prec,	500 k	1/2W	1%
2 ea	309-0148-00	Resistor, prec,	1 M	1/2W	1%
2 ea	309-0260-00	Resistor, prec,	100 k	1/2W	1%
2 ea	309-0444-00	Resistor, prec,	200 k	1/2W	1%
1 ea	316-0104-00	Resistor, comp,	100 k	1/4W	10%
2 ea	348-0031-00	Grommet, poly, 1/4			
1 ea	406-0720-00	Bracket, switch end			
1 ea	406-0724-00	Bracket, SELECTOR switch			
1 ea	(175-0522-00)	Wire, #22 solid,	6 in.	white-black-violet	
1 ea	(162-0504-00)	Tubing, plastic, #20,	1-1/4 in.	black	
1 ea	(175-0522-00)	Wire, #22 solid,	2-1/2 in.	white-yellow	
1 ea	(176-0005-00)	Wire, #22 solid,	4 in.	bare	

INSTRUCTIONS

NOTE: The following method is used to identify the SELECTOR switch terminals:

The wafers are numbered from front to rear.

The contact positions are numbered 1 through 12 relative to the index key as shown in Fig 1.

The contacts have an 'F' or 'R' suffix which denotes that they are on the front or rear of the wafer.

Example: W2-7R (denoted by * on drawing) is contact #7 on the rear of wafer 2.

(TYPICAL SWITCH CONFIGURATION)

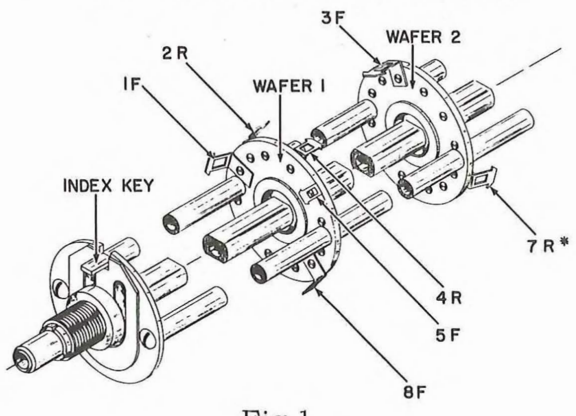


Fig 1

TO REMOVE THE OLD SWITCH:

( ) 1. Remove tubes V5524, V5534, and V5543.

( ) 2. Remove the two screws securing the potentiometer bracket to the rear panel, and bend the bracket forward slightly.

NOTE: Save all hardware until installation is completed.

( ) 3. Remove the screws securing the front and rear mounting brackets of the "A" SELECTOR switch.

( ) 4. Remove the knobs from the switch shaft.

( ) 5. Remove the front panel switch shaft bushing.

( ) 6. Unsolder all wires going to the switch and remove the switch from the plug-in unit.

TO INSTALL THE NEW SWITCH:

( ) 7. Place the front panel bushing nut (removed from old switch) on the shaft of the SELECTOR switch from the kit.

( ) Place the switch in position in the instrument. DO NOT MOUNT YET.

8. Solder the wires to the switch as indicated below (see Fig 2):

( ) white-green to W3-12F
( ) white-yellow to W3-1F
( ) solid white to W3-2F
( ) white-black-red to W3-3F
( ) white-blue to W2-2R
( ) white-violet to W2-3R
( ) white-orange to W2-4R
( ) white-red (no.26) to W2-5R
( ) white-gray to W1-3F
( ) white-red (no.22) to W1-8F
( ) white-black-orange to W4-3R
( ) coaxial to W4-6R

( ) 9. Replace the front panel bushing.

( ) Secure the switch mounting brackets, using the hardware removed in step 3.

( ) Remount the potentiometer bracket (removed in step 2).

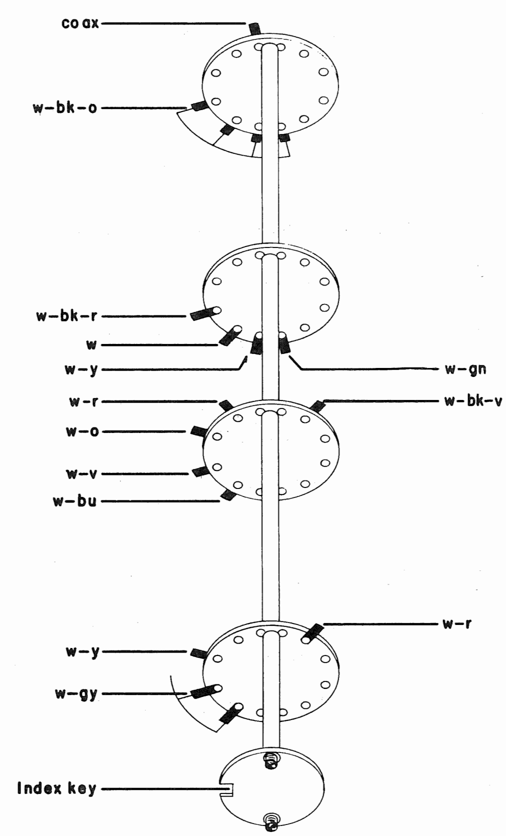


Fig 2

TO ALIGN THE KNOBS:

( ) 10. Place the larger (Z_f) knob on the shaft and tighten just enough to allow rotation of the switch.

( ) 11. Rotate the switch until the rotor contacts of W1 and W2 are at position 9 (a 1 Meg resistor is connected between W1 and W2 at this position.

( ) 12. Loosen the set-screw, align the knob index to indicate R = 1 Meg, and tighten it securely.

( ) 13. Place the Z_i knob on its shaft and tighten temporarily.

( ) 14. Rotate the switch until the rotor contacts of W3 and W4 are at position 7 (a 1 Meg precision resistor is connected between W3 and W4 at this position)

( ) 15. Loosen the set-screw, align the knob index to indicate R = 1 Meg, and tighten it securely.

TO MODIFY AMPLIFIER "A":

16. Unsolder and remove the following components and wires associated with the "A" INTEGRATOR LF REJECT switch:

( ) 100 k 1/4W resistor from a switch terminal to the OUTPUT coaxial connector.

( ) white-yellow wire from a switch terminal (other end was formerly connected to the old SELECTOR switch).

( ) 1 Meg 1/4W resistor between two switch terminals.

( ) wire strap between two switch terminals.

( ) 17. Unsolder the end of the 0.0022 μF capacitor from the switch terminals.

( ) Resolder it to the terminal indicated in Fig 3.

( ) 18. Clean the excess solder from the switch terminals.

( ) 19. Locate the white-black-violet wire from the "A" ±GRID SEL switch that was unsoldered from the SELECTOR switch.

( ) Remove the nylon spacing clip between the white-black-violet and the white-yellow wires.

( ) 20. Dress the wire to the LF REJECT switch.

( ) Cut off the excess wire and solder, along with the white-black-violet wire from the new SELECTOR switch, W2-8R, to the switch terminal indicated in Fig 3.

( ) 21. Resolder the components and wires indicated in Fig 3.

TO MODIFY AMPLIFIER "B":

22. Locate the 100k 1/4W resistor on the "B" SELECTOR switch, soldered between W1-5R and W2-8R.

( ) Unsolder the end from W2-8R.

( ) Resolder it to W1-8F.

23. Unsolder and remove the following components and wires associated with the "B" INTEGRATOR LF REJECT switch:

( ) 100k 1/4W resistor from a switch terminal to the OUTPUT coaxial connector.

( ) 1 Meg 1/4W resistor between two switch terminals.

( ) wire strap from a switch terminal to the "B" SELECTOR switch, W1-4R.

( ) wire strap between two switch terminals.

( ) 24. Unsolder the end of the 0.0022 μF capacitor from the switch terminal.

( ) Resolder it to the terminal indicated in Fig 3.

( ) 25. Clean the excess solder from the switch terminals.

( ) 26. Solder the 2-1/2 in. length of white-yellow wire (from kit) between the "B" SELECTOR switch, W2-8R, and the switch terminal indicated in Fig 3 (in parenthesis).

( ) 27. Solder a 1-1/2 in. piece of bare wire (from kit) to the switch, W1-4R.

( ) Place the length of insulated tubing (from kit) over the wire and dress the wire to the switch terminal indicated in Fig 3. DO NOT SOLDER YET.

( ) 28. Resolder the components and wires to the switch as indicated in Fig 3.

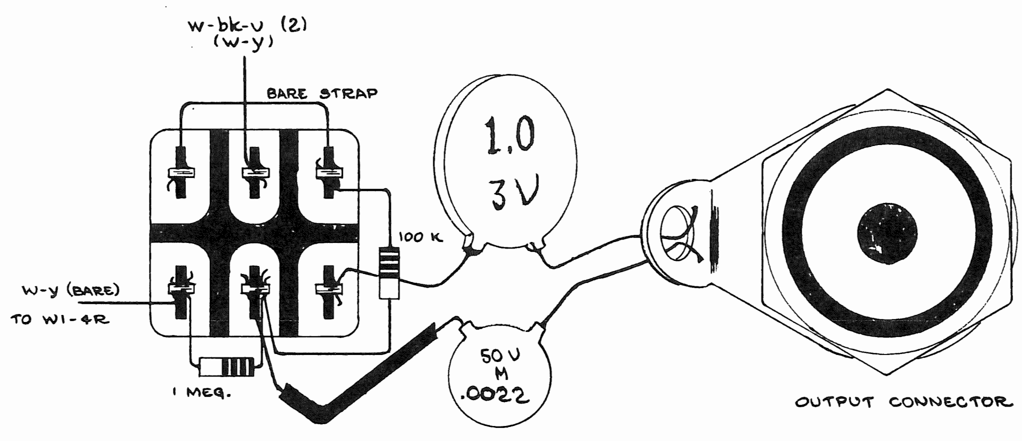


Fig. 3

NOTE: Fig. 3 shows the wiring for amplifier "A". Amplifier "B" wiring differs only where shown in parenthesis.

THIS COMPLETES THE INSTALLATION.

( ) Check wiring for accuracy.

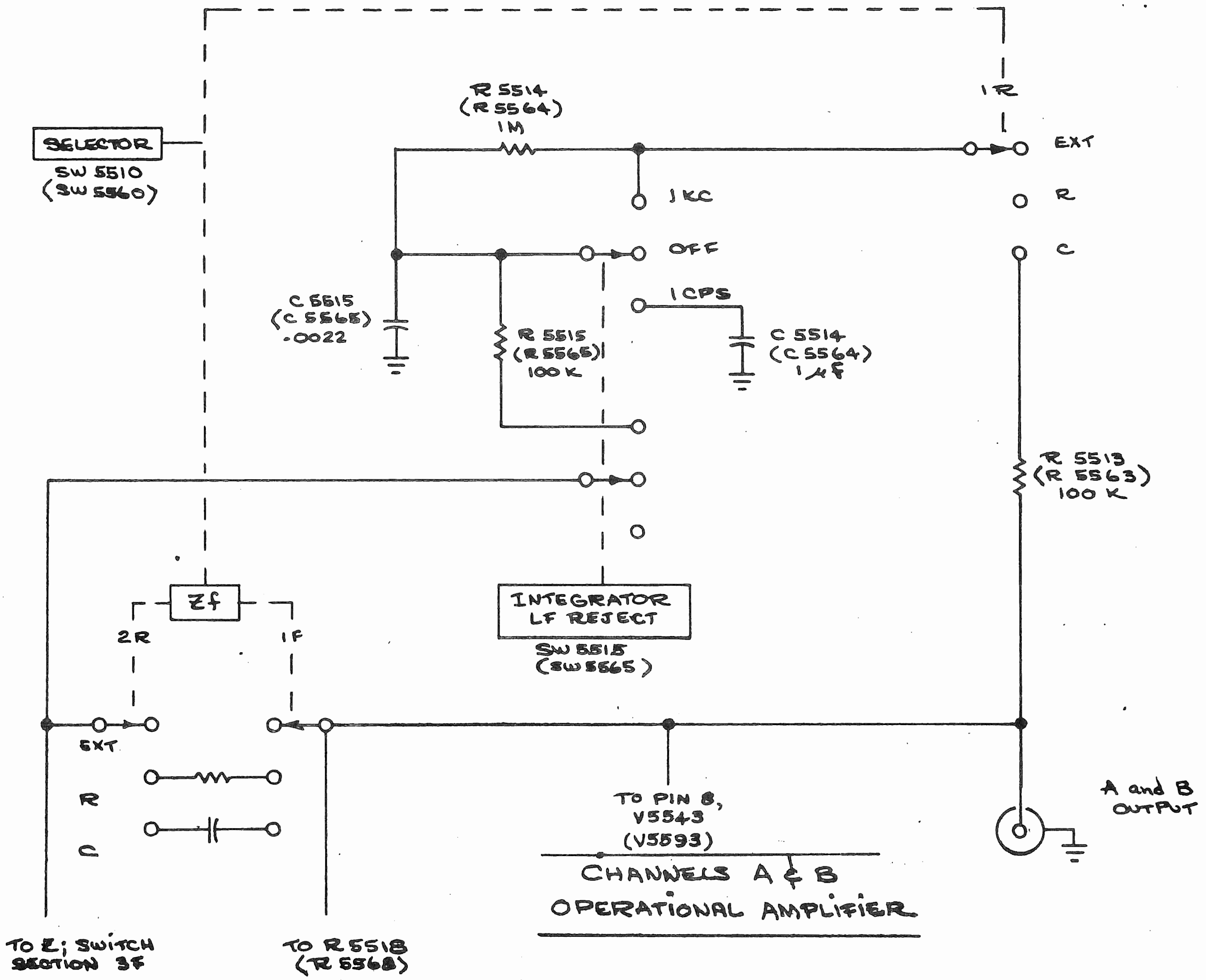
( ) Adjust the SELECTOR switch compensating capacitors as described in your Instruction Manual.

( ) If you are using the Preliminary Instruction Manual, fasten in it the partial schematic (the upper circuit symbols refer to amplifier "A", the lower symbols refer to amplifier "B").

( ) If you have received your permanent Manual, make a notation in it to refer to the amplifier schematics for s/n 319-up. Also, add the following Parts List information:

SW5510 260-0431-00 (unwired) 262-0518-00 (wired)
Rotary SELECTOR A

CH:cet



CHANNELS A & B
OPERATIONAL AMPLIFIER

PARTS REPLACEMENT KIT

SELECTOR "B" SWITCH

For Tektronix Type "O" Plug-in Units
Serial numbers 101-318, with exceptions

050-0075-00

DESCRIPTION

The new operational amplifier "B" SELECTOR switch 262-0519-00 replaces 262-0425-00, previously used.

The new switch is changed to accommodate a modification which improves the accuracy of the output amplitude with a Z_f of 10 pF and the INTEGRATOR LF REJECT off.

NOTE: If the serial number of your instrument is above those listed, or if this kit has been installed, disregard the instructions as P/N 262-0519-00 is a direct replacement.

Publication:
Instructions for 050-0075-00
March 1966

Supersedes:
September 1963

PARTS LIST

Qty.	Part Number	Description			
(1 ea)		Assembly, SELECTOR switch (modified 262-0519-00), consisting of:			
1 ea	210-0012-00	Lockwasher, int. 3/8 x 1/2			
2 ea	210-0406-00	Nut, hex, 4-40 x 3/16			
1 ea	210-0413-00	Nut, hex, 3/8-32 x 1/2			
1 ea	210-0840-00	Washer, steel, 0.390 x 9/16 x 0.020			
1 ea	260-0430-00	Switch, raw			
2 ea	281-0031-00	Capacitor, cer, var,	3-12 pF		
2 ea	281-0048-00	Capacitor, cer, var,	5-25 pF		
1 ea	281-0504-00	Capacitor, cer,	10 pF	500V	±1 pF
2 ea	281-0528-00	Capacitor, cer,	82 pF	500V	±8.2 pF
1 ea	281-0537-00	Capacitor, cer,	0.68 pF	500V	±0.136 pF
5 ea	281-0538-00	Capacitor, cer,	1 pF	500V	±0.2 pF
1 ea	301-0244-00	Resistor, comp,	240 k	1/2W	5%
1 ea	301-0514-00	Resistor, comp,	510 k	1/2W	5%
2 ea	309-0100-00	Resistor, prec,	10 k	1/2W	1%
2 ea	309-0140-00	Resistor, prec,	500 k	1/2W	1%
2 ea	309-0148-00	Resistor, prec,	1 M	1/2W	1%
2 ea	309-0260-00	Resistor, prec,	100 k	1/2W	1%
2 ea	309-0444-00	Resistor, prec,	200 k	1/2W	1%
1 ea	316-0104-00	Resistor, comp,	100 k	1/4W	10%
1 ea	348-0031-00	Grommet, poly, 1/4			
1 ea	406-0720-00	Bracket, switch end			
1 ea	(162-0504-00)	Tubing, plastic, #20	1-1/4 in.	black	
1 ea	(175-0522-00)	Wire, #22 solid,	6 in.	white-black-violet	
1 ea	(175-0522-00)	Wire, #22 solid,	2-1/2 in.	white-yellow	
1 ea	(176-0005-00)	Wire, #22 solid,	1-1/2 in.	bare	

INSTRUCTIONS

NOTE: The following method is used to identify the SELECTOR switch terminals:

The wafers are numbered from front to rear.

The contact positions are numbered 1 through 12 relative to the index key as shown in Fig 1.

The contacts have an 'F' or 'R' suffix which denotes that they are on the front or rear of the wafer.

Example: W2-7R (denoted by * on Fig 1) is contact #7 on the rear of wafer 2.

(TYPICAL SWITCH CONFIGURATION)

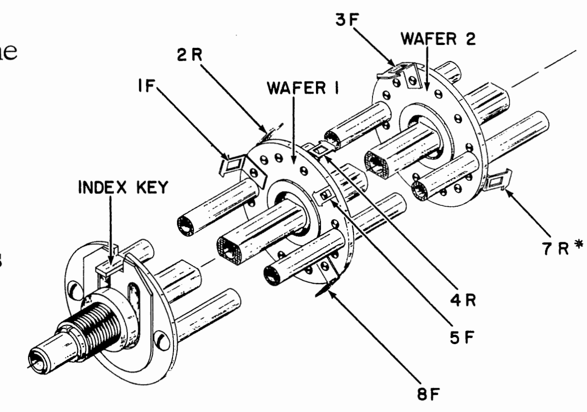


Fig 1

INSTRUCTIONS (cont)

TO REMOVE THE OLD SWITCH:

( ) 1. Remove the screw securing the rear mounting bracket of the amplifier "B" SELECTOR switch.

( ) 2. Remove the knobs from the SELECTOR switch.

( ) Remove the mounting nut.

( ) 3. Unsolder the white-red wire from R5596 going to the "B" OUTPUT coax connector.

( ) Remove the wire from the grommet and dress it out of the way.

( ) 4. Unsolder all necessary wires from the switch and remove the switch from the instrument.

TO INSTALL THE NEW SWITCH:

( ) 5. With a lockwasher on the shaft, place the new switch (from kit) in position in the instrument. DO NOT MOUNT YET.

( ) 6. Solder the wires to the switch terminals indicated below (see Fig. 2).

( ) white-gray to W1-2F
( ) white-blue to W2-2R
( ) white-violet to W2-3R
( ) white-orange to W2-4R
( ) white-red to W2-5R
( ) white-green to W3-2F
( ) white-yellow to W3-3F
( ) solid white to W3-4F
( ) white-black-red to W3-5F
( ) white-black-orange to W4-2R
( ) coaxial to W4-8R
( ) bare wire from OUTPUT coax to W1-11F

( ) 7. Replace the switch mounting nut, making sure the index key is properly seated and that the rear mounting bracket is aligned.

( ) 8. Replace the rear switch bracket mounting screw.

( ) 9. Dress the white-red wire (unsoldered in step 3) through the grommet in the rear switch bracket.

( ) Resolder to R5596.

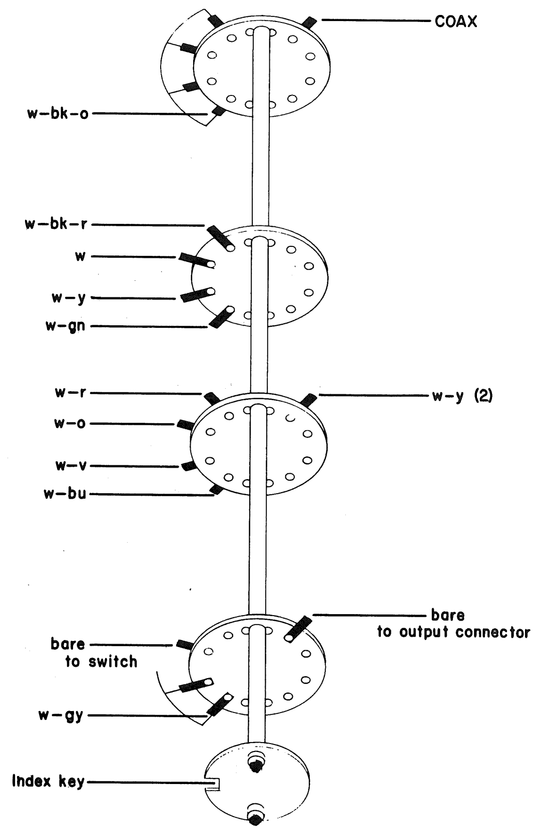


Fig. 2

INSTRUCTIONS (cont)

TO ALIGN THE KNOBS:

( ) 10. Place the larger (Z_f) knob on the shaft and tighten just enough to allow rotation of the switch.

( ) 11. Rotate the switch until the rotor contact of W2 is at position 1.

( ) Loosen the set-screw, align the knob index to indicate R = 0.01 Meg, and tighten it securely.

( ) 12. Place the Z_i knob on its shaft and tighten temporarily.

( ) 13. Rotate the switch until the rotor contacts of W3 and W4 are at position 1.

( ) Loosen the set-screw, align the knob index to indicate R = 0.01 Meg, and tighten it securely.

TO MODIFY AMPLIFIER "B":

14. Unsolder and remove the following components and wires associated with the "B" INTEGRATOR LF REJECT switch:

( ) 100k 1/4W resistor from a switch terminal to the OUTPUT coaxial connector.

( ) 1 Meg 1/4W resistor between two switch terminals.

( ) wire strap from a switch terminal to the "B" SELECTOR switch, W1-4F.

( ) wire strap between two switch terminals.

( ) 15. Unsolder the end of the 0.0022 μF capacitor from the switch terminal.

( ) Resolder it to the terminal indicated in Fig 3.

( ) 16. Clean the excess solder from the switch terminals.

( ) 17. Solder one end of the 2-1/2 in. white-yellow wire (from kit) to the switch terminal indicated in Fig 3 (in parenthesis, upper center in figure).

( ) Solder the other end, along with the white-yellow wire from the Amplifier "B" ±GRID SEL switch, to the SELECTOR switch, W2-11R.

( ) 18. Solder a 1-1/2 in. piece of bare wire (from kit) to the SELECTOR switch, W1-4R. Place the length of insulated tubing (from kit) over the wire and dress the wire to the switch terminal indicated in Fig 3. DO NOT SOLDER YET.

( ) 19. Resolder the components and wires to the switch as indicated in Fig 3.

20. Unsolder and remove the following components and wires associated with the "A" INTEGRATOR LF REJECT switch:

( ) 100k 1/4W resistor from a switch terminal to the OUTPUT coaxial connector.

( ) white-yellow wire from a switch terminal to the "A" SELECTOR switch, W1-4R.

( ) 1 Meg 1/4W resistor between two switch terminals.

( ) wire strap between two switch terminals.

( ) 21. Unsolder the end of the 0.0022 μF capacitor from the switch terminals.

( ) Resolder it to the terminal indicated in Fig 3.

( ) 22. Clean the excess solder from the switch terminals.

23. Locate the white-black-violet wire from the "A" ±GRID SEL switch that is soldered to the "A" SELECTOR switch, W2-8R.

( ) Unsolder this wire from the "A" SELECTOR switch.

( ) 24. Remove the nylon spacing clip between the white-black-violet and the white-yellow wires.

25. Locate the 100k 1/4W resistor on the "A" SELECTOR switch, soldered between W1-5R and W2-8R.

( ) Unsolder the end from W2-8R.

( ) Resolder it to W1-8F.

( ) 26. Dress the white-black-violet wire, unsoldered in step 23, to the "A" LF REJECT switch.

( ) Cut off the excess wire and solder, along with the white-black-violet wire (from kit) to the switch terminal indicated in Fig 3.

( ) 27. Resolder the remainder of the components and wires indicated in Fig 3.

( ) 28. Dress the other end of the white-black-violet wire (step 26) through the grommet in the "A" SELECTOR switch bracket.

( ) Solder to the SELECTOR switch, W2-8R.

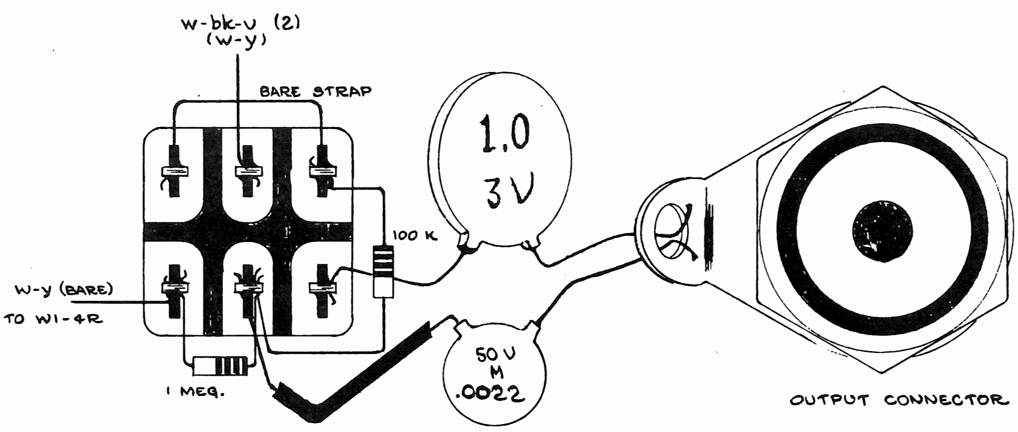


Fig. 3

NOTE: Fig. 3 shows the wiring for amplifier "A". Amplifier "B" wiring differs only where shown in parenthesis.

THIS COMPLETES THE INSTALLATION

( ) Check wiring for accuracy.

( ) Adjust the SELECTOR switch compensating capacitors as described in your Instruction Manual.

( ) If you are using the Preliminary Instruction Manual, fasten in it the partial schematic (the upper circuit symbols refer to amplifier "A", the lower symbols refer to amplifier "B").

( ) If you have received your permanent Manual, make a notation in it to refer to the amplifier schematics for s/n 319-up. Also, add the following Parts List information:

SW5560	260-0430-00 (unwired)	262-0519-00 (wired)
Rotary	SELECTOR B	

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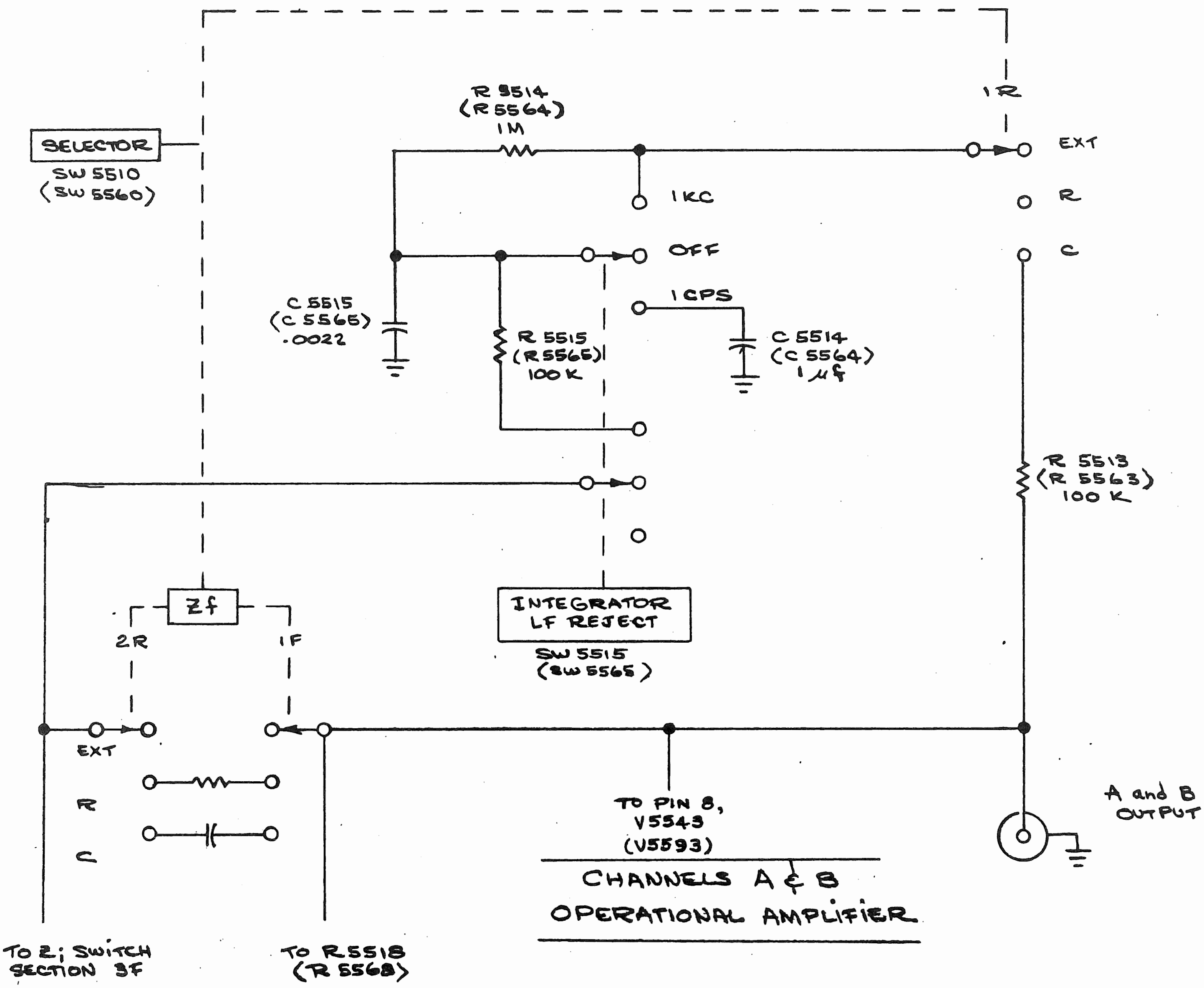

SELECTOR
SW 5510
(SW 5560)
R 5514
(R 5564)
1M
1R
EXT
R
C
1KC
OFF
1CPS
C 5515
(C 5565)
.0022
R 5515
(R 5565)
100 K
C 5514
(C 5564)
1 µf
R 5513
(R 5563)
100 K
Zf
2R
1F
INTEGRATOR
LF REJECT
SW 5515
(SW 5565)
EXT
R
C
TO PIN 8,
V5543
(V5593)
A and B
OUTPUT
CHANNELS A & B
OPERATIONAL AMPLIFIER
TO 2; SWITCH
SECTION 3F
TO R5518
(R 5568)

ELECTRICAL PARTS LIST

Values are fixed unless marked variable.

Bulbs

Ckt. No.	Tektronix Part No.	Description	S/N Range
B5517	Use 150-027	Neon, Type NE-23	101-3279
B5517	150-0030-00	Neon, Type NE-2 V	3280-up
B5567	Use 150-027	Neon, Type NE-23	101-3279
B5567	150-0030-00	Neon, Type NE-2 V	3280-up

Capacitors

Tolerance ±20% unless otherwise indicated.
Tolerance of all electrolytic capacitors are as follows (with exceptions):

3 v— 50 v = —10% +250%
51 v—350 v = —10% +100%
351 v—450 v = —10% + 50%

C5509A †		Selected					
C5509B †		Selected					
C5509C †		Selected					
C5509D †		Selected					
C5510A		1 μf					
C5510B		.1 μf					
C5510C		.01 μf					
C5510D	*291-033	.001 μf					
C5510E		1 μf		Z_i — Z_f Series		±1%	
C5510F		.1 μf					
C5510G		.01 μf					
C5510H		.001 μf					
C5511C †		Selected					
C5511D †		Selected					
C5511E	281-504	10 pf	Cer.		500 v	10%	
C5512A	281-528	82 pf	Cer.		500 v	10%	
C5512B	281-048	5-25 pf	Cer.	Var.			
C5512C	281-031	3-12 pf	Cer.	Var.			
C5512E	281-528	82 pf	Cer.		500 v	10%	
C5512F	281-048	5-25 pf	Cer.	Var.			
C5512G	281-031	3-12 pf	Cer.	Var.			
C5514	283-017	1 μf	Disc Type		3 v		
C5515	283-028	.0022 μf	Disc Type		50 v		
C5528	283-057	.1 μf	Disc Type		200 v		
C5529	283-002	.01 μf	Disc Type		500 v		
C5533	283-002	.01 μf	Disc Type		500 v		
C5535	283-003	.01 μf	Disc Type		150 v		X319-up
C5538	283-004	.02 μf	Disc Type		150 v		X814-up
C5543	283-002	.01 μf	Disc Type		500 v		
C5544	283-000	.001 μf	Disc Type		500 v		
C5559A †		Selected					
C5559B †		Selected					
C5559C †		Selected					
C5559D †		Selected					

† These capacitors are installed when necessary for optimum performance. Nominal values are from 0 to 2 pf.

Capacitors *(continued)*

Ckt. No.	Tektronix Part No.	Description					S/N Range
C5560A	*291-033	1 μf		Z_i — Z_f Series		±1%	
C5560B	*291-033	.1 μf		Z_i — Z_f Series		±1%	
C5560C	*291-033	.01 μf		Z_i — Z_f Series		±1%	
C5560D	*291-033	.001 μf		Z_i — Z_f Series		±1%	
C5560E	*291-033	1 μf		Z_i — Z_f Series		±1%	
C5560F	*291-033	.1 μf		Z_i — Z_f Series		±1%	
C5560G	*291-033	.01 μf		Z_i — Z_f Series		±1%	
C5560H	*291-033	.001 μf		Z_i — Z_f Series		±1%	
C5561C†		Selected					
C5561D†		Selected					
C5561E	281-504	10 pf	Cer.		500 v	10%	
C5562A	281-528	82 pf	Cer.		500 v	10%	
C5562B	281-048	5-25 pf	Cer.	Var.			
C5562C	281-031	3-12 pf	Cer.	Var.			
C5562E	281-528	82 pf	Cer.		500 v	10%	
C5562F	281-048	5-25 pf	Cer.	Var.			
C5562G	281-031	3-12 pf	Cer.	Var.			
C5564	283-017	1 μf	Disc Type		3 v		
C5565	283-028	.0022 μf	Disc Type		50 v		
C5578	283-057	.1 μf	Disc Type		200 v		
C5579	283-002	.01 μf	Disc Type		500 v		
C5583	283-002	.01 μf	Disc Type		500 v		
C5585	283-003	.01 μf	Disc Type		150 v		X319-up
C5588	283-004	.02 μf	Disc Type		150 v		X814-up
C5593	283-002	.01 μf	Disc Type		500 v		
C5594	283-000	.001 μf	Disc Type		500 v		
C6501	Use *285-0672-00	.1 μf	PTM		600 v		
C6507	281-543	270 pf	Cer.		500 v	10%	
C6508	281-007	3-12 pf	Cer.	Var.			
C6508C	281-010	4.5-25 pf	Cer.	Var.			
C6508D	281-501	4.7 pf	Cer.				
C6509B	281-010	4.5-25 pf	Cer.	Var.			
C6509C	281-007	3-12 pf	Cer.	Var.			
C6510B	281-010	4.5-25 pf	Cer.	Var.			
C6510C	281-010	4.5-25 pf	Cer.	Var.			
C6510E	283-508	150 pf	Mica		500 v	10%	
C6513A	281-505	12 pf	Cer.		500 v	10%	
C6513B	281-010	4.5-25 pf	Cer.	Var.			
C6513C	281-005	1.5-7 pf	Cer.	Var.			
C6513E	283-543	250 pf	Mica		500 v	5%	
C6518	283-002	.01 μf	Disc Type		500 v		
C6519	283-002	.01 μf	Disc Type		500 v		
C6521	Use 281-036	3-12 pf	Cer.	Var.			
C6523	283-023	.1 μf	Disc Type		10 v		X105-2949X
C6525	281-0509-00	15 pf	Cer.		500 v	10%	X2950-up
C6539	290-164	1 μf	EMT		150 v		
C6541	Use 281-036	3-12 pf	Cer.	Var.			
C6545	281-0509-00	15 pf	Cer.		500 v	10%	X2950-up

†These capacitors are installed when necessary for optimum performance. Nominal values are from 0 to 2 pf.

Capacitors *(Cont'd)*

Ckt. No.	Tektronix Part No.	Description					S/N Range
C6546	281-0063-00	9-35 pf	Cer.	Var.			X2950-up
C6562	283-0065-00	.001 μf	Disc Type		100 v	5%	X2950-up
C6564	281-027	.7-3 pf	Tub.	Var.			
C6565	283-028	.0022 μf	Disc Type		50 v		101-2949X
C6572	283-0065-00	.001 μf	Disc Type		100 v	5%	X2950-up
C6574	281-027	.7-3 pf	Tub.	Var.			
C6576	283-002	.01 μf	Disc Type		500 v		
C6579	283-004	.02 μf	Disc Type		150 v		101-2949X
C6582	283-002	.01 μf	Disc Type		500 v		
C6584	283-002	.01 μf	Disc Type		500 v		
C6586	283-002	.01 μf	Disc Type		500 v		
C6589	Use 283-057	.1 μf	Disc Type		200 v		

Diodes

Ckt. No.	Tektronix Part No.	Description				S/N Range
D5528	152-087	Zener IN3044B		100 v		101-813X
D5529	152-087	Zener IN3044B		100 v		101-813
D5529	152-055	Zener 1N962B	.4 w	11 v	5%	814-up
D5578	152-087	Zener IN3044B		100 v		101-813X
D5579	152-087	Zener IN3044B		100 v		101-813
D5579	152-055	Zener 1N962B	.4 w	11 v	5%	814-up
D6576	152-016	Zener RT6				101-2949
D6576	152-0076-00	Zener IN4372	.4 w	3 v	10%	2950-up

Inductors

Ckt. No.	Tektronix Part No.	Description			S/N Range
L5520	276-532	Bead, Ferrite			X2651-3429
L5520	276-0507-00	Core, Ferramic Suppressor			3430-up
L5530	276-532	Bead, Ferrite			X2651-3429
L5530	276-0507-00	Core, Ferramic Suppressor			3430-up
L5570	276-532	Bead, Ferrite			X2651-3429
L5570	276-0507-00	Core, Ferramic Suppressor			3430-up
L5580	276-532	Bead, Ferrite			X2651-3429
L5580	276-0507-00	Core, Ferramic Suppressor			3430-up
L6524	*114-149	.2-.325 μh	Var.	Core 276-506	101-2949
L6524	*114-0043-00	.5-1 μh	Var.	Core 276-0506-00	2950-up
L6544	*114-149	.2-.325 μh	Var.	Core 276-506	101-2949
L6544	*114-0043-00	.5-1 μh	Var.	Core 276-0506-00	2950-up
L6564	*114-043	.5-1 μh	Var.	Core 276-506	101-2949X
L6565	276-528	Core, Ferrite			X2429-2949X
L6574	*114-043	.5-1 μh	Var.	Core 276-506	101-2949X
L6575	276-528	Core, Ferrite			X2429-2949X

Resistors

Ckt. No.	Tektronix Part No.	Description					S/N Range
Resistors are fixed, composition, ±10% unless otherwise indicated.							
R5509A	309-148	1 meg	½ w		Prec.	1%	101-3519
R5509A	323-0481-01	1 meg	½ w		Prec.	½%	3520-up
R5509B	309-140	500 k	½ w		Prec.	1%	101-3799
R5509B	323-0740-01	500 k	½ w		Prec.	½%	3800-up
R5509C	309-444	200 k	½ w		Prec.	1%	101-3799
R5509C	323-0414-01	200 k	½ w		Prec.	½%	3800-up
R5509D	309-260	100 k	½ w		Prec.	1%	101-3519
R5509D	323-0385-01	100 k	½ w		Prec.	½%	3520-up
R5509E	309-100	10 k	½ w		Prec.	1%	101-3519
R5509E	323-0289-01	10 k	½ w		Prec.	½%	3520-up
R5509F	301-514	510 k	½ w			5%	
R5509G	301-244	240 k	½ w			5%	
R5511A	309-148	1 meg	½ w		Prec.	1%	101-3519
R5511A	323-0481-01	1 meg	½ w		Prec.	½%	3520-up
R5511B	309-140	500 k	½ w		Prec.	1%	101-3799
R5511B	323-0740-01	500 k	½ w		Prec.	½%	3800-up
R5511C	309-444	200 k	½ w		Prec.	1%	101-3799
R5511C	323-0414-01	200 k	½ w		Prec.	½%	3800-up
R5511D	309-260	100 k	½ w		Prec.	1%	101-3519
R5511D	323-0385-01	100 k	½ w		Prec.	½%	3520-up
R5511E	309-100	10 k	½ w		Prec.	1%	101-3519
R5511E	323-0289-01	10 k	½ w		Prec.	½%	3520-up
R5513	316-104	100 k	¼ w				
R5514	316-105	1 meg	¼ w				
R5515	316-104	100 k	¼ w				
R5517	316-104	100 k	¼ w				
R5518	316-105	1 meg	¼ w				
R5519	316-103	10 k	¼ w				
R5520	316-470	47 Ω	¼ w				101-2650X
R5521	302-473	47 k	½ w				101-813
R5521	309-354	45 k	½ w		Prec.	1%	814-up
R5522	311-164	50 k		Var.		OUTPUT DC LEVEL	
R5523	310-070	33 k	1 w		Prec.	1%	101-813
R5523	305-393	39 k	2 w			5%	814-3239
R5523	305-0273-00	27 k	2 w			5%	3240-up
R5524	302-102	1 k	½ w				
R5525	309-354	45 k	½ w		Prec.	1%	101-813
R5525	323-347	40.2 k	½ w		Prec.	1%	814-up
R5526	305-223	22 k	2 w			5%	101-813
R5526	308-241	22 k	7 w		WW	1%	814-up
R5529	Use 306-473	47 k	2 w				
R5530	316-470	47 Ω	¼ w				101-2650X
R5531	302-473	47 k	½ w				101-813
R5531	309-354	45 k	½ w		Prec.	1%	814-up
R5532	311-153	10 k		Var.		DC LEVEL RANGE	

Resistors *(continued)*

Ckt. No.	Tektronix Part No.	Description					S/N Range
R5533	302-102	1 k	½ w				
R5534	305-123	12 k	2 w			5%	X155-up
R5535	311-171	5 k		Var.			X155-up
R5538	315-102	1 k	¼ w			5%	X814-up
R5539	304-184	180 k	1 w				X814-3239
R5539	305-0823-00	82 k	2 w			5%	3240-up
R5540	316-104	100 k	¼ w				
R5541	316-470	47 Ω	¼ w				
R5542	316-101	110 Ω	¼ w				
R5543	302-221	220 Ω	½ w				
R5544	315-100	10 Ω	¼ w			5%	
R5546	308-069	12 k	8 w		WW	5%	
R5547	316-154	150 k	¼ w				
R5548	311-068	500 k		Var.		OPEN LOOP GAIN	
R5559A	309-148	1 meg	½ w		Prec.	1%	101-3519
R5559A	323-0481-01	1 meg	½ w		Prec.	½%	3520-up
R5559B	309-140	500 k	½ w		Prec.	1%	101-3799
R5559B	323-0740-01	500 k	½ w		Prec.	½%	3800-up
R5559C	309-444	200 k	½ w		Prec.	1%	101-3799
R5559C	323-0414-01	200 k	½ w		Prec.	½%	3800-up
R5559D	309-260	100 k	½ w		Prec.	1%	101-3519
R5559D	323-0385-01	100 k	½ w		Prec.	½%	3520-up
R5559E	309-100	10 k	½ w		Prec.	1%	101-3519
R5559E	323-0289-01	10 k	½w		Prec.	½%	3520-up
R5559F	301-514	510 k	½ w			5%	
R5559G	301-244	240 k	½ w			5%	
R5561A	309-148	1 meg	½ w		Prec.	1%	101-3519
R5561A	323-0481-01	1 meg	½w		Prec.	½%	3520-up
R5561B	309-140	500 k	½ w		Prec.	1%	101-3799
R5561B	323-0740-01	500 k	½ w		Prec.	½%	3800-up
R5561C	309-444	200 k	½ w		Prec.	1%	101-3799
R5561C	323-0414-01	200 k	½ w		Prec.	½%	3800-up
R5561D	309-260	100 k	½ w		Prec.	1%	101-3519
R5561D	323-0385-01	100 k	½ w		Prec.	½%	3520-up
R5561E	309-100	10 k	½ w		Prec.	1%	101-3519
R5561E	323-0289-01	10 k	½ w		Prec.	½%	3520-up
R5563	316-104	100 k	¼ w				
R5564	316-105	1 meg	¼ w				
R5565	316-104	100 k	¼ w				
R5567	316-104	100 k	¼ w				
R5568	316-105	1 meg	¼ w				
R5569	316-103	10 k	¼ w				
R5570	316-470	47 Ω	¼ w				101-2650X
R5571	302-473	47 k	½ w				101-813
R5571	309-354	45 k	½ w		Prec.	1%	814-up

Resistors *(continued)*

Ckt. No.	Tektronix Part No.	Description					S/N Range
R5572	311-164	50 k		Var.		OUTPUT DC LEVEL	
R5573	310-070	33 k	1 w		Prec.	1%	101-813
R5573	305-393	39 k	2 w			5%	814-3239
R5573	305-0273-00	27 k	2 w			5%	3240-up
R5574	302-102	1 k	½ w				
R5575	309-354	45 k	½ w		Prec.	1%	101-813
R5575	323-347	40.2 k	½ w		Prec.	1%	814-up
R5576	305-223	22 k	2 w			5%	101-813
R5576	308-241	22 k	7 w		WW	3%	814-up
R5579	Use 306-473	47 k	2 w				
R5580	316-470	47 Ω	¼ w				101-2650X
R5581	302-473	47 k	½ w				101-813
R5581	309-354	45 k	½ w		Prec.	1%	814-up
R5582	311-153	10 k		Var.		DC LEVEL RANGE	
R5583	302-102	1 k	½ w				
R5584	305-123	12 k	2 w			5%	X155-up
R5585	311-171	5 k		Var.			X155-up
R5588	315-102	1 k	¼ w			5%	X814-up
R5589	304-184	180 k	1 w				X814-3239
R5589	305-0823-00	82 k	2 w			5%	3240-up
R5590	316-104	100 k	¼ w				
R5591	316-470	47 Ω	¼ w				
R5592	316-101	100 Ω	¼ w				
R5593	302-221	220 Ω	½ w				
R5594	315-100	100 Ω	¼ w			5%	
R5596	308-069	12 k	8 w		WW	5%	
R5597	316-154	150 k	¼ w				
R5598	311-068	500 k		Var.		OPEN LOOP GAIN	
R6502	302-220	22 Ω	½ w				
R6507	302-220	22 Ω	½ w				
R6508C	309-140	500 k	½ w		Prec.	1%	
R6508E	309-148	1 meg	½ w		Prec.	1%	
R6509C	309-141	750 k	½ w		Prec.	1%	
R6509E	309-139	333 k	½ w		Prec.	1%	
R6510C	309-142	900 k	½ w		Prec.	1%	
R6510E	309-138	111 k	½ w		Prec.	1%	
R6513C	309-145	990 k	½ w		Prec.	1%	
R6513E	309-135	10.1 k	½ w		Prec.	1%	
R6516	302-270	27 Ω	½ w				
R6517	309-148	1 meg	½ w		Prec.	1%	
R6518	316-104	100 k	¼ w				
R6519	316-334	330 k	¼ w				
R6520	*308-141	1 Ω	½ w		WW	5%	X105-2949X
R6521	316-470	47 Ω	¼ w				
R6522	316-221	220 Ω	¼ w				X105-2949

Resistors *(continued)*

Ckt. No.	Tektronix Part No.	Description					S/N Range
R6522	301-0331-00	330 Ω	½ w			5%	2950-up
R6523	316-560	56 Ω	¼ w				101-104
R6523	*308-141	1 Ω	½ w		WW	5%	105-2949X
R6524	318-083	200 Ω	⅛ w				
R6530	*311-259	710 Ω		Var.	WW	VARIABLE VOLTS/CM	
R6531	303-153	15 k	1 w			5%	
R6532	303-153	15 k	1 w			5%	
R6533	311-171	5 k		Var.		DC BAL.	
R6536	Use 311-392	10 k	2 w	Var.	WW	GAIN ADJ.	
R6538	306-103	10 k	2 w				
R6539	302-562	5.6 k	½ w				
R6541	316-470	47 Ω	¼ w				
R6544	318-083	200 Ω	⅛ w		Prec.	1%	
R6546	311-0337-00	20 k		Var.			X2950-up
R6550	311-028	2x100 k	2 w	Var.		POSITION	
R6551	302-104	100 k	½ w				
R6552	302-104	100 k	½ w				
R6556	302-104	100 k	½ w				
R6557	311-301	100 k	2 w	Var.		VERT. POS. RANGE	101-154
R6557	311-088	100 k	0.2 w	Var.			155-up
R6558	302-104	100 k	½ w				101-2949X
R6560	303-0103-00	10 k	1 w			5%	X2950-up
R6562	315-0221-00	220 Ω	¼ w			5%	X2950-up
R6563	302-392	3.9 k	½ w				101-2949X
R6564	318-083	200 Ω	⅛ w		Prec.	1%	101-2949X
R6565	307-023	4.7 Ω	½ w				101-2949X
R6567	321-0097-00	100 Ω	⅛ w		Prec.	1%	X2950-up
R6568	319-050	119 Ω	¼ w		Prec.	1%	101-2949
R6568	321-0059-00	40.2 Ω	¼ w		Prec.	1%	2950-up
R6569	301-912	9.1 k	½ w			5%	101-2949X
R6570	303-0103-00	10 k	1 w			5%	X2950-up
R6572	315-0221-00	220 Ω	¼ w			5%	X2950-up
R6573	302-392	3.9 k	½ w				101-2949X
R6574	318-083	200 Ω	⅛ w		Prec.	1%	101-2949X
R6575	307-023	4.7 Ω	½ w				101-2949X
R6576	306-392	3.9 k	2 w				101-2949X
R6577	321-0097-00	100 Ω	⅛ w		Prec.	1%	X2950-up
R6578	315-0822-00	8.2 k	¼ w			5%	X2950-up
R6579	301-912	9.1 k	½ w			5%	101-2949X
R6580	304-152	1.5 k	1 w				
R6582	302-101	100 Ω	½ w				
R6586	302-101	100 Ω	½ w				

Switches

	Unwired	Wired			
SW5510A&B	*260-430	Use *050-074	Rotary	SELECTOR A	101-318
SW5510A&B	*260-430	*262-518	Rotary	SELECTOR A	319-up
SW5515	260-396		Toggle	INTEGRATOR LF REJECT A	
SW5517	260-398		Toggle	GRID SEL	
SW5520	260-397		Toggle	DC LEVEL ADJ A	

Switches *(continued)*

Ckt. No.	Tektronix Part No. Unwired	Wired		Description	S/N Range
SW5560A&B	*260-431	Use *050-075	Rotary	SELECTOR B	101-318
SW5560A&A	*260-431	*262-519	Rotary	SELECTOR B	319-up
SW5565	260-396		Toggle	INTEGRATOR LF REJECT B	
SW5567	260-398		Toggle	GRID SEL	
SW5570	260-397		Toggle	DC LEVEL ADJ B	
SW6500	*260-432	Use *262-634	Rotary	VERTICAL DISPLAY	
SW6502	260-248		Push Button	ZERO CHECK	
SW6510	*260-428	*262-426	Rotary	VOLTS/CM	

Transistors

Q5523	*151-0096-00	Selected from 2N1893	X814-3879
Q5523	151-0150-00	Silicon 2N3440	3880-up
Q5573	*151-0096-00	Selected from 2N1893	X814-3879
Q5573	151-0150-00	Silicon 2N3440	3880-up
Q6564	151-067	2N1143	101-2949
Q6564	*151-0120-00	Selected from 2N2475	2950-up
Q6574	151-067	2N1143	101-2949
Q6574	*151-0120-00	Selected from 2N2475	2950-up

Electron Tubes

V5524, V5534 †	Use *157-071	8426	
V5529	154-370	ZZ1000	X814-up
V5539	154-370	ZZ1000	X814-up
V5543	154-187	6DJ8	
V5574, V5584 †	Use *157-071	8426	
V5579	154-370	ZZ1000	X814-up
V5589	154-370	ZZ1000	X814-up
V5593	154-187	6DJ8	
V6524, V6544 †	Use *157-077	8426	

† Selected pair. Furnished as a unit.

Ⓔ

TYPE O TENT SN 3880

INSTRUCTION MANUAL

PARTS LIST AND SCHEMATIC ADDENDUM

CHANGE TO:

Q5523	151-0150-00	Silicon	2N3440
Q5573	151-0150-00	Silicon	2N3440

This insert is placed in its appropriate position in your Product Reference Book and printed on colored paper to expedite retrieval. In a standard manual, it will be filed at the back of the manual.

M11,567/867

FACTORY TEST LIMITS

QUALIFICATION

Factory test limits are qualified by the conditions specified in the main body of the factory calibration procedure. Instruments may not meet factory test limits if calibration or checkout methods and test equipment differ substantially from those in the factory procedure.

These limits usually are tighter than advertised performance requirements, thus helping to insure the instrument will meet or be within advertised performance requirements after shipment and during subsequent recalibrations. Instruments that have left the factory may not meet factory test limits but should meet catalog or instruction manual performance requirements.

OUTPUT DC LEVEL +67.5V ±2.5V

DC BALANCE

control must be centered ±90°

VERTICAL POSITION RANGE

control must be centered ±90°

GRID CURRENT AND MICROPHONICS

Grid current	2mm, max.
Microphonics	2mm, max.; no ringing type

DC SHIFT 1mm, max.

GAIN

GAIN ADJ Range	±10%, min.
Set GAIN ADJ	exactly 2cm.

VOLTS/CM ACCURACY

Accuracy	±2%, max.
Variable VOLTS/CM range	2.5 to 1, min.

PREAMPLIFIER SPIKES

Dress leads minimum spikes

TEK 0 PRB 1-4-66

INPUT AND NEUTRALIZING CAPACITORS

Adjust input and neutralizing capacitors for best flat top and square-wave

VOLTS/CM COMPENSATIONS

Flat top deviation ±.5 mm, max

HIGH FREQUENCY COMPENSATION

adjust for best square corner

RISETIME

13.5 ns, max
ringing: 1%, max

FREQUENCY RESPONSE

Check response dc to 26 MHz at -3 dB

OUTPUT DC LEVEL

OUTPUT DC LEVEL ADJ	adjusted to electrical center
DC LEVEL Range	adjusted to 0 volts

OPEN LOOP GAIN

Range Adjust to 3000

OUTPUT CONNECTORS, VOLTAGE AND CURRENT

Check output voltage	±50 V, min
Check output current	5 mA, min.

GRID CURRENT

Grid current, -grid	.3 nanoamp, max
Grid current, +grid	.15 nanoamp, max

NOISE

Check noise .5 cm, max

Zi AND Zf VALUES

Check Zi and Zf resistors must match within ±1%

Check Zi and Zf capacitors must match within ±1%

Test Limits - continued

Zi - Zf EQUALIZATION

Select capacitors	less than 5% roll-off, hook, or overshoot
Unity gain bandwidth	750 kHz, min

GAIN - BANDWIDTH PRODUCT

Check gain-bandwidth product	15 MHz, min

COMPENSATED BANDWIDTH

Check bandwidth	-3 dB at 10 MHz

OUTPUT IMPEDANCE

$\leq 35\ \Omega$

CROSSTALK

A to B amplifier	500:1, min
B to A amplifier	500:1, min

MISCELLANEOUS CALIBRATION INFORMATION

Leonard Bell to Geoff Gass

9-6-63

September 25 I am giving a lecture to our local "troops" on the O Unit. There are a number of questions we would like answers for before the 25th if possible. Since we haven't seen very many O Units here at the Union office, some of the items may be "old hat" by now.

1. On O Unit, S/N 690, I find that the chassis markings for the input caps C-6541 and C-6521 seem to be reversed. Using the Factory Cal. Procedure, dated Jan. 26, 1962, step No. 14d says to tweek C-6541 for Ext DC + and C-6521 for Ext DC -. The cap marked C-6541 on this particular instrument tweeked the Ext DC - and vice versa.

2. In connection with No. 1 above, I checked the preamp schematic in two manuals. One manual has two operational amp schematics, one for S/N 101 thru 318 and one for S/N 318 and up. The other manual has operational amp schematics for S/N 101 thru 813 and 813 and up. In both manuals the preamplifier schematics appear to have errors in the vertical display switch wafer 3F and 4R area. Both schematics are marked MRH 2-19-62, but one has A at bottom center and the other A_1. In neither schematic can I trace a signal from the junction of R6517 and R6518 to either input grid. The way they are drawn it appears that both input grids are shorted together in all input positions and that the input is left floating.

3. While adjusting the attenuator compensations, I found a "stacking" problem on the 10 V/cm and 20 V/cm positions. There was a leading spike on both these positions even though the X100, X2, and X4, attenuators by themselves looked good. I tried "fudging" back and forth but couldn't come up with a satisfactory compromise that would make the compensations look good when used singly and stacked. Is this normal or is there a cure that you know of?

   P.S. The problem was worst in the 20 V/cm position.

4. Cal Procedure step No. 21c has a trap in the wording. First sentence should read "Remove Calibrator signal from EXT INPUT and apply it *through the special AC coupler* to A INPUT." I moved only the patch cord and thus got into a lot of trouble until Bob Keyes and I figured out that the AC coupler has to be moved also.

5. It seems that there must be a later Cal Procedure than the one I have--is there? If so, could I get nine copies?

6. What are the latest specs for grid current? Noise specs?

7. Do you have any pictures of what the best 10 kc square wave looks like when it leaves production? I am referring to lead dressing for optimum square wave.

8. Could you send eight copies of your O Unit LF Reject Circuits write-up? If not, then perhaps at least one?

9. Do you have an ABC type of explanation for the operation of R6568, 6569, 6579, and C6579, i.e. how do they compensate for thermal changes?

Any help you can give will certainly be greatly appreciated.

Here's what I can dig up for you:

1. It's the FCP, not the silk-screening that's in error. If you'll look in the manual, you'll see that the photos on page 7-4, either in edition A or revised edition A1, show C6521 as the marking for the CCW (+Input) capacitor of the pair. Current production is the same. This is consistent with the schematic and the cal procedure in the manual.

2. The preamp input switching was finally straightened out in the schematic edition A_2, which is contained in the revised manual (coded 563 or higher on the title page). Wud be a good idea if you ordered current manuals for your class.

3. Attenuator Tweeking--10 v/cm, 20 v/cm: Ralph Smith in Plant 1 Test says about a millimeter of overshoot or a little less is normal for the stacked attenuator positions at the high end. If you have about this much (≈2%), rest easy. If more, you may need new input jugs (157-077).

4. This part of the FCP *is* quite confusing. The Manual procedure assumes you don't have a special coupler, but gets more complicated.

   There's really no point in putting the AC coupler on the Ext input that I can see. Just use +AC. But anyway, you're trying to prove in this step that the O can hack ±50 v and (part d) ±5 ma output, which you accomplish by changing the unidirectional 100 v cal signal to a ±50 v signal. The FCP should make this more clear, I agree. Cal symmetry should be pretty close for this test.

5. So far as I know, these corrections have never been made--the 1-26-62 version of the FCP is the latest available.

6. Latest specs for grid current are:

(Us)	300 pA on the -grid 150 pA on the +grid
(Customer)	Less than 500 picoamperes, *adjustable* to the above spex.

   In the FCP (see also Manual, page 7-7) it should mention that the -grid grid current should deflect the spot downward with the input selector at +A (+B) and the grid current at the +grid should deflect the spot upward.

   For noise, same as always: set 'er up for X100 gain and don't get more than 25 mv p-p output noise (250 μv input noise test spec). Customer type number is 500 μv. The other number in the manual (3 mv) is the *additional* output noise when Z_f is 1 meg. This is to account for the fact that when $Z_f = Z_i = 1\,M$, the equivalent input noise is not 1/2 mv, but more like 2 or 3 mv because the feed-back isn't compensated, and all of the output noise doesn't get back to the input. With $Z_f = 10\,k$, you don't see this noise.

7. Best 10 kc square wave? That's easy--on 10 k and 10 k!

   Seriously, the enclosed photos from S/N 1522 (rotation instrument) show a big 5% on 1 meg/1 meg and in two different ways. Channel A has a spike and a slow tilt, Channel B just has lots of spike. There's an unfortunate spike on the 100 k position of Ch A; this could be fixed.

   The third waveform photo shows (upper half, middle trace and lower half, middle trace) what just *micro* tweaking of the C5509 and C5559 (1 pf each) lead dress can do. The final trim of compensation is normally done by pushing these leads around.

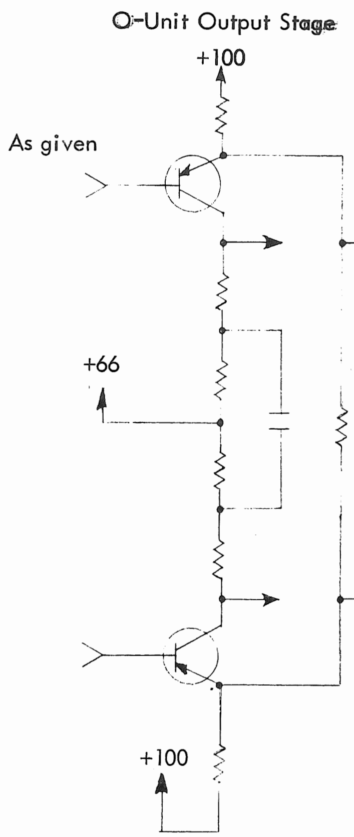
   The upper and lower traces of each set of three shows what fiddling with the LF Reject switch will do. In the 1 cps position you roll it off; in the 1 kc position you spike it up. Channel A did not quite make bandwidth in the 1 cps reject position, but this is *not* technically a material consideration--if you make bandwidth in the "off" position. The effect of the LF Reject switch is simply a miniscule change in the millipicofarads between the -grid and the input and output circuits. Trying to control this more closely than we do, without lousing up the 10 pf integrator or something else, would be pretty tough. So we live with it.

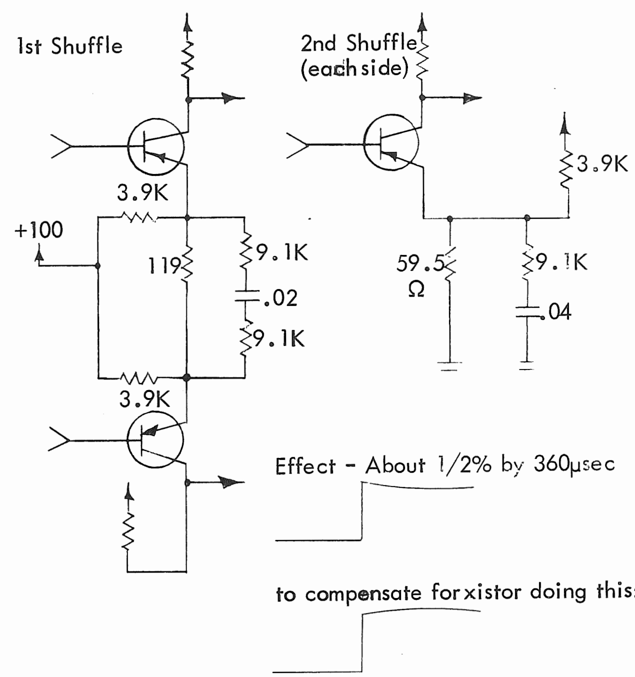
   The actual compensation of the 1 M/1 M amplifier is not too stable. The difference between the first two and the third set of waveforms did not amount to more than perhaps 1/64" shift of an inch of wire. So when you ask for "best", I almost have to ask, how many minutes and feet away from the calibration bench? A few seconds with a spudger will convince you, I think, Leonard.

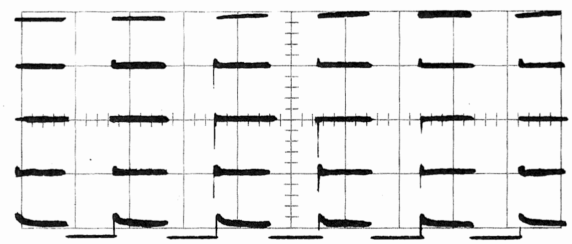
Certainly if an O came out of a repair center looking like the third photo in the 1 M/1 M position, that would be as much as anyone could ask.

8. Rather send you only one. It's a long, talky and unwieldy thing, and I'd like to cut it down to about a tenth size before making a general distribution via printing -- confining the "prepublication" distribution to those who have some urgent need to get the info in whatever awkward shape it's in. Maybe you can talk around it and the guys' notes can be the essential abstract.

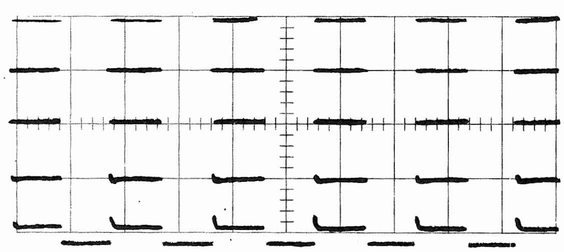
9. I think if you look on the output emitter network as "DC Shift" compensation and throw out that "thermal" nomenclature, you get a better picture of the purpose. The compensation is not for ambient temperature in the plug-in, but for the change in a transistor's characteristics that happens when you turn it on hard. It starts out with normal gain, then picks up a little. It would be better, I think, for understanding purposes if--in your mind's eye--you turn the transistors over so their emitters ("cathodes") are looking at each other, and then split 'em apart as shown on the enclosed sketch. It's front-corner boost, or flat-top sag, or whatever you want to call the rummy little time-constants that get to ya when you work with transistors or vacuum tubes. In this case, the compensation looks like about 1/2% by 360 μsec, or 0.3 mm in 6 cm. Pretty small.



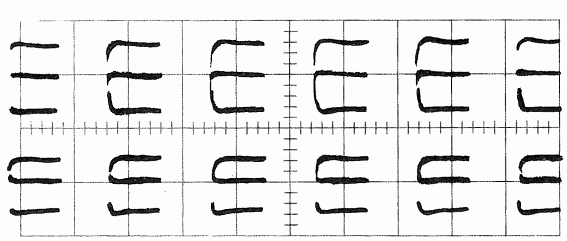




50 µsec/cm O - Unit R_i R_f compensation (1)



50 µsec/cm O - Unit R_i R_f compensation (2)



Comp Cap Lead Dress, Effects of L.F. REJ sw. (3)

CALIBRATION/TEST LIMIT ADDENDUM

OPEN LOOP GAIN

PMSE 4-7-65

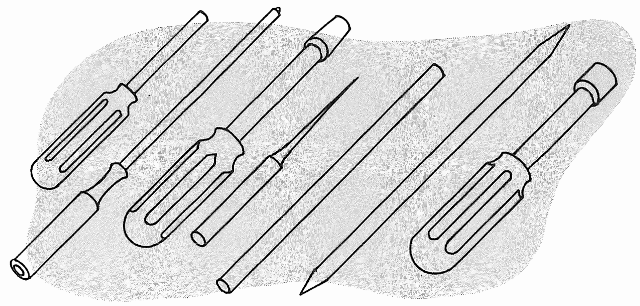
O FCP, July, 1964

Page 4, Step 21 change to:
OPEN LOOP GAIN
Range: Adjust to 3000.

Page 14, step 21 change test limit to:
OPEN LOOP GAIN   adjust to 3000

Page 14, step 21, last sentence change to:
Adjust Open Loop Gain for exactly 3 cm of deflection.

Reason for change:
Test limits were too tight and did not take into consideration differences in indicators.



SECTION 7
CALIBRATION PROCEDURE

Introduction

The following procedure should be used to calibrate the Type O Operational Amplifier Plug-In Unit. The instrument should not require frequent calibration, but occasional adjustments will be necessary when tubes and other components are changed. Also, a periodic recalibration is desirable from the standpoint of preventive maintenance.

Apparent troubles in the instrument are occasionally the result of improper calibration of one or more circuits. Consequently, calibration checks should be an integral part of any troubleshooting procedure. Abnormal indications during calibration checks will often aid in isolating troubles to a definite circuit or stage.

In the instructions that follow, the steps are arranged in the proper sequence for a complete calibration of the unit. Each step contains the information required to make one check or adjustment or a series of related checks or adjustments. The steps are arranged to avoid unnecessary repetition of checks or adjustments.

EQUIPMENT REQUIRED

NOTE

It may be necessary to use BNC to UHF adapters, on some instruments, to be able to calibrate the instrument using the following procedure. If your instrument is equipped with BNC connectors, then the Input Time Constant Normalizer with BNC connectors (011-0068-00) must be used. If a UHF Input Time Constant Normalizer is used along with an adapter, then the unit will be normalized to approximately 45 pF. The adapter adds about 2 pF to the Input Time Constant Normalizer.

The following equipment or its equivalent is required to perform a complete calibration of the Type O Unit.

1. An accurate DC voltmeter with a sensitivity of 5000 ohms per volt or better.

2. An ohmmeter.

3. An oscilloscope having a bandpass of at least 30 MHz in which to insert the Type O Unit during calibration, such as a Tektronix Type 540, 550 or 580 series oscilloscope. The test oscilloscope square-wave transient adjustment must be correct.

4. Standard amplitude calibrator. Accuracy within 0.25%; signal amplitude, 5 millivolts to 50 volts; output frequency of approximately 1 kHz. Tektronix calibration fixture 067-0502-00 recommended.

5. Square-wave generator. Frequency, 1 kHz and 500 kHz; risetime, 20 ns or faster from high-amplitude output; 1 ns or faster from first-rise output. High-amplitude output variable from 0.2 volts to 12 volts into a 50-ohm load, about 7 volts to 120 volts unterminated. Fast-rise output variable from 0.2 volts to .5 volts into a 50-ohm load. Tektronix Type 106 Square-Wave Generator recommended.

6. (Optional). Constant amplitude signal generator. Capable of generating sine wave of constant amplitude from 350 kHz through 30 MHz with an output impedance of 50 ohms. Tektronix Type 191 Constant Amplitude Signal Generator recommended.

7. A 47-pF Input Normalizer. On plug-in units with UHF type connectors order Tektronix Part No. 011-0030-00; for units with BNC type connectors order Tektronix Part No. 011-0068-00.

8. A 6-inch plug-in extension, Tektronix Part No. 013-0055-00.

9. A 50-ohm 42-inch coaxial cable with BNC connectors on each end. Order Tektronix Part No. 012-0057-01.

10. Coaxial cable. Impedance, 50 ohms; Type RG8/213; length, five nanoseconds; connectors, GR874. Tektronix Part No. 017-0502-00. Supplied with items 5 and 6.

11. In-line termination. Impedance, 50 ohms; connectors, GR input with BNC male output. Tektronix Part No. 017-0083-00. Supplied with items 5 and 6.

12. 10× attenuator. Impedance, 50 ohms; connectors GR-Type. Tektronix Part No. 017-0078-00.

13. Adapter, GR to BNC male. Tektronix Part No. 017-0064-00.

14. Additional adapter (for plug-in units with UHF connectors). UHF male to BNC female. Tektronix Part No. 103-0015-00.

15. One 18-inch patch cord. BNC to banana plug connectors. Tektronix Part No. 012-0091-00.

16. One 18-inch patch cord. Banana plug connectors. Tektronix Part No. 012-0031-00.

17. Double banana plug adapters (three). General Radio Type 274-MB.

18. Capacitor. 1 μF, 100 V.

19. Two 6-inch patch cords. For plug-in units with BNC connectors use Tektronix Part No. 012-0088-00; for UHF connectors, use Tektronix Part No. 012-0023-00.

20. A 10 kΩ, 1%, ½-watt resistor

21. A 27-ohm, 10%, 1-watt resistor.

22. Grid Current Checker, Tektronix Part No. 067-0507-00. A very low-leakage capacitor and a SPST switch may be substituted for the O Unit Grid Current Checker. The Capacitor and switch, if used, should be wired as per Fig. 4-3 in this manual. Note: Polystyrene and mylar type capacitors have very low leakage.

23. Assorted alignment tools:

1—Jaco No. 125 insulated low-capacitance-type screwdriver with $1\frac{1}{2}$ inch long shank and $\frac{1}{8}$ inch wide metal tip. Total length is 5 inch. Tektronix Part No. 003-0000-00.

1—Low-capacitance alignment tool, Tektronix Part No. 003-0007-00;

Consisting of:

1—Gray nylon insert with wire pin, Tektronix Part No. 0308-00.

1—White cymac insert with wire pin, Tektronix Part No. 003-0309-00.

1—Gray nylon insert with metal screwdriver tip, Tektronix Part No. 003-0334-00.

1—$\frac{5}{64}$ inch hexagonal wrench insert, Tektronix Part No. 003-0310-00.

1—Nylon handle, Tektronix Part No. 003-0307-00.

PRELIMINARY PROCEDURE

Make a complete visual check of the plug-in unit. Be careful not to change any of the lead dress around the Z_i—Z_f SELECTOR switches.

Make resistance checks from each interconnecting plug terminal to ground. The resistance values should be approximately as listed in Table 7-1.

TABLE 7-1

Interconnecting Plug Terminal Resistance to Ground

Pin Number	Resistance	Pin Number	Resistance
1	3.9 kΩ	9	0.5 M
2	0	10	1.5 kΩ
3	3.9 kΩ	11	4.7 kΩ
4	Infinite	12	0.5 M
5	Infinite	13	Infinite
6	Infinite	14	Infinite
7	Infinite	15	70 Ω
8	Infinite	16	Infinite

CHECK AND ADJUSTMENT PROCEDURE

Calibration of the Type O Unit requires that the line voltage be at the value indicated at the rear of the oscilloscope, near the power cord.

Install the Type O Unit in the oscilloscope and turn on the power. Let the instrument warm up for a few minutes.

Preamplifier

1. Adjust Front-Panel Controls

Set the Type O and oscilloscope controls as follows:

Type O Unit

Preamplifier controls

POSITION	Midrange
VERTICAL DISPLAY	+DC
VARIABLE	CALIBRATED
VOLTS/CM	.05

Operational Amplifier controls (both channels)

±GRID SEL	(—)
Z_i SELECTOR	1 MEG
Z_f SELECTOR	1 MEG
INTEGRATOR LF REJECT	OFF

Type 545B Oscilloscope

Stability	Preset
Triggering Level	As is
Triggering Mode	Auto
Trigger Slope	+Int
Time/Cm	1 mSec
Variable (Time/CM)	Calibrated
Horizontal Display	A
5× Magnifier	Off
Horizontal Position and Vernier	Midrange
Amplitude Calibrator	Off
Intensity	CCW

NOTE

The Type 545B Oscilloscope utilizing Time Base A was used in this Calibration Procedure. If a substitute oscilloscope is used, adapt the control settings given in the procedure to the oscilloscope being used.

2. Check Vertical System Electrical Center

Turn the Intensity control clockwise to produce a visible trace. Remove the oscilloscope left side panel. Locate the vertical system electrical center by shorting between the two test points marked [1], Fig. 7-1 or Fig. 7-2, with the 27- ohm resistor. Position the trace with the POSITION control for no movement of the trace as the test points marked [1] are shorted and then unshorted. Record the trace location for future reference.

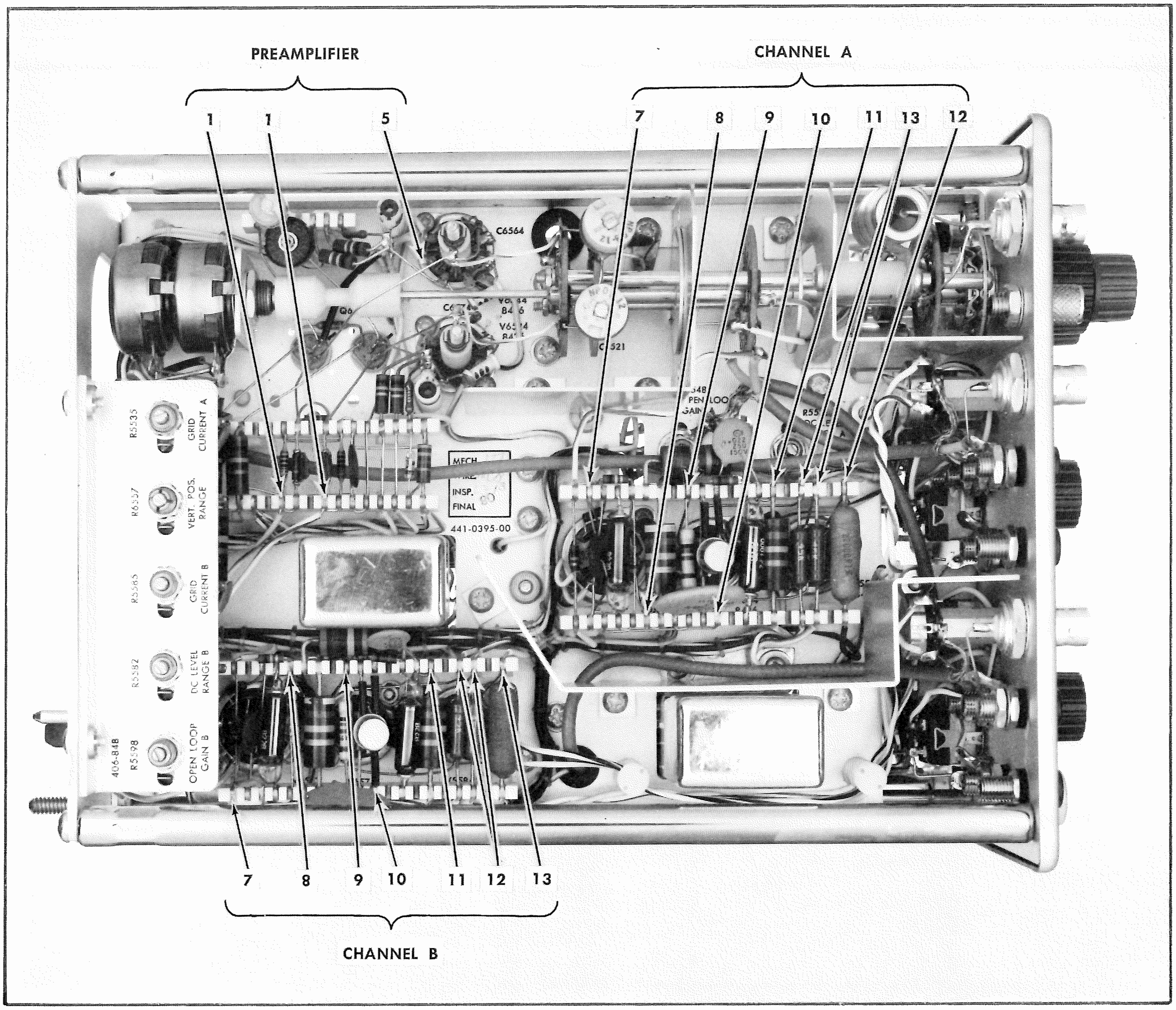


Fig. 7-1. Voltage test point locations for units SN 2950 and up. These test point are identified on the appropriate schematic.

3. Check Preamplifier DC Output Level

The voltage at the output leads of the plug-in unit (test point [1]) must be between 65 and 70 volts above ground. The preferred value for the output voltage is +67 volts. If an output transistor is shorted, the output voltage will be near 72 volts for the lead connected to the shorted transisor.

4. Adjust DC Balance

With the trace centered, adjust the DC BAL. control until there is no trace movement as the VARIABLE (VOLTS/CM) control is rotated. Return the VARIABLE control to the CALIBRATED position.

5. Adjust Vertical Position Range

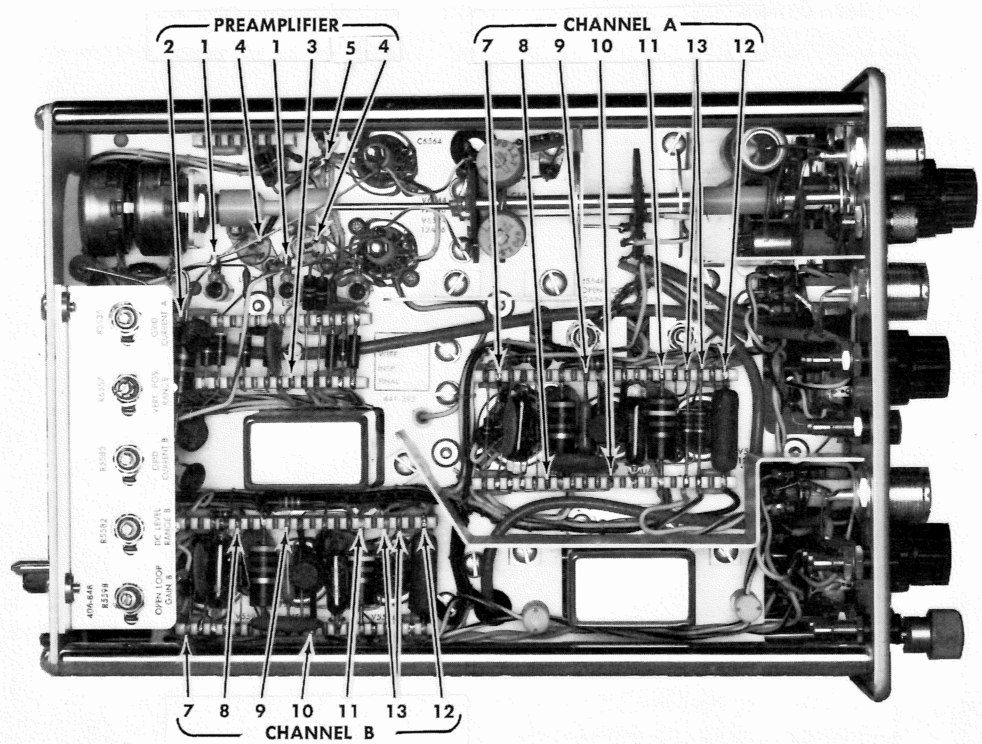
Set the O Unit POSITION control to midrange. Then adjust the VERT. POS. RANGE control, R6557, until the trace is at the previously determined electrical center. The VERT. POS. RANGE control is illustrated as (5), Fig. 7-3.

6. Gain Adjust

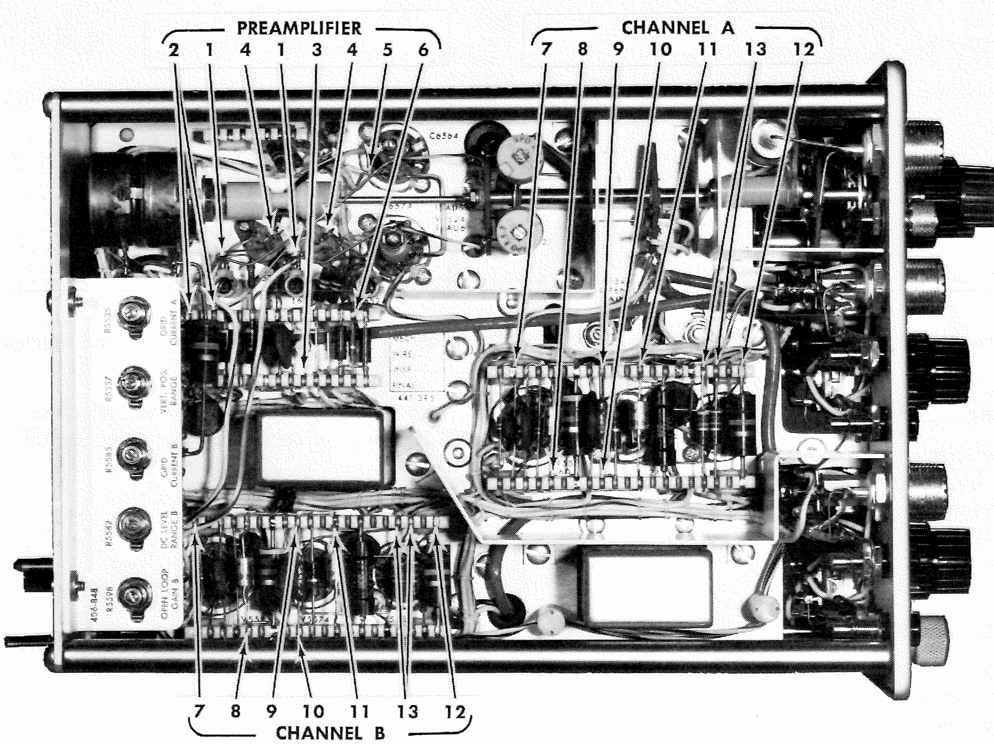
a. Set the VERTICAL DISPLAY switch to +DC.

b. Set the Standard Amplitude Calibrator controls as follows:

Amplitude	.1 Volt
Mode	Square Wave
Mixed	Up
×100 Amplifier	Not applicable
Power	On



(a) Type O Unit from SN 814 to 2949.



(b) Type O Unit from SN 101 to 813.

Fig. 7-2. Voltage test point locations for units SN 101 to 2949. These test points are identified on the appropriate schematics.

O-Unit

Provisional Manual Cal Procedure Changes

For S/N 2950-Up

Step 10* (Input C and Neutralizing Capacitors): Set up as before (105, 50Ω Termination, 47 pF Standardizer).

a. Set O-Unit VARIABLE VOLTS/CM to minimum amplitude, increase output from 105 for 3.5 cm display. With VERTICAL DISPLAY switch at EXT + DC, adjust* C6521 for optimum flat top display of 1 kHz square waves.

b. Set the VERTICAL DISPLAY switch to EXT – DC and adjust* C6541 for optimum flat bottom on display.

c. Set the VERTICAL DISPLAY switch back to EXT + DC, turn VARIABLE VOLTS/CM to CALIBRATED, and reduce output from 105. Adjust C6574 for optimum flat top.

d. Switch to EXT – DC and adjust C6564 for optimum flat bottom.

e. Switch to EXT + DC, return the VARIABLE control to minimum gain and increase 105 output amplitude. Adjust C6521 if necessary for optimum flat top.

f. Work back and forth between C6521 and C6574 as necessary for optimum flat top at both extremes of the VARIABLE VOLTS/CM control, in EXT + DC.

g. Repeat steps e. and f. for C6541 and C6564, with the VERTICAL DISPLAY switch set to EXT – DC, observing the bottom of the displayed waveform.

*If adjustments are made in this step, also perform step 11.

Step 12. Adjust HF Peaking.

For routine recalibration, proceed with Step 13 (Check Preamplifier Risetime).

If peaking is needed, see Step 13A.

Step 13A. Adjust High Frequency Peaking.

If the risetime measured in Step 13 is greater than 14 nsec (with 107) or the transient response has excessive aberrations on the positive side, adjust peaking as follows:

Preset: L6524 and L6544 so the core is just below the winding (on the chassis side). Set R6546 and C6546 to mid-range (C6546 is at mid-range when the silvered area is over the arrow). Equipment and control settings as in Step 13.

a. Adjust R6546 and C6546 for optimum flat top (EXT + DC) and flat bottom (EXT – DC) while viewing the waveform at 1 μsec/cm to 5 μsec/cm. R6546 will affect primarily the first 250 nsec after the leading edge and C6546 the second 250 nsec. (Ignore the negative side of the waveform when in EXT + DC and the positive side when in EXT – DC: these aberrations are in the generator).

b. Increase equally the inductance of L6524 and L6544 (by turning the slugs so they enter the coil winding area) in small increments, observing the front corner of the square wave (top left corner in EXT + DC, bottom left corner in EXT – DC), switching back and forth between EXT + and –. The coils will affect about 1 mm at the start of the waveform viewed at 0.2 μsec/cm. Increase inductance for shortest risetime without excessive overshoot in either EXT + DC or EXT – DC.

c. Check risetime; repeat 13A-a and 13A-b (omit presets) if necessary.

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c. Apply the 0.1-volt signal from the Standard Amplitude Calibrator through a 50-ohm cable to the Type O EXT. INPUT connector.

d. Adjust the GAIN ADJ. front panel control for 2 cm of vertical deflection, centered about the graticule centerline. Be sure to measure from top to top or bottom to bottom on the trace, making certain the trace thickness does not become part of the amplitude.

e. Remove the standard Amplitude Calibrator signal from the O Unit EXT. INPUT connector.

7. Check Input Tubes Grid Input

The grid current effect of the input tubes, V6524 and V6544, can be checked as follows:

a. Ground the EXT. INPUT connector with a short patch cord.

b. Make sure the VOLTS/CM switch is set to .05, and the VARIABLE control to CALIBRATED.

c. Switch the VERTICAL DISPLAY switch for +DC to —DC. The trace shift should be less than 1 mm. If the shift is more than 1 mm, replace V6524 and V6544. See the Parts List for proper replacements.

d. Return the VERTICAL DISPLAY switch to +DC.

8. Check Input Tubes Microphonics

a. Gently tap the O Unit front panel. A light tap on the front panel should not produce microphonics of the short duration form greater than 2 mm, nor any prolonged ringing microphonics.

b. Disconnect the patch cord.

9. Check Attenuator Tolerances

a. Check the setting of the following O Unit controls.

VERTICAL DISPLAY	+DC
VARIABLE	CALIBRATED
VOLTS/CM	.05

b. Apply the 0.1-volt signal from the Standard Amplitude Calibrator to the O Unit EXT. INPUT connector.

c. Check for proper CRT vertical deflection according to Table 7-2.

NOTE

Should any position of the VOLTS/CM switch (except the .05 position) produce a vertical deflection more than the allowable tolerance of ±3%, it will be necessary to remove and measure on a precise bridge the resistors for that particular switch position. Replace one or both resistors if necessary and recheck for proper CRT deflection.

TABLE 7-2

VOLTS/CM Switch Setting	Standard Amplitude Calibrator Output	Vertical Deflection	Maximum Error For ±3% Accuracy
.05	.1 Volt	2 cm	Previously adjusted in Step 6.
.1	.2 Volt	2 cm	0.6 mm
.2	.5 Volt	2.5 cm	0.75 mm
.5	1 Volt	2 cm	0.6 mm
1	2 Volts	2 cm	0.6 mm
2	5 Volts	2.5 cm	0.75 mm
5	10 Volts	2 cm	0.6 mm
10	20 Volts	2 cm	0.6 mm
20	50 Volts	2.5 cm	0.75 mm

d. Remove the Standard Amplitude Calibrator signal from the O Unit.

10. Adjust Input Capacitance and Neutralizing Capacitors

Set the front-panel controls of the O Unit and Type 106 Square Wave Generator as follows:

Type O Unit

VARIABLE	Minimum amplitude (slightly clockwise out of detent)
VOLTS/CM	.05

Type 106 Square Wave Generator

Repetition Rate Range	1 kHz
Multiplier	Less than 1
Symmetry	Midrange
Amplitude	Fully CCW
High Amplitude Fast Rise Switch	Hi Amplitude
+Transition Amplitude	Not applicable
—Transition Amplitude	Not applicable
Power	On

At this point in the procedure the remaining front-panel controls should be at the following positions.

Type O Unit

POSITION	Midrange
VERTICAL DISPLAY	+DC

Operational Amplifier Controls (both channels)

±GRID SEL	(—)
Z_i SELECTOR	1 MEG
Z_f SELECTOR	1 MEG
INTEGRATOR LF REJECT	OFF

Type 545B Oscilloscope

Stability	Preset
Triggering Level	As is
Triggering Mode	Auto
Trigger Slope	+Int
Time/Cm	1 mSEC
Variable (Time/Cm)	Calibrated
Horizontal Display	A
5× Magnifier	Off
Horizontal Position and Vernier	Midrange
Amplitude Calibrator	Off
Intensity	Normal trace brightness

Apply the 1 kHz signal from the Type 106 high amplitude output connector through a 10× attenuator, 51ns coaxial cable, 50-ohm in-line termination and a 47-pF Time Constant Normalizer to the 0 Unit EXT. INPUT connector. Connect all items in the order that was given.

Set the Type 106 Amplitude control to produce a display amplitude of 3.5 cm.

NOTE

Maintain 3.5 cm of display amplitude throughout the remaining portion of this step.

Procedure For O Units SN 2950 and UP

a. Using Table 7-3 as a guide, adjust the variable capacitors as directed in steps b through g in this procedure for units SN 2950 and up. Fig. 7-3A shows the location of the capacitors.

TABLE 7-3

VERTICAL DISPLAY Switch	VARIABLE Control	Adjust	For Optimum
+DC	Minimum amplitude CALIBRATED	C6521 C6574	Flat top
—DC	Minimum amplitude CALIBRATED	C6541 C6564	Flat bottom

b. Adjust C6521 for an optimum flat top appearance of the displayed waveform.

c. Move the VERTICAL DISPLAY switch to the —DC position and adjust C6541 for an optimum flat bottom appearance of the displayed waveform.

d. Set the VERTICAL DISPLAY switch to the + DC position, turn the VARIABLE control to CALIBRATED (reduce the Type 106 signal amplitude) and adjust C6574 for an optimum flat top appearance of the displayed waveform.

e. Move the VERTICAL DISPLAY switch to —DC and adjust C6564 for an optimum flat bottom appearance of the displayed waveform.

f. The adjustments of C6541 and C6564 tend to interact; turn the VARIABLE control to the minimum amplitude position, readjust C6541, turn the VARIABLE control to CALIBRATED, and readjust C6564. Repeat this procedure until an optimum flat bottom waveform appearance is obtained at the two extremes of the VARIABLE control range.

g. Move the VERTICAL DISPLAY switch to +DC, turn the VARIABLE control to the minimum amplitude position and readjust C6521. Turn the VARIABLE control to CALIBRATED and readjust C6574. Alternately readjust these two capacitors until an optimum flat top waveform appearance is obtained at the two extremes of the VARIABLE control range.

Procedure For O Units SN 101 to 2949

Set VARIABLE (VOLTS/CM) control to CALIBRATE.

a. Adjust C6521 [(10-a) Fig. 7-31B] for an optimum flat-top display.

b. Set the VERTICAL DISPLAY switch to —DC and adjust C6541 [(10-b), Fig. 7-3B] for an optimum flat-top display.

c. Set the VERTICAL DISPLAY switch back to +DC and set the VARIABLE (VOLTS/CM) control for minimum gain. Increase the Type 106 output for 3.5 cm of CRT display. Adjust C6574 [(10-c), Fig. 7-3B] for an optimum flat-top display.

d. Return the VARIABLE control to the CALIBRATED position and lower the Type 106 output to 3.5 mm of CRT display. It may be necessary to readjust C6521.

e. Work betwen C6521 and C6574 for best square-wave flat top at the two extremes of the VARIABLE control.

f. Repeat this procedure first with C6541 and then C6564 [(10-f), Fig. 7-3B] with the VERTICAL DISPLAY switch at —DC.

11. Adjust Attenuator Compensation

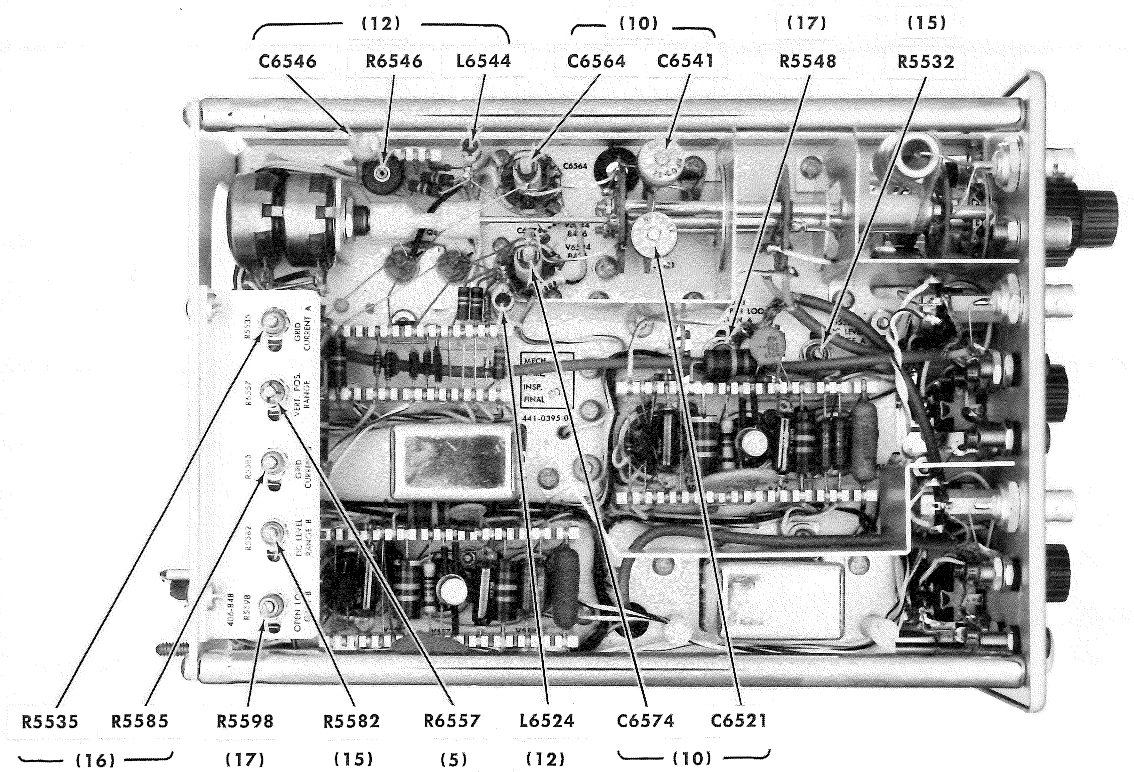
a. Temporarily remove the signal from the O Unit EXT. INPUT connector. Turn off the oscilloscope. Insert the plug-in extension (item 8) between the O Unit and oscilloscope. Reconnect the signal to the EXT. INPUT connector. Turn on the oscilloscope and wait about two minutes before proceeding.

b. Set the oscilloscope Time/Cm switch to .5 mSec.

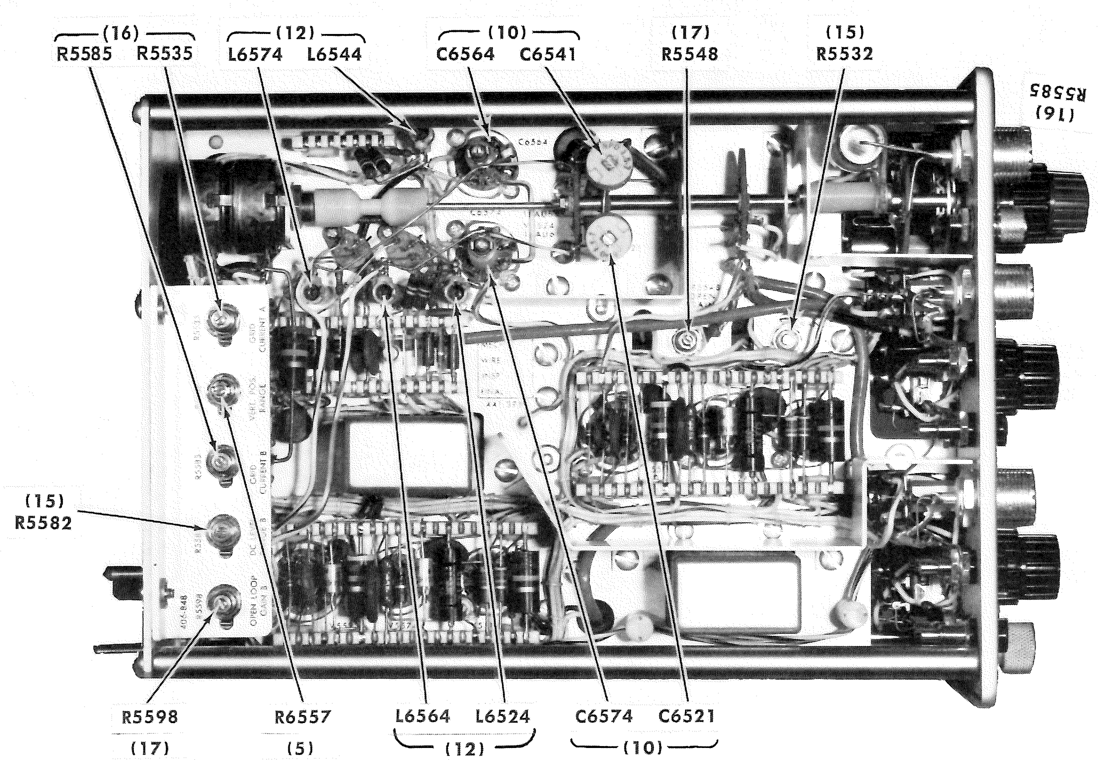
c. Check that the O Unit controls are set as follows:

VERTICAL DISPLAY	+DC
VARIABLE	CALIBRATED
VOLTS/CM	.05

d. Using Table 7-4 as a guide, set the VOLTS/CM switch to the positions given in the table and adjust the indicated capacitors. Adjust the "B" capacitors for optimum flat top, and the "C" capacitors for optimum square corners. Fig. 7-4 (11) shows the location of the adjustments. When using Table 7-4, remove the 10× attenuator and then the 50-ohm termination when necessary to obtain more signal drive. (Use GR to BNC adapter when 50-ohm termination is removed.) Maintain a display amplitude of about 3.5 cm during the adjustment procedure.



(a) Type O Unit for SN 2950 and up.



(b) Type O Unit for SN 101 to 2949.

Fig. 7-3. Location of calibration adjustments. Numbers in parenthesis refer to steps in the adjustment procedure.

TABLE 7-4

VOLTS/CM Switch	Adjust for Optimum	
	Flat top	Square Corner
.1	C6508B	C6508C
.2	C6509B	C6509C
.5	C6510B	C6510C
5	C6513B	C6513C

NOTE

The waveform on the above adjustments will appear slightly spiked when the adjustment is properly made. This effect will be due to the change in capacitance when the unit is extended from the oscilloscope.

e. Disconnect the Type 106 signal from the O Unit. Turn off the oscilloscope, remove the plug-in extension and put the O Unit back in the oscilloscope. Turn on the oscilloscope and wait for about a two minute warm-up period before proceeding.

12. Adjust High-Frequency Peaking

Connect the +Output of the Type 106 Square Wave Generator through a 5-ns coaxial cable and a 50-ohm in1line termination to the O Unit EXT. INPUT connector. Connect all items in the order given.

Set the front-panel controls as follows:

Type O Unit

VOLTS/CM	.05

Type 545B Oscilloscope

Time/Cm	.5 μSec

Type 106 Square Wave Generator

Repetition Rate Range	100 kHz
Multiplier	5
Amplitude	Fully CCW
Hi Amplitude Fast Rise switch	Fast Rise
+Transition Amplitude	As required for 3.5 cm of vertical deflection

At this point in the procedure the remaining controls should be at the following positions:

Type O Unit

POSITION	Midrange
VERTICAL DISPLAY	+DC
VARIABLE	CALIBRATED

Operational Amplifier Controls (both channels)

±GRID SEL	(—)
Z_i SELECTOR	1 MEG
Z_f SELECTOR	1 MEG

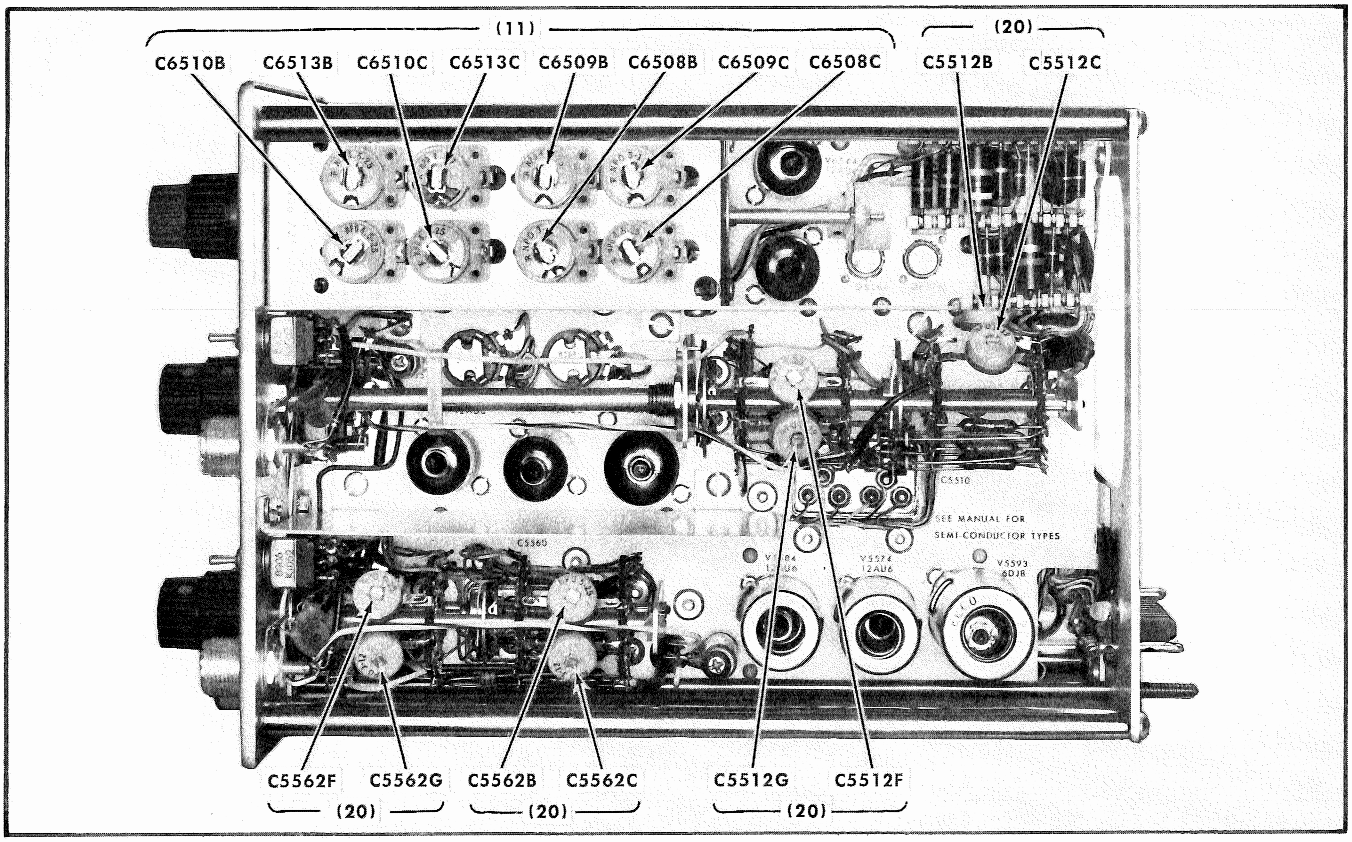


Fig. 7-4. Location of calibration adjustments. Numbers in parenthesis refer to steps in adjustment procedure. Applies to O Units SN 101 and up.

Type 545B Oscilloscope

Stability	Preset
Triggering Level	As is
Triggering Mode	Auto
Trigger Slope	+Int
Variable (Time/Cm)	Calibrated
Horizontal Display	A
5X Magnifier	Off
Horizontal Position and Vernier	Midrange
Amplitude Calibrator	Off
Intensity	Normal trace brightness

Procedure For O Units SN 2950 and Up

a. Preset the high-frequency peaking adjustments by moving the cores of L6524 and L6544 toward the chassis until they are out of the coil winding area and adjust R6546 and C6546 to mid-range. C6546 is at mid-range when the silvered area is aligned with the arrow. Fig. 7-3A shows the location of these adjustments.

b. Adjust R6546 and C6546 for optimum flat-top waveform. (R6546 affects the first 250 ns after the leading edge; C6546 affects the second 250 ns.)

c. Change the VERTICAL DISPLAY switch to — DC, move the bottom of the waveform near the graticule center with the POSITION control, and adjust R6546 and C6546 for optimum flat-bottom waveform.

Since adjusting R6546 and C6546 affects both the flat-top and flat-bottom waveform it may be necessary to repeat this procedure a few times to obtain an equally flat-top and flat-bottom waveform.

d. Set the VERTICAL DISPLAY switch on +DC. Adjust L6424 for an optimum square top-front corner of the waveform. Change the VERTICAL DISPLAY switch to — DC and adjust L6544 for an optimum square bottom front corner.

NOTE

Alternately adjust L6524 and L6544 in equal amounts and approach their final settings in small increments.

Procedure For O Units SN 101 to 2949

a. Adjust L6524 and L6564 for optimum square top-front corner Fig. 7-3B shows the location of these adjustments.

b. Move the VERTICAL DISPLAY switch to —DC and adjust L6544 and L6574 for optimum square bottom-front corner. Fig. 7-3B shows the locations of these adjustments.

NOTE

Check for interaction between the +DC and —DC adjustments. Repeat this procedure as necessary to obtain an optimum square corner waveform in the +DC and —DC positions of the VERTICAL DISPLAY switch.

13. Check Preamplifier Risetime

a. Set the front-panel controls as follows:

Type O Unit

POSITION	Midrange
VERTICAL DISPLAY	+DC

Type 545B Oscilloscope

Triggering Level	As required for stable display
Triggering Mode	AC LF Reject
Time/Cm	.1 μsec
5X Magnifier	On
Horizontal Position	As required to position rising part of waveform near center of graticule

Type 106 Square Wave Generator

+Transition Amplitude	2 cm of vertical deflection

b. Measure the rise-time of the waveform from the 10% to the 90% points of the rise. Rise-time should be no greater than 14 ns (7 mm) and is typically less than 14 ns (assuming the CRT geometry is correctly adjusted).

c. Disconnect the Type 106 signal from the O Unit.

14. (Optional) Check Preamplifier Sine-Wave Frequency Response

a. Set the front-panel controls as follows:

Type 545B Oscilloscope

Stability	Fully clockwise
Triggering Level	Fully clockwise
Triggering Mode	AC
Time/Cm	.1 mSec
5X Magnifier	Off
Horizontal Position and Vernier	Midrange

Type 191 Constant Amplitude Signal Generator

Frequency Dial	Fully Clockwise
Frequency Range	50 kHz Only
Amplitude	15 (or as required)
Variable	As required for 3 cm of display
Amplitude Range	50 - 500 mV
Power	On

b. Connect the output of the Type 191 through a 5-ns coaxial cable and a 50-ohm in-line termination to the O Unit EXT. INPUT connector. Connect all items in the order given.

c. Observe the CRT display and adjust the VARIABLE control of the Type 191 for exactly 3 cm of vertical display. The display should be a solid band, not a stable sine-wave presentation of the Type 191 output signal.

d. Change the Frequency Range switch of the Type 191 to the 18 - 42 MHz range. Without changing the output amplitude, increase the frequency from the Type 191 until the vertical display decreases in amplitude to 2.1 cm. This is the point at which the high frequency response is down 3-dB from the 50-kHz reference signal.

e. Note the frequency dial setting of the Type 191. It should be 25 MHz or higher.

f. Disconect the Type 191 signal from the O Unit.

This completes calibration of the preamplifier section.

A and B OPERATIONAL AMPLIFIERS

The following calibration procedure is applicable to the A and B Operational Amplifiers. The amplifiers are identical except for part numbers and location. Either amplifier can be selected by rotating the VERTICAL DISPLAY switch to the desired OUTPUT position.

The calibration procedue lists both A and B OUTPUT positions for the VERTICAL DISPLAY switch setting; however, only one amplifier at a time can be selected. If you start with Amplifier A (VERTICAL DISPLAY switch set to + or — OUTPUT) complete the entire procedure before starting Amplifier B (VERTICAL DISPLAY switch set to + or — B OUTPUT).

15. Adjust Output DC Level

a. Set the front-panel controls as follows:

Type O Unit

POSITION	Midrange
VERTICAL DISPLAY	+A OUTPUT +B OUTPUT
VOLTS/CM	.5

At this point in the procedure the remaining controls should be at the following settings:

Type O Unit

VARIABLE	CALIBRATED

Operational Amplifier Controls (both channels)

±GRID SELECTOR	(—)
Z_i SELECTOR	1 MEG
Z_f SELECTOR	1 MEG
INTEGRATOR LF REJECT	OFF

Type 545B Oscilloscope

Stability	Fully Clockwise
Triggering	Fully Clockwise
Triggering Mode	AC
Triggering Slope	+Int
Time/Cm	.1 mSec
Variable (Time/Cm)	Calibrated
Horizontal Display	A
5× Magnifier	Off
Horizontal Position and Vernier	Midrange
Amplitude Calibrator	Off
Intensity	Normal trace brightness

b. Set the OUTPUT DC LEVEL control (front-panel) to midrange position.

c. Push ZERO CHECK button in and hold; move the trace to graticule center with the POSITION control then release the ZERO CHECK button. Push the OUTPUT DC LEVEL switch to ADJ. position and hold; adjust DC LEVEL RANGE control R5532 (R5582 Amplifier B); see Fig. 7-3, to position the trace near graticule center. Release OUTPUT DC LEVEL switch.

d. Push ZERO CHECK button in and hold, if necessary recenter trace with POSITION control, then release ZERO CHECK button. Push the OUTPUT DC LEVEL switch to ADJ. position and hold it there while adjusting the OUTPUT DC LEVEL control (front-panel) to position the trace at graticule center. Release the OUTPUT DC LEVEL switch. Any drift that occurs during this adjustment is due to the 100× gain when the OUTPUT DC LEVEL switch is in the ADJ. position.

16. Adjust Input Vacuum Tubes Grid Current

a. Set the front-panel controls as follows:

Type O Unit

VOLTS/CM	1

Operational Amplifier Controls (both channels)

Z_i SELECTOR	EXT.
Z_f SELECTOR	EXT.
±GRID SEL	(+)

Type 545B Oscilloscope

Time/Cm	1 Sec

b. +Grid. Plug the Grid Current Checker (push buttons facing up) into the banana jacks on the front of the O Unit. (If the low-leakage capacitor and toggle-switch test fixtures are used refer to Fig. 4-3B for their connections.)

As the spot moves past a vertical graticule line push the +Grid current button and hold. (Open the toggle switch when using the test set-up as in Fig. 4-3B.) The spot should not move vertically more than 2½ cm in 5 cm of horizontal travel. Release the +Grid Current button. The amount of vertical travel indicates the rate at which current is charging the .001-μF capacitor. The most stable grid current condition is when the spot moves upward.

If the spot moves more than 2½ cm vertically adjust the GRID CURRENT control R5535 (R5585 for Amplifier B—see Fig. 7-3) to reduce grid current and vertical spot travel. Typically the vertical excursion should not exceed 3/4 cm in 5 cm of horizontal travel.

Adjustment of the +Grid current affects the —Grid current.

c. —Grid. If the low-leakage capacitor and toggle-switch test fixtures are used, connect as in Fig. 4-3A.

As the spot moves past a vertical graticule line push the —Grid button and hold. (Open the toggle-switch when using the test set-up as in Fig. 4-3A.) The spot should not move vertically more than 2½ cm in 5 cm of horizontal travel. Release the —Grid Current button. The most stable grid current condition is when the spot moves downward.

If the spot moves vertically more than 2½ cm exchange the two input tubes (V5524 and V5534). Typically the vertical excursion should not exceed 1½ cm in 5 cm of horizontal travel. (Exchanging the input tubes may bring the grid current within allowable limits.) If the input tubes were changed repeat steps 15 and 16 as interaction occurs between the GRID CURRENT and OUTPUT DC LEVEL controls when the input tubes are interchanged or replaced.

NOTE

In the Grid Current adjustments just described, two limits of vertical spot excursion were stated for each grid, a maximum limit and a typical limit. The typical limit corresponds to a practical grid current value of 0.15 nanoamperes for the +Grid, and 0.3 nanoamperes for the —Grid. (The procedure for calculating grid current is on page 4-2, Grid Current Calculations.)

d. Remove the Grid Current Checker from the O Unit.

17. Adjust Open Loop Gain

a. Set the front-panel controls as follows:

Type O Unit

±GRID SEL	(—)

Type 545B Oscilloscope

Stability	Preset
Triggering Level	Midrange
Triggering Mode	Auto
Time/Cm	.5 mSec

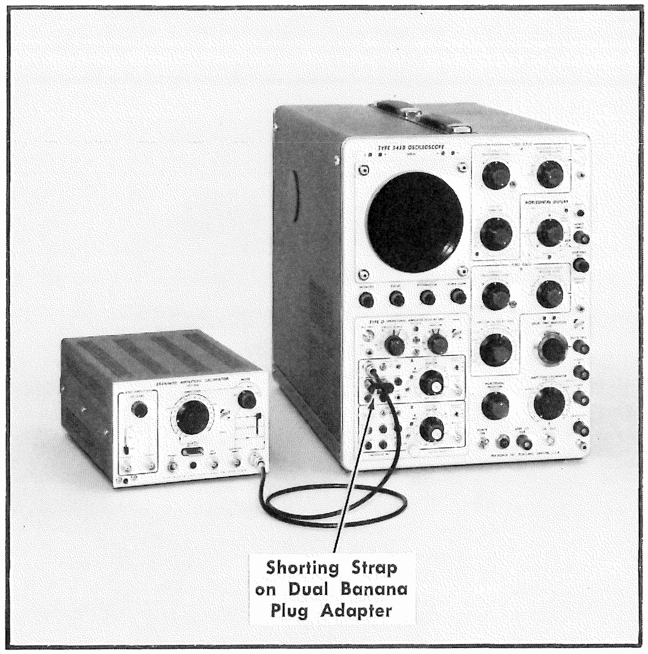


Fig. 7-5. Initial test setup for step 17.

Standard Amplitude Calibrator

Amplitude	1 mVolt
Mode	Square Wave
Mixed	Up
×100 Amplifier	Not Applicable
Power	ON

b. Install a shorted double banana plug adapter between the INPUT and —GRID banana jacks (see Fig. 7-5) and apply a 1 mVolt signal from the Standard Amplitude Calibrator to the appropriate O Unit INPUT connector. Set Z_f on the unused amplifier to 1 MEG; otherwise stray signals may deflect the display offscreen. Position the display to graticule center with the POSITION control. If necessary, readjust OUTPUT DC LEVEL control (front-panel) to move the display onto the graticule.

c. Adjust the OPEN LOOP GAIN control R5548 (R5598 for Amplifier B—see Fig. 7-3) for 2½ cm of vertical display.

d. Remove the shorted double banana plug adapter and the Standard Amplitude Calibrator signal.

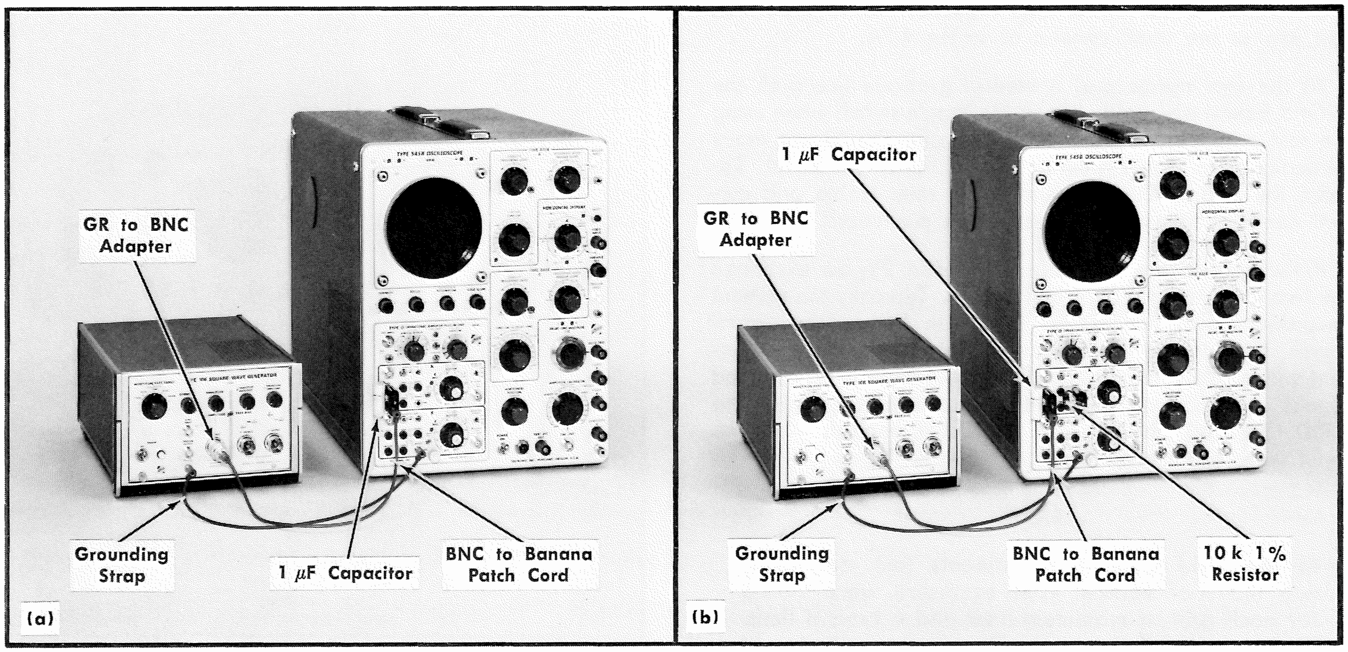


Fig. 7-6. Initial test setup for step 18.

18. Check Output Voltage and Current

a. Set the front-panel controls as follows:

Type O Unit

VOLTS/CM	20

Operational Amplifier Controls (both channels)

Z_i SELECTOR	1 MEG
Z_f SELECTOR	1 MEG

Type 106 Square Wave Generator

Repetition Rate Range	1 kHz
Multiplier	Less than 1
Symmetry	Midrange
Amplitude	CCW
Hi Amplitude Fast Rise switch	Hi Amplitude

b. Connect the signal from the Type 106 through the externally mounted 1 μF coupling capacitor (See Fig. 7-6A). Adjust the Type 106 Hi Amplitude control for 5 cm of vertical display (100 volts).

c. Install the 10-kΩ, 1% resistor between the OUTPUT banana jack and GND (See Fig. 7-6B). Display should remain 5 cm in amplitude.

d. Turn Amplitude control fully CCW; disconnect the Type 106 signal and test fixtures.

19. Check Z_i and Z_f Components

This step checks the tolerances of the resistors and capacitors in the Z_i and Z_f circuits.

a. Set the front-panel controls as follows:

Type O Unit

VOLTS/CM	1

Operational Amplifier Controls (both channels)

Z_i SELECTOR	.1 MEG
Z_f SELECTOR	.1 MEG

Type 106 Square Wave Generator

Repetition Rate Range	100 Hz
Multiplier	4.5
Amplitude	As required for 4 cm display

At this point in the procedure the remaining controls should be at the following settings:

Type O Unit

Preamplifier Controls

POSITION	Midrange
VERTICAL DISPLAY	+A +B
VARIABLE	CALIBRATED

Operational Amplifier Controls (both channels)

±GRID SEL	(—)
INTEGRATOR LF REJECT	OFF

Type 545B Oscilloscope

Stability	Preset
Triggering Level	Midrange
Triggering Mode	Auto
Trigger Slope	+Int
Time/Cm	.5 mSec
Variable (Time/Cm)	Calibrated
Horizontal Display	A
5× Magnifier	Off
Horizontal Position and Vernier	Midrange
Amplitude Calibrator	Off
Intensity	Normal trace brightness

Type 106 Square Wave Generator

Hi Amplitude Fast Rise switch	Hi Amplitude
Symmetry	Midrange
+Transition Amplitude	CCW
—Transition Amplitude	CCW
Power	On

b. Connect the Output of the Type 106 through a 5 ns coaxial cable and a 50-ohm in-line termination to the appropriate INPUT connector on the Type O Unit. Adjust the Type 106 Output Amplitude control for a vertical display of 4 cm.

c. Use Table 7-5 as a guide and check for correct vertical deflection.

TABLE 7-5

Z_i	Z_f	VOLTS/CM Switch	Deflection
.01 MEG	.01 MEG	1	4 cm
.1 MEG	.1 MEG	1	4 cm
.2 MEG	.2 MEG	1	4 cm
.5 MEG	.5 MEG	1	4 cm
1 MEG	1 MEG	1	4 cm
1 MEG	.5 MEG	.5	4 cm
1 MEG	.2 MEG	.2	4 cm
1 MEG	.1 MEG	.1	4 cm
.2 MEG	.01 MEG	.05	4 cm
.01 MEG	.1 MEG	10	4 cm
.1 MEG	1 MEG	10	4 cm
.2 MEG	1 MEG	5	4 cm
.5 MEG	1 MEG	2	4 cm

d. Set the Type O Unit front panel controls as follows:

Z_i SELECTOR	.1 μF
Z_f SELECTOR	.1 μF
VOLTS/CM	1
INTEGRATOR LF REJECT	1 CPS

Without changing the Type 106 Amplitude control the vertical display should be 4 cm.

e. Use Table 7-6 as a guide and check for correct vertical deflection.

TABLE 7-6

Z_i	Z_f	VOLTS/CM Switch	Deflection
1 μF	1 μF	1	4 cm
1 μF	.1 μF	1	4 cm
.01 μF	.01 μF	1	4 cm
.001 μF	.001 μF	1	4 cm
.01 μF	.001 μF	10	4 cm
.1 μF	.01 μF	10	4 cm
1 μF	.1 μF	10	4 cm
.001 μF	.01 μF	.1	4 cm
.01 μF	.1 μF	.1	4 cm
.1 μF	1 μF	.1	4 cm

f. Remove the Type 106 signal from the Type O Unit.

20. Adjust Variable .001-μF and 10-pF Z_i and Z_f Capacitors

Turn the oscilloscope off and install the plug-in extension between the Type O Unit and the oscilloscope. Turn the oscilloscope on and wait for about two minutes warm-up period. Adjust the OUTPUT DC LEVEL before proceeding. (Refer to step 15d.)

a. Set the front-panel controls as follows:

Type O Unit

VOLTS/CM	1

Operational Amplifier Controls (both channels)

Z_i SELECTOR	.01 μF
Z_f SELECTOR	.001 μF

Type 545B Oscilloscope

Triggering Level	As required for stable display
Triggering Mode	AC
Trigger Slope	+Int
Time/Cm	20 μSEC

Type 106 Square Wave Generator

Repetition Rate Range	10 kHz
Multiplier	Less than 1
Amplitude	As required for 4 cm of vertical display

b. Connect the Output of the Type 106 through a 10× attenuator, a 5-ns coaxial cable, and a 50-ohm in-line termination to the appropriate INPUT connector on the O Unit. Adjust the Amplitude control of the Type 106 for 4 cm of vertical display.

c. Using Table 7-7 as a guide, set the Z_i - Z_f switches as indicated and adjust the appropriate variable capacitor for 4 cm of vertical display. Where indicated, remove or replace test equipment fixtures and readjust the Type 106 Amplitude control for 4 cm of vertical display. (See Fig. 7-4 for location of these adjustments.)

NOTE

Capacitors shown in parenthesis are adjustments for the B Operation Amplifier.

TABLE 7-7

Z_i	Z_f	Adjust	Deflection
.001 μF	.0001 μF	C5512F (C5562F)	4 cm
.001 μF[1]	.01 μF	Type 106	4 cm
.0001 μF	.001 μF	C5512B (C5562B)	4 cm
10 pF	.0001 μF	C5512C (C5562C)	4 cm
.01 μF[2]	.001 μF	Type 106	4 cm
.001 μF	10 pF	C5512G (C5562G)	4 cm

[1]**Remove 50-ohm termination and 10× attenuator.**
[2]**Reinstall the 50-ohm termination and 10× attenuator.**

d. Disconnect the Type 106 signal from the O Unit. Turn the oscilloscope off, remove the plug-in extension and replace the O Unit in the oscilloscope. Turn the oscilloscope on, wait for about a two minute warm-up period, then adjust the OUTPUT DC LEVEL (refer to step 15d) before proceeding to the next step.

21. (Optional) Check Gain-Bandwidth Product

a. Set the front-panel controls as follows:

Type O Unit

VERTICAL DISPLAY	+DC
VOLTS/CM	.5

Operational Amplifier Controls (both chanenls)

Z_i SELECTOR	1 μF
Z_f SELECTOR	10 pF
INTEGRATOR LF REJECT	1 KC

Type 545B Oscilloscope

Stability	Fully Clockwise
Triggering Level	Fully Clockwise
Time/Cm	1 mSec

Type 191 Constant Amplitude Signal Generator

Frequency dial	Fully clockwise
Frequency Range	50 kHz Only
Amplitude	10
Variable	Cal
Amplitude Range	.5-5 V
Power	On

b. Connect the Output of the Type 191 through a 5 ns coaxial cable and a 50-ohm in-line termination to the O Unit EXT. INPUT connector. If necessary, adjust the Type 191 Cal and Amplitude settings to produce 2 cm of vertical deflection.

c. Change the Frequency Range switch of the Type 191 to 8-18 MHz.

CAUTION

Do not change amplitude settings of the Type 191.

d. Move the 50-ohm termination and coaxial cable from the EXT. INPUT connector to the A (or B) INPUT connector. Change the O Unit VERTICAL DISPLAY switch to +A (or +B) OUTPUT position.

e. Increase the output frequency of the Type 191 until the vertical deflection is at 2 cm. The output frequency of the Type 191 should be 15 MHz or greater.

NOTE

If there is an unstable ripple on the top and bottom of the display, measure between the average values of the ripple.

f. Disconnect the Type 191 signal from the O Unit. To calibrate the other Operational Amplifier repeat steps 15 through 21.

MAINTENANCE NOTES
TEXT ADDENDUM

TYPE O

Section 7 Calibration Procedure

Page 7-8, Step 19.

Second Paragraph:

CHANGE 'INTEGRATOR LF REJECT 1 CPS' to read;
'INTEGRATOR LF REJECT OFF'.

Step 19b.

CHANGE to read;

b. Without changing the Type 105 output, set Z_i to .1 µf, Z_f to .1 µf, and the INTEGRATOR LF REJECT switch to 1 CPS. With the VOLTS/CM switch set at 1 there should be 4 cm of crt display.

Page 7-9, Step 20.

Third Paragraph:

CHANGE 'INTEGRATOR LF REJECT 1 cps' to read;
'INTEGRATOR LF REJECT 1 CPS'.

This insert is placed in its appropriate position in your Product Reference Book and printed on colored paper to expedite retrieval. In a standard manual, it will be filed at the back of the manual.

SCHEMATICS

0

Publication:
NOT STOCKED
November 1963

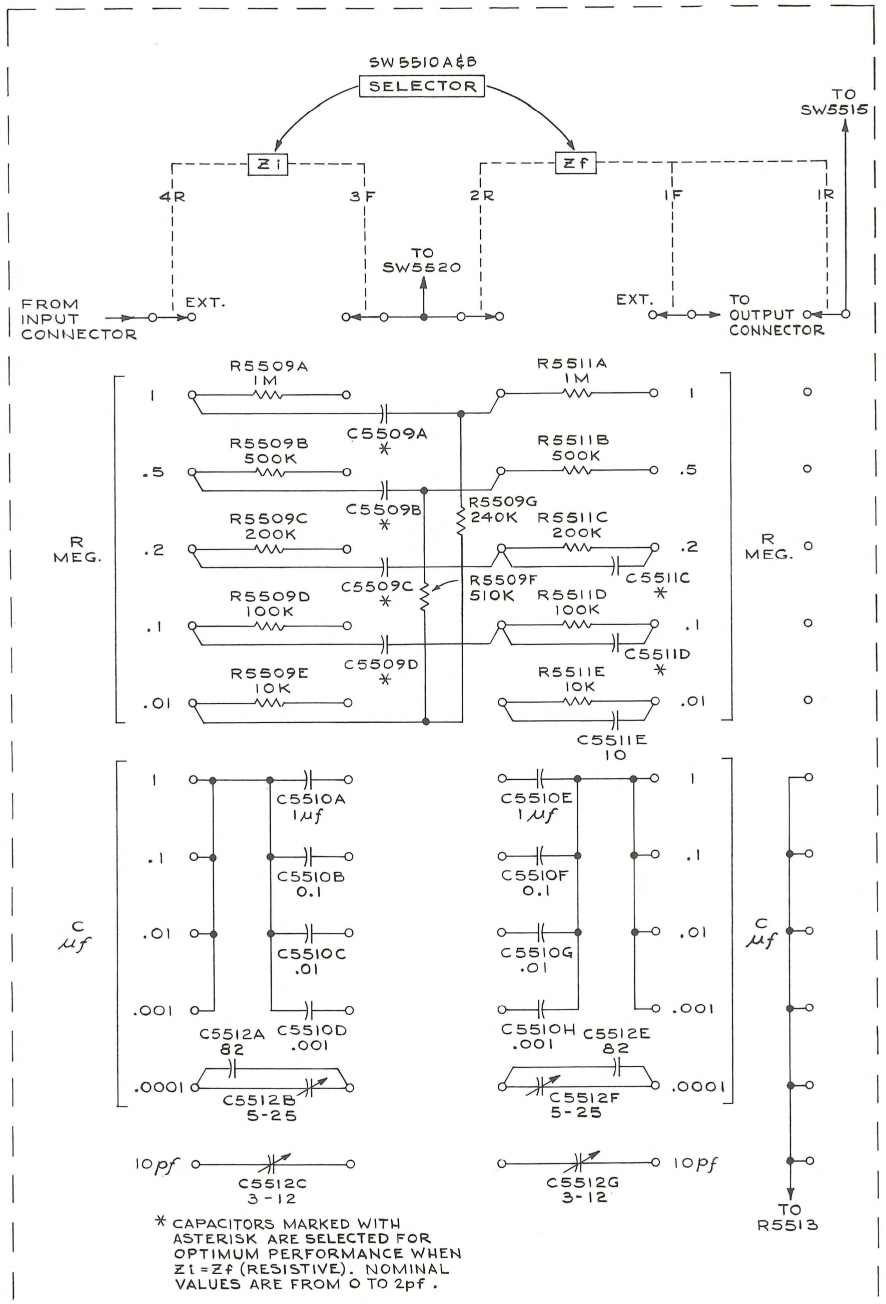
For all serial numbers.

CONTENTS:

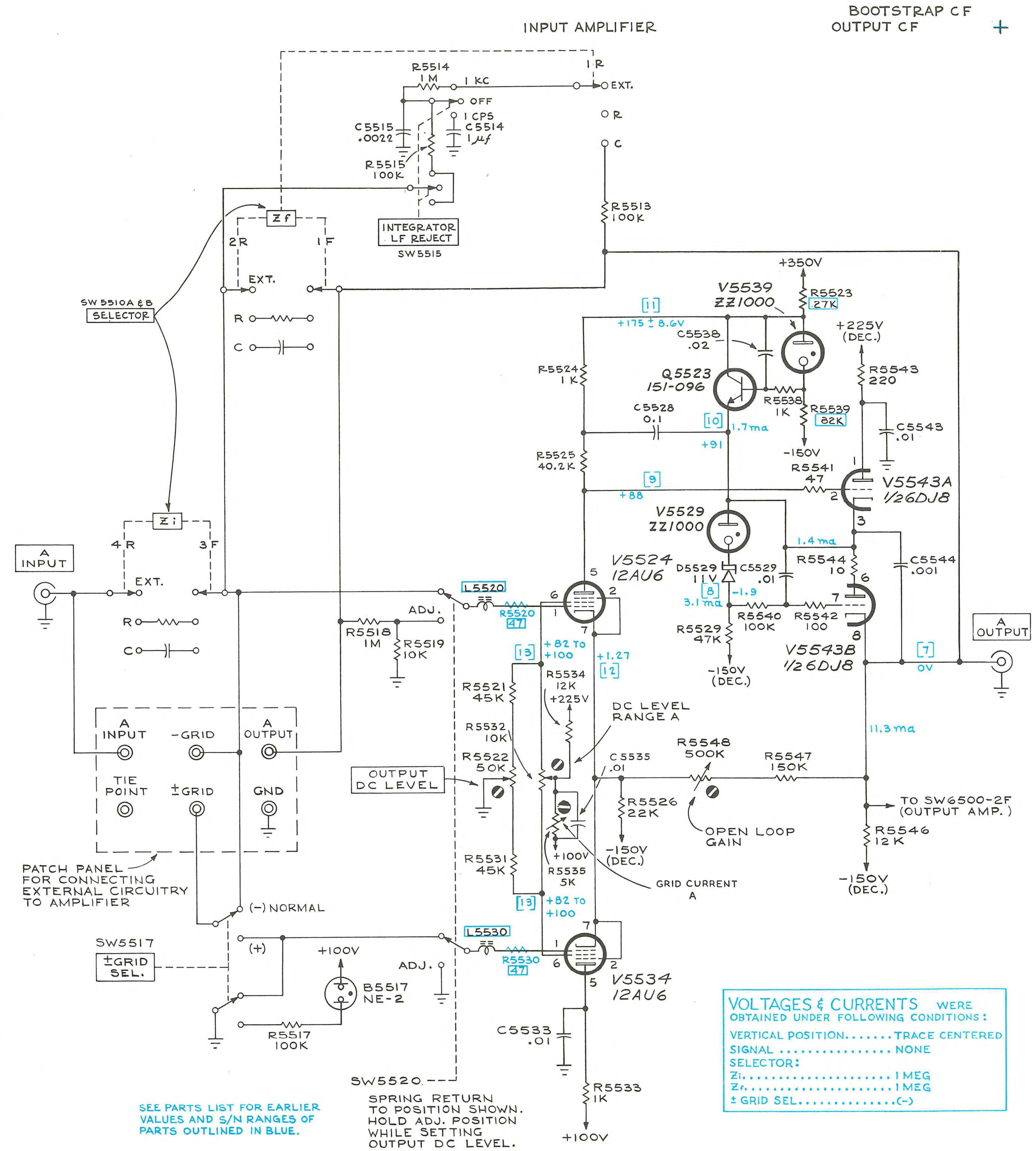
title	print symbol	schematic date
BLOCK DIAGRAM	A	2-19-62
CHANNEL A sn 101 to 813	B	2-16-62
CHANNEL A sn 814 up	A1	7-16-62
CHANNEL B sn 101 to 813	B	11-22-61
CHANNEL B sn 814 up	A1	7-22-62
PREAMPLIFIER	B	1163

ABBREVIATIONS:

cer	ceramic
comp	composition
emc	electrolytic, metal cased
gmv	guaranteed minimum value
h	henry
k	kilo (10^3)
k	kilohm
m	milli (10^{-3})
ma	milliamp
meg	megohm
mh	millihenry
mpt	metalized, paper tubular
mt	mylar, tubular
mv	millivolt
μ	micro (10^{-6})
μf	microfarad
μh	microhenry
μsec	microsecond
n	nano (10^{-9})
nsec	nano second
Ω	ohm
p	pico (10^{-12})
pbt	paper, "bathtub"
pcc	paper covered can
pf	picofarad ($\mu\mu$f)
piv	peak inverse voltage
pmc	paper, metal cased
poly	polystyrene
prec	precision
pt	paper, tubular
ptm	paper, tubular molded
sn or S/N	serial number
tub	tubular
v	working volt, dc
var	variable
w	watt
WW	wire wound



DETAILS OF Zi - Zf SELECTOR SWITCH



TYPE O PLUG-IN UNIT

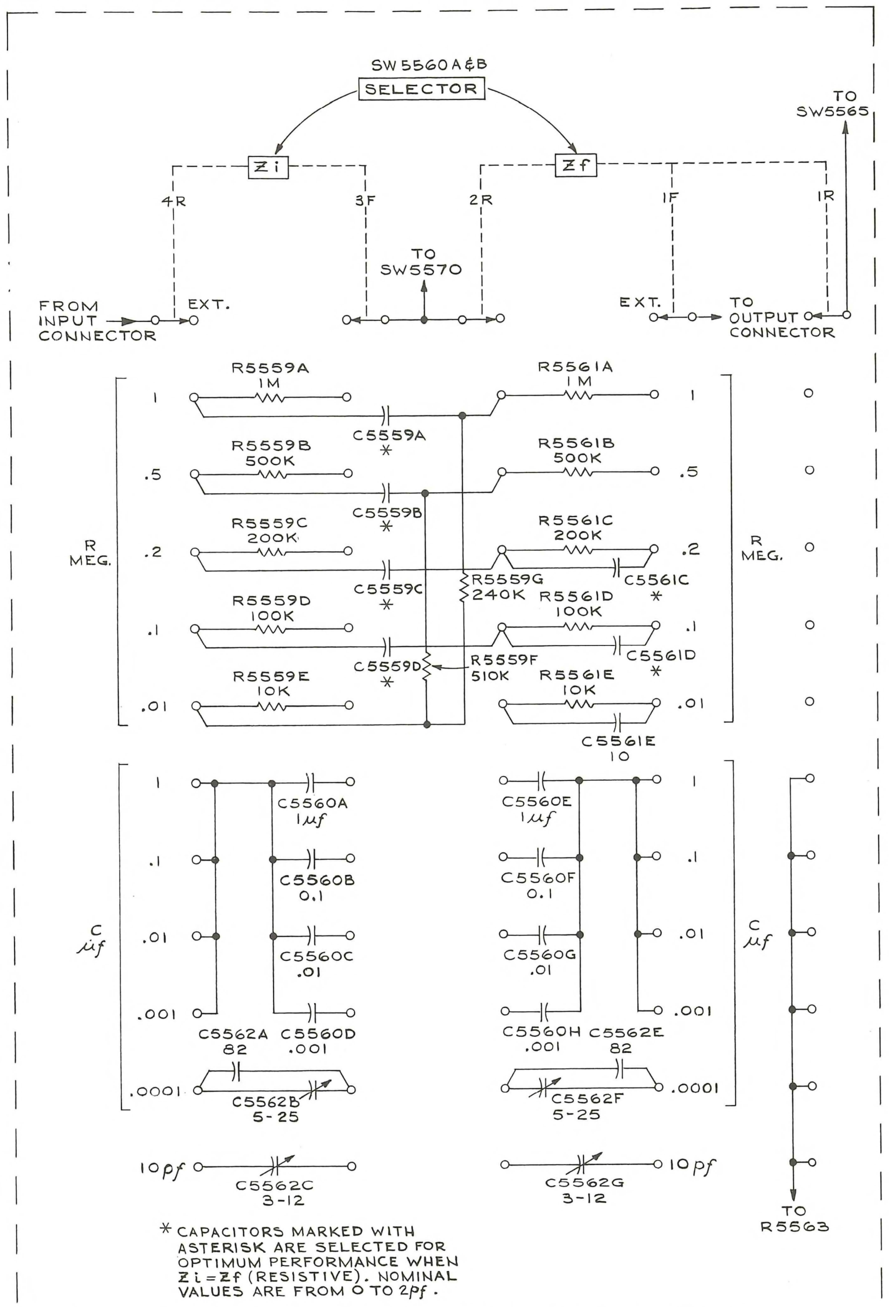
B1

MRH
566

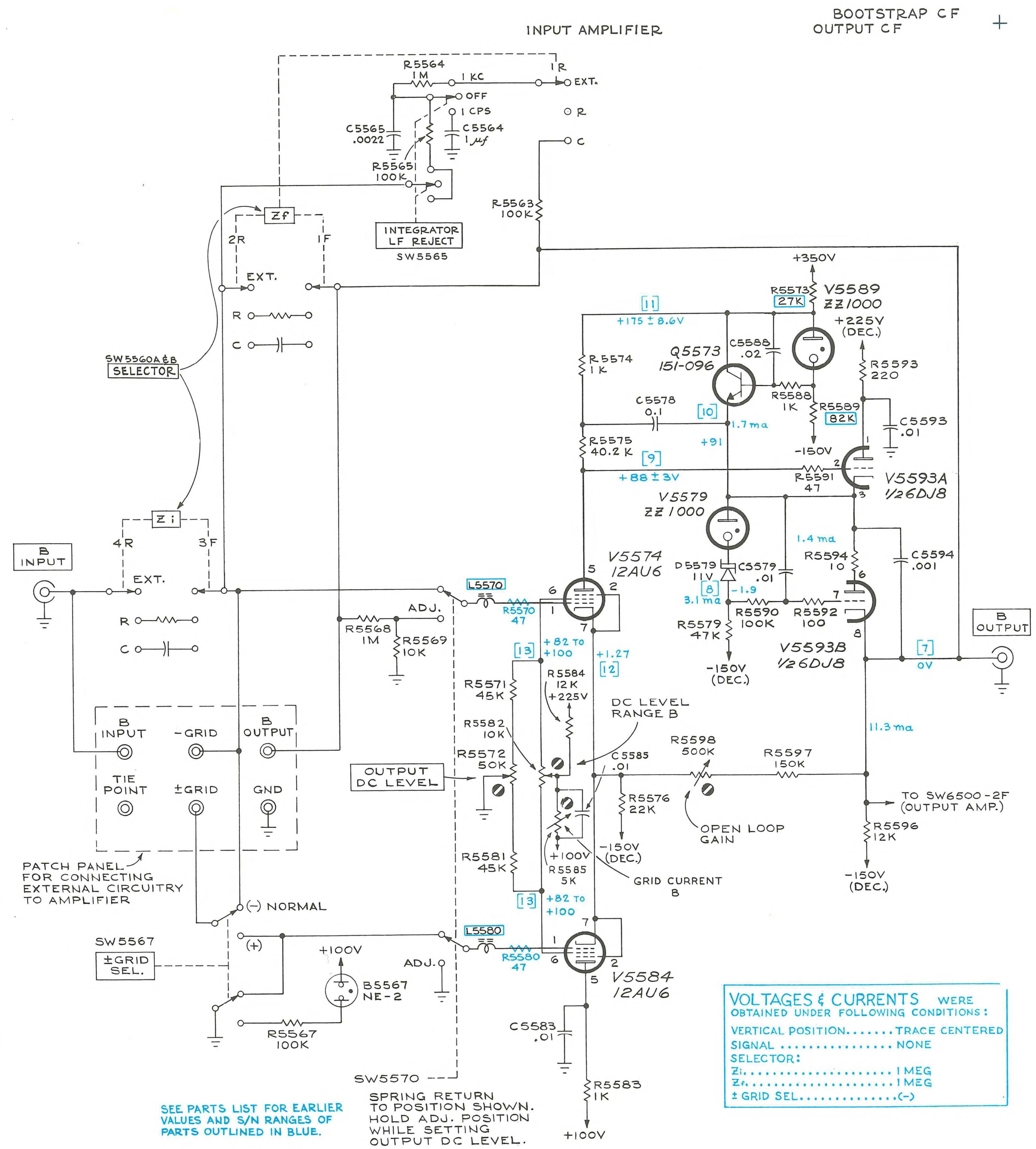
CHANNEL A
OPERATIONAL AMPLIFIER

EFF. S/N 814 & UP

S/N 814-up



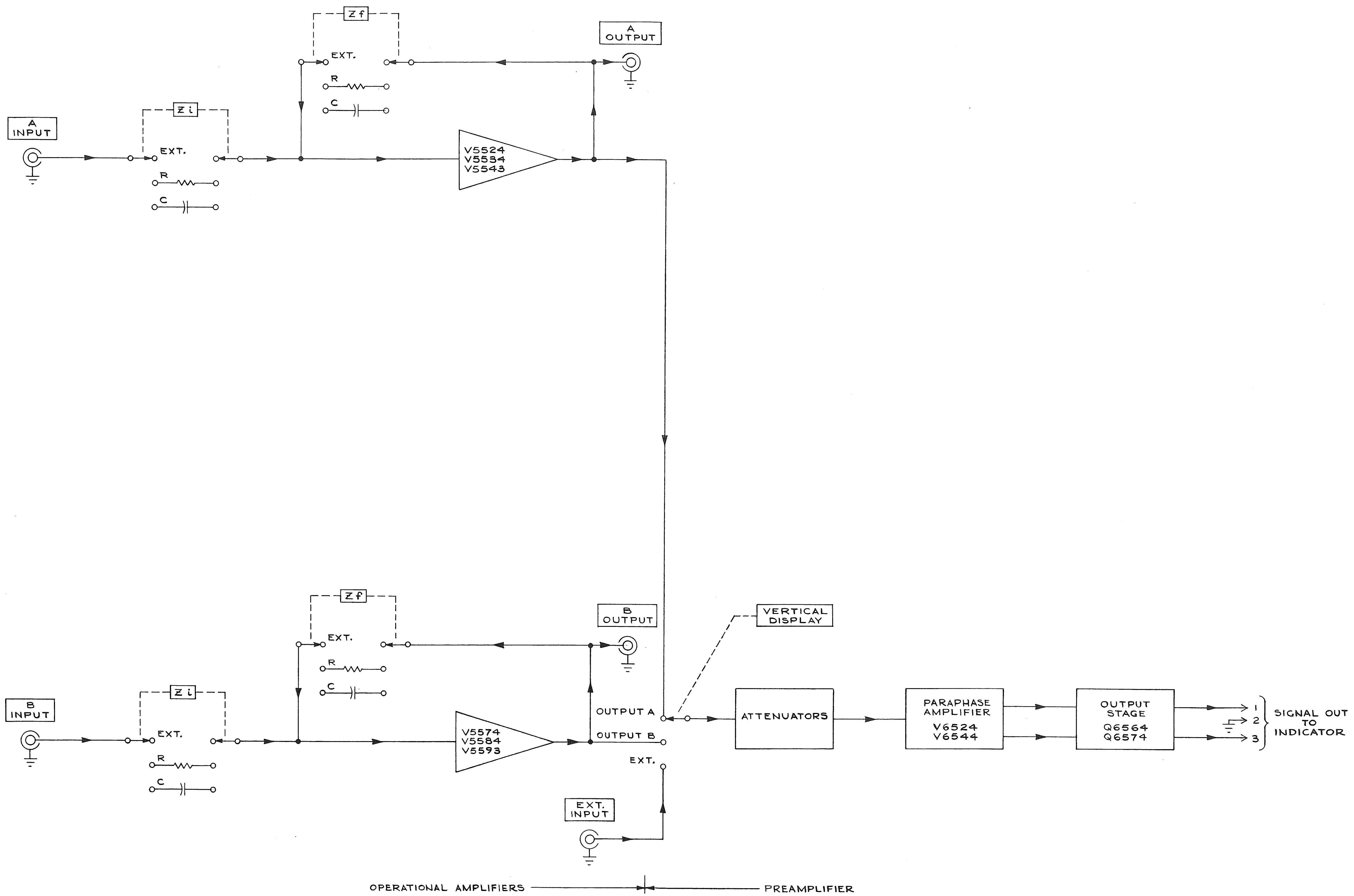
DETAILS OF Zi-Zf SELECTOR SWITCH



MRH
566

CHANNEL B
OPERATIONAL AMPLIFIER
EFF. S/N 814 UP

CHAN B AMP
S/N 814-up

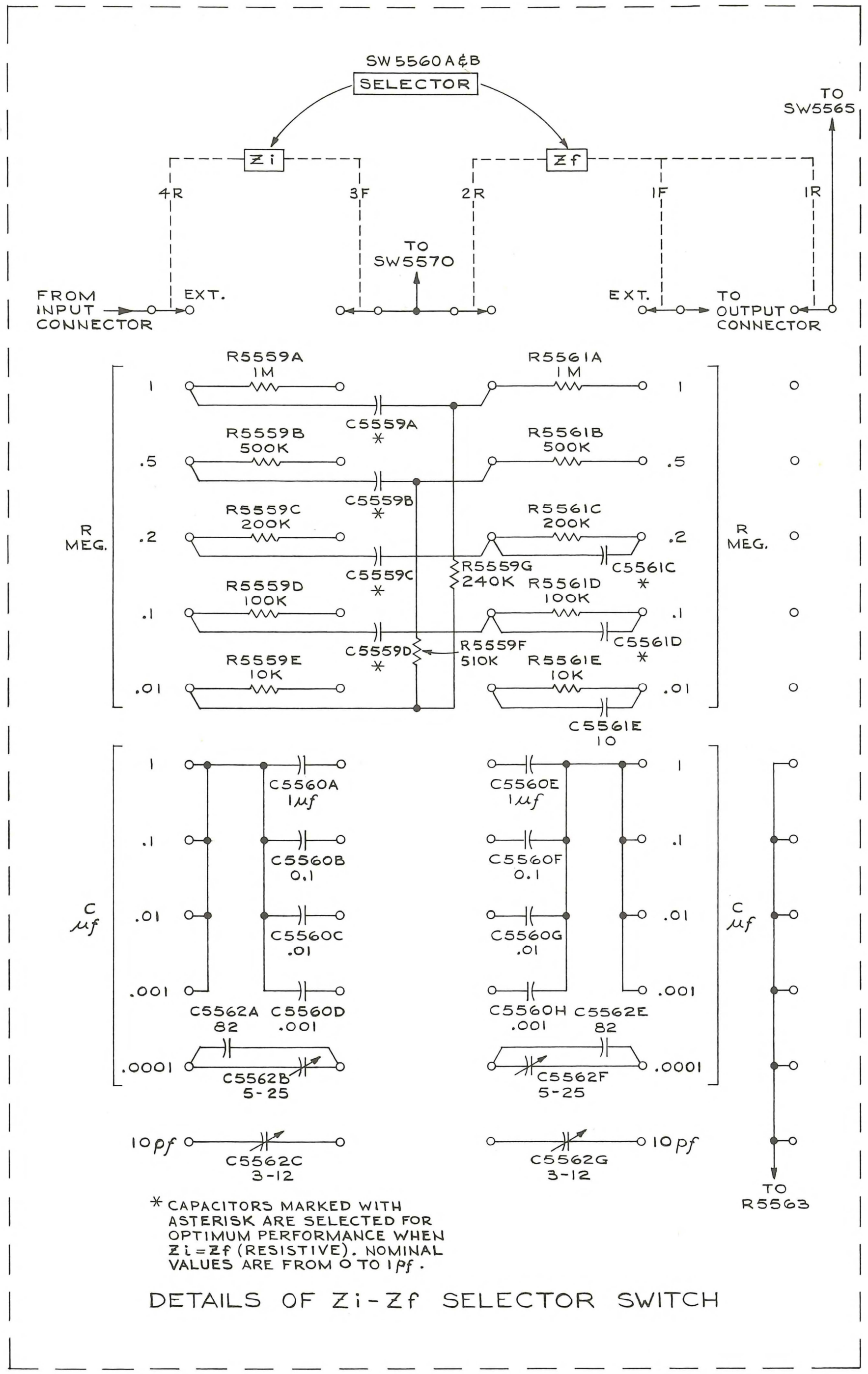


MRH
2-19-62

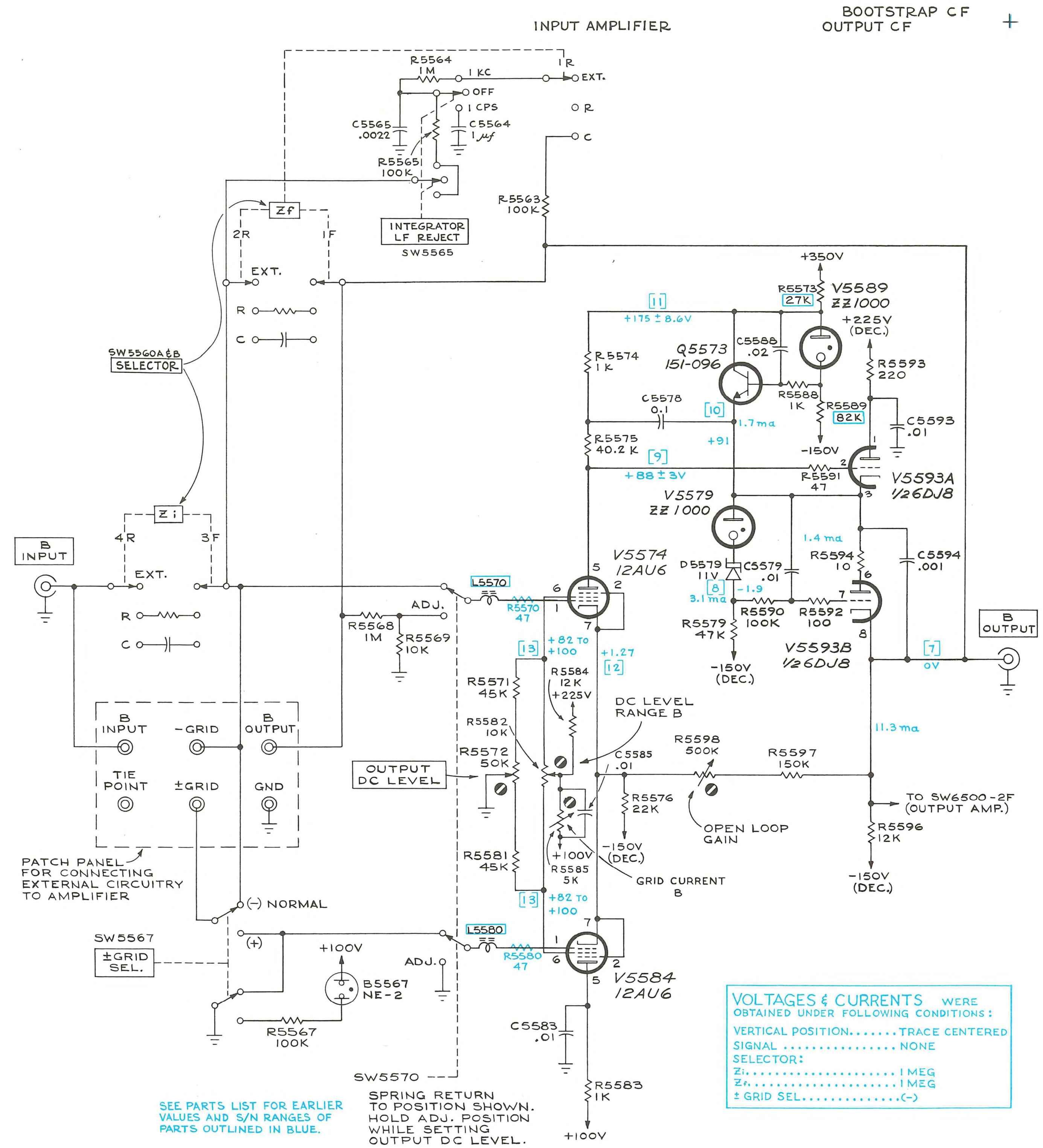
TYPE O PLUG-IN UNIT

A

BLOCK DIAGRAM



DETAILS OF Zi-Zf SELECTOR SWITCH

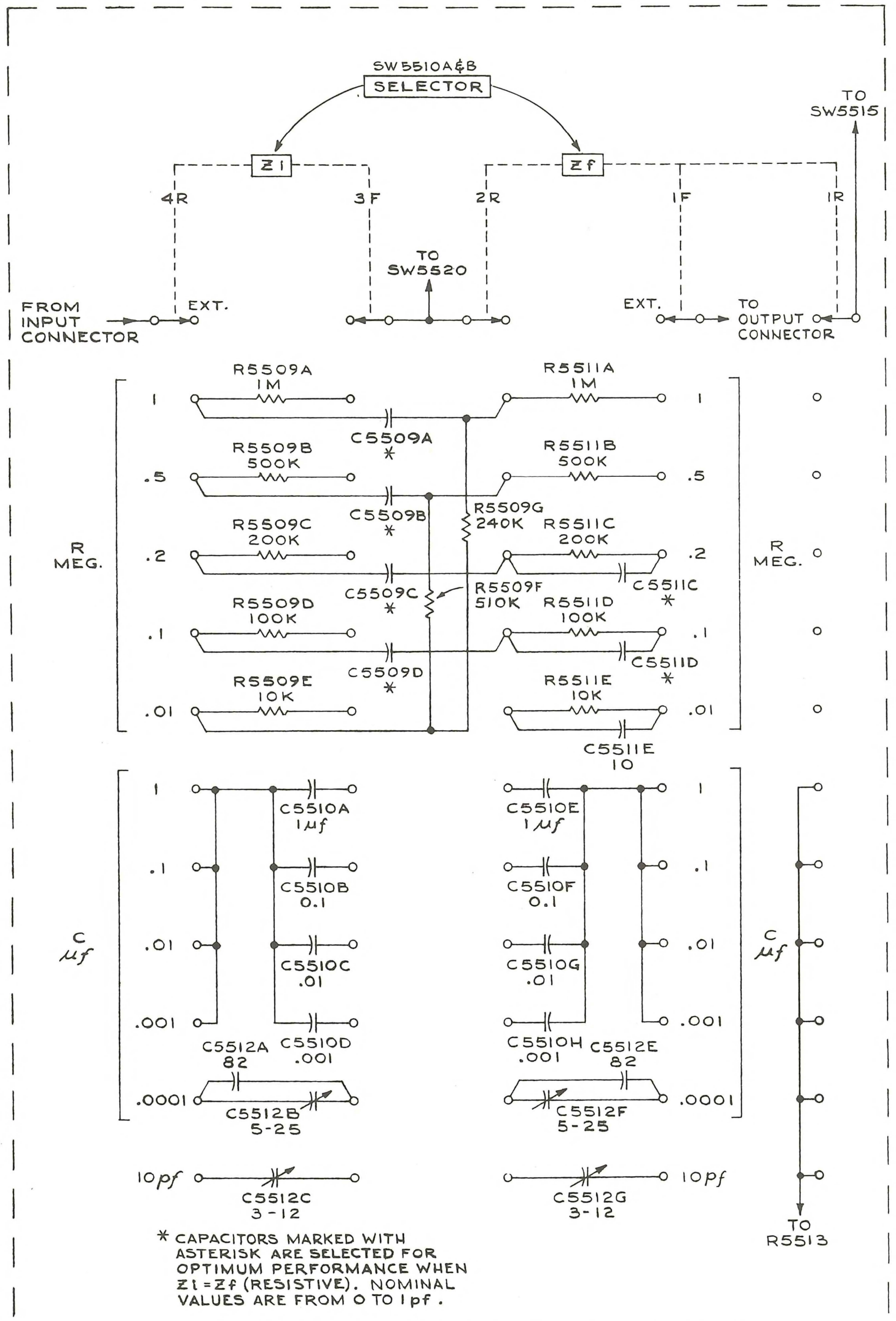


VOLTAGES & CURRENTS WERE OBTAINED UNDER FOLLOWING CONDITIONS:

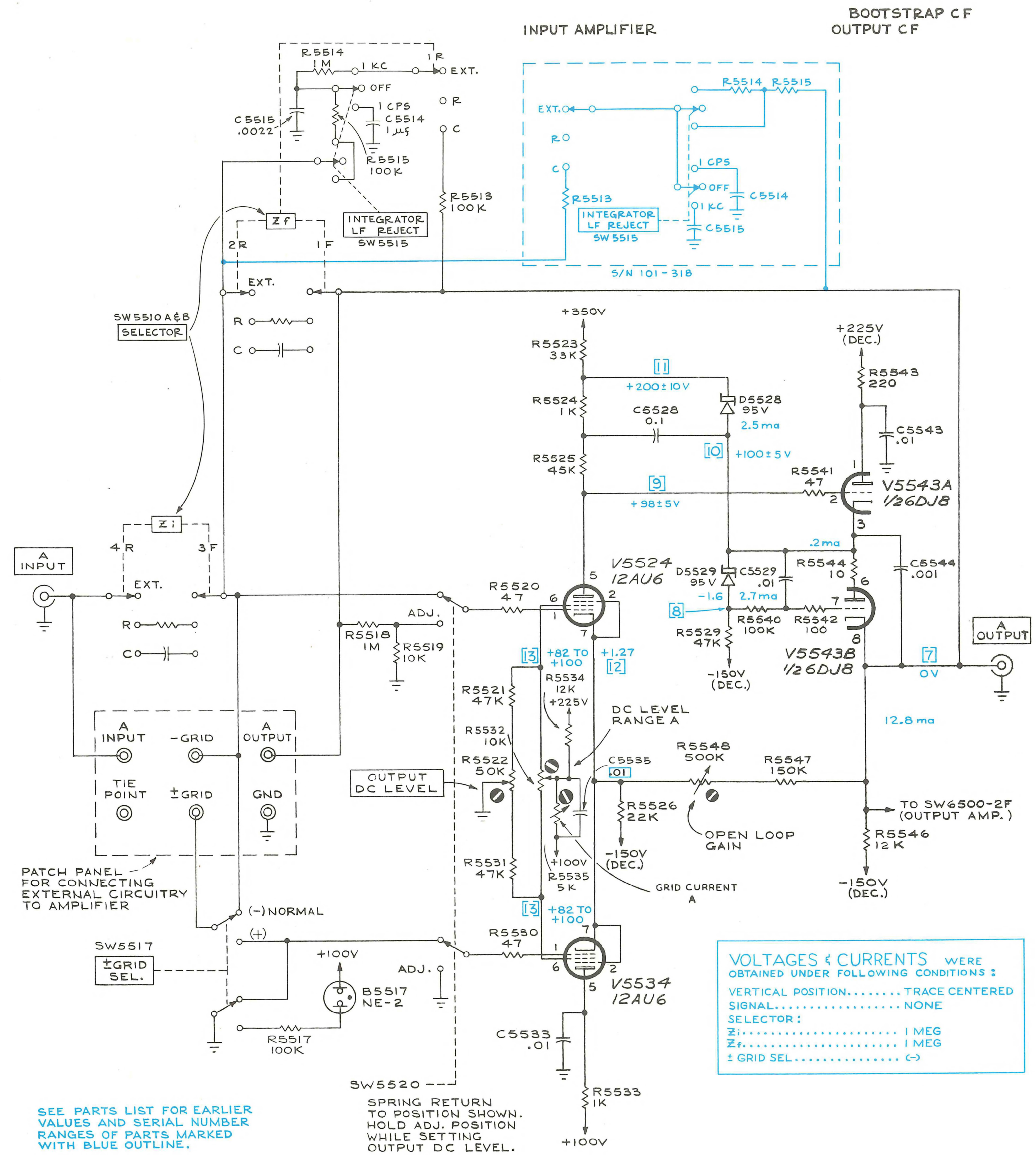
VERTICAL POSITION.......TRACE CENTERED
SIGNAL ....NONE
SELECTOR:
Zi......................1 MEG
Zf......................1 MEG
± GRID SEL..............(-)

MRH
566

CHANNEL B
OPERATIONAL AMPLIFIER
EFF. S/N 814 UP



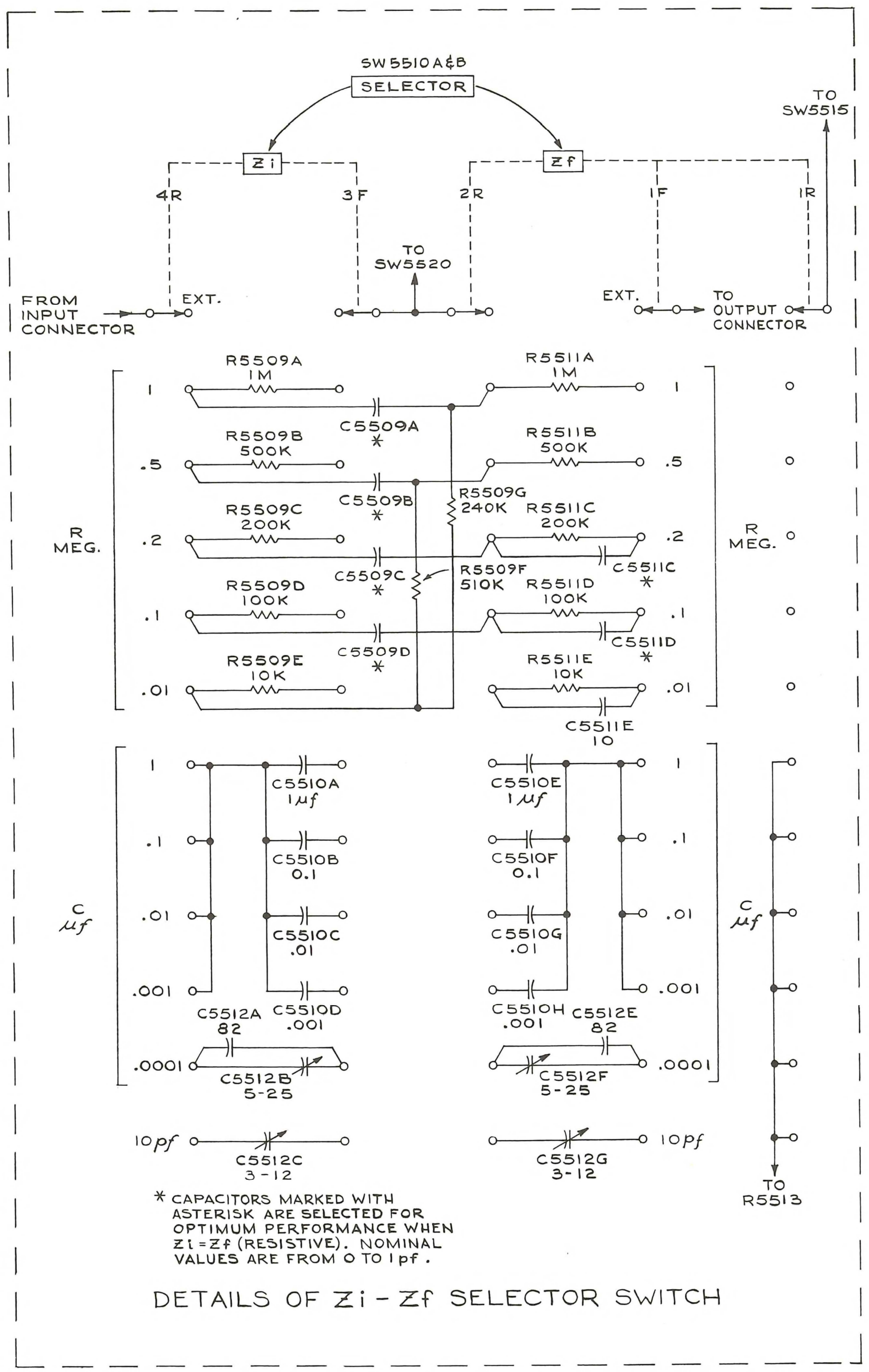
DETAILS OF Zi - Zf SELECTOR SWITCH

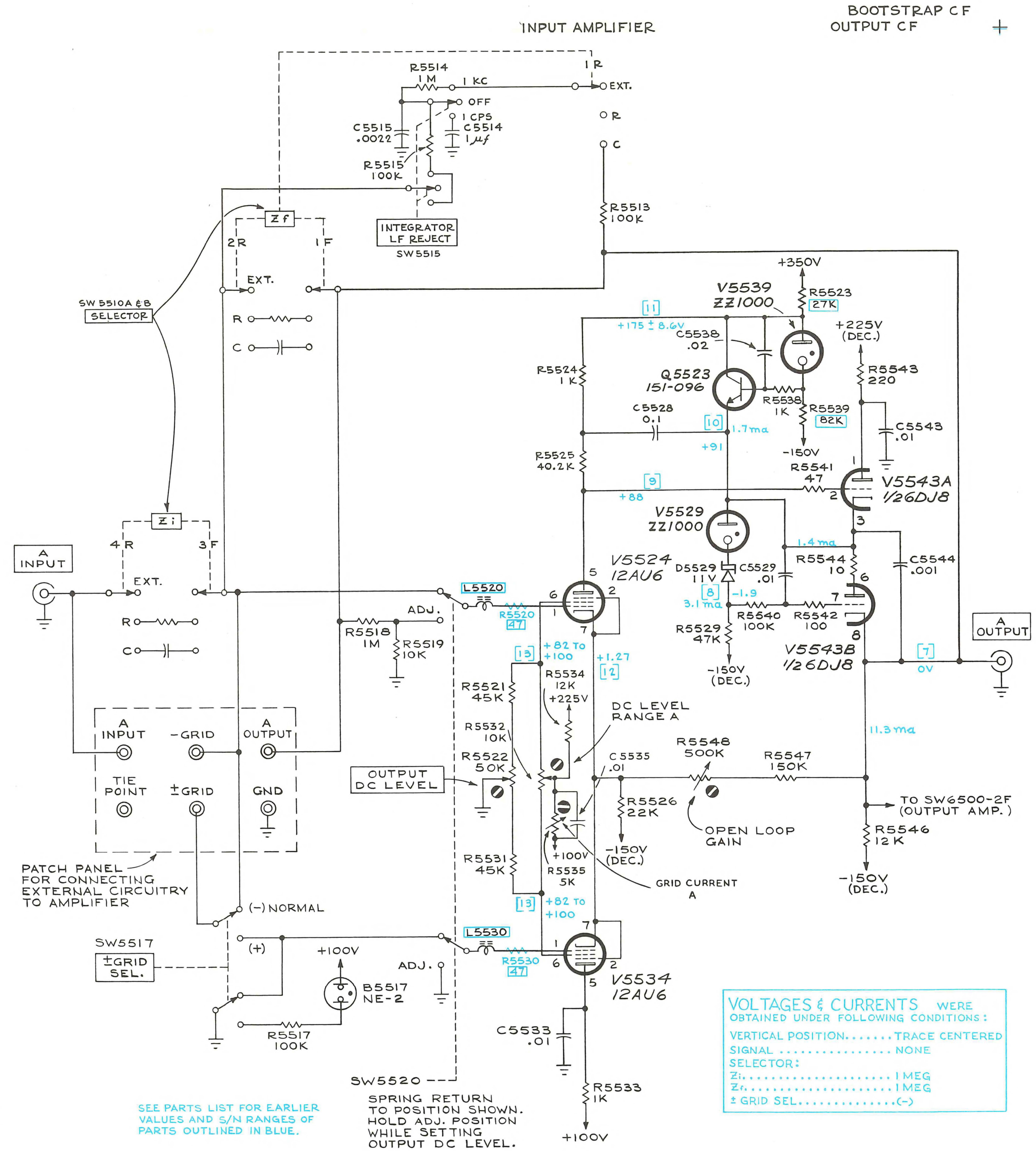


MRH
364

CHANNEL A
OPERATIONAL AMPLIFIER
EFF. S/N 101-813

CHANNEL A

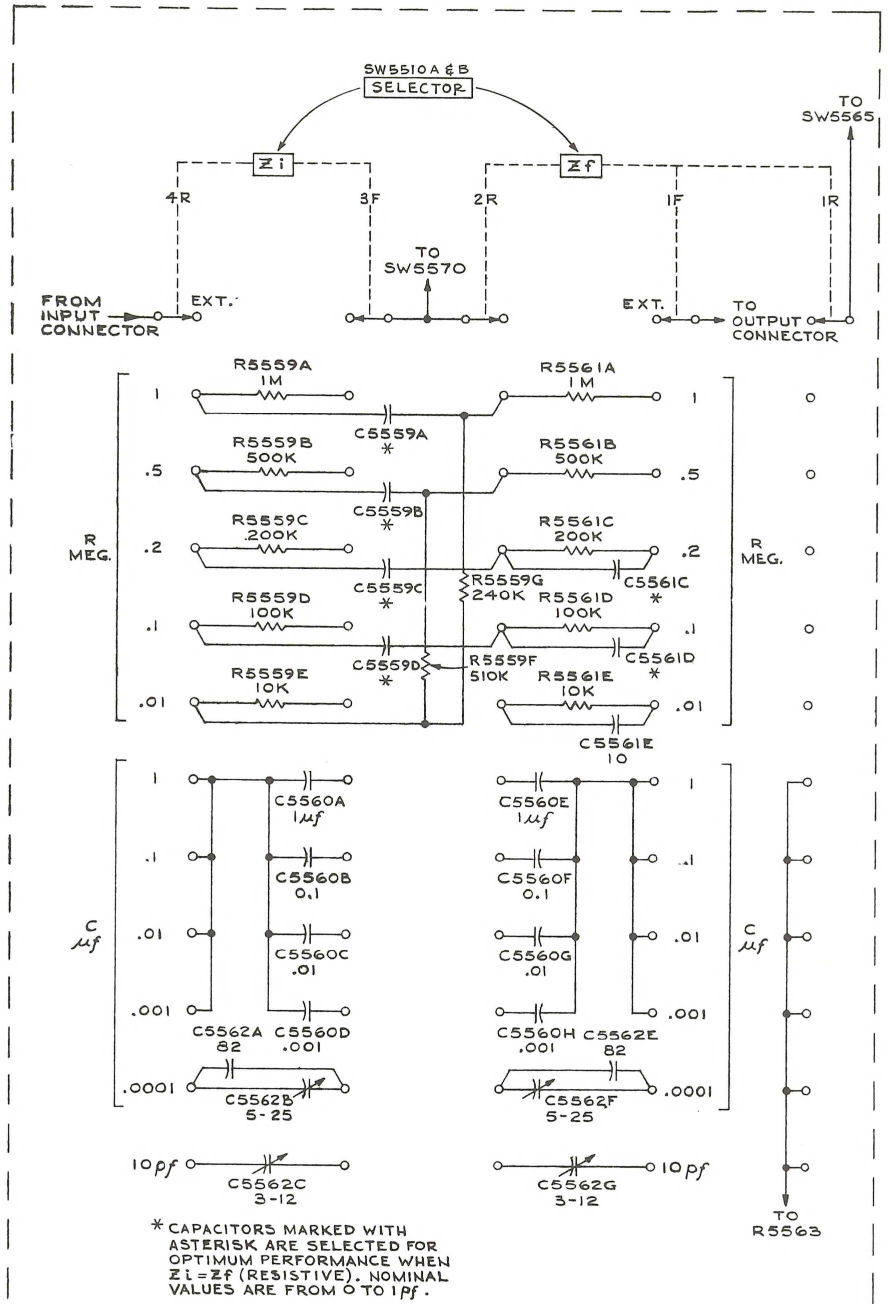




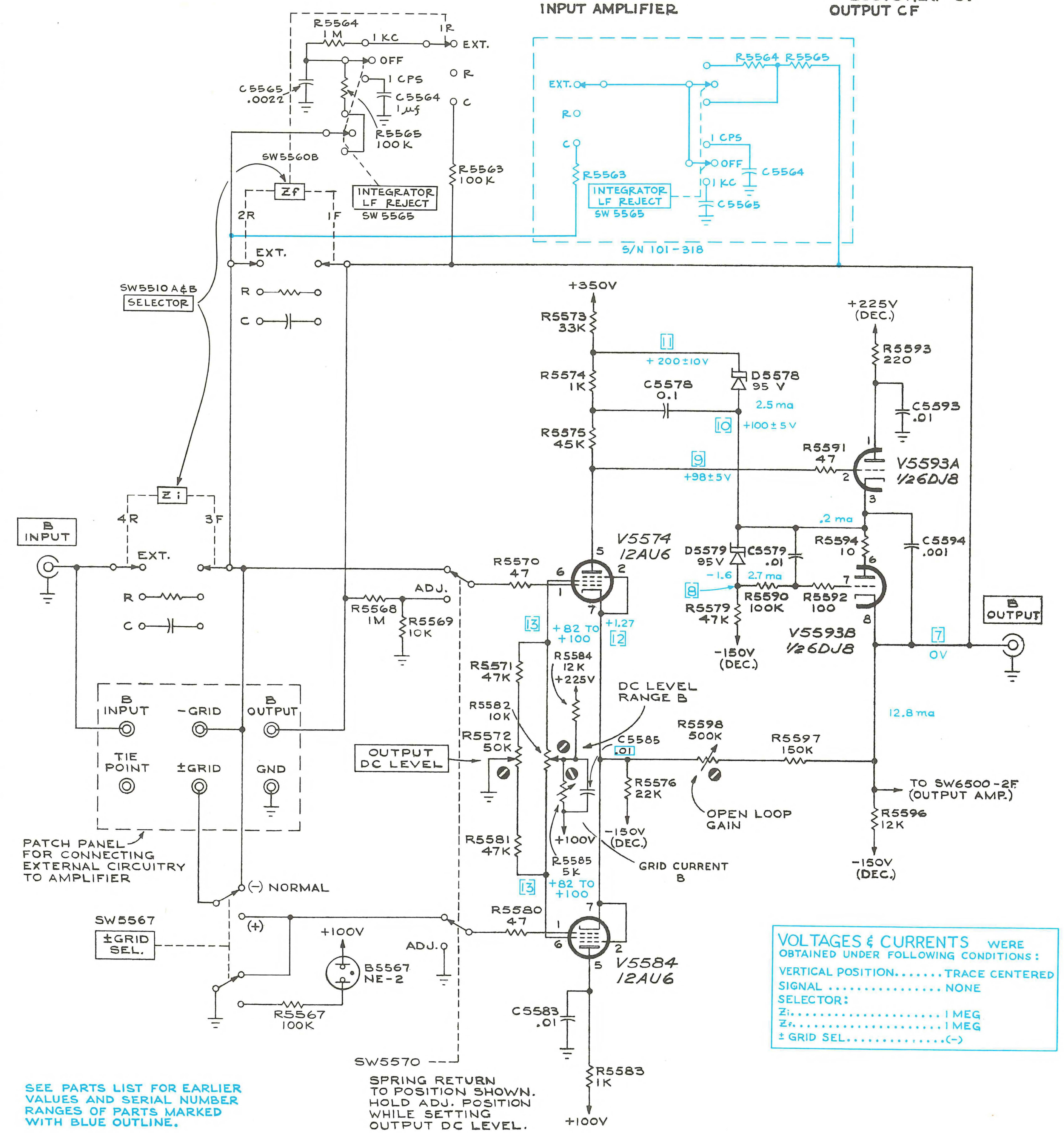
MRH
566

CHANNEL A
OPERATIONAL AMPLIFIER
EFF. S/N 814 & UP

CHANNEL A



DETAILS OF Zi-Zf SELECTOR SWITCH



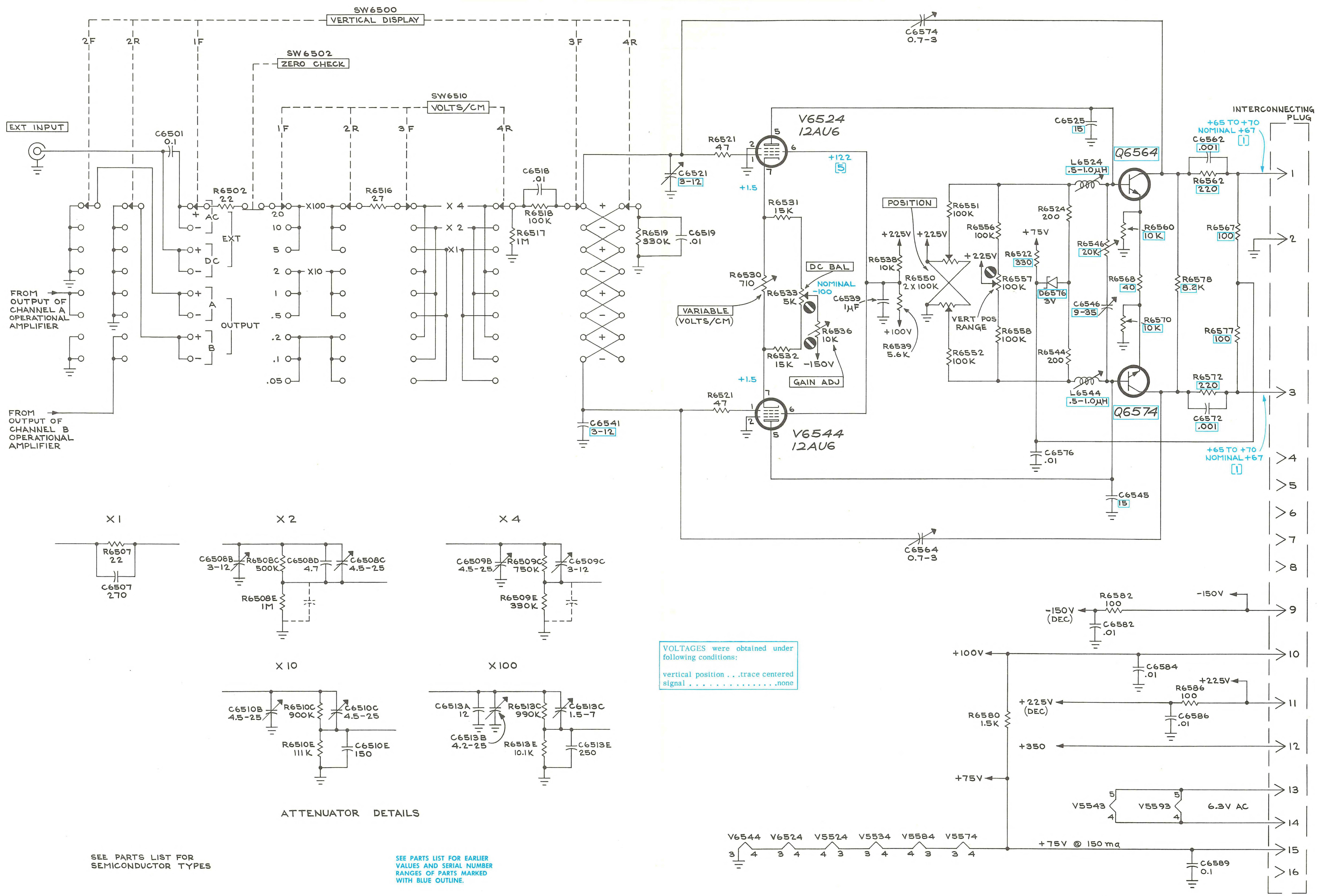
MRH
364

TYPE O PLUG-IN UNIT

B2

CHANNEL B
OPERATIONAL AMPLIFIER
EFF. S/N 101-813

CHANNEL B



SEE PARTS LIST FOR SEMICONDUCTOR TYPES

SEE PARTS LIST FOR EARLIER VALUES AND SERIAL NUMBER RANGES OF PARTS MARKED WITH BLUE OUTLINE.

1165

TYPE O PLUG-IN UNIT

A1

PREAMPLIFIER
S/N 2950 & UP

PREAMPLIFIER