Title
DO-IT-YOURSELF"" LOCK-IN AMPLIFIERS, RADIO AND AUDIO FREQUENCY

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ABSTRACT

We describe two specific lock-in amplifiers, one designed for radio frequency, the other for audio. These amplifiers were developed for specific applications, but subsequent experience has indicated that they offer promise in others. In view of the general utility of these amplifiers, we have included detailed schematics, photographs, templates, and other data useful in duplicating either unit. Both have been built and tested by chemists with limited electronics backgrounds. Proper testing requires an oscilloscope and a signal generator. Design philosophy is presented in detail for those who wish to extend performance or design their own.

The audio-frequency lock-in amplifier was designed for an ESR spectrometer, locking a klystron's frequency to that of a resonant cavity. The lock-in unit is now being adapted to NMR work. More specifically, it is to be part of an NMR fluxmeter, locking the frequency of an rf oscillator to the resonant frequency of a paramagnetic sample. Another application involves automatic frequency tracking when the magnetic field is varied. The af unit was designed around commercial circuit modules, so that much of its circuitry need not be fabricated. Performance compares favorably with specifications published for some commercial units. It is designed for 40 to 20,000 hertz, but its range is easily extended. A small preamplifier for extending sensitivity to microvolt levels is also described.

The radio-frequency lock-in amplifier was designed for measuring differential pressures on the order of 0.1 micron with an accuracy of several percent. In achieving the design objective, we found that we were able to detect capacitance changes on the order of 10^{-18} farads. Our rf unit operates at 2.7 MHz but can be adapted to other radio frequencies. It uses Nuvistor tubes to minimize the instability problems normally encountered at radio frequencies.

I. INTRODUCTION

This report describes two specific lock-in amplifiers,[1] one designed for radio frequency (2.7 MHz), the other for audio. Although these lock-ins were originally developed for specific applications, subsequent experience has indicated that they offer promise in other applications (see Secs. III. F and IV. F). With this in mind, we have included photographs, templates, and other data useful in duplicating either unit. Both units have been built and tested by chemists with limited electronics backgrounds. Proper testing requires an oscilloscope and a signal generator. Design philosophy is emphasized for those who wish to extend performance or to design their own.

High-performance lock-in units are commercially available, but in many cases the price is prohibitive for those experimenters with limited funds. Furthermore, some of those units incorporate features unnecessary to a given application, and the additional complexity is undesirable. Also interesting, commercial lock-ins do not encompass the entire radio-frequency field. In some cases, therefore, one has no alternative but to design his own. In any event, the experience gained in building and trouble-shooting one's own lock-in is invaluable in understanding the instrumentation requirements generally associated with it.

Concerning the audio-frequency lock-in (af unit), its principal merit is simplicity. This simplicity stems from its construction; i. e., it is designed around commercial circuit modules, so that much of its circuitry need not be fabricated. The unit is solid-state, and includes its own power supply. The components cost approximately $200. Also described is a small preamplifier for extending sensitivity to microvolt levels (uses two circuit modules). Performance compares favorably with specifications published for some commercial lock-in amplifiers, but ours is less flexible in selecting frequency. It is best used in applications in which frequency changes are seldom made. but it can be adapted to variable-frequency operation. Fixed-frequency operation is quite suitable for many applications, however, and reduces complexity of the instrument. Some experimenters will probably choose to establish frequency and sensitivity requirements with a commercial lock-in unit before building their own.

The radio-frequency lock-in amplifier (rf unit) does not employ circuit modules, but uses Nuvistor tubes instead. Nuvistors, which are much smaller than conventional tubes, offer advantages at radio frequencies, (as discussed in Sec. IV. A). In some ways, the rf unit is easier to build than the af unit. Component cost for the rf unit is approximately $100.

II. GENERAL CONSIDERATIONS

Lock-in amplifiers are discussed elsewhere,[2] so this discussion will be somewhat limited. Basically, a lock-in amplifier does much the same thing as a tuned amplifier, but it does it better. The principal difference is that lock-in amplifiers are phase-sensitive. Also, they can operate at bandwidths which are very much narrower than are possible with a conventional tuned amplifier operating at the same frequency. The reasons that lock-in amplifiers can operate at such narrow bandwidths are (a) the information ("signal") sought is amplitude modulated by a reference, usually of audio frequency or higher; (b) the reference also gates a synchronous detector that responds to the gating frequency only; (c) if the reference frequency changes, the gating changes in correspondence, so that the lock-in always remains "in tune."

Lock-in techniques are commonly used to recover signals 40 db below the ambient noise level. These techniques are very effective because: (a) they minimize noise generated by the amplifying devices used;[3] (b) they permit the use of much narrower bandwidths than are usually possible, reducing white noise[4] associated with the signal inversely as the square root of bandwidth (the degree to which bandwidth can be reduced is related to the highest frequency of the information sought); and (c) they discriminate against noise at the "tuned" frequency, but of random phase.

Lock-in techniques minimize noise generated by the amplifying devices used because: (a) the signal is modulated to translate its spectra from a band centered around zero frequency to a band centered about a higher frequency (the modulation frequency);[5] (b) most amplifying devices have noise spectra that vary as the reciprocal of frequency;[2] (c) translation to a higher frequency moves the signal to a frequency where less noise is generated by the amplifying devices used. In general, the modulating frequency should be greater than 100 hertz when vacuum-tube amplifiers are used, and greater than 1000 hertz when transistor amplifiers are used.[6]

III. AUDIO-FREQUENCY LOCK-IN AMPLIFIER

A. General Considerations

Principal components of our lock-in amplifier are shown in the block diagram of Fig. 1. The modulator, which is external to the af unit, takes many forms and can be electrical or mechanical.

The af unit uses three linear-amplifier modules, for which detailed specifications from the manufacturer are included in Appendix I. A. One amplifier in conjunction with a network comprises the oscillator. Another type of module includes three emitter-followers in one package. Only two are used, but the extra one might prove useful to some experimenters.

The frequency-determining elements, a parallel-T for the first amplifier stage and a network for the oscillator, are fabricated into blank containers supplied by the module manufacturer. Plug-in elements facilitate frequency change, but some experimenters may prefer switches.

B. Circuit Description

NOTE: Refer to the schematic diagrams, Figs. 2 and 3.

1. Signal Amplifier

The signal amplifier comprises two stages, the first one tuned to prevent noise and spurious pickup (such as 60 hertz) from saturating it. Bandwidth of the system, however, is normally determined by the time constant (Fig. 1) following the detector.[7]

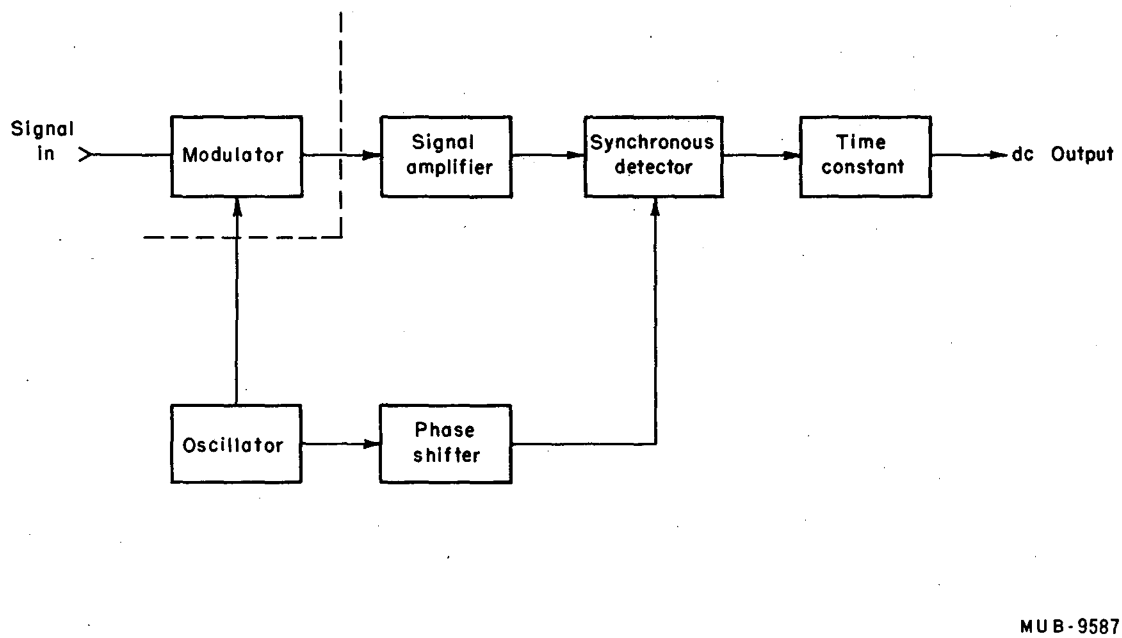


Fig. 1. Block diagram of the af unit. The modulator, which is external to the af unit, takes many forms, electrical or mechanical.

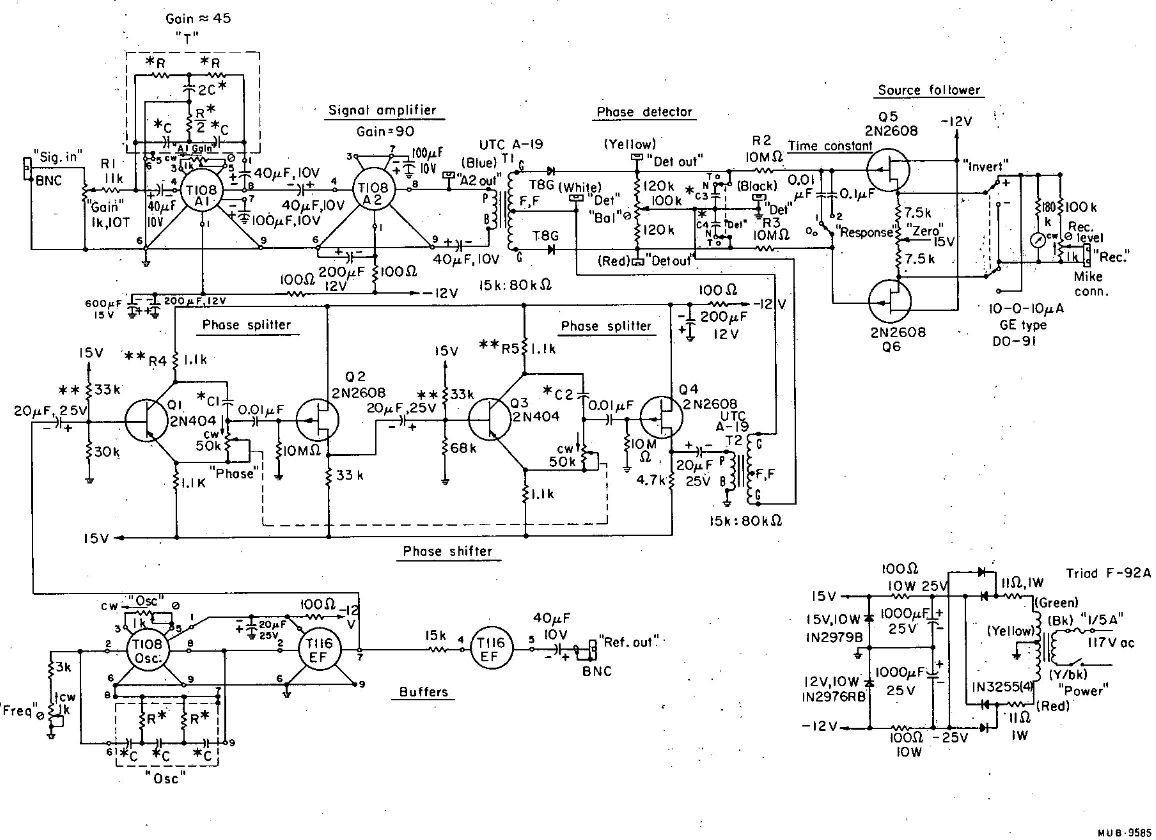
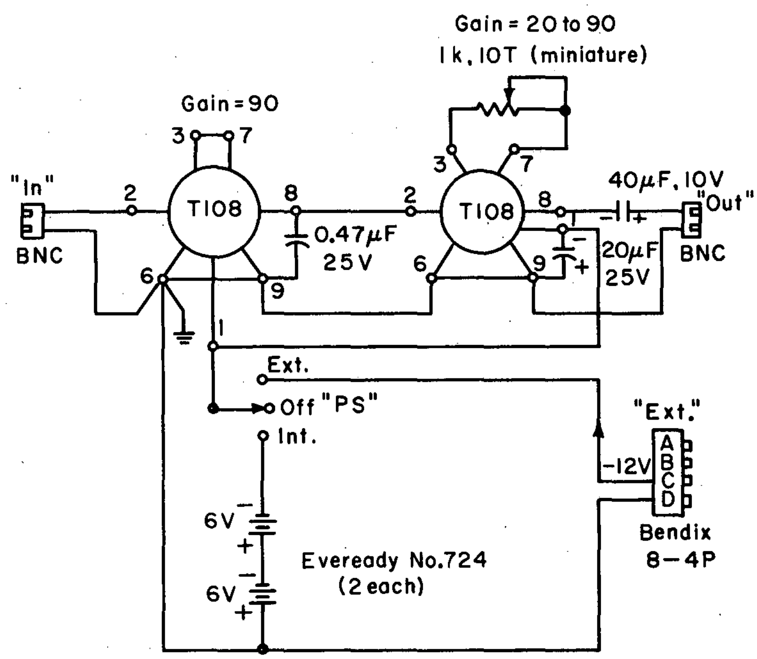
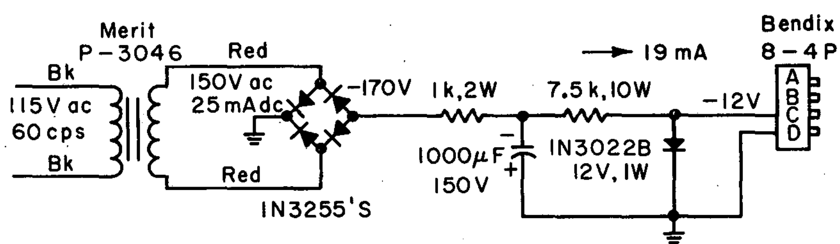


Fig. 2. Schematic diagram of the af unit. Module specifications are given in Appendix I. A. Module grounds should be made as indicated to avoid ground loops. All resistors are $\frac{1}{4}$ W, 5% carbon, unless indicated otherwise. Phase control is a 2 gang, 2 W, Ohmite CCU-5031. All capacitors not polarized are Mylar. BNC connectors are insulated from ground, DAGE 4890-1. Components indicated by * are frequency sensitive; those indicated by ** may require special attention.



(a)



(b)

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Fig. 3. (a) Schematic diagram of audio-frequency preamplifier. BNC connectors are DAGE 4890-1. First stage bypass (pins 8 to 9) is subject to frequency. (b) Power supply for preamplifier.

Frequency characteristics of the first stage are achieved with negative feedback through a parallel-T network. Characteristics measured when the af unit is tuned to 560 hertz are indicated in Fig. 4. Asymmetry is not due to the network per se, but to the divider action of R1 and the network at high frequencies (Fig. 2). Figure 4 indicates that the high-frequency response falls off more than the gain of A1 (Fig. 2). One would expect the relative response to change by no more than the gain of the module used.[8] Additional rolloff at high frequencies is due to the divider action of R1 (Ref. 9) and the network, which reduces the signal actually applied to the module's input. Asymmetry is not objectionable, however; the reduced response at high frequencies enhances lock-in performance (by reducing noise) even more than if it were symmetrical.

Design equations for the parallel-T network are:[10]

$$b = k/(k+1) \tag{1}$$

$$RC = 1/\omega_o , \tag{2}$$

where the parameters represented are indicated in Fig. 5, R_s and R_L are the source and load impedances, respectively, and ω_o is 2π times the frequency to which the parallel-T is tuned.

The value of R (Fig. 5) must be compatible with the source and load impedances used with the parallel-T. If it is not, the values of beta (percentage feedback) at frequencies remote from the center frequency might be too small, and attenuation would accordingly be reduced to undesired frequencies. The value of R is frequently chosen with symmetry considerations in mind,[11] but we are not concerned with that effect, as indicated above. Our principal concern is that beta be reasonably high at frequencies far from the center frequency. Referring to the equivalent circuits of Fig. 6, we see that the values of beta can be computed if the network, source, and load parameters are known. We chose an R of 3 kilohms, and a k of 1. In Fig. 2, the load presented to the parallel-T is determined by the input impedance of A1 and any resistance in shunt with its input. The shunt resistance is always greater than R1. Assuming 11k for R1, 700 Ω for the source impedance (the measured output impedance of a T-108 module), and 30k for the module's input impedance (see manufacturers' specifications, Appendix I. A), we obtain betas of 0.55 and 0.50 at the low and high frequencies, respectively.

In view of asymmetry due to the divider action of R1 and the network, one might consider reducing R (Fig. 5) so as to obtain a larger beta at low frequencies. A simple calculation, however, reveals that R must be reduced considerably in order that beta be increased appreciably. A reduced R demands a proportionate increase in capacitor sizes; however, it is sometimes difficult to fabricate larger capacitors into the blank containers used. At any rate, if one understands the requirements imposed upon R, he can design a network with properties suitable to the circuit used.

The value of R is chosen to compromise opposing influences; that is, a change in R that increases beta at low frequencies reduces beta at high frequencies, and vice versa. This explains why R is usually chosen to be the geometric mean of the load and source impedances.

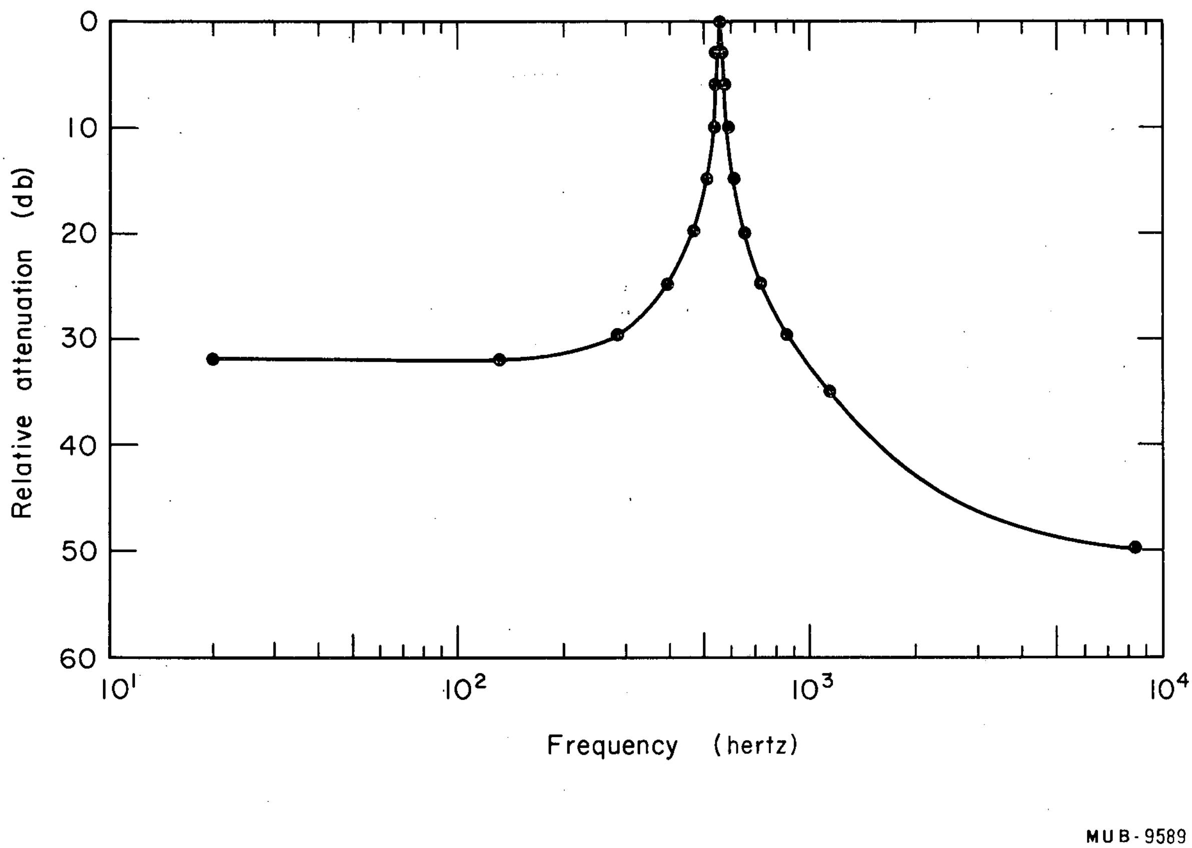

Fig. 4. Frequency characteristics of A1 (Fig. 2) with parallel-T tuned to 560 hertz. Asymmetry is explained in text.

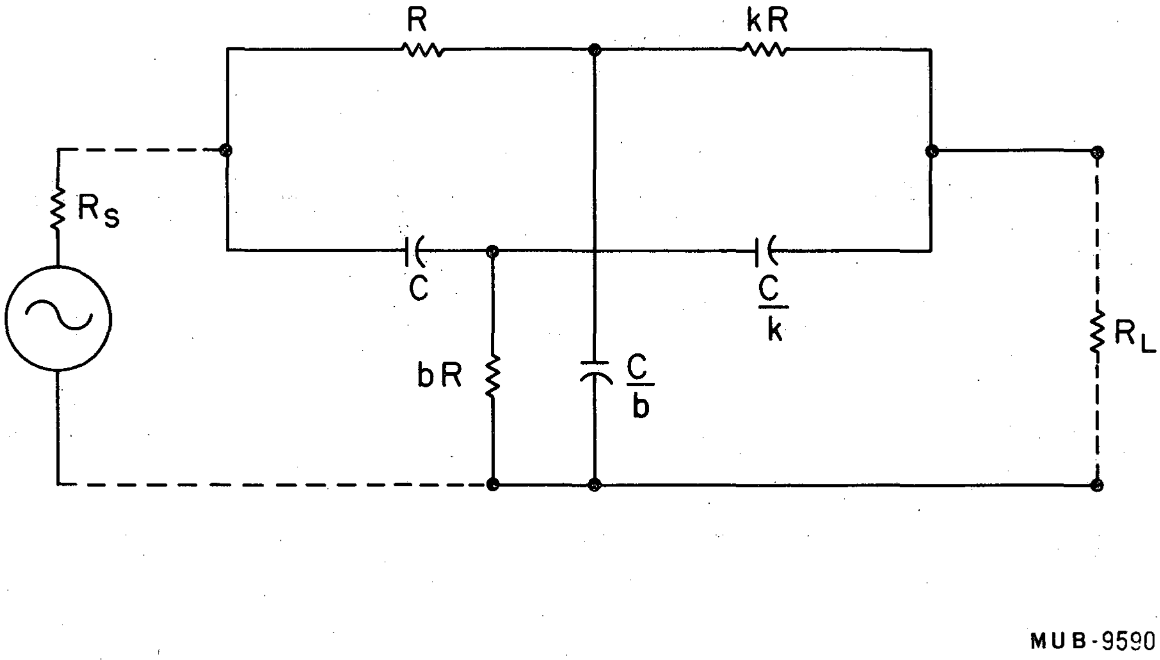

Fig. 5. Parallel-T network. The source and load are connected as shown by dashed lines.

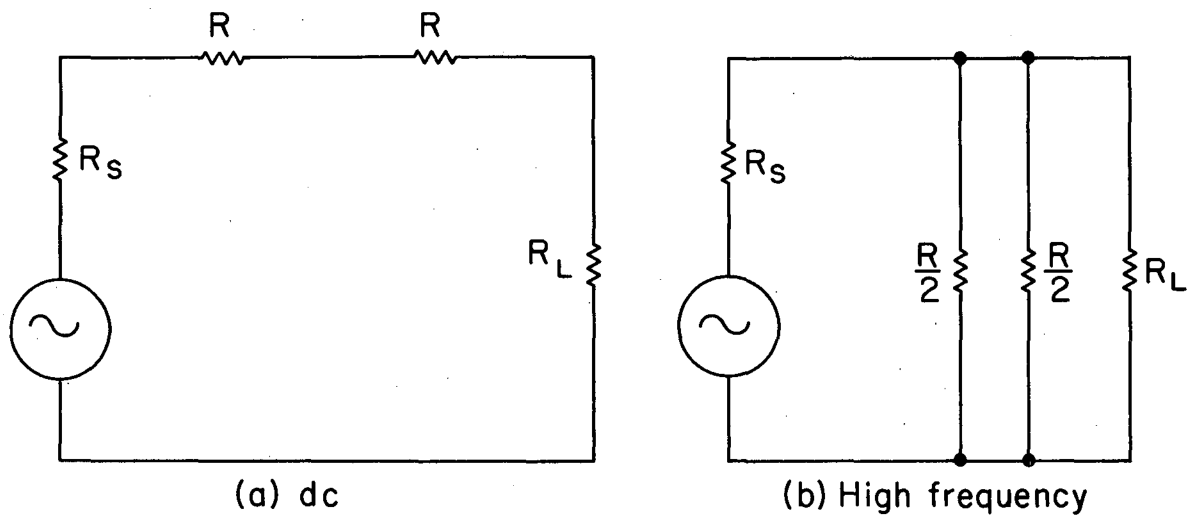

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Fig. 6. Equivalent circuits of the parallel-T at (a) dc, and (b) high frequency. R is the network resistance, as indicated in Fig. 5. R_s and R_L are the source and load resistances, respectively.

2. Synchronous Detector

The detector is a synchronous type[12] employing two diodes, both of which are gated in-phase by the reference so that they conduct during the same half cycle. The gating waveform is sinusoidal.

The detector responds to inputs at the reference frequency only. Inputs of any other frequency produce a beat frequency that is filtered out by the time constant following the detector. Since the detector is phase sensitive, inputs of improper phase produce little or no output, the amount depending upon deviation from the required phase relationship.

As indicated earlier, the signal sought is modulated at the reference frequency. Signal frequency is therefore correct, but its phase is not necessarily optimum. In order to meet this phase requirement, the af unit includes a phase shifter that permits us to change phase of the gating sinusoid.

The phase shifter is normally adjusted for maximum response to the signal so that the modulated signal appears in phase with the gating sinusoid at one diode, out of phase at the other diode. The net result is increased output from one diode, reduced output from the other. The diode load-resistors (Fig. 2) are connected in series, polarities being such that the output is zero in the absence of signal.

Detector response is very linear because the gating amplitude chosen is larger than any signal normally applied to the detector. This ensures that both diodes operate about a linear region of their characteristic curves.

Since the detector is balanced, it is relatively insensitive to changes in gating amplitude. This configuration also minimizes changes in zero resulting from changes in the phase control. This can be explained as follows: The gating sinusoid is derived from phase-shifter output. Ideally, its amplitude remains constant and the waveform is not distorted when phase is altered. This idealization has been approached, but the sinusoid distorts slightly. Distortion does not cause zero shift with this configuration because both diodes are gated to conduction by the same half cycle. Distortion, therefore, influences both diodes identically.

If the diodes were gated by alternate half cycles, asymmetric distortion would cause zero shift because the gating sinusoid is applied to the detector through a transformer. The secondary waveform therefore adjusts about a level at which areas above and below zero voltage are equal.[13] If the secondary waveform is symmetrical, the positive and negative peaks have equal amplitude.

If, however, the phase shifter introduces asymmetric distortion, the secondary waveform adjusts at a different level when the phase is changed. The new level causes the positive and negative peaks to be unequal. If the diodes were gated by alternate half cycles, one would conduct more, the other less, so that zero would change.

Some might wonder why transformers are used, especially in view of their frequency limitations. The answer lies in the application for which the af unit was originally designed--an automatic frequency control that locks a klystron's frequency to that of a resonant cavity. If the frequencies differ (owing to such changes as cavity temperature or klystron drift), the af unit develops a feedback voltage that brings them back into correspondence. This feedback voltage, obtained from the detector output, is applied in series with the voltage normally applied to the klystron's reflector (approximately 500 volts above ground). Transformers isolate the phase detector from ground.[14]

For most applications one can dispense with transformers, substituting dc coupling and the appropriate phase-splitter instead.[1] These changes would extend the frequency limits considerably.

3. Oscillator

The oscillator is a three-section, phase-lead type. Oscillation frequency, ω_{osc}, is given by[15]

$$\omega_{osc} = \frac{1}{RC\sqrt{3 + \frac{2}{a} + \frac{1}{a^2} + \frac{R_s}{R}\left(2 + \frac{2}{a}\right)}} \tag{3}$$

where the parameters represented are shown in Fig. 7.

The minimum gain necessary to sustain oscillation, A_r, is given by[15]

$$A_r = -\left[8 + \frac{12}{a} + \frac{7}{a^2} + \frac{2}{a^3} + \frac{R_s}{R}\left(9 + \frac{11}{a} + \frac{4}{a^2}\right) + \left(\frac{R_s}{R}\right)^2\left(2 + \frac{2}{a}\right)\right]. \tag{4}$$

Referring again to Fig. 7, we chose an R of 3.6 kilohms, an a of 1. Here R is considerably greater than R_s (700Ω) in order to reduce the influence of R_s upon oscillator frequency. Larger R's also reduce gain requirements of the amplifier, as indicated by Eq. (4). If R is increased, changes in the module's input impedance (30k) influence oscillator frequency to a greater extent. Small frequency changes are not harmful, but excessive drift is detrimental because the signal amplifier is tuned.

Oscillation is achieved with positive feedback around a linear amplifier. The af unit is designed for single-frequency operation, so the amplitude regulation normally associated with variable-frequency oscillators[16] has been omitted.

4. Phase Shifter

The phase shifter may seem unduly complicated, but the reasons for this complexity will soon become evident. Its basic requirement is that it shift the gating 180 deg. It is desirable that its output amplitude remain constant in order to minimize zero shift resulting from any detector unbalance. Our phase shifter satisfies both requirements.

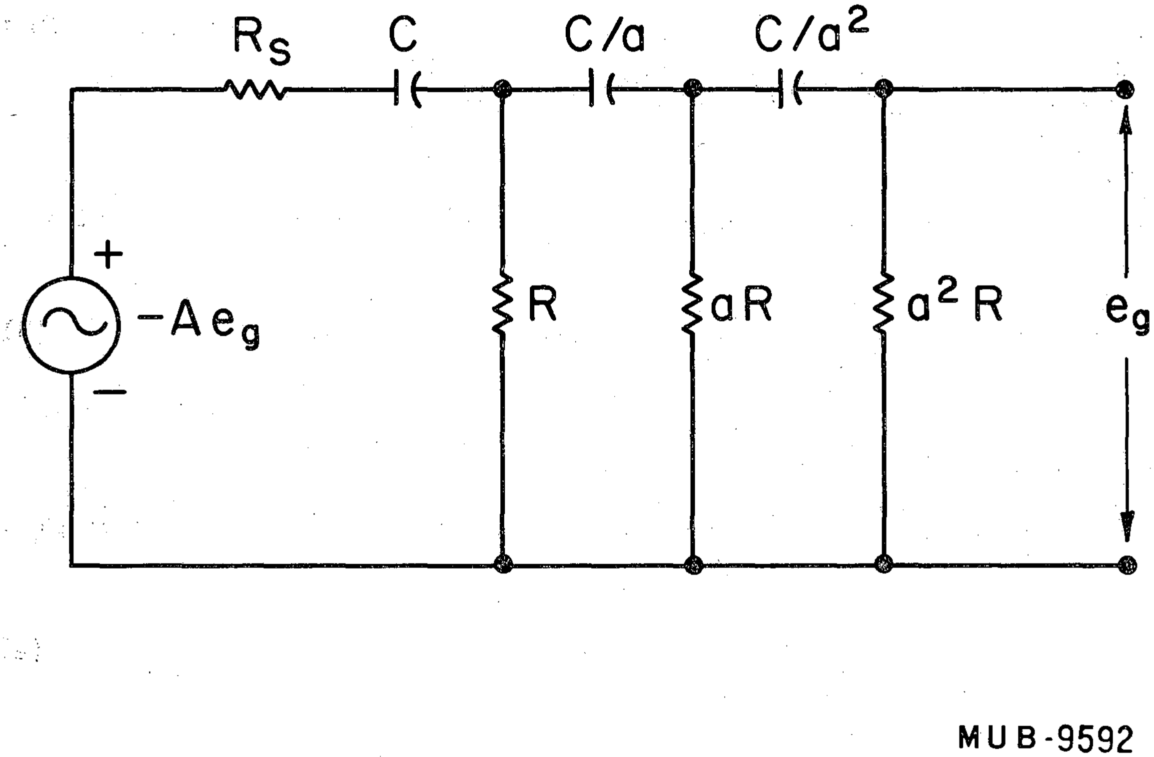

Fig. 7. Equivalent circuit of three-section phase-lead oscillator.

Output amplitude remains constant with phase because: The networks (Fig. 2) are driven by two voltages of equal amplitude but opposite phase. In the equivalent circuit of Fig. 8 and the vector diagrams of Fig. 9, note that the networks are excited by two voltages in series, E1 and E2. Part of the exciting voltage is developed across the capacitor (E_C), the rest across the resistance (E_R). Because these components are always in quadrature, they can be represented within a circle, as shown in Fig. 9. When phase is adjusted, E_C and E_R change according to the ratio of reactance to resistance. Figure 9 indicates the distribution of voltage for two different settings. Because output is taken between points P1 and P2, the output vector is drawn from the center of the circle (Fig. 9). When R changes (Fig. 8), the output vector rotates relative to the applied voltage, but it does not change in amplitude.

If the reactance of C (Fig. 8) could be reduced to zero, the output vector would rotate 180 deg when R was changed from zero to some finite resistance. But the reactance cannot be reduced to zero because (a) such a reduction would require a capacitor of infinite size, and (b) zero reactance would short-circuit the generator driving the network.

Since the source has limited current capability, the reactance of C must be large enough to ensure that the source is not overloaded when R is reduced to zero resistance. With the phase-splitter chosen (Fig. 2), we determined that distortion results unless the reactance of C is greater than 40 kilohms.

Inasmuch as distortion limits the value of C, and since the source always includes some resistance, it is not generally possible to obtain 180-deg shift with one network. Consequently we employ two networks, obtaining more than 90-deg phase shift from each. The networks are ganged, so that one control adjusts the phase of both (see C1 and C2, Fig. 2).

The foregoing discussion indicates that the reactance of C must be greater than 40 kilohms. Since reactance is a function of frequency, the value of C must be adapted to the operating frequency chosen.

In order for phase to be shifted at least 90 deg, the value of R must be approximately 50 kΩ. If R is 50 kilohms, however, phase-shifter output amplitude changes with phase when a low-impedance load is connected to the phase-shifter output. (This is more apparent from the equivalent circuit of Fig. 10.) Loading is a problem if the phase shifter is followed by conventional (bipolar) transistors. One solution to loading is bootstrapping;[17] our solution is field-effect transistors (FET's) for Q2 and Q4 (Fig. 2). FET's have a very high input impedance, literally hundreds of megohms. As used here, their input is shunted by 10 megohms, but that is relatively high when compared with the value of R used (50 kilohms). FET's are also useful because they permit coupling with smaller capacitors.

5. Source Follower

The source follower, not required for some applications, is necessary when the phase detector must drive a low-impedance load. The phase detector is not suitable for driving low-impedance loads because (a) detector

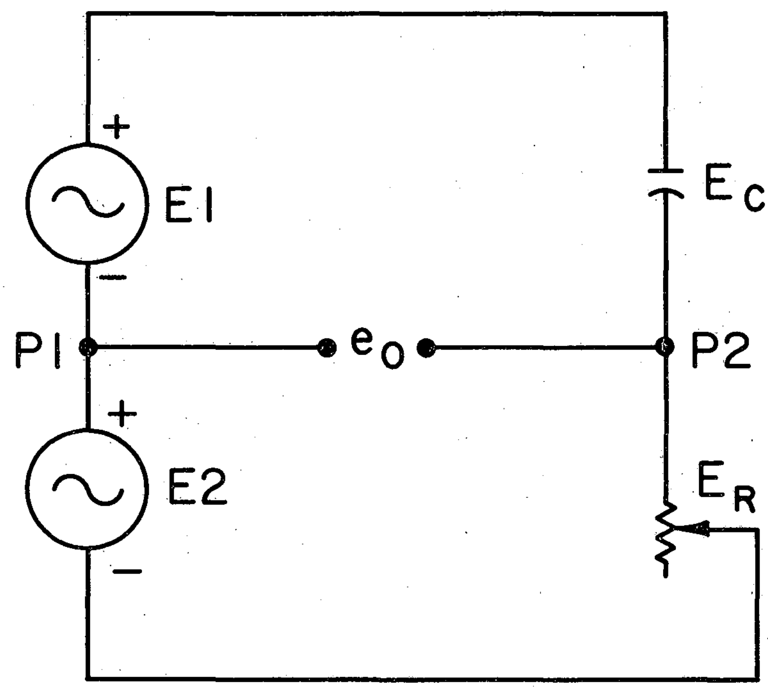

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Fig. 8. Equivalent circuit of each phase-splitter (Fig. 2). E1 and E2 are equal in amplitude, but of opposite phase. Output is taken between points P1 and P2.

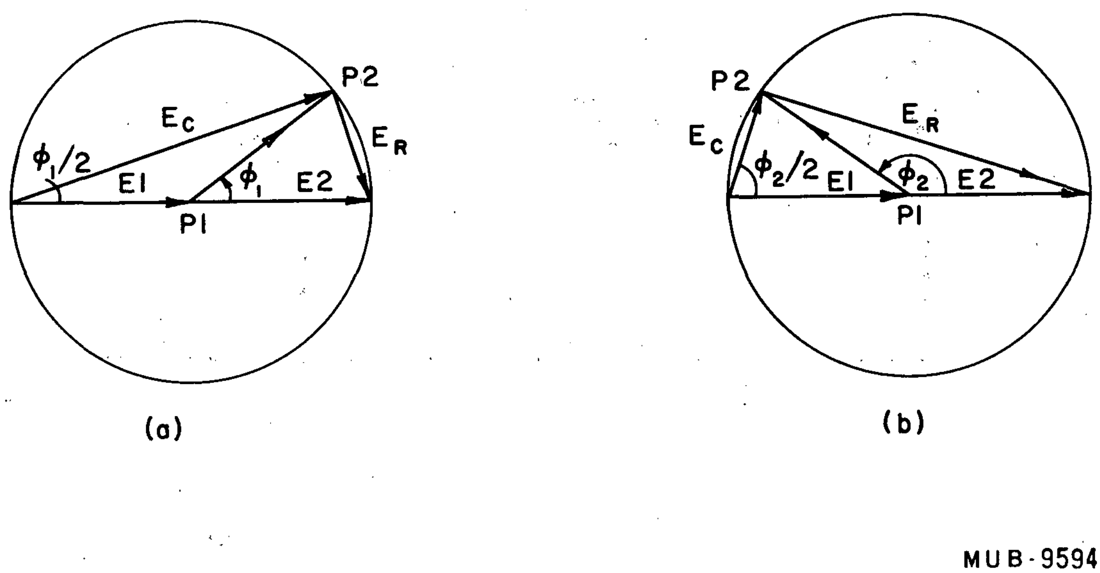

Fig. 9. Vector diagrams illustrating why the phase-splitters yield an output of constant amplitude but variable phase: (a) Phase control at an arbitrary setting of R. (b) R increased to a higher value.

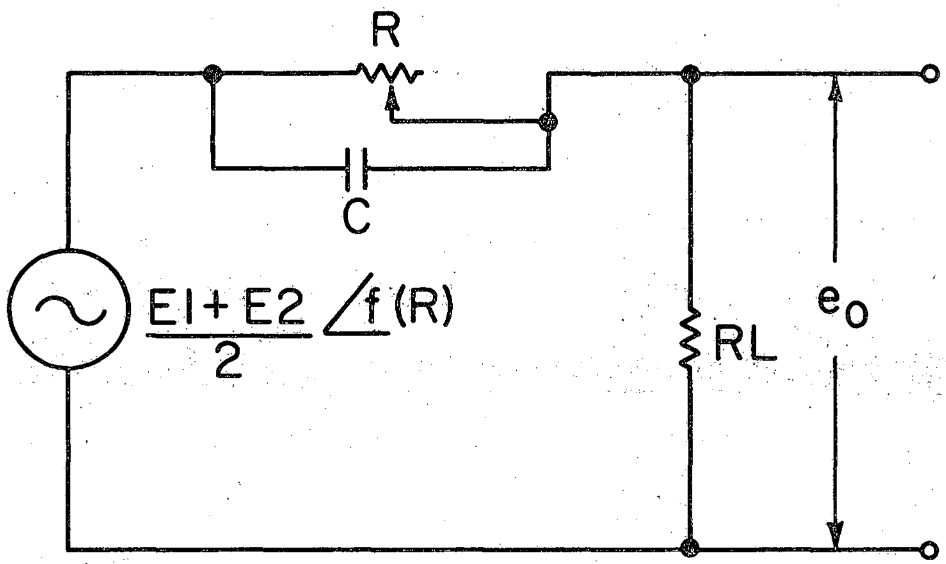

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Fig. 10. Equivalent circuit of phase-splitter illustrating why output amplitude changes with phase unless R_L is much greater than R.

linearity suffers, and (b) output voltage is reduced too much for many applications. Also important, the time constant responsible for system bandwidths (Fig. 1) is connected between phase-detector output and source-follower input. The source follower permits us to reduce the capacitor size associated with a given time constant. Because the source-follower input impedance is very high (uses FET's), we can use large resistors for R2 and R3 of Fig. 2, thereby reducing the capacitance required for a given time constant. We had difficulty balancing this source follower (Q5 and Q6 of Fig. 2) with one make of transistor, but not with the one made by Siliconix.

6. Preamplifier

A small preamplifier has been designed for those applications requiring microvolt sensitivity. This preamplifier, which uses two linear-amplifier modules, was constructed separately from the audio unit; its schematic is given in Fig. 3. Power was obtained from (a) internal batteries, or (b) a separate power supply, also shown in Fig. 3. Maximum gain from the preamplifier is about 8000, with the actual value depending upon factors such as load impedance.

C. Construction

As indicated earlier, we have attempted to simplify construction by including photographs (Figs. 11 through 15), templates (Figs. 31 and 32, Appendix I. E), and other useful data.

The entire af unit is built on a commercial chassis box, LMB type 17, which is 13 by 4-1/8 by 2-5/8 in. Layout is simplified if the master template (Fig. 31) is folded around the chassis box. This template indicates the hole centers and should be pierced at the appropriate locations. Another template, Fig. 32, facilitates phase-shifter layout.

Some will find it easier to wire the unit in phases, making the appropriate tests as each phase is completed. This approach is more practical if the power supply is wired first. After the power supply is operative, a practical wiring sequence is (a) signal amplifier, (b) oscillator plus buffers, (c) phase shifter, (d) phase detector, (e) source follower. The appropriate tests can be deduced from Sec. III. D.

The preamplifier and its power supply are so simple to construct that photographs and templates are not included.

D. Testing

Appendix I includes detailed descriptions of measurements relative to gain, noise, drift, and linearity. The more obvious tests are not discussed here, but a few key points should prove helpful:

1. Signal Amplifier

   a. Removal of the parallel-T will facilitate amplifier testing and will not interfere with amplifier operation.

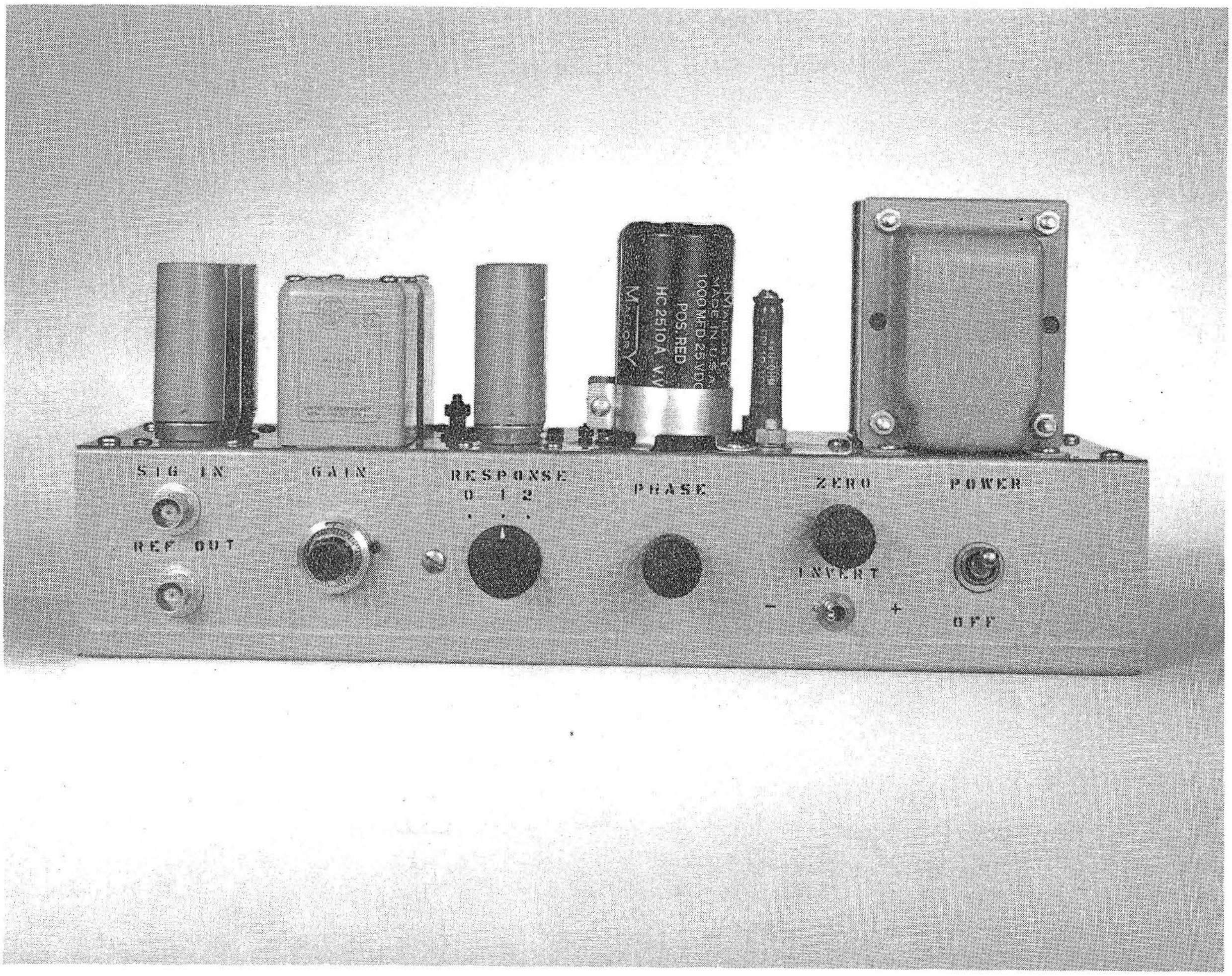

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Fig. 11. Front view of af unit. The circuit modules (both sides of transformer) are plug-in. Power supply (filter, resistors, and transformer) is to the right. BNC connectors are insulated from chassis. The audio transformers are enclosed by magnetic shields.

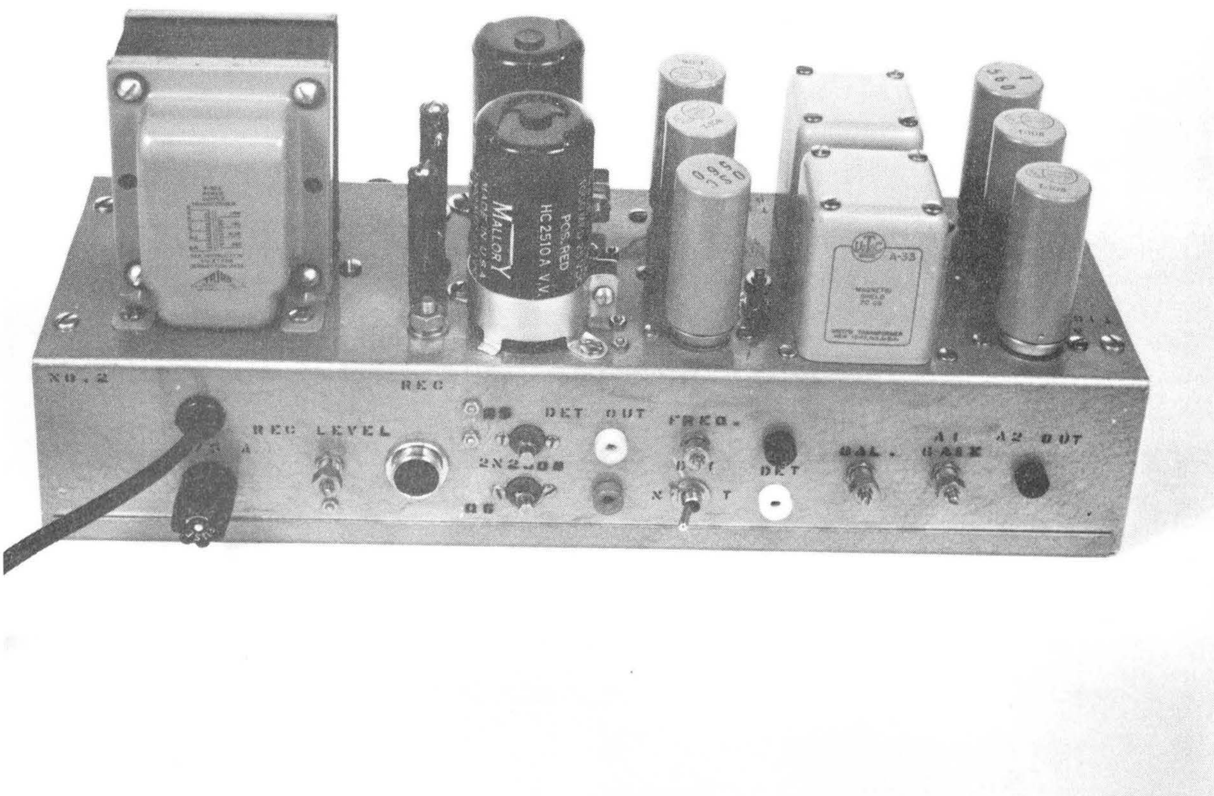

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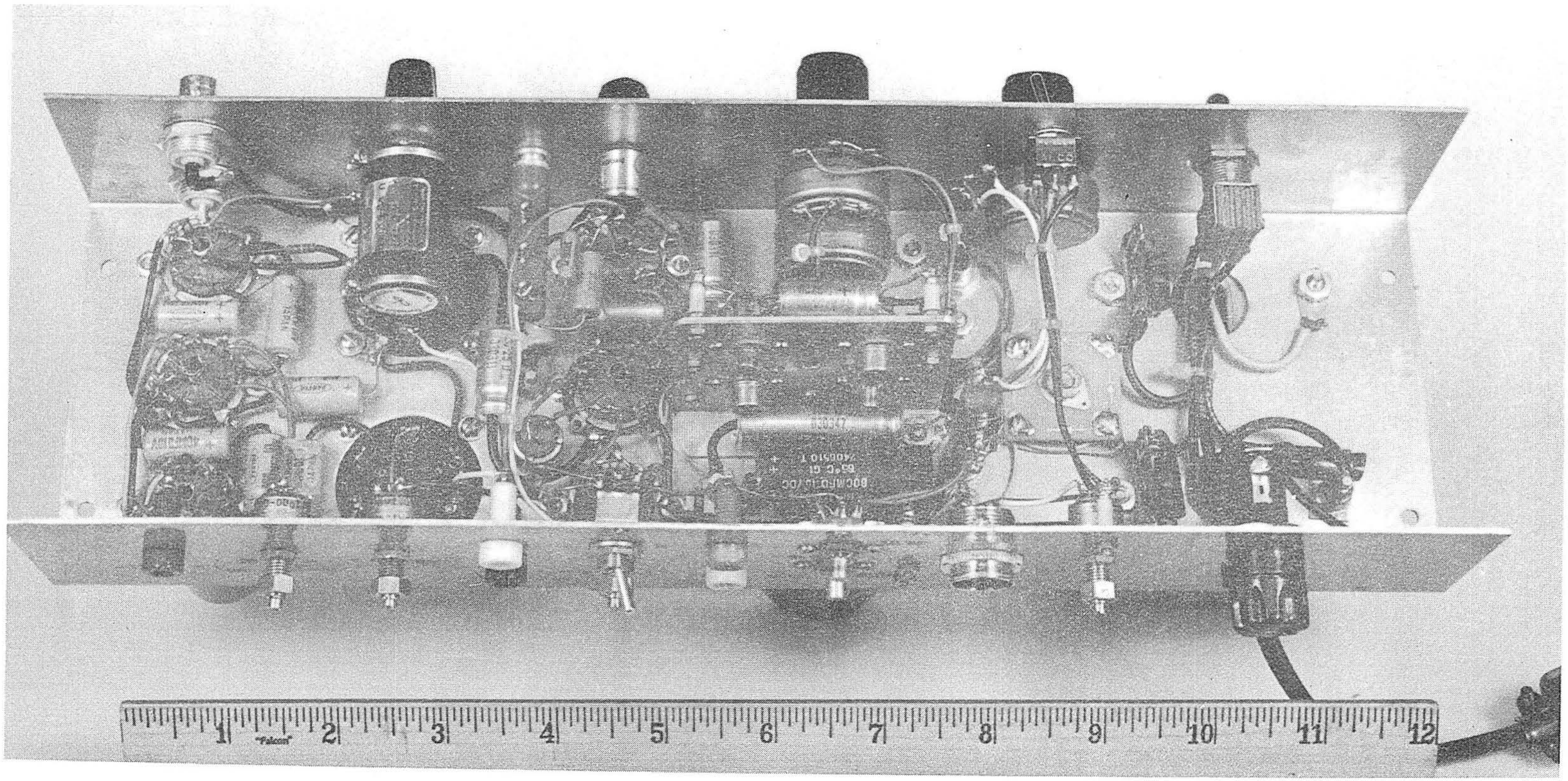
Fig. 12. Top rear view of af unit. Modules housing the parallel-T and oscillator networks have their frequency stencilled on top. OSC control is the screw-driver adjustment next to the oscillator network. All adjustments and test points are accessible.



ZN-5381

Fig. 13. Bottom rear view of af unit showing components and wiring in detail (bottom cover plate removed). Most phase-shifting circuitry is mounted on the small plate in the center (template in Appendix I. E); transistors Q1 through Q4 are clearly visible.

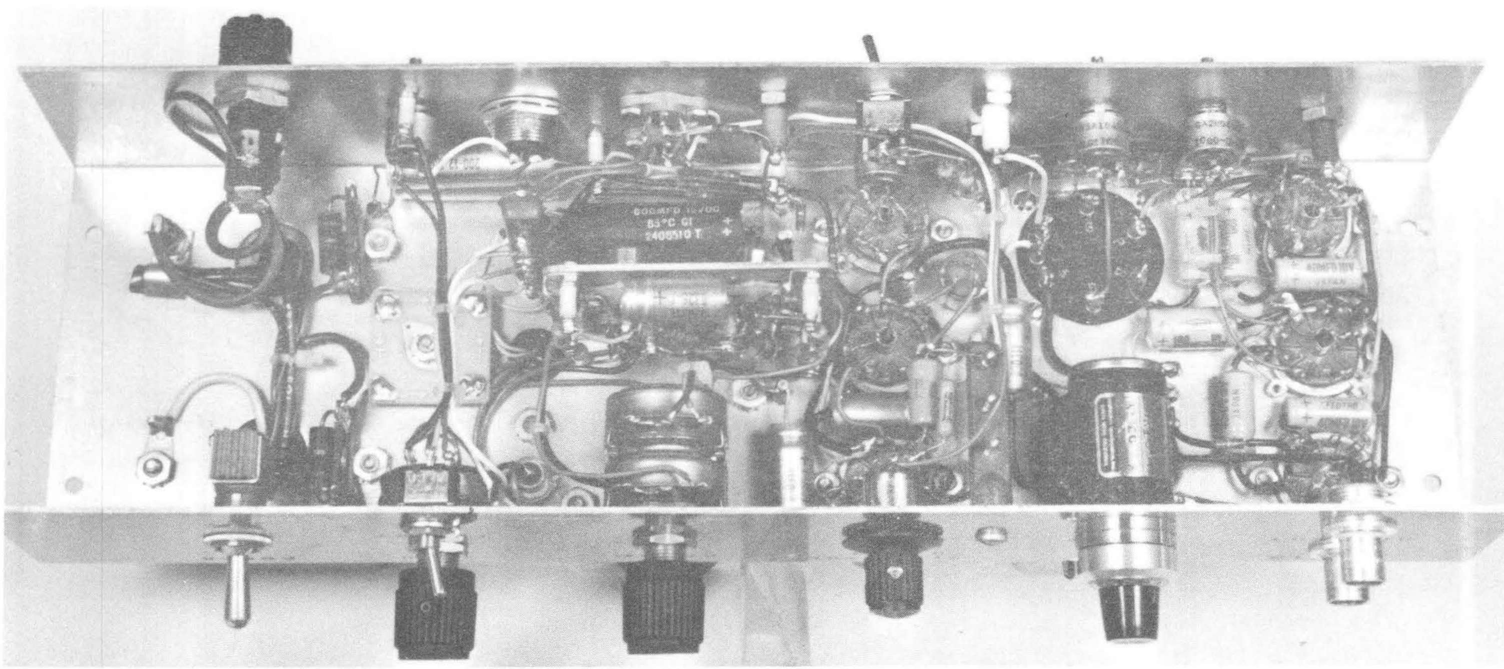


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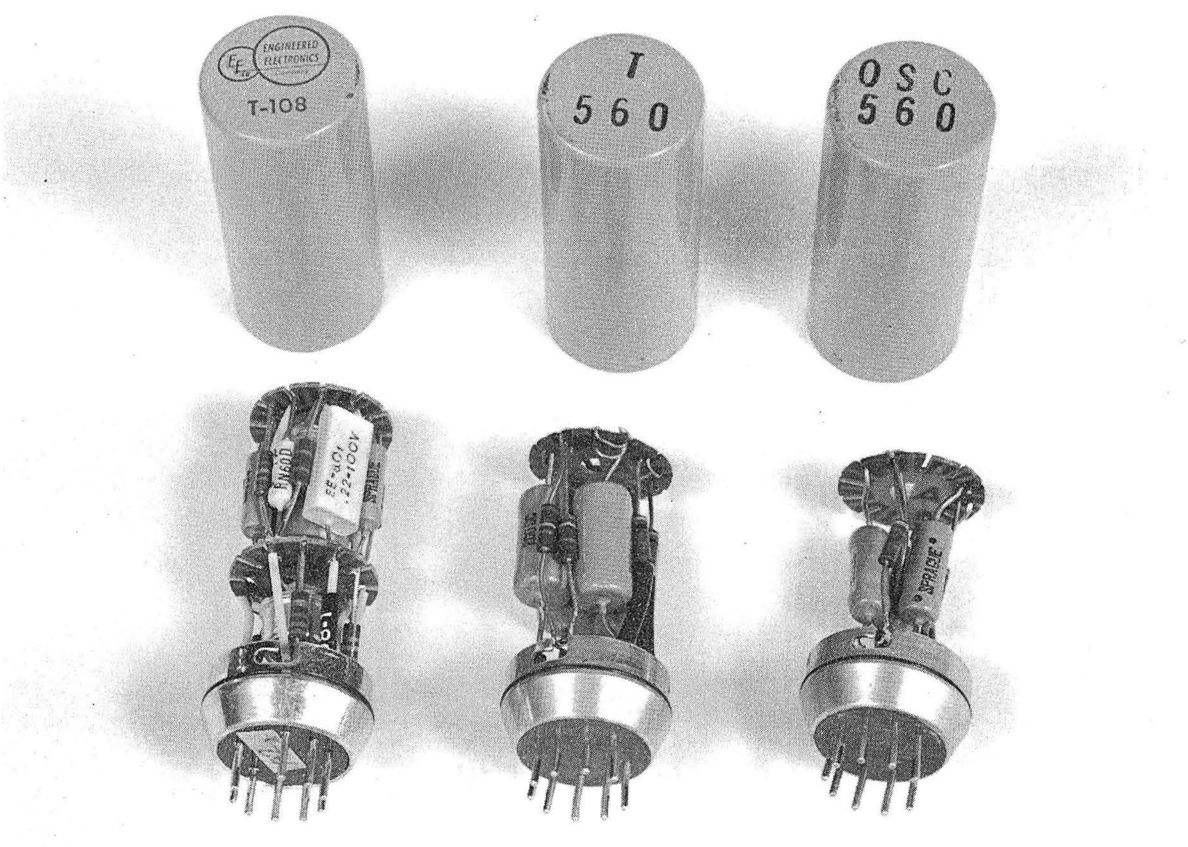
Fig. 14. Bottom front view of af unit showing components and wiring not visible in Fig. 13.



ZN-5382

Fig. 15. Plug-in modules. A linear amplifier, T-108, is shown at the left. At center, T560 is a parallel-T fabricated into a blank module. At right, OSC560 is an oscillator network, also fabricated.

b. Amplifier gain with the parallel-T plugged in should be slightly less than with it removed. If the parallel-T increases gain, the first stage becomes regenerative, and therefore susceptible to oscillation.

c. The value of R for the parallel-T (Fig. 5) should be approximately 3k. The capacitor values required can be computed from Eq. (1); if these values differ only slightly from standard ones, it is easier to select the value of R accordingly. If the frequency characteristics of A1 are not sharp enough ($Q \approx 25$), or if the gain of A1 is reduced too much when the parallel-T is inserted, change the value of bR (Fig. 5) to one that yields the desired characteristics.

d. It is generally necessary to reduce A1 GAIN slightly below its maximum value when the parallel-T is plugged in. If GAIN is not reduced, the first stage oscillates. Adjust A1 GAIN to the highest stable value in order to increase selectivity.

e. The signal at the blue test point (Fig. 2) should not exceed 2 V peak-to-peak. Above this value, linearity falls off.

2. Oscillator

a. The value of R (Fig. 7) should be approximately 3.6k. Capacitor values can be computed from Eq. (3). If these values differ only slightly from standard resistances, it is easier to select the value of R accordingly.

b. The oscillator frequency must match that of the parallel-T. In order to achieve this, apply an attenuated output (see Appendix I. B for attenuator) from REF OUT to SIG IN, then adjust the FREQ trimpot for maximum output at the blue test point. If the FREQ trimpot has insufficient control, modify the oscillator network accordingly.

c. Adjust the OSC trimpot for approximately 6 volts, peak-to-peak, at REF OUT. The waveform should be sinusoidal and free of distortion.

3. Phase Shifter

As indicated in Sec. III. B. 4, the phase-shifting capacitors (C1 and C2 of Fig. 2) should have a minimum reactance of 40k. Below this figure, phase-shifter output becomes distorted for some settings of the PHASE control. If the reactance is considerably greater than 40k, it is not possible to shift phase 180 deg if R (PHASE CONTROL, Fig. 2) is only 50k.

To check phase-shifter performance, monitor Q4's emitter (Fig. 2) with an oscilloscope (trigger scope externally from REF OUT). If network parameters are correct, the phase control permits 180-deg shift. The output waveshape should be sinusoidal and its amplitude should be insensitive to phase. If the waveform is distorted, alter bias resistors indicated by a double asterisk in Fig. 2.

To balance the first phase-splitter (Q1 of Fig. 2), select a collector load resistance that results in equal voltages at emitter and collector. Balance Q3 in the same manner.

4. Synchronous Detector

The synchronous detector, like many of the other components, is subject to frequency considerations. For example, dc output level is related to the size of capacitors C3 and C4 (Fig. 2). For each frequency there is an optimum value dependent upon ripple tolerable and output level required. These requirements are not severe, however, and one is permitted considerable latitude in choosing C3 (and C4).

The synchronous detector is a form of rectifier. Its performance depends upon two different time constants, charge and discharge. Both involve the same capacitor (C3 for example), but a different resistance.

The charging resistance is determined by the output impedance of A2 (Fig. 2) and the properties of transformer T1; most of it is due to the winding resistances of T1. Resistance components contributed by the output impedance of A2 and each diode are negligible by comparison. In practice, the charging resistance is approximately 10k.

The charging time constant with C3 (and C4) selected for 560 hertz is approximately 100 μsec. Compare this with 450 μsec, the time required for a sinusoid to reach peak value at the frequency indicated. If the capacitance of C3 is reduced, it charges to a higher amplitude.

The discharge time constant, determined by C3 and the resistance in shunt with it, is approximately 1700 μsec. This constant determines the level to which C3 discharges in the time interval between charging cycles. The discharge time constant determines ripple at the detector output.

Some may choose to find the value of C3 empirically, selecting a value that gives the greatest output (meter current) for a given signal input. Others may be more concerned with ripple. To obtain comparable performance at another frequency, scale C3 (and C4) inversely with frequency.

The detector (Fig. 2) includes a switch labelled DET, which has two positions, T and N (Test and Normal). The Test position is convenient for monitoring detector performance with C3 and C4 disconnected. If, for example, the black test point is grounded and one monitors the yellow (or red) test point with an oscilloscope probe, half-wave rectification can be observed. If a signal is applied, the amplitude of the rectified waveform increases or decreases, depending upon signal phase relative to the reference.

E. Specifications

Performance is best described in terms of specifications, which are listed below:

FREQUENCY RANGE: 40 to 20,000 hertz. The frequency range can be extended if the transformers are omitted.

SIGNAL CHANNEL Q: ≈ 28 at 560 hertz.

GAIN: (dc out/rms in) approximately 3000. Additional gain (more than 70 db) can be obtained with the preamplifier described.

EQUIVALENT INPUT NOISE: Approximately 1 microvolt rms with the input terminals shorted and a 2-second time constant. Measured at 560 hertz.

LINEARITY: Better than $\pm 1\%$ of full scale.

ZERO DRIFT: Less than $\pm 1\%$ of full scale per hour, maximum.

OUTPUT:

LEVEL: Maximum dc output is ± 2 V (linear range).

IMPEDANCE: 1 k when taken from RECORDER, 300 k at PHASE DET OUT.

INPUT IMPEDANCE: 1 k (approximately 30 k with preamplifier).

Appendix I(C and D) includes detailed descriptions of gain, bandwidth, noise, drift, and linearity measurements.

F. Applications

The af unit was designed for an ESR spectrometer now under development. Two lock-in amplifiers are used, one for automatic frequency control, another for the signal channel.

We are now adapting the af unit to NMR work. Specifically, we are designing an NMR fluxmeter and will use the af unit for locking the frequency of an rf oscillator to the resonant frequency of a sample. A future application involves automatic frequency tracking when the magnetic field is varied.

IV. RADIO-FREQUENCY LOCK-IN AMPLIFIER

A. General Considerations

Commercial rf lock-in amplifiers do not yet encompass the entire radio-frequency range. This is probably because (a) there is little demand, and (b) frequency flexibility is hard to come by. As requirements become better defined, commercial units can be expected to follow suit.

As stated earlier, the rf unit was designed with a specific objective in mind: to measure pressure differentials on the order of 0.1 μ accurate to several percent. At that time, commercial micromanometers of sufficient sensitivity were not available, so we designed our own. For reasons described elsewhere,[18, 19] we elected to sense pressure difference with a membrane manometer, constructed like a differential capacitor. The capacitor formed two legs of a resonant-bridge network excited by a 2.7 MHz source. Bridge output was amplified and detected with the rf unit described.

The frequency of 2.7 MHz was chosen for several reasons, some theoretical and some practical. Those reasons, not discussed here, are explained in Ref. 18. At any rate, we ended up with an rf unit that could offer promise to others. If adapted to other rf frequencies, the LC ratio used must be compatible with Nuvistor characteristics.

Some may wonder why the rf unit also was not designed around circuit-modules. The main reason was that the rf unit was developed before the af unit, so we had not yet thought of using circuit modules. Even so, we might have rejected the modular concept for the rf unit because: (a) Most modules with adequate frequency response are subject to oscillation so we might not have been able to realize as much gain. (b) The signal amplifier should be tuned, so there is no particular advantage to wide-band devices. The gain-bandwidth product is wasted. (c) High-frequency modules are more expensive. If modules were used, the rf unit would have cost about 3 times as much as it did with Nuvistors. (d) Vacuum tubes operate at higher voltages, so the output levels are higher and the need for a dc amplifier is eliminated in some cases. In spite of the reasons listed, some might find the modular concept worthy of investigation. At any rate, our rf unit performs well and is easy to duplicate.

The rf unit uses Nuvistor tubes because: (a) It was to be duplicated and operated by those with little or no electronic background. Transistors were not used because their "loose" tolerances pose duplication problems. Good design can compensate for this, but engineering funds were limited. (b) Nuvistors are very small and generate little heat; these factors permit us to confine them to well-shielded compartments, thus minimizing instability problems. (c) Their combination of high transconductance with low interelectrode capacitance permits considerable gain without neutralization of the amplifier stages.

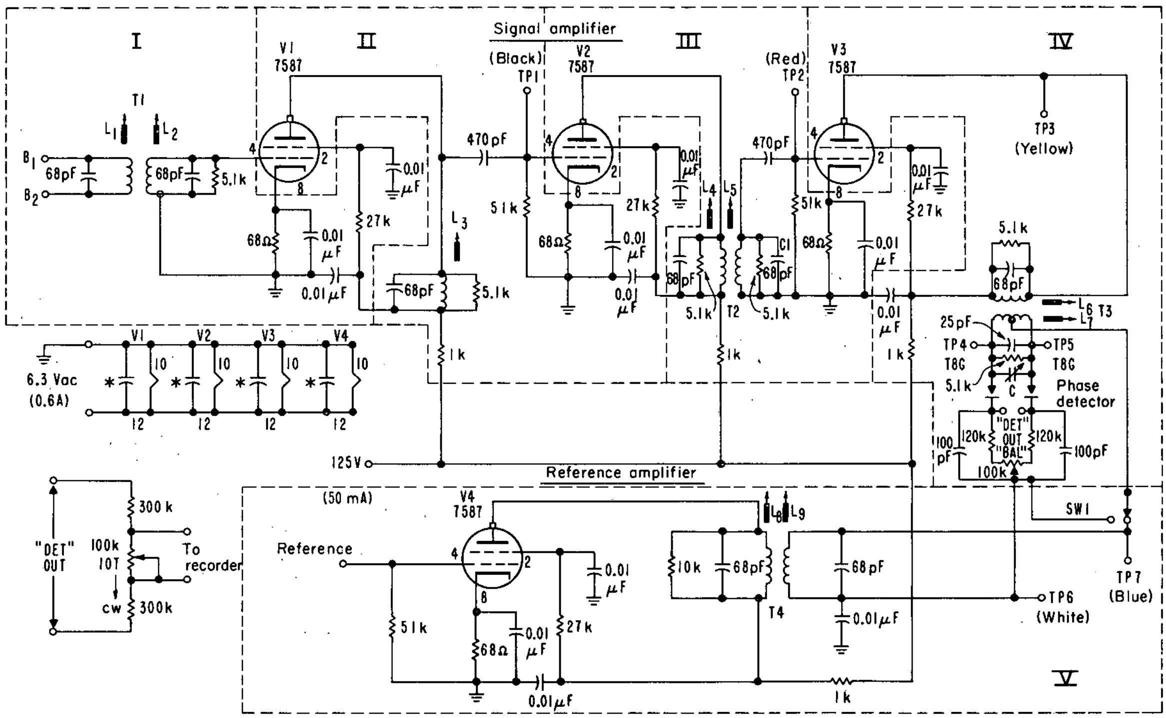
B. Circuit Description

NOTE: Refer to the schematic diagrams, Figs. 16, 17, and 18.

1. Signal Amplifier

The signal amplifier comprises three stages, each tuned to 2.7 MHz. Because all three stages employ Nuvistors, more gain can be realized than is possible with most conventional tubes.[20] The tuned circuits are shunted with 5.1k resistors that reduce regeneration and increase bandwidth, thus minimizing the effects of frequency drift.

Some tuned circuits are inductively coupled to form transformers. In Fig. 16, T1 (for example) is used to match the input impedance of V1 to that of the source. We now show that its primary should be series resonant in some cases, parallel resonant in others, with the choice depending upon source resistance.



MUB-9586

Fig. 16. Schematic diagram of the rf lock-in detector unit. The 27 k resistors are $\frac{1}{2}$ W; all others are $\frac{1}{4}$ W. Asterisks denote 1000-pF ceramic feed-through capacitors, C is a 7 to 45 pF ceramic trimmer, all other pF capacitors are silver-mica, and all μF capacitors are Mylar. All inductors comprise 64 turns of No. 32 Formvar wound on National Radio Corp. form XR50; L7 is center-tapped. The roman numerals indicate the compartments (Fig. 21) in which each component is mounted. A low-impedance potentiometric recorder can be driven by a divider across the DET OUT, as shown in insert.

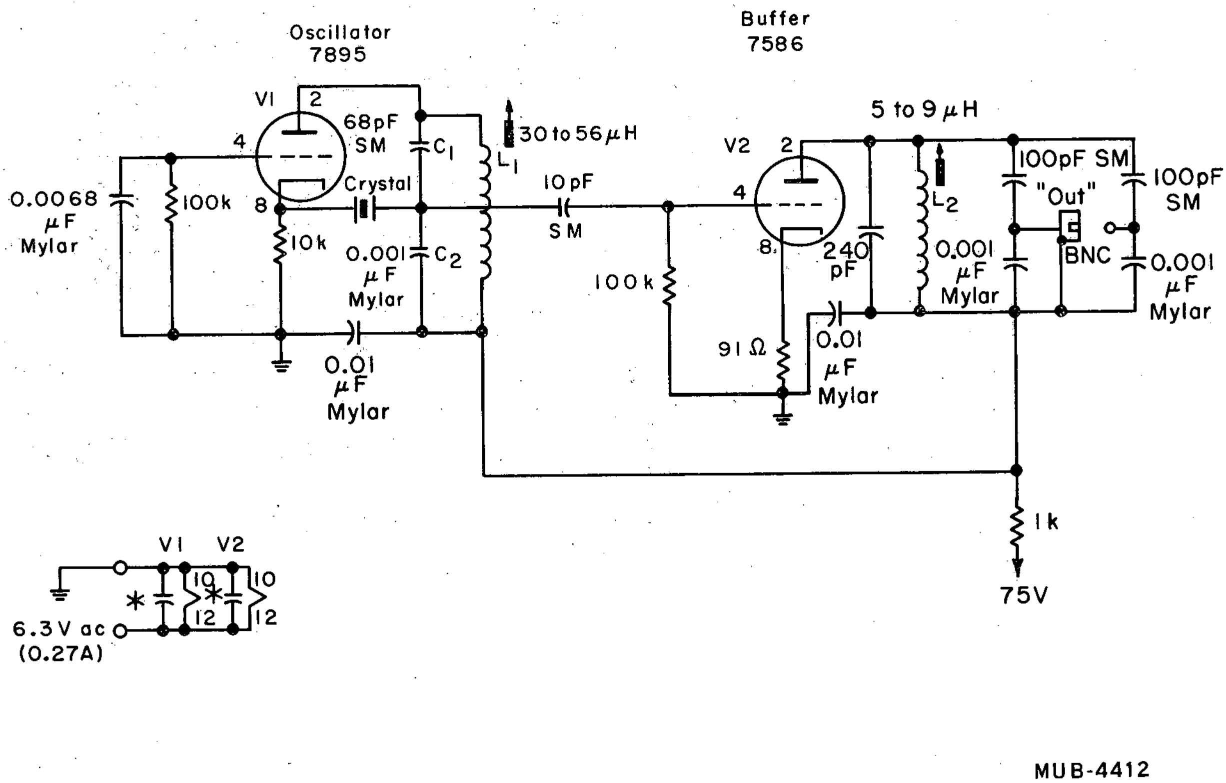
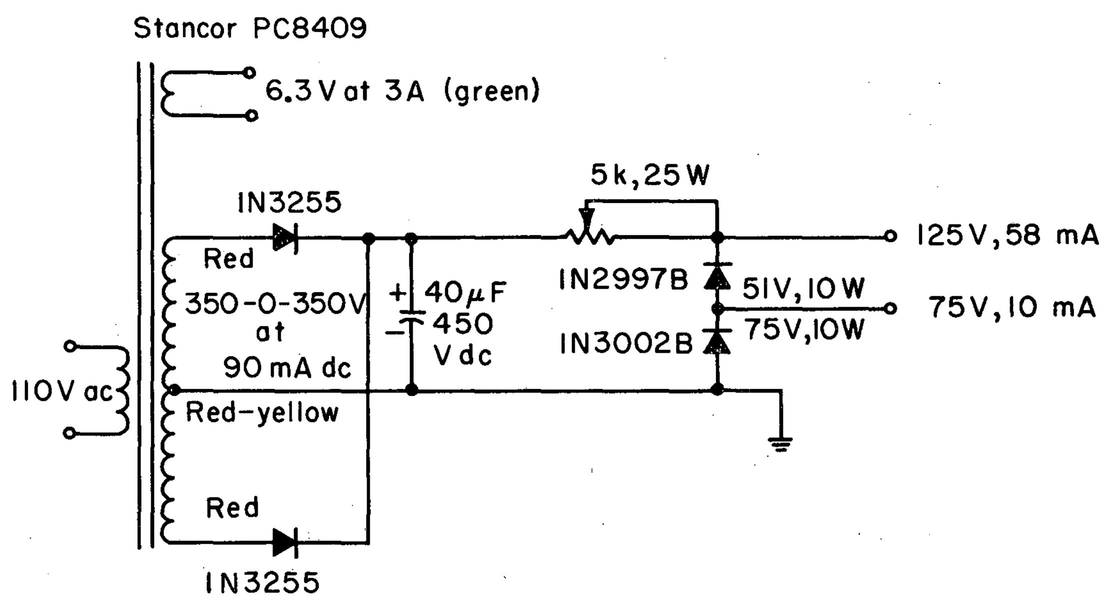


Fig. 17. Schematic diagram of the rf lock-in generator unit. All resistors are $\frac{1}{4}$ W. Inductance L1 comprises 64 turns of No. 32 Formvar wound on National Radio Corp. form XR50; L2 is 25 turns of No. 22 Formvar wound on form XR50. Coil measurements are included in Appendix II. The crystal (type CR-18/U) operates at 2762.500 kHz. Capacitors marked with an asterisk are 1000-pF ceramic feed-through type; those indicated by SM are silver-mica.



MUB-9596

Fig. 18. Schematic diagram of the rf unit power supply.

Assuming that we wish to obtain maximum gain from T1, we write the equation as[21]

$$G_{max} = E_c/E = \tfrac{1}{2}(Q_p\, Q_s)^{\frac{1}{2}}\,(L_s/L_p)^{\frac{1}{2}}\,, \qquad (5)$$

(with critical coupling assumed) where the parameters represented are indicated by Fig. 19; here Q_p and Q_s represent the primary and secondary Q's, respectively.

Assuming further that L_s and L_p are equal (which they are), and that Q_s is fixed, we reduce Eq. (5) to

$$G_{max} = K(Q_p)^{\frac{1}{2}} \qquad (6)$$

where K is some proportionality constant. Therefore, G_{max} is proportional to the square root of Q_p.

The value of Q_p can be expressed by two equations, one for series resonance, the other for parallel resonance:[22]

$$Q_p = \omega L1/Req \qquad \text{(series resonance)} \qquad (7)$$

$$Q_p = Req/\omega L1, \qquad \text{(parallel resonance)} \qquad (8)$$

where $\omega L1$ is the inductive reactance of L1, and Req is the equivalent resistance lowering primary Q.

If we assume that losses due to source resistance are much greater than the equivalent coil losses (6 Ω), we can substitute the source resistance for Req into Eqs. (7) and (8). Each equation shields a different Q_p;[23] we select the resonance type giving the highest value.

An example should prove helpful. With 2.7-MHz operation and 40 μH assumed for L1, its inductive reactance is approximately 700 Ω. We would use series resonance if the source resistance were less than 700 Ω, parallel resonance if it were more.

Physically, this can be explained as follows: Secondary voltage is directly proportional to primary current, I_p.[24] If the primary is series resonant, I_p and line current (I_L) are the same. If the primary is parallel resonant, I_p is Q_p times I_L. The important quantity in either case is the value of I_p.

At first, it might appear that parallel resonance is always best because I_L is multiplied by Q_p. Remember, however, that parallel resonance increases the input impedance of the primary, reducing I_L accordingly. If I_p is to be increased by parallel resonance, the value of Q_p must be greater than the factor by which line currents are reduced; this factor is related to the impedance of the primary relative to the resistance of the source. The preceding derivation indicates that the transition point occurs when the source resistance is equal to the inductive reactance of L_p.

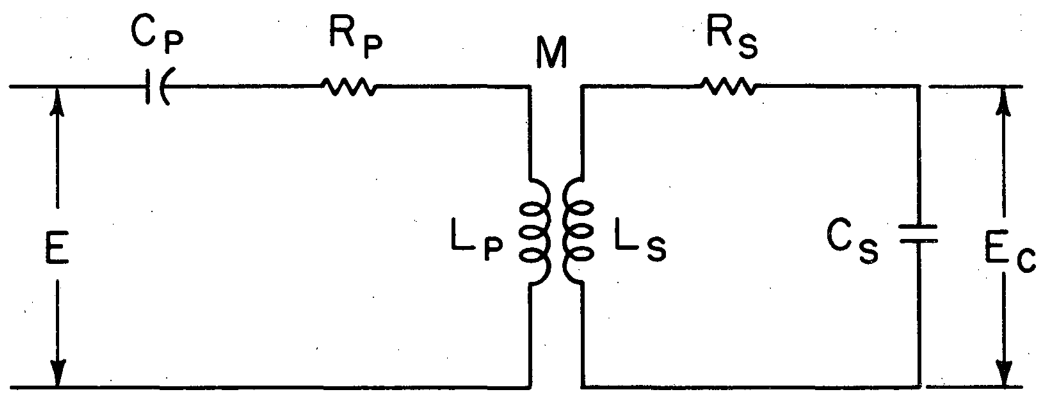


Fig. 19. Equivalent circuit useful in proving that transformer T1 (Fig. 16) should be operated series resonant in some cases, parallel resonant in others, with the choice depending on source resistance.

As to the maximum gain obtainable from T1, substitution of the design parameters into Eq. (5) reveals that the maximum value is approximately 16. In most cases, the value realized will be considerably less, the amount depending upon source resistance.

In Fig. 16, note that V2 and V3 are inductively coupled also. This represents a change (from the original version) that improves stability. Inductive coupling permits a common ground return for C1 and the input capacity of V3. (The circuits of Figs. 16, 17, and 18 include other changes, but these are not discussed.)

2. Synchronous Detector

With the exception of tuned circuits, the detector configuration is identical to that used for the af unit. For a discussion, refer to Sec. III. B. 2. The rf unit does not include a source follower and special time constants, such as those used in the af unit. If necessary, these can be added as shown in Fig. 2. Common-mode voltage at the detector output is greater for the rf unit, however, and bias to the source follower must be adjusted to compensate for their differences. If large time constants are not required, a low-impedance potentiometer recorder can be driven by a divider connected across the DET OUT terminals, as shown in Fig. 16.

3. Reference Amplifier

The reference amplifier involves only one stage. Amplifier output is coupled to the phase detector with transformer T4. The rf unit does not include a phase control per se. Phase is adjusted with L9. Some may prefer a more elaborate phase shifter, such as discussed in Sec. III. B. 4.

4. Generator

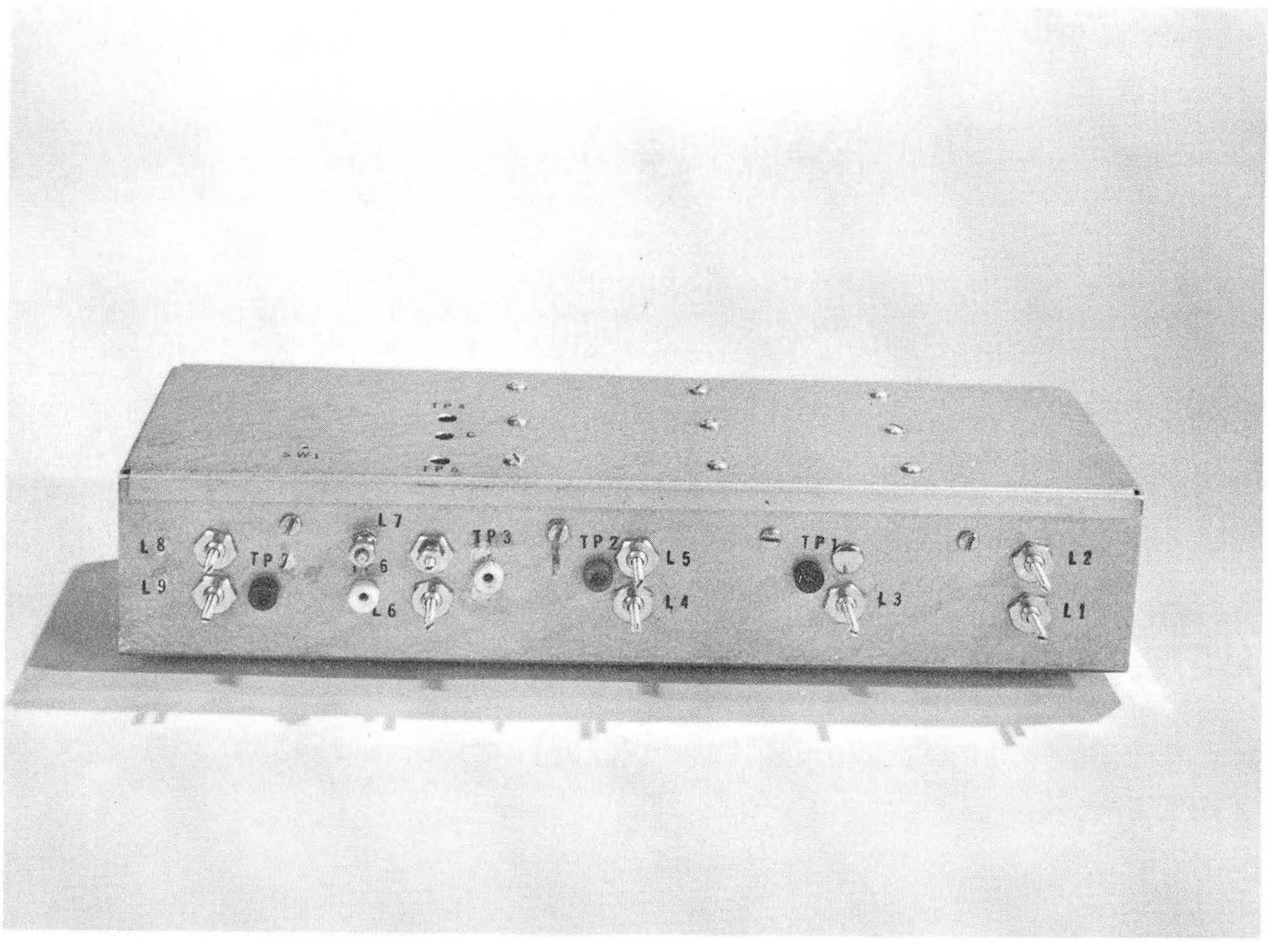
The generator (Fig. 17) comprises a crystal-controlled oscillator and a buffer. (A self-excited oscillator might have proved adequate, but we preferred the stability associated with crystal control.) The crystal is operated series resonant with feedback determined by the capacitance ratio of C1 to C2. The divider ratio and bias are adjusted to produce a minimum of distortion. The divider also reduces oscillator loading, thus improving frequency stability.

Buffer excitation is low to prevent V2 from drawing grid current. The coupling capacitor and bias are adjusted for purity of waveform.

C. Construction

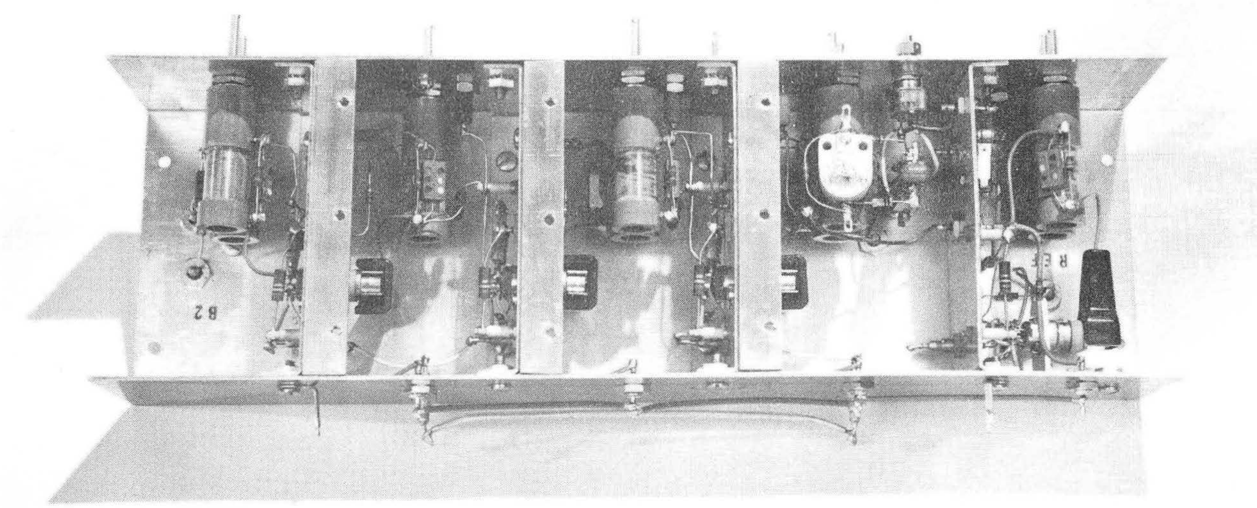
We have attempted to simplify construction by including photographs, templates, and other useful data. Photographs are presented as Figs. 20 through 25, but do not show the power supply.

The layout templates, Figs. 34 through 36, are presented in Appendix II. G.



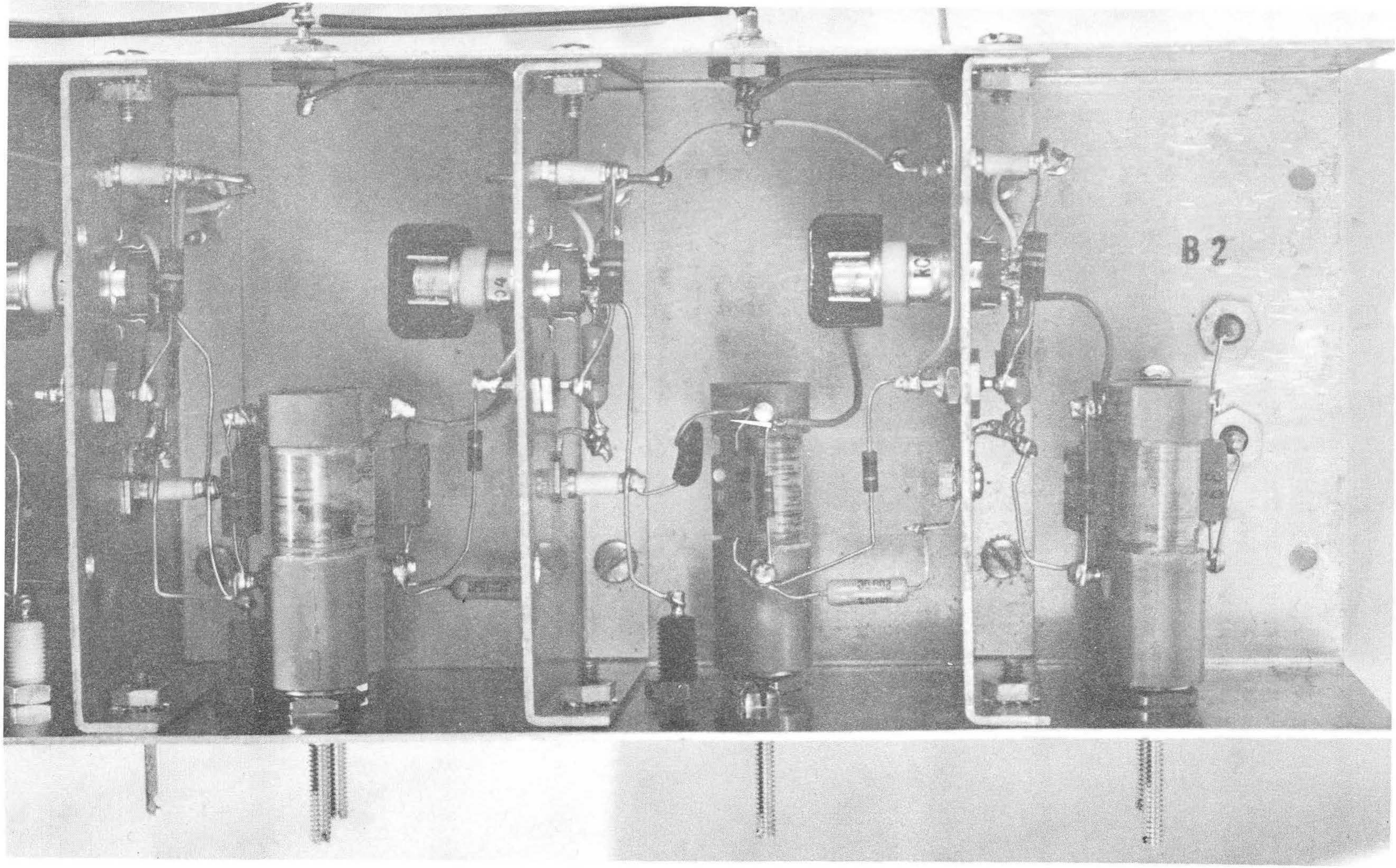
ZN-5384

Fig. 20. Top front view of rf lock-in detector unit. Note the proximity of L1 and L2. These coils and others are mounted to form transformers. Test points allow monitoring with the cover plate on. Holes in the cover plate provide access to C and SW1.



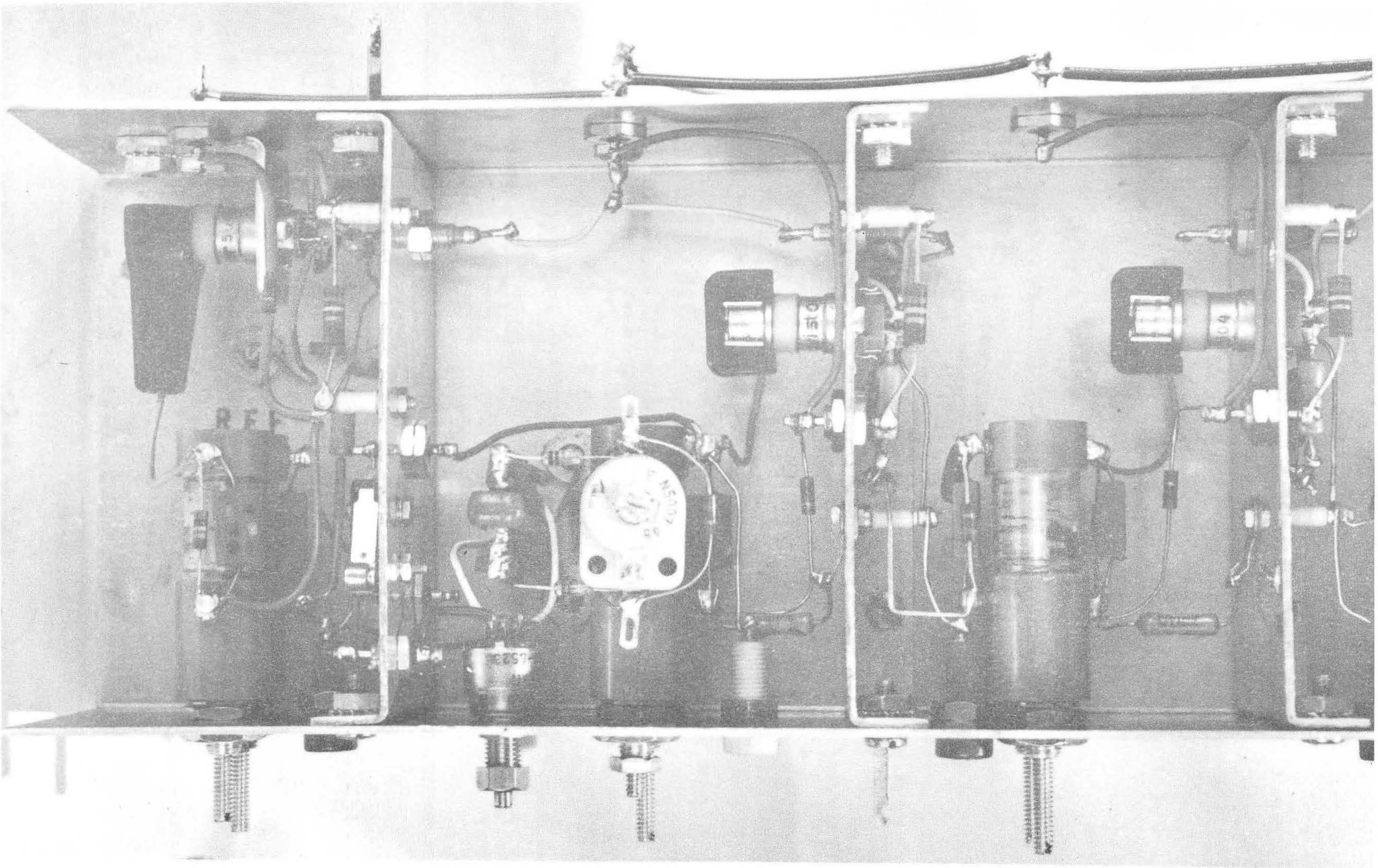
ZN-5385

Fig. 21. Top rear view of rf detector unit with the top cover plate removed. The three compartments to the left house the signal amplifier; the right-hand compartment houses the reference amplifier. The phase-sensitive detector occupies the compartment between amplifiers. Note the three angle brackets, tapped to receive screws holding the top cover-plate down.



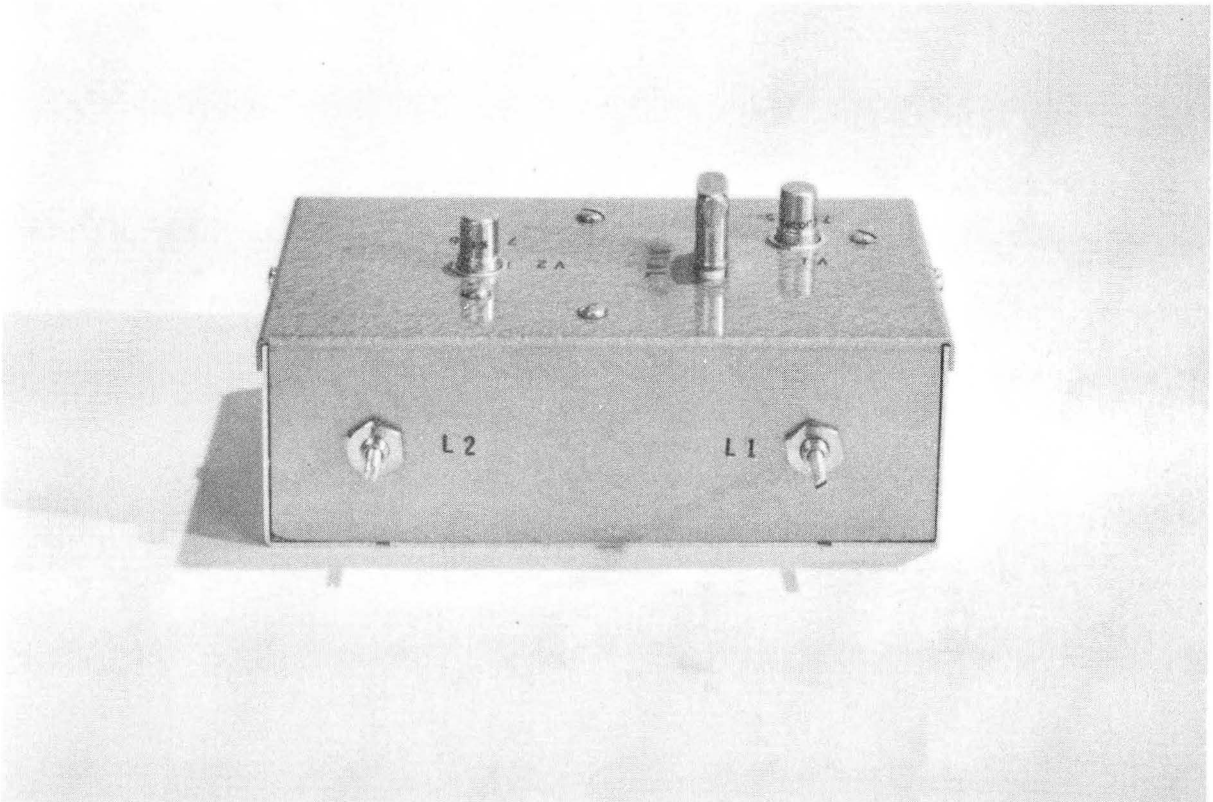
ZN-5388

Fig. 22. Top view of rf detector unit showing signal amplifier only. The angle brackets (Fig. 21) have been removed to reveal the wiring in greater detail. Lead positioning and ground returns are critical. The intercompartment dividers are fabricated from a spare chassis box.



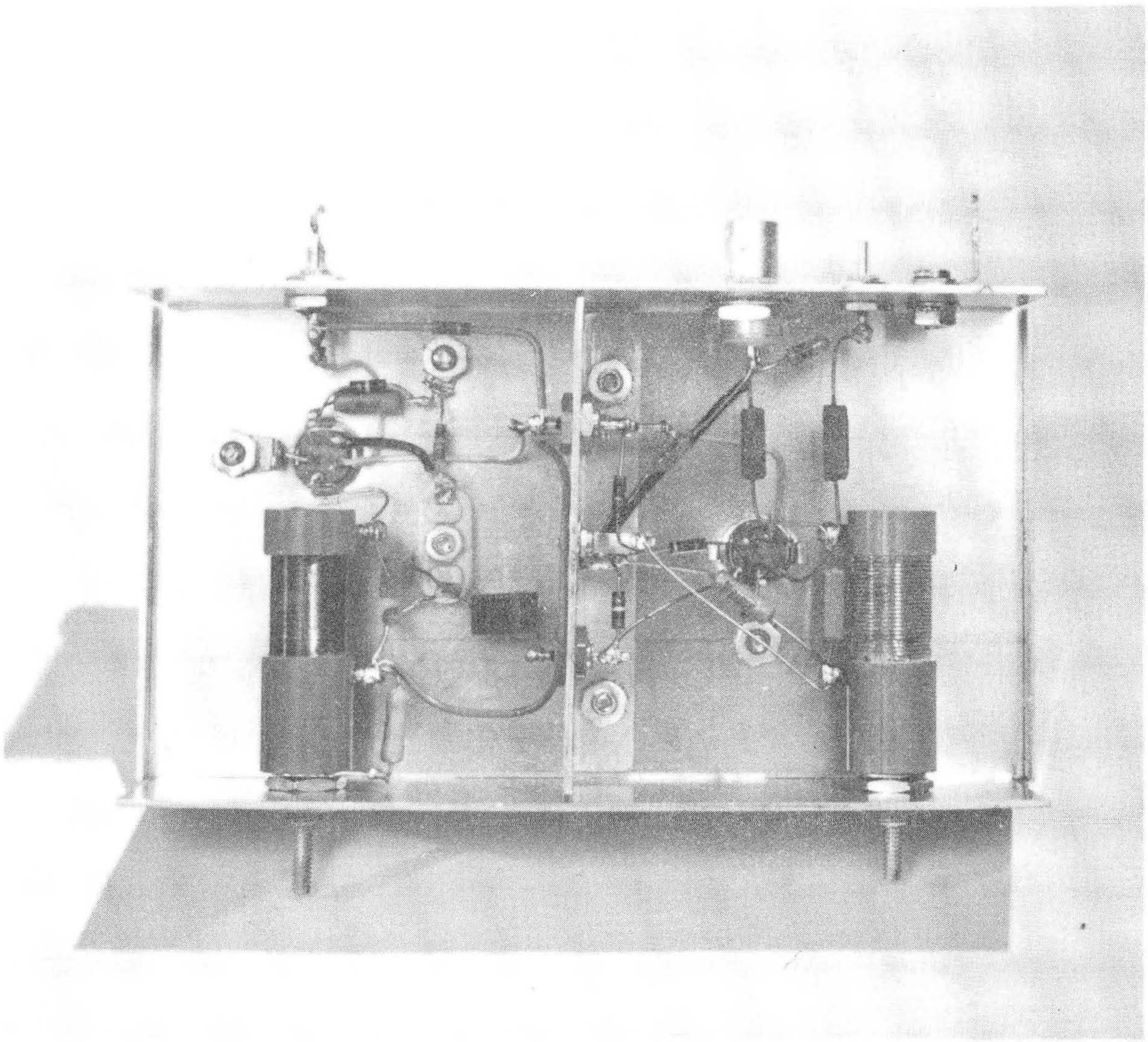
ZN-5387

Fig. 23. Top view of rf detector unit, showing the reference amplifier, left-hand compartment, and the phase-sensitive detector at right.



ZN-5378

Fig. 24. Top front view of rf lock-in generator. The Nuvistors and crystal are mounted on top. Appendix II. F includes layout templates.



ZN-5386

Fig. 25. Bottom view of rf lock-in generator unit. Note lead positioning and ground returns. The BNC connector is insulated from the chassis.

The rf lock-in is built on two commercial chassis boxes--one is LMB type 17 (13 by 4-1/8 by 2-5/8 in.) and the other is LMB type 138 (6-1/4 by 3-1/2 by 2-1/8 in.). Layout is simplified if the appropriate template is folded around each chassis box. This permits one to mark the hole centers with an automatic punch, piercing through the templates at the appropriate places.

The dividing plates, Fig. 35, were fashioned from other miniboxes, LMB type 17. Repeated flexing simplifies separation. Construction is simplified if the plates are wired and stencilled prior to installation. Note that in Fig. 21 angle brackets are mounted on three of the dividing plates. The brackets are cut from aluminum angle, 1/2 by 1/2 by 1/8 in., tapped to receive screws holding the top cover plate down. Intercompartment shielding is critical; tapping is necessary.

D. Tuning Procedure

Appendix II includes detailed descriptions of gain, bandwidth, and linearity measurements.

1. Generator

Follow the instructions in this section (IV. D) by referring to Fig. 17.

a. Monitor either OUTPUT with an oscilloscope probe.
b. Tune L1 and L2 for maximum.

2. Detector Unit

Follow these instructions by referring to Fig. 16.

a. Mount the cover plate, making sure that it is held down with all screws both above and below the chassis box.

b. Depress SW1, monitor the blue test point, TP7, with the oscilloscope probe, then adjust L8 and L9 for maximum at the monitor.

c. Apply an attenuated signal (see Fig. 33 for attenuator) from the generator OUTPUT to input terminals B1 and B2, as shown in Fig. 16.

d. Transfer the oscilloscope probe to the yellow test point, TP3; tune L1 through L6 for maximum. If the signal at TP3 exceeds 20 V, peak-to-peak, more attenuation should be added.

e. Remove probe from yellow test point. Connect a 10-0-10 microampere meter and a series resistor (330 k) between one DET OUT terminal and the white test point, TP6.

f. Tune C for maximum meter deflection. Recheck the tuning of L1 through L6.

g. Transfer the meter and series resistor so that they monitor the voltage across the DET OUT terminals. Release SW1, then tune L9 for maximum.

h. Disconnect signal to B2 input (disconnect cable at attenuator); the meter reading should fall to zero. If it does not, adjust the BAL control until it does. If it is not possible to obtain zero, adjust the position of the L7 slug until zero can be obtained with the BAL control.

i. A change of L7 may require readjustment of C. To check this, repeat steps (c) through (h), above.

E. Specifications

Appendix II includes detailed descriptions of the gain, bandwidth, and linearity measurements.

Performance, as related to the original application of the rf unit, is indicated in Ref. 19. The important specifications for the modified rf unit are listed below:

FREQUENCY: 2.7 MHz.

SIGNAL CHANNEL BANDWIDTH: 220 kHz.

GAIN: (dc out/rms in) approximately 2700, dependent upon source resistance.

EQUIVALENT INPUT NOISE: Not measured. (We were able to detect capacitance changes on the order of 10^{-18} F with the unit of Ref. 19.)

LINEARITY: Approximately 1% of full scale.

ZERO DRIFT: Not measured.

OUTPUT: Linear to at least 2.4 V.

F. Applications

The original application was for a micromanometer, as discussed in Ref. 19. In achieving the design objective, we found that we were able to detect capacitance changes on the order of 10^{-18} F. This unusual sensitivity to dielectric properties could be useful in other areas also; a current application involves the detection of light modulated at 2.7 MHz. We are also considering its application to diagnostic studies of gaseous media.

ACKNOWLEDGMENTS

The af unit was designed for an ESR spectrometer now being developed at this Laboratory. Spectrometer development is under the direction of Professor Harold S. Johnston, Inorganic Materials Research Division.

The rf unit was designed under the direction of Professor D. N. Hanson, Department of Chemical Engineering. This unit was developed in collaboration with Dr. Peter Rony, now affiliated with the Monsanto Chemical Company, St. Louis, Missouri.

Module specifications are reproduced through the courtesy of Engineered Electronics Co., Santa Ana, California.

All work was performed under the auspices of the U. S. Atomic Energy Commission.

APPENDIX I. Audio-Frequency Lock-in Amplifier

A. Module Specifications

The specifications in this section were furnished by Engineered Electronics Company and used with their permission. Figures 26 through 29 were redrawn from data supplied by EEC.

SPECIFICATION: EMITTER FOLLOWER, TRIPLE, NPN TRANSISTOR

I. GENERAL

This is a transistorized assembly containing three independent NPN transistor emitter follower circuits. The purpose is to provide current gain, circuit isolation, or for increasing the load-driving capability of an input signal.

The NPN emitter follower is used principally for driving low impedance loads with positive-going signals. Its best characteristic is its ability to drive capacitive loads such as flip-flops, one shots, etc. This circuit is not used for operating into D. C. logic.

This assembly is contained in a cylindrical plug-in package which inserts into a standard 9-pin miniature tube socket.

II. ELECTRICAL SPECIFICATIONS

A. Input: (Each Emitter Follower)

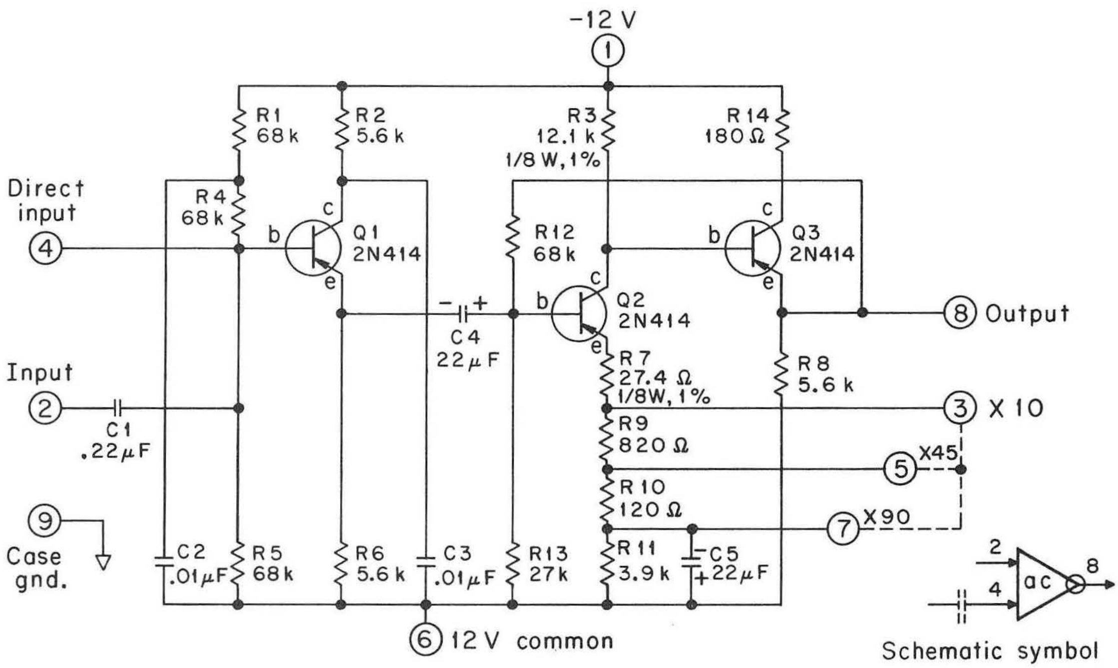
1. Signal Frequency Range: 0 to 250 KC.
2. Signal Amplitude: The standard input is an 8VDC level shift from -11VDC to -3VDC nominal.
3. Input Impedance: Nominally 30K under loaded conditions.
4. Pulse inputs up to 12V peak amplitude may be applied through a suitable capacitor. The input pin must be biased by a resistor return to Pin 1. Such an R. C. input network is also useful for producing pulses from voltage steps.
5. Frequency range extends to 500 KC with reduced loading.

B. Output: (Each Emitter Follower)

1. Amplitude: Equal to input signal. Level shift is 1/4 V in the negative direction.
2. Rise Time: Normally not deteriorated by more than 0.1 μsec referred to input. Under maximum capacitive loading, rise time deterioration will not exceed 0.2 μsec.
3. Loads: Typical, up to 4 paralleled flip-flop inputs. At frequencies to 125 KC, capacitive loading can be as great as .0015 μf. Greater capacitive loads may be imposed at correspondingly lower frequencies. Maximum should not exceed .005 μf.
4. Output Impedance: 150Ω for positive-going signal; 1.8K maximum for negative-going signal.

C. Power Requirements: (Each Emitter Follower)

1. -12VDC at 1 to 7 ma. depending on load. Pin 1 negative with



NOTE: Unless otherwise specified:

1. Resistors are 1/4 W, 5 %.
2. Amplifier voltage gains of X10, X45, X90 selectable by jumpering pins as indicated.
3. Mark: amplifier per A-95695.

MUB-9634

Fig. 26. Schematic of the T-108 linear amplifier module. Redrawn from data supplied by Engineered Electronics Company.

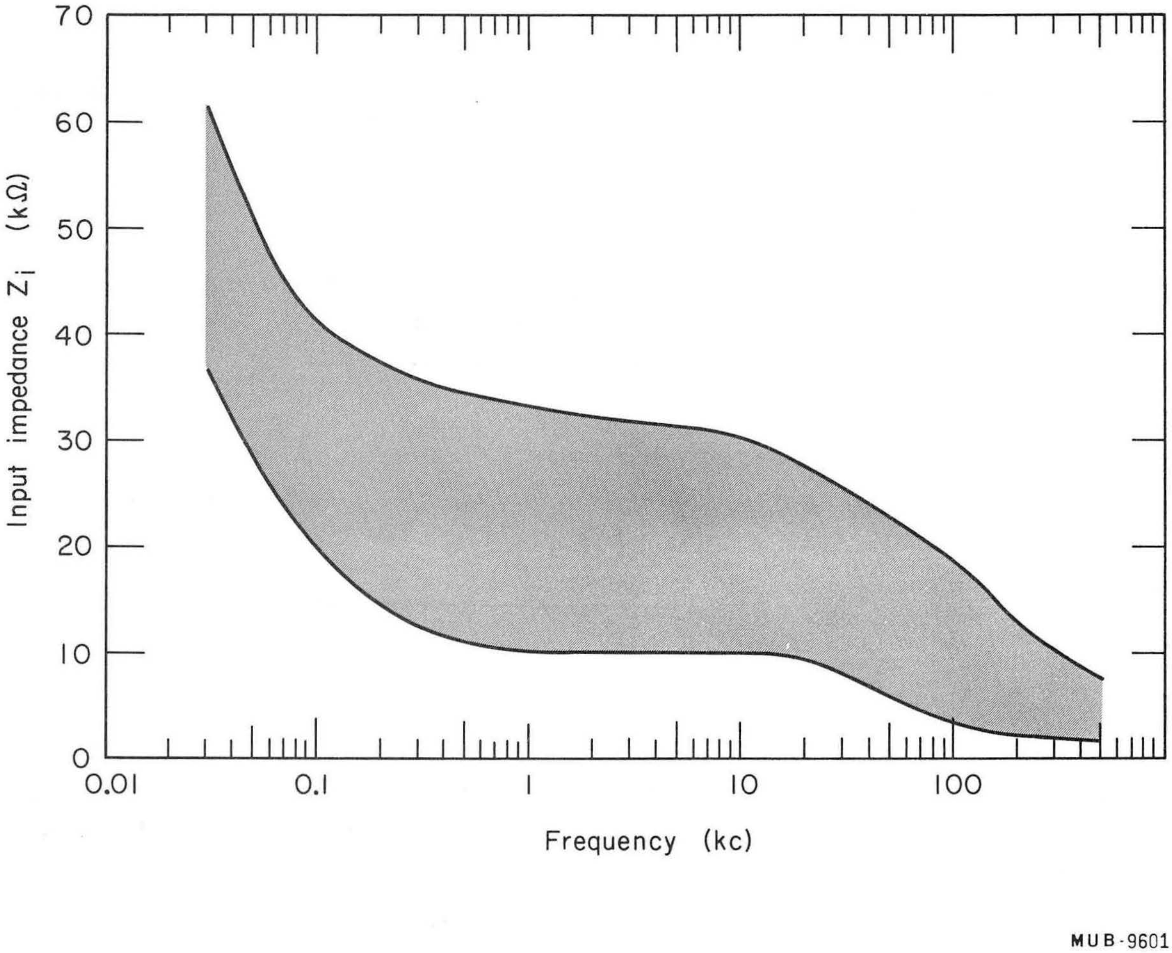


Fig. 27. Input impedance limits (pin 2) vs frequency for T-108 linear amplifier module. Shaded area represents range of input impedance encountered under all conditions outlined in linear amplifier specifications. Redrawn from data supplied by Engineered Electronics Company.

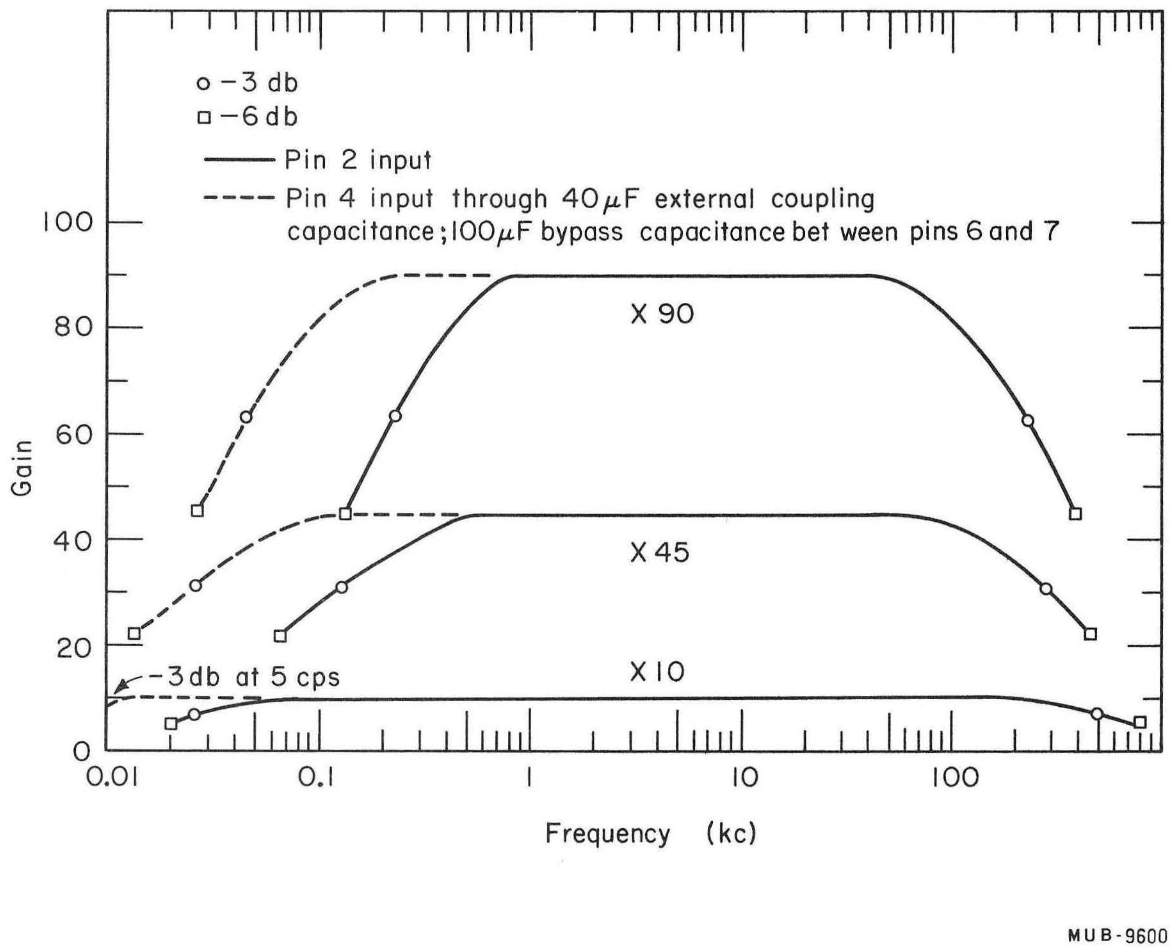


Fig. 28. Gain vs frequency of T-108 linear amplifier module. Redrawn from data supplied by Engineered Electronics Company.

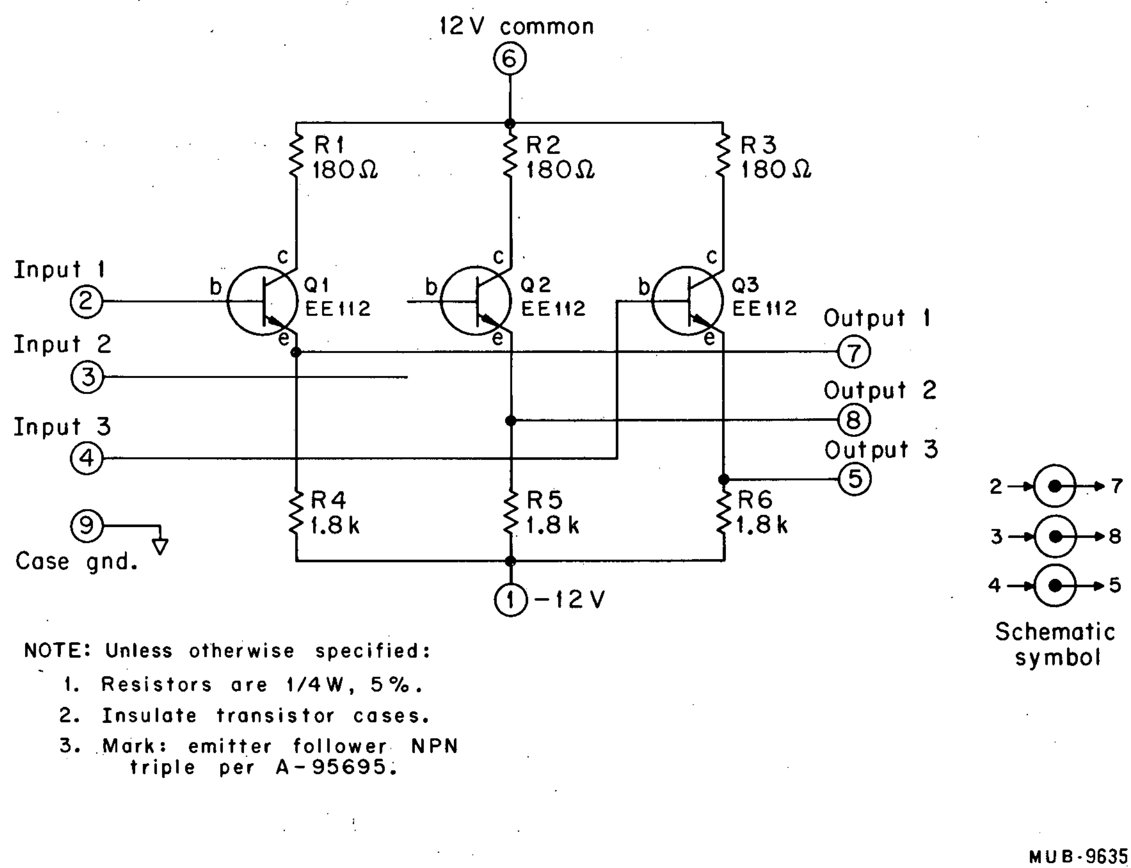


Fig. 29. Schematic of the T-116 emitter-follower module. Redrawn from data supplied by Engineered Electronics Company.

respect to Pin 6.

2. Supply Voltage Tolerance: ± 10%.

III. MECHANICAL SPECIFICATIONS

A. Dimensions: This unit is completely contained within a cylindrical package of 29/32" diameter by 2-1/2" long including the plug-in base. The removable cover is attached to the base by means of a bayonet arrangement, and a locking ring.

B. Mounting: Assembly is mounted by inserting into a standard 9-pin miniature tube socket. Where mechanical retention is required, a standard 2-3/8" noval tube shield is used. Either a J-slot type or a snap-on type shield may be used.

C. Pin Connections are so arranged that power and grounds may be in-line wired.

D. Operating Temperature Range: -54° C. to +71° C.

IV. REFERENCES

A.	Schematic:	SA-95520
B.	Bill of Material:	MA-95519
C.	Application Chart:	ZA-95578

SPECIFICATION FOR LINEAR AMPLIFIER, TRANSISTOR, T-108

I. GENERAL

This unit is a transistorized assembly containing three germanium transistors in a linear amplifier circuit. The purpose is to increase the amplitude of small input signals. Both A.C. and D.C. feedback are incorporated for stability of operation.

Inputs can be sine waves, square waves, pulses or any complex wave forms provided the frequency components of the signal are within the response range of the amplifier.

Inputs can be derived from a variety of external sources including voltage pickups and low level tranducers. Signals as small as 50 mv PP can be amplified to a 5 V PP level.

Features include high input impedance, low output impedance, three selectable fixed gains, provisions for obtaining any intermediate gain by use of external resistors and provisions for extending low frequency response by use of external capacitors.

This assembly is contained in a cylindrical plug-in package which inserts into a standard 9-pin miniature tube socket.

II. ELECTRICAL SPECIFICATIONS

A. Input:

1. Signal Frequency Range: 30 cps to 500 kc sine wave. Low

frequency limit is extended when using large external capacitors. (See Page 44 for response curves.)

2. Amplitude Range: 5 mv to 0.6 V PP depending upon gain, for linear operation.
3. Maximum Amplitude: 12 V PP.
4. Input Impedance: 10KΩ or better at 1 kc at any gain. See Fig. 27 for limits of input impedance versus frequency.
5. Wave Form: Sinusoidal, or complex if the frequency components are within response range of the amplifier.

B. Output:

1. Type: One emitter follower output. D.C. level is -7.0 V D.C. nominal.
2. Phase: Output signal is inverted (180°) relative to input.
3. Amplitude: 6 V peak to peak maximum undistorted.
4. Gain: Any of three voltage gains may be obtained by jumpering socket pins as indicated below:

Selected Gain	Pins Jumpered
10	None
45	3 and 5
90	3 and 7

Variable gain may be obtained by connecting the arm of a potentiometer to Pin # 7 and one end to Pin # 3. Fixed intermediate gains may be obtained by connecting a suitable resistor between Pin # 3 and Pin # 7.

6. Loads: Typical load is a squaring amplifier unit.
7. Response: The frequency response at each gain is shown in Fig. 28. Low frequency response may be extended by using large external input capacitor to direct input (Pin #4) and by adding external capacitor between Pin #6 and Pin # 7. Correct capacitor polarity must be observed.
8. Random Output Noise: 20 mv PP.

C. Power Requirements:

1. -12 V DC at 2.5 ma.
2. Supply voltage tolerance: ±10%.

III. MECHANICAL SPECIFICATIONS

A. Dimensions: This unit is completely contained within a cylindrical package of 29/32" diameter by 2-1/2" long including the plug-in base. The removable cover is attached to the base by means of a bayonet arrangement and a locking ring.

B. Mounting: Assembly is mounted by inserting into a standard 9-pin miniature socket. Where mechanical retention is required, a standard 2-3/8" noval tube shield is used. Either a J-slot type or a snap-on type shield may be used.

C. Pin connections are so arranged that power and grounds may be in-line wired.

D. Operating Temperature Range: -45° C. to +65° C.

IV. REFERENCES

A. Schematic: SA-95496

B. Bill of Material: MA-95495

C. Application Chart: ZA-95578

B. Test Attenuator

A schematic diagram of the test attenuator is shown in Fig. 30.

C. Gain and Bandwidth Measurements

Most of the following measurements were made with a Hewlett-Packard oscillator, Model 200 AB. The test attenuator (Fig. 30) was connected between oscillator output and SIG IN. The signals were monitored with a Tektronix Oscilloscope, Model 545, with a type C preamplifier and X10 probe.

1. First Stage

a. Parallel-T removed, A1 GAIN set at maximum clockwise. Maximum gain measured was 35. This figure includes attenuation losses due to the divider formed by R1 (Fig. 2) and the input impedance of a T-108 module. (Divider output is approximately 75% of input with the parallel-T removed.)

b. Parallel-T, 560 hertz inserted. The first stage became regenerative (gain increased with parallel-T in), so A1 GAIN was reduced until the signal at the blue test point was the same (2 volts pp) with the T in or out. This adjustment reduced gain from 35 to 26. (Effective Q with the new adjustment measured was 28; i.e., the center frequency was 560 hertz, the 70% frequencies were 550 and 570 hertz.)

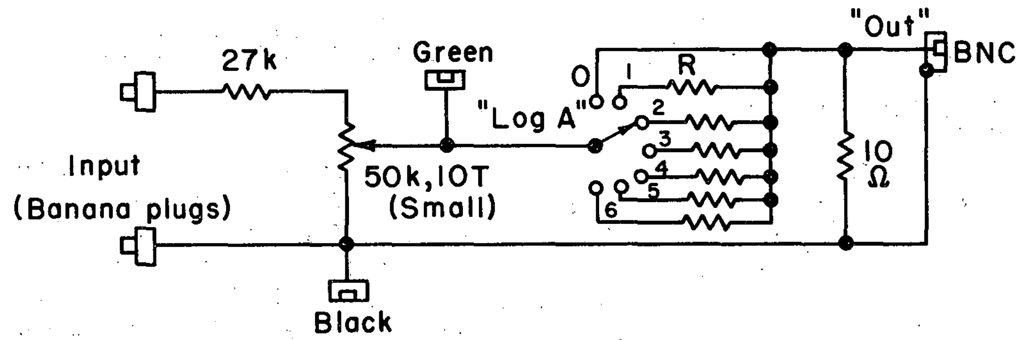
2. Second Stage

a. The gain was 90.

3. Phase-Sensitive Detector and Transformer T1

a. Voltages at the red and yellow test points measured 2.5 volts dc (relative to the black) without an input signal. The output voltage at each test point varied approximately 0.1 volt when the PHASE control was adjusted about its limits. These voltage measurements were made with a 10-0-10 microampere meter and a series resistor of 330 k (E = IR). DET was at N, C3 and C4 of Fig. 2 were 0.01 μF, and the source-follower transistors (Q5 and Q6) were removed.

b. The meter and 330 k resistor were transferred to monitor the voltage across the red and yellow test points, and the BAL control adjusted for zero. The test attenuator (Fig. 30) was connected from REF OUT to SIG IN, and its attenuation adjusted for 2 volts pp at the blue test point. When the PHASE control was adjusted for maximum meter deflection, DET OUT was 0.9 volt. Therefore dc out/rms in measured 1.3, the combined gain of the detector and transformer T1.



Log A setting	R
0	0
1	100 Ω
2	1 k
3	10 k
4	100 k
5	1 MΩ
6	10 MΩ

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Fig. 30. Schematic diagram of attenuator used for testing af unit. The attenuation values are approximate. The attenuator input resistance is high to prevent overloading at REF OUT. All resistors are $\frac{1}{4}$ W carbon.

4. Overall Gain

Combining the gains of 1 through 3, above, we obtained an overall gain, dc out/rms in, of approximately 3000.

D. Noise, Drift, and Linearity Measurements

1. Noise Measurements

The GAIN control was set at maximum clockwise, the test attenuator (Fig. 30) connected between REF OUT and SIG IN, and the attenuation adjusted for 2 volts pp at the blue test point (560 hertz parallel-T still in). A recording potentiometer (1 mV full-scale deflection) was connected to monitor the source-follower output (REC). TIME CONSTANT was set for 2 sec, and the PHASE control adjusted for maximum recorder deflection. REC LEVEL was then adjusted for a chart deflection of 94 divisions. The input signal was disconnected, SIG IN was shorted to ground, and the resulting fluctuations were recorded. When so adjusted, each chart division corresponded to approximately 3 μV rms, referred to lock-in input. The subsequent recording fluctuated approximately one-third of a division, indicating a 1-μV rms noise level, referred to the input. If converted to a bandwidth of 1/40th hertz (noise inversely proportional to square root of bandwidth), this value corresponds to approximately $\frac{1}{2}$ μV rms.

2. Drift Measurements

Zero drift was measured in the same manner as noise. The recordings indicated that drift was less than 1 chart division per hour, or less than 1% of full scale per hour.

3. Linearity Measurements

Linearity was measured with the set-up indicated for noise measurements. The test attenuator (Fig. 30) was adjusted for 2 V pp at the blue test point; REC LEVEL was then adjusted for a chart deflection of 96 divisions. The attenuator was adjusted for various output levels (never more than 2 V pp at the blue test point), and the recorder deflection noted. Output linearity measured better than 1%.

Linearity is limited by the signal amplifier, not by the phase detector. If output at the blue test point is appreciably greater than 2 V pp, the signal amplifier becomes nonlinear.

E. Templates

In this section are shown templates (Figs. 31 and 32) designed to aid in construction of the audio-frequency lock-in amplifier. (These templates are on paper that has been perforated for easy removal from the manual.)

II. Radio-Frequency Lock-In Amplifier

A. Test Attenuator

A schematic diagram of the test attenuator is shown in Fig. 33.

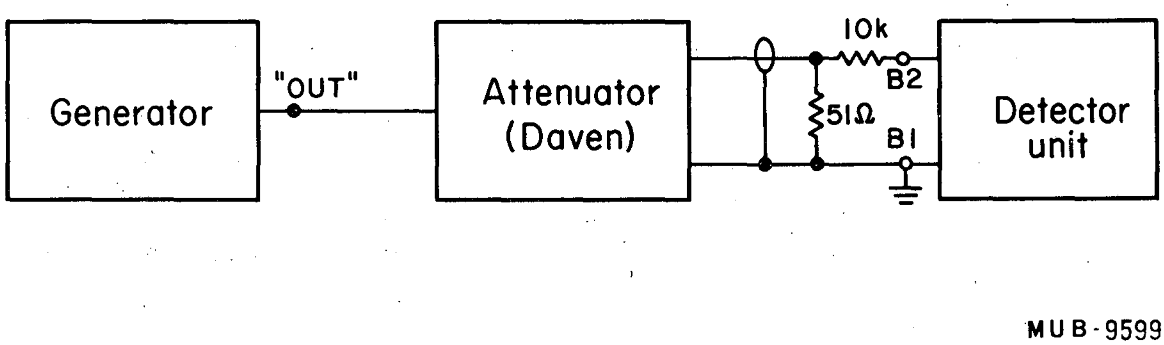

Fig. 33. Attenuator circuit useful for alignment, gain, and linearity tests of rf detector unit.

B. Gain and Bandwidth Measurements

Most of the following measurements were made with a Measurements Corporation Standard Signal Generator, Model 82. The signal generator output cable was terminated with a 51-Ω resistance. Signals developed across the 51-Ω resistor were applied to B2 through a series resistance of 10 k (B1 grounded). L1C1 was connected parallel resonant as shown in Fig. 16. The signals were monitored with a Tektronix oscilloscope, Model 545, with a type C preamplifier and X10 probe.

1. First Stage

a. The gain of V1, excluding that from transformer T1, measured 46.

b. Transformer T1 gives voltage stepdown when operated with the impedance levels selected (10 k series resistor to generator and 5.1 k across L2). The voltage across L2 measured 0.22 times that appearing at the generator output. If L1 were series resonant, and if the 10-k resistor were shorted, G_{max} of T1 would approach 5 when the source resistance was 51 Ω.[25]

c. Gain of the first stage, V1 plus transformer T1, measured 8.8

2. Second Stage

a. The gain of V2, excluding that from transformer T2, measured 41.

b. Transformer T2 gives voltage stepdown when operated with the impedance levels shown in Fig. 16 (5.1 k resistors across L4 and L5). The voltage across L5 measured 0.41 times that appearing across L4.

c. Gain of the second stage, V2 plus transformer T2, measured 15.

3. Third Stage

a. The gain of V3, excluding that from transformer T3, measured 50.

4. Combined Gain and Bandwidth of First Three Stages

a. Gain up to the yellow test point measured ≈ 7000.

b. We obtained maximum response at the yellow test point with the signal generator tuned to 2.86 MHz. The 70 percent frequencies measured 2.76 and 2.98 MHz.

5. Synchronous Detector

a. Voltages at the DET OUT terminals measured relative to the white test point were 7.4 volts dc without an input signal. These measurements were made with a 10-0-10 microampere meter and two series resistors, 510 k each (one resistor to each side of meter).

b. The meter (plus two 150 k series resistors) was transferred to monitor the voltage across the DET OUT terminals, and the BAL control adjusted for zero. A radio-frequency attenuator was connected as shown in

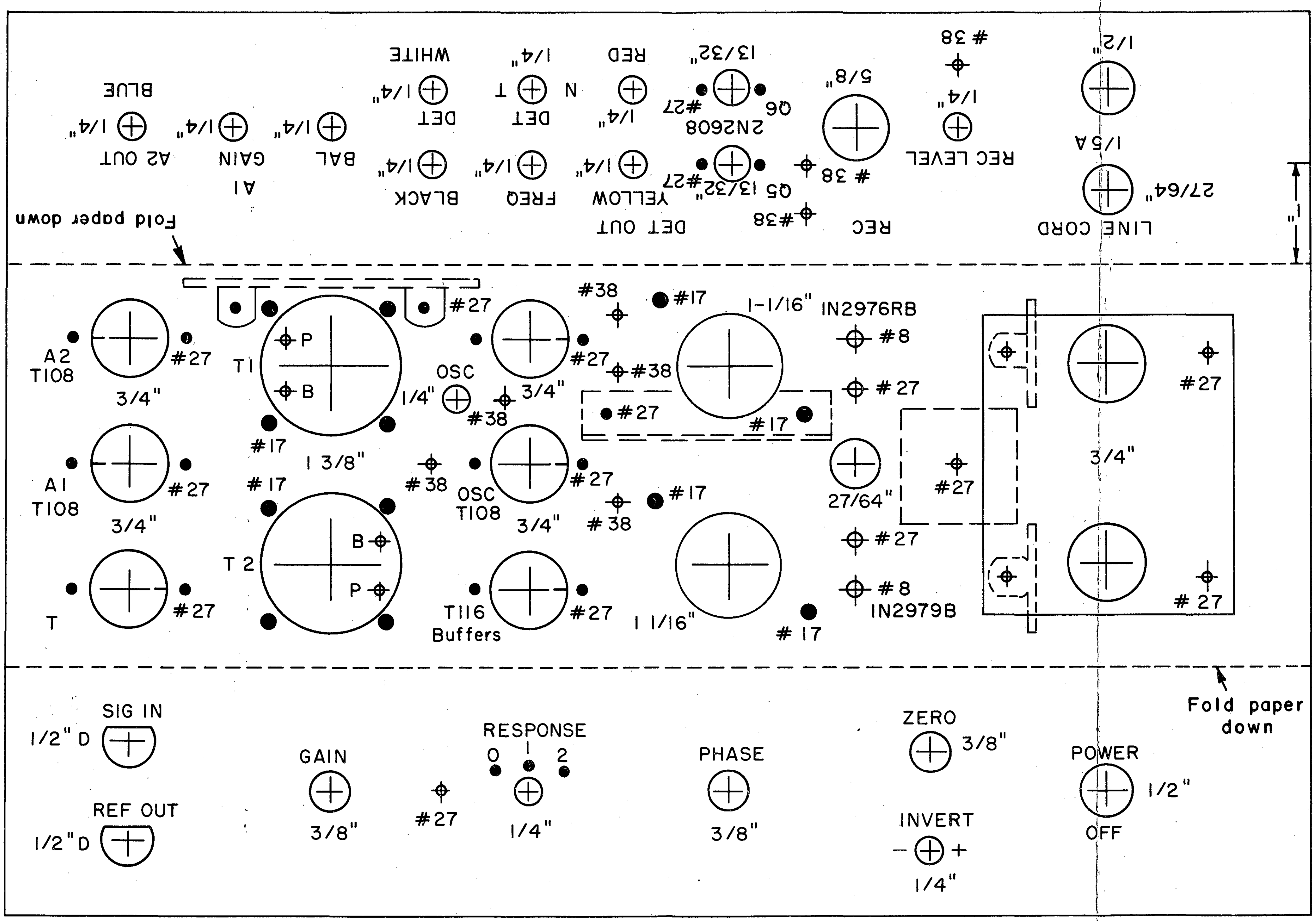

Fig. 31. Layout template for af unit. The template is folded around the chassis box as shown. Hole centers are marked with an automatic punch, which pierces through the template at the appropriate places. The darkened holes must match other components; their location should be verified prior to drilling. Drill sizes are indicated by numbers shown. The sockets and transformers should be oriented as shown.

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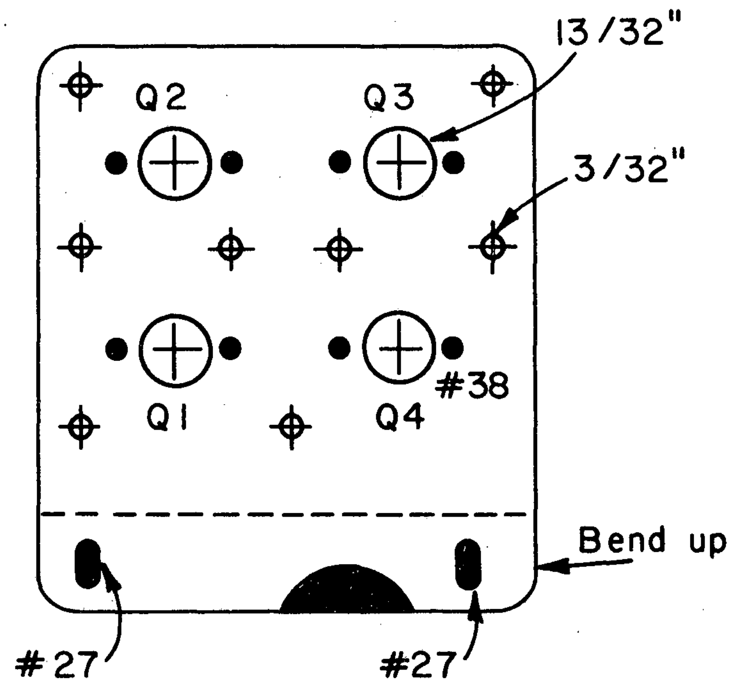

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Fig. 32. Layout template for af lock-in phase shifter. The phase shifter should be wired and stencilled prior to mounting. The darkened holes must match other components. Drill sizes are indicated by numbers shown.

Fig. 33, and its attenuation adjusted for 19 V pp at the yellow test point (attenuation 42 db). When L9 was adjusted for maximum meter deflection, DET OUT measured 2.4 volts. Therefore dc out/rms in measured 0.36, the combined gain of the detector and transformer T3.

c. Transformer T3 gives a voltage stepdown when operated with the impedance levels shown in Fig. 16 (5.1 k resistors across L6 and L7). The voltage across L7 measured 0.44 times that appearing across L6.

6. Overall Gain of Signal Channel

a. Combining the gains of 4 and 5 above, we obtain an overall gain, dc out/rms in, of approximately 2400. The actual gain depends upon the source resistance that drives the rf lock-in unit.

b. The overall gain was also measured in one operation, as follows:

1) The radio-frequency attenuator was connected as shown in Fig. 33, and its attenuation adjusted for approximately 20 volts pp at the yellow test point (attenuation was 44 db). The attenuator input, monitored with a scope probe, measured 0.4 volt pp.

2) The meter (10-0-10 microampere and two 150 k series resistors) was connected to monitor the voltage across the DET OUT terminals. When L9 was adjusted for maximum meter deflection, DET OUT measured 2.4 volts dc (E = IR). When the attenuation is taken into consideration, dc out/rms in measured 2700.

7. Reference Amplifier

a. Reference input measured 0.5 volt pp.

b. Blue test point measured 27 volts pp with SW1 depressed and L8 and L9 tuned for maximum.

c. Reference amplifier gain, V4 plus transformer T4, therefore measured 54.

C. Linearity Measurements

Linearity was measured with the set-up described in Appendix II. A. 6. b. The attenuator was set to the values indicated below, and the resultant meter deflections recorded, as follows:

Attenuator setting (db)	Meter current (μA)	Theoretical current (μA)
47	6.25	6.25
57	1.9	1.965
67	0.6	0.625
77	0.2	0.1965
87	0.005	0.006

It is difficult to read the meter with precision, but the above figures indicate that linearity is approximately 1%.

D. Generator Voltage and Current Measurements

The following measurements are reproduced from Ref. 18. The measurements were not repeated after modification, but the values from both the modified and unmodified versions should be reasonably close.

1. Current Measurements

   a. Total plate current – 11.5 mA.

   b. Oscillator plate current – 0.25 mA.

   c. Buffer plate current – 11.25 mA.

2. Radio-Frequency Voltage Measurements (531 scope, CA plug-in) and (7 pF × 10 probe)

   a. Maximum possible voltage at plate of V1 = 27 volts pp.

   b. Maximum possible voltage at grid of V2 = 0.9 volt pp, 1.35 volts pp if V2 removed.

   c. Maximum possible voltage at plate of V2 = 8.8 volts pp.

   d. Maximum possible voltage at OUTPUT = 0.8 volt pp.

E. Coil Measurements

1. 25 turns, No. 22 Formvar, National Radio XR50 form

   a. Inductance range: 4.9 to 8.7 μH

   b. Q measurements

   Boonton Q meter, type 160 A. Values of L are not correct except at 7.9 MHz.

   1) Q = 190 at 7.9 MHz
      L = 5.45 μh
      C = 74.3 pF
   2) Q = 144 at 3.14 MHz
      L = 5.45 μH
      C = maximum at Q meter, over 450 pF
   3) Q = 117 at 2.765 MHz
      L = not measured
      C = maximum at Q meter, over 450 pF
   4) Q = 111 at 2.765 MHz
      L = not measured; resonates 450 pF to frequency
      C = 450 pF

2. 64 turns, No. 32 Formvar; National Radio XR50

   a. Inductance range: 30 to 56 μH

   b. Q measurements

   Boonton Q meter, type 160 A. Values of L are not correct except at 2.5 and 7.9 MHz.

   1) Q = 131 at 2.5 MHz
      L = 57.5 μH
      C = 70.7 pF
   2) Q = 131 at 2.75 MHz
      L = 57.5 μH

C = 56.8 pF
3) Q = 155 at 2.765 MHz — C set at 68 pF, L adjusted for resonance
L = 59.5 μH
C = 68 pF
4) Q = 152.5 at 2.5 MHz
L = 48 μH
C = 85 pF

F. rf Transformer Measurements

Inductance was measured on two coils, each with 64 turns of No. 32 Formvar wire wound on National Radio form XR50; coils spaced 5/8 in. between centers. Inductance measurements were made with a Tektronix LC meter, type 130, and with coil slugs set at various positions. (M = mutual inductance; K = coefficient of coupling = $M/\sqrt{L_1 L_2}$; L_1 is inductance of the primary, L_2 inductance of the secondary.)

1. Both slugs fully out

a. Measurements

$L_1 = L_2 = 32.5\ \mu H$ (inductance, each coil)

$L_1 + L_2 + 2M = 66.5\ \mu H$ (inductance, series-aiding)

$L_1 + L_2 - 2M = 62\ \mu H$ (inductance, series-opposing)

b. Conclusions

$M = 1.125\ \mu H$

$K = 0.0346$

2. a. Measurements

$L_1 = L_2 = 34\ \mu H$ (inductance, each coil)

$L_1 + L_2 + 2M = 72\ \mu H$ (inductance, series-aiding)

$L_1 + L_2 - 2M = 66\ H$ (inductance, series-opposing)

b. Conclusions

$M = 1.5\ \mu H$

$K = 0.044$

3. a. Measurements

$L_1 = L_2 = 40\ \mu H$ (inductance, each coil)

$L_1 + L_2 + 2M = 82\ \mu H$ (inductance, series-aiding)

$L_1 + L_2 - 2M = 74\ \mu H$ (inductance, series-opposing)

b. Conclusions

$M = 2\ \mu H$

$K = 0.05$

4. a. Measurements

$L_1 = L_2 = 50\ \mu H$ (inductance, each coil)

$L_1 + L_2 + 2M = 103\ \mu H$ (inductance, series-aiding)

$L_1 + L_2 - 2M = 90\ \mu H$ (inductance, series-opposing)

b. Conclusions

$M = 3.25\ \mu H$

$K = 0.065$

G. Templates

In this section are templates (Figs. 34 through 36) designed to aid construction of the rf unit. (These templates are on paper that has been perforated for easy removal from the manual.)

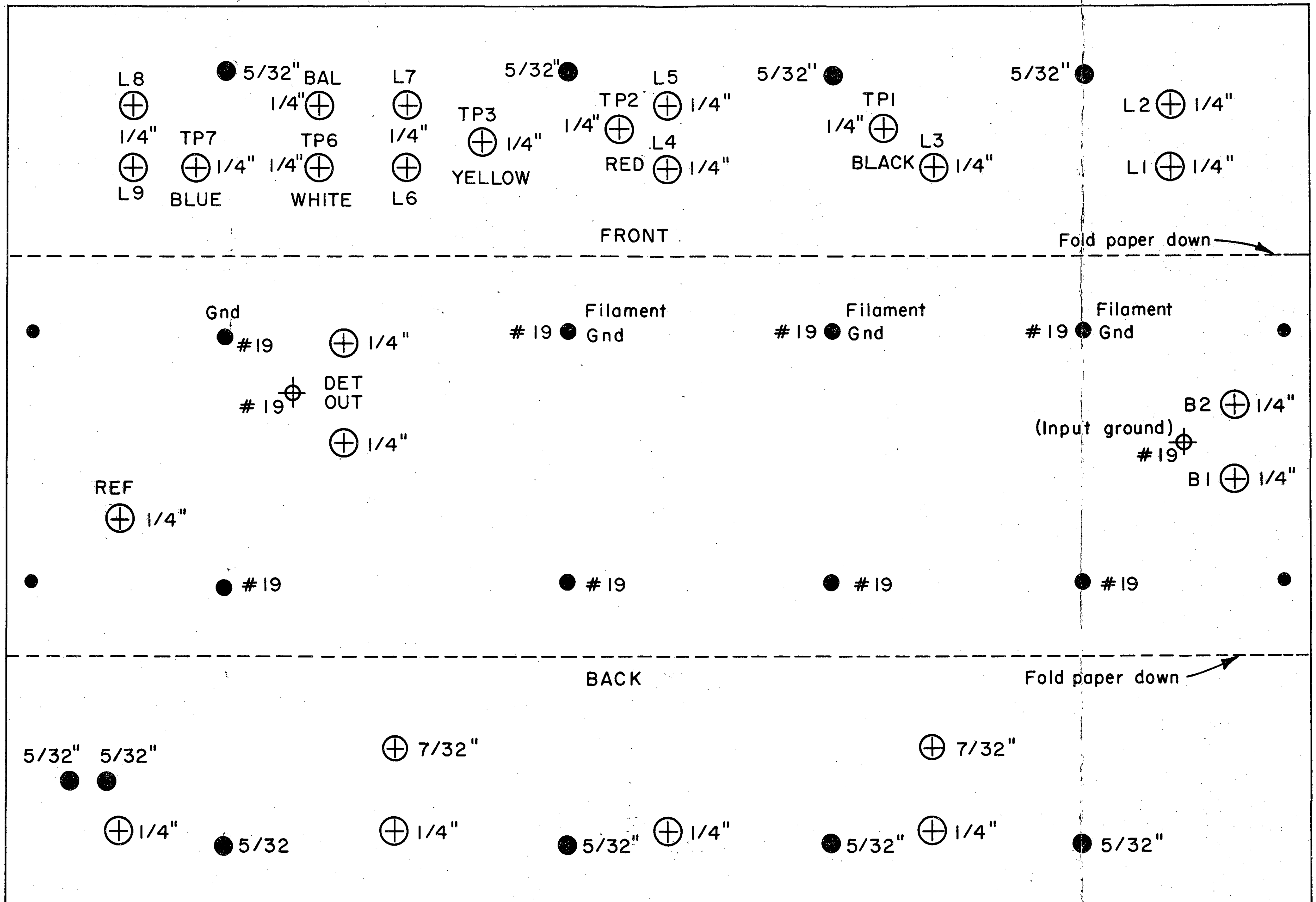

Fig. 34. Layout template for rf lock-in detector unit. The template is folded around the chassis box as shown. Hole centers are marked with an automatic punch, which pierces through the template at the appropriate places. The darkened holes must match other components; their location should be verified prior to drilling. Drill sizes are indicated by numbers shown.

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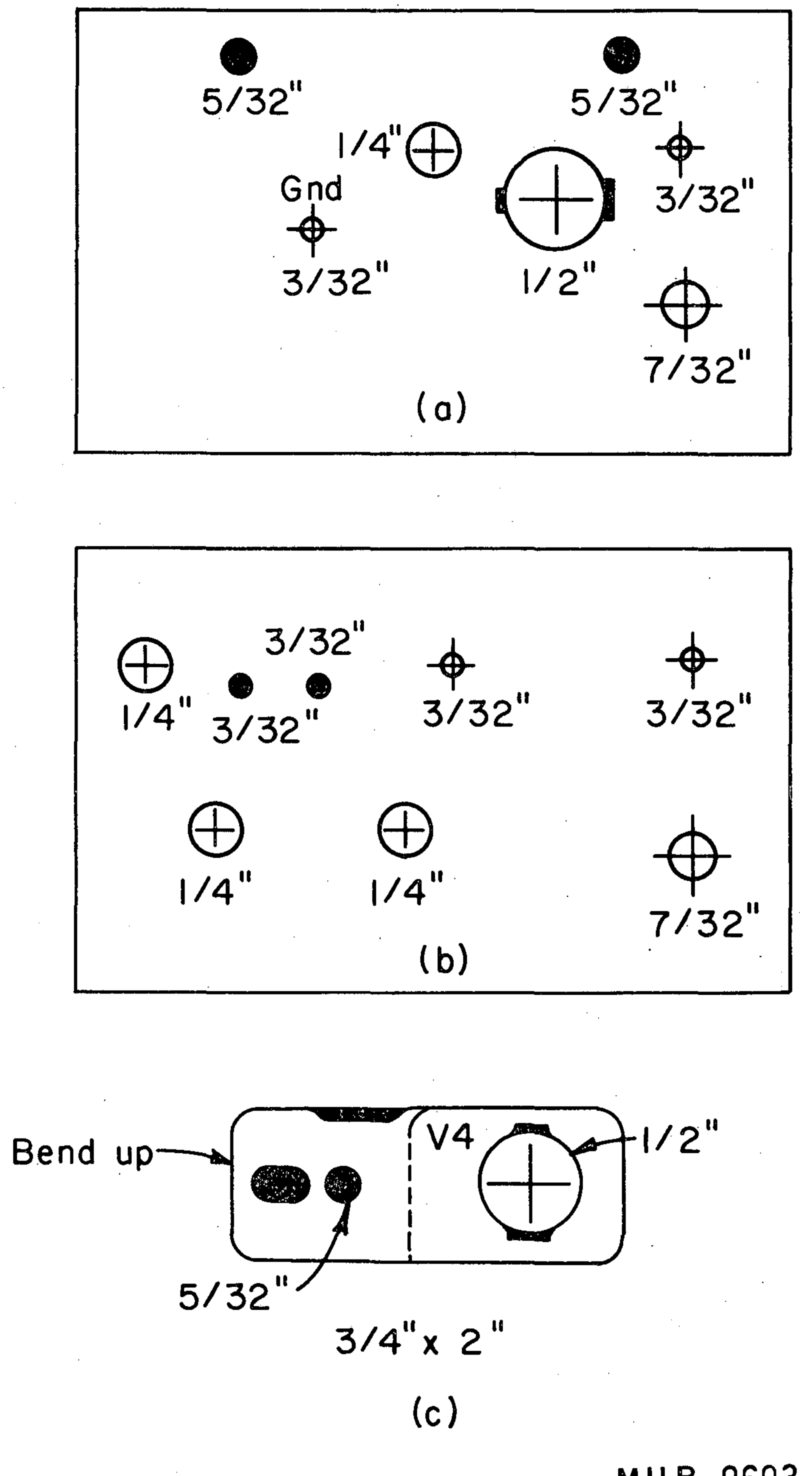

Fig. 35. Layout templates for rf lock-in detector unit: (a) and (b) intercompartment plates, and (c) V4 bracket. Darkened holes must match other components; their location should be verified before drilling. Drill sizes are indicated by numbers shown. Construction is simplified if the plates are wired and stencilled before installation.

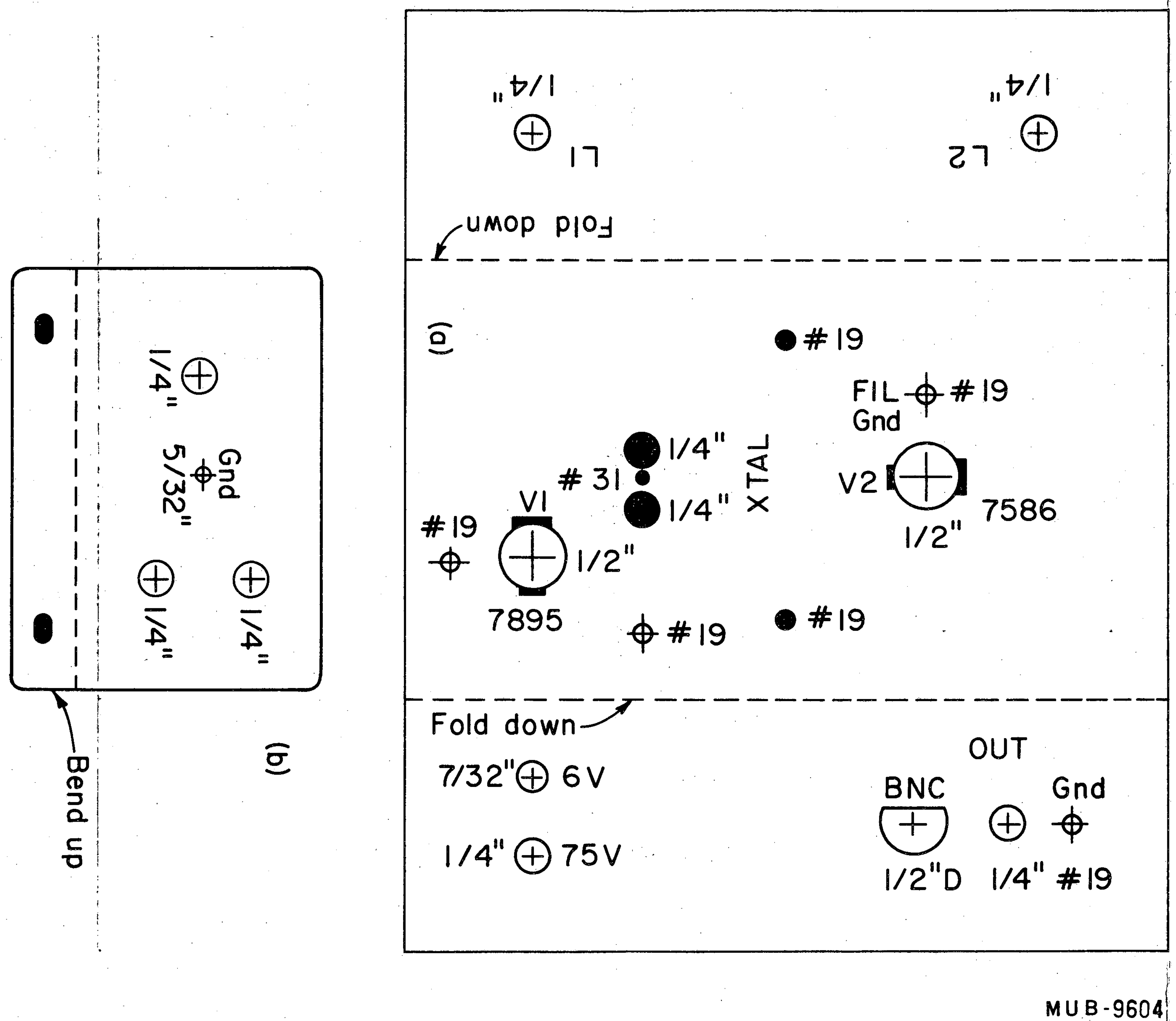

Fig. 36. Layout templates for rf lock-in generator unit: (a) generator, and (b) intercompartment plate.

FOOTNOTES AND REFERENCES

1. R. D. Moore, Lock-in Amplifiers for Signals Buried in Noise, Electronics (June 8, 1962).

2. R. D. Moore and O. C. Chaykowsky, Modern Signal Processing Technique for Optimal Signal to Noise Ratios, Princeton Applied Research, Technical Bulletin 109 (1963).

3. Lock-in amplifiers are commonly driven by other devices that generate considerable noise, a receiver for example. The lock-in technique minimizes noise generated by devices preceding it, provided that those noises are not modulated by the reference.

4. Noise per unit bandwidth is the same at all frequencies.

5. Harold S. Black, Modulation Theory (D. Van Nostrand Company, Inc., New York, 1953), p. 39.

6. ESR spectrometers are commonly modulated at 100 kHz in order to minimize 1/f noise generated by crystal detectors preceding the lock-in.

7. G. Feher, Sensitivity Considerations in Microwave Paramagnetic Resonance Absorption Techniques, Bell System Technical Journal, 480, (March 1957).

8. This can be explained as follows: $G = A/(1 + A\beta)$. With a perfect network, beta is zero at the tuned frequency, so G is A. At frequencies far removed from zero, beta approaches 1 and G approaches $A/(1 + A)$. Maximum gain change is approximately A.

9. R1 is added to ensure that the minimum impedance shunting the input of A1 is a reasonable value.

10. R. Landee, D. Davis, and A. Albrecht, Electronic Designers' Handbook, (McGraw-Hill Book Company, Inc., New York, 1957), pp. 16-24.

11. Selection of Twin-T Networks for Feedback Amplifiers, White Instrument Laboratories, Network Notes No. 1, (July 1955).

12. Application Notes on Synchronous Detection Systems and Associated Instrumentation, Triconix, Inc., Waltham, Mass. (after 1960).

13. A transformer does not couple a dc component, nor does capacitive coupling.

14. When so operated, the source-follower transistors (Q5 and Q6 of Fig. 2) are removed. The black test point is disconnected from ground, and output is taken from DET OUT terminals.

15. See Ref. 10, pp. 6-40.

16. Oscillators and Audio Signal Generators, Hewlett-Packard Catalog No. 25, p. 248, 1965. Palo Alto, California.

17. Handbook of Operational Amplifier Applications, Burr-Brown Research Corp., Tucson, Arizona (1963), p. 57.

18. Kenneth W. Lamers, The Design, Construction, and Operation of a Differential Micromanometer. Part I. Electronics, UCRL-11218, (October 1964).

19. Peter R. Rony, Design, Construction, and Operation of a Differential Micromanometer. Part II. Theory and Operational Characteristics, UCRL-11218, (April 1965).

20. See Ref. 10, pp. 7-53.

21. F. E. Terman, Radio Engineering, Third Edition (McGraw-Hill Book Company, Inc., New York, 1947), p. 63.

22. Ibid, pp. 32 and 357.

23. Except when the source resistance is equal to the reactance of L1.

24. See Ref. 21, p. 54.

25. If $L_s = L_p$, Eq. (5) can be reduced to $G_{max} = \frac{1}{2}\sqrt{\frac{R2}{R}}$, where R2 is the equivalent resistance in shunt with L_s, and R is the equivalent resistance in series with L_p.