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Author
Lamers, Kenneth W.

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University of California

Ernest O. Lawrence
Radiation Laboratory

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Kenneth W. Lamers

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ABSTRACT

We describe two specific lock-in amplifiers, one designed for radio frequencies, the other for audio. Both have been built and tested by chemists with limited electronics backgrounds. Design philosophy is emphasized for those who wish to extend performance or to design their own.

The af unit was designed for an electron-spin-resonance spectrometer, locking a klystron's frequency to that of a resonant cavity. It has been adapted to nuclear-magnetic-resonance work by combining it with a marginal oscillator. The af unit is designed around circuit modules, so that much of its circuitry need not be fabricated. Performance compares favorably with specifications published for some commerical lock-in amplifiers. The af unit is designed for 40 to 20,000 Hz, but its range is easily extended. A small preamplifier for extending sensitivity to microvolt levels is also described.

The rf unit was designed for measuring differential pressures on the order of 0.1 μ with an accuracy of several percent. In achieving the design objective, we were able to detect capacitance changes on the order of 10^{-18} F. Our rf unit operates at 2.7 MHz, but it can be.

adapted to other radio frequencies. It uses Nuvistor tubes to minimize

the instability problems normally encountered at radio frequencies.

I. INTRODUCTION

High-performance lock-in amplifiers are commercially available, but in many cases the price is prohibitive for experimenters with limited funds. Furthermore, some of these units incorporate features unnecessary to a given application, and the additional complexity is undesirable. Also, commerical lock-in amplifiers do not encompass the entire radio-frequency spectrum. In some cases one has no alternative but to design his own. In any event, the experience gained in building and trouble-shooting one's lock-in is invaluable in understanding the instrumentation requirements generally associated with it. The lock-in amplifiers described here are relatively easy to build and test. Proper testing requires an oscilloscope and a signal generator, but our chemists have had little trouble in adapting to these instruments. Additional information in the form of photographs, templates, construction details, and test procedures is available.[1]

The principal merit of the af unit is its simplicity. This simplicity stems from its construction; i.e., it is designed around commerical circuit modules, so that much of its circuitry need not be fabricated. The unit is solid-state, and includes its own power supply. We also describe a small preamplifier for extending sensitivity to microvolt levels.

Performance of the af lock-in amplifier compares favorably with specifications published for some commercial units, but ours is less flexible in selecting frequency. It is best used in applications where 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 probably will choose to establish frequency and sensitivity requirements with a commercial lock-in amplifier before building their own.

The radio-frequency lock-in amplifier 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 later.

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 does it better. The principal difference is that lock-in amplifiers are phase sensitive. Also, they can operate at bandwidths that 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; and (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 reduce 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] most amplifying devices have noise spectra that vary as the reciprocal of frequency;[6] and (c) translation to a higher frequency moves the signal to a frequency where less noise is introduced by the amplifying devices used. In general, the modulating frequency should be greater than 100 Hz when vacuum-tube amplifiers are used and greater than 1000 Hz when transistor amplifiers are used.

III. AUDIO-FREQUENCY LOCK-IN AMPLIFIER

A. General

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 (Fig. 2) uses three linear-amplifier modules. 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. Most frequency-determining elements (a parallel-T network for the

first amplifier stage and a network for the oscillator) are fabricated

into blank containers supplied by the module manufacturer.

B. Circuit Description

1. Signal Amplifier

The signal amplifier comprises two stages, the first one tuned

to prevent noise and spurious pickup (such as 60 Hz) from saturating

the amplifier stage that follows it. Bandwidth of the system, however,

is normally determined by the time constant following the detector

(Fig. 1).[7] Frequency characteristics of the first stage are achieved

with negative feedback through a parallel-T network. Characteristics

measured when it is tuned to 560 Hz are indicated in Fig. 3. 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 3 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 (added to ensure that the minimum

impedance shunting the input of A1 is a reasonable value) and the

network which reduces the signal actually applied to the input of A1.

Asymmetry is not objectionable, however; the reduced response at

high frequencies enhances lock-in performance (by reducing noise)

even more than if the response were symmetrical.

Design equations for the parallel-T network are:[9]

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

$$RC = \omega_0^{-1}, \tag{2}$$

where the parameters represented are indicated in Fig. 4, R_S and R_L are the source and load impedances, respectively, and ω_0 is 2π times the frequency to which the parallel T is tuned.

The value of R (Fig. 4) must be compatible with the source and load impedances used with the parallel T. If it is not, the percentage feedback β, at frequencies remote from the center frequency might be too small, and attenuation to the undesired frequencies would therefore be reduced. The value of R is frequently chosen with symmetry considerations in mind,[10] but we are not concerned with that effect, as indicated above. Our principal concern is that β be reasonably high at frequencies far from the network's frequency. Referring to the equivalent circuits of Fig. 5, we see that the values of β can be computed if the network, source, and load parameters are known. We chose an R of 3 kΩ and a k of 1 (see Fig. 4). 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 11 kΩ for R1, 700Ω for the source impedance (the measured output impedance of a T-108 module), and 30 kΩ for the module's input impedance (manufacturer's data), 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. 4) so as to obtain a larger β at low frequencies. A simple calculation, however, reveals that R must be reduced considerably in order that β 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 β at low frequencies reduces β at high frequencies, and vice versa. This explains why R is usually chosen to be the geometric mean of the source and load impedances.

Referring again to Fig. 4, our parameters at 560 Hz are k = 1, b = 1/2, R = 3kΩ, and C = 0.1μF. In general, the value of bR must be adjusted to obtain the required frequency characteristics. (The absolute frequency is generally not critical, so it is easier to adjust bR for the desired selectivity even though the tuned frequency is changed somewhat.) For example, we found that bR should be 1.3 kΩ shunted by 15 kΩ. When one selects the value of bR, the following key points should prove helpful:

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

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 regenera-
      tive, and therefore susceptible to oscillation.

c.   The value of R for the parallel T (Fig. 4) should be
        approximately 3kΩ. The capacitor values required
        can be computed from Eqs. (1) and (2); if 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 ≈ 25), or
        if the gain of A1 is reduced too much when the parallel
        T is inserted, change the value of bR (Fig. 4) 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 2V peak-to-peak. Above this value, linearity
        falls off.

2. Synchronous Detector

The detector is a synchronous type[11] 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 relation-

ship.

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. [The blue test point (Fig. 2) should be monitored with an oscilloscope. If the monitored signal is 2V or less, peak-to-peak, detector operation is linear.]

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, the gating amplitude remains constant, and its 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.[12] For a symmetrical waveform, 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.

One may 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 their 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 (several hundred volts). Transformers isolate the phase detector from ground.[13] For most applications one can dispense with transformers, substituting dc coupling and the appropriate phase-splitter instead.[2] These changes would extend the frequency limits considerably.

The detector also is subject to frequency considerations. For example, its 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 these capacitances.

Since the synchronous detector is a form of rectifier, its per-
formance depends upon two different time constants, charge and dis-
charge. 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), the forward resistance of the diodes,
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 the diodes are negligible by comparison. In
practice, the charging resistance is approximately 10kΩ.

The charging time constant with C3 selected for 560 Hz is
approximately 100 µsec. Compare this with 450 µsec, the time required
for a sinusoid to reach peak value at the frequency indicated.

The discharge time constant, determined by C3 and the
resistance in shunt with it, is approximately 1700 µsec. This time
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. We chose 0.01 µF
at 560 Hz; for comparable performance at another frequency, one should
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 dis-
connected. 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.

3. Oscillator

The oscillator is a three-section, phase-lead type. Oscillation

frequency is given by[14]

$$\omega_{osc} = \{RC[3 + 2/a + 1/a^2 + (R_S/R)(2 + 2/a)]^{1/2}\}^{-1} \qquad (3)$$

where the parameters represented are shown in Fig. 6. The minimum

gain necessary to sustain oscillation is given by[14]

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

Referring again to Fig. 6, we chose an R of 3.6 kΩ and an

a of 1. 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 (30 kΩ) 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 has been omitted.[15]

Referring to Fig. 6, our parameters at 560 Hz are a = 1,
R = 3.6 kΩ, C = 0.033 μF. (a^{2}R is the frequency adjustment shown in
Fig. 2.) The parameters for other frequencies are more easily
determined if the following points are kept in mind:

a. The value of R (Fig. 6) should be approximately 3.6 kΩ.
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. To achieve this, apply an attenuated output from REF
OUT to SIG IN, then adjust the FREQ trimpot for maxi-
mum output at the blue test point (the attenuator should
present more than 30 kΩ to REF OUT). If the FREQ
trimpot has insufficient control, modify the oscillator
network accordingly.

c. The OSC trimpot is adjusted for approximately 6 V peak-
to-peak at REF OUT. The waveform should be sinusoidal
and free from distortion.

4. Phase Shifter

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

Output amplitude remains constant with phase for the following
reason. The networks (Fig. 2) are driven by two voltages of equal
amplitude but opposite phase. In the equivalent circuit of Fig. 7 and
the vector diagrams of Fig. 8, 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. 8. When phase

is adjusted, E_C and E_R change according to the ratio of reactance to resistance. Figure 8 indicates the distribution of voltage for two different settings of R. Because output is taken between points P1 and P2, the output vector is drawn from the center of the circle (Fig. 8). When R changes (Fig. 7), the output vector rotates relative to the applied voltage, but it does not change in amplitude.

If the reactance of C (Fig. 7) 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 kΩ.

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 kΩ. Since reactance is a function of frequency, the value of C must be adapted to the operating frequency chosen—we use 6800 pF at 560 Hz. For a phase shift of at least 90 deg, the

value of R must be approximately 50 kΩ. If R is 50 kΩ, 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. 9.) Loading is a problem if the

phase shifter is followed by bipolar transistors. One solution to

loading is bootstrapping;[16] our solution is field-effect transistors for

Q2 and Q4 (Fig. 2). These transistors have a very high input impedance,

literally hundreds of megohms. As used here, their input is shunted

by 10 MΩ, but that is relatively high when compared with the value of

R used (50 kΩ). Field-effect transistors are also useful because they

permit the use of smaller coupling 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 linearity suffers, and (b) output voltage is reduced

too much for many applications. Also important, the time constant

responsible for system bandwidth (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 field-

effect transistors), we can use large resistors for R2 and R3 (Fig. 2),

thereby reducing the capacitance required for a given time constant.

6. Preamplifier

A small preamplifier has been designed for those applications

requiring microvolt sensitivity. This preamplifier, which uses two

linear-amplifier modules, is constructed separately; its schematic
is given in Fig. 10. Power is obtained from (a) internal batteries,
or (b) a separate power supply, also shown in Fig. 10. Maximum gain
is about 8000, with the actual value depending upon such factors as
load impedance.

E. Specifications

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

SIGNAL CHANNEL Q: approximately 28 at 560 Hz.

GAIN: (dc out/rms in) $\approx$ 3000. Additional gain (more than 70 dB)
   can be obtained with the preamplifier described.

EQUIVALENT INPUT NOISE: approximately 1 μV rms with the input
   terminals shorted and a 2-sec time constant. Measured at
   560 Hz.

LINEARITY: better than ±1% of full scale.

ZERO DRIFT: less than ±1% of full scale per hour, maximum.

OUTPUT:

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

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

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

F. Applications

The af unit was designed for an ESR spectrometer now being
developed.   Two lock-in amplifiers are used, one for automatic fre-
quency control (12 kHz), another for the signal channel (560 Hz).

We have adapted the af unit to nuclear-magnetic-resonance
work, using the lock-in amplifier with a marginal oscillator to detect
the nuclear-magnetic-resonant frequency of a sample exposed to a
magnetic field.   This frequency, directly proportional to field
strength, can be measured with precision and affords an accurate
measurement of field intensity.   Another 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 spectrum.   This is probably because there is
little demand and frequency flexibility is hard to come by.   As require-
ments 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
0.1μ accurate to several percent.   When the unit was designed, com-
mercial micromanometers of sufficient sensitivity were not available,
and so we designed our own.   For reasons described elsewhere,[17,18]
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 lock-in detector described

(Fig. 11). 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. 17.

Some readers may wonder why the rf unit also was not designed

around circuit modules. The main reason is that the rf unit was

developed before the af unit, so we had not yet thought of using cir-

cuit 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-band-

width product is wasted; (c) high-frequency modules were more expen-

sive; if modules were used, the rf unit would have cost about three

times as much as it did with Nuvistors; and (d) Nuvistors operate at

higher voltages, so the output levels are higher and the need for a dc

amplifier is eliminated in some cases.

The rf unit uses Nuvistor tubes because: (a) it was to be dupli-

cated 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, which permitted us to confine them to well-shielded

compartments, thus minimizing instability problems; and (c) their

combination of high transconductance with low interelectrode

capacitance permits considerable gain without neutralization of the

amplifier stages.

B. Circuit Description

1. Signal Amplifier

The signal amplifier comprises three stages, each tuned to

2.7 MHz. (See Fig. 11.) Because all three stages employ Nuvistors,

more gain can be realized than is possible with most conventional tubes.[19]

The tuned circuits are shunted with 5.1 kΩ resistors that reduce regen-

eration and increase bandwidth, thus minimizing the effects of frequency

drift.

Some tuned circuits are inductively coupled to form transformers.

In Fig. 11, 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.

Assuming that we wish to obtain maximum gain from T1, we

write the equation as[20]

$$G_{max} = E_C/E = 1/2(Q_P/Q_S)^{1/2} (L_S/L_P)^{1/2} \qquad (5)$$

(with critical coupling assumed), where the parameters represented are

indicated in Fig. 12; here Q_P and Q_S represent the primary and secon-

dary Q's, respectively.

If we assume further that L_S, L_P, and Q_S are constant, Eq. (5)

reduces to

$$G_{max} = K Q_P^{1/2}, \qquad\qquad (6)$$

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:[21]

$$Q_P = \omega_0 L1/R_{eq} \qquad \text{(series resonance)} \qquad (7)$$

$$Q_P = R_{eq}/\omega_0 L1 \qquad \text{(parallel resonance)} \qquad (8)$$

where $\omega_0 L1$ is the inductive reactance of L1, and R_{eq} is the equivalent resistance lowering primary Q.

If we assume that losses due to the source resistance are much greater than the equivalent coil losses (6Ω), we can substitute the source resistance for R_{eq} in Eqs. (7) and (8). Each equation yields a different Q_P;[22] 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 700Ω. We would use series resonance if the source resistance were less than 700Ω, parallel resonance if it were more. Physically, this can be explained as follows: Secondary voltage is directly proportional to primary current, I_P.[23] 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 occurs when the source resistance equals the inductive

reactance of L_p. Substitution of the design parameters into Eq. (5)

reveals that the maximum gain obtainable from T1 is approximately 16.

In most cases, the value realized is considerably less, the amount

depending on source resistance.

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. 11.

3. Reference Amplifier

The reference amplifier involves only one stage. Amplifier

output is coupled to the phase detector through 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 that .

discussed in Sec. III. B. 4.

4. <u>Generator</u>

The generator (Fig. 13) 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 and C2. The ratio and bias are adjusted to produce a minimum of distortion. This capacitance 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.

E. <u>Specifications</u>

Performance, as related to the original application of the rf unit, is described in Ref. 18. The important specifications for the <u>modified</u> rf unit are listed below:

FREQUENCY: 2.7 MHz.

SIGNAL CHANNEL BANDWIDTH: 220 kHz.

GAIN: (dc out/rms in) $\approx$ 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. 18.)

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. 17. In achieving the design objective, 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; we are considering its application to diagnostic studies of
gaseous media. Another application involved the detection of light
modulated at 2.7 MHz.

ACKNOWLEDGMENTS

FOOTNOTES AND REFERENCES

*Work done under the auspices of the U.S. Atomic Energy Commission.

1. Kenneth W. Lamers, Lawrence Radiation Laboratory Report
   UCRL-16632, January 1966.

2. R. D. Moore, Lock-in Amplifiers for Signals Buried in Noise,
   Electronics 35, No. 23, 40 (1962).

3. Lock-in amplifiers are commonly driven by other devices that
   generate 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. R. D. Moore and O. C. Chaykowsky; Modern Signal Processing
   Technique for Optimal Signal to Noise Ratios, Princeton Applied
   Research Technical Bulletin 109, 1963.

7. G. Feher, Sensitivity Considerations in Microwave Paramagnetic
   Resonance Absorption Techniques, Bell System Tech. J. 36, 480
   (1957).

8. This can be explained as follows: $G = A/(1 + A\beta)$. With a perfect
   network, β is zero at the tuned frequency, so G is A. At frequen-
   cies far removed from ω_0, β approaches 1 and G approaches
   $A/(1 + A)$. Maximum gain change is approximately A.

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

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

11. Application Notes on Synchronous Detection Systems and Associated
    Instrumentation, Triconix, Inc., Waltham, Mass, n.d.

12. A transformer does not couple a dc component, nor does capa-
    citive coupling.

13. 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 the DET OUT terminals.

14. See. Ref. 9, pp. 6-40.

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

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

17. Kenneth W. Lamers, The Design, Construction, and Operation of a
    Differential Micromanometer, Part I. Electronics, Lawrence
    Radiation Laboratory Report UCRL-11218, October 1964.

18. Peter R. Rony, Design, Construction, and Operation of a Differ-
    ential Micromanometer, Part II. Theory and Operational Char-
    acteristics, Lawrence Radiation Laboratory Report UCRL-11218,
    April 1965.

19. See. Ref. 9, pp. 7-53.

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

21. Ibid., pp. 32 and 357.

22. This is true except when the source resistance is equal to the reactance of L1.

23.   See. Ref. 20, p. 54.

FIGURE LEGENDS

Fig. 1.   Block diagram of the af unit. The modulator, which is

external to the af unit can take many forms, electrical or mechani-

cal.

Fig. 2.   Schematic diagram of the af unit. Modules T-108 and T-116

are manufactured by Engineered Electronics Company, Santa Ana,

California. Module grounds should be made as indicated to avoid

ground loops. All resistors are 1/4W, 5% carbon, unless

indicated otherwise. Phase control is a 2-gang, 2W, Ohmite

CCU-5031. All capacitors not polarized are Mylar. BNC connectors

are insulated from ground, DAGE 4890-1. Components indicated by

* are frequency sensitive; their values can be deduced from the

text. Components indicated by ** may require special attention.

We found that when operating at 12 kHz, the capacity unbalance of

T1 caused detector unbalance and zero shift with phase. This was

corrected by substituting UTC type A-22 for T1.

Fig. 3.   Frequency characteristics of A1 (Fig. 2) with parallel T tuned

to 560 Hz. Asymmetry is explained in text.

Fig. 4.   Parallel-T network. The source and load are connected as

shown by dashed lines.

Fig. 5.   Equivalent circuits of the parallel T at (a) dc, and (b) high

frequency. R is the network resistance, as indicated in Fig. 4.

R_S and R_L are the source and load resistances, respectively.

Fig. 6.   Equivalent circuit of three-section phase-lead oscillator.

Fig. 7.   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. 8. Vector diagram 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.

Fig. 9. Equivalent circuit of phase-splitter illustrating why output amplitude changes with phase unless R_L is much greater than R.

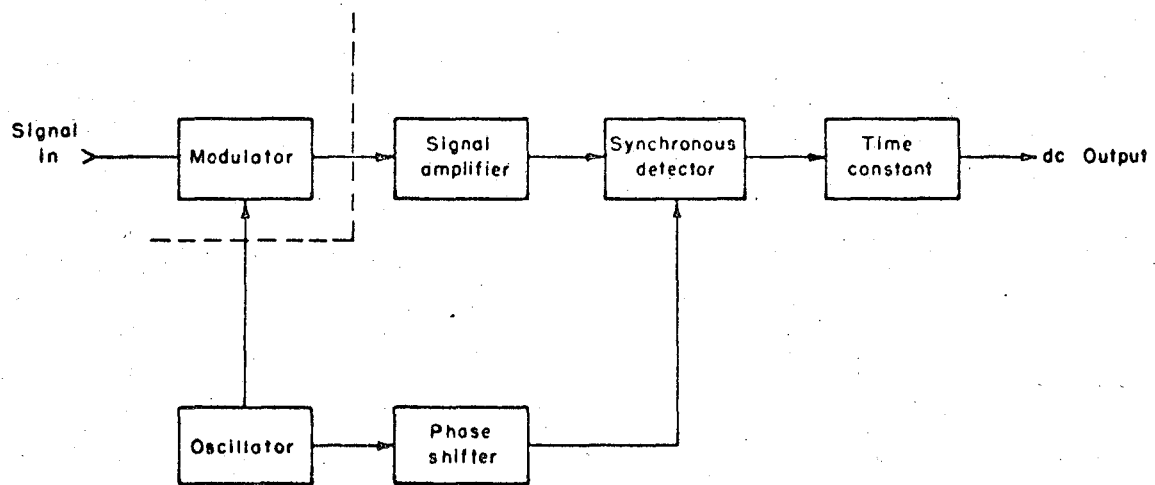Fig. 10. (a) Schematic diagram of the af preamplifier. BNC connectors are DAGE 4890-1. First-stage bypass (pins 8 to 9) is subject to frequency. (b) Power supply for preamplifier.

Fig. 11. Schematic diagram of the rf lock-in detector unit. The 27 kΩ resistors are 1/2W; all others are 1/4W. Asterisks denote 1000-pF ceramic feed-through capacitors, all other pF capacitors are silver-mica, and all μF capacitors are Mylar. C is a 7- to 45-pF ceramic trimmer. 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 in which the components are mounted. A low-impedance potentiometric recorder can be driven by a divider across the DET OUT terminals, as shown. The power supply is not shown.

Fig. 12. Equivalent circuit useful in proving that transformer T1 (Fig. 11) should be operated series-resonant in some cases and parallel-resonant in others, with the choice depending on source resistance.

Fig. 13. Schematic diagram of the rf lock-in generator unit. All resistors are 1/4W. 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. The crystal (type CR-

Fig. 13. (con't)

18/U) operates at 2762.500 kHz. Capacitors marked with an

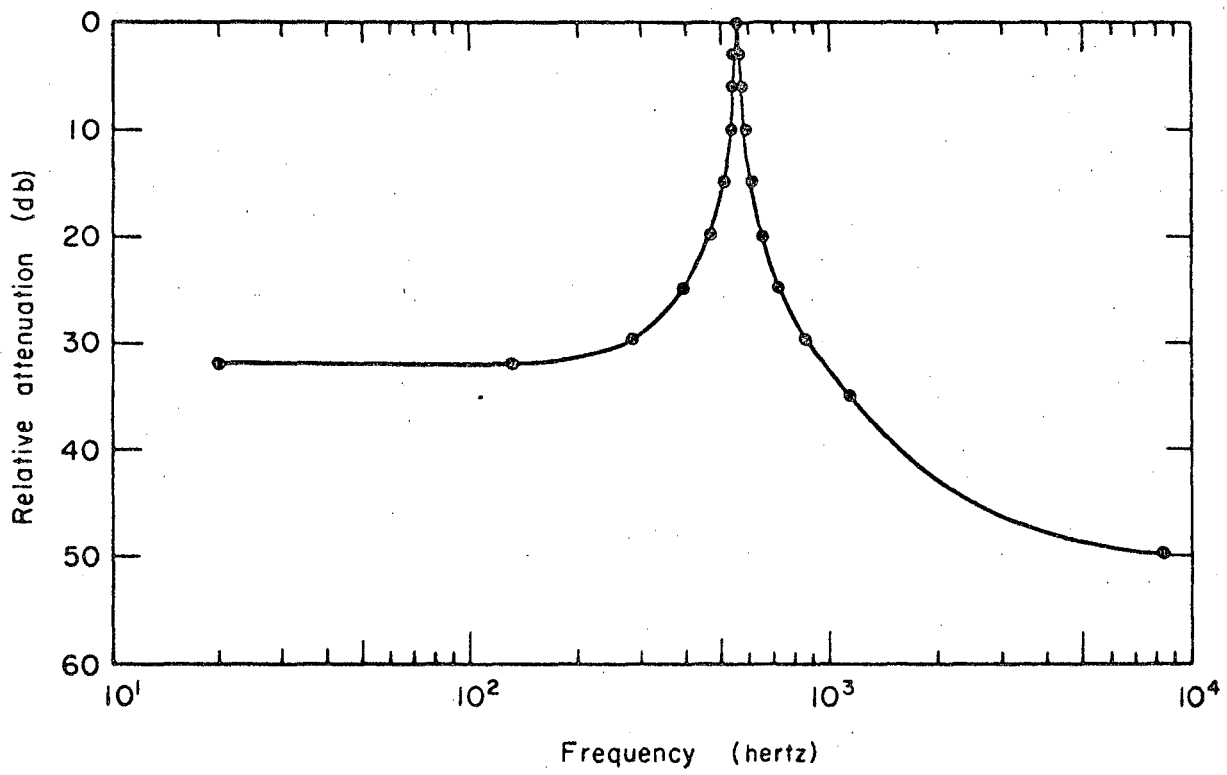asterisk are 1000-pF, ceramic feed-through type; those designated
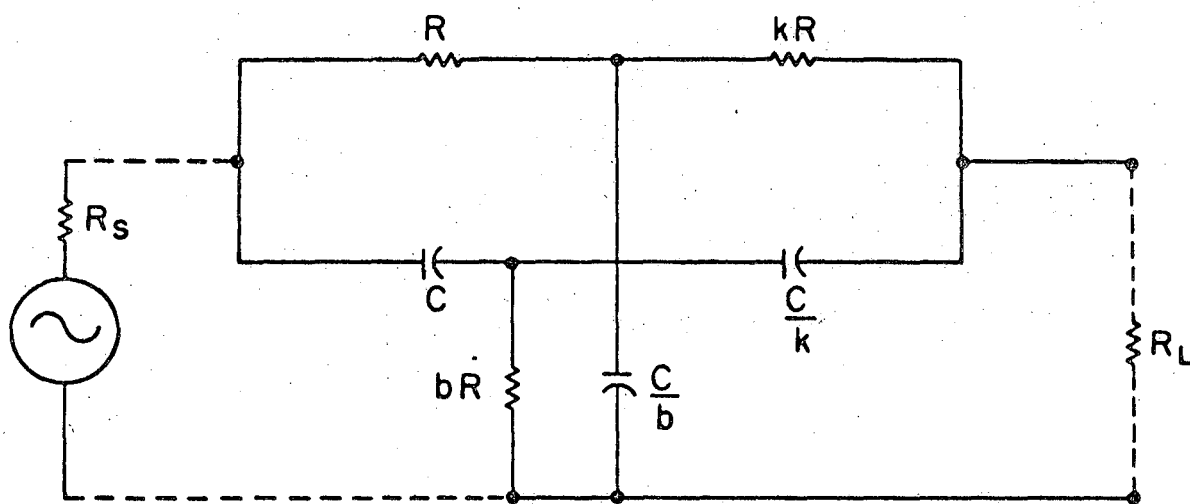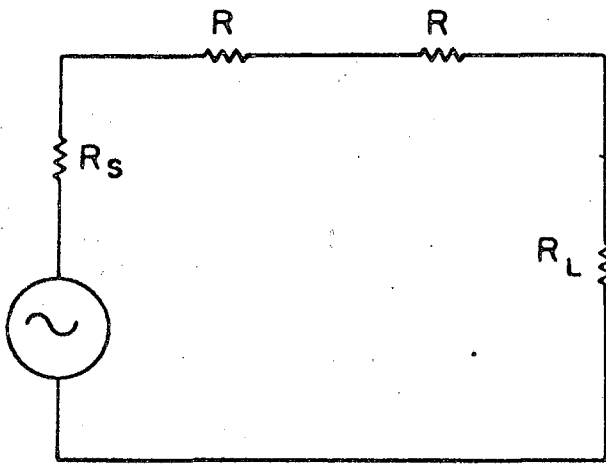
SM are silver-mica.

MUB-9587

Oscillator
7895

Buffer
7586

V1

2

68pF
SM

C₁

30 to 56 μH
L₁

4

Crystal

8

10k

0.001
μF
Mylar

C₂

10 pF
SM

100k

0.0068
μF
Mylar

100k

0.01
μF
Mylar

5 to 9 μH

V2

2

L₂

100pF SM

"Out"

100pF
SM

4

8

240
pF

0.001
μF
Mylar

BNC

0.001
μF
Mylar

100k

91 Ω

0.01
μF
Mylar

1 k

75V

V1   V2

10   10

12   12

6.3 V ac
(0.27A)
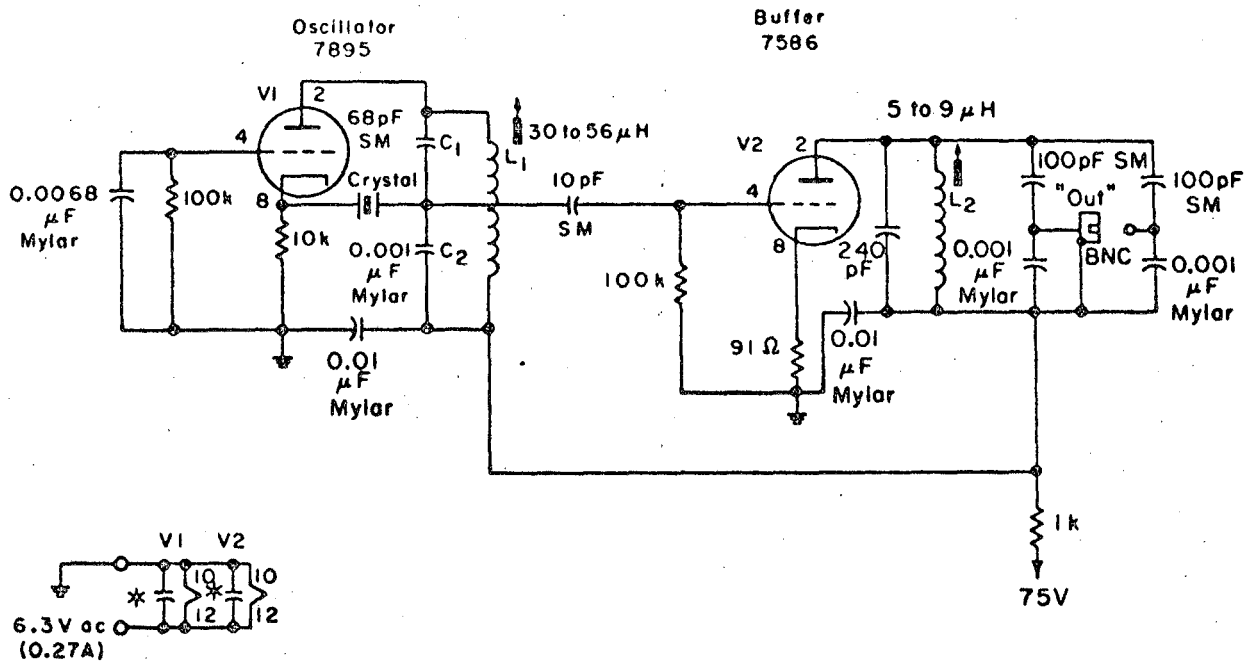
MUB-4412

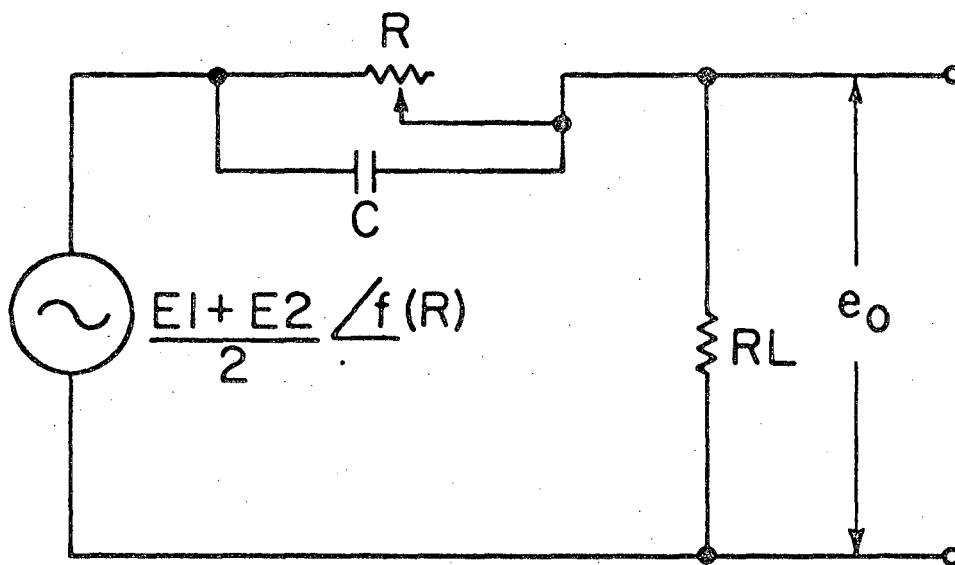MUB-9589

MUB-9590
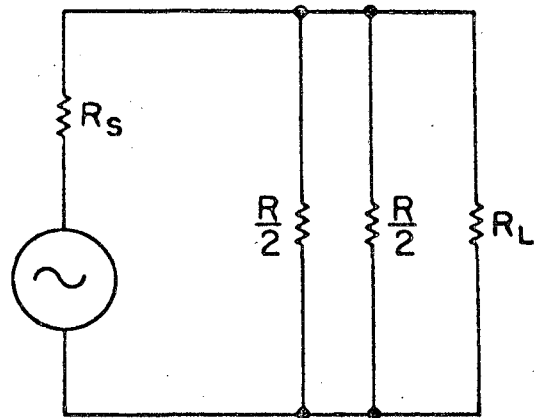
(a) dc   (b) High frequency

MUB-9591
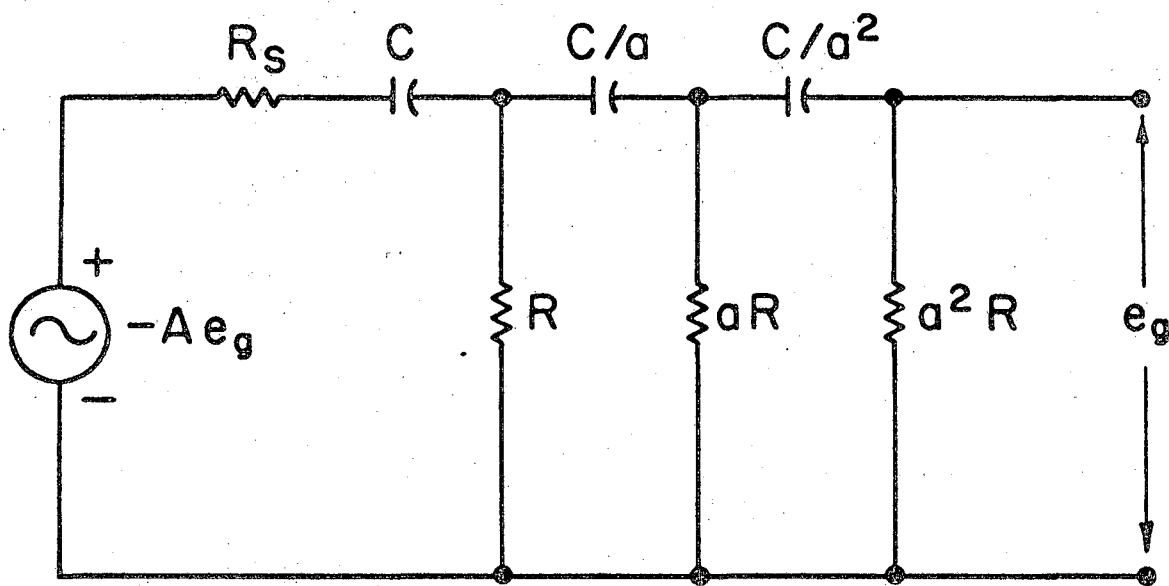
MUB-9592

MUB·9593

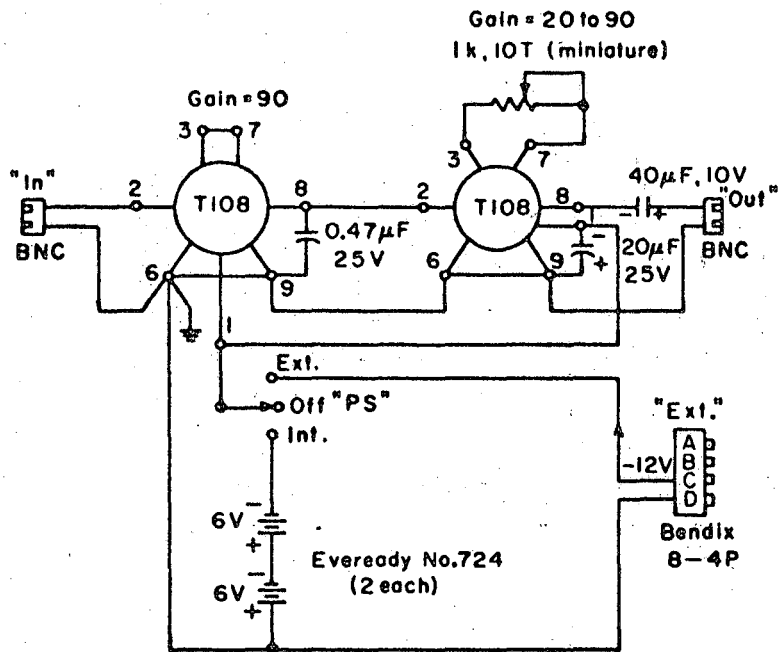(a)

(b)

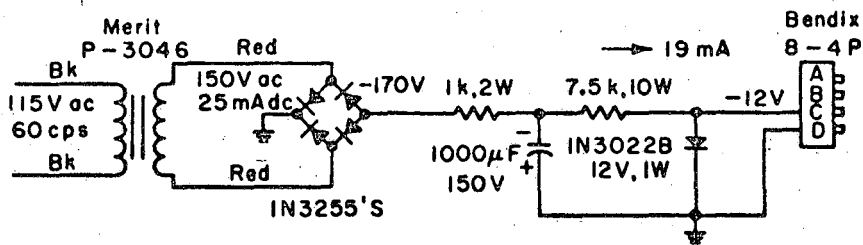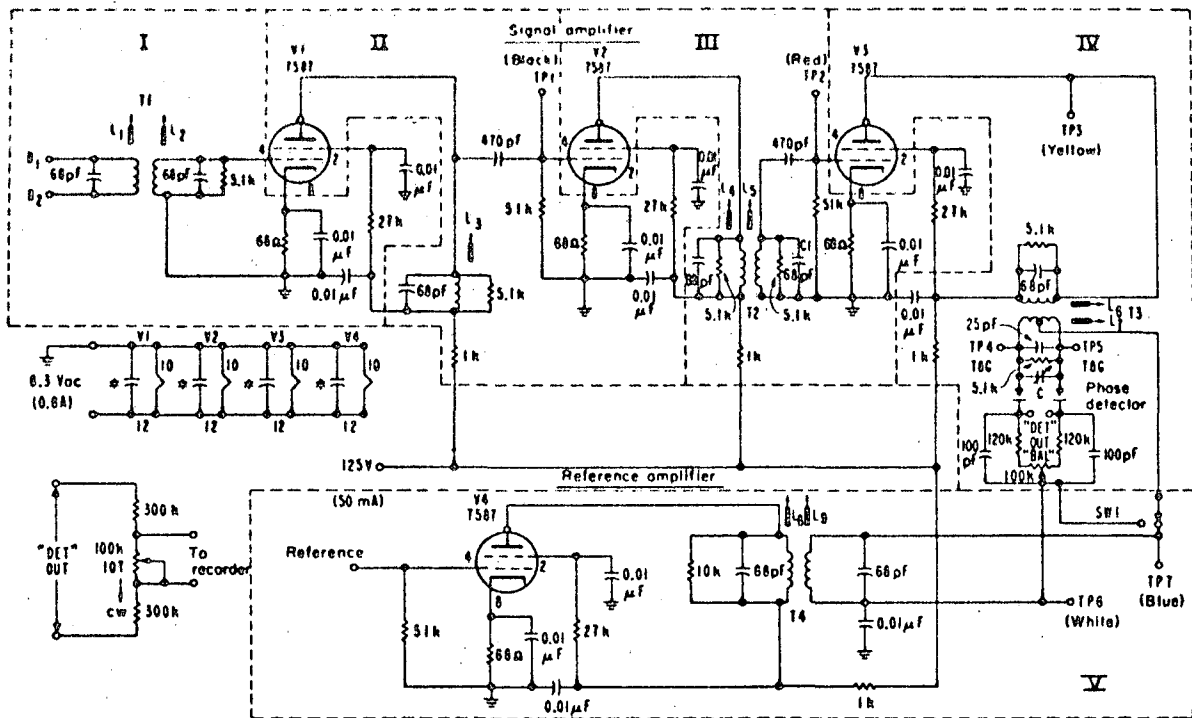MUB-9594

$$\frac{E1 + E2}{2} \angle f(R)$$

R

C

RL

e_0

MUB-9595

Gain = 20 to 90
1k, 10T (miniature)

Gain = 90

3 7

"In" 2   8   2   T108   8 $40\mu F, 10V$ "Out"
         T108   BNC
BNC
      6   9   6   9 $20\mu F$
                                       25V
              $0.47\mu F$
              25V

Ext.

Off "PS"   "Ext."
Int.   A
                                      -12V   B
                                             C
6V   D
                                          Bendix
6V   Eveready No.724   8-4P
           (2 each)

(a)

Merit   Bendix
P-3046   Red   → 19 mA   8-4P
Bk   A
     150V ac -170V 1k,2W   7.5k,10W -12V B
115V ac   C
60 cps 25mAdc   D
Bk $1000\mu F$ 1N3022B
     Red   150V   12V,1W
     1N3255'S

(b)

MUB-9566

MUB·9597