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AUTOMATIC REDUCTION OF WIND TUNNEL DATA

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for the period

November 17, 1951, to February 17, 1952

Project
M938

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PROGRESS REPORT NO. 4

AUTOMATIC REDUCTION OF WIND TUNNEL DATA

INTRODUCTION

This progress report will be chiefly concerned with the work of this project during the period November 17, 1951, to February 17, 1952. Details of the individual components of the information channel and the description of the d-c operational amplifier were presented in Progress Report No. 3; therefore, this report contains only a general description of these items in addition to the work performed during this particular reporting period.

In general, the work performed during this reporting period has resulted in a final design of the information channel and completion of the 2-kc carrier supply and 400-cps power source.

Several design changes have been made in the information channel. The Class A a-c signal amplifier was redesigned so that more common tubes could be used (Figure 6). To permit the utilization of a greater range of phase angles from the phase-shifting network, a cathode follower was placed at the output of the phase-shifting network (Figure 8). In order to permit automatic compensation of greater initial unbalances in the sensing unit, the zero-error balancing network has also been redesigned (Figure 8). These changes have not altered the over-all operating characteristics of the information channel but have merely increased its utility.

Progress and present status of each unit in the information channel will be reported as sections of the block diagram, Figure 1, and subsidiary units, such as the 2-kc carrier, 400-cps power, d-c operational amplifier, B^+ supplies, and operational procedures for the use of this equipment.

BACKGROUND

This system of data reduction will use analog method of computation for calculation of desired measurements in wind tunnel work. The system will take the output of sensing units, operate on them, correct for tares, position, etc., and produce either the desired measurement or coefficient to be applied to some type of d-c voltage sensing recorder. It is thought that it might be valuable to apply a digital conversion to this analog output, but this is not part of this contract.

In studying a sample computation, let us assume that we have two strain gage elements on a cantilever sting of a balance system. The output of these two units will be proportional to moments while the desired measurements are force and center of pressure. Each of these strain gage elements will become the end instruments for a separate channel of information. The gages will be incorporated into bridges, operated with a 2-kc carrier, and the output will be an amplitude-modulated signal with a 180-degree phase reversal for negative or positive forces. Figure 3 is a block diagram of this system.

This signal will be amplified through the class A amplifier (reference Figure 1) of which the gain control will represent a composite of constant multiplying terms determined by calibration for each new test setup. This composite multiplying term can be determined by applying a known force and varying the gain of each channel until its output directly represents the applied force or moment.

The output will be summed with a carrier so that the total output will be amplitude modulated without phase reversal. This summed signal then will be rectified directly. At the same time, a carrier equal to that summed in will be rectified through a separate channel and subtracted from the above rectified signal. This yields a d-c voltage proportional to the actuating quantity times a lumped constant of the system.

This d-c signal will be applied through weighting factors to a d-c operational amplifier used for summing the two hypothetical parallel channels. The weighting factors will be determined by the position from a reference point on the sting, tares, tunnel corrections, and the decision of whether center of pressure or force is desired. The output of the d-c operational amplifier will be the desired quantity. It is possible to perform both calculations simultaneously by utilizing two d-c amplifiers, as shown schematically in Figure 3.

Balance in normal strain-gage bridges and other sensing units will be accomplished in two steps. The first will be the initial correction which will be applied directly to the strain-gage bridge to bring the zero point within the limits of the automatic balancing system and will be manual in the initial test setup. The second step will take any normal unbalances due to temperature and other causes and correct for them by changes in the bias voltage. This will be accomplished, only when the channel is not computing, by sensing the d-c output of the rectifiers, and chopping and amplifying the error signal to a servo-motor that will vary the bias to bring the error signal to zero.

The d-c operational amplifiers will have their own automatic a-c balancing loop.

INFORMATION CHANNEL

General

A prototype of the information channel has been completed and, except for very minor changes in the physical layout of the individual components, all information channels will be reproductions of it. Photographs and a diagram of an information channel are shown in Figure 4.

Each channel will be a plug-in unit which is completely interchangeable with all other channels. All power is brought in through the plug at the front, while the signal input and output plugs are located at the rear of the chassis. The adapter box, shown at the rear of the chassis, is used when forces are to be measured. It contains two legs of a Wheatstone bridge- the other two legs being strain gages. One of the legs contained in the box is a variable resistance which provides a coarse, manual adjustment for balancing the bridge initially. When a pressure-sensing unit is employed, the adapter box is not used. It should be noted that the input to the adapter box is a four-prong socket, while the input to the chassis is a six-prong socket. By using a four-prong plug for the strain gages and a six-prong plug for the pressure-sensing element, it was believed that the possibility of using by mistake only two resistances instead of a complete Wheatstone bridge would be eliminated. The information channel accepts signals from a strain-gage Wheatstone bridge or a Schaevitz Differential Transformer (pressure-sensing element) with equal facility. In general, any sensing unit which will amplitude modulate the 2-kc carrier can be used.

It might be noted here that the type of strain-gage bridge mentioned above - one with two fixed arms - is not the only type that can be used with this system. If the wind tunnel test configuration is of such a

nature that four active gages can be used, certain advantages are obtained. Among these are increased bridge sensitivity - a factor of four if four active gages are used instead of one - and the possibility of measuring differential strains, i.e., comparison of the strains at two points. Due to the increased sensitivity, the relative magnitude of the temperature effects is reduced, since not only will any over-all variation in bridge characteristics because of temperature variations be smaller, relative to the total signal, but also the positioning of four gages in the same general location exposes all gages to the same thermal conditions. In addition to the above, the symmetrical bridge arrangement permits the wiring associated with the bridge to be symmetrical also, thereby minimizing the resistive effects of leads and connections. Basically, it is better engineering practice to make use of four active gages whenever possible, due to the fact that all adverse effects are minimized when this configuration is used.

As can be seen from Figure 4, Vector turrets have been used wherever possible to make identical components of each information channel completely interchangeable. The use of these plug-in units greatly decreases the loss of running time due to breakdown of an information channel. If a channel should become faulty, it can be replaced immediately by a spare channel, thereby making the complete system operative again. The plug-in units of the faulty channel can then be tested individually and the inoperative unit repaired at the leisure of the operator or technician concerned. These cans will be color-coded along with the octal sockets in the chassis to assist in replacing the units in the proper socket.

In addition to the above considerations, the cans were used to reduce pickup within the information channel not only by restraining the entrance of stray voltages into a particular unit but also by reducing the amount of radiation from the units themselves. In short, the cans have greatly reduced the possibility of crosstalk among individual units and in doing so have aided in the suppression of positive feedback, which can be bothersome in relatively high-gain amplifiers.

Although no extensive testing has been made on the prototype, preliminary tests indicate that the servo-balancing system is slightly underdamped. This was believed to be desirable; therefore, no additional electronic or mechanical damping will be placed in the servo-loop. Back-lash in the gear train between the servo-motor and balancing potentiometer is approximately ± 0.05 per cent of the total travel of the potentiometer. This indicates that mechanically the servo-loop will balance to within approximately ± 6 millivolts at the channel output. Tests have not as yet been made to determine the sensitivity, linearity, and balancing accuracy. They will be made as soon as the power supplies are completed so that the characteristics of the system as an entity - information channel and its power supply - can be determined.

The signal amplifier gain and phase-shift controls are located adjacent to the power plug at the front end of the information channel chassis. These controls are screwdriver adjustments which are accessible through holes in the rack panel. This configuration, it is believed, will prevent the setting of either control from being accidentally changed at any time. A change in the position of one of these controls, especially the gain control, will necessitate a complete recalibration of that channel since it will no longer be zero-balanced. It should be noted that a change in the gain of any unit in the information channel, except the servomotor amplifier, will introduce a shift in the zero-balance position. Therefore, the system must be rebalanced before accurate results can be obtained.

Dial lights will be placed in the rack panel immediately ahead of each information channel. One of these lights will indicate when the servomotor is operating and another will show if the zero-balancing potentiometer has been moved to either limit of its travel. When the latter light is on it will indicate that a manual readjustment of the sensing element is necessary to bring the system within the range of the automatic balancing system.

The present plan is to mount ten information channels in one relay rack. The physical layout of the rack and associated information channels has begun, and construction should be completed in the very near future.

60-cps Parallel "T" Filter

This filter is located at the input of the information channel. Its purpose is to attenuate greatly any stray 60-cps pickup voltages which may originate in the sensing unit and particularly in the leads between the sensing unit and the information channel. At 60 cps, it has an attenuation of approximately 60 db, while at frequencies greater than 1 kc it has a constant attenuation of approximately 2.5 db. The circuit diagram is shown in Figure 5.

Class A A-C Signal Amplifier

As described in Progress Report No. 2, the estimated maximum gain required from the a-c signal amplifier is approximately 75 db. At the time of publication of that report a satisfactory amplifier had been designed, but subsequently it was decided to redesign it so that vacuum tubes which are cheaper and more easily obtainable could be used. This redesign led to the amplifier shown schematically in Figure 6. Again the gain control was placed in the negative feedback loop.

The open-loop frequency response of this redesigned amplifier, without the input transformer, is shown in Figure 7. The maximum open-loop gain is approximately 80 db at 1000 cps. Including the input transformer, the over-all open-loop gain is approximately 100 db. This is approximately 25 db more gain than was thought to be necessary, but with this additional gain the amplifier can be operated with some negative feedback at all times. Using the amplifier in this manner, the frequency response will be flat over a greater range of frequencies than that required by the expected band width.

The components used in the feedback loop are such that the maximum negative feedback obtainable in the system is 48 db. However, the stability of the prototype was checked using 65 db of negative feedback; no instability occurred. With this margin of safety, it is reasonable to expect this system to remain stable for normal changes of tube characteristics and circuit variations.

As in the previous design, the gain control is in the cathode circuit of the first stage, Figure 6. Minimum over-all gain is obtained when the negative feedback is fed directly into the cathode, and maximum gain when the feedback-loop output is grounded at the cathode. The cathode-follower output stage is again employed as an impedance-matching stage to prevent the amplifier from being overloaded by the network into which it operates.

In the expected operating frequency range, the maximum undistorted output is approximately 80 volts peak, which is well beyond the maximum desired input voltage to the operational amplifier, namely 50 volts.

It must be kept in mind that the gain of the a-c signal amplifier must remain fixed after the system has been balanced, since this amplifier is not located in the closed-loop balancing network, and any change in gain will appear as a change in error signal, appearing at the output of the information channel. It is expected that this will introduce no hardship, since the gain of the amplifier will be adjusted during calibration and will then be allowed to remain at a fixed setting over a relatively long period of time, whereas the balance circuit will be activated before and after each run.

Bias Summing

The output of the sensing units will have a phase reversal of 180 degrees when passing through the zero-actuating balance point. However, the diode-bridge rectifier is not a phase-sensitive device; its d-c output will have the same polarity regardless of the phase of the input a-c signal.

It is therefore necessary to add a 2-kc carrier as bias at the output of the a-c signal amplifier. This bias carrier will be in phase with the signal carrier for positive signals and of sufficient amplitude to keep the input to the rectifier in phase with the bias carrier regardless of the phase of the a-c signal amplifier output, thus yielding a biased signal that does not require a phase-sensitive rectifier.

The d-c output of the diode-bridge rectifier will be linearly proportional to the magnitude of this biased signal. The d-c equivalent of the 2-kc bias carrier is obtained by applying the 2-kc bias carrier to a separate diode bridge. This d-c output is then subtracted from the d-c equivalent of the biased signal, giving a d-c voltage linearly proportional to the input signal.

The bias-summing circuit proposed in Progress Report No. 3 has been modified. This modification occurred as the result of a redesign of the zero-error balancing loop to increase the range of the latter. This change has reduced the cost per channel by eliminating one transformer from the circuit. The new bias-summing circuit is shown in Figure 9.

Zero-Error Balancing

For a no-load condition on the sensing units, any unbalances will appear as a d-c voltage at the output of the diode-bridge rectifier. To balance out this error the value of the d-c equivalent of the bias carrier can be increased or decreased. This is done by having the servo-motor in the automatic zero-error balancing loop drive a ten-turn potentiometer which varies the magnitude of the bias carrier to be subtracted from the biased signal voltage.

Previously, the balancing potentiometer was located in one leg of a bridge circuit, but it was found that acceptable accuracy could be obtained if the potentiometer was used in a simple voltage-dividing network. Therefore, since the range of unbalance that could be compensated for by the dividing network was greatly increased, it was decided to utilize the greater range with its reduced, but still acceptable, accuracy. This new balancing network is shown in Figure 8.

Phase Shifting

The proposed bias-voltage phase control, reported in Progress Report No. 3, was a passive phase-shifting network with an attenuation factor of 2. If such a circuit were operating into an infinite impedance the range of phase shift available would be would be 0 to 180 degrees. However,

the redesigned bias-summing circuit presents an impedance of only 15,000 ohms. This reduces the range of control to little over 90 degrees. Therefore a cathode follower was placed at the output of the passive phase-shifting network to improve the impedance matching. The total phase shift available now is 0 to 170 degrees. A type 12AU7 tube with both triode sections paralleled was chosen for the cathode follower, since this type of tube is already being used in other sections of the information channel. In keeping with the general practice on this project, the phase-shifting network and cathode follower were placed in a Vector turret socket. The schematic layout of this circuit is shown in Figure 8.

400-cps, DC-AC Chopper

The information channel at present is being operated with a Stevens-Arnold No. 268 chopper. However, information on other 400-cps choppers, which have a break-before-make type of contact action, is being gathered. A final choice will be made as soon as sufficient data are obtained from the literature and life tests being conducted by this project.

The life tests are being conducted on two 400-cps choppers, namely, a Stevens-Arnold Model No. 268 and an Airpax Model A493-2. These choppers have been operating for approximately 700 hours with 1.5 volts at approximately 300 microamperes across the contacts. The Airpax was found to have more inherent noise initially, but the Stevens-Arnold chopper appears to have an increasing amount of noise as the length of operating time increases.

Initially it was realized that this test would not conclusively determine which one of the two choppers is better, but assuming that each is a representative unit of the particular manufacturer, a test of this type will in general indicate what can be expected from each of the choppers under similar operating conditions.

Servo-Motor Amplifier

The class AB_1 type servo-motor amplifier explained in detail in Progress Report No. 3 remains unchanged.

Briefly, the servo-motor amplifier consists of three stages of voltage amplification and a phase-inverter stage using type 12AU7 tubes. This is followed by two 6AQ5 tubes in push-pull with an impedance-matching transformer at the output. The schematic circuit diagram of the servo-motor amplifier and its frequency response are shown in Figures 10 and 11, respectively.

Servo-Motor

Each information channel contains one servo-motor amplifier driving a Bendix Eclypse-Pioner CK-2 motor (Signal Corps, No. MO-18A). The CK-2 is a 400-cps two-phase motor rated at 26 volts on the fixed phase and 0-40 volts on the variable phase. The motor is used to drive a 20,000-ohm Micro-pot in the zero-error balancing loop.

For a more detailed description, reference is made to Progress Report No. 3.

SUBSIDIARY UNITS

2-kc Carrier System

Since the accuracy of the whole Wind Tunnel Data Reduction System depends upon the stability and accuracy of the transducer output, it is necessary to have a carrier that has amplitude stability of an order of magnitude greater than the desired accuracy of the system.

The 2-kc carrier amplitude control must provide amplitude regulation of the carrier of 0.1 per cent to obtain the over-all system accuracy of 1 per cent.

This unit, shown schematically in Figure 14, accomplishes this by slicing the sine wave generated by the Wien Bridge oscillator, passing the resultant wave through a low pass filter and amplifying the carrier to the desired amplitude and power level. The output of the power amplifier, i.e., the carrier to be applied to the transducer, is also applied to a rectifier bridge through a step-down transformer and compared to a variable d-c reference. This error signal is amplified in a d-c amplifier, the last stage of which governs the plate supply of the slicer, and therefore regulates the amplitude of the carrier.

The system has been constructed as shown in Figures 15, 16 and 17. The complete system has been tested both open-loop and closed-loop and is in operation for further testing of other component parts of the computer system.

The control available by full variation of a 22-1/2-volt d-c reference is better than 20 per cent of the normal operating value of the output to the transducers.

This system was tested closed-loop by applying step variation of the output of the oscillator into the slicer and noting the variation of output to the transducers. These step variations were twice the normal output and half the normal output of the oscillator; they yielded no measurable variation of the output on meters that would show a 1 per cent change in the output. Variations of the 110-volt a-c power to the power amplifier B^+ supply of better than 10 per cent also yielded no variation in the output. Further and more precise tests will be conducted before completion of the project, but it can be concluded now that this unit has proved to be more than satisfactory.

400-cps Power

The 400-cps power will be used to drive the choppers and fixed phase of the servo-balancing loop. This unit has the same general circuit design as that used for the 2-kc carrier system, except that the amplitude control requirements are not as stringent and a servo-control loop is not used. The oscillator and driver amplifier (Figure 18) are mounted on the same chassis as the 2-kc carrier control.

The power amplifier and the B^+ supply are identical to that used for the 2-kc carrier, Figures 16 and 17.

This unit has been tested and is in use, testing other components of the computer network.

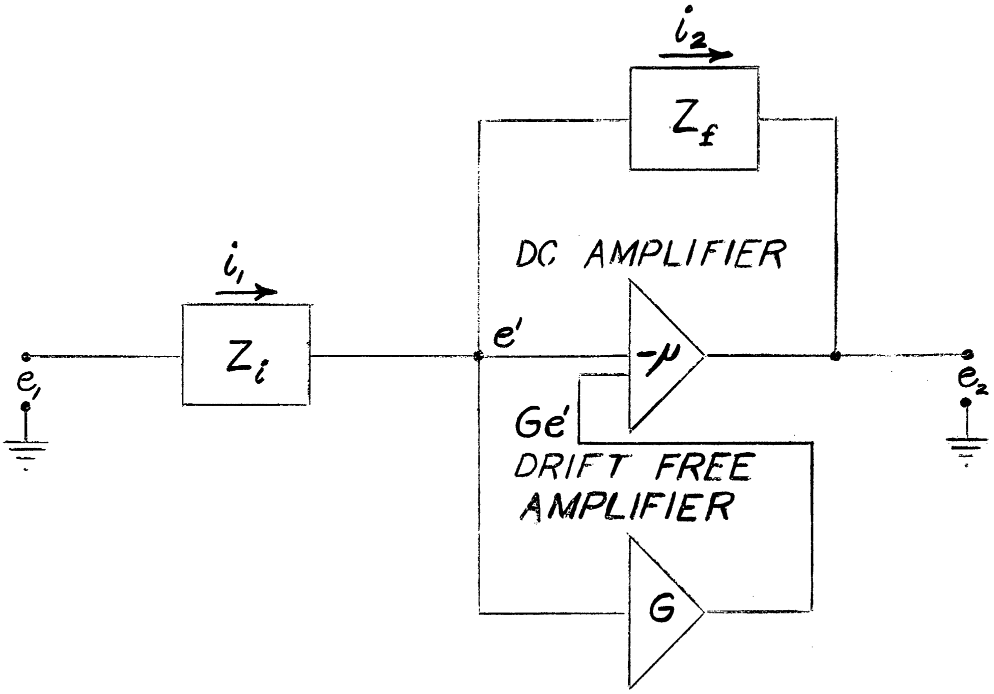
All units of the 2-kc carrier system and the 400-cycle power supply are mounted in one relay rack. The B^+ supply for both the 2-kc and the 400-cps oscillator, control loop, and driver amplifiers will follow the general design utilized in many ordinary regulated power supplies, and will be contained in this rack, which requires only 110-volt a-c power input and 2-kc, 400-cps output leads to it.

DRIFT-STABILIZED D-C OPERATIONAL AMPLIFIER

The design principles and considerations of the d-c operational amplifier were presented in Progress Report No. 3. In the present report the basic d-c operational amplifier will be considered to be available, and therefore emphasis will be placed upon the drift stabilization and its contribution to the characteristics of the drift-stabilized d-c operational amplifier when the latter is considered as an entity.

Theory of Operation

One of the difficulties inherent with any d-c amplifier is the zero-drift problem, and the operational amplifiers used with electronic differential analyzers are no exception. Since the operational amplifiers employ a large amount of feedback, the zero-drift is very much less than it would be with an open-loop d-c amplifier; yet it can still be appreciable. To cite a specific example, it has been found that with an electronically regulated power supply (VR tube reference), an operational amplifier with a gain of unity (input resistor equal to feedback resistor) will drift the order of 0.01 volt per hour after it has been thoroughly warmed up. Naturally this figure is very approximate, but it gives an idea of the amount of drift which can be expected. This drift can be reduced by a large factor by using the additional drift-free a-c amplifier of gain G shown in Figure A. In this drift-stabilizing amplifier the input voltage e' is chopped into a-c by a 60-cycle vibrator. The a-c signal is amplified through capacitor-coupled stages and then reconverted to a d-c output, Ge', by the same 60-cycle vibrator. Since the actual amplification is accomplished by an a-c amplifier, there is no d-c drift introduced. The output, Ge', of the drift-free amplifier is fed into the main d-c amplifier along with e', so that the net input into the amplifier is e' (1 + G). The output voltage, e_2, is



Note: Ground connections on amplifiers omitted for clarity.

Figure A

$-\mu(1 + G)e'$, where $-\mu$ is the gain of the main d-c amplifier. Thus by using the auxiliary drift-free amplifier we have increased the gain e_2/e' of the open-loop system by a factor $1 + G$ without introducing any additional d-c drift.

If the currents into the amplifiers are negligible when compared with the currents i_1 and i_2 through the input and feedback impedances, respectively, then $i_1 = i_2$, and

$$\frac{e_1 - e'}{Z_i} = \frac{e' - e_2}{Z_f} . \qquad (1)$$

Remembering that $e_2 = -\mu(1 + G)e'$, we have

$$e_2 = -\frac{Z_f}{Z_i} \frac{1}{1 + \frac{1}{\mu(1 + G)}\left(1 + \frac{Z_f}{Z_i}\right)} e_1 . \qquad (2)$$

For the actual drift-stabilized amplifier which will be used, $\mu \cong 30{,}000$ and $G = 330$, so that the total d-c gain $\mu(1 + G) \cong 10^7$. Hence it is evident that for Z_f/Z_i the order of 100 or less

$$e_2 = -\frac{Z_f}{Z_i} e_1 , \qquad (3)$$

to a high degree of approximation.

Effect of Drift-Stabilizing Amplifier on D-C Unbalance

In order to assess accurately the effectiveness of the additional a-c stabilizing loop in reducing the d-c unbalance or drift, we must write down the equations for the output e_2 of the operational amplifier for zero input ($e_1 = 0$). In general, this output will not be zero, due to unbalance in the d-c amplifier proper as well as unbalance in the drift-stabilizing loop. The latter results from any a-c pickup at the chopping frequency (60 cycles in our case) coming either from the chopper contacts themselves (contact potentials), the chopper-driving coil, or the B^+ supply voltages for the a-c amplifier tubes. Since d-c is used for the filaments, no pickup should be introduced by the heaters. Any 60 cps a-c pickup gets amplified in the a-c amplifier and converted to a d-c signal before being fed

back into the main operational amplifier. This d-c unbalance will be constant as long as the pickups causing it remain constant.

Let the d-c unbalance of the drift-free amplifier, referred to its input, be c volts. If e' is the input to the amplifier, then its output is Ge' + Gc, where G is the gain of the drift-free amplifier. Also let b volts be the d-c unbalance referred to the input of the main d-c amplifier so that its output e_2 is given by

$$e_2 = -\mu(e' + Ge' + b + Gc) \quad . \tag{4}$$

Eliminating e' from equations (1) and (4), we have

$$e_2 = -\frac{Z_f}{Z_i}\,\frac{1}{1 + \frac{1}{(1 + G)}\left(1 + \frac{Z_f}{Z_i}\right)}\,e_1 - \left[\frac{b}{(1 + G)} + c\right]\left(1 + \frac{Z_f}{Z_i}\right) \quad . \tag{5}$$

The first term above is just the output which was derived earlier in equation (2); the second term is the unbalance in the output because of the d-c amplifier unbalance of b volts and the drift-free amplifier unbalance of c volts (both referred to the respective inputs). For no drift stabilization (G = 0) the unbalance in the output reduces to $b(1 + Z_f/Z_i)$. The addition of the drift-stabilizing loop decreases this unbalance by a factor of 1 + G, where G is the gain of the drift-stabilizing amplifier; it also introduces an additional unbalance c due to a-c pickup which gets converted to d-c.

By increasing the gain G of the drift-stabilizing amplifier, the effect of unbalance b in the main d-c amplifier is proportionately reduced, but the unbalance c in the drift-stabilizing amplifier will not be reduced. Thus there is no advantage in making G so large that $b/(1 + G) \ll c$ for the values of d-c amplifier unbalance b normally encountered. Our experience has been that b is the order of 0.01 volt after sufficient warm-up time, while c is the order of 0.0001 volt for the drift-stabilizing circuits encountered. Evidently a value of G of 100 will reduce the drift b in the main d-c amplifier to the same order as the steady unbalance c (or 0.0001 volt). Higher values of G will reduce further the effect of drifts in b but will not change the intrinsic unbalance c. It is often feasible to adjust manually the unbalance b in the main d-c amplifier so that the net unbalance $[b/(1 + G) + c](1 + Z_f/Z_i)$ is zero at the output.

Drift-Stabilizing Circuit and Performance Data

The circuit developed for the drift-stabilizing amplifier is shown in Figure 13. The d-c input is passed through a low-pass filter before being chopped into a-c by a Stephens Arnold Type 252 DPDT chopper. Each set of the DPDT contacts is used for one drift-stabilizing amplifier, so that one chopper serves two operational amplifiers. The a-c input signal is amplified by a 6AU6 pentode and one-half of a 12AX7 twin triode. The a-c output is reconverted to d-c by the chopper and then passed through another low-pass filter. The input and output filters are similar to those given by Vance et al. for the typhoon amplifiers. The second-half of the 12AX7 is available for additional a-c amplification of the error signal if an external indication of excess error signal is incorporated later.

The over-all gain of the drift-stabilizing system is 330, from d-c in to d-c out. The d-c unbalance c referred to input is of the order of 0.0001 volt. The unit operates very satisfactorily with the d-c operational amplifier described previously in Progress Report No. 3 and shown schematically in Figure 12. For an operational gain of unity the unbalance in the output voltage remained within 300 microvolts of zero over a test period of 16 hours.

Power Supplies

The design of the information channel, d-c operational amplifier and drift stabilizer having been completed, the power required by a ten-channel system is now known. Hence the preliminary design of the required power sources, namely, +300, -350, and -190 volts regulated and +300 volts unregulated, has begun. The proposed power supplies will occupy one relay rack, with the associated regulation circuits being located in another rack to reduce greatly the effects of radiation from the power supplies upon the regulating systems.

After considering the voltage regulation requirements of the system, it was decided that super-regulation was not needed in the +300, -350, and -190-volt supplies, since the a-c signal amplifier will be operated with a relatively large amount of negative feedback, the cathode-follower is relatively insensitive to B^+ fluctuations, and the d-c operational amplifiers are drift stabilized. Therefore just an A regulation loop, voltage-regulating tube reference, will be used in the B^+ supplies of the above-mentioned components. Variation of the B^+ voltage of the servo-amplifier will simply cause the sensitivity of the balancing system to vary. This will not greatly affect the balancing of the system; therefore the B^+ supply

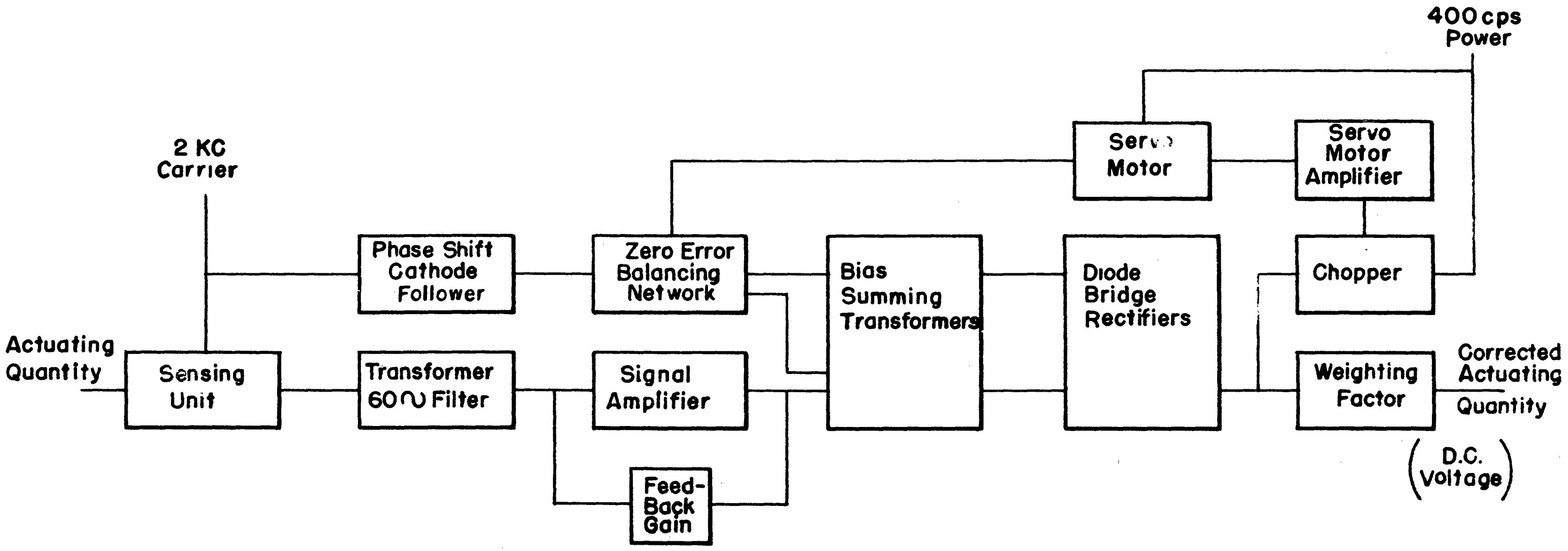
will have no regulation except that provided by a choke-input filter on the output of the power supply.

Components Test Chassis

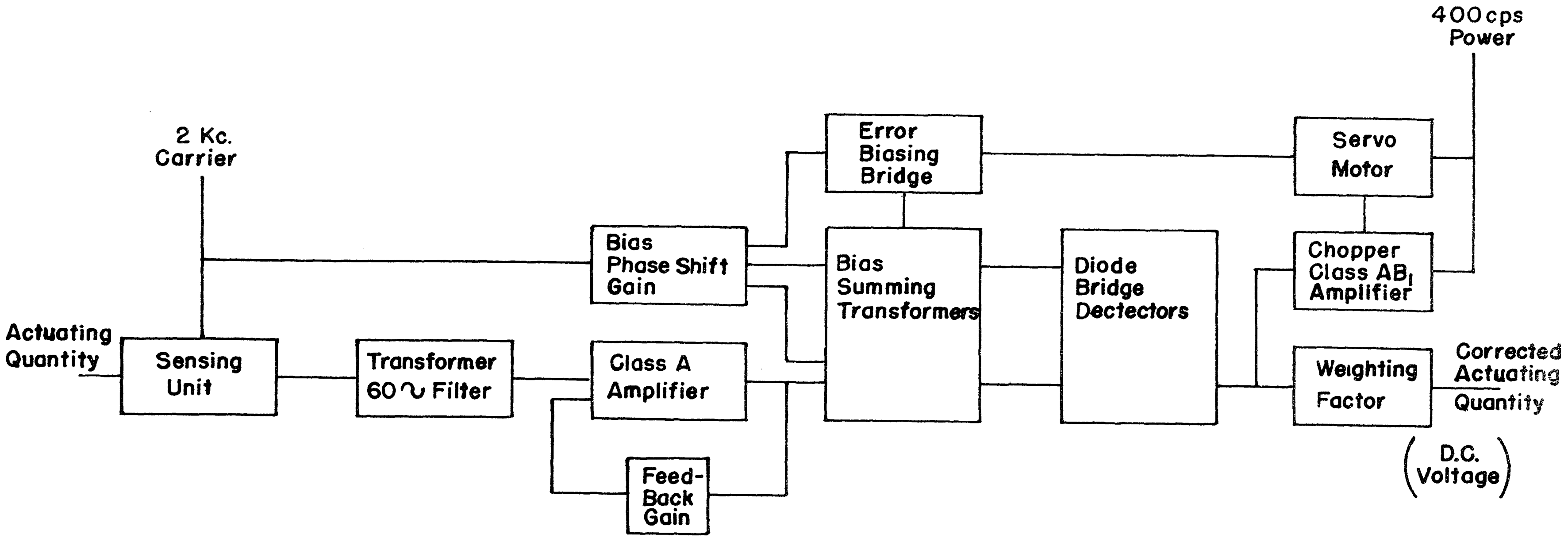
The information channels were designed, as much as possible, with each component in a separate plug-in unit. To facilitate testing each plug-in unit for proper operation as construction was completed, a test chassis was designed and built. Test of the various plug-in units was thus reduced to a process of plugging the unit into its individual receptacle on the test chassis and measuring input and output. An equally important function of this chassis is the simplification of trouble-shooting. If an information channel is suspected of improper operation, each plug-in unit can be tested individually, thus isolating the source of trouble.

The test chassis is capable of testing the following units: 60-cps filter, 400-cps chopper, type 1N40 tube, phase-shift and cathode-follower unit, both units of the Class A type input-signal amplifier individually or simultaneously, both units of the Class AB_1 type servo-motor amplifier individually only. The chassis layout is shown schematically in Figure 19.

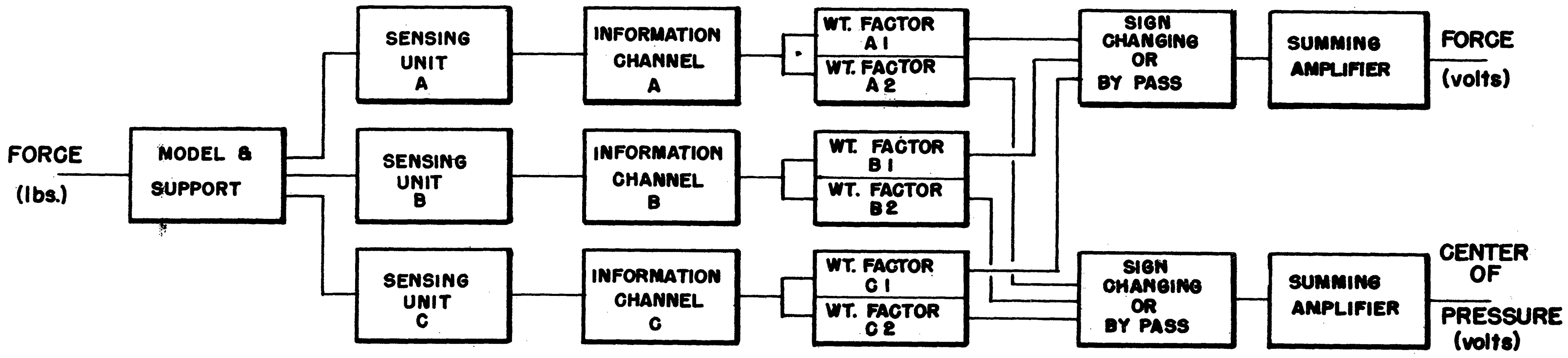
The test chassis contains its own power source, supplying necessary 300-VDC B^+ and 6-VAC filament voltage. The remaining voltages required for the various units, such as 400-cps, 2-kc, and d-c must be applied from some external source through the appropriate plug-in jacks on the chassis.



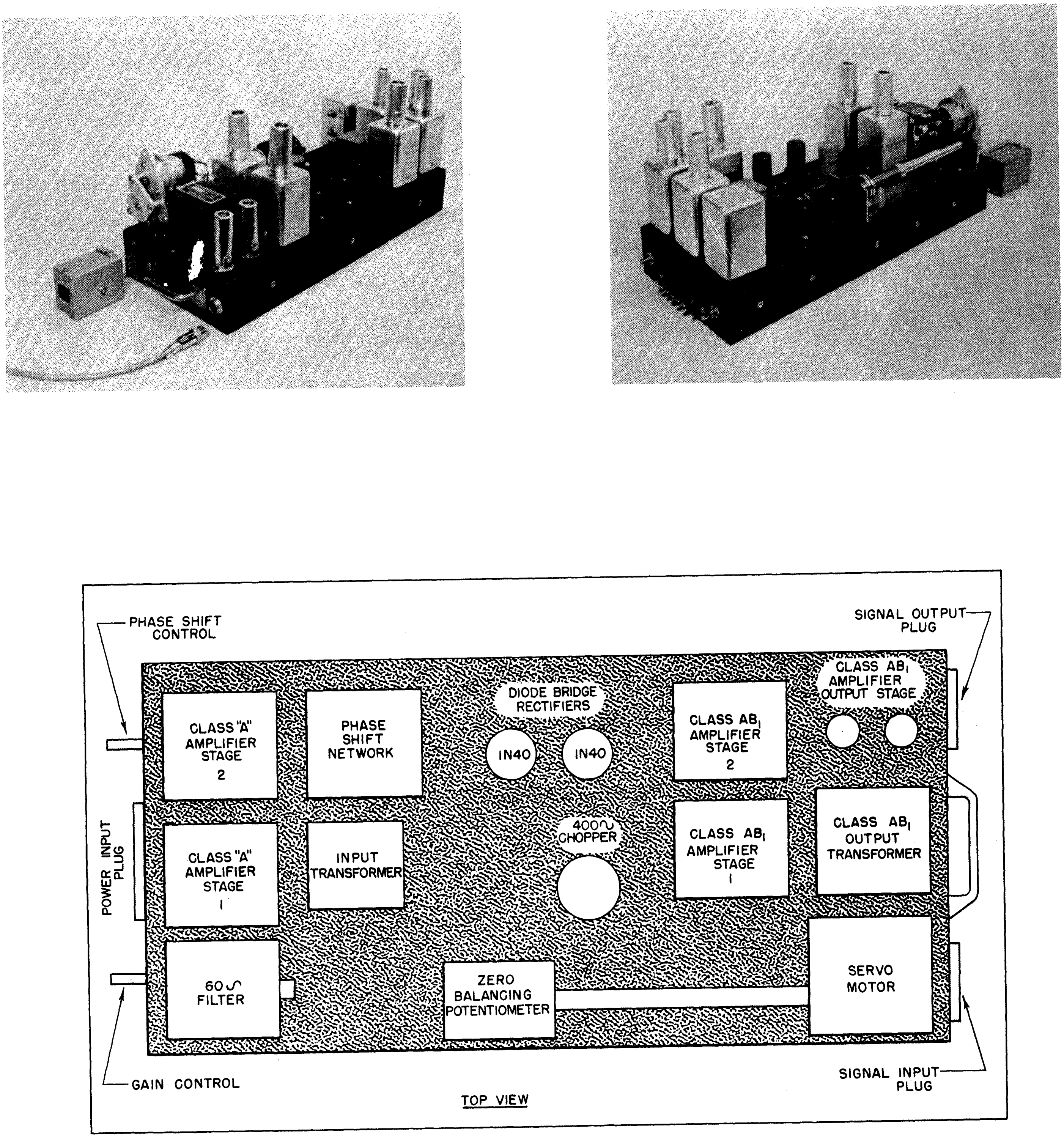
BLOCK DIAGRAM OF ONE CHANNEL OF INFORMATION
MARK I MOD. 3
FIGURE I



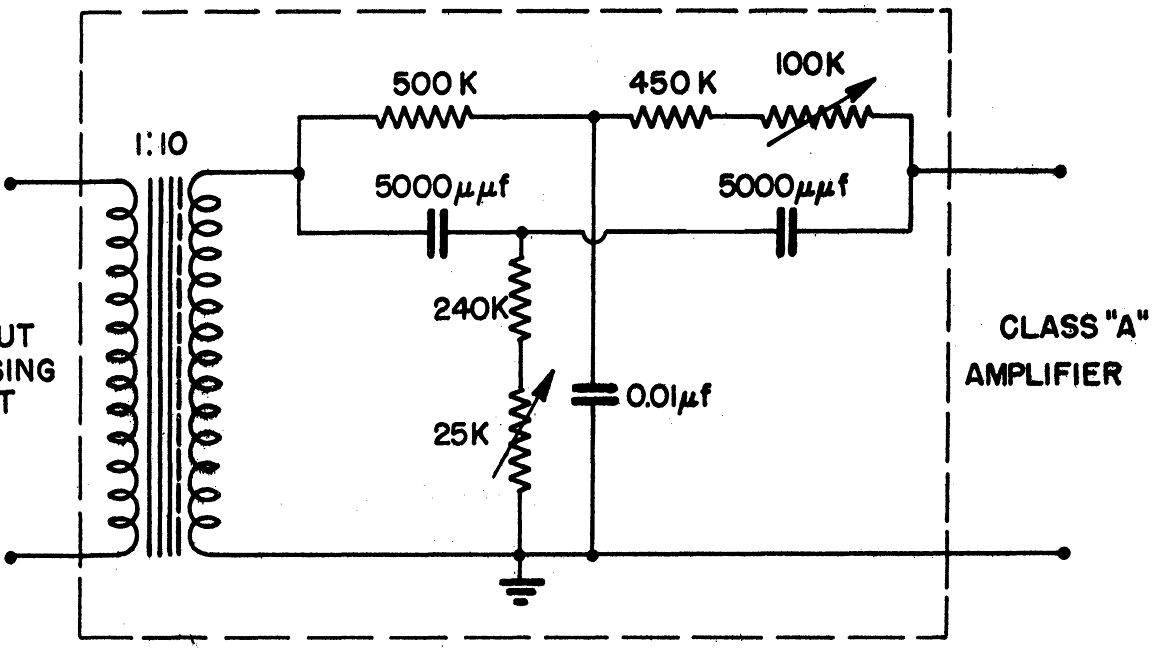
BLOCK DIAGRAM OF ONE CHANNEL OF INFORMATION
MARK I MOD. 2
FIGURE 2



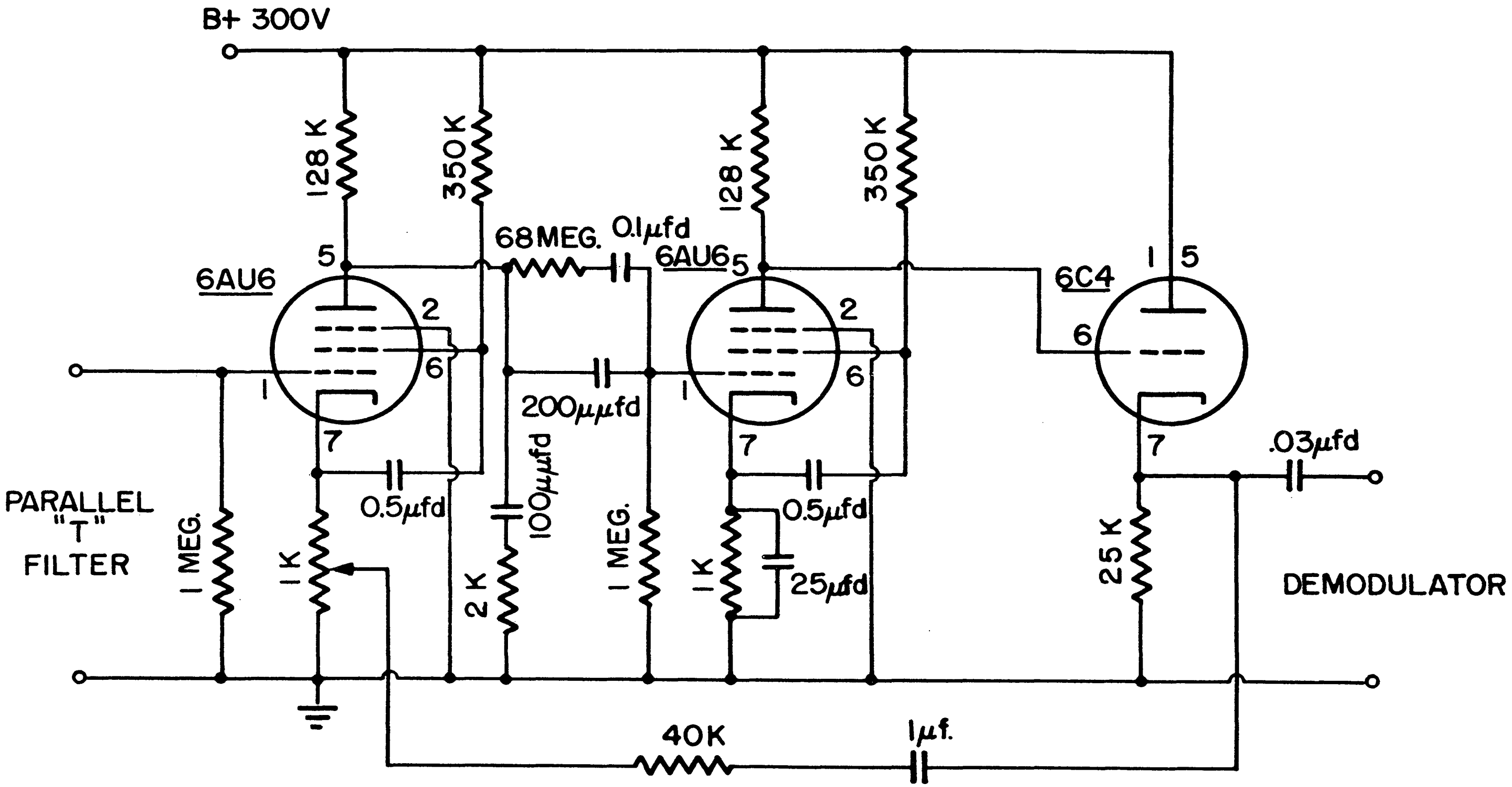
BLOCK DIAGRAM OF FORCE SYSTEM
FIGURE 3



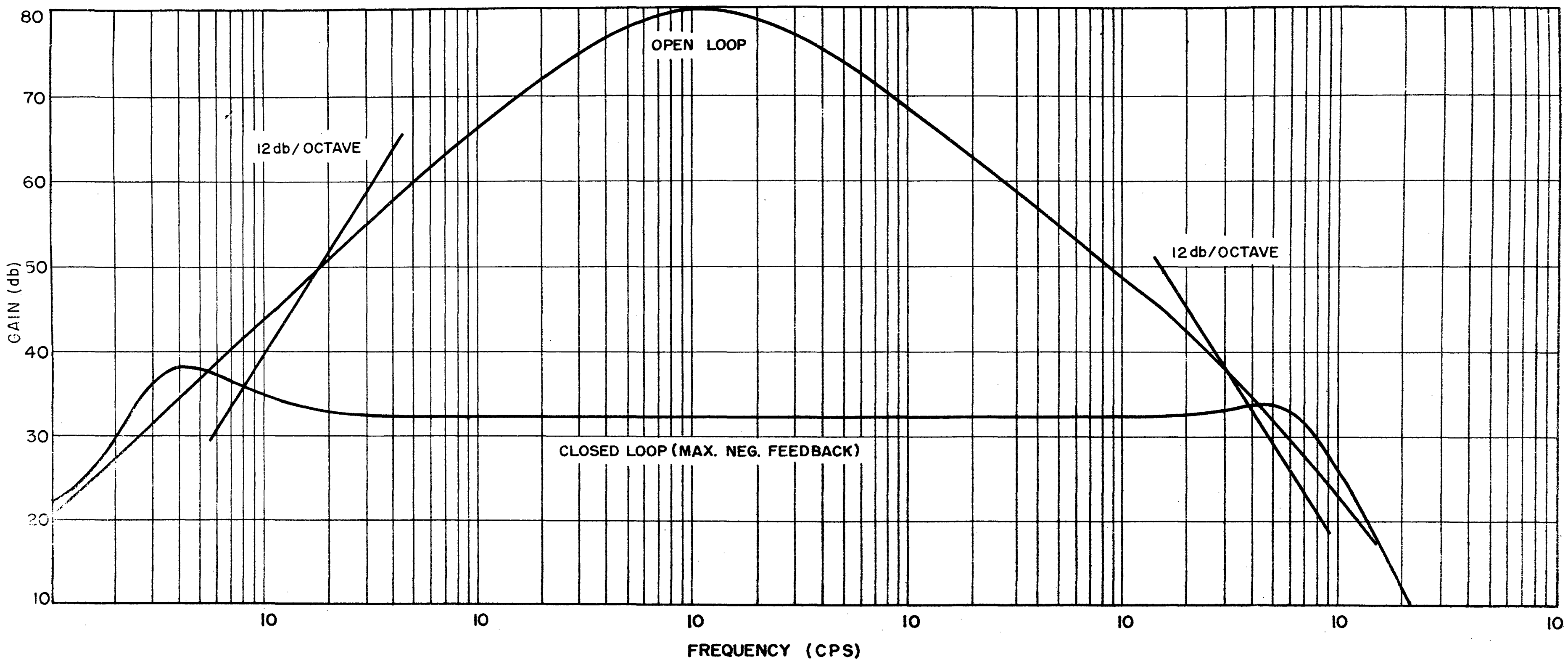
INFORMATION CHANNEL
FIGURE 4



"60" CPS PARALLEL "T" FILTER
FIGURE 5



CLASS "A" AC SIGNAL AMPLIFIER
FIGURE 6



FREQUENCY RESPONSE OF CLASS "A" AC SIGNAL AMPLIFIER
FIGURE 7

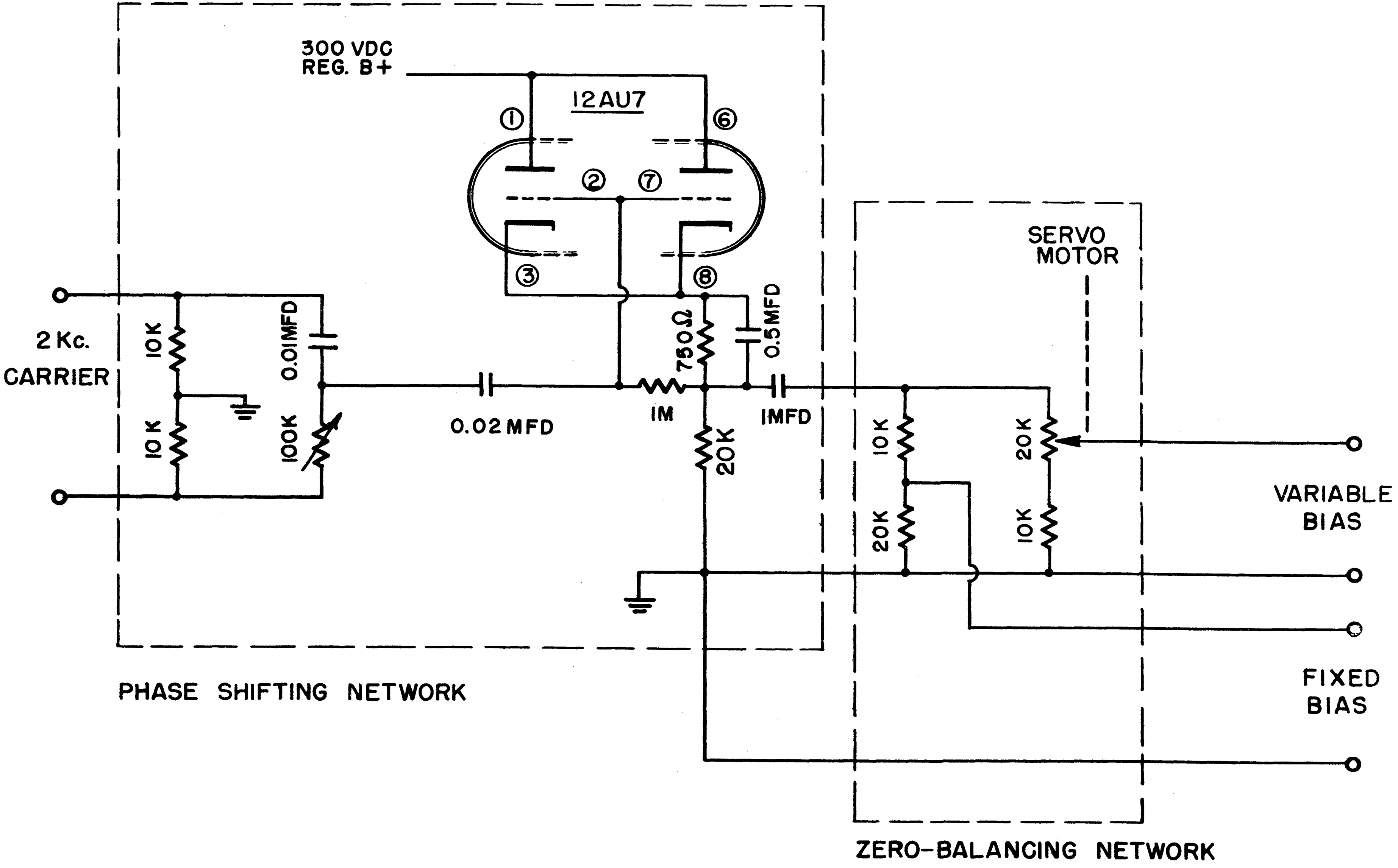

300 VDC
REG. B+
12AU7
①
⑥
②
⑦
③
⑧
2 Kc.
CARRIER
10K
10 K
0.01MFD
100K
0.02 MFD
IM
750Ω
0.5MFD
IMFD
20K
SERVO
MOTOR
10K
20K
20K
10K
VARIABLE
BIAS
FIXED
BIAS
PHASE SHIFTING NETWORK
ZERO-BALANCING NETWORK

FIGURE 8

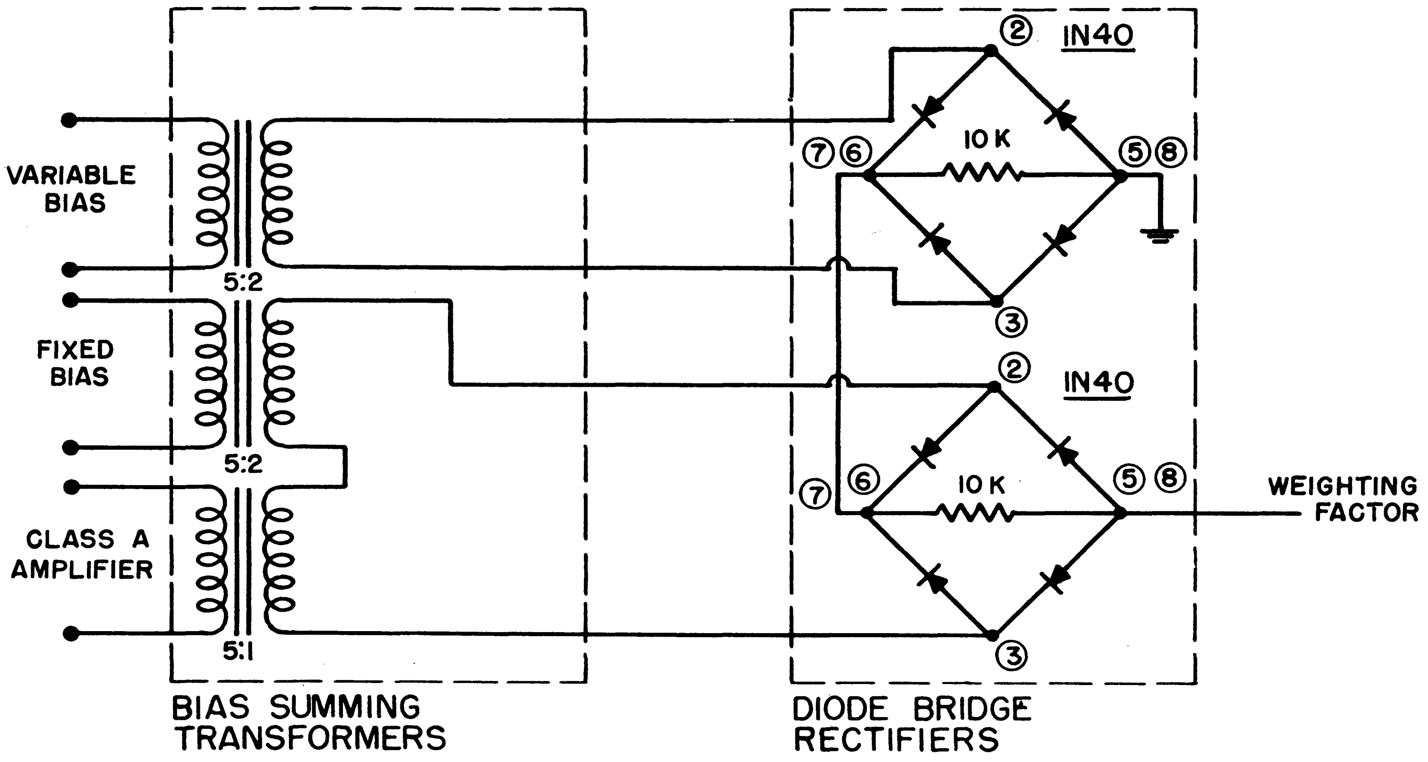
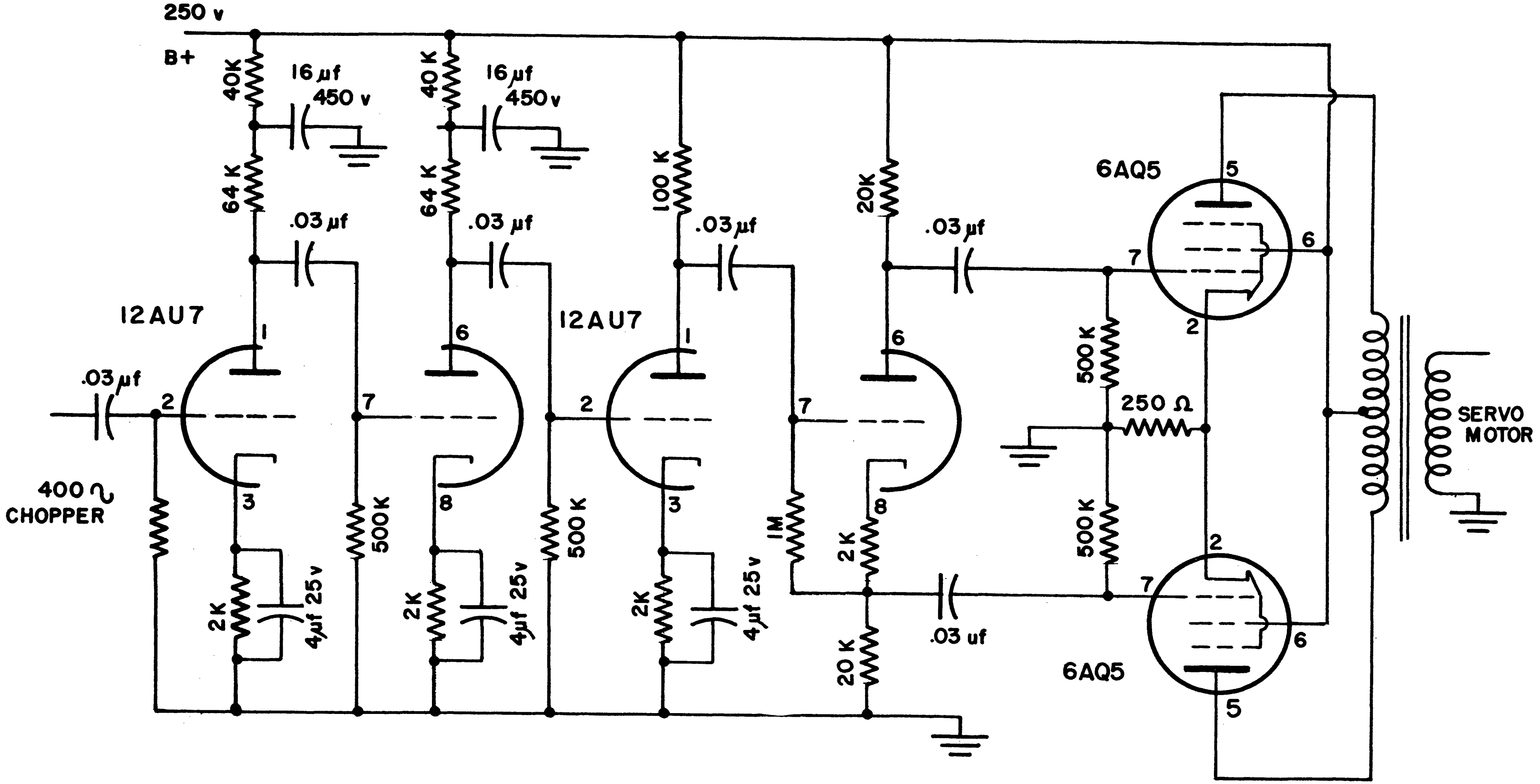
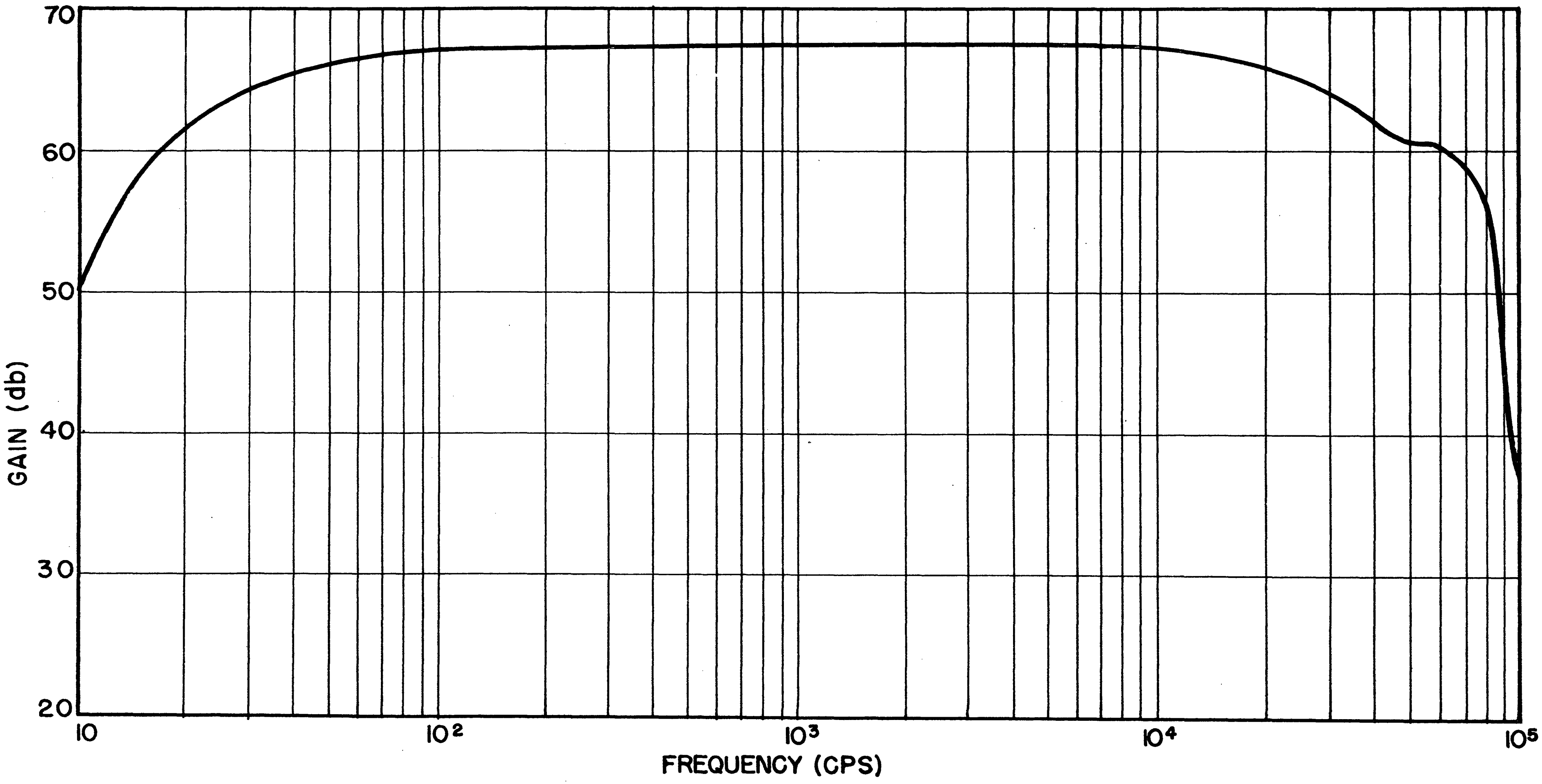

VARIABLE BIAS
FIXED BIAS
CLASS A AMPLIFIER
5:2
5:2
5:1
BIAS SUMMING TRANSFORMERS
② IN40
⑦ ⑥
10 K
⑤ ⑧
③
② IN40
⑦ ⑥
10 K
⑤ ⑧
③
WEIGHTING FACTOR
DIODE BRIDGE RECTIFIERS

FIGURE 9



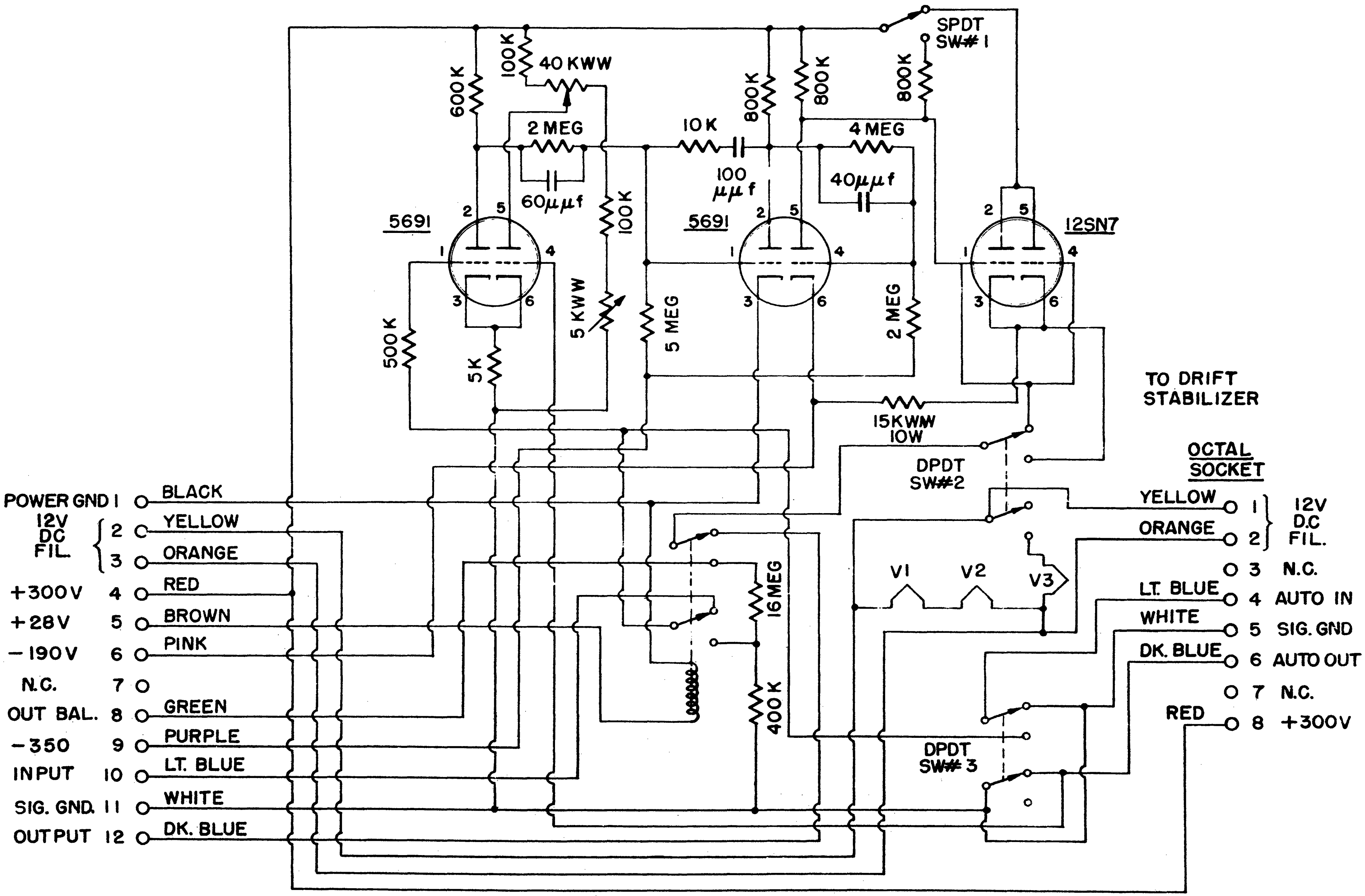
CLASS AB_1 AMPLIFIER

FIGURE 10

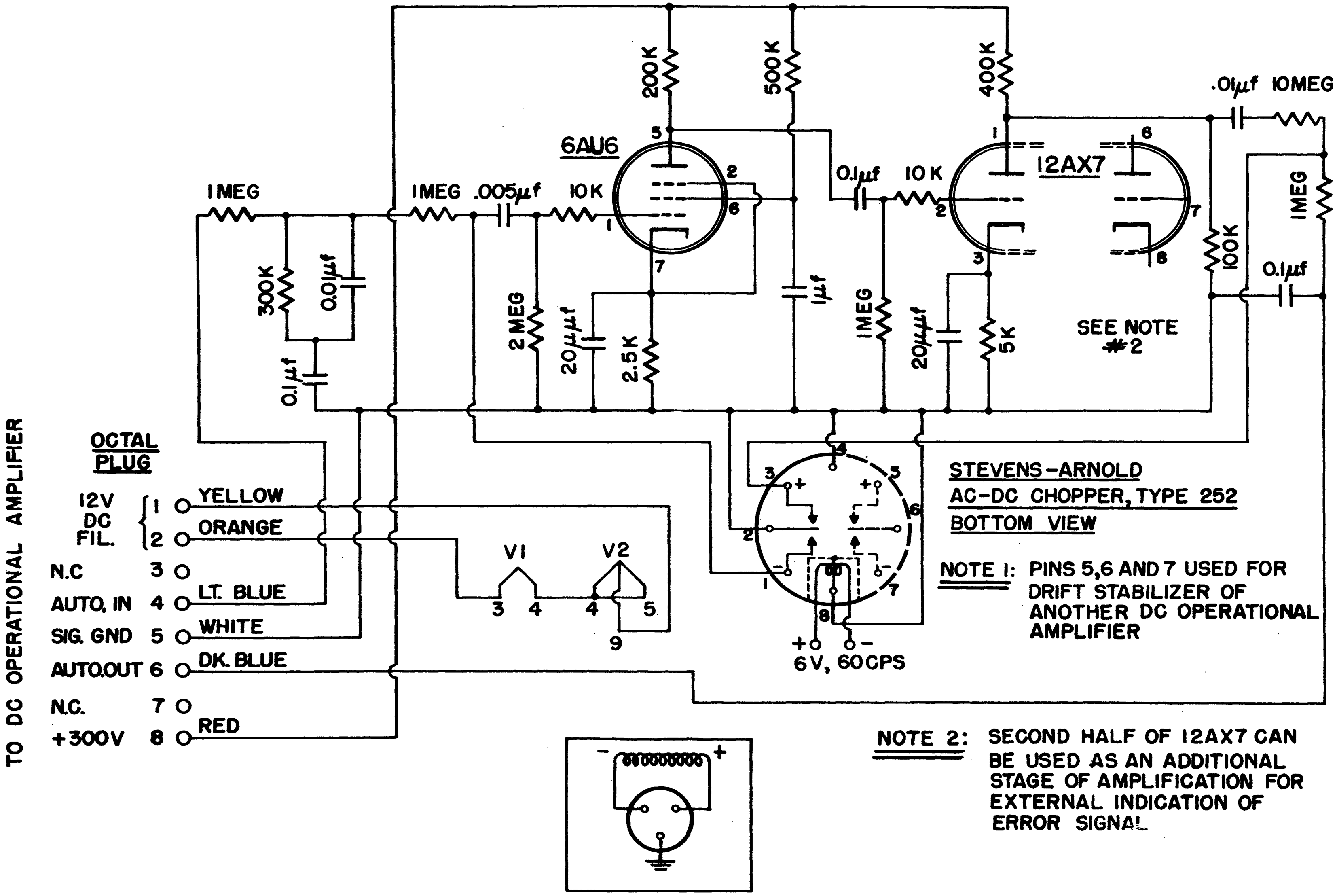


FREQUENCY RESPONSE OF CLASS AB$_1$ AMPLIFIER

FIGURE 11



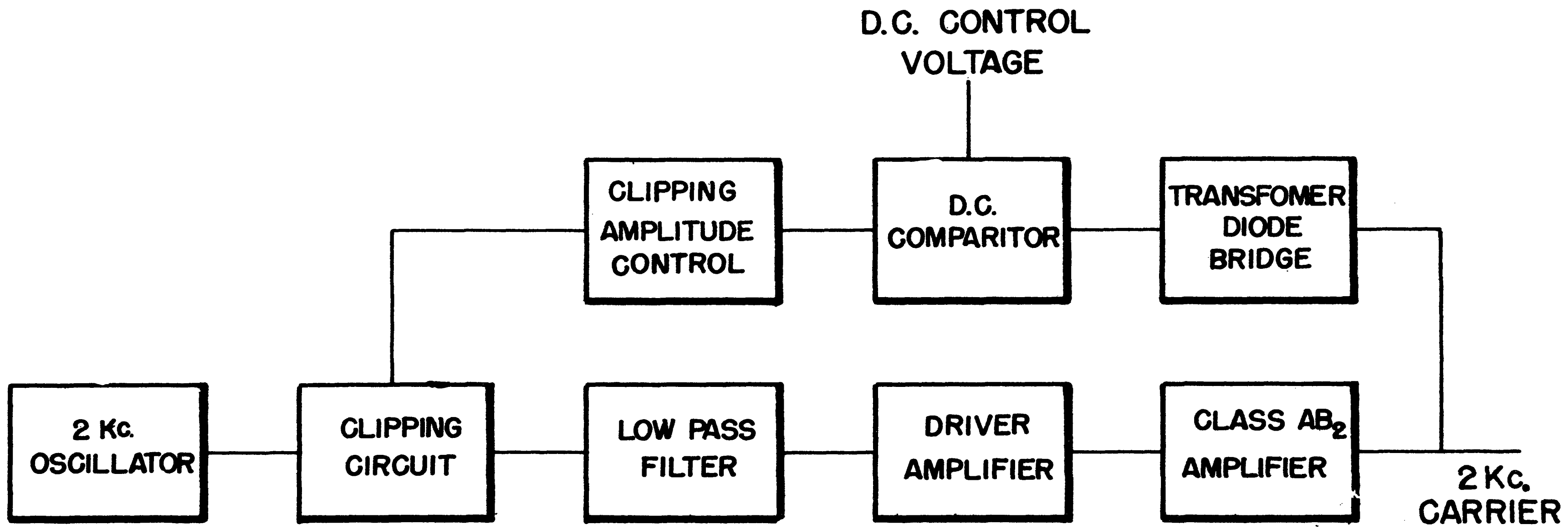
D.C. OPERATIONAL AMPLIFIER
AIR COMPUTER MOD.4
FIGURE 12



NOTE 1: PINS 5,6 AND 7 USED FOR DRIFT STABILIZER OF ANOTHER DC OPERATIONAL AMPLIFIER

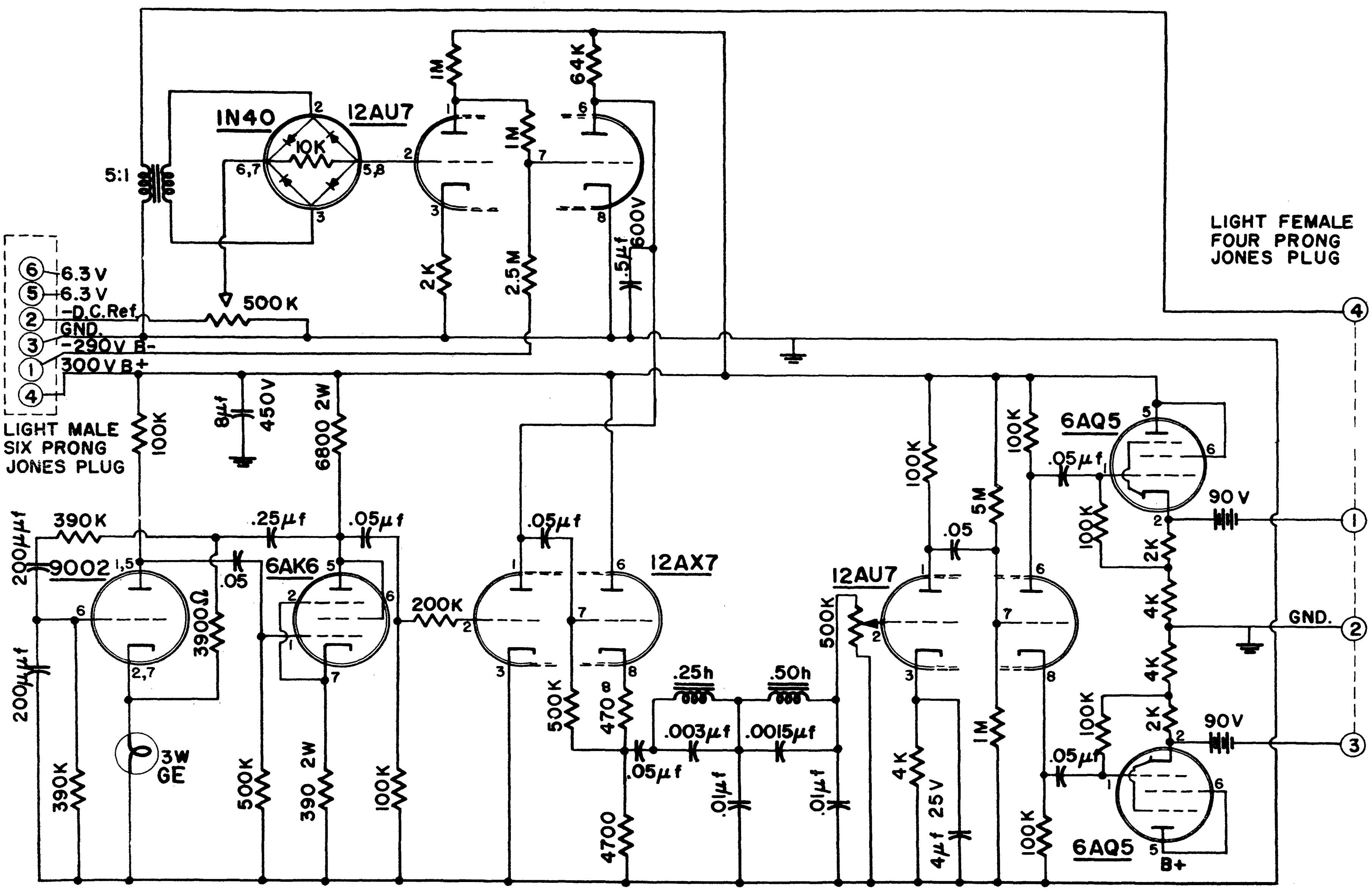
NOTE 2: SECOND HALF OF 12AX7 CAN BE USED AS AN ADDITIONAL STAGE OF AMPLIFICATION FOR EXTERNAL INDICATION OF ERROR SIGNAL

DRIFT STABILIZER
AIR COMPUTER MOD. 4
FIGURE 5

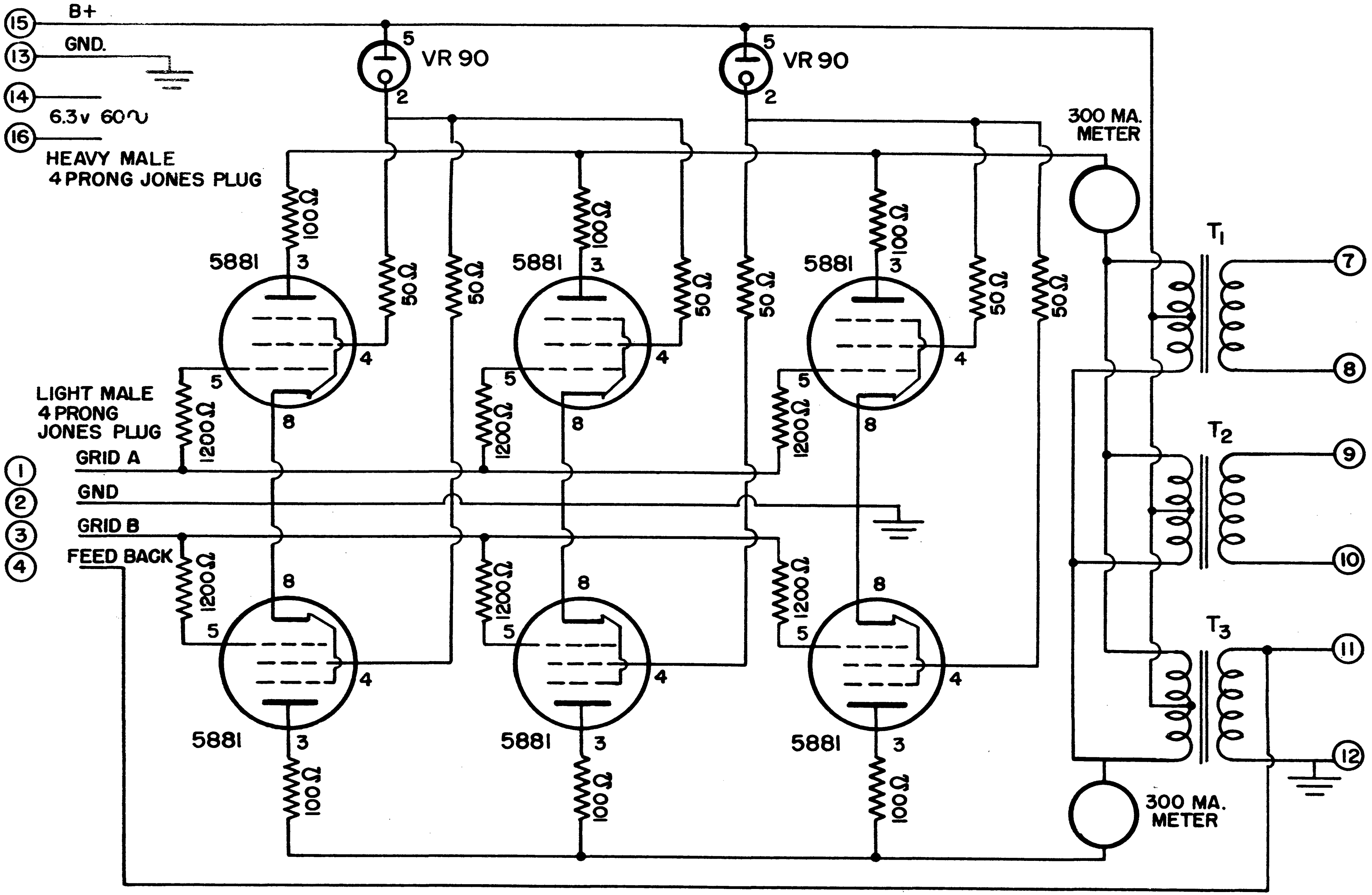


BLOCK DIAGRAM OF 2 Kc. CARRIER AMPLITUDE CONTROL

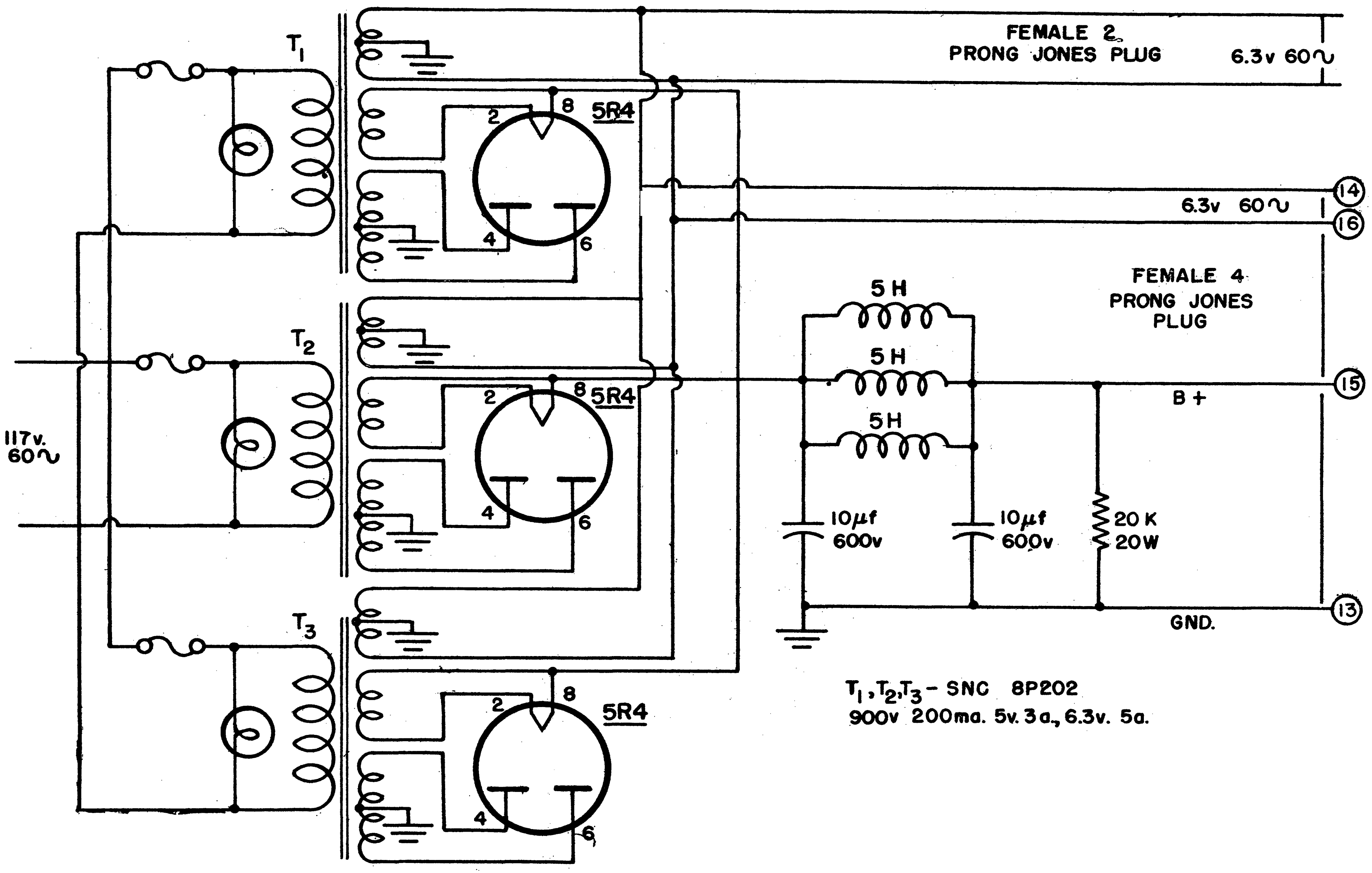
FIGURE 14



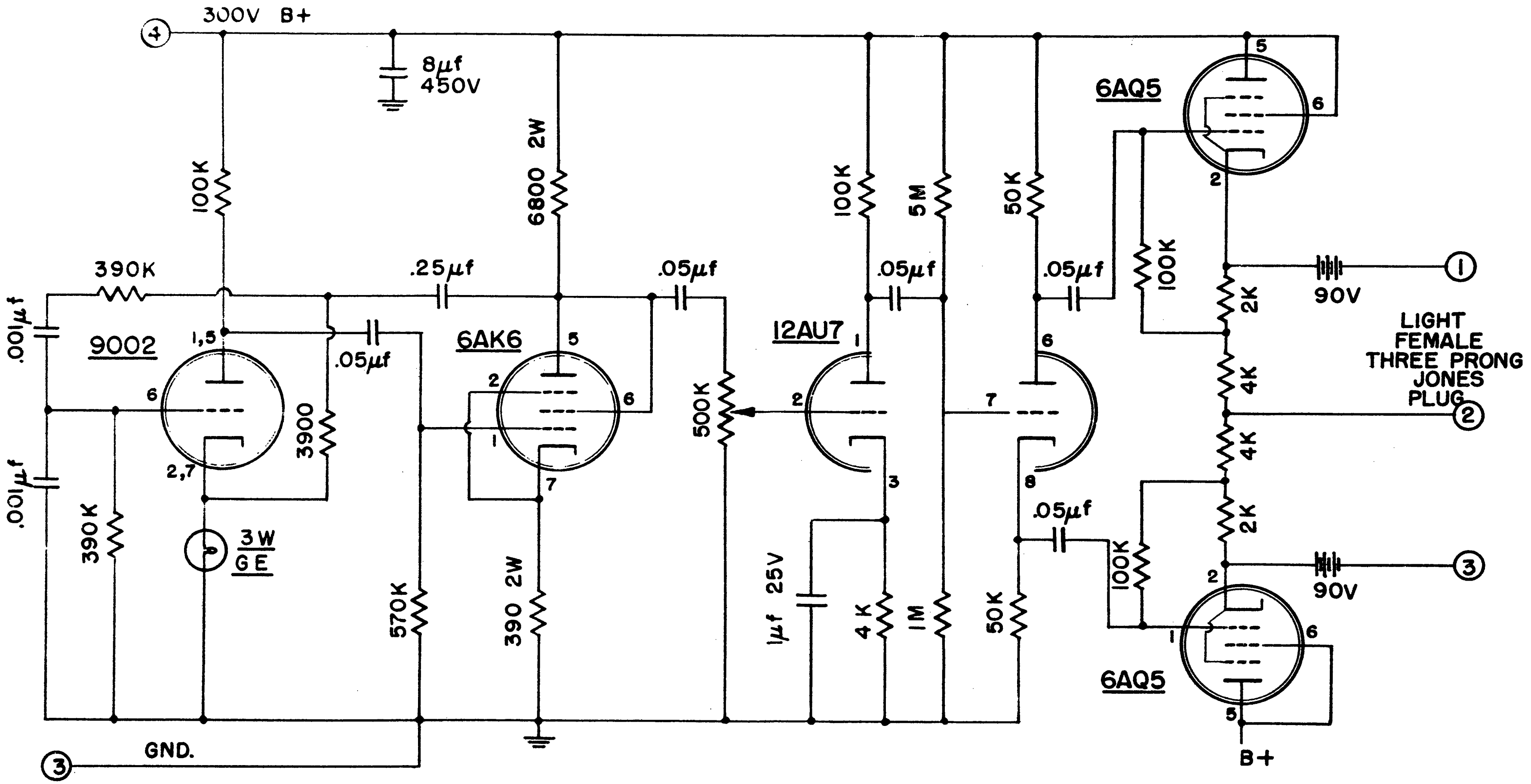
2KC CARRIER OSCILLATOR AND AMPLITUDE CONTROL
FIGURE 15



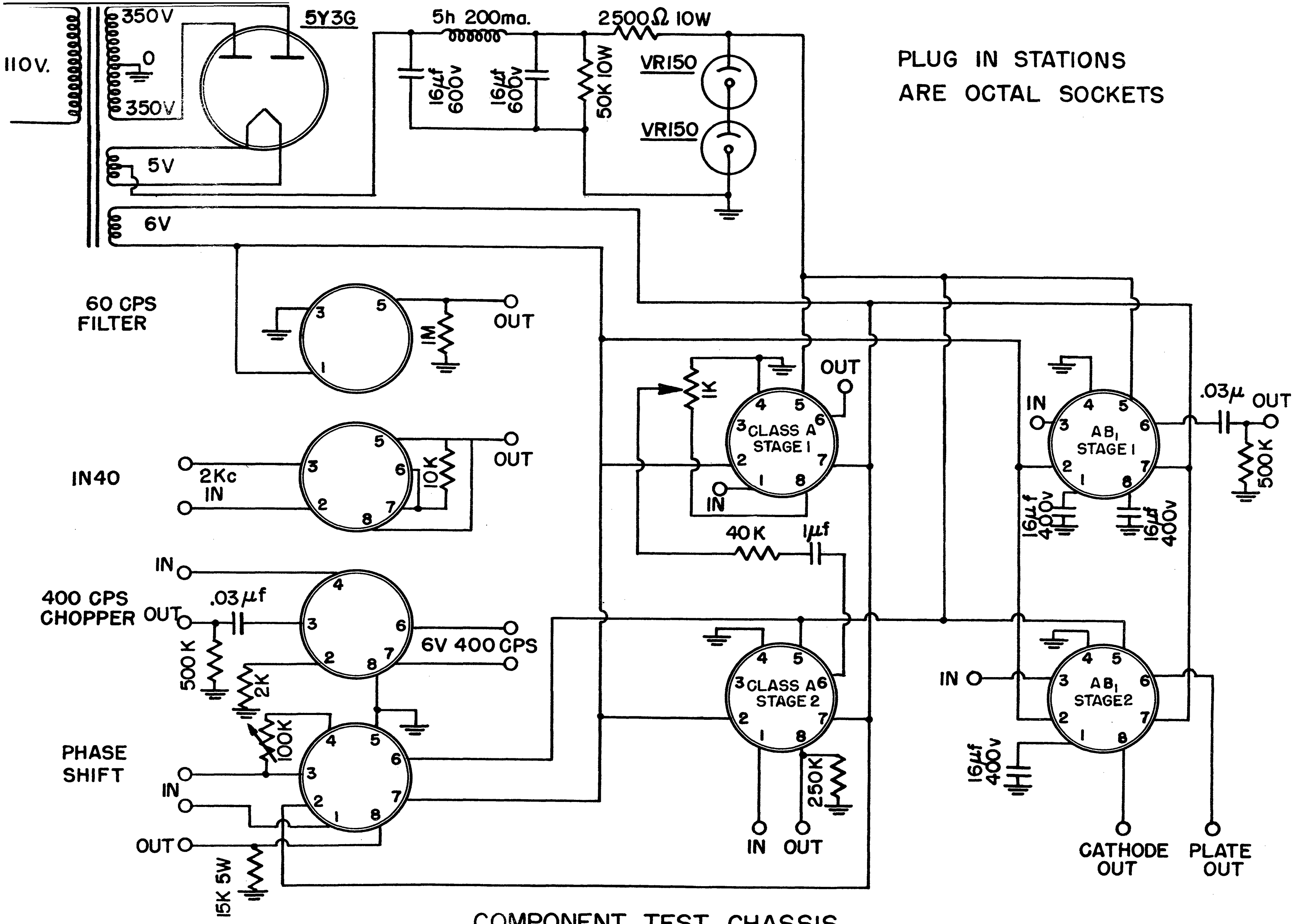
2Kc. POWER AMPLIFIER
FIGURE 16



B SUPPLY 2Kc. POWER AMPLIFIER
FIGURE 17



400 CPS OSCILLATOR AND DRIVER AMPLIFIER
FIGURE 18



PLUG IN STATIONS
ARE OCTAL SOCKETS

COMPONENT TEST CHASSIS
FIGURE 19