The Apt 1 Amplifier
Service Manual

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1 Introduction

The Apt 1 Amplifier is a power amplifier for driving loudspeakers in a wide range of home music systems. It provides a flexible complement to the Apt/Holman preamplifier, but it is meant to allow an optimum match, without detrimental interactions, to any other possible system components. The chosen design supplies the maximum useful power presently attainable without concessions in reliability or cost-effectiveness. It features novel approaches to protection circuitry and power supply design, as well as dc and ac stability. It incorporates psychoacoustically optimized output monitoring indicators, switching for both bridging (monaural) operation and loudspeaker impedance matching. A power transformer with low external flux minimizes the amplifier's radiated hum field.

The flat, U–I lamination transformer has also allowed a convenient mechanical design, with a horizontal component layout making all parts accessible at the same time. Most servicing points can be reached by removing just the top cover and the two amplifier-heat-sink modules. The circuit locations of the front panel monitoring indicators have been strategically chosen to provide considerable information about the nature of a possible malfunction even before the case is removed.

2 Specifications

2.1 General.

2.1.1 The outside dimensions of the case are 3.12 × 16.90 × 10.19" (7.92 × 42.93 × 25.88 cm). Connectors and cables require up to an additional 1.75" (4.44 cm) depth from the rear surface. The unit weight is 22 lb (10 kg). The packed weight is 26 lb (11.8 kg).

2.1.2 The front panel finish is instrumentation gray baked enamel with permanent baked-epoxy white markings; the top cover finish is neutral gray wrinkle baked enamel.

2.1.3 Input power line voltage is 120 Vac nominal. Internal rewiring for operation on nominal 100, 200, 220, or 240 Vac may be made by a qualified service technician. Any line voltage may be supplied at 50 Hz or 60 Hz power line frequency. "Brown-out" conditions will result in reduced power output capability, but other aspects of performance remain unaffected down to 95 Vac (on the 120 Vac input, other line voltage inputs have proportional "brownout" performance).

2.1.4 Data are valid for the range 20 Hz to 20 kHz unless otherwise noted.

2.1.5 Voltages in dBV are referred to 1.0 Vrms.

2.1.6 Outputs under test are loaded by 8 ohms unless otherwise noted.

2.1.7 The inputs and outputs are unconditionally stable with all source and load impedances.

2.1.8 The inputs and outputs have identical polarity, i.e., a positive-going signal at an input will produce a positive-going signal at the output.

2.1.9 All specifications are valid simultaneously unless obviously conflicting conditions obtain (e. g., bridged mode vs stereo).

2.1.10 We reserve the right to make changes as technical progress warrants; but we will not engage in capricious updating. Specifications are subject to change without notice. All specifications conform to the requirements of IHF Standard A-202; specifications to other standards, or where no standard is stated, are not comparable. *In particular, noise and distortion measurements are not comparable. N.B.: The reference level for noise measurements is 1 watt and not the more commonly employed full rated output of the amplifier.*

2.2 Levels, Gains, and Impedances.

2.2.1 The input sensitivity for 1 watt output is 91 mVrms; for 100 watts it is 908 mVrms.

2.2.2 The input impedance is 50 kohms in parallel with 300 pF.

2.2.3 The low-frequency damping factor is greater than 200. The broadband damping factor is greater than 50.

2.3 Power Output Ratings.

2.3.1 The continuous average power output per channel with both channels driven is greater than 100 watts per channel in either 4 ohm or 8 ohm loads, greater than 140 watts per channel in 2 ohm loads, and greater than 75 watts per channel in 16 ohm loads at less than 0.03% total harmonic distortion for 4 to 16 ohm loads and less than 0.05% for 2 ohm loads.

The continuous average power output in the mono bridging mode is greater than 200 watts in either 8 or 16 ohm loads, and greater than 250 watts in 4 ohm loads at less than 0.03% total harmonic distortion in 8 to 16 ohm loads, and less than 0.05% for 4 ohm loads.

2.3.2 The dynamic headroom at 8 ohms is 3 dB; at 4 or 16 ohms it is 2 dB; at 2 ohms it is 1.5 dB. In the bridging mode the dynamic headroom at 4 and 8 ohms is 2 dB; at 16 ohms it is 3 dB.

2.3.3 The transient-overload recovery time is unmeasurably short.

2.3.4 The reactive load rating is +3 dB.

2.3.5 The capacitive load rating is from zero to 0.8 μF.

2.4 Amplitude and Group Delay Response vs. Frequency.

2.4.1 The frequency response is within +0.0, −0.25 dB, 10 Hz to 30 kHz.

2.4.2 The group delay response with respect to frequency is less than 3 μsec, dc to 20 kHz. The differential group delay between channels is less than 300 nsec. The transit time is less than 200 nsec.

2.5 Distortion.

2.5.1 No one distortion test fully characterizes all known distortion mechanisms. Therefore, a combination of tests is necessary to reveal all forms of distortion, including both static and dynamic types. It has been shown that the combination of T.H.D., S.M.P.T.E. (60 Hz and 7 kHz) I.M., I.H.F. (Difference-Tone) I.M., and T.I.M. tests will be maximally sensitive to the widest possible range of distortions.

2.5.2 All distortion measurements (with the exception of ones which require greater bandwidth) are made with a Hewlett-Packard Model 3580A Spectrum Analyzer. Distortion measurements made with notch-type distortion analyzers will be inaccurate compared to spectrum-analyzer measurements since they null a chosen freqency and measure the residue in the system under test, whether that residue is distortion or noise. Also, most such distortion analyzers use an average-reading voltmeter to measure the residue. Average-responding voltmeters are inaccurate on any but simple sinusoid waveforms; therefore, they are inaccurate when reading a mixture of sinusoids (the distortion components 2f, 3f, . . .).

2.5.3 A total-harmonic distortion test consists of applying a pure sine wave to the input and examining the output for the presence of distortion products at 2f, 3f, ... The total harmonic distortion in various conditions is given under Power Output Ratings as required by the U. S. Federal Trade Commission.

2.5.4 A S.M.P.T.E.-intermodulation-distortion test consists of applying to the input 60 Hz and 7.0 kHz sine waves mixed with an amplitude ratio of 4:1 and examining the output for intermodulation products spaced at 60 Hz intervals about the 7.0 kHz tone. The distortion percentage is the rms sum of all such sidebands compared to the amplitude of the 7.0 kHz sine wave. The S.M.P.T.E. I.M. distortion at rated output level or lower is less than 0.01%.

2.5.5 An I.H.F. difference-tone-intermodulation test consists of applying to the input two high-frequency sine waves mixed with an amplitude ratio of 1:1 and examining the output for the difference product at $f_2 - f_1$. The distortion percentage is the ratio of the distortion products compared to the amplitude of the sum of the high-frequency sine waves. With 19 kHz and 20 kHz mixed 1:1 at rated output level or lower, the difference-tone intermodulation is less than 0.01%.

2.5.6 A transient-intermodulation-distortion (T.I.M.) test consists of applying to the input a symmetrical square wave at 3.18 kHz and a 15 kHz sine wave mixed with an amplitude ratio of 4:1 peak-to-peak, low-pass filtered at 6 dB/octave above 100 kHz (T.I.M. 100), and examining the output for the presence of any intermodulation products in the audio band. The T.I.M. 100 distortion is less than the measurement residual of −84 dB (0.006%) at rated output level or lower.

2.5.7 All stages meet the Jung-Stephens-Todd criteria for negligible measurable and audible slewing-induced distortion. The criteria is that the amplifier have a linear-transconductance input stage, symmetrical slewing, and adequate speed.

2.5.8 The slew factor is a measure of the amount of slew rate headroom beyond that required for a full-power 20 kHz sine wave. The slew factor is greater than 10.

2.6 Noise and Crosstalk.

2.6.1 Unless otherwise noted, all noise measurements are made with a true-rms-reading audio noise meter with psychometric weighting for the annoyance value of noise to listeners (ANSI A-weighting).

2.6.2 The output noise with a standardized input termination of 1 kohm is less than 80 dB below 1 watt (typically -90 dBV). Compared with more typical noise specifications (output noise below the full power output of the amplifier) the typical signal-to-noise ratio is thus 110 dB.

2.6.3 The 60 Hz hum components and its harmonics are below the weighted noise spectrum level. The radiated hum field as measured with a Perfection Mica pickup and a true-rms voltmeter is less than 2 mVrms at 2 in. from the amplifier to the reference mark on the pickup.

2.6.4 Crosstalk between the channels is less than 70 dB at 1 kHz and 50 dB at 20 kHz.

2.7 Dynamic Range.

2.7.1 It follows from the data above that the output dynamic range is in excess of 103 dB.

3 Test Instrumentation

3.1 **General Advice.** For additional information concerning instrumentation and procedures for audio servicing, see *The Apt/Holman Preamplifier: Service Manual*, Chapters 3, 8, and 13.

3.2 **Recommended Equipment.**

3.2.1 *Oscillator.* The test oscillator must be flat (< ±0.2 dB from 10 to 30 kHz) and must show no distortion products or noise at greater than -85 dB for 1 V rms output.

Recommended:
- *Hewlett-Packard 329A*
- *Krohn-Hite 4400*
- *Sound Technology 1700 series*
- *Tektronics SG505*

3.2.2 *Oscilloscope.* 15 MHz or greater bandwidth, preferably dual trace, with dc input.

Recommended:
- *Hewlett-Packard 1220A*
- *Tektronics 922*

3.2.3 *Volt-Ohmmeter.* 20 kohm/volt or better.

Recommended:
- *Fluke 2020*
- *Simpson 260*
- *Triplett 302*

3.2.4 *Ac Voltmeter.* This meter, used for establishing frequency response and noise levels, should be flat from 10 Hz to 30 kHz, should compute a true rms average, should allow A-weighted noise measurements, and should read accurately to -100 dBV. Many meters will require calibration or the addition of a special preamplifier, described in Chapter 13.

Recommended:
- *Radford ANM3*
- *Fluke 8050A. Add input preamp, Chapter 13.*
- *Sound Technology 1701A or 1710 Distortion Measurement System with 1200A Stereo Test Panel. Correct for ac averaging, Section 3.3.2.*
- *Sennheiser UPM 550; RV 55. Correct for ac averaging, Section 3.3.2, and add input preamp, Chapter 13.*
- *Digital Meter Research DMR-100. Add input preamp, Chapter 13.*

3.2.5 *Spectrum Analyzer/Distortion Meter.* Authentic measurements of harmonic distortion, independent of noise and spurious signals, can only be carried out on a spectrum analyzer (see IHF Standard A-202, Section 2 9.6). Nevertheless, in most cases the most useful servicing measurements can be made with a high quality distortion meter capable of measuring to -85 dB (0.006 %) in the range 100 Hz to 50 kHz and equipped with low and high pass filters (e. g., 400 Hz and 80 kHz) and a distortion output connector for oscilloscope observation.

Recommended: *Hewlett-Packard 4580A Signal Analyzer. A most desirable and versatile instrument.*

Hewlett-Packard 339A

Krohn-Hite 6800-2

Sound Technology 1700B; 1701A; 1710A.

Tektronics AA501

3.2.6 *Variable Autotransformer.* For 120 (or 240) Vac, fused and equipped with moving vane ac ammeter, ranges 0-1 and 0-10 amperes.

3.2.7 *Test Loads and IHF Standard Load.* The normal test load for servicing purposes consists of four non-inductive 8 ohm, 250 watt resistors. Wire-wound rheostat-type resistors should not be used, except in the Load Impedance Adjust set-up procedure (Section 10.8).

Recommended: *Dale RH-250*

The optional Reactive-Load Test described in Section 10.11 requires a standard reactive load, as specified in IHF Standard A-202, capable of absorbing a 30 Vac input without showing saturation effects. The Standard specifies the parallel combination of an 18.3 ohm ±1 percent resistor, a 12.5 mH ±2 percent inductor, and an 800 μF ±2 percent capacitor; the combination shall be in series with a 5.4 ohm ±1 percent resistor. Since this network is hard to construct for the power level required to test the Apt 1, it is suggested that you contact Apt Corporation for further advice about construction methods.

3.3 Calibration and Use.

3.3.1 *Frequency/amplitude calibration.* Using the 1.0 Vac range of your voltmeter, set the output of your signal generator to read 0.0 dB at 1.0 kHz. Record the readings at 10, 20, 30, and 300 Hz, as well as at 10, 20, and 30 kHz. The loaded response of your generator may vary from the unloaded response, so it may be useful to repeat this test with the IHF standard line level load, consisting of 10 kohm in parallel with 1000 pF. (The input impedance of the Apt 1 is about 35 kohm in parallel with 250 pF.)

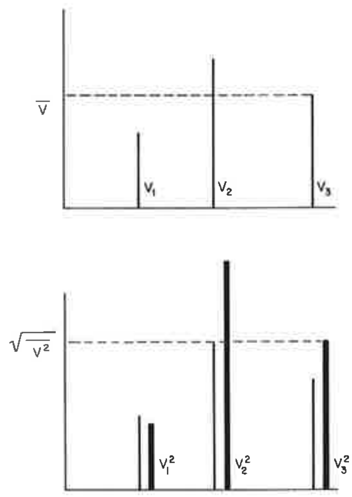


Fig. 3-1. Arithemetic vs. rms Mean.

3.3.2 *Averaging vs. rms meters.* Averaging ac voltmeters are calibrated to give the reading on a sine wave input that a true rms meter would give, and for measuring sine waves they are just as accurate. However, for other waveforms this calibration will no longer be valid (since the arithmetic average shown by the averaging meter will not equal the root mean square registered by the rms meter). With A-weighted white noise, the averaging meter will read about -1.1 dB relative to an rms meter, even though sine waves will give the same reading. (Because the sensation of loudness itself approximates a power, or rms response in most cases, the rms measurement is the preferable basis for comparison.)

If you are using an averaging ac voltmeter, add 1.1 dB to the A-weighted noise measurement.

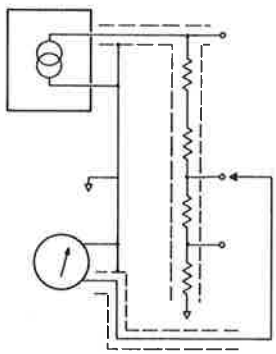


Fig. 3-2. Linearity Test.

3.3.3 *Noise Calibration.* The noise test (Section 10.10) requires reliable measurement of levels down to −100 dBV (10 μV). In addition, psychoacoustically meaningful measurement requires a weighting function such as the ANSI "A" curve. If your ac voltmeter is not capable of reading these voltages accurately, or if it does not allow for A-weighting of the input, refer to the construction project in the Appendix 13.A.

Even very good voltmeters sometimes show some non-linearity at very low levels (often because of internal noise adding to the input signal). Check your voltmeter's lowest range with a low level signal fed into a well-shielded string of four 1 kohm resistors. Relative to the unattenuated signal the center tap of the string should give -6 dB; the one-fourth division, −12 dB.

When the A-weighting filter is inserted, no change in level should be observed at 1 kHz relative to the unweighted measurement.

4 Circuit Description

4.1 Power Supplies.

4.1.1 *Transformer.* The power transformer T400 is a special design, wound on a U-I core to produce minimal leakage flux and allow an easily serviced, flat physical construction of the amplifier. The transformer primary may be connected for 100, 120, 200, and 240 Vac, 50 or 60 Hz line voltages

4.1.2 *Main Supplies.* Electrically, the power supply is the simplest part of the amplifier, but its design presents one of the most difficult tasks, since it must take into account a number of complex factors. Simply put, the designer's problem is to allow the greatest average movement of power to the load at lowest cost under the statistical conditions defined by typical musical program material. Very roughly speaking, cost determines a maximum average power level that can be accommodated by the power supply and output stage together. However, once this level has been set (arbitrarily, alas, for one never has too much power), the behavior of the supply must be fitted to the demands imposed by its intended use. Fig. 4-1 illustrates the behavior of three conceivable power supplies, all of roughly equal total capability, over varying ranges of steady and intermittent load conditions. The well-regulated, 'hard' supply can deliver (in this example) about 150 watts continuously, but power levels slightly higher than this it can provide only at infrequent intervals. The loosely-regulated, 'soft' supply, by contrast, delivers a lower power level under continuous (100 percent) demand, but at low usage rates (such as, for example, the 20 ms pulse every half second prescribed by the IHF Dynamic Headroom test, a usage rate of 4 percent) it can deliver more than twice this continuous power level.

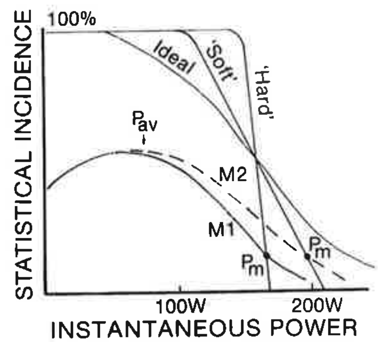


Fig. 4-1. Statistical Incidence of Demand vs. Instantaneous Power.

The power demands made by musical program sources are rarely continuous, but rather have a statistical structure. That is, if we break a typical segment of program into 20 ms segments and measure the average power in each segment, we will obtain a statistical distribution of power levels such as shown by the curve M1 in Fig. 4-1. No one power level is called for 100 percent of the time; instead, a middle range of levels is required fairly often, and higher levels are demanded less and less frequently. Theoretically, the power curve is asymptotic to the power axis, but (especially for recorded material) it is reasonable to identify a minimum level of statistical demand, say 4 percent (as in the Dynamic Headroom test) to be accommodated, and associate with it a corresponding maximum required power level, P_m. The power supply's task may then be defined as that of providing the full range of power levels up to P_m at (at least) the usage rates given by the incidence-power level curve. It must be able to supply the maximum value, P_m, but only a few percent of the time, and lower power levels at increasing usage rates.

Since turning up the volume control has the effect of expanding the curve M1 to, say, M2, it is evident that what matters most is the point where the 4 percent usage capability of the supply and the 4 percent incidence-power level of the program material intersect. The 100 percent usage power level does not by itself tell just where this point will fall. For a given cost, the best supply is the one which, like the 'soft' curve, allows the highest volume control setting. (The demand capability of the supply speculatively named 'ideal' in the Figure follows not the *average* incidence curve M2 but M2 + $(k/P)\sigma$, where k is an empirical constant and σ is the standard deviation of the incidence of each power level P, to increase the probablility that any program material will be accommodated as the volume control is turned down.)

Since a given transformer delivers maximum power at only one, unique ratio of average current to average voltage, high impedance loudspeakers in effect require a different power transformer than low impedance speakers. Switch 400 therefore changes the wiring of the transformer secondary to accommodate two loudspeaker impedance ranges. The transformer has a section of secondary windings that can be connected either in series or in parallel to deliver higher maximum voltage to high impedance speakers or higher maximum current (due to the windings' lowered internal resistance) to low impedance speakers.

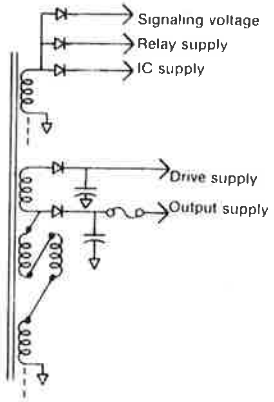


Fig. 4-2. Power Supply Levels

At high power levels a substantial signal-related ac voltage is developed across the supply capacitors C401 and C402. To prevent this voltage from reaching the drive circuitry and causing interchannel crosstalk, envelope modulation of the signal, or low-frequency stability problems, a second bridge-capacitor network, D211, C209, C210 supplies the drive circuits. Thus, while the output stage is fed from a 'soft' supply, this second network, floating some 18 volts above this level, provides a relatively 'hard' supply for the drive circuitry and realizes the advantages attributed to transistor circuits with 'hard' supplies. Its input is obtained from an additional section of secondary windings connected in series with the output supply windings; the capacitors, however, are referred to ground, so that the rectifying action of the second bridge can insulate this supply from sudden drops in voltage in the output supply. The higher voltage in the drive supply permits the output stage to be driven to its maximum voltage swing without approaching saturation of the transistors Q9 and Q11 in the voltage amplifier section.

The output devices have very low collector-to-emitter leakage, which means that the 10,000 μF output supply capacitors can remain charged for a long time after turn-off. This improves the turn-on characteristics of the supply by reducing the normal initial current transient, but presents a danger not to be overlooked in servicing.

4.1.3 *Auxiliary Supplies.* A number of peripheral circuits, including the relay, protection, level indicator, and dc offset circuits, require stable, low-level dc voltages. To anticipate turn-on and turn-off conditions, these circuits also require a signaling voltage with rise and fall times less than those of the supply voltages.

A separate, smaller secondary winding on the power transformer feeds a ±12 volt dual supply (D200) for the operational amplifier ICs and dc offset reference (R63). D201–C203 form a half-wave supply to the output relay coil K200; D203–C204 form a second half-wave supply with a small time constant to signal turn-on and turn-off (more at Section 4.7).

4.2 **Design of the Amplifier Circuit.** The power amplifier is based on the usual differential-input, complementary-symmetry design. No output current or voltage limiting is applied within the circuit, since it has been found that the side effects of such circuits invariably interfere with an amplifier's sonic performance. Instead, external sensing circuits reduce the signal level or disconnect the load when unsafe operating conditions arise. However, a number of measures have been taken to improve the linearity and stability of the conventional design which warrant special discussion.

A differential pair with high-impedance, Darlington inputs has been chosen for input impedance, non-saturation, and slewing characteristics; it also allows precise control of the amplifier's dc offset level. Normally only one leg of this differential circuit is used to pass the signal; its inherent symmetry is not exploited. The familiar complementary-symmetry circuit shown in Fig. 4-3 thus obtains a single ac voltage from the differential amplifier and passes it to the common-emitter drive transistor, Qd, which generates the output voltage presented to the speakers. Hence, the complementary-symmetry output stage in fact only functions as a unity-gain current buffer to what is really a *single*-ended voltage amplifier stage. The class-A drive transistor, Qd, must operate over a wide range of collector voltages and currents, and the variations with voltage in its operating parameters—especially internal capacitance and beta—can induce distortions, so that the choice of this transistor becomes highly critical. The advantages of symmetrical, push-pull design sought in the complementary current amplifier can be compromised by this high signal level, single-ended stage.

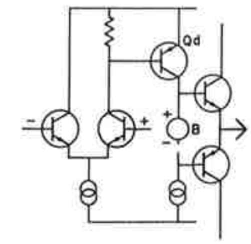


Fig. 4-3. Conventional Drive Circuit.

The fully symmetrical drive circuit in the Apt 1 adds a second device, Qd2, to the opposite end of the voltage amplifer (see Fig. 4). (Qd1 and Qd2 are realized in the actual circuit as cascode pairs, Q8-Q9 and Q11-Q12, for improved linearity, higher collector impedance, and higher drive capability.) The choice of these transistors becomes less critical, since their non-linearities tend to cancel in push-pull (class A) operation. Because current can be supplied to the output buffer from either end, the average current in the drive loop can be kept low and small-signal drive transistors with high cutoff frequencies and other desirable characteristics may be used. Furthermore, the normally unused leg of the differential amplifier may be employed to drive Qd2. A current mirror, formed by Q5 and Q6 and inserted in the unused leg, inverts the phase of the ac current in this loop and makes it available from a high impedance source to drive a second 680 ohm resistor referred to the opposite supply rail. Symmetrical design is thus realized throughout the amplifier.

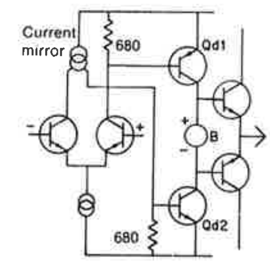


Fig.4-4. Fully Symmetrical Drive Circuit.

Another departure from conventional complementary-symmetry design is found in the Baker clamp formed by diodes D5-D10. Normally, the output swing of an amplifier is limited by the available supply voltages: when a signal waveform comes to within less than a volt of the supply rail, the transistor saturates and clips the signal voltage. At this point the equations governing the open and closed loop behavior of the amplifier cease to be valid; other, sometimes highly erratic mechanisms come into play which can produce ringing, oscillation, recovery problems and other distortions much in excess of simple clipping.

Because the drive circuitry in the present design is supplied at a level considerably beyond the output stage's voltages, it could drive the output quite far into saturation, where considerable charge would be stored in the base regions. Before the transistors can recover from this saturated condition, the charge must be bled back into the drive circuit, resulting in a short interval after the end of clipping where the transistors are unable to respond to their inputs. The Baker clamp precludes this by diverting the excess drive current into the supply rails and maintaining a voltage of at least one diode drop across the base-collector junction of the predriver transistor. While the output stage can be driven to within a few tenths of a volt of saturation, clipping results from the relatively simple behavior of the diode clamps. Overload recovery time is thus unmeasurably low, and the amplifier remains unconditionally stable under all drive conditions.

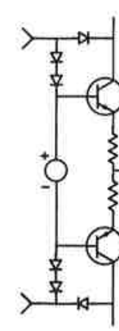


Fig. 4-5. Baker Clamp.

The circuit design and choice of components emphasize linearity as well as open-loop gain and frequency response. It is often felt that stability and freedom from slewing distortion must be bought at the expense of open-loop gain; however, the slewing capabilities of modern transistor amplifiers generally exceed by many times the requirements imposed by any conceivable audio program source. The one real threat to the amplifier's slewing ability—radio frequency interference—has been countered by an RF filter, R2-C2, at the input. The use of very fast transistors, even in the output stage, in the present design has allowed an open-loop gain considerably in excess of 100 dB, with single-pole compensation, set by C4 and C9 acting in parallel. The gain-bandwidth product is about 30 MHz; the phase margin is approximately 85 degrees.

While this large bandwidth was made possible by the recent development of a fast pnp complement to the npn output transistor, bandwidth was not the first consideration in the design. The decision not to use the even faster V-FET complementary devices now available—with their many indisputable advantages—was made in favor of the ability to interface well with difficult loads. The bipolar devices chosen offer a larger safe operating area than could be obtained with FETs, and thus greater drive capability when confronted with large out-of-phase currents. With proper drive and protection circuitry, they can be made to behave as well as FETs, and at present they can provide some seven times as much safe operating area for a given cost.

4.3 **Operation of the Amplifier Section.** Input signals are applied to JA200 and JA201. Optimum RF grounding is obtained through a chassis ground (⏛) connection to the mounting rivet at the side of the left channel jack. The remaining two shield pins of the jack connect to the ground foil on the protection board, which connects to the capacitor ground bus, or common ground (⏚). The input jack chassis ground is the *only* chassis ground in the amplifier.

C1 blocks dc signals from the input which could trigger the dc protection circuit. (However, very low frequency signals will be passed on to the amplifier and may activate the dc detector, Q202, to open the output relay. A phonograph preamplifier used without any infrasonic filtering may trigger this protection at high levels.) For monaural, bridging operation, the Stereo/Mono switch, S100, replaces the right channel signal with the left channel input, inverted by IC101.

R2-C2 form an RF filter to exclude possible RFI sources. R5, the "Moxie" thermal resistor, is mounted on the heat sink bracket, opposite the biasing transistor Q10. When the heat sink temperature exceeds 85° C, the resistance of this device falls rapidly from around 130 kohms to less than 100 ohms, reducing the level of the input signal. Q1 provides a high impedance, Darlington input to the differential amplifier Q2-Q3. The total current in this section is set at 4 mA by D2 and R16 in the current sink Q7. The base circuit of the current sink operates between the output supply -B and the driver supply -D; the diode D17 prevents C26 from discharging abruptly at turn-off.

The two legs of the differential amplifier trade between them the 4 mA supplied to the current sink so that the ac signal currents are in opposite phase. (Since a positive-going input to Q2 will cause *more* current to flow in this transistor, the voltage across R10 rises; the base-emitter voltage in Q3 falls, and *less* current flows in Q3. An input applied to Q3 will have an equal but opposite effect to one applied to Q2; the current in R11 thus develops a voltage proportional to the *difference* between the voltages at the two bases.)

Because it is in opposite phase, the current in the other leg of the differential amplifier is normally ignored; but in this design it is inverted by the current mirror Q5–Q6 and fed to a second 680 ohm resistor, R28, which feeds a second drive transistor, Q12. The result is a complementary, push-pull drive rather than the more usual single-ended drive terminated with a current sink. The point between Q9 and Q11 represents a high impedance node in the circuit, where the high output impedance of the collectors of Q9 and Q11 faces the high input impedance of the Darlington pre-drivers of the current amplifier. Accordingly, the currents here are normally very low, even though the full output voltage (as much as 120 volts peak-to-peak) is present. (A fault causing loss of gain in the current amplifier, of course, will change this.) The open loop voltage gain of the circuit as a whole is rolled off by the two 25 pF capacitors C4 and C9, which feed back a portion of the collector voltage to R11 and R28. The 1 nF capacitors C3 and C10 provide phase lead but do not affect the open-loop amplitude response.

The voltage present at the output to the loudspeakers, test point O, is fed back through a voltage divider, R30–R20, which sets the closed loop gain at 31. This voltage gain is obtained entirely in the differential amplifier and voltage amplifier, so that the full output voltage can be seen at the bases of Q13 and Q14, and at any point beyond. When current flows out of the amplifier, it does so through the string of current multipliers Q13–Q17; when it flows into the amplifier, it flows through Q19–Q14. This means that an oscilloscope probe referenced to ground will show the ac output voltage at any point in the current amplifier's signal path, but when placed across any of the emitter resistors, it will show current flowing in that side of the current amplifier only slightly more than half of the time (see Fig. 4–6), and that the currents increase dramatically from one stage of the current amplifier to the next.

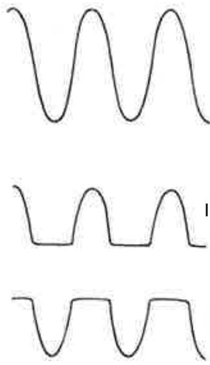


Fig. 4–6. Output Currents and Voltages.

When the output devices warm up, the current through them tends to increase. The biasing transistor Q10, mounted on the heat sink, then lowers the bias voltage between the two sides of the amplifier to maintain a constant quiescent current level. The bias level must be held high enough to ensure a smooth transistion from one side to the other in the current amplifier. At low frequencies, the circuit is fast enough to 'jump' between the two sides without evidencing any crossover notch in an underbiased condition; thus, it is important to examine the output for crossover distortion at high frequencies, where a break in the waveform (often not visible on the oscilloscope, but readily apparent as a spike in the distortion output of a distortion meter—see Fig. 10–2) will reveal an underbiased condition. Minimum distortion is obtained with about 20 mA of quiescent current in each of the output devices, or about 40 mA per channel. (More bias current actually *raises* the distortion. The gains of the pnp and npn devices differ, but they are current dependent, and at around 20 mA the gains find their best match.)

The output inductor L1 is necessary to prevent a condition which can arise with some loudspeakers or with cables which present capacitive loads at high frequencies. Under this condition, where the output sees a low, largely capacitive impedance at very high (radio) frequencies, the signal presented to the feedback loop lags the input by a substantial phase angle in the radio frequency region; added to the amplifier's own compensation phase lag, this additional lag can produce a small net positive feedback component. A radio frequency oscillation can arise, supplying enormous currents that can destroy the amplifier or the load. The 1.3 μH inductance used here is large enough to hide the feedback loop from a capacitive load at radio frequencies but has no effect in the audio band. Capacitors C23 and C403 help to exclude radio frequency interference from the feedback loop.

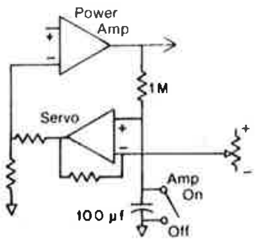


Fig. 4-7. Offset dc Servo.

4.4 **Servo (dc) Amplifier.** Normally, dc stabilization of a direct-coupled amplifier is obtained through the amplifier's feedback loop. A large capacitor terminates the feedback divider, reducing the dc gain to unity and minimizing the inherent offest, which is then further reduced with a trim voltage that brings the output voltage to about zero. However, for some loudspeakers, especially those with transformer inputs having very low (dc) resistance, the dc control provided in this way is not sufficient: even a slight offset will cause a large dc current to flow into the speaker, impairing the performance of the speaker and of the amplifier. In addition, the capacitor in the feedback loop represents a departure from strict direct-coupled design; it causes severe voltage swings at turn-on and turn-off and can introduce distortion, particularly envelope distortion. For these reasons, dc control of the Apt 1 has been obtained through an external servo amplifier not involved in the signal path. Since the amplifier's own feedback loop is not affected, the characteristics of the servo can be chosen at will.

One half of the operational amplifier IC1 functions as a high-gain, non-inverting voltage multiplier, feeding the power amplifier's dc level back to its inverting input. The output signal reaches the non-inverting input to the op amp through a low-pass filter composed of R57 and C16, having a time constant of about 100 seconds. When the amplifier is turned off, C16 discharges to ground through the FET Q23; however, as soon as the negative supply voltage -D appears, the FET is biased beyond cutoff and opens this ground connection. R58 and R56 set the voltage gain of the amplifier at about 280; C24 causes the ac gain to fall to unity to provide further ac filtering of the feedback and reference voltages. (Failure of this capacitor can thus introduce at the inverting input a low level hum from the reference voltage source.) For dc voltages the servo amplifier effectively replaces the 27 kohm resistor in the main amplifier's feedback loop and causes the amplifier output to appear 280 times larger—this amplified voltage, divided by 909/47 K, and applied to the inverting input, yields an overall dc closed-loop gain of only 0.19, so that a compensating trim for dc offset is not necessary.

Note: the offset trimmer, R53, does *not* trim the dc offset level at the loudspeaker output—its voltage range is much too small—but rather presents at turn-on, while the power amplifier is becoming operational, a voltage which anticipates the later operating levels of the loop and thus minimizes the initial settling time. It is trimmed for minimum settling time and not for minimum offset.

4.5 **Signal Indicator.** To indicate the presence of an audible signal at the loudspeaker terminals (*after* the relay), the output is monitored by the signal indicator circuit IC300. Its input is clipped by diodes D301 and D302, and positive-going signals are passed through D304 to charge C301 and illuminate the green LED. Current is limited internally by the IC. The gain, set on the positive halfcycle by R303 and R301, gives this amplifier a sensitivity of about 15 mV at the speaker outputs. D303 lowers the gain of the amplifier on the unused, negative half-cycle to about unity.

4.6 **Overload Detector.** Because there are no significant phase shifts in the power amplifier to discount in making a distortion measurement, the amplifier's distortion performance can be assessed very simply by comparing instantaneous voltages at the input and output terminals. As is often the case in psychoacoustic measurement, however, this raw difference signal does not always correspond well to the ear's sensitivity to distortion. In particular, the ear can tolerate distortion signals of very short duration and large magnitude much more easily than those of low magnitude but relatively long (>10 ms) duration. The overload indicator is constructed to respond in a meaningful way, somewhat ahead of audible distortion, but not to give false alarms.

The second half of the dc servo, IC1, serves as a difference amplifier to detect discrepancies between the input and output of the power amplifier. A precision divider, R16–R61, reduces the output signal exactly to input level; what remains at the output of the difference amplifier is then the distortion. The threshold and clipping amplifier, IC300, has an exponential gain, which causes it to ignore very small signals, but once either D320 or D321 begins to conduct, the gain rises with amplitude, as the effective resistance of the diode falls. The output is then clipped by the IC's internal limiting. In the next stage of processing (see Fig. 4-8), the clipped signal passes to the rectifier bridge formed by D306–D309, and the rectified voltage, integrated by the filter C302–R305 lights the red half of the indicator LED, D310. The condition for a distortion indication is thus not simply the presence of a discrepancy at the comparator output, but rather a discrepancy which approaches meeting the criteria set by the ear for audibility (and by the eye for visibility). (Because of the poorer common mode rejection of the IC at high frequencies, however, very high frequency signals will sometimes produce enough comparator output to light the distortion indicator before clipping actually occurs.)

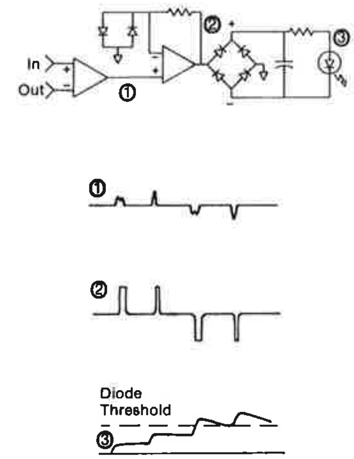


Fig. 4-8. Distortion Indicator.

4.7 **Output Relay.** Under a number of conditions it is unsafe to have loudspeakers connected to the output of a large power amplifier. The relay control circuit is meant to accept signals from a number of sense circuits to disconnect the speakers when large transients could be passed to them, when dc is present at the output, or when the amplifier's own safe operating conditions have been exceeded.

The relay coil, K200, connects the loudspeakers when energized by Q201, which requires that Q200 be biased. When the unit is turned on, the network R204, C205/C214 delays turn-on of the relay for a few seconds so that turn-on transients from other components will not be passed to the loudspeakers. At turn-off, the small bias supply capacitor C204 discharges very rapidly into the negative supply through R204; the relay will open in much less than a tenth of a second after turn-off. C205/C214 store momentary "off" signals from other inputs, particularly the SOA monitor, to make their effects cumulative.

4.8 **Offset (dc) Detector.** Under normal conditions, the dc servo amplifier applies a correction voltage which holds the offset at the output within a few millivolts of zero. To protect both the loudspeakers and output devices from a gross dc error, the dc detector will open the relay when an offset of more than ±1.5 volts occurs. The filter R209–C207/C217 attenuates the ac signal, but the time constant has been chosen to make the detector also respond to very large, very low frequency signals, such as may arise from record warp, acoustic feedback, or FM station tuning, and which could damage the low frequency elements in a loudspeaker. When more than about 1.5 volts appears at any of the diodes D205 to D208, a current will flow in the string formed by one of the diodes and the two base-emitter junctions of Q202 and Q203, so that current will flow to the base of Q205, turning off Q200 to open the relay.

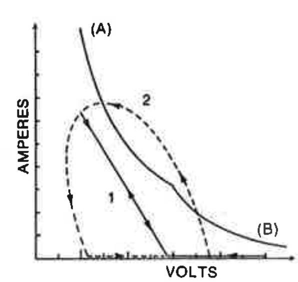


Fig. 4-9. Safe Operating Area.
1 - Resistive Load
2 - Reactive Load

4.9 **SOA Protection Circuits.** The output devices must be monitored for safe operating voltages and currents. When an output transistor is driving a resistive load, the current and collector-emitter voltage vary in a predictable, inverse way, so that the dissipation of the transistor remains within the boundary defined by A–B in Fig. 4-9. On the non-conducting half cycle, the voltage across the transistor is high, but the current is zero, as shown by line 1 in the Figure. But with a reactive load—and especially with the complicated reactances many loudspeakers present—high out-of-phase currents may be called for at times when the collector-emitter voltage is also high (line 2). These combinations of current and voltage may lie well outside the transistor's safe operating area, even at less than maximum power levels and even with very conservative overall design; and some mechanism must be provided to shut the amplifier down before burn-out of the transistors occurs.

An electronic SOA circuit, incorporated in the amplifier, can respond instantaneously to overload conditions by removing drive current from the output stage, but it inevitably impairs the sonic quality of the amplifier, adding spurious signals under clipping or very brief overload conditions (which can occur even at relatively low average levels when the program source has a wide dynamic range). Fortunately, transistor burn-out does not happen instantaneously, and there is no need to shut off the amplifier at once. The SOA curve shown above is in reality a family of curves related by a third, time dimension; and for short impulses, there is considerably more safe operating area available. Thus, safe area protection by means of an output relay external to the amplifier circuit avoids interfering with the amplifier's operation, and it can take advantage of the temporal dimension of the SOA curve by modeling the temporal as well as the current and voltage parameters.

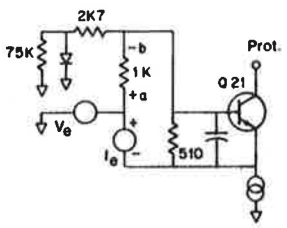


Fig. 4-10. SOA Sense Loops.

The SOA detector consists of two transistors, placed on both sides of the output stage. They function as simple voltage threshold detectors, and when the threshold is exceeded, they allow current to flow from the base of the SOA switch, Q204.

The condition which turns on either Q21 or Q22 is a combination of excessive current and voltage on either side of the output stage. When a large current flows in the 0.47 ohm emitter resistor, R39 (or R41), the voltage produced at the 510 ohm resistor (see Fig. 4-10) may turn on the protection transistor. At low collector-emitter voltages—when the emitter is close to the supply rail voltage—this current can safely be quite high. However, when the voltage from the supply rail to the emitter is high, a smaller current can be dangerous.

Instead of multiplying I_c and V_{ce} sense components, the circuit creates a threshold voltage proportional to V_{ce}, and then turns on if the emitter current passes this threshold. The voltage present at the output devices' emitters (V_e) passes through the 1 kohm resistor to ground (or through the 75 kohm resistor to the opposite SOA circuit, to approximate the hyperbolic SOA curve). Hence, point -b is at a negative voltage relative to +a. Before a positive current can flow in the current sense loop through the 510 ohm resistor, the I_e voltage must exceed this. When the emitter is close to the supply rail, the voltage on the 1 kohm resistor is large, and a lot of emitter current can be drawn before a positive voltage appears across the base-emitter junction of Q21. When V_e is near zero, or negative, a much smaller current sense voltage will suffice to turn on the protection sense function. C11 and C12 integrate the current and voltage sense impulses to model the time dimension of the SOA surface.

4.10 Load Impedance Detector. Above a certain ratio of current to voltage supplied to the left channel loudspeaker, the Load Impedance Adjust indicator shows that a low impedance load is present at the left channel, advising the user to switch the Load Impedance Adjustment switch to the lower range. This circuit consists of a reference voltage source, voltage and current detectors, and a comparator. The reference voltage source, R218, charges capacitor C215 to a certain trim voltage (about 0.1 V), set in the load-impedance set-up procedure (see Section 10.8). The voltage detector, D213, presents the positive half-cycle of the output (divided by 10) to the inverting input of the comparator. The current detector, Q206, sends a current equal to 0.001 of the current being supplied to the positive side of the output stage through the 0.1 ohm resistor, R223, to R227. Hence, the voltage developed on R227 is 0.001 × 560 or about 0.5 volts per ampere of current in the output stage during the positive half-cycle. A 5 ohm load, drawing (in amperes) 0.2 times the applied voltage,

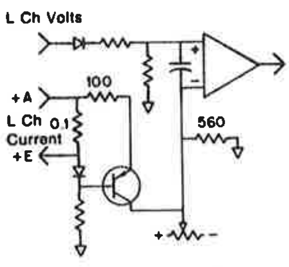
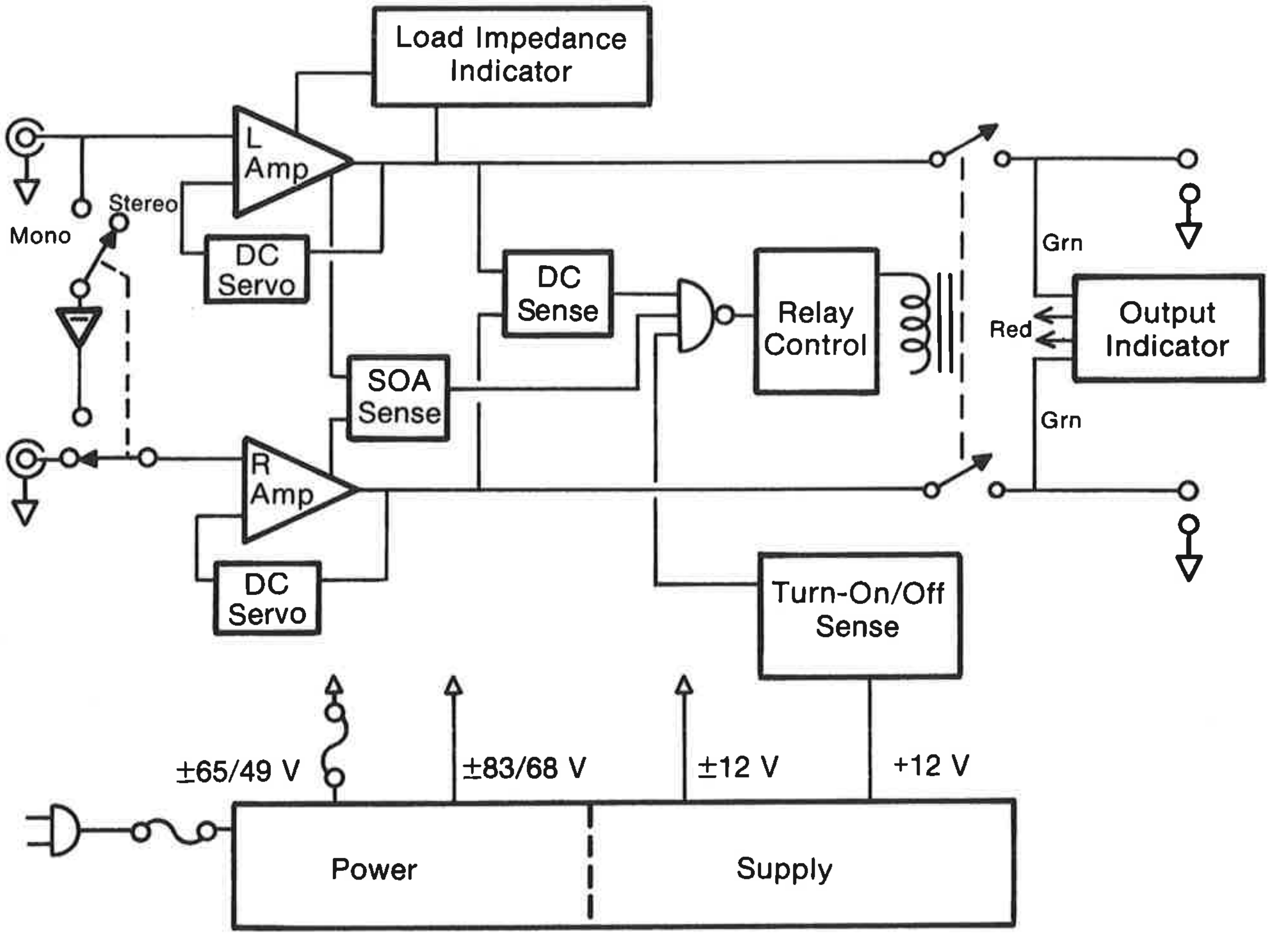


Fig. 4-11. V/I Comparator.

will therefore place a voltage equal to the output voltage on the non-inverting input. A lower impedance load will impose an even larger voltage. As soon as the current-derived voltage exceeds the loudspeaker input voltage plus the small trim voltage on C215, the comparator swings from negative to positive, signalling the presence of a low impedance load. If, in addition, the Load Impedance Adjustment switch is in the 8 ohm position, current will be supplied to the base of Q302 to light the Load Impedance Adjust indicator.

5 Block Diagram



6 Mechanical Disassembly

6.1 Tools:

Number 2 Phillips screwdriver (magnetized)
Number 1 Phillips screwdriver (magnetized)
3/16 in. screwdriver, long shaft
1/16 in. hex driver
1/4 in. ignition wrench
11/32 hex driver

6.2 Top Cover and Front Panel. The front panel will never need to be removed separately, except for replacement. Instead, the top cover and front panel should be removed from the amplifier as a unit. Remove the flat 4–40 × 3/8 flat head cap screw marked A in Fig. 6-1 with a 1/16 in. hex driver. Remove the two no. 6 sheet metal screws at each end of the amplifier (not shown) with a no. 2 Phillips driver and the seven no. 6 sheet metal screws marked A in Fig. 6-2. To avoid damage to the LED power indicator, lift the *rear* of the top cover about half an inch, then lift the cover off. Before proceeding further,

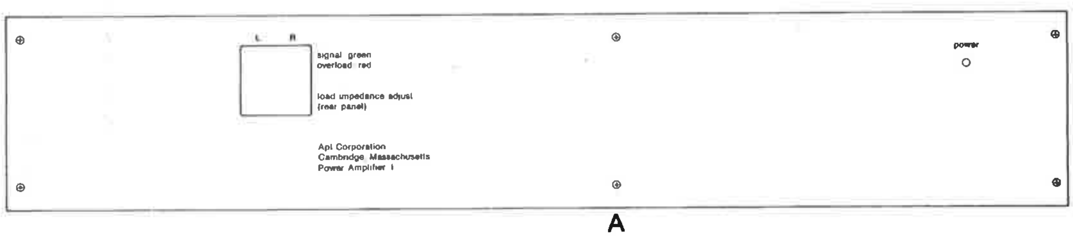


Fig. 6-1. Front Panel.

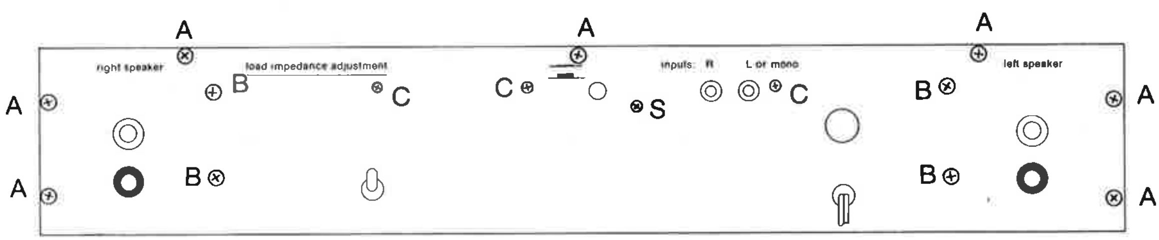


Fig. 6-2. Back Panel.

w/ 120 VAC IN

Test Points	*8-16 ohms*	*2-4 ohms*
+A	+65 V	+49 V
-B	-65 V	-49 V
+C	+84.5 +83 V	+68 V +69
-D	-84.5 -83 V	-68 V -69

Noise and ripple should be as follows:

+A	60 mV p-p
-B	60 mV p-p
+C	1.4 V p-p
-D	1.4 V p-p
±12	0.5 V p-p

The positive side of the ±12 V supply will appear different from the negative side on the oscilloscope, showing alternating large and small sawtooth spikes because of the additional current drawn by the half-wave relay supply, D201, on the positive half-cycle.

7.4 **Amplifier Section.** In general, failure of a resistor or semiconductor will produce a dc offset, and identification of abnormal dc levels or currents is the most important troubleshooting procedure. The dc servo, IC1, should be disconnected at point X (Figs. 9-1 and 9-2) when you are checking dc levels.

7.4.1 *Test Voltages.* Proper operation of the differential input stage can be checked at test points E-G; of the push-pull voltage amplifier, at points -H-; and of the output stage (current amplifier) at points -J-. Points with single arrows on the schematic are voltages referred to ground. Test points written with dashes (-J-) indicate voltages not referred to ground but measured across the identified resistor.

E	-1.25 V
-F-	1.36 V
-G-	4.8 V
-H-	0.7 V
-J-	8 mV
O	±20 mV (servo connected)

7.4.2 *Faulty dc Offset at Output.* Unless the dc fault is obvious, you will need to separate the dc servo amplifier from the power amplifier's feedback loop. Disconnect the end of the 47 ohm resistor, R59, at point X (see Foil Overlays, Chapter 9). If a large offset (more than ±2 V) remains at the output, point O (before the relay), check the amplifier section. Check particularly the test points -F- and -H- for similar voltage drops: differing values at points -F-, the two 680 ohm collector resistors in the input section, may point to a fault in the differential amplifier or in the current mirror, Q5-Q6, while different values of -H-, the 360 ohm resistors in the cascode drivers, may indicate that excessive current is being drawn into one side or the other of the current amplifier section (e. g., because of an open collector at Q13 or Q14).

If the offset voltage does not remain, the servo section is the probable culprit. An originally observed offset of around ±12 volts may indicate that the servo IC (IC1) has an internal short, in which case a large positive or negative voltage will be seen at R59. The voltage at point Z, the overload detector output, should be within a few millivolts of zero. 'Battery effect' or dielectric absorption in a damaged electrolytic capacitor in the servo loop may also cause erratic offset.

7.4.3 *Faulty ac Performance*. First check the overload outputs, point Z. (These are also accessible on the display board, before the amplifier modules have been removed, at IC301, pin 5, for the left channel, and at IC300, pin 6, for the right channel.) The overload comparator, IC1, divides the output of the amplifier by exactly the amplifier's gain (within one percent) and compares it to the input. Hence distortion products above about 0.5 percent and other spurious signals will be immediately evident at the comparator output. (The one percent resistors in the comparator divider, R61 and R18, are chosen in production to keep the normal comparator output below the threshold of the overload indicator. If a feedback resistor must be replaced, you may need to try a few one percent parts until the right match is found. Replace *signal path* feedback resistors with *metal film* types only.)

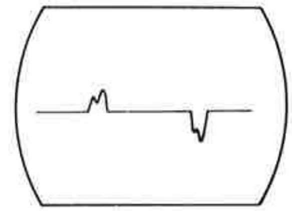


Fig. 7-1. Comparator Output at Clipping.

Apply a 2 kHz, 1.3 Vac signal to the amplifier input. Adjust to clipping level, which should occur at about 120 V peak-to-peak in the 8–16 ohm mode with no load, and at about 80 volts peak-to-peak with an 8 ohm load. Clipping should be perfectly symmetrical. Shorting the input to ground, *after* the 1 kohm input resistor, R1, to remove the input signal, you should observe no dc level 'bounce' when the signal is removed or restored. (Some signal generators will exhibit 'bounce' when their outputs are switched: shorting is the only way to preclude this.) The clipping products should appear at the comparator output, as in Fig. 7–1, without ringing or oscillation. As a further test for stability, place a 0.1 μF, 400 V mylar capacitor across the loudspeaker output jacks; apply a 25 kHz signal to the input at low level and raise the level gradually to clipping, watching both waveform and current drain for irregularities. Under hard clipping the output should appear as in Fig. 7–2. The less than critically damped ring evident here may appear whenever a reactive load is driven with a clipped, high frequency signal: it is an oscillation excited in the output inductor, L1, rather than an amplifier instability and may be ignored

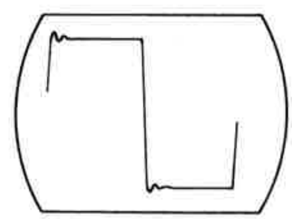


Fig. 7-2. 25 kHz Capacitive Load Test.

A number of factors can contribute to high frequency instability: faulty mounting of the bias transistor or Moxie device; an internal fault in either of these; faulty compensation capacitors, C4 or C9; insufficient gain or slewing capability (e. g., a fault in the current source, Q7, open collectors in the current amplifier section, or a failure in one of the cascode pairs, Q8–Q9 or Q11–Q12). (Consult Section 4.3 for a functional description of the amplifier section.)

Normally, you will be able to identify a fault discovered in an ac test by checking dc levels more carefully. A further, marginally useful test is to observe ac voltages and currents. Virtually no ac signal should be seen at points E or G; equal ac signals should be seen at points -F- and points -H-. Within the current amplifier, each of the base-emitter resistors should show the full sinusoidal waveform with the probe referred to ground, but you should find complementary half-waveforms (as in Fig. 4–6) of roughly equal amplitudes when looking at the currents in these resistors. (When the amplifier is driving a reactive load, current will be out of phase with voltage, and the waveform will be split at different points from those shown in Fig. 4–6.) Using a 350 mV, 1 kHz signal to obtain a 10 Vrms output into an 8 ohm load (28 V peak-to-peak), you should observe about 120 mV peak-to-peak across R34 and R43, around 170 mV peak-to-peak across R36 and R44, and about 0.5 V peak-to-peak across the output emitter resistors, test points -J-.

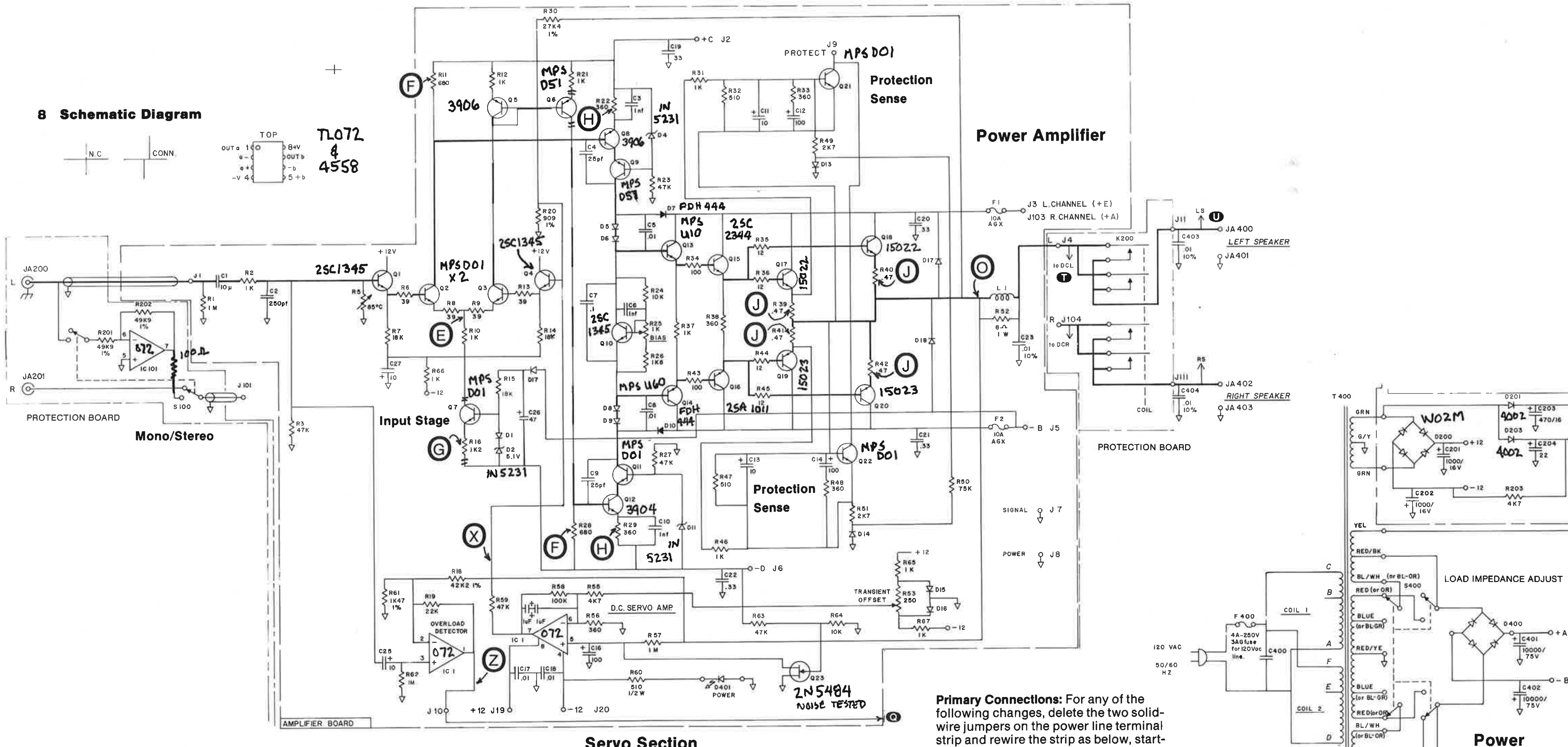
Excessive current drain, especially when no load is connected to the outputs, is a possible sign of high frequency oscillation, especially if dc voltages appear to be correct. If the leads of the Moxie device, R5, make contact with the can when the module is reassembled, a dangerous 2 MHz oscillation may arise. A defective Moxie device or contact between the bias transistor's leads and its mounting clamp may also cause this condition.

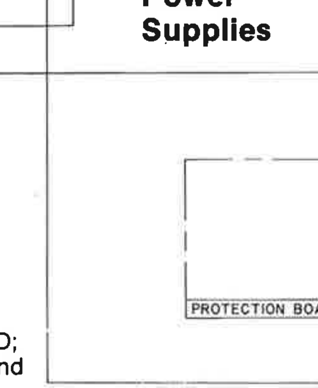
Note that the *only* chassis ground connection is at the input jacks on the protection circuit board.

7.5 **Relay and Protection Circuits.** Check the operation of the relay and its drive circuit by connecting the base of Q205, test point R, to the large ground foil in the center of the board through a 1 kohm resistor. The relay should open immediately (green signal indicators will go off) and close again in 3 to 10 seconds after the short is removed. (C205/C214–R204 integrate all protection impulses, and cause the protection relay to close a second time more slowly than the first.)

Protection circuits may be isolated from the relay circuit by removing the lead at point P on the Power Amp boards. Consult Sections 4.5, 4.6, and 4.10 for operation of the Signal/Overload and Load Impedance indicators. Test point M, the non-inverting input to the V/I Comparator, should show from 20 to 150 mV under idle conditions and should show the positive-going half-cycle of an output signal at a low amplitude when a load is present.

8 Schematic Diagram





Servo Section

Schematic Notes: Only the left channel power amplifier is shown—the right channel amplifier is identical.

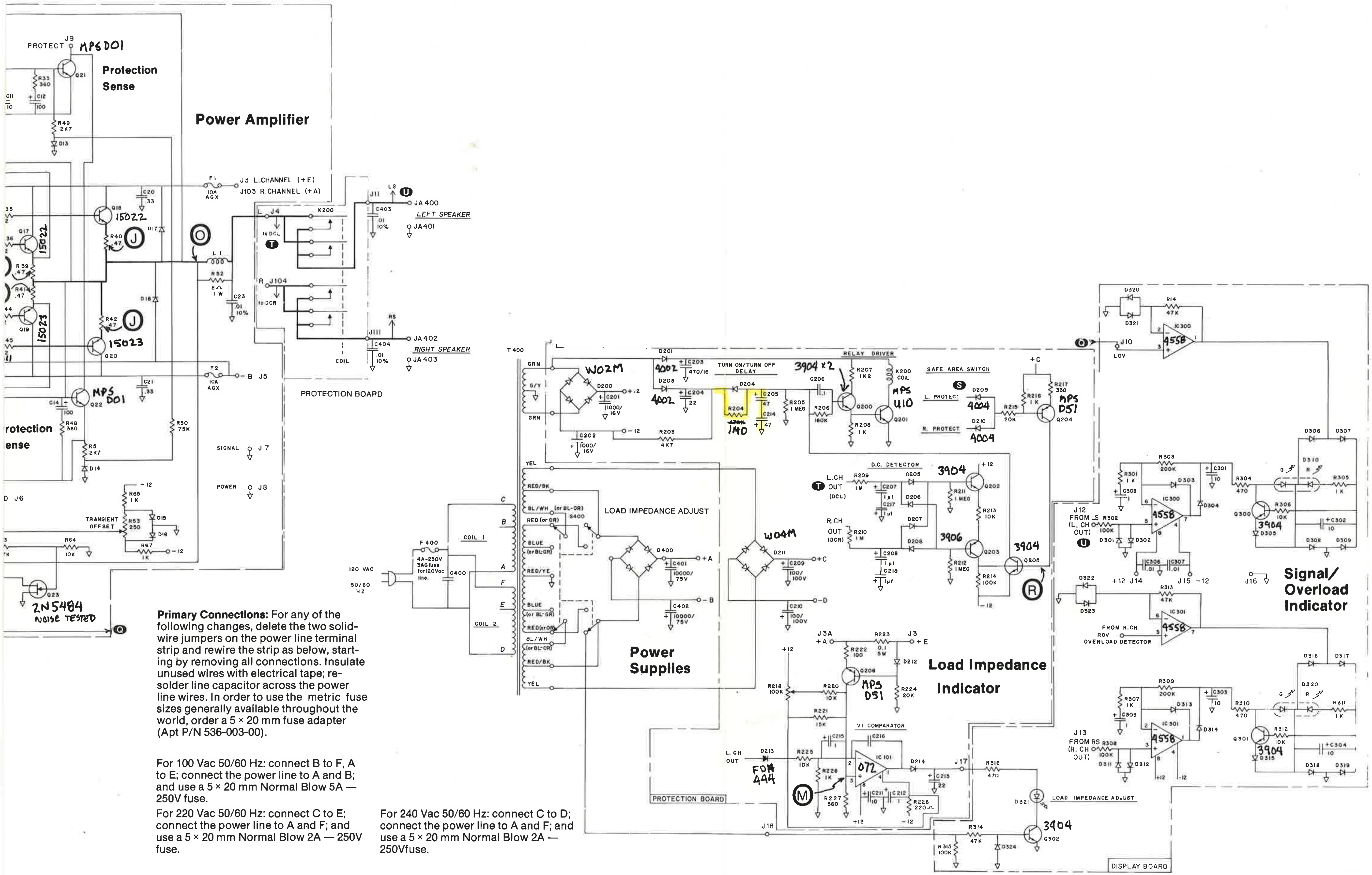
Parts Numbering: Parts numbered 1-99 are on the left channel amplifier circuit board; parts numbered 100-199 are on the right channel amplifier circuit board; parts numbered 200-299 are on the protection circuit board; parts numbered 300-399 are on the display circuit board; and parts numbered 400-499 are on the chassis.

Primary Connections: For any of the following changes, delete the two solid-wire jumpers on the power line terminal strip and rewire the strip as below, starting by removing all connections. Insulate unused wires with electrical tape; resolder line capacitor across the power line wires. In order to use the metric fuse sizes generally available throughout the world, order a 5 × 20 mm fuse adapter (Apt P/N 536-003-00).

For 100 Vac 50/60 Hz: connect B to F, A to E; connect the power line to A and B; and use a 5 × 20 mm Normal Blow 5A — 250V fuse.

For 220 Vac 50/60 Hz: connect C to E; connect the power line to A and F; and use a 5 × 20 mm Normal Blow 2A — 250V fuse.

For 240 Vac 50/60 Hz: connect C to D; connect the power line to A and F; and use a 5 × 20 mm Normal Blow 2A — 250Vfuse.



Primary Connections: For any of the following changes, delete the two solid-wire jumpers on the power line terminal strip and rewire the strip as below, starting by removing all connections. Insulate unused wires with electrical tape; re-solder line capacitor across the power line wires. In order to use the metric fuse sizes generally available throughout the world, order a 5 × 20 mm fuse adapter (Apt P/N 536-003-00).

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For 240 Vac 50/60 Hz: connect C to D; connect the power line to A and F; and use a 5 × 20 mm Normal Blow 2A — 250Vfuse.

9 Foil Overlays

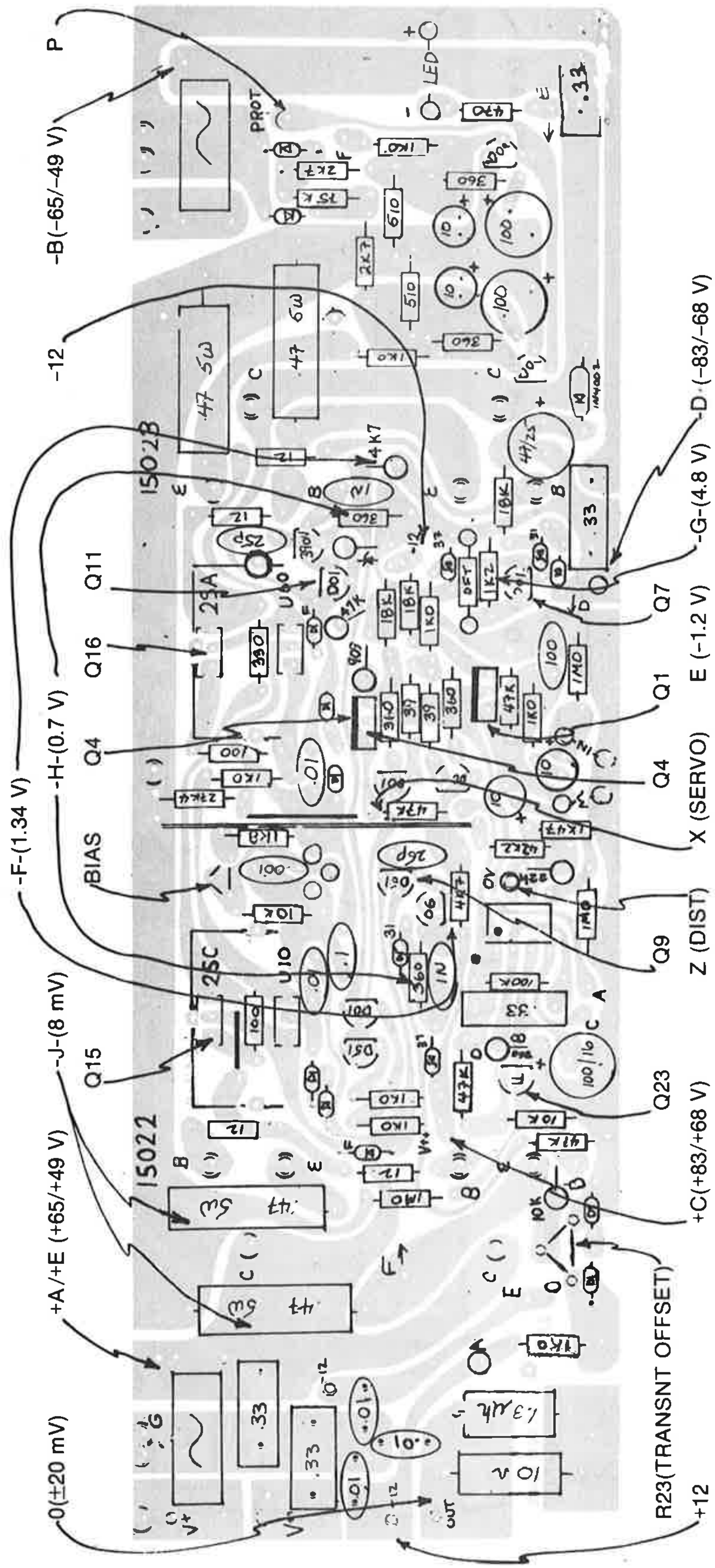


Fig. 9-1. Power Amp Circuit Board, Component Side, Rev 1-2 (Early Units Only).

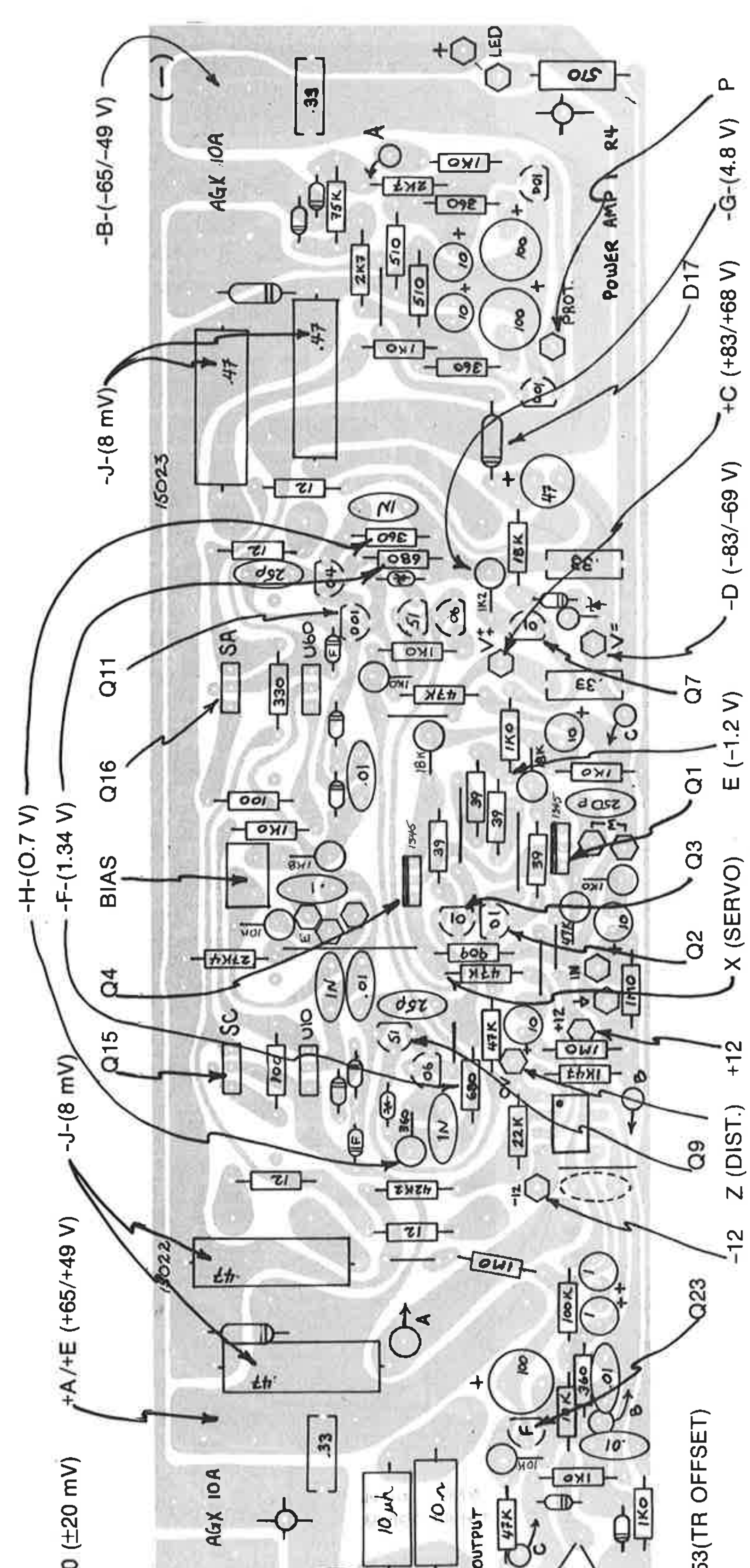


Fig. 9-2. Power Amp Circuit Board, Component Side, Rev 3-6. #6 -9
(Minor layout variations will be found.)

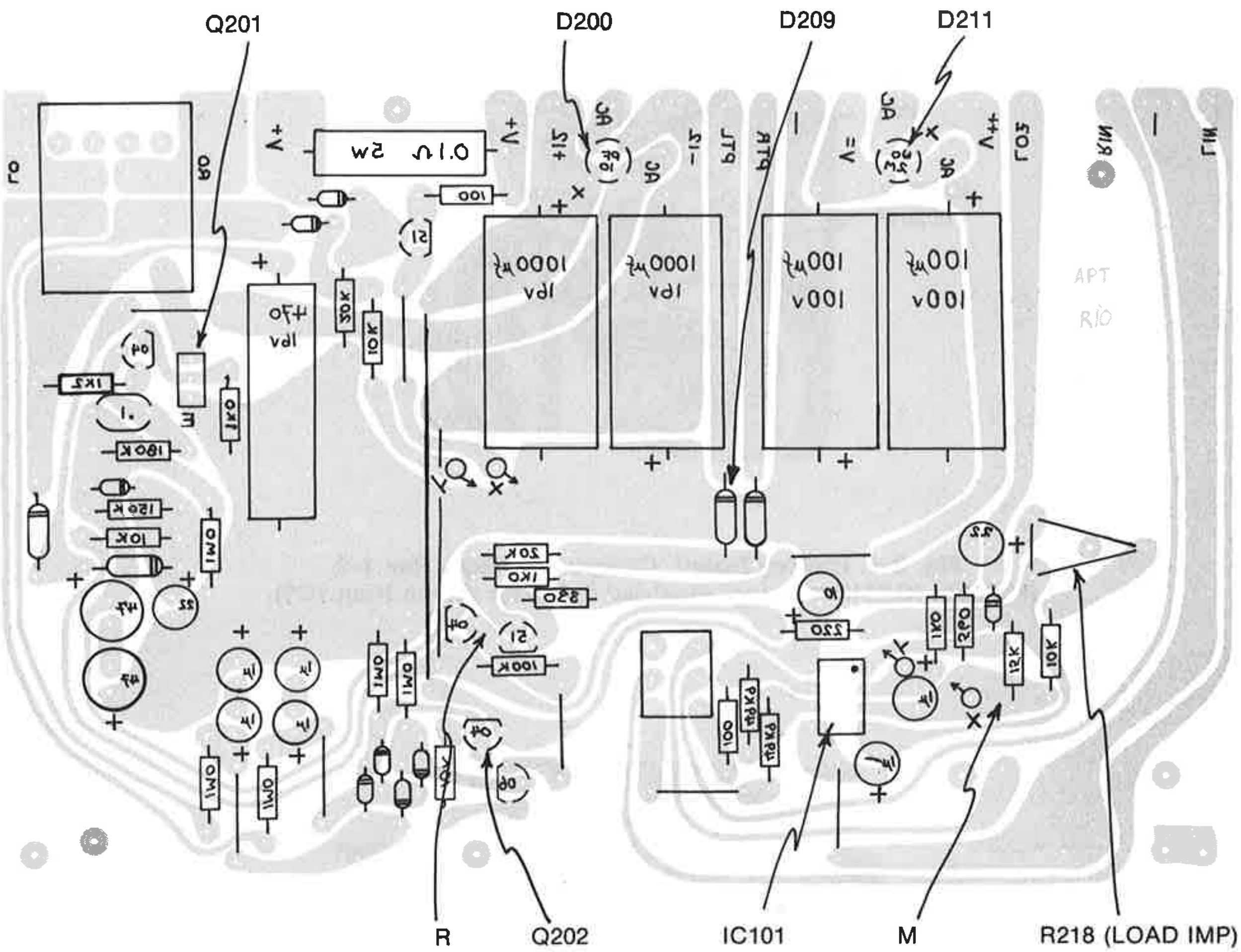


Fig. 9-3. Protection Circuit Board.

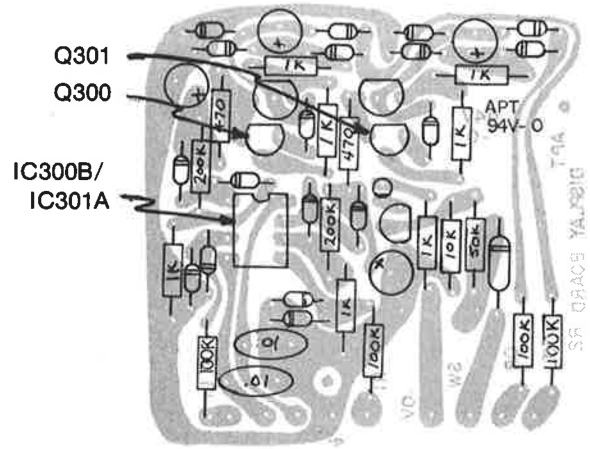


Fig. 9-4. Display Board, Component Side, Rev 1-2 (IC300A, IC301B missing: overload indicator driven from IC1).

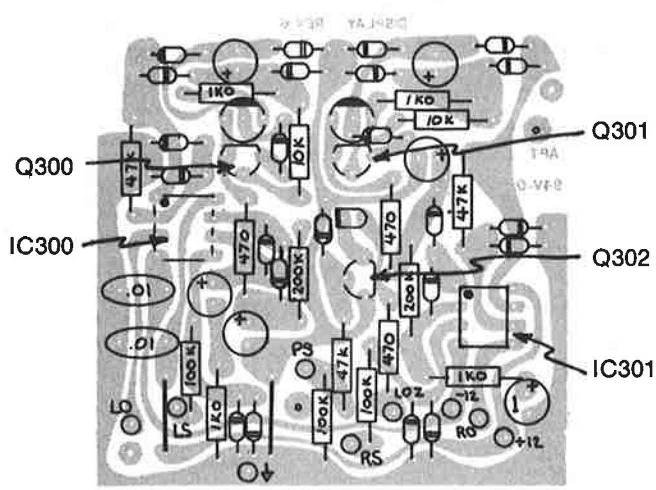


Fig. 9-5. Display Circuit Board, Component Side, Rev 3-6.

10 Final Test of Specifications

10.1 The following procedure should be performed after all repairs or adjustments have been completed to ensure that the amplifier meets its published specifications. It is similar to the procedure which generates the Final Test Report packed with the new unit.

10.2 **Test Set-Up and Instrumentation.** Reassemble the amplifier, except for the top cover. Check carefully for pinched wires and shorts in the leads to the Moxie devices and bias transistors. Switch the metered variable autotransformer to "off" and connect the amplifier's line cord. Discharge the main supply capacitors with an 8 ohm load resistor, connecting the hot side of the fuse (blue or red supply wire) to ground through the resistor (see Fig. 6-5.)

Set the Stereo/Mono switch to the Mono (out) position. Set the Load Impedance Adjustment switch to the 8-16 ohm (up) position. Attach a voltmeter probe to test point O (output before relay) of the left channel, and connect the ac output of the voltmeter to one input of the oscilloscope. Connect a cable from the signal generator (oscillator) to the left channel input (the inversion of the oscillator's output will be sent to the right channel through the bridging connection). Do not connect load resistors at this stage.

10.3 **Power and Relay.** Set the autotransformer to 0.0 Vac and turn it on. Apply a 2 kHz, 0.325 Vac signal to the input and bring up the autotransformer voltage slowly. The current will peak each time the voltage is advanced, settling to less than 300 mA. If it remains above this level, and the amplifier otherwise appears to be working normally, reduce the bias current by turning the bias trimmers (see Fig. 9-1 or Fig. 9-2) counterclockwise. At some point between 30 and 60 Vac the signal should appear abruptly. By 70 Vac all signs of oscillation should be gone; by 90 Vac the relay should pull in.

10.4 **Overload and Signal Lights.** Reduce the 2 kHz signal to about 0.5 mV and note that the green signal lights begin to go out around this level (15 mV output level). Alternatively, turn off the signal generator, note that the indicators extinguish, then apply a 20 Hz signal at a low level, and observe a 20 Hz flicker in both indicators.

Raise the 2 kHz input signal voltage to about 1.3 Vac; you should see about 39 Vac at the output or about 110 volts peak-to-peak. Raise the level further, until the first overload (red) indicator lights. Check that the second overload light comes on within 0.5 dB of this level.

Reduce the output level to about 14 Vac (to avoid blowing a fuse) and apply a short to the left channel output jacks. The relay should open at once and click at intervals of one-half to one second, as the SOA circuit rediscovers the short and opens the relay again. Repeat for the right channel. Remove the signal.

10.5 **Transients.** Connect an oscilloscope probe directly to the left channel output jacks. Set the vertical gain to 100 mV per division, dc coupled. Turn the amplifier off; turn it on again. Observe that the offset voltage transient passed to the output jacks (*after* the relay) does not exceed

11 Options

11.1 **Alternate Line Voltages.** Maximum utility from the Apt 1 requires a good match to the prevailing mains voltage. The transformer has been wired for the voltage stamped on the rear panel near the line cord; if a different nominal voltage is required, the transformer connections should be changed according to Fig. 11–1.

Remove the top cover as described in Section 6.2. Remove the two number 8 nuts which anchor the transformer's fish paper cover at the right channel end. Raising the fish paper will reveal a four-lug terminal strip at the left channel end of the transformer. The lugs in Fig. 11–1 are numbered from the front of the amplifier to the rear. The 4.7 nF capacitor should always be connected from the hot to the neutral power input wire.

11.2 **Rack Mount.** The optional rack-mounting front panel kit consists of a replacement front panel with exterior dimensions conforming to 19 inch (EIA) American standards.

The original front panel is held on by six 4–40 flat head socket screws. The replacement rack-mount panel duplicates the dimensions of the amplifier for hole layout, but its outside dimensions are expanded to 19 × 3-15/32 in. (usually called 3½ in.). Since not all racks are built to this standard, you may encounter some difficulty with mounting in foreign-made racks.

11.2.1 *Removal of the Original Panel and Top Cover.* Remove the top cover according to section 6.2 (Figs. 6–1 and 6–2). Remove the front panel hex socket screws with the 1/16 in. socket hex key supplied and set them aside. Press the LED (Power indicator) with your thumb, and ease the panel off.

11.2.2 *Installation of Rack Mount Panel.* In the rack-mounted configuration, the full weight of the power transformer must be borne at the two forward corners of the case; for this reason, reinforcing brackets have been provided with the rack-mount kit, which require that you drill out the threads in the holes at either end of the top cover.

1. Enlarge the four holes at either end of the top cover by drilling through them with the supplied drill bit and smooth them off with a small file. Do not enlarge the top center mounting hole.

Warning: Drill bits may shatter. Always wear proper eye protection in the vicinity of their use.

2. Mount the handles to the front panel with the 6–32 × 1/2 hex socket screws, using the allen key supplied.

3. Attach the rack panel to the top cover first with the 4–40 × 3/8 hex screw, inserting it in the undrilled center hole of the top cover. Then place one of the mounting plates behind the mounting flange at one end of the top cover so that the tapped holes align with the holes in the front panel. Insert two 6–40 × 1/2 hex socket screws through the front panel into the mounting plate; screw them in but do not yet tighten. Repeat for the other end.

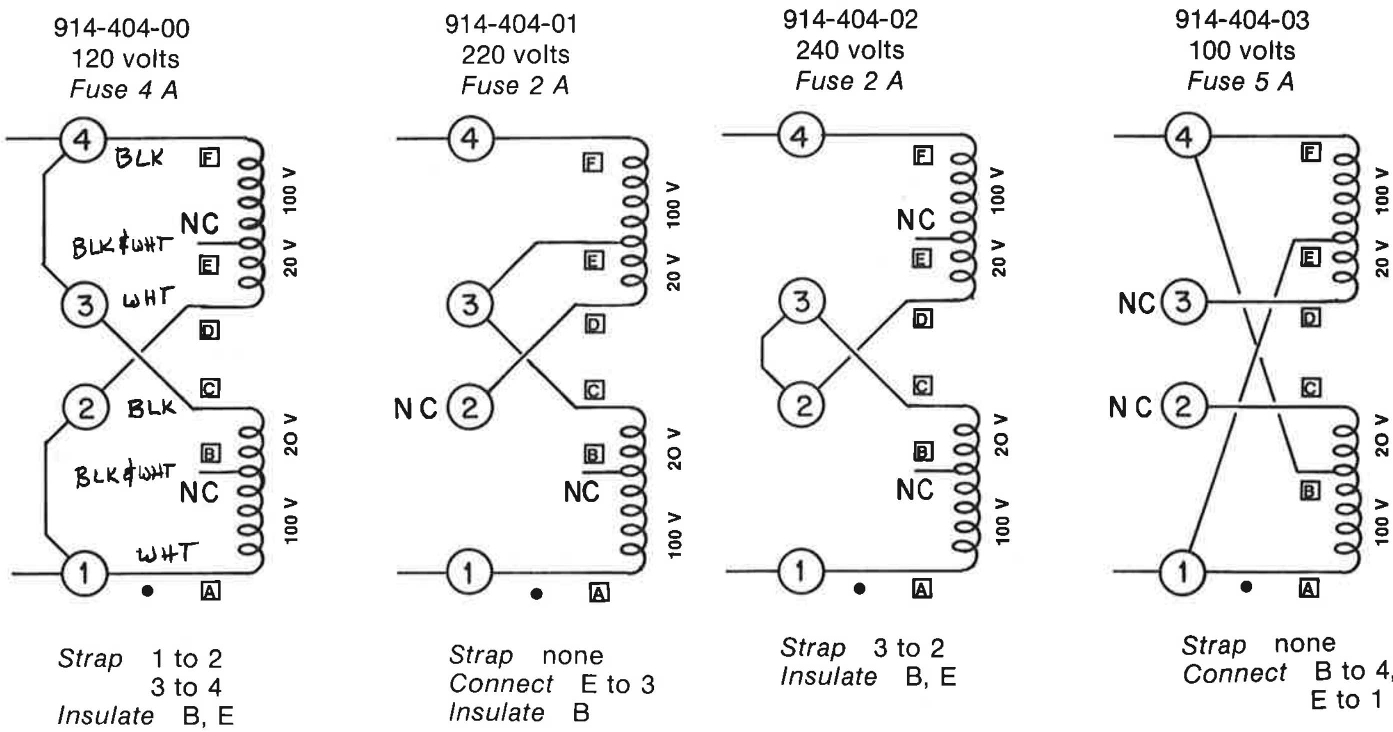
4. Peel off the protective paper from the back side of the display window in the front panel.

5. Place the top cover on the amplifier and secure it with the lower center hex socket screw. Replace the four sheet metal screws in the sides of the cover and the seven remaining screws in the rear. Then tighten all of the hex socket screws in the front panel.

6. Peel off the the protective paper from the front of the display window.

7. Use the 10-32 × 1/2 in. machine screws and corresponding fibre washers to mount the amplifier in the 19 in. rack.

8. Since the rack installation adds an additional ground connection to the amplifier, ground loops may be formed in the system, injecting hum. If this happens, consult the section on hum in the *Owner's Manual*.



O – Solder lugs

□ – Primary leads

Connect primary power through fuse to hot terminal (4). Neutral side of ac mains connects to (1).

Fig. 11-1. Transformer Connections.

723-001-00	Diode, power, 1N4002
723-002-00	Diode, power, 1N4004
724-001-00	Diode bridge, Vearo VE27 (WO2M)
724-003-00	Diode bridge, MDA 3502
724-004-00	Diode bridge, 400 PIV, WO4M
725-002-00	Diode, zener, 5.1 V, 1N5231A
731-001-00	Transistor, npn, Hitachi 2SC1345E
731-002-00	Transistor, npn, 2N3904
731-003-00	Transistor, npn, Motorola MPS-D01
732-002-00	Transistor, pnp, 2N3906
732-004-00	Transistor, pnp, Motorola MPS-D51
733-001-00	Transistor, npn, Motorola MPS-U10
733-002-00	Transistor, npn, Panasonic 2SC2592/2SC 3344
733-003-00	Power transistor, npn, Motorola MJ15022
734-001-00	Transistor, pnp, Motorola MPS-U60
734-002-00	Transistor, pnp, Panasonic 2SA1112/2SA1011
734-003-00	Power transistor, pnp, Motorola MJ15023
741-001-00	Field effect transistor, 2N5484
751-001-00	Integrated circuit, Texas Instr. TL072CP
751-002-00	Integrated circuit, Raytheon RC4558CP
913-402-00	Light emitting diode (Power)
781-002-00	Light emitting diode, TIL216-1 (Load Imped.)
781-003-00	Light emitting diode, Stanley SPRG 4311, red/green

12.5 Passive Components.

555-125-00	Trimpot, 250 ohm, standup (offset)
555-210-00	Trimpot, 1 kohm, standup, sealed (bias)
555-410-00	Trimpot, 100 kohm, laydown (Load Imp)
610-001-00	Inductor, 1.3 μH (L1)
640-004-00	Fuse, 4 A, 250 V, AGC
640-005-00	Fuse, 10 A, AGX
673-190-01	Resistor, 909 ohm, 1% metal film
673-214-01	Resistor, 1K47 ohm, 1% metal film
673-327-03	Resistor, 27.4 kohm, 2% metal film
673-342-01	Resistor, 42K2, 1% metal film

673-349-01	Resistor, 49K9, 1% metal film
674-151-01	Resistor, 510 ohm, 5%, ½W
674-810-01	Resistor, 0.10 ohm, 10%, 5W
674-847-01	Resistor, 0.47 ohm, 10%, 5W
674-982-01	Resistor, 8.2 ohm, 10%, 1W
770-001-00	Temperature sensitive resistor, ca. 150 kohm at 20°C, Moxie TS3-85-B3
Resistors	*All other resistors may be ordered by specifying the following:* R number from schematic (e. g., R91) Value Axial (flat lying) or p. c. (standup) Tolerance (usually 5%)
681-710-04	Capacitor, electr., 100 μF, 100 V
681-810-02	Capacitor, electr. 1000 μF, 16 V
681-910-01	Capacitor, electr. 10,000 μF 75 V
682-433-01	Capacitor, mylar, 0.33 μF
683-025-01	Capacitor, ceramic, 25 pF, 10%
683-125-01	Capacitor, ceramic, 250 pF, 10%
683-310-01	Capacitor, ceramic, 0.01 μF
683-247-01	Capacitor, ceramic, 4.7 nF, U.L. and CSA listed C400
Capacitors	*All other capacitors may be ordered by specifying:* C number from schematic (e. g., C21) Value in μF or pF Electrolytic, mylar film, or ceramic disc Axial (flat) or p. c. (standup)

12.6 Switches, Relays, Assemblies.

914-001-01	Display circuit board, tested
914-002-01	Protection circuit board, tested
914-003-01	Power transformer assembly, with terminal strip and 0.0047 line capacitor
914-004-01	Left amplifier module with heat sinks, tested
914-005-01	Right amplifier module with heat sinks, tested
431-002-00	Pushbutton, black
551-004-00	Switch, slide, 2P2T
551-008-00	Switch, toggle, 4P2T, 10 A
712-002-00	Relay, 12 V, 4P2T
511-003-00	Accessory: phono shorting plugs (2)
694-003-01	Accessory: dual phono cable with plugs, 3 ft. (1 meter) length
694-004-01	Accessory: single shielded cable, with plugs, 30 ft. (9 meters) length

13 Appendixes

13.A Voltmeter Preamplifier Project.

The noise test described in Section 8.10 requires reliable measurement of levels as low as -100 dBV (10 μV). In addition, psychoacoustically meaningful measurement requires a weighting function that takes into account the varying sensitivity of the ear with frequency and energy distribution. For the purpose of noise measurement, a good approximation to the ear's response is provided by the ANSI A-weighting characteristic. While this curve does show some considerable deviations from the curve one would obtain by following closely the low-level sensitivity of the ear, particularly in the area around 3 kHz and at low frequencies, a closer fit would require an electrically more complex filter. And although the A-weighting function overestimates the ear's sensitivity to noise at low frequencies, it does give a useful approximation to its sensitivity to sinusoidal signals, such as power supply hum and its harmonics.

For ac voltmeters which are not capable of reading low voltages accurately, or which do not incorporate an A-weighting filter, the circuit described here will add the facilities needed for accurate noise measurements.

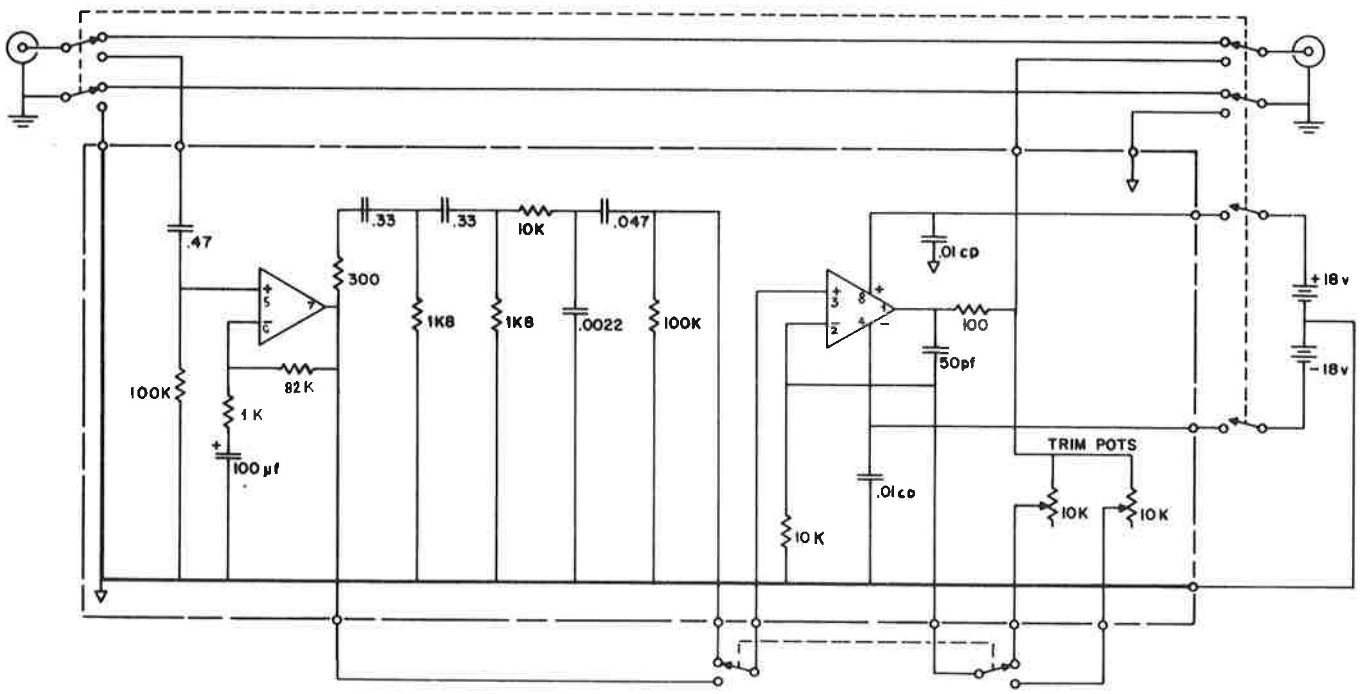


Fig. 13-1. Voltmeter Preamplifier Schematic.

The circuit shown in Fig. 13-1 uses one half of a TL072 operational amplifier to raise the input signal level 40 dB. It drives the weighting filter from the required low impedance (see Howard M. Tremaine, *Audio Cyclopedia* [Indianapolis: Howard W. Sams, 1969], p. 260). The second half of the TL072 presents the high impedance required at the output of the filter and allows 0 to 6 dB of gain to overcome the filter's insertion loss. A 4PDT switch permits the input signal to be fed directly through to the meter when additional gain is not needed and at the same time disconnects the battery power supply. Because the TL072 draws only about 3 mA, the preamp can be expected to run more than 100 hours on a set of batteries, and the battery power supply ensures freedom from ground loops, which can be quite troublesome at such low levels. A second switch allows the A-weighting filter to be removed from the circuit and inserts a second feedback resistor to maintain the overall gain of 40 dB.

Construct the filter and amplifier with careful attention to separtion of inputs and outputs. The 0.01 ceramic disc capacitors must be located as close as possible to pins 8 and 4, and must have short leads to ground. The unit must be installed in a grounded, shielded case to prevent hum.

Calibrate the "unweighted" gain by feeding a 10 mV (–40 dBV), 1 kHz signal directly to your voltmeter. Trim the oscillator level to read exactly -40 dBV on the voltmeter. Set the meter to its 1 V range, insert the meter preamp between the voltmeter and the oscillator, and adjust the preamp's "unweighted" gain trimmer to give a reading of exactly 0 dB on the voltmeter. If you are using an rms voltmeter, you may calibrate the "A-weighted" trimmer in the same way. If yours is an *averaging* meter, set the gain of the "A-weighted" trimmer to give a reading of +1.1 dBV. This will correct for the error incurred in measuring noise with an averaging rectifier (see Section 3.3.2); it will mean, of course, that sine waves measured through the A-weighting filter (not that one would have much cause to do this) will read 1.1 dB higher than their true (rms) A-weighted values.

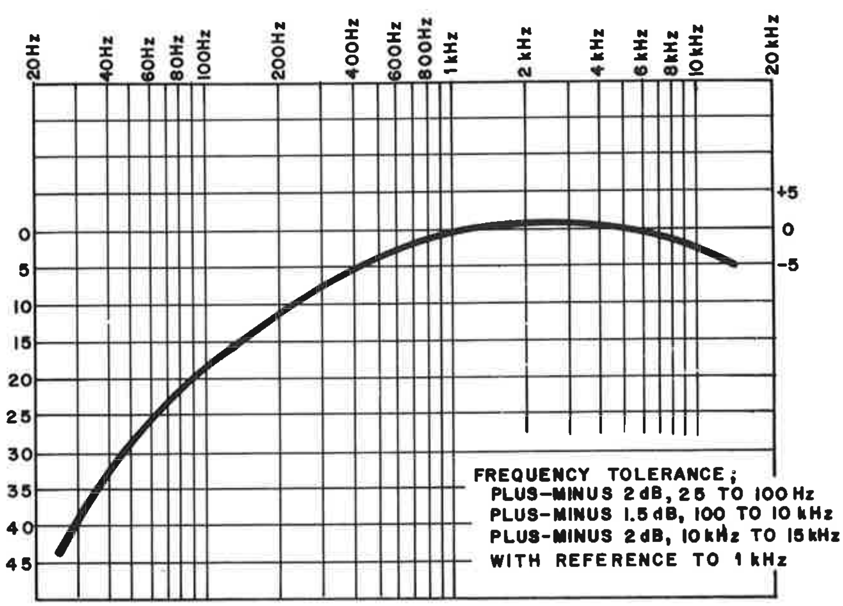


Fig. 13-2. A-Weighting Curve.