A HIGH EFFICIENCY – HIGH QUALITY AUDIO FREQUENCY POWER AMPLIFIER*

Alexander B. Bereskin **
The Baldwin Co.
Cincinnati, Ohio

The development of the power amplifier discussed in this paper was undertaken with the objective of providing a large amount of good quality audio power in a small package at relatively low cost. The size of an amplifier package and its cost are dependent largely on the efficiency of operation and the power sensitivity of the output stage. Beam power tubes satisfy these two requirements more readily than triodes. The quality of the audio power is improved by the use of push-pull operation and large amounts of feedback. The use of a suitable amount and type of feedback with beam power tubes eliminates the advantage of inherent low output impedance obtained with triodes.

In Class A or Class AB operation the dynamic characteristics of the two tubes may be matched to obtain a reasonably linear characteristic, and while feedback is desirable its use may sometimes be avoided. In Class B operation it is not possible to match dynamic characteristics to get linear operation, and therefore a large amount of feedback must always be used. The problems encountered in Class B operation are considerably more severe than those occurring in Class A and Class AB operation, and the question naturally arises as to whether this type of operation is worth the extra effort required. This question can be resolved rather conclusively by the following example:

Two 6L6 tubes operating Class A_1 push-pull with 30% efficiency, and having an allowable total plate dissipation of 40 watts, would be capable of developing an output power of $\frac{30}{70} \times 40 = 17.1$ watts.

The same two 6L6 tubes operating Class B_1, push-pull with 60% efficiency would be capable of developing $\frac{60}{40} \times 40 = 60$ watts. The possibility of getting 60 watts of output power in Class *B* operation for the same investment in power tubes which produced only 17.1 watts in Class *A* operation provides a real incentive to solve the severe problems encountered in Class *B* operation. The advantage of high efficiency operation also reflects advantageously on the size of the power supply necessary to operate the power amplifier.

The basic power amplifier circuit shown in Fig. 1 is capable of employing large amounts of feedback with good stability. This circuit has been operated with 36 *db* of feedback without showing any trace of instability. However, since the driving voltage required under these conditions is too great, the circuit is normally used with only 24 *db* of feedback. In this circuit the 12AX7 tube is used as a phase inverter – amplifier – driver stage. It is direct-coupled to the beam power output tubes and the

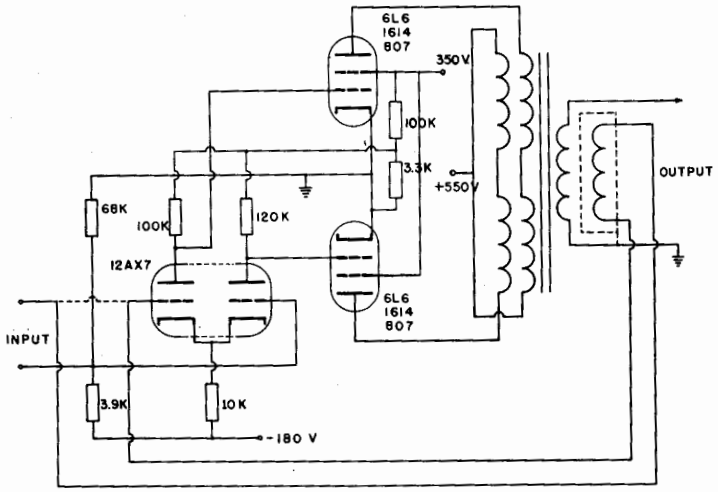


Fig. 1 – Basic Bereskin power amplifier circuit.

operation is essentially Class B_1 since the high impedance driver stage is incapable of driving the power tube grids positive. An added advantage of the direct-coupled driver is that it eliminates the possibility of blocking due to excessive input signal. The two output-tube cathodes are returned to ground and therefore any combination of screen and plate supply voltage may be used. The values shown on the diagram will keep both the screen and plate dissipation below the rated values for full signal Class B_1 operation with either the 1614 or the 807 tubes. The bias for the output tubes is supplied by the direct-coupled phase inverter – amplifier – driver and is normally adjusted to produce a zero signal plate current of about 15 ma per tube. The high value of cathode resistance employed makes the driver circuit fundamentally stable. A test with 6 different 1614 tubes and 12 different 12AX7 tubes, of different manufacturers and chosen at random, produced a zero-signal plate current variation ranging from 10 to 25 ma per tube. The full signal operation was substantially independent of the choice of 12AX7 and beam power output tubes. The feedback winding used is electrostatically shielded from the secondary but very closely coupled to it. The elimination of the electrostatic shield greatly reduces the amount of feedback that can be used successfully.

*Manuscript received February 2, 1954. Paper delivered at Cincinnati IRE Section, November 17, 1953.

** Professor of Electrical Engineering, University of Cincinnati.

One of the major problems associated with Class *B* operation is due to the energy stored in the leakage reactance between the two primary windings. A. P. Sah[1] showed how this stored energy gave rise to a conduction transfer notch which must be eliminated before Class *B* operation can be used successfully. Several different winding schemes will reduce the leakage reactance below the critical value but the most successful one is that of using a bifilar winding for the two primary sections.

The use of bifilar windings introduces new problems which were not previously important. One of these problems is that appreciable voltage may exist between the adjacent wires of a bifilar winding and sufficient insulation must be provided to withstand this voltage. Adequately insulated wires are now available commercially so this problem is seldom a serious one. A second and more significant problem is that appreciable capacitance exists between the adjacent wires in a bifilar winding and that charging current must be supplied to this capacitance before any voltage can be developed between the wires. This charging current must usually be supplied through the output stage tubes and is one of the major factors limiting the high frequency power-delivering capacity of an amplifier.

This problem can be understood more readily by examining the circuit of Fig. 2A. In this circuit the bifilar primary has been separated into two sections and the

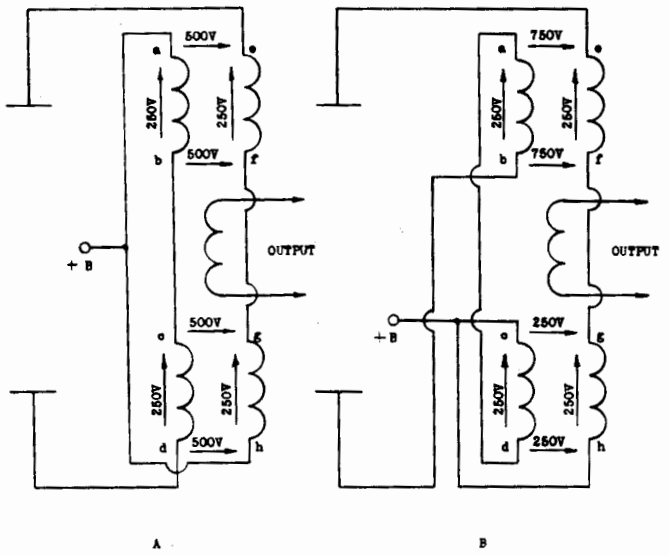


Fig. 2 – Primary interwinding voltage relations.

secondary has been sandwiched between them to keep the leakage reactance between the primaries and secondary at a low value. The four primary sections have been symmetrically interconnected and there is negligible dc voltage between these windings. If a peak signal voltage of 250 v is assumed on each of the primary sections, as shown by the vertical arrows, it is observed that this will give rise to a peak signal voltage of 500 v between all adjacent points on the bifilar winding as shown by the horizontal arrows. The undesirable feature here is that before this voltage can appear between the two primary windings the interwinding capacitance must be suitably charged, and the charging current must flow through one of the two tubes. An experimental transformer of this type, wound with adjacent-wire layer bifilar winding, using No. 28 Heavy Formvar wire, was found to have a capacitance of 0.045 microfarads between the two primary windings. A peak charging current of 1.5 amperes is required to charge this capacitance at 10KC with a sine wave voltage of 500 volts peak across the primaries. An ordinary tube used in a circuit of this type would normally supply only about 0.30 amperes peak so that at 2 kc the peak current capacity of the tube would be required to furnish the charging current for the interwinding capacitance. The amplifier's power-delivering capacity to a resistance load would therefore be down 3 db at 2 kc.

A natural step at this point is to consider the possibility of using a different interconnection of the primary sections to reduce the interwinding voltages, thereby reducing the charging current required for the interwinding capacitance. Fig. 2B shows a different connection of the primary sections. The voltage between the lower sections of the bifilar winding has been reduced to 250 v, but the voltage between the upper sections has been correspondingly increased to 750 v. The total interwinding charging current is the same as before but the insulation burden on the upper section is greater than it was before. As a matter of fact, further sectionalization and reconnection of the primary does not reduce the charging current problem and will, in most cases, increase the burden on the insulation.

A different type of interconnection is shown in Fig. 3. In this case half of the sectionalized bifilar

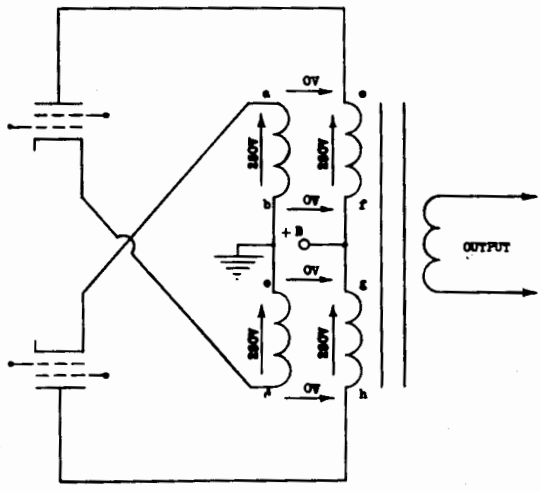


Fig. 3 – Basic McIntosh power amplifier.

[1] A. Pen-Tung Sah, "Quasi-transients in Class B audio-frequency push-pull amplifiers," *Proc. I.R.E.*, vol. 24, no. 11, pp. 1522-1541; November, 1936.

primary winding is connected in the plate circuit, as before, and the other half, with proper consideration for the signal polarities, is connected in the cathode circuit. Now points b, c, f, and g are all at zero ac signal potential. For signal conditions the same as those considered before, when a is instantaneously positive by 250 v with respect to b, e is positive by the same amount with respect to f. This means there will be zero ac signal voltage between points e and a. The same is seen to be true for points h and d and for all other adjacent points on the two bifilar windings. We now require zero charging current for the primary interwinding capacitance regardless of the value of this capacitance. Since point h is at the same ac signal potential as d, but at a much higher dc potential, it could be connected to the screen of the upper tube to supply a constant screen-cathode voltage for this tube. In a like manner point e could be connected to the screen of the lower tube. The screen and plate supply voltages in this circuit will be equal unless special circuitry is provided to make them different. This circuit also requires a grid drive voltage greater than 50% of the output transformer primary voltage. When this circuit is combined with a suitable driving circuit and feedback networks it becomes the McIntosh Power Amplifier[2] and is capable of delivering 50 w of high quality power over an exceptionally large frequency range.

A different solution to the problem is provided by the basic Sinclair-Peterson[3] circuit shown in Fig. 4A and its transformer coupled equivalent, using beam power tubes, in Fig. 4B. In the transformer coupled case both

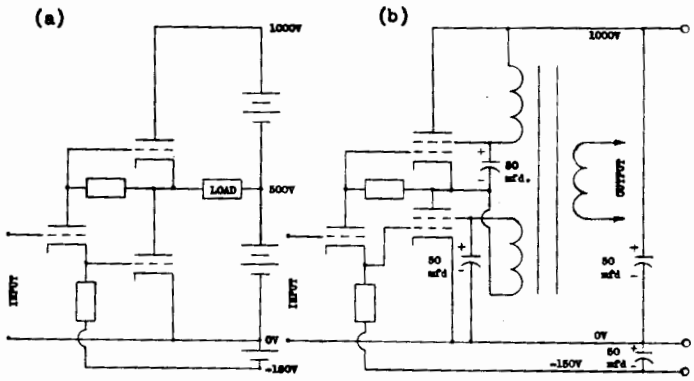


Fig. 4 — Basic Sinclair–Peterson power amplifier.

primary sections of the transformer work at all times so they need not be bifilar wound. These circuits present an appreciable burden to the phase inverter-driver for, if a peak primary or load voltage of 450 volts is assumed, it is seen that the phase inverter-driver plate supply voltage fluctuates between 1100 volts and 200 volts during the course of one cycle. Suitable points are available in Fig. 4B for supplying the screen voltages of the beam power tubes. A disadvantage of this connection is that the plate and screen supply voltages are forced to be equal unless special additional circuitry is used. The basic Sinclair-Peterson circuit has been incorporated in two complete power amplifier circuits which have been described in the literature[4,5]. In these circuits the screens are fed through suitable voltage dropping VR tubes to insure that the screen voltage will be less than the average plate supply voltage. The published data on these amplifiers indicate that their performance is substantially the same as that of the McIntosh amplifier mentioned previously. Neither the McIntosh nor the Sinclair-Peterson circuits use feedback around the output transformer.

The design of the transformer used in the circuit of Fig. 1 represents another solution to the problem introduced by the primary interwinding capacitance. In the general case of two isolated parallel circular wires an increase of the spacing between the surfaces from 10% to 20% of the diameter of the wire will reduce the capacitance between these wires by approximately 30%. An increase of this spacing from 10% to 100% of the diameter of the wire will reduce the capacitance by approximately 70%.

In the transformer winding even though we are not dealing with two isolated parallel wires but with many wires in a small space, the same general principles apply. Each wire will have capacitance to the two wires on each side of it in the same layer and also to the wires in the layers above and below it. The capacitance between wires in the same layer can be cut in half by transposing the two wires of the bifilar wire at every turn. The capacitance between wires in adjacent layers will not be modified by this process. In the non-transposed winding, assuming the same spacing between layer centers that exists between adjacent wire centers in a layer, and uniform dielectric material, the capacitance between the wires in the layers accounts for approximately two thirds of the total capacitance, while the capacitance between wires in adjacent layers accounts for one third of the total capacitance. Since transposition of the wires can be expected to cut in half the capacitance between wires in a layer without disturbing the capacitance between wires in adjacent layers, it should reduce the total capacitance to two thirds of its original value. The use of insulating materials such as cotton, varnish, and wax, and the accumulation of moisture will all tend to increase the capacitance. A mechanism has been developed which automatically transposes the bifilar winding at every turn so that this type of winding is no more

[2] F. H. McIntosh & G. J. Gow, "Description and analysis of a new 50 watt amplifier circuit," *Audio Eng.*, vol. 33, no. 12, pp. 9-11, 35-40; December, 1949.

[3] A. Peterson and D. B. Sinclair, "A single ended push-pull audio amplifier," *Proc. I.R.E.*, vol. 40, no. 1, pp. 7-11, January, 1952.

[4] A. P. Peterson, "A new push-pull amplifier circuit," *The General Radio Experimenter*," vol. 26, no. 5, pp. 1-7; October, 1951.

[5] H. W. Lamson, "A high power toroidal output transformer," *The General Radio Experimenter*, vol. 26, no. 6, pp. 5-8; November, 1951.

difficult to make than any ordinary bifilar winding.

An alternative to the transposed bifilar winding is a random wound bifilar winding. The random winding is not as consistent as the transposed winding but appears on an average to produce an increase in the capacitance of approximately 15%. Other than this the two windings are equivalent.

The introduction of space between the adjacent primary wires will tend to increase the leakage reactance between the two primary sections. The safety factor provided by the bifilar type of winding has so far been sufficiently large to avoid the appearance of the conduction transfer notch.

Fig. 5 is a schematic diagram of the output transformer used to make the tests discussed in the remainder of this paper, and Fig. 6 shows the coil buildup that was used. Fig. 6 is drawn in proper vertical scale but the horizontal scale has been modified to show the relative positions of the windings without showing the true coil width. This transformer winding was designed to be used with two grain-oriented Hipersil C Cores (Moloney ME-31 Hipercores or equivalent). The nominal impedance levels were intended to be 4, 8, and 16 ohms but the optimum levels obtained with this transformer were 4.63,

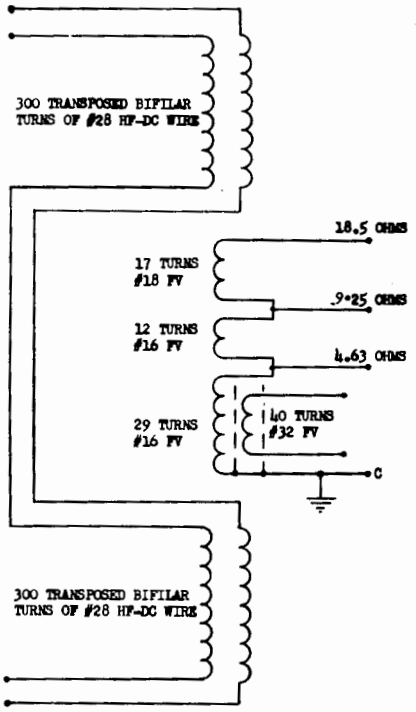


Fig. 5 – Output transformer winding arrangement.

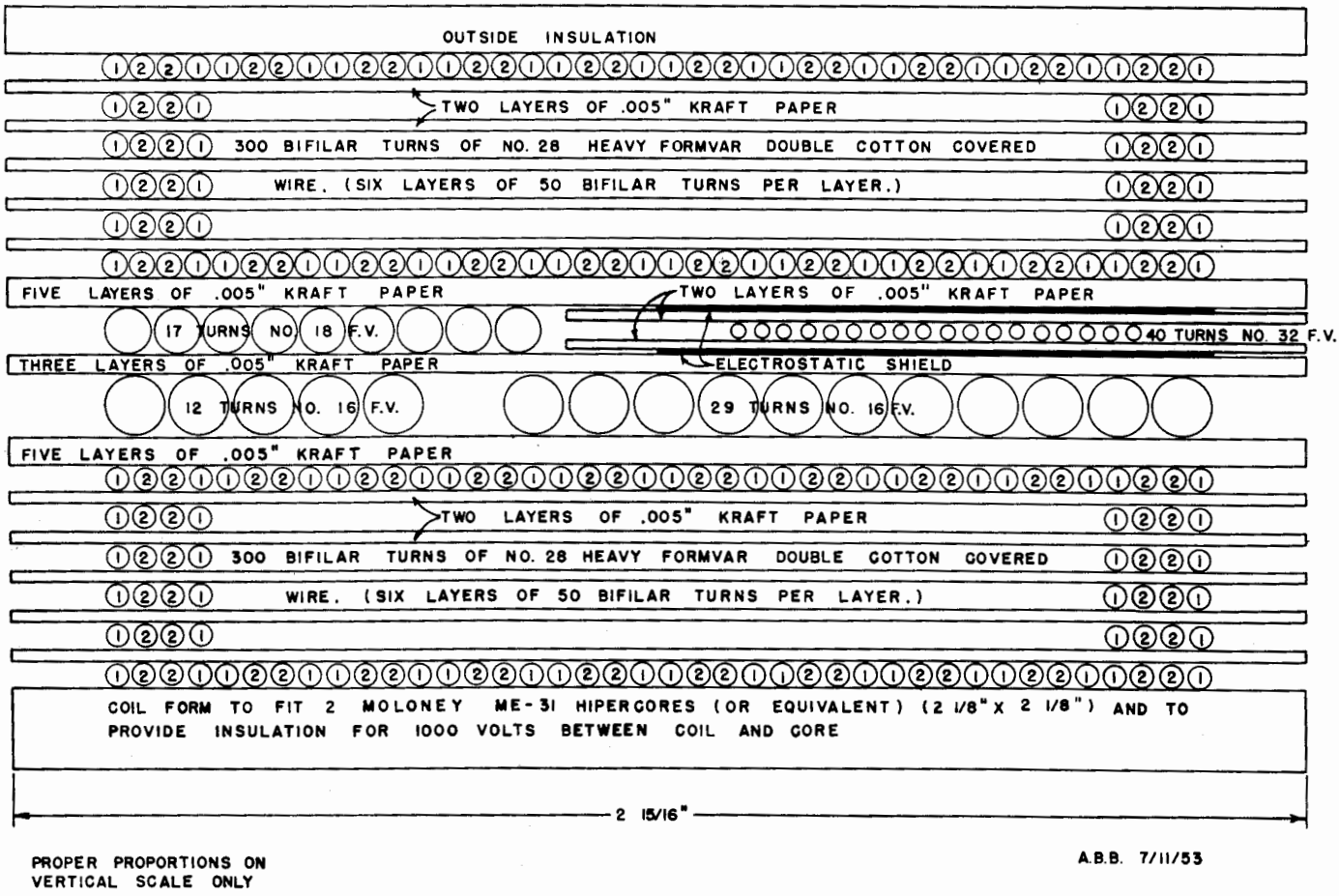


Fig. 6 – Output transformer coil buildup for Bereskin 50 watt 1614 tube power amplifier.

9.25, and 18.5 ohms. An optimum value resistor was used on the 9.25 ohm tap in all of the succeeding tests.

The transformer of Fig. 6 had a primary interwinding capacitance of 0.010 microfarads when it was vacuum impregnated with GE Type 9700 clear baking varnish and baked for the prescribed amount of time. This transformer was potted and has maintained the same value of capacitance since that time. Another transformer winding differing from this one only in the fact it had an ordinary layer bifilar winding had a primary interwinding capacitance of 0.015 mf. A third transformer winding in which No. 28 quadruple Formvar wire was used with the same spacing between wire centers and an ordinary layer bifilar winding had a primary interwinding capacitance of 0.012 microfarads. Additional paper was used between layers in this case to achieve the same spacing between wire centers in adjacent layers. This last winding has been found to have less tendency to pick up moisture from the atmosphere than the other two windings although potting eliminates this trouble in all cases.

In order to obtain the same low frequency power delivering capacity with an ordinary non grain-oriented core it is necessary to increase both the core cross section and the number of turns by about 25%. The combined effect of these increases is that the primary interwinding capacitance increases by approximately 50%.

The circuit of Fig. 1 was used with the transformer of Fig. 6 with transformer coupled input. The input transformer had an electrostatic shield and this shield was connected to ground. All filaments were operated from a common filament supply, one end of which was connected to ground. With well regulated screen and plate supply voltages, the residual hum in the output was 96 db below 50 w. For this particular amplifier, without any special attempt to obtain good balance, a ripple voltage of 42 volts inserted in series with the plate supply, or 9 v inserted in series with the screen supply, was necessary to bring the residual hum level in the output to 80 db below 50 w.

The power delivering capacity of this amplifier was tested by setting the input at the value necessary to produce 2% distortion in the output. Fig. 7 shows the results of this test. In the range below 30 cs the output was limited by the inability of the 1614 tubes to supply adequate magnetizing current to the transformer. Between 30 and 3000 cs the output was limited by peak clipping due to the inability of the 12AX7 tubes to drive the 1614 tube grids positive. Above 3000 cs the output was limited by the inability of the 1614 tubes to supply the charging current required by the primary interwinding capacitance. At the low end the amplifier has a drop-off rate of 9db/octave while at the high end it approaches a drop-off rate of 6 db/octave.

Most of the power in speech, song, and music is contained in the fundamental tones with frequencies below 3000 cs. The power levels of the higher frequency fundamental tones and of the harmonics of the lower frequency fundamentals drop off at a greater rate than the power-delivering capacity of this amplifier. It will be shown in the appendix that the power-delivering capacity of this amplifier is fully adequate for all audio frequency signals.

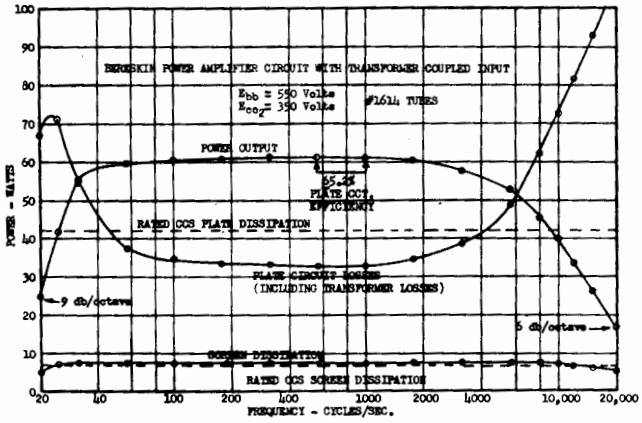


Fig. 7 — Two per cent distortion — power relations.

The amplifier develops its full power output of 60 watts over most of the middle-frequency range with total plate circuit losses, including transformer losses, considerably lower than the rated CCS values. The screen dissipation exceeds the rated CCS value by about 7%, but this is only because the amplifier is being over driven to obtain the required 2% distortion. A reduction of 3 w in the output power, over most of the range, brings the harmonic distortion below 1% and the screen dissipation safely below the rated CCS value. The maximum plate circuit efficiency occurred in the 500-1000 cs. range and was 65.2%. This value includes the output transformer losses, and is ramarkably close to the ideal value of 78.5% for Class B operation, which does not include the transformer losses.

Since transformer coupled input is not usually available for amplifiers of this type a circuit including a preamplifier and power supply was developed and is shown in Fig. 8. This circuit was designed to deliver 50 w of high quality power over most of the middle frequency range. Since the output stage is very insensitive to ripple in the screen and plate supply circuits, very simple power supply filter circuits were adequate. Filter chokes were not necessary in either the plate or screen supply circuits. A single 5U4—G rectifier tube, operating within the manufacturer's ratings, was adequate to supply the power required by the plate circuits of the 1614 tubes. One 6X4 tube is used to supply the power required by the preamplifier and the screens of the 1614 tubes, while another 6X4 tube is used to supply the negative voltage required by the 12AX7 tube. The two 6X4 tubes could be replaced by a single rectifier tube with separate cathodes and plates, used as two single phase half-wave rectifiers, but the tubes available cost more than the two 6X4 tubes put together and, in addition, full-wave rectification in

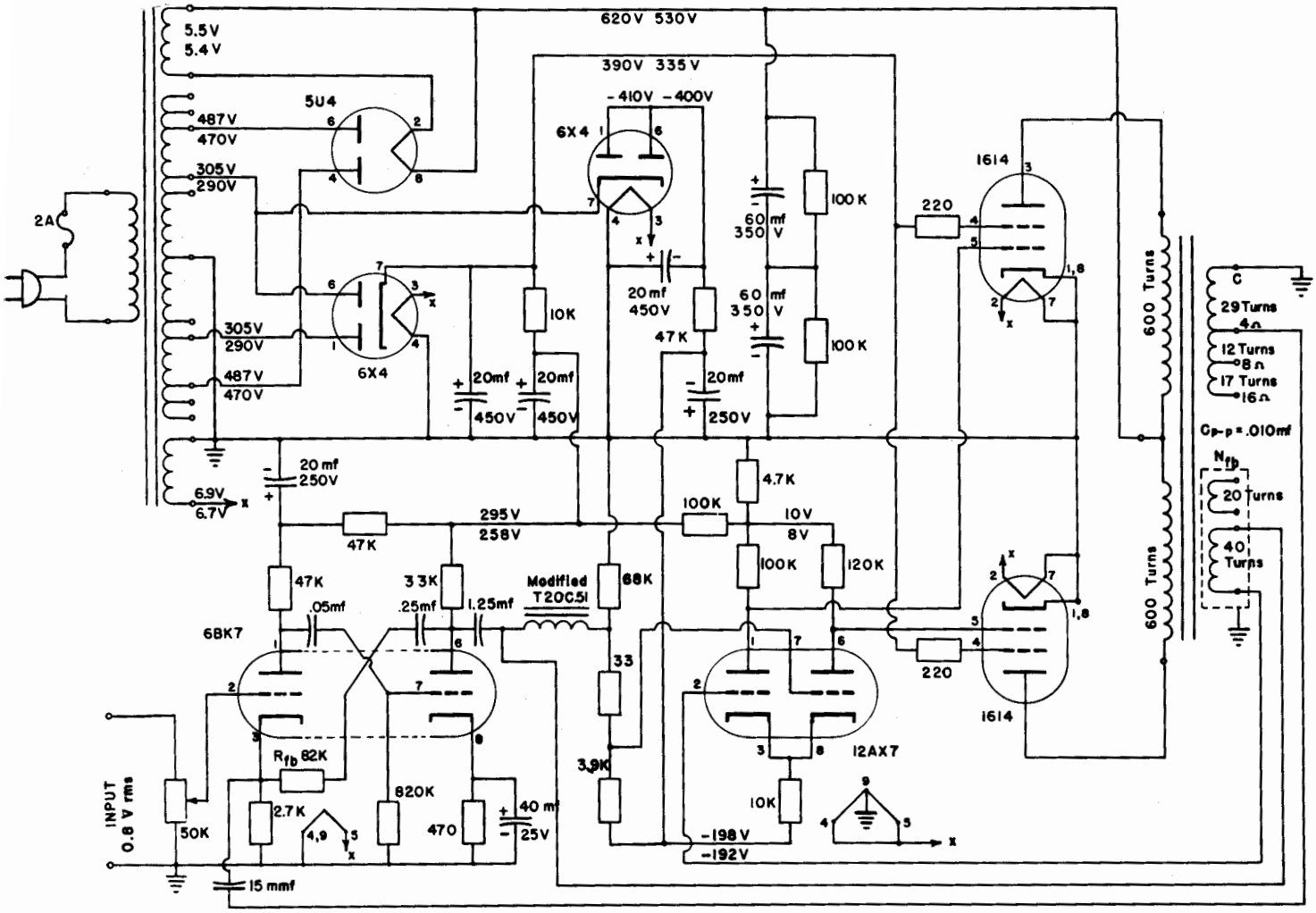


Fig. 8 – Bereskin 50 watt 1614 tube power amplifier.

the screen supply permits the use of a lower value of filter capacitance. The voltages indicated at various locations on the diagram are the measured no-load and full-load values.

The preamplifier consists of a two stage resistance-capacitance coupled amplifier with feedback between the second plate and the first cathode. This feedback provides good wave shape and low output impedance on the preamplifier. The preamplifier is coupled to the 12AX7 grid with a 1.25 μf capacitor and a modified Thordarson T20C51 choke. The modification consists in interleaving the laminations of this choke. A low dc resistance is necessary in this circuit because the 12AX7 has appreciable grid current when the grid voltage becomes more positive than -1 volt and this grid current must not be allowed to change the bias relations of the phase inverter. This coupling circuit has a low Q resonance between 10 and 15 cs.

Feedback from the secondary of the transformer is incorporated in a manner similar to that used before, but additional overall feedback has been incorporated from the 4.63 ohm tap to the first cathode in the preamplifier. It was found that a complex Bridged T network produced the best high frequency square wave response but that the square wave response was perfectly adequate when a simple 15 $\mu\mu$f capacitor was substituted in the overall feedback circuit. This capacitor has no effect on the low frequency response but reduces the tendency of the amplifier to ring slightly with sharp rise time square wave inputs. The photographs of Fig. 9 show the manner

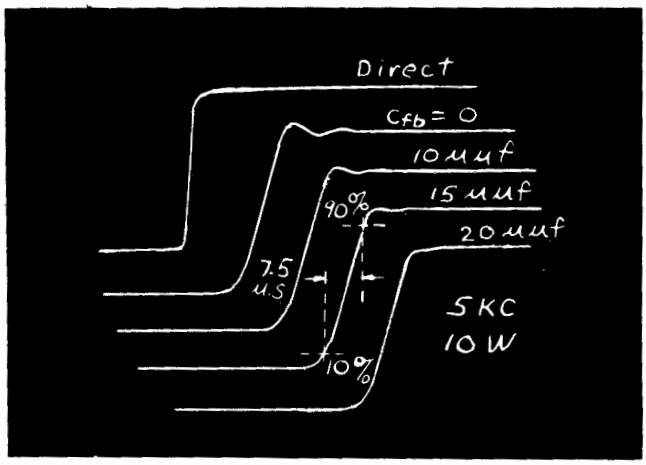


Fig. 9 – Leading edge analysis of 5 KC 10 watt square wave.

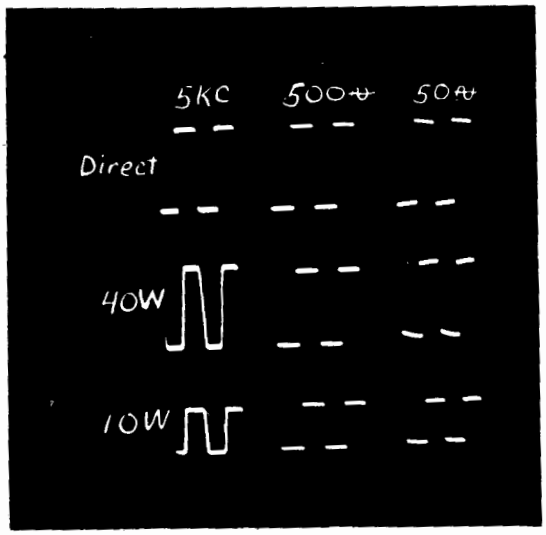


Fig. 10A – Amplifier square wave response.

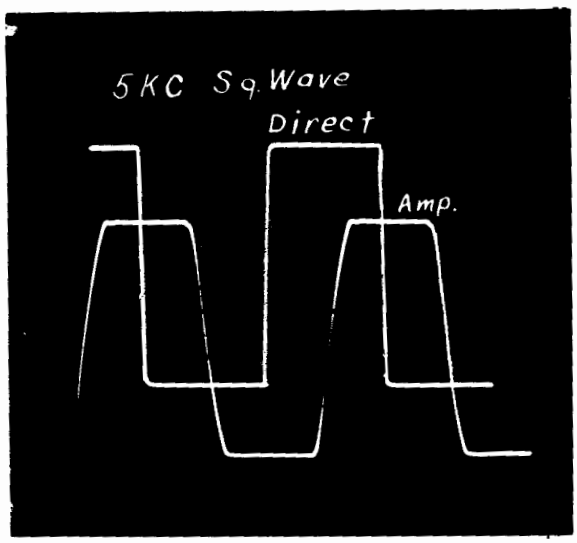


Fig. 10B – Enlarged view of amplifier response to a 40 watt 5 kc square wave.

in which the ringing, following the leading edge of a 10 w - 5 kc square wave, is modified by varying the value of this capacitor. The rise time of the leading edge is approximately 7.5 μs between the 10% and 90% points and the ringing frequency is approximately 100 kc. The complete square wave response at 50, 500, and 5000 cs. and power levels of 10 and 40 w is shown on Fig. 10A together with the output of the square wave generator connected directly to the CRO. A greatly expanded view of the 40 w - 5000 cs case is shown on Fig. 10B. The gain control settings were not changed during these tests.

The results of tests made to determine the best balance between the various types of feedback are shown in Fig. 11. In this figure curves 1 and 2 have inadequate feedback turns to correct for Class B operation at low power levels. The Bridged T overall feedback network produced 6 db feedback at operating frequencies so that curve 1 is lower than curve 2 at high power levels. Curve 3 has much less low level distortion than curve 2 because of the additional feedback turns but it requires more drive from the preamplifier and therefore has higher distortion than curve 2 at high power levels. Curve 4 uses approximately 6 db additional feedback in the preamplifier and its distortion is quite satisfactory at both low and high power levels. The 15 $\mu\mu$f overall feedback has no effect at these frequencies. The values of feedback turns and preamplifier feedback resistance used here represent a practical compromise between low input signal and low output distortion. Additional reduction in distortion could be obtained by increasing the number of turns on the feedback winding and reducing the value of the feedback resistor in the preamplifier. Both of these changes would increase the input voltage required to produce full output power. The conditions specified for curve 4 are the ones shown in the circuit diagram of Fig. 8 and are the ones used in all succeeding tests. Fig. 12 shows the curves of plate circuit losses (in-

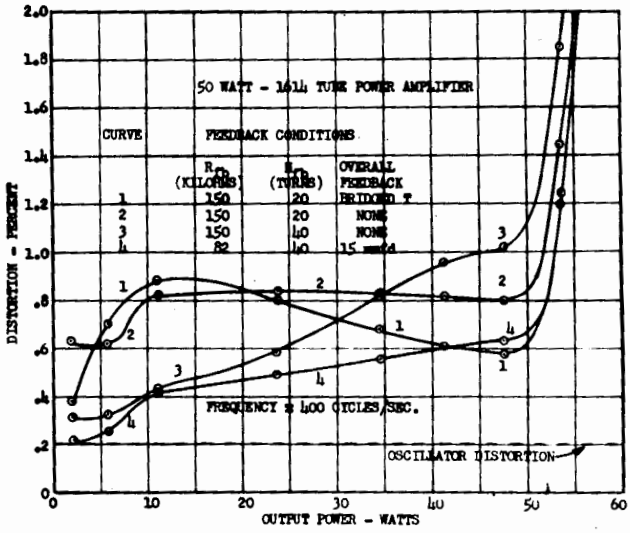


Fig. 11 – Feedback – distortion characteristics.

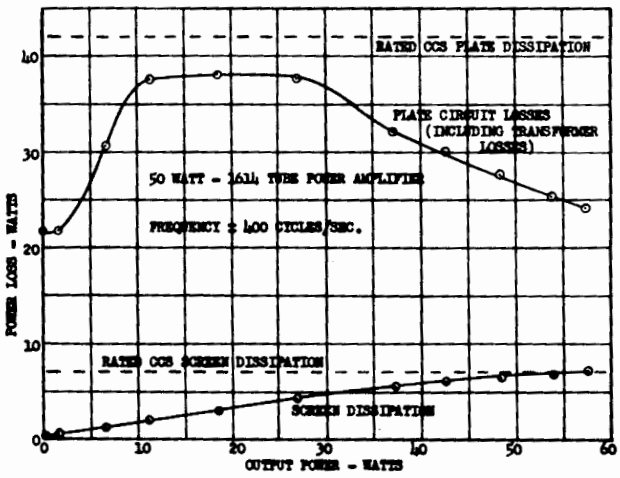


Fig. 12 – Power loss characteristics.

cluding transformer losses) and screen dissipation as the output power is varied. The plate circuit losses are less than the rated CCS values for all operating conditions. The screen dissipation becomes equal to the rated CCS value at the highest power levels shown but is less than the rated value at lower levels.

It can be seen from Fig. 11 that once the amplifier starts over-loading, the distortion increases very rapidly. For comparison purposes a 2% distortion level was considered a satisfactory standard value for determining the power delivering capacity of the amplifier. 1% distortion would have been adequate at 400 cs, where the oscillator distortion was only 0.2%, but not at 20 and 20,000 cs where the oscillator distortion was 0.8% and 0.6% respectively. Fig. 13 shows typical 2% distortion wave shapes at 20, 500, and 10,000 cs, for this amplifier.

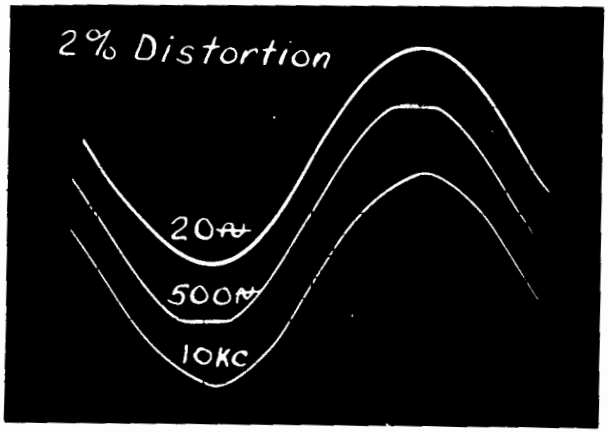


Fig. 13 — Two per cent distortion wave shapes.

The 2% distortion power handling capacity together with the corresponding plate circuit losses and screen dissipation are shown in Fig. 14. The highest plate

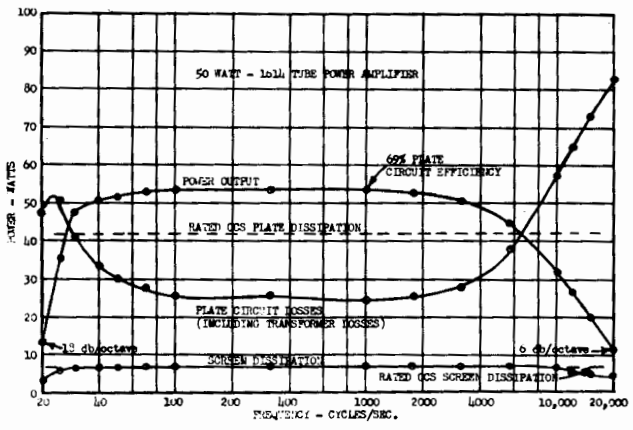


Fig. 14 — Two per cent distortion — power relations.

circuit efficiency, including transformer losses, occurred at 1000 cs and was 69%. The plate circuit losses are below the rated CCS value for the tubes alone and the screen dissipation is equal to the rated CCS value over most of the range. It should be emphasized that this curve represents the power delivering capacity of the amplifier and not the linearity of response with frequency variations. The frequency response characteristics, together with the power delivering capacity curve, plotted to a db scale, are shown in Fig. 15. It can be seen from

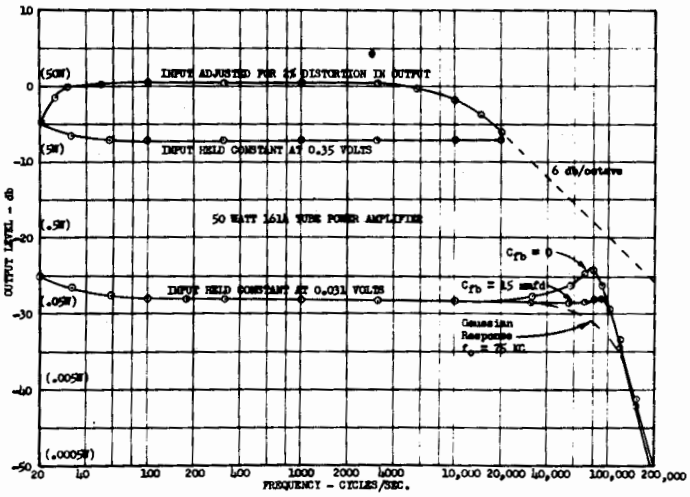


Fig. 15 — Frequency response characteristics.

this diagram that as long as the operating level is below the 2% distortion curve the response is perfectly flat between 100 and 20,000 cs. Below 100 cs the response rises slightly due to the series resonance in the impedance coupled circuit. The amount of this rise can be controlled by modifying the values of the coupling capacitor and choke. At low levels the response above 20,000 cs depends on the amount of feedback capacitance used. The curve for $C_{fb} = 0$ is seen to rise to a maximum at about 85 kc, and then to drop off very rapidly. The curve for $C_{fb} = 15$ micromicrofarads is seen to be almost perfectly flat to 95 kc, after which it also drops off very rapidly. A curve for Gaussian response with a -3 db point at 75 kc has also been shown for comparison purposes. Gaussian response would produce minimum rise time consistent with zero overshoot. It is seen that the ringing frequency of approximately 100 kc corresponds closely to the region of maximum deviation of the actual response characteristic from the Gaussian response. This also shows why the use of $C_{fb} = 15$ micromicrofarads reduced the ringing amplitude obtained with a square wave input signal.

Fig. 16 shows the results of an intermodulation distortion test using a 4:1 combination of 60 and 1500 cs. As is customary, the resulting distortion is plotted as a per cent of the smaller of the two signals. The distortion values shown here should be divided by 5 if they are to be compared with the harmonic distortion values discussed previously. The values shown are acceptable up to at least 112% peak-to-peak equivalent input. An intermodulation distortion test was also performed for a 4:1 combination of 60 and 15,000 cs. and the results are

shown in Fig. 17. The purpose of this test was to find out if the high frequency roll off of the power handling capacity had any appreciable effect on the intermodulation distortion. The values obtained in this test were slightly higher than those shown in Fig. 22 but the performance is acceptable up to at least 105% peak-to-peak equivalent input.

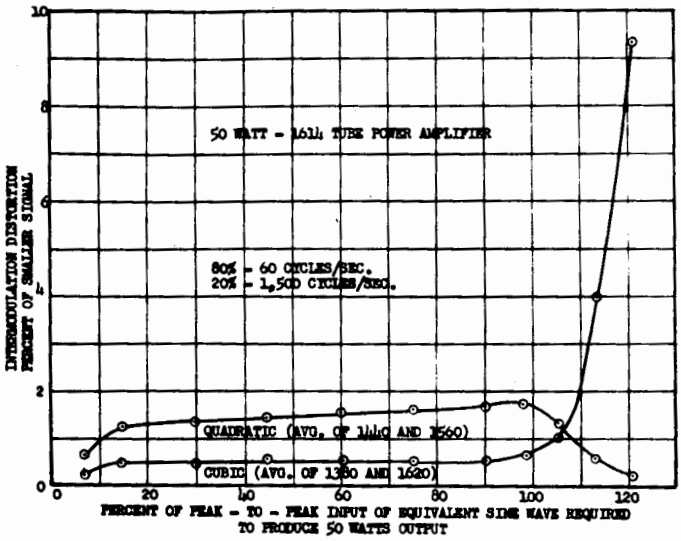


Fig. 16 – Intermodulation distortion characteristics.

Good transient response of a loudspeaker requires a low output impedance source. The manner in which the output impedance of this amplifier varies with frequency is shown in Fig. 18. The output impedance remains relatively constant with frequency and is about 10% of the nominal impedance of the tap. This output impedance was relatively independent of the current used to make the test.

A short circuited secondary test was performed on the amplifier to determine its susceptibility to accidental damage. The results of this test are shown in Fig. 19. The screen dissipation of the tube reaches 4.5 w at relatively low signal levels but remains below 5 w even with 120% of normal full load signal. With the same signal the plate circuit input approaches 120 w. This is considerably greater than the rated value but the plate structure is quite rugged and capable of handling these powers for short periods of time. The screen grid of the tube is quite fragile compared to the plate structure so it is fortunate in this situation that it only has to handle about 70% of its rated dissipation. This amplifier was operated for 10 minutes at the highest signal condition shown on the graph. At the end of the ten minute period the short circuit was removed and the amplifier operated normally. No damage could be detected in any part of the amplifier or its tube complement.

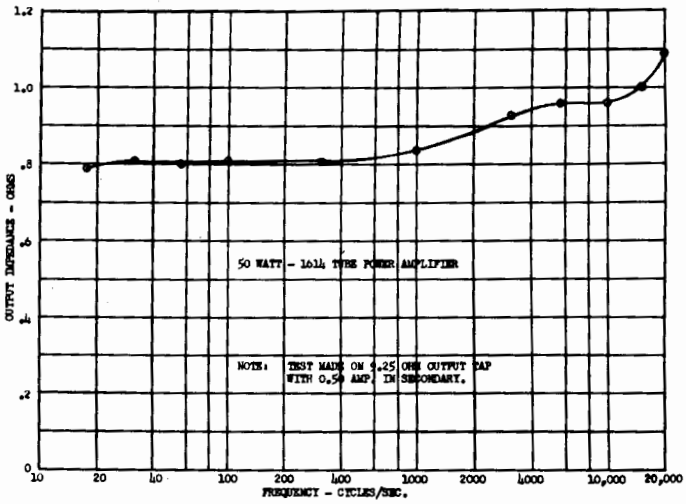


Fig. 18 – Output impedance characteristic.

The amplifier can also operate with full signal applied under open secondary conditions. The only noticeable change is that the output voltage rises by about ten per cent when the load is disconnected. This

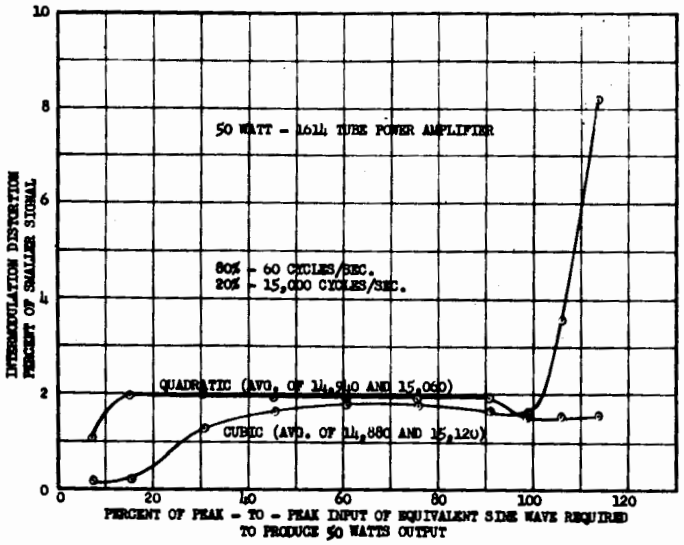


Fig. 17 – Intermodulation distortion characteristics.

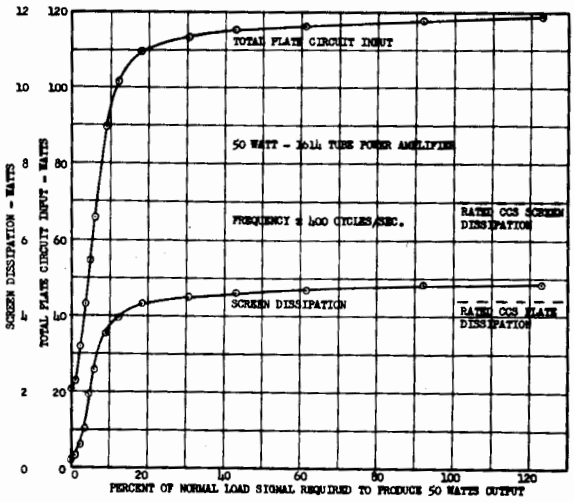


Fig. 19 – Short circuited output – power relations.

control is due to the close coupling used between the secondary and the feedback winding.

The residual hum in the complete amplifier is slightly greater than that measured in the basic amplifier with transformer coupled input. For the complete amplifier the residual 120 cycle hum is about 80 db below 50 w. The residual 60 cycle hum is about 66 db below 50 w. Due to its lower frequency, the 60 cycle hum is no more serious than the 120 cycle hum and both have been found to be completely negligible in most cases. The 60 cycle hum is picked up inductively by the unshielded modified T20C51 choke from the power transformer. If further reduction of the 60 cycle hum is desirable and the amplifier is operated remotely from other equipment, this choke can be oriented for minimum 60 cycle pickup. Approximately 20 db reduction has been obtained by this means but a clumsy mounting position was required for the choke. If the equipment is operated in proximity to other equipment whose relative position may change from time to time, a better solution would be to use a fully shielded choke. It should be emphasized, however, that in most cases further hum reduction is not necessary.

Fig. 22 is a photograph of a developmental model of the Bereskin 50 watt - 1614 Tube Power Amplifier. The output transformer is shown mounted on the chassis in its normal position but not potted.

APPENDIX

On the basis of the characteristics shown in Fig. 15, the 2% distortion power-delivering capacity of this amplifier is down approximately 6 db at 20,000 cycles while the low level response is flat to approximately 100,000 cycles. Theoretical and experimental investigations on both a qualitative and quantitative basis were made to determine the adequacy of this power-delivering capacity. It is the purpose of the appendix to discuss these investigations.

Sivian, Dunn, and White[6] and others have shown that most of the power in speech, song, and music is contained in the fundamental tones with frequencies below 3000 cs. The power levels of the higher frequency fundamental tones and of the harmonics of the lower frequency fundamentals decrease rapidly as the frequencies become greater than 3000 cs. Curve D in Fig. 20 is an adaptation of data presented by Sivian, Dunn, and White with regards to a 75 piece orchestra playing 4 different types of musical selections. In the original data the audio frequencies were divided into suitable but not equal bands, and for each of these bands the instantaneous peak power was recorded. This was done for each of the four musical selections and for many other sources of sound. The adaptation of this data consisted in taking the peak instantaneous power required in each of the four musical selections and calling this value 0 db. All four musical selections produced 0 db in the 250-500 cs band. The value plotted in any of the other bands was chosen from the musical selection in which this band had the minimum deviation from its own 0 db level. This curve is believed to be a close approximation to the most severe requirements for an audio frequency power system and has therefore been called the "MAXIMUM SEVERITY" composite. Curve E has been obtained by converting the

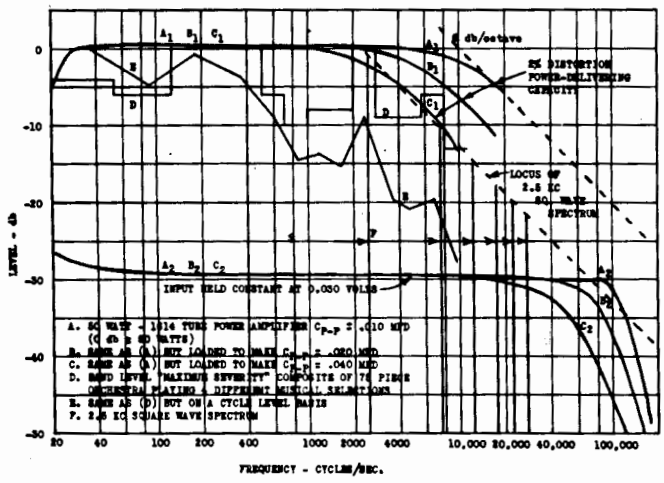


Fig. 20.

data represented by curve D to a cycle level. Sound is in general an integrated composite of many frequencies and these two curves show the relative contributions to be expected, under very severe conditions, from various frequency bands and ranges.

The 50 Watt – 1614 Tube Power Amplifier lends itself readily to the control of its power-delivering capacity without appreciably affecting its low level frequency response or its middle frequency distortion characteristic. Curve A_1 in Fig. 20 is the 2% distortion power-delivering capacity of the amplifier. Curves B_1 and C_1 are experimental 2% distortion power-delivering capacity curves for this same amplifier when it was loaded with .005 microfarad and .015 microfarad capacitors respectively between each of the 1614 tube plates and B+. For curve A the total primary interwinding capacitance is .010 microfarads, for Curve B it is .020 microfarads, and for Curve C it is .040 microfarads. Curves A_2, B_2, and C_2 are the corresponding low level frequency response characteristics obtained with a constant input voltage. The 400 cs harmonic distortion for all three cases was identically the same as that of Curve 4 in Fig. 11. The loading capacitors were connected with a rotary switch which made it possible to switch from any one condition to any other condition without going through the remaining condition.

Several full orchestra passages were recorded from LP records to tape with a Magnecord Tape Recorder operating at 15 inches/second. The tapes were cut into

[6] Sivian, Dunn, and White, "Absolute amplitudes and spectra of certain musical instruments and orchestras," *Jour. Acous. Soc. Amer.*, vol. 11, no. 3, pp. 330-371; January, 1931.

four foot lengths which were spliced into rings so that they could be played over and over with a one second dead interval in between to permit switching of the amplifier capacitance load. The Magnecord output was used as an input to the power amplifier and was also connected to the horizontal deflection system of a cathode-ray oscilloscope. The vertical deflection system of the cro was connected to the power amplifier output and the various gain controls were adjusted for a 45° trace on the face of the cro. Distortionless operation was characterized by a straight diagonal trace with slight tendency towards an ellipse due to the higher audio frequencies. Middle frequency distortion was characterized by horizontal extensions at the tips of the diagonal lines. High frequency distortion was characterized by small loops, similar to musical half note marks, at the tips of the diagonal lines.

For each section of tape the input level to the power amplifier was adjusted so that the amplifier just failed to clip peaks at the highest level passage of this section when the condition of Curve A was used. Switching the amplifier to the conditions of Curves B and C produced correspondingly larger indications of high frequency distortion. Reductions in input level ranging between ½ and 1 db were necessary to eliminate the distortion for the condition of Curve B. Reductions in input level ranging between 2 and 3 db were necessary to

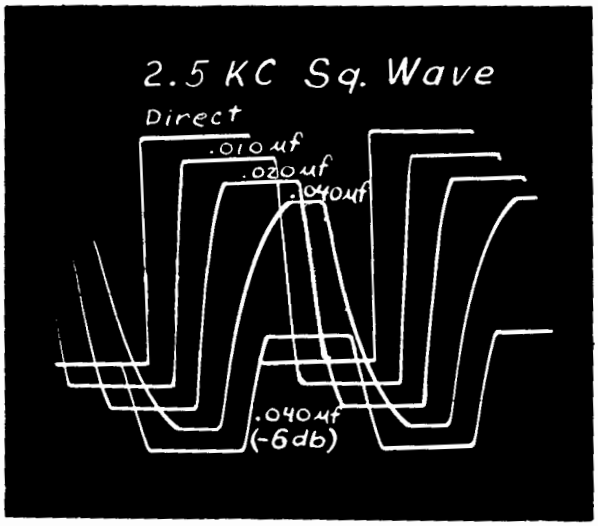


Fig. 21 — Effect of capacitance loading on square wave response of amplifier.

eliminate the distortion for the condition of Curve C. In all cases the distortion shown by the condition of Curve A was of the middle frequency peak clipping variety. Listening tests made at full level for these conditions showed in general that a slight difference could be detected between the conditions of Curves A and C but there was no conclusive agreement as to which represented the better operating condition. In the region of 2500 cycles/second Curve B is seen to be approximately 1 db below Curve A and Curve C is seen to be approximately 3 db below Curve A. This same region is seen to correspond to humps in both Curves D and E. These humps are probably exaggerated by the manner in which the "Maximum Severity" data was adapted from the original data but there is ample evidence that the 2000 to 2800 cycle/second band makes an appreciable contribution to the power content of the composite signal. Beyond this band Curves A_1 and B_1 safely override the band level "Maximum Severity" composite curves. All three of the Curves A_1, B_1, and C_1 are seen to drop off at a lower rate than the cycle level "Maximum Severity" composite in the high frequency range.

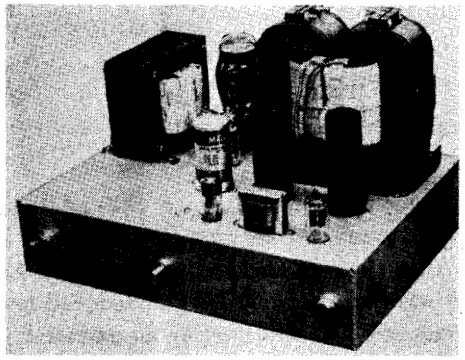


Fig. 22

This amplifier was used in listening tests with several combinations of full frequency range equipment and LP records containing full orchestra passages. In some of these tests the input level was adjusted to just fail to produce peak clipping during the most severe full orchestra passages with the condition of Curve A. In the remainder of the tests a high level was maintained but this level was below the peak clipping region. The transition between conditions A, B, and C was noiseless and in no case was the audience able to detect any difference between the three conditions.

The spectrum of a 2500 cs square wave, having a total power of 0 db, is shown up to the 11th harmonic by the arrows marked F in Fig. 20. The components above 27.5 kc have been left out to avoid confusion in the diagram but their effect can be understood by recognizing that only odd harmonics are present and that the spectrum level of a square wave drops off at a 6 db/octave rate. Several of the harmonics of this wave are in the audible frequency range and the drop-off rate of the harmonics does not differ too much from the band level and cycle level "Maximum Severity" composite characteristics. The behaviour of the amplifier conditions to a full level square wave input signal could be used for comparison purposes to interpret the high frequency power delivering capacity of other power amplifiers.

The oscillograms in Fig. 21 were obtained by the application of a 2.5 kc square wave to the amplifiers corresponding to Curves A, B, and C in Fig. 20. The input signal used in the upper 4 waves was equal to that required to produce 50 w output at 250 cs. The power

present in the output was slightly less than 50 w because of the sloping sides of the output waves. The input signal used for the lowest wave was 6 db below that used for the other curves. The top curve is used for reference and represents the application of the output of the square wave generator directly to the cro with the same gain control settings that were used in all of the other tests.

The rise times between the 10% and 90% points are tabulated below:

C_{P_P} (microfarads)	Signal Level	Rise Time (10% to 90%) (microseconds)	(% of period)
.010	0 db	22	5.5
.020	0 db	44	11.0
.040	0 db	92	23.0
.040	-6 db	36	9.0
(direct from sq. wv. gen. to CRO)	0 db	2.5	.63

The wave for condition (C_{P_P} = .040 mfd) shows a relatively large amount of rise time for full signal operation but this rise time is reduced to less than half its original value when the signal level is reduced by 6 db. All of the waves are characterized by an extremely small amount of ringing which is an indication of good low level frequency response.

It appears from the tests discussed previously that the condition of Curves B (C_{P_P} = .020 mfd) represents a transition range where it is possible to detect incipient loss of high frequency program material with instruments but the loss involved is still insufficient to be detected from listening tests. Although further investigations will be made, it appears reasonable at this point to assume that an amplifier will have adequate high frequency power-delivering capacity if it can reproduce a 2.5 kc square wave at full signal level with a rise time of less than 40 microseconds between the 10% and 90% points of the wave. To avoid the possibility of audible intermodulation components due to the combination of ringing and high audio frequencies, the ringing amplitude should be relatively small and the ringing frequency should be relatively high.

It has been shown that the amplifier discussed in this paper has adequate high frequency power-delivering capacity for all normally encountered audio frequency signals. The middle and low frequency power-delivering capacity and the transient response are all excellent. The harmonic and intermodulation distortion are uniformly low and could be further reduced at the expense of requiring additional input signal. All of these performance features have been achieved with a structure which is compact and relatively inexpensive and capable of withstanding unusual abnormal conditions, such as full signal input with short circuited load, without damage to the amplifier itself.