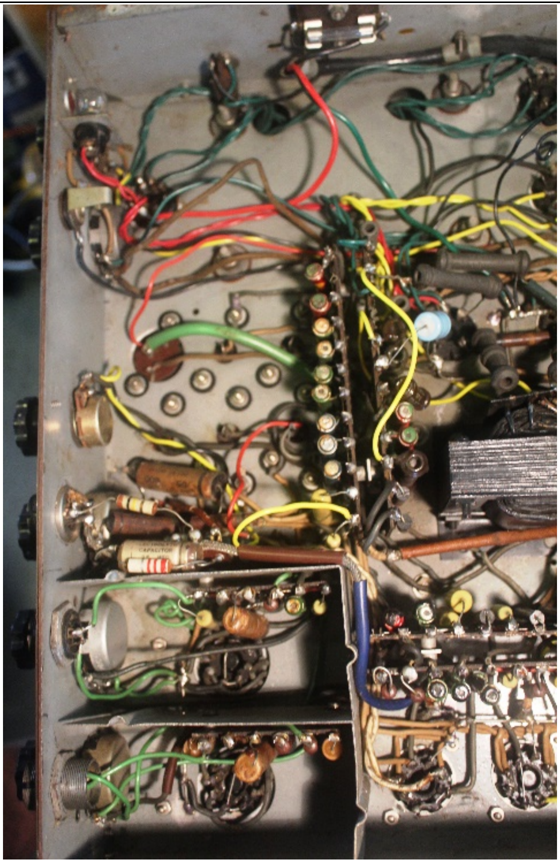
1. Summary

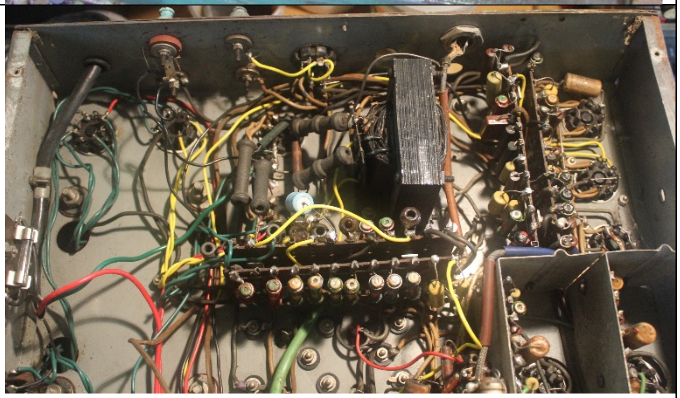
Philips Type 952 valve amplifier. S.N. 4. eBay Jan 2012

Three input (2x MIC, 1x PU) channel PA amplifier. 2x 6J7 MIC amps in screened, isolated enclosure, EF86 MIC amp. 4 input 6J7 mixer. 6J7 amp with output feedback. 6V6 SE driver for inter-stage transformer to feed 807 PP class AB2, and separate screen 100R 5W. Separate monitor and feedback windings. Power supply transformer with separate plate and screen/preamp windings and 5V4 rectifiers, including low impedance back-bias supply for output stage class B grid-bias. Capable of up to 75W output with AB2 or B2 mode operation, with treble roll-off above 7kHz, and 0.25mV MIC sensitivity for 50W

Output Transformer	Type No. 80W nominal 5,500Ω PP (51+57Ω)
	Output winding sections 0,30R,75R,125R,200R.
Power Transformer	0-120-220-240-260V; 450-0-450V ?mA; 270-0-270V ?mA;
	5V 2A; 5V 2A; 6V3 ~4A (BLK,BLK);
Driver Transformer	231: Pri: Sec with CT
POTs	IRC BHO; PD5
Caps	Ducon Aerovox 10580; UCC superseal; Ducon HS
	Ducon electrolytic 3669, 127
	UCC Superseal 85 type PPS Australian made
	Mustards 114H, 108H, 125M, 088H, 029H
Resistors	Mix of old cylinder/dot code and carbon comp and WW.
	IRC WW AA 6811, AA, AB
Valves	807 x2: Brimar brown base : double D getter; 1 cup getter
	6J7G x4: Miniwatt 5K 6C;
	EF86 x1: Mullard ?5F
	6V6GT x1: Sylvania
	5V4G x2: Philips R54; Radiotron ER

Good general condition. Side handles missing. RC snubber on PP. Shielded and vibration isolated enclosure for input 6J7G x2 (but very difficult to replace valves). Most cable insulation and mounts deteriorated. Added input socket and gain pot, and possible EF86 stage. Two MIC input sockets possibly changed (larger size). Rear PU input socket changed. Many wax caps replaced with mustards. Transformer bolts un-insulated. OT dipped in tar. Main 450V HT too high for 5V4G. 20Ω 5W WW grid stoppers for 807 (only available resistor?). PT bad (shorted windings). Vol pot poor contact. Base plate with rust areas.



2. Aim

- Triode mode 6J7 input stages – either g2 and g3 to anode or g2 to anode and g3 to cathode. Stage gain perhaps around 10x per triode. Operate at about 150V plate, with 250V supply and 100k load and 1mA – needing about 6V bias (5k6Ω with bypass).
- Operate 807 PP as class AB2 with fixed bias, with nominal 280V on screens, and up to 600-700V on plates. Aim for 80Vpp (grid-grid) drive. 807 fixed biasing requires about -30V back-biasing with low source resistance to allow 20-30mA grid conduction. Screen conduction current swings will modulate the resistive back-bias – especially as grid-conduction rapidly increases screen current, and for lower frequencies where the back-bias bypass capacitance is less effective. Regulating back-bias suppresses minor compensation of anode idle dissipation with change in mains voltage. A diode and Zener string is the simplest form of back-bias regulation.

- 6V6 SE inter-stage driver operating in class A, triode-mode at 300V, and about 30mA (about 10W plate dissipation), and needing about 12-20V fixed bias. Plate resistance about 2.4kΩ, and voltage gain about x9. Driver transformer voltage transfer 2.5:2 to 807 Vpp (grid-grid), requires 80x1.25 = 100Vpp across driver primary (35Vrms). Loadline is initially horizontal and transitions to ~ 6kΩ as 807 grid-conduction starts (approx. secondary loading of 20Ω + 20V/20mA ~1k).
- Replace 5V4G with GZ34 or 5AR4, or alternatively add 2x UF4007 in series with each anode for the main B+ supply. Increase first filter cap to 50uF (peak current remains under 3.5Apk). Use steering diode from screen supply to avoid loss of VS1 causing screen 807 distress.
- Use CLC 400V cap filter for approx. 310VDC screen/preamp supply, with Zener diodes to pin the negative end to about 30V below 0V ground for biasing the output stage grids with low impedance, and use a pot to derive bias for driver stage. 6V6 driver VS2 requires about 30mA; 807 screen supply increases from 5mA idle to about 22mA; input stages draw a few mA; so VS2 supply range is about 40mA to 60mA. Output stage bias current increases from 0 to about 20-30mA (estimate). Zener needs to conduct up to 60mA max (no bias supply current), so max dissipation is about 1.8W. Zener will achieve rapid bias turn-on as VS2 starts to rise and draw current.
- Fender style tone-stack using treble pot and bass 3-pos switch.
- Restrict bandwidth – target 60-70Hz stage CR coupling corner frequencies (4N7 & 500kΩ). High frequencies attenuated by tone stack, and then across driver transformer primary.
- Feedback via monitor winding – plugging in monitor jack can then disconnect feedback. Adjust feedback resistor for about 6dB feedback. CR coupling to driver stage, and driver transformer primary shunting can roll-off gain before output transformer effects are seen.
- 10Ω current sense resistors for 807 cathodes. Worst-case current waveform estimated at 200mA squarewave, so 100mArms causing 0.1W dissipation.
- Configure the output transformer speaker windings to present an 8Ω level at speaker terminals. Speaker should be able to handle 100W.
- Check if hum is alleviated by heater elevation.

3. Modifications

- Cleaned baseplate of rust and painted and fitted new feet. Fitted handle at one end.
- Megger checked transformers ok.
- Fitted IEC socket/fuse/switch combo. MOV on PT primary.
- 2x IN4007 in series with each 5V4 plate. PP stage supply CT fused.
- EC13 choke for VS2 CLC supply.
- 30V Zener for back-bias using 15V + 2x 5V6 zeners + 6x 1N4004 in series with output switched back to a Zener/diode node, and 15V Zener always in circuit. Set bias for 20W dissipation (500V x 40mA).
- 6V6 fixed bias taken from pot across back-bias.
- Steering diode (2x 1N4004) from VS2 to VS1.
- Replaced electrolytic and foil caps.
- Replaced isolation mounts.
- 807 grid stoppers replaced by 20Ω 1W.
- 612Vdc 1mA MOV's across each primary half-winding.
- Monitoring connections made to 'impedance matching' octal on rear panel. 180k+180k:3k9//68k.
- Power transformer replaced with TEK 120-0117. Although this transformer has multiple spare windings to allow fixed bias, the original back-bias was retained.

- Cleaned vol pot track.
- Added Fender style tone-stack with fixed mid-range, and fail-safe across Bass switch.

Replacement power transformer - TEK 120-0117-00 Power Transformer for 502A Oscilloscope – see spec sheet.
Primary: 110,117,124,220,234,248V (50 to 400Hz). Main windings 1,2,3,4 are 117V and 234V, with A,B,C,D buck/add windings. DCR = 5.5.

Heaters (loadings from tubes identified in 502A schematic)

terminals;	voltage;	heater current loading	Spec sheet rating	
12-13	5V	2A [large terminals]	2.0A	5.2V
27-28	5V	2A [large terminals]	2.0A	5.2V
5-6;	6.3V	~0.6A	2.0A	6.5V
22-23	6.3V	3A	3.0A	6.5V
24-25	6.3V	?A 3kV elevated	0.9A	6.5V
18-19	6.3V	9.7A [large terminals]	9.0A	6.5V
10-11-21	9.5-0-9.5V	0.6A at 6.2VDC heaters	0.6A	9.7V

HT windings

29-17-30:	395-0-395V	~200mA	DCR=88	0.18A 383-0-383V
7-8-9:	175-0-175V	~100mA	DCR=176	0.03A 170-0-170V
14-15-16:	220-0-220V	~150mA	DCR=86	0.09A 213-0-213V

[anticipated ratings based on comparison with other transformer ratings and DCR]
Screens

20:	primary side earth screen.
26:	not connected.

Core size: 95x115mm, 65mm stack.

3.1 Frequency response and feedback design

Original frequency response was purposefully restricted to circa 7kHz max, along with separate bass control (switched screen bypass caps for feedback stage V4) and treble control (feedback path). Bass roll-off also by inter-stage coupling caps. Treble roll-off also by output transformer primary PP RC, and 6V6 anode bypass, and inter-stage transformer response, and feedback RC's. Original level of mid-band feedback unknown.

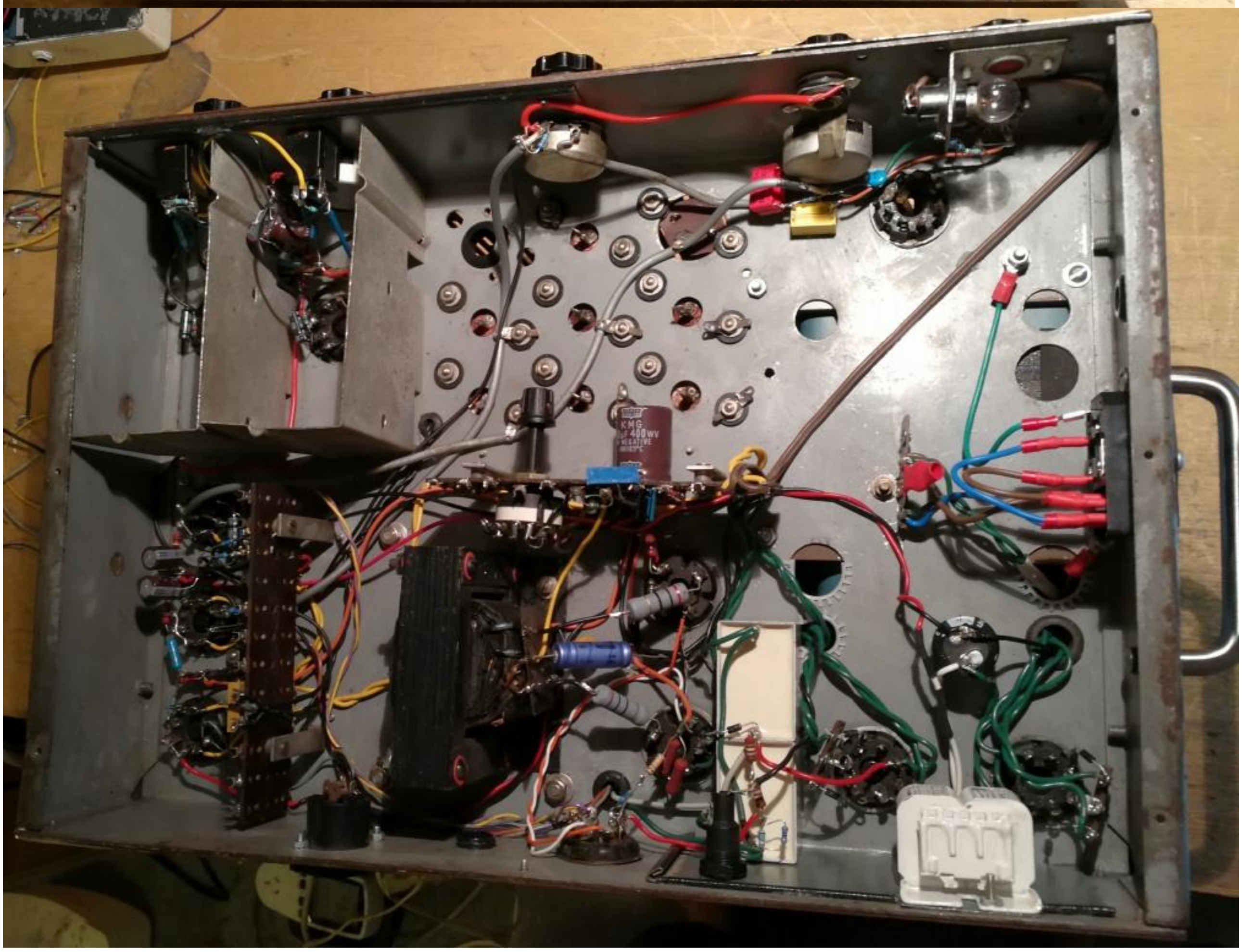
Modified amp uses wide-bandwidth, low gain (6J7 in triode mode) input stages each with their own output volume control, that can be used sequentially. Then a buffer stage feeds a Fender style tone stack, with output volume pot feeding a triode-mode stage with cathode global feedback node. Output stage PP RC, and driver stage anode bypass, and feedback path RC's removed, and output stage changed to fixed bias. Inter-stage 6V6 bias changed from interactive to fixed. Feedback winding changed to Monitor winding (TBD), which can be disabled. Output stage grid-grid load retained at assumed 10k.

Tuning the output transformer primary PP RC was a known method to compensate for the rising PP stage gain above circa 1kHz, which can significantly degrade stability margin. RCA 1937 AN-68 'A 55-Watt Amplifier Using Two Type 6L6 Tubes' characterises the compensation by such an RC, where R=1.3Rpp (ie. circa 7k5), and C is adjusted for flat gain out past 10kHz (eg. circa 47nF). STC 1954 807 application report appears to apply a 5k-1.5nF RC PP shunt for the described 75W amp, where OPT has 4k5PP primary – which appears to dampen a 13kHz peak, but no significant details.

RDH4 p590 discusses adding shunt C to each inter-stage half secondary, to resonate with leakage L and cause a gain drop beyond about 20-30kHz [E.18], or alternately add shunt C to primary side anode.

IRE 1936 McLean describes negative resistance effect influence on unstable operation, and points to IRE 1935 Fyler with a diode based mechanism to alleviate the effect.


PHILIPS AMPLIFIER TYPE 952
OPERATING INSTRUCTIONS
IMPEDANCE MATCHING
EARTH



4. Measurements

Megger test 1kV on PT mains, and 500V on OT primary and secondary – all ok. PT has a shorted turn(s) – draws abt 0.7A on primary winding.

TEK PT: 245V, 30mArms. 405-0-405; 224-0-224: 5.66; 5.66; 6.8.

Voltage rail regulation.

Rail	245Vac mains	45W output	60W cranked
	Bias set for 20W per 807		
VS1	540V to 487V idle, 7.7Vac 41+42mA, 20W + 20W VS5: -20.3V	445V	435V
VS2	(250V, 1.7Vac) 246V	242V	223V
V3	238V Vak, 28mA, 6.7W VS6: -15.7V		
VS3	227V		
V4, V5	128Va, 5.6Vk; 123Va, 5.7Vk		
VS4	209V		
V6, V7	114Va, 5.3Vk; 114Va, 5.3Vk		
Heater	6.5V		

12VAC 50Hz nominal applied to output transformer

Winding	Voltage rms	Turns ratio; Impedance for 5.65K pri; Spec level; Notes
Pri P-P: DCR winding	49.2	
Sec: 200Ω	9.26 0.7	; Ω; 200Ω; 1000T ‘1’
Sec: 125Ω	7.4 0.7	; Ω; 128Ω; 800T ‘2’
Sec: 75Ω	5.56 1.1	; 0Ω; 72Ω; 600T ‘3’
Sec: 40Ω	3.7 1.3	; Ω; 32Ω; 400T ‘4’
Sec: COM	0	0T
Sec: FB	1.86 4.1	; Ω; 8Ω; 200T
Sec: 0Ω	0	0T
Sec: MON	0.124	; Ω; 0.04Ω; 13T

Output transformer primary DC resistance: 55+50Ω.
Output transformer tap locations: Anode; B+; Anode; space; Mon; 75Ω; 200Ω
COM; space; FB+; FB-; space; 40Ω; 125Ω

Wiring: Com=Grn ; FB+=Pnk ; FB-=Blk ; 40R=Red ; 125R=Pur ; 75=Tan.
Red=Gry to Speaker - ; Tan to Speaker + ; So 40R to 75R winding to speaker (for 8Ω output).
Pnk and Blk to feedback.
Monitor, COM and 200R taps not connected. Other brought out for convenience.

Three of the four main sections of windings, and the feedback winding, have 20% turns for 8Ω, but the main sections can’t be separated into independent windings. The FB winding has a higher DCR than the three main sections with 20% turns, so using just one section of 20% turns for speaker 8Ω output (ie. 40 to 75R section). The Mon tap is relative to the FB-, which is connected to signal ground. The feedback winding remains the same, and can be swapped for phasing as it is an independent winding.

Driver transformer: primary DCR=266Ω.
Measured stage gain: V7, V6 each about +22dBV (13x). V5 gain with tone at max and bass at max is about +15dB (6x).

V6 stage FR showed bass roll-off, so 4n7 increased to 15nF; gain of 14x.
V6+V5 stage FR showed +2dB rise at 5kHz (flat to 12kHz) with original 220pF top tone stack. 30pF tone FR had no peak, and fall starting 6kHz, and -3dB at 12kHz. Tone pot at max, and Bass at max, and 100pF gives flat FR to 10kHz. 150pF top tone stack chosen – marginal +1dB rise at 5kHz, with 0dB at 12kHz, and -2dB at 21kHz.

Frequency response has LF roll-off from coupling caps of around 80-100Hz corner, and HF roll off above about 10kHz. Tone and Bass switch have reasonable variation. Preamp frequency response extends from abt 70Hz to 22kHz at -3dB.

Relatively clean out to 45W into 8Ω (4.4% THD), with <28mVrms input (couldn't max out all pots – tone at max). Cranked to 60W continuous with clipping. 56mV MIC1 into 10V MIC2 Vol top out. See measurement waveforms/spectrums.

Output noise floor (Keithley 197):

1.7mVrms with all pots at min	1.9mVrms with Vol at max
9.4mVrms with MIC2 and Vol at max	144mVrms with MIC1, MIC2 and Vol at max.

Weight 21.4kg

FR with REW from top of Vol pot input (so only V4, V3, output stage) to 8R load, no f/b. Bass roll off below 1kHz (500Hz -3dB). Increased V4-V3 coupling cap from 4n7 to 47n – lowered -3dB to 200Hz – maybe enough before adding f/b. Noticeable resonance abt 85-90kHz, but otherwise smooth to 40kHz, and midband HD at 1W dominated by 2H and 1.5% thd, and below 2% from 800Hz to 12kHz.

3dB f/b with 63R – flattens FR but 12dB peak at 70kHz using Monitor winding – suppress hf peak then see how much feedback is available from winding. Swapped to output for f/b (needed to swap inter-stage primary phase), and set for 3.2dB - similar 12dB peak at 60kHz. Hf peak constrained by adding shunt C from V3 anode-cathode. 3.2dB f/b was flat with abt 3n3F. 10dB f/b showed some minor smooth peaking at circa 50kHz with 10nF at 1V 8R, but this got a bit ragged at 4V.

10dB f/b with Picoscope and FRA from 1Hz to 1MHz shows dominant 50kHz resonance, and then further 100Khz resonance, but nothing else significant. Shunt 10nF across V3 anode-cathode flattens response out to 40kHz for rated 8R load, but higher f/b or higher load R causes peak to reappear. SQW shows anticipated response.

Feedback and speaker windings modified. Speaker from just 40-75R winding. Feedback winding only used for feedback. Monitor and other winding section not connected to.

Further increased 47nF V4-V3 coupling to alleviate bass roll off with no f/b.

Added shallow shelf filter to V4, but would need to stop phase change above eg 50kHz.

Added some phase advance out at 50-100kHz. Checked that 1nF//3k2 (50kHz) showed some benefit – too much more caused oscillation. For no load, the 20kHz resonance was suppressed from +15 to +10dB with 680pF comp across f/b 3k.

Varied the loading across the feedback winding to see influence. Feedback path imposes ~3k.

Zobel RC across speaker, to avoid no load at 20-40kHz and the resultant peaking. Zobel load can be circa 20-50R, and cap circa 220nF. 20R-100nF drops peak about 3dB and lowers freq from 19k to 17k. 20R 220nF just lowers and broadens peak.

Instability worse with 6V6 cap to 0V, compared to cathode. Added a 100R screen stopper, but no apparent change.

Added 15k-220pF shelf network across 100k for 7kHz start, and 50kHz stop, and reduced shunt 10nF to 3N3 – resonant peak at 45kHz, and unstable with 6dB for >220Vac.

Lowered 10k PP grid load with //2k2 – which flattened FR and was stable at 210Vac, but showed +7dB peak at 40kHz at 240Vac. Added 560pF across each 807 grid winding – no noticeable change – then increased to 2n8 and increased peaking.

Lowered 10k PP to 10k//1k, and step to 10k-330pF for 5kHz start, and 50kHz stop, and 4n7 shunt, and 6dB f/b at 240Vac gave +4dB peak at 35kHz, and flat down to ~80Hz.

463V, 235Vdc, 35+35mA, 19.5mA. 5dB f/b, 0.36V 8R for 78mV input to top of pot (gain 4.6x). Say 18Vrms for 40W, so 40x gain needed from pre-amps for 40W.

MIC2 V6 stage provides clean gain to 20Vac across MIC2 Vol pot.
Tone stage provides clean gain to 15Vac across PU Vol pot, with slightly asymmetric soft clipping above – out to strong clipping above 20Vac. 10Vac at 200mV sine input and max Mic2 pot.
8R output clipping above abt 5 (for clean signal at top of PU Vol) with f/b on.

With feedback off, the 6V6 cathode current waveform shows it firstly bottoms at 0V and then the top starts to clip, at 210Vac Ibias is 11.8mA – 240Vac 20.0mA. Increased to 25mA, Vout to 8V before 6V6 cathode current touches zero. Increased to 30mA (229V), Vout to 8V before 6V6 cathode current touches zero – with now symmetric clipping – with 1k//10k on 807 grid PP.

With just 10k PP 807 grid loading, the 6V6 cathode current shows clipping on high side before low side approaches 0mA with about 18-19Vac out (40W). 807 cathode current waveform shows Class B starts about 8Vout, then obvious waveform discontinuities initiated at cross-overs.

Monitor input jack rewired as input to top of PU Gain pot, for testing direct into V4 feedback stage. Switch in OPT f/b winding, then through 10T pot to 220R. 6V6 screen stopper, and 4n7 anode-cathode bypass. 47nF coupling from V4 to V3. 10k-330pF step across V4 anode 100k. 10k loading of 807 g-g. 807 V1 (closest to rear and 76R driver side) to brown primary (swapped during feedback changes).

Modified 6V6 drive capability – see interstage section.

SQW and GP at different Pout without f/b. Ringing at ~45kHz for f/b >8dB (no comp). With f/b, minor ringing snivets at 45kHz start at 5Vrms at peak of waveform, with snivets moving to two regions on either side of peak as signal output increases and of low disturbance with rapid dampening. Then above abt 12Vrms, more substantial snivets start at top of waveform and take longer to dampen. Pico output 2V limit at 15Vrms output, so need to use preamp stage. GP responses at 1 to 7V with 7.8dB f/b, shows the 45kHz resonance peak reducing in level compared to the rising mid-band output level. SQW shows those snivet disturbances on top edge of waveform forming as SQW level increases, but not on bottom edge. The snivets do not appear to be related to quasi stability, but to disturbance induced from AB1-2 transition on one side of waveform.

807 bias=41+40mA, 22.5V, 473V, 19.3W. 6V6 bias 20mA, 18.4V, 250V, 5.0W. Swapped 807's, and no oscillations showed up at 7.7Vac and 17.7Vac output into 8R (39W). Unsure what mechanism caused the oscillations at current =0 transitions.

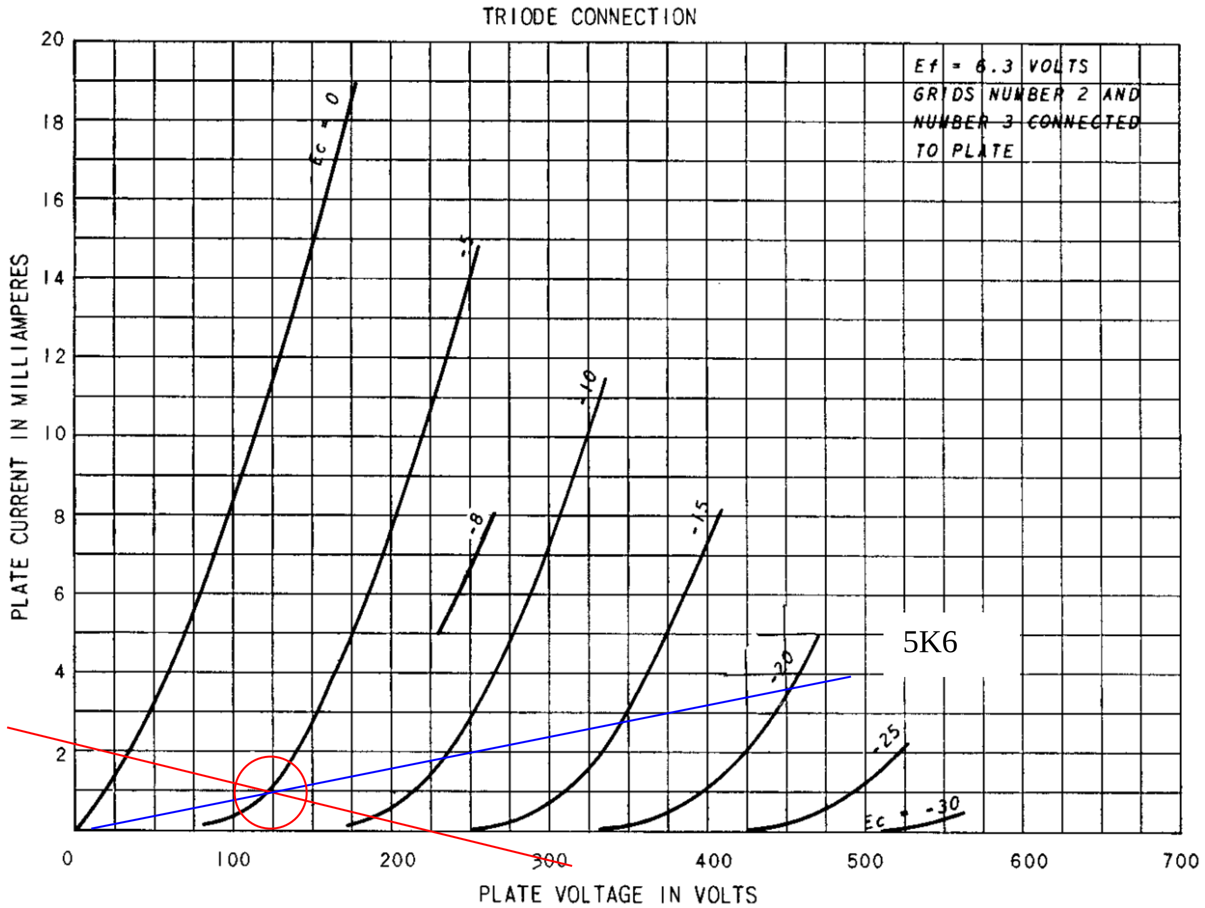
Check 807 borderline stopper conditions.
135R screen stopper #1 - //10R – no noticeable difference.
Short both 135R screen stoppers with //1R - no noticeable difference.
Short both 20R grid stoppers with //1R - no noticeable difference.
Raise VS1 using bridged anodes - no noticeable difference.

Checked stability with GP plots. Some comp cap is better – set with 680pF. No load gives quite a peak, and needs cap comp for stability.

To do:
Check bias variation/regulation with output power/AB2.

4.1 Pre-driver Stage

6J7, V5; VS3 = 230V; Va=125V; Rk=5k6; Vk=5V; Ia=1.0mA; RLdc=100k. Available swing is 220-30=190V, ie. >50Vrms.



4.2 Interstage driver transformer

This Class A single-ended driver stage uses the 6V6GT in triode mode with fixed bias. A 15.6kΩ impedance is presented to the 6V6 plate by the interface transformer (the value of 807 grid-grid load was not measured, but photo indicates 10kΩ, which would reflect back 15.6k to primary). The secondary voltage swing is about 20Vpk for class AB1, and perhaps 40Vpk into AB2 of 807 output stage. With a 2.5:1:1 turns ratio, the primary swing may need to reach 2.5x40=100Vpk, although the primary side loading will increase as 807 grid conduction kicks in.

6V6 plate DC voltage will be lower than VS2 by about 10V; ie. OPT primary resistance of about 268Ω with idle current of 35mA, and 10Ω cathode current sense. Grid fixed-bias voltage VS6 has an idle bias of about -16V (derived from the output stage back-bias). So effective plate-cathode idle voltage is about 246-10=236V.

The maximum 6V6 bias current allowed is dependent on the maximum recommended plate and screen dissipation of 12+2W: Ibias(max) = Pd / Vb = 12W / 240V = 50mA. Given a 35mA idle, and class A operation, the peak dissipation during AB2 appears ok.

Loadline of 15k6 across 6V6 indicates a 236-30V=206Vpk primary swing into 10kPP load (807's in class A region). When 807 grid start to conduct, the loadline would bend as indicated (for 2k loading), and suggests idle should be set at a higher current than 35mA to allow a more symmetric swing into the cut-off region.

The 6V6 anode voltage waveform probed with Picoscope 10:1 probe with 600Vpk rating, although derate to at most 400Vpk, as 4224A has 40Vpk input spec. So using 10:1 resistor divider to reduce 600Vpk signal on plate to 60Vpk on 10:1 probe. Divider uses 2x200k PRO2 upper, and 39k+5k6 lower arm, for total loading on anode of 445k, and idle probe input voltage of 33Vdc.
183Vdc=18.0Vdc.

For 240Vac mains, VS1=468V, VS2=227V, 807s 30.4+30.5mA (14W); 6V6 30mA.
500Hz 18Vac output into 8R (40W), 6V6 idle with VS2=227V, and 30mA cathode. Sine clipping on 6V6 anode waveform at -100Vdc negative swing from idle, with positive swing compressed but no clip. This indicates 6V6 is going into grid conduction.

807 bias=30mA at 22.2V VS5 and 14W diss. Bias=37mA at 20.0V VS5 and 17W diss.

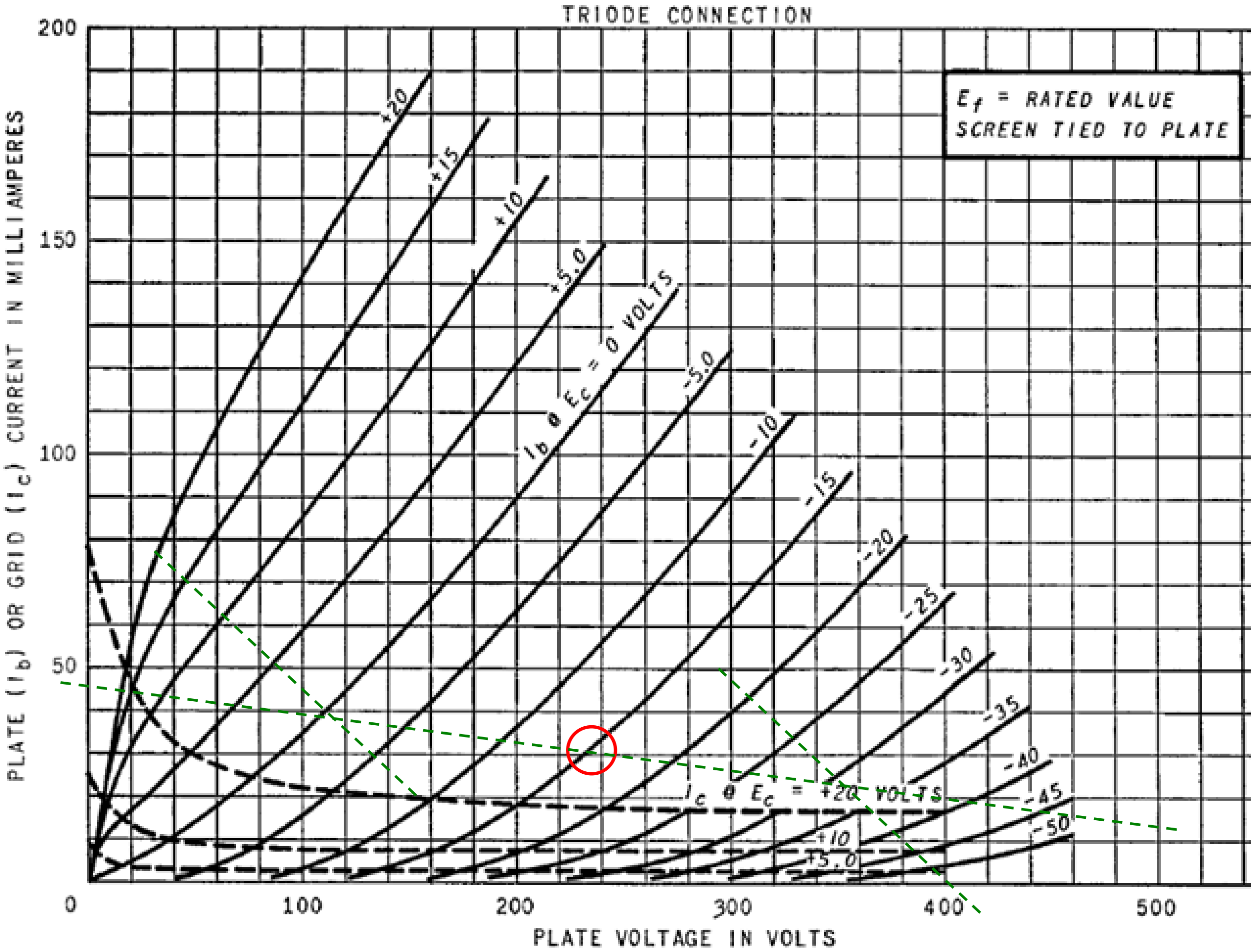
Lower 6V6 idle bias from 30mA to 25mA:
807 bias=37+37mA, 21.5V, 472V, 17.5W. 6V6 bias 28mA, 14.0V, 234V, 9.4W. No clip at 18.2Vrms 41W; noticeable neg swing clip at 19.6Vrms 48W.

807 bias=38+37mA, 21.9V, 473V, 17.5W. 6V6 bias 20.2mA, 17.2V, 240V, 4.8W. No clip at 19.0Vrms 45W; noticeable neg swing clip and pos clip at 19.4Vrms 47W.

Raise 10k pp secondary load to 10+39k:
807 bias=37+36mA, 21.9V, 468V, 17.5W. 6V6 bias 19.5mA, 17.2V, 237V, 4.8W. Pos side compression first at 18.7Vrms 44W; noticeable neg swing clip starting at 19.7Vrms 48.5W.

Bridge the VS2 5V4 rectifier anodes:
807 bias=40+39mA, 22.0V, 468V, 18.7W. 6V6 bias 21.6mA, 17.2V, 246V, 5.3W.
807 bias=43+41mA, 22.0V, 470V, 18.7W. 6V6 bias 20.0mA, 18.0V, 248V, 5.3W.
AB1-2 transition noticeable from abt 5.5Vrms 4W at neg peak of 6V6 anode, moving to the sides of sine (~200Vdc) at higher output level. Neg clipping at 20.3Vrms 51.5W.

Now the 6V6 drive appears sufficient to force 807 PP class AB1-2 transition into 6V6 linear region. So return to assessing class AB1-2 transition and ringing in 807 output stage and feedback.



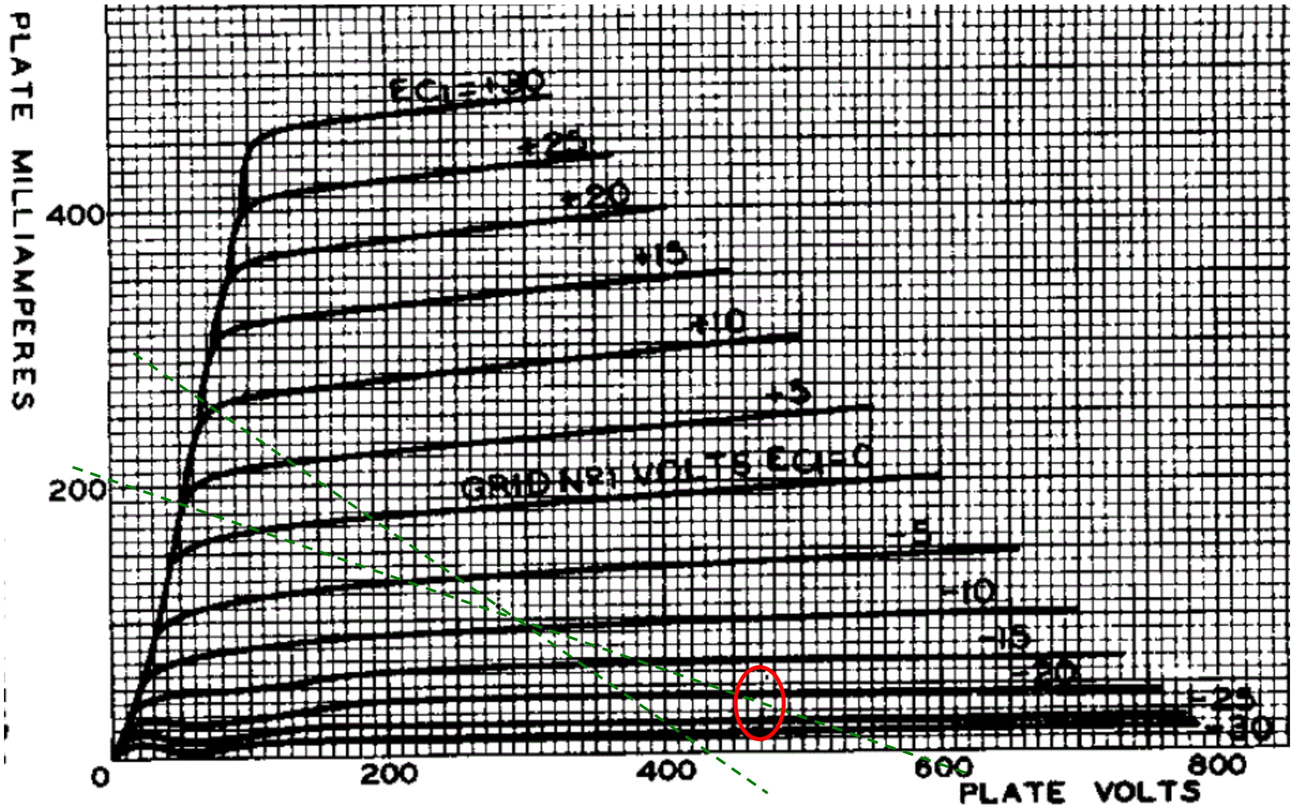
4.3 *Output stage*

807 fixed bias push-pull stage operating into Class AB2. 10Ω cathode sense per tube. 20Ω grid stopper and 135Ω screen stopper on each base terminal. 612Vdc 1mA MOV across each primary half-winding. Feedback winding loaded. 40R to 75R winding section to Speakon socket, with one side grounded. 5k6 PP reflected from 8Ω load, so class A loadline of 2k8, and class AB1-2 loadline of 1k4.

Idle bias of 41+40mA, 22.5V, 473V, 19.3W. Vak lower than VS1 by 10+55Ω x 0.04=2.6V. VS2 250V screen sag may peak at 7V across 135Ω 6W screen stopper if screen current gets to 50mA, but VS2 likely to sag to.

Just starting to clip with f/b at 16.5V 34W, with VS1 sagging from 464V to 415V.

Bridging 5V4 anodes for VS1: Idle bias of 36+37mA, 474V, 17.5W. Clip with f/b at 17.0V 34W, with VS1 sagging from 474V to 438V.



807 loadline with 5k6PP. Screen 250V.

4.4 Power Supply – secondary DC side – TEK PT

Plate HT 383-0-383V 0.18A (was 450-0-450V):	29-17-30
Screen-driver HT 213-0-213V 0.09A (was 270-0-270V):	14-15-16
5V heater loading – 5V4G : 2A each.	27-28; 12-13.
6.3V heater loading: 2x 0.9A, 4x 0.3A, 1x 0.2A, 1x 0.45, 1x 0.32A= 4A.	18-19
Not used: 5-6, 22-23, 24-25, 10-11-21.	

The power supply generates two HT supplies from two PT winding groups. Each supply is a full-wave rectified type using double diode 5V4G, with added protection ss diodes, and capacitor input filtering. Plate supply is CT fused. Screen supply has no fuse, and is diode connected to Plate supply to avoid only screen supply.

The plate HT supply has an effective input resistance of the transformer is about 5.5 x $(383/240)^2$ + 44Ω = 58Ω. Two 1N4007 in series with each anode to allow 383VAC supply (375Vac max rating of 5V4). VS1 is about 500VDC with 60mA idle with 21Vpp ripple, sagging to 470V at 120mA with 41Vpp ripple and 0.52Apk continuous diode (=max rated) and winding current approaching rated 0.18A. 20uF first filter capacitance has 120mA load hot-start peak current of 1.9A (so below 3.5A max rated). CT with 220mArms at max 120mA continuous loading.

VS1 fuse in CT; idle load hot-start; 0.22Arms max load continuous IEC F fuse

Simulate period in PSUD2	10ms	50ms	continuous
Simulated RMS current	1.1A	0.54A	0.22A
Multiplier (based on 0.315A fuse rating)	3.5	1.7	0.67
IEC60127-2 Quick-acting F min limit multiplier	4	2.75	1

VS1 fuse in CT; idle load hot-start; 0.22Arms max load continuous IEC T fuse

Simulate period in PSUD2	20ms	150ms	600ms	continuous
Simulated RMS current	0.77A	0.33A	0.17A	0.22A
Multiplier (based on 0.25A fuse rating)	3.1	1.4	0.7	0.84
IEC60127-2 Quick-acting F min limit multiplier	10	4	2.75	1

The 0.25A T is preferred as it has a lower continuous current rating.

The screen/driver/preamp supply VS2 provides about 50mA idle, but may increase for AB2 operation. 807 screen current is likely <5mA each; driver idle current is about 35mA; preamp currents total to about 5mA. Using the 220-0-220V PT winding, the total output is about 260V, of which about 20V is used for back-bias of the output stage, and so VS2 was about 240V. A 3.3H choke has been added to significantly filter ripple from the screen/driver voltage, and the output stage bias voltage. Winding current is well under 100mA.

The back-bias VS5 is generated by the switchable Zener voltage setting of about 22V (17.5V min, up to 10+7.5+ 7x 1N4004 in series) in series with 10Ω to 0V. This circuit replaces the original 500Ω dropper to the VS5 R//C (5kΩ//16uF).

Heater loading is 2A on 5V 3A winding; 2x0.9+0.3=2.1A on 6V3 3A winding; and 0.3+0.3+0.3+0.3=1.2A on 6V3 3A winding.

Idle loading from VS4 is ~1.2+1.2=2.4mA; VS3 is ~ 2.4+1.3+1.3=5mA; VS2 is ~ 5+35+5+5=50mA, (24V drop from VS3 to VS4 via 10k; 24V drop from VS2 to VS3 via 4k7).

Supply rail bleeds: 360k from VS1 monitor; 150k from VS2 monitor.

4.5 Power supply - AC primary side

Primary fuse: initially use IEC 2A T time-delay (VBS UTE 250V) – T2AL250V.

4.6 Protection

2x series VE13 0421K MOV across each OT half-primary – 2x612=1225V 1mA rating.
1x VE13 0421K MOV across interstage driver primary – 612V 1mA rating.
Series diodes with each 5V4 anode.
Secondary side CT fuse.
Steering diode from 807 Screen supply (VS2) to Anode supply (VS1).

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https://www.diyaudio.com/community/threads/807-unstable-operation-related-to-class-ab2-and-negative-resistance-region.438631/#post-8256999

I am revisiting a vintage Philips 952 PA amp (from circa 1947) that I basically restored 8 years ago. This has an 807 PP output stage meant to operate into class AB2 for up to 75W output, and initial testing had certainly shown clean output to 45W and clipping output to 60W. It uses a 6V6 SE driver stage through an inter-stage transformer for 807 PP grid driving in pentode mode, along with feedback from a separate output transformer winding back to the stage before the driver. So my focus now is to revisit the feedback, and managing the driver and 807 grids to avoid any unstable operation. [The amp was described in the Jan 1947 Philips Technical Communication - a scan at Kevin Chant's website - https://www.kevinchant.com/philips4.html, and a schematic with as-found part values is here: https://dalmura.com.au/static/Philips 952 amp schematic.pdf].

Inter-stage transformer drive/PI has not been in common use since the 1940's for PA or hi-fi application. Vintage books/articles discuss the risk of unstable operation from parasitic oscillation in an 807 PP output stage from operation in the negative resistance region due to the secondary emission effect within the 807. Vintage references, and typical advise is that grid and screen stoppers, and perhaps even anode stoppers, suppress the 807 operating characteristic in that negative resistance region and typically avoid any operating instability. That's all fine, as I have been careful to use stoppers for the other four 807 PP related audio amps I've restored, although all those amps used more typical 'modern' PI stages. One good discussion of stoppers is David Newkirk's article at https://dpnwritings.nfshost.com/ej/beam_power_tube_parasitics/.

However the topic of using an interstage transformer for class AB2 operation is interesting, so I started digging around for articles on not just basic ways to design the inter-stage transformer, but also how to manage transformer resonance and extend frequency response (eg. out past 7-10kHz). McLean 'An analysis of distortion in class B amplifiers' Proc IRE March 1936 (in worldradiohistory.com) indicates that the inter-stage transformer leakage inductance, and any capacitor related tuning of the 807 grid circuit (to suppress resonance peaks) may benefit from loading the 807 grid-grid (pp.502) - and the Philips 952 used a 10k loading resistance, but no 807 grid capacitor tuning, but it did use a 6V6 anode bypass cap. The RCA 1937 application note AN-68 'A 55-Watt Amplifier Using Two Type 6L6 Tubes' provides a good example of tuning the RC across the 807 PP to flatten speaker load and output transformer related response interaction. So I will certainly be looking at those aspects for compensating frequency response and avoiding resonances on the way to adding back in some global feedback.

But McLean also pointed to an earlier Sept 1935 Proc IRE article by Fyler that shunts the 807 grid-cathode with a diode-resistor circuit, so as to pass current within the negative resistance region, and so change the resistance characteristic away from a negative slope - somewhat similar to what stoppers can achieve. Back in 1935 there were no miniscule 1N4004's to retrofit, so Fyler used a Dynatron rectifier - which at that time was a Kenotron (basically a vacuum diode tube). So perhaps I'll try and see if I can induce unstable operation, and then see if that can be managed with an 807 grid-cathode shunt diode-R circuit.

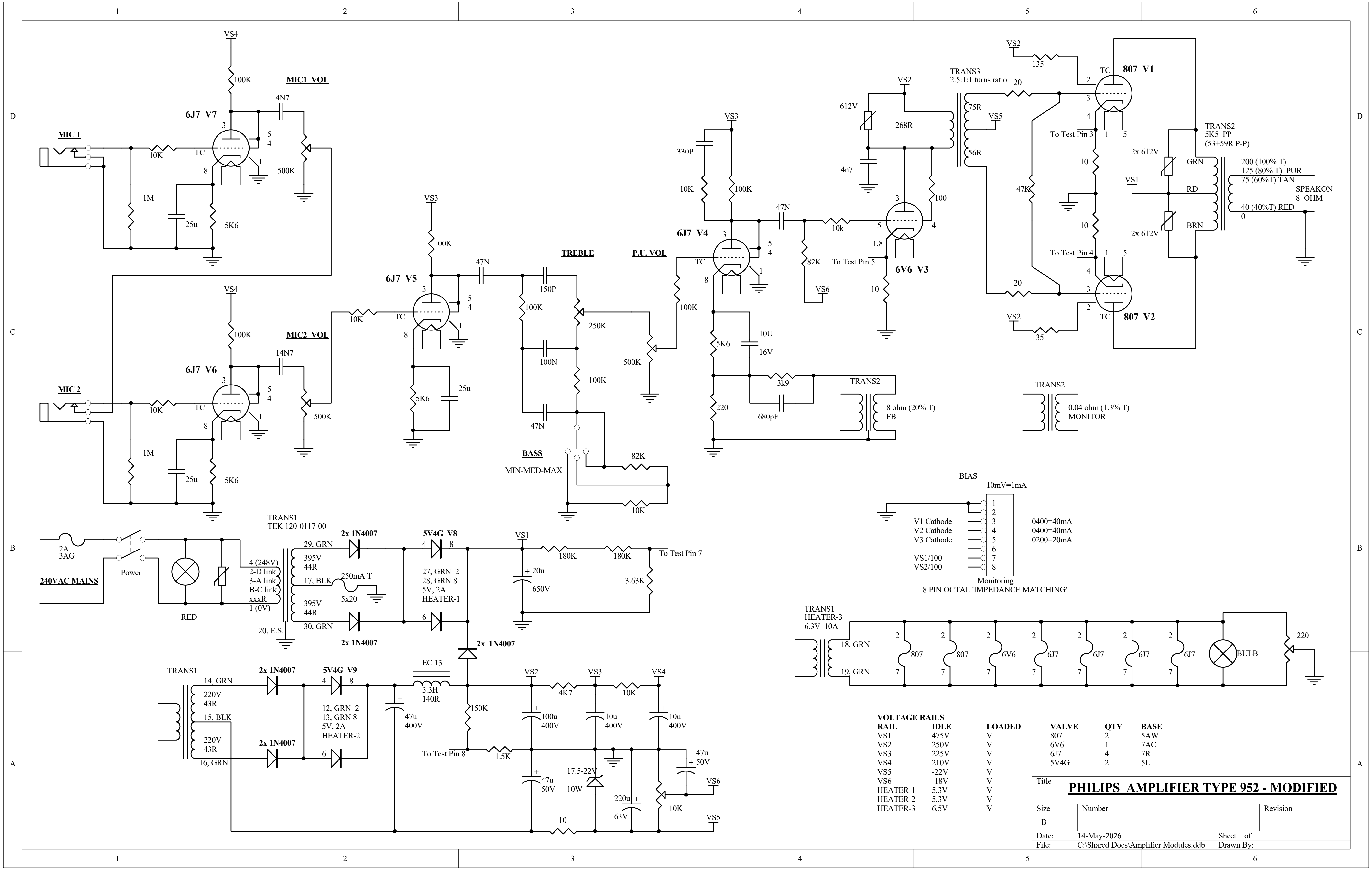
I've done a quick assessment of feedback stability based on the attached modified schematic, and testing with 8 ohm, 16 ohm, and no load using 6dB of feedback, and low signal levels. 10nF shunt across driver V3 anode to cathode gave a flat frequency response out to 30kHz with 8R load, and no feedback showed the response dropping from about 3kHz due to the 10nF. There is a dominant resonance at circa 40kHz, and another at 100kHz, with nothing observable beyond that out to 1MHz. The 40kHz resonance starts to peak with lower loading, rising to about +15dB with no load. That peak can be suppressed somewhat with feedback cap comp, and with some zobel loading, but for now was left as is.

I then set up a short tone burst to see if I could elicit instability. I used REW's tone burst, as I'd found that convenient a few years back doing overload testing on a PP amp that forced the output transformer half-primaries into over-voltage transients to test MOV protection. But instead of using a digital TEK scope, I'm using a Picoscope 4224A to acquire the amp output waveform in search of instability. The test setup works, with the first plot example below showing regions of instability when the output is driven into clipping territory for a 15 ohm load. Blue trace is input signal (to top of Volume pot, with preamp stages disconnected), and red trace is speaker output level (16 ohm load, with clipping above about 30Vpk or 28W continuous). Instability is circa 40kHz ring, using zoom, and active for just a portion of the 150Hz tone cycle, but starting as one 807 is recovering from operation in saturation. There is asymmetric clipping.

Some more testing has shown that the AB1 transition is quite imposing on operation, as the output transformer dominant resonance is energised on the half-primary winding of the 807 entering cut-off. The primary half-winding leakage inductance is likely much more than typically found in later audio amps, although I haven't measured it yet. The discontinuity in the half-winding current, as it hits zero, then transfers dI/dt energy into the other half-winding etc.

Subsequent testing indicates one of the 807's was initiating a ringing disturbance as its cathode current transitioned to and from zero - as seen in post #9 plots. Prior to that, I had tweaked the 6V6 driver stage to make sure it wasn't limiting the output stage from being pushed into class AB2. Relocating the 807's in the other's socket stopped any such disturbance - not sure why.

So I then tried to make AB2 operation initiate any disturbances or instability by effectively shorting the screen stoppers, and shorting the grid stoppers, but nothing was observed.



VOLTAGE RAILS

RAIL	IDLE	LOADED
VS1	475V	V
VS2	250V	V
VS3	225V	V
VS4	210V	V
VS5	-22V	V
VS6	-18V	V
HEATER-1	5.3V	V
HEATER-2	5.3V	V
HEATER-3	6.5V	V

VALVE	QTY	BASE
807	2	5AW
6V6	1	7AC
6J7	4	7R
5V4G	2	5L

Title
PHILIPS AMPLIFIER TYPE 952 - MODIFIED

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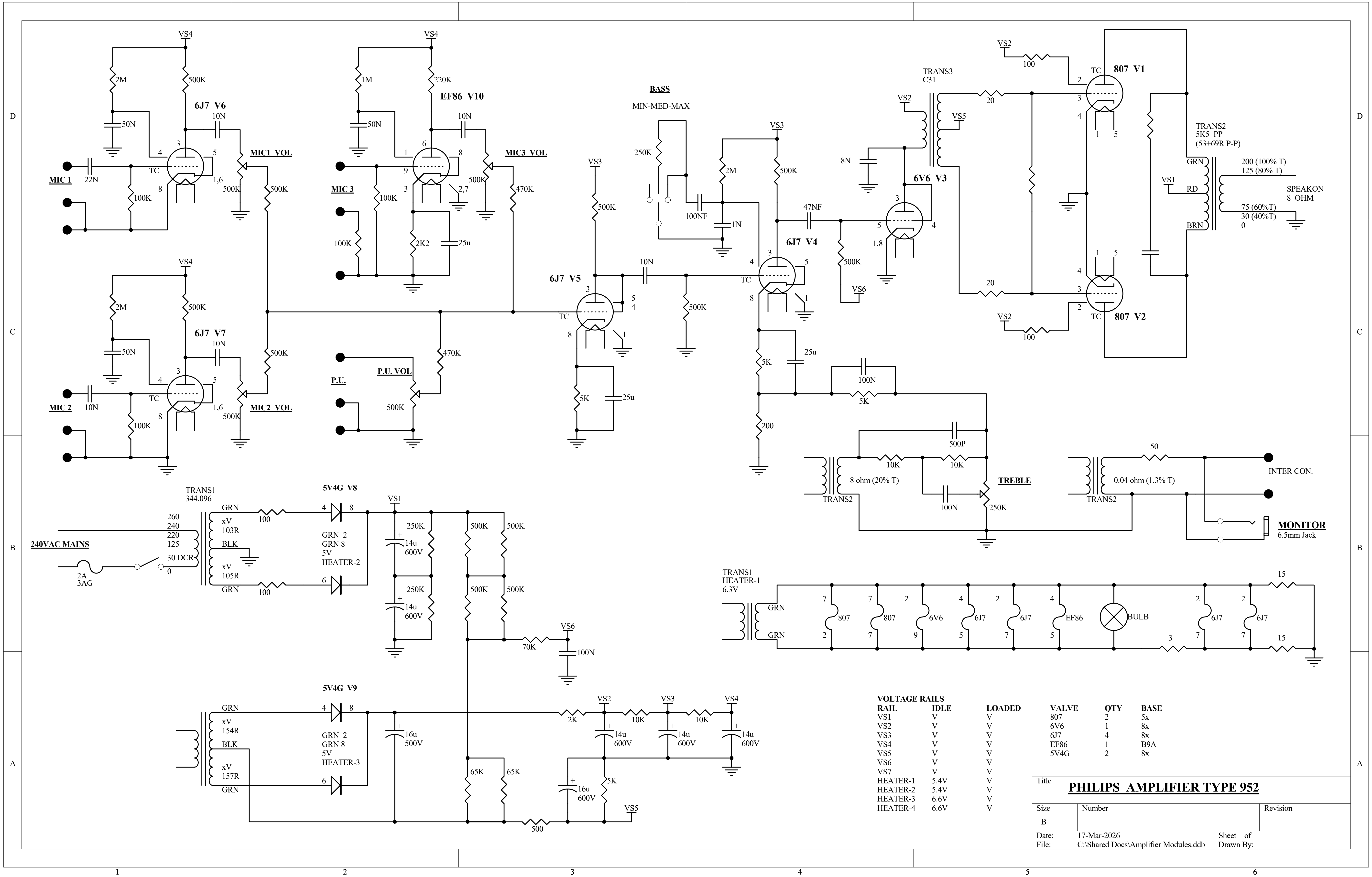

VOLTAGE RAILS

RAIL	IDLE	LOADED
VS1	V	V
VS2	V	V
VS3	V	V
VS4	V	V
VS5	V	V
VS6	V	V
VS7	V	V
HEATER-1	5.4V	V
HEATER-2	5.4V	V
HEATER-3	6.6V	V
HEATER-4	6.6V	V

VALVE	QTY	BASE
807	2	5x
6V6	1	8x
6J7	4	8x
EF86	1	B9A
5V4G	2	8x

Title: **PHILIPS AMPLIFIER TYPE 952**

Size	Number	Revision
B		

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CONTENTS

CORRECTIONS

December 1946. No. 7. Page 16.

Fig. 2. Delete the words "PBF Beam forming plates."

Please read the last six lines of the first paragraph on page 18—Valve Type PE04/15A as Plate Modulated R.F. Power Amplifier as follows:—

DC Cathode Current	55	—	mA
Total Effective Grid Circuit Resistance	1	—	Megohm
Plate Dissipation	9	7.5	watts
Driving Power (Approx.)	—	64	milliwatts
Power Output (Approx.)	—	8	watts

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CONTENTS

A MODERN A.F. POWER AMPLIFIER SYSTEM

Performance figures are given for a typical modern audio frequency power amplifier type 952.

A constant voltage loudspeaker distribution system is described which enables the maximum power output of the individual speakers to be pre-selected by the choice of the speaker transformer impedance.

A new circuit arrangement for utilising the driver tube to stabilise the high voltage supplies of a class AB2 or B2 types by means of a degenerative d.c. network is described and its method of functioning explained.

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A Modern A.F. Power Amplifier System

1. Introduction.

While it is probably correct to say that the fundamentals involved in the design of power amplifiers and loudspeaker systems have altered very little during the past ten years or so, modern designs for power amplifiers incorporate refinements and facilities, developed as a result of experience, which render a modern power amplifier capable of a performance and universality of application which are amazing.

Perhaps the most important of the improvements introduced during the past few years are the use of negative feedback circuits, beam and high slope pentode valves, voltage or current stabilising circuits, tone control circuits and perhaps, last but not least, a system of wiring a number of loudspeakers which minimises the problems of impedance matching and at the same time reduces the danger of shock from the circuits external to the amplifier.

2. Typical Modern Power Amplifier

An appreciation of the practical improvements which have taken place in power amplifier design are best illustrated with the aid of a typical amplifier and for the purposes of this paper an amplifier available to the general public has been adopted [1]).

The performance figures for a typical power amplifier of this type are given hereunder:—

Power Output and Distortion

Measured at 1,000 c.p.s. with bass and treble controls set to give a flat frequency response curve.

Output	Distortion
1 watt	1%
10 watts	1%
25 watts	1%
50 watts	1.7%
60 watts	4.3%
70 watts	7.6%
75 watts	15%

[1]) Amplifier Philips Type 952

Some of the distortion at low outputs was inherent in the beat frequency oscillator used for making the measurements. A measurement made at 100 c.p.s. gave 50 watts for 4.3% distortion and at 5,000 c.p.s. the amplifier gave 50 watts for 1.3% distortion.

Sensitivity.

Measured at 1,000 c.p.s. with all controls at maximum:—

Microphone 1 ...... 250 microvolts input gives 50 watts output.

Microphone 2 ...... 250 microvolts input gives 50 watts output.

Pickup ...... 0.15 volts input gives 50 watts output.

These sensitivities correspond to a gain of about 90 dB on pickup and 156 dB on microphone.

Frequency Response.

Response curves are shown in figs. 1 and 2. The high frequency response is deliberately attenuated as quickly as possible after 7 Kc/s, as higher frequencies do not improve intelligibility, but make distortion sound extremely unpleasant if a good speaker is used.

Hum and Noise.

Figures from the microphone channel vary considerably with the input valve used, but average figures with all tone controls giving maximum response are:—

All gain controls at zero, output = 0.1 volt = 0.05 milliwatt = 60 dB down from 50 watts.

Pickup control at maximum, output = 0.1 volt = 0.05 milliwatt = 60 dB down from 50 watts.

One microphone control at max. — with short circuited input — output = 2.5. volts = 31 milliwatts = 32 dB down from 50 watts.

One microphone control at max. — with open circuit input — output = 3.2 volts = 50 milliwatts = 30 dB down from 50 watts.

Much better hum and noise figures can of course be obtained with the tone controls turned down.

Power Consumption.

Varies with output as follows:—

Output	Power Consumption
0 watts	115 watts
50 watts	200 watts
75 watts	225 watts
maximum	240 watts

Input Arrangements.

Pickup terminals are provided at the back of the amplifier and jacks for two independent microphones are on the front panel.

Output Arrangements.

A plug type switch, which is on the rear of the unit adjusts the output impedance to either 200, 125, 75 or 30 ohms. A monitoring jack is provided on the front panel for a pair of headphones. Interconnection terminals are also provided at the back of the amplifier for feeding additional amplifiers if required.

Controls.

Independent gain controls are provided for both microphones and for the pickup. In addition, separate bass and treble controls are fitted. A circuit diagram of this amplifier is given in Fig. 3.

3. Loudspeaker System.

In order to obtain the maximum undistorted output which the amplifier is capable of delivering, the load driven by the amplifier should be matched to the impedance indicated by the output impedance, adjusting plug switch in accordance with common practice. The highest output impedance is 200 ohms and this value has been selected for reasons explained in the following.

When an amplifier is in operation the maximum voltage which can develop across the output terminals for a given power output is determined by the output impedance of the amplifier and, provided that an adequate degree of negative feedback is used in the amplifier, this voltage will not be materially increased by an increase in the impedance of the load to which the amplifier is connected. The following table gives the maximum voltage at the output terminals of an amplifier matched to deliver 50 watts into loads of various impedances, calculated from the well known expression, Watts = E^2/R.

Impedance	Maximum Voltage
100 ohms	77
200 ohms	100
500 ohms	158
1,000 ohms	223
5,000 ohms	500
10,000 ohms	770

Voltages in excess of 100 volts should be regarded as dangerous to human life and should only be permitted if the wiring carrying the voltages is very carefully insulated and installed. It is a fact that many loudspeaker installations are of a temporary nature and quite often the wiring to the loudspeakers is most unsuitable for carrying high voltages. Furthermore, many radio mechanics are accustomed to regard the voltages delivered to loudspeakers as harmless and would not hesitate to connect a loudspeaker to the wiring while the amplifier is switched on.

By limiting the output impedance of this amplifier to 200 ohms the designers have limited the maximum voltage across the output terminals to the comparatively safe value of 100 volts and thereby protected users of the amplifier from the risk of fatal accident through carelessness.

This limitation of output voltage has another very convenient consequence. The maximum power which can be delivered to any loudspeaker connected to the output terminals is limited by the impedance of the speaker transformer. This not only reduces the risk of damaging a loudspeaker by overloading it, but enables the maximum volume which can be produced by any loudspeaker to be limited by using a transformer of suitable impedance. For example, if a volume of sound corresponding to 2 watts is required from a loudspeaker in a particular location and 4 watts in another, the two speakers may be wired in parallel to the output terminals of the amplifier and if one of the speakers has an impedance of 5,000 ohms and the other an impedance of 2,500 ohms, their power will be limited as desired. Loudspeakers can therefore be rated according to power output by means of the impedance of their transformers.

Suitable impedances for various power ratings of loudspeakers are given in the following table:—

Power	*Impedance*
0.25 watt	40,000 ohms
0.5 watt	20,000 ohms
1.0 watt	10,000 ohms
2.0 watts	5,000 ohms
5.0 watts	2,000 ohms
10.0 watts	1,000 ohms

4. Impedance Matching and Frequency Response.

Communication engineers generally regard it as axiomatic that a transmission line must be matched to its load if frequency selection by the line is to be avoided and at first it may appear that the system of wiring loudspeakers described in the foregoing section would adversely affect the frequency response characteristic of the whole system, but this is not so in normal installations.

It is only necessary to match the impedance of a transmission line to avoid reflections if the length of the line is comparable with one quarter of the wavelength corresponding to the highest frequency to be considered. In this case the highest frequency is 7,000 c/s and the corresponding quarter wavelength is approximately 27 miles so the effects of reflections may be disregarded for normal installations.

When audio frequency currents are transmitted over short lines the higher frequencies suffer from attenuation more than the lower frequencies mainly as a consequence of the total capacitance of the cable. The circuit arrangement resembles a resistance capacitance filter in which the resistance is the parallel impedance of the amplifier and the load while the capacitance is that of the cable. The resistance is likely to be less than 100 ohms and with this value for a 3 dB drop at 7,000 c/s, the capacitance of the wiring would have to be approximately .25 μF and such a value is improbable in a normal amplifier installation.

When only a few high impedance speakers are connected to the amplifier the output transformer will be mismatched. However here again no difficulties arise. Negative feedback in the amplifier keeps the gain constant and minimises distortion in the amplifier due to mismatch. So far as frequency distortion due to the transformers is concerned, it should be noted that the output transformer is coupled to a load higher than the correct matching value while the loudspeakers are connected to an input load lower than the correct value.

Frequency distortion in a transformer due to mismatch of the secondary impedance when the primary is loaded only occurs when the output load resistance is too small for correct matching. In such a case the leakage inductance of the transformer leads to attenuation of the higher frequencies. In the case under consideration the output load impedance is too large, and not too small, so no difficulties arise.

With loudspeakers the position is the reverse. If the input load impedance is too large the speaker will be insufficiently damped and attenuation of the lower frequencies may occur, but with this system the input load impedance is too small and not too large so again the arrangement is satisfactory.

The only loss due to the mismatch is that the maximum undistorted output of the amplifier will be reduced, but if the impedance of the speakers is selected to reduce the power handled by the individual speakers, this is not a disadvantage. The voltage of the output is approximately constant regardless of the load. On the other hand if the total impedance of the loudspeakers connected to the amplifier becomes less than the output impedance of the amplifier, distortion will be increased and the amplifier will be unable to deliver its rated power to the speakers. The arrangement will then be one which exceeds the power capacity of the amplifier and it will be necessary to reduce the output impedance of the amplifier in order to match the load. This will reduce the maximum voltage across the output terminals and the power to all speakers will be reduced in the same proportion.

5. Individual Volume Control of Loudspeakers.

Very few loudspeaker installations will require the full power which the amplifier is capable of delivering and this enables a very simple method of individually adjusting the volume of sound from the speakers to be used. Each speaker may be fed from a simple potentiometer connected across the output line and having a resistance equal to about twice the impedance of the speaker it is intended to control. The potentiometers may be used to control the volume from individual loudspeakers without noticeably affecting the volume from the others.

Such a system is very convenient for a house or building having many small rooms. A single radio receiver together with microphones and a pickup if desired, may be used to feed the amplifier which in turn drives a loudspeaker in each of the rooms. The impedance of the speakers is selected according to the size and noise level of the rooms and individual volume control potentiometers may be fitted.

6. Audio Frequency Negative Feedback and Tone Controls.

Negative feedback for the amplifier is derived from a separate secondary winding on the output transformer T3 (Fig. 3) of the amplifier and applied to the cathode of the valve V1D. This is a circuit arrangement much favoured in equipment designed by Philips in preference to feedback from the anode of the output valve. Not only does this feedback reduce distortion due to non-linearity of the valve characteristics but it also reduces distortion due to non-linearity or frequency selection of the output transformer. Furthermore, as clearly explained in a recent article in another journal [2]) feedback from the secondary of the output transformer reduces hum from the power supply while feedback from the anode of the output valve can actually increase the hum. (It will be seen from the circuit diagram that no smoothing chokes are used in this amplifier. See under sub-heading 11). On the other hand, the design must be carefully carried out as the phase shifts in the transformer must be taken into account if instability is to be avoided. With care use can be made of this phase shift to give bass boost to the amplifier.

Included in the feedback circuit is a network of resistances and capacitances including the treble tone control potentiometer R10A. Feedback at very high frequencies is maintained for all adjustments of the tone control by means of the capacitors C7A and C2C. Resistor R6B is so proportioned with respect to capacitor C2C that the feedback is reduced at the very low frequencies. The feedback at the medium higher frequencies can be attenuated by the tone control potentiometer R10A and capacitor C1H. When the arm is moved to the top of the potentiometer, capacitor C1H bypasses resistor R6C and increases the feedback at the higher frequencies thus attenuating the treble response of the amplifier. On the other hand, when the potentiometer arm is moved towards earth, this capacitor C1H bypasses the higher frequency feedback potentials to earth, reduces the feedback at these frequencies and so elevates the treble response of the amplifier.

[2]) "Wireless World," May 1946.

A curve showing the frequency characteristics of the feedback circuits is given in Fig. 4 for a condition when the treble and bass tone controls are set for maximum response and the attenuation of the feedback at about 30 and 5,000 c/s is clearly shown. This curve should be compared with the corresponding response curves of Figs. 1 and 2 and the impedance curve of Fig. 10.

When designing an amplifier such as this the designer is faced with the choice of placing tone controls either in the feedback path or outside that portion of the amplifier straddled by the feedback. If they are outside the feedback path, the total gain of the amplifier with feedback functioning must be sufficient to give adequate gain to the frequencies attenuated by the tone control, and adjustment of the tone controls will cause a noticeable change in the volume of sound. The effect of an increase in gain at certain frequencies is obtained by reducing the gain at others. If located in the feedback path, the tone control circuits vary the gain of the relative frequencies themselves. The effect of the tone control on the apparent volume is reduced and also the number of stages necessary in the amplifier is reduced because the gain of the amplifier is at a minimum for all frequencies when the tone controls are adjusted for a flat frequency response.

7. Bass Tone Control.

In this typical amplifier, bass control is effected by varying the size of the screen grid bypass capacitor by means of the switch S1A. In effect this is equivalent to a form of feedback at the lower frequencies applied through the screen of this valve. The screen voltage is permitted to change in synchronism with the lower frequencies applied to the control grid of the valve but in opposite phase and this reduces the effective slope of the valve at these frequencies.

8. Output and Driver Stages.

The output stages of this amplifier use two 807 beam power valves (V4A and V4B) driven by a 6V6 valve (V3A) per medium of an interstage transformer T2A. The output valves operate in class AB_2.

In addition to acting as a driver valve the 6V6 valve V3A provides negative feedback for d.c. and so minimises changes in the d.c. potential applied to the screen of the output stages.[3]) This action, which largely contributes to the overall performance of the amplifier is explained in more detail in section 11.

[3]) Devised by Messrs. G. Smith, W. Storm and E. Watkinson.

9. Microphone Input Valves.

The microphone input valves are of the 6J7 type connected as triodes. As the input voltage derived from a microphone is very small it has been possible to bias these valves by means of a very large grid leak. When a valve is left with its grid insulated the grid accumulates a negative charge and the potential of the grid can then be controlled by means of the grid leak. If the resistance of the grid leak is sufficiently large the control grid is biased to a value a little less than that produced by the charge accumulated on the isolated grid.

10. Volume Control and Fading Circuits.

The microphones are fed directly to the grids of the microphone input valves and their respective volume controls are located in the outputs of these valves. This permits a very high input impedance to be maintained at the microphone terminals so that crystal microphones may be used if desired.

The pick-up terminals feed the pickup volume control directly. The outputs of the three volume controls are connected to the control grid of the valve V1C via isolating resistances so that any volume control may be adjusted without affecting materially the adjustment of the others.

11. Power Supply Circuits.

Perhaps the most interesting feature of this amplifier from a technical point of view is the power supply and voltage stabilising circuits. Two rectifiers are used, one supplying the anodes of the output stages and the other the screens of the output stages and also the anodes and screens of the driving and voltage amplifying stages. The two power supplies are interconnected in such a manner that the driving valve controls the bias of the output stages in such a manner as to compensate for changes in the screen potentials. No smoothing chokes are used as the overall design is such that these are unnecessary and power which would otherwise be dissipated in chokes is saved.

A simplified form of the voltage stabilisation circuits is shown in Figure 5 and it will be seen from this figure that the driver stage power supply is linked to the power supply for the anodes of the 807 valves by a direct current connection from its control grid to the junction of two resistors R3 and R4 which connect the positive terminal of the 500 volt supply to the negative terminal of the 300 volt supply. The earth circuit of the 300 volt supply is made via the resistors R2 and R1, the voltage across the latter being used to bias the output valves. The bias applied to the 807 valves therefore depends upon the current flowing through the 6V6 driver valve and this current in turn depends upon the voltage of the main 500 volt supply because of the d.c. connection to its control grid. If the voltage of the main supply falls the potential applied to the grid of the driver valve will move in a negative direction thus decreasing the current through this valve and so decreasing the bias applied to the output 807 valves. If the voltage of the 500 volt supply increases the opposite action takes place and the bias to the 807 valves is increased.

The regulation of the 500 volt supply is not perfect and when the output valves commence to function as class B amplifiers, and their consumption of d.c. anode current increases, the effective voltage of the 500 volt supply will fall. In the absence of any compensation network, the output power of the stages would also be decreased. Due to the network the bias voltage also changes and so operating conditions are maintained near to those for maximum output.

Similarly the network also adjusts the screen voltage of the 807 valves to assist in compensating for a fall in anode voltage. The effective screen voltage is the voltage of the 300 volt supply less the voltage drop in the resistors R1 and R2. This voltage drop is partially determined by the current through the driver valve and preliminary stages and partly by the screen current itself. When the voltage of the 500 volt supply decreases the current through the driver 6V6 also decreases thus decreasing the voltage drop in the resistors R1 and R2 and thereby increasing the screen voltage which is 300 volts minus the voltage drop.

The curve in Figure 6 shows the relationships between the bias applied to the 6V6 bias voltage and cathode current and the output voltage of the amplifier. Figure 7 shows a similar relationship between the anode voltage and current of the 807 valves and the voltage output.

Figures 8 and 9 show the relationships for 807 screen current and bias and also the low voltage supply current and voltage (to ground) with respect to output voltage. The extent of the compensation caused by the circuit can be gathered from these figures.

12. Output Impedance.

The output impedance of an amplifier with feedback is largely determined by the magnitude of the feedback and when the frequency response of an amplifier is controlled by means of negative feedback the output impedance of the amplifier varies with frequency. The shape of this curve will depend on the adjustment of the tone controls of the amplifier and is illustrated in Fig. 10.

—*E.G.B.*

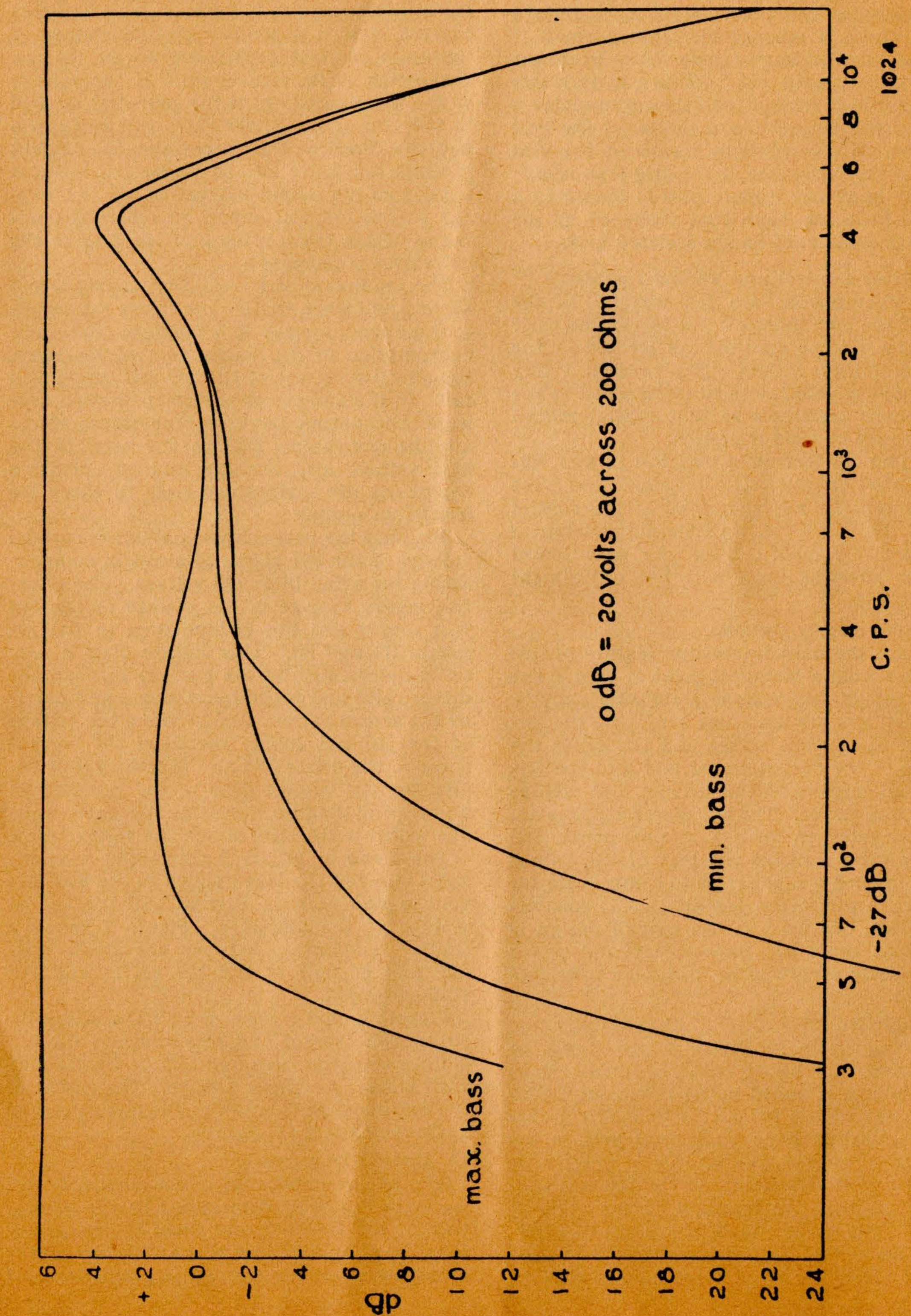

Fig. 1. Microphone response

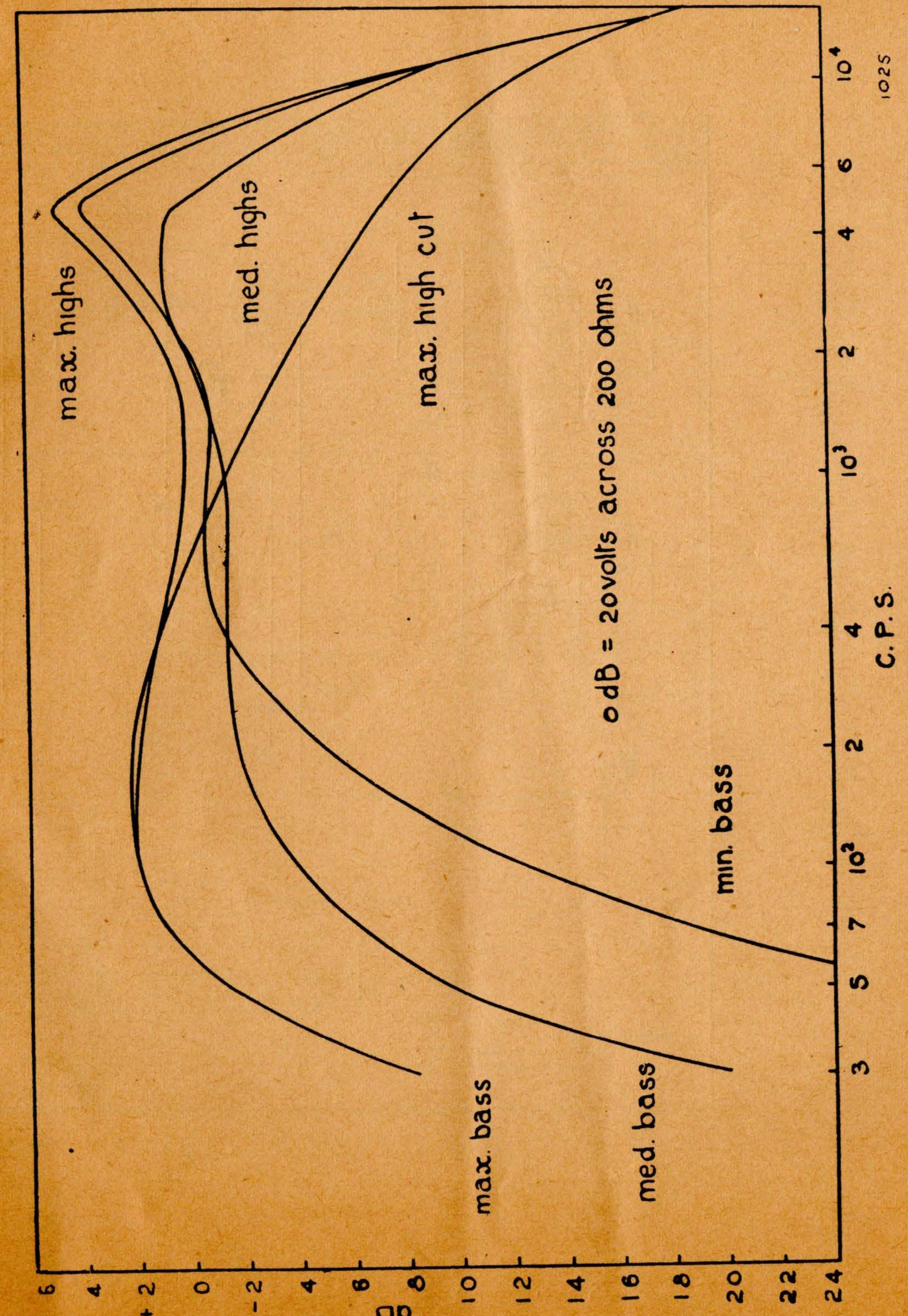

Fig. 2. Pickup response

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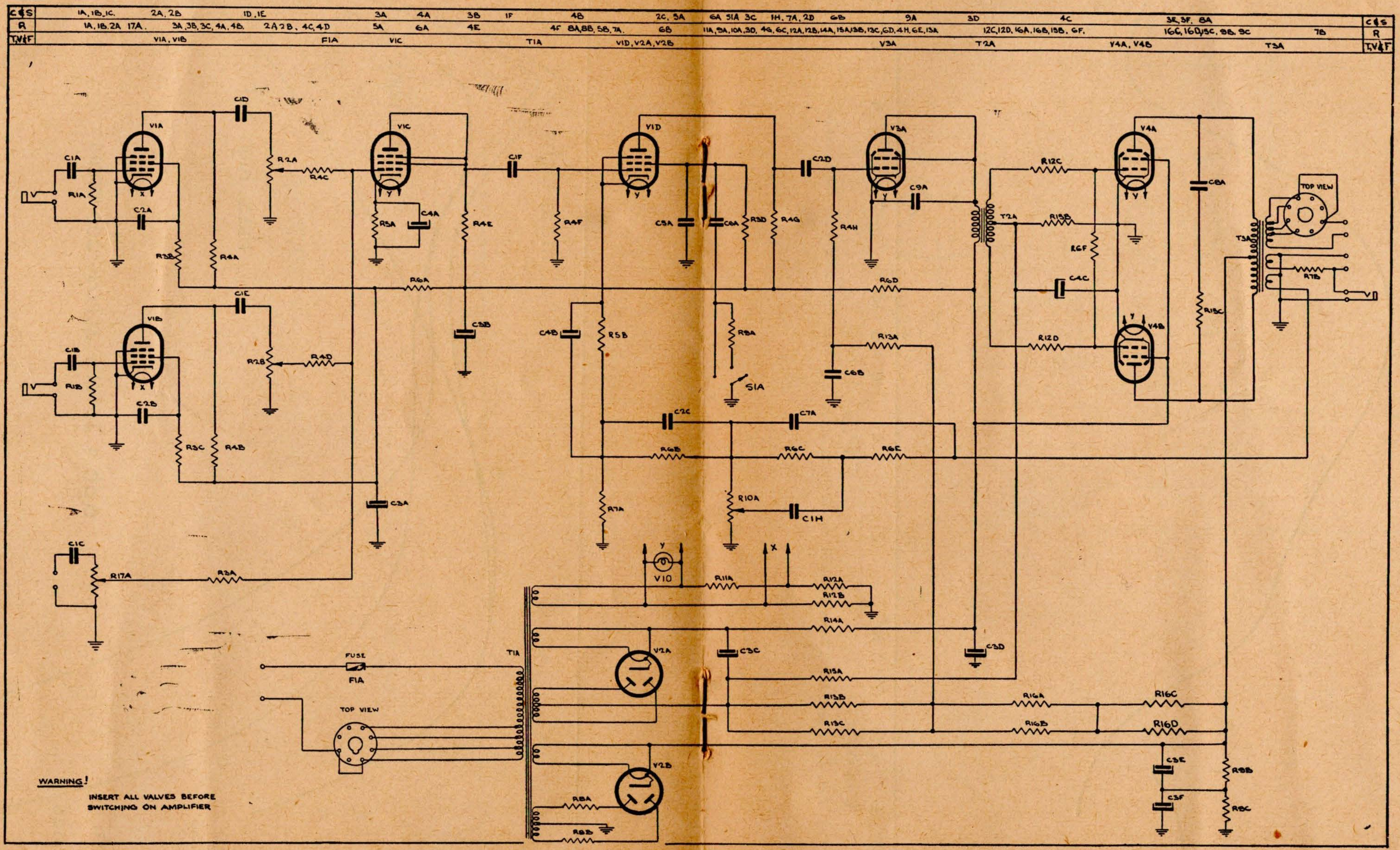

Fig. 3. Circuit Diagram of 952 Amplifier.

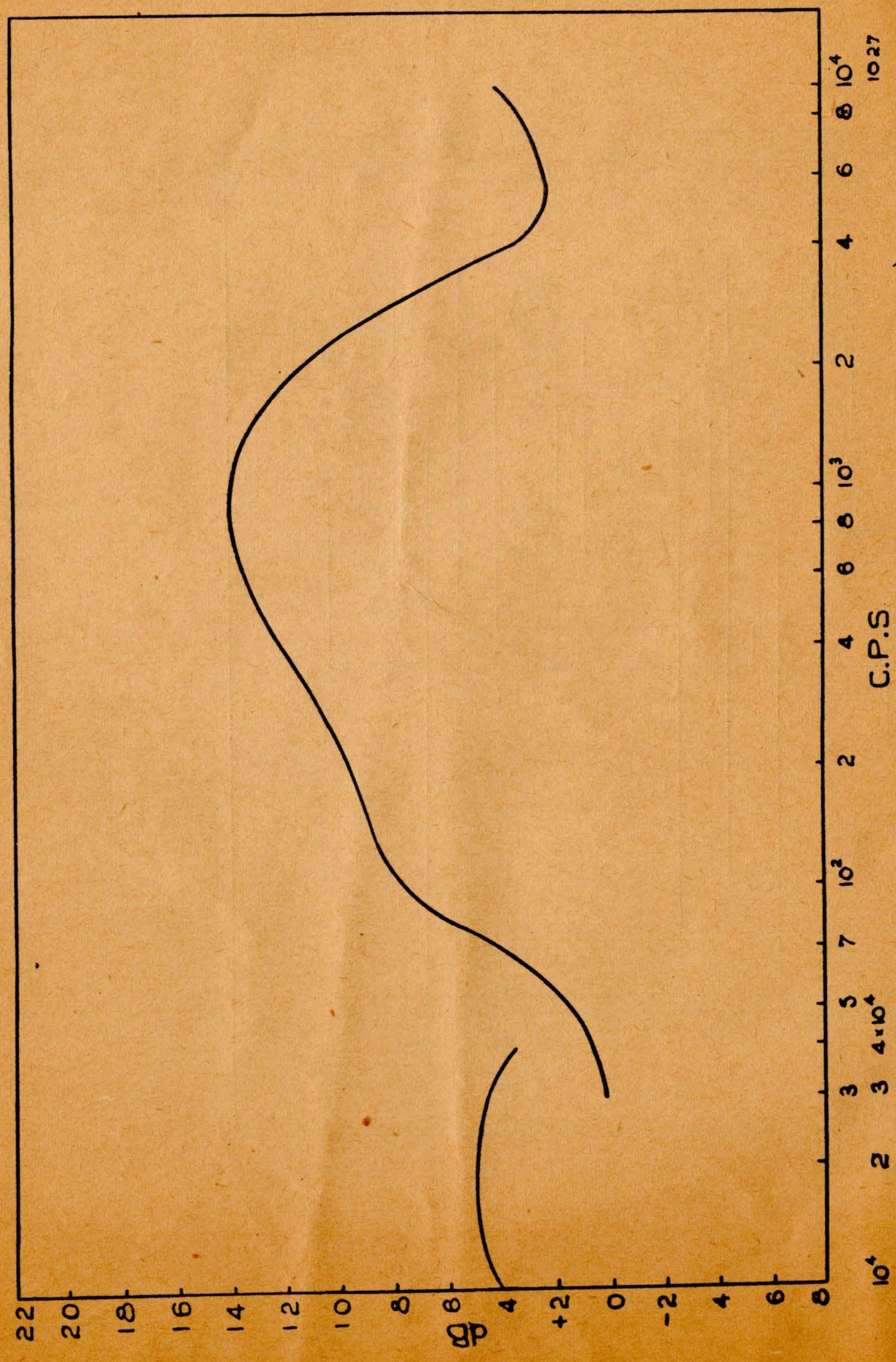

Fig. 4. Feedback (Tone control set for maximum high and low frequency response).

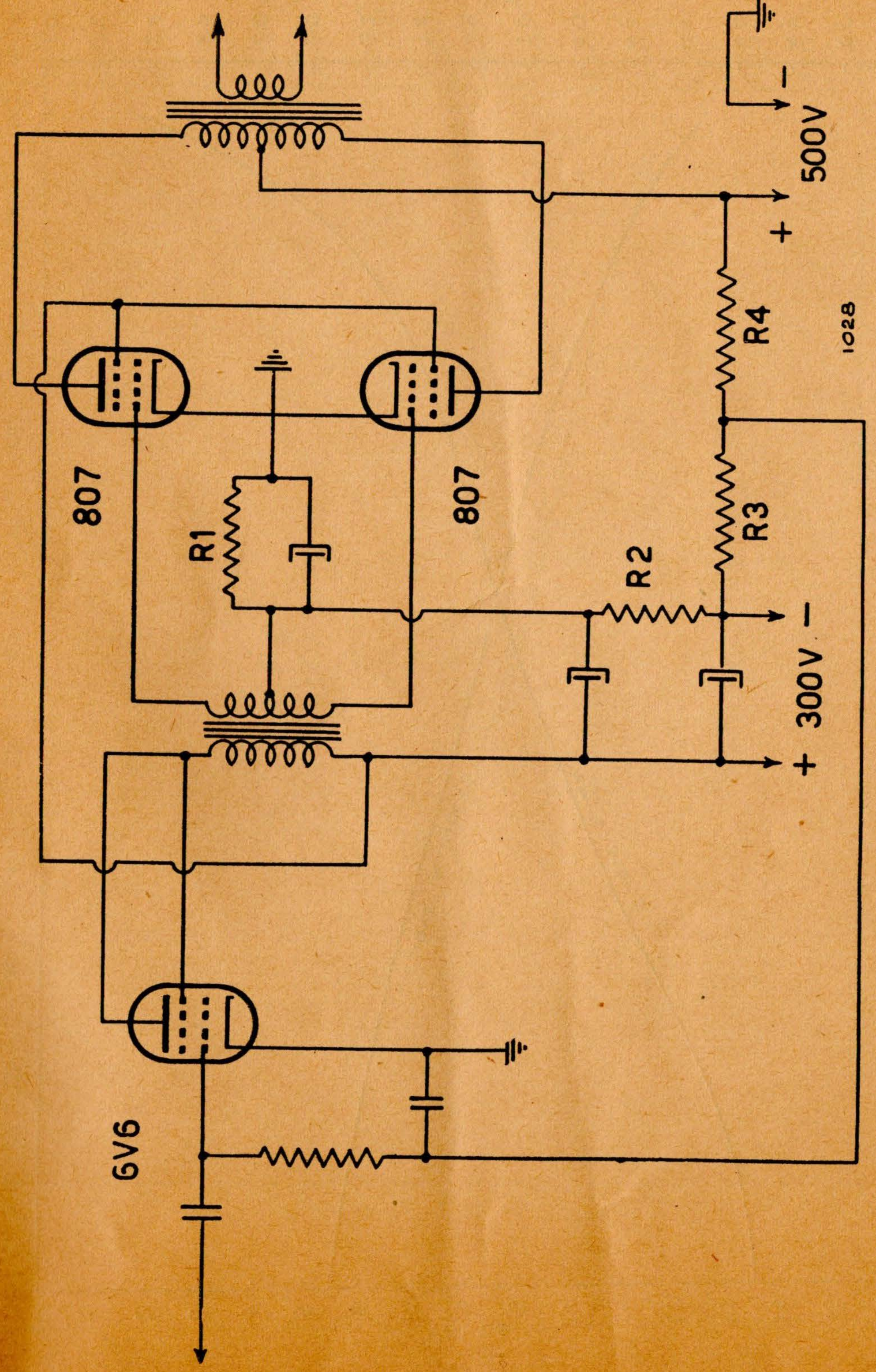

Fig. 5. Voltage stabilisation circuit.

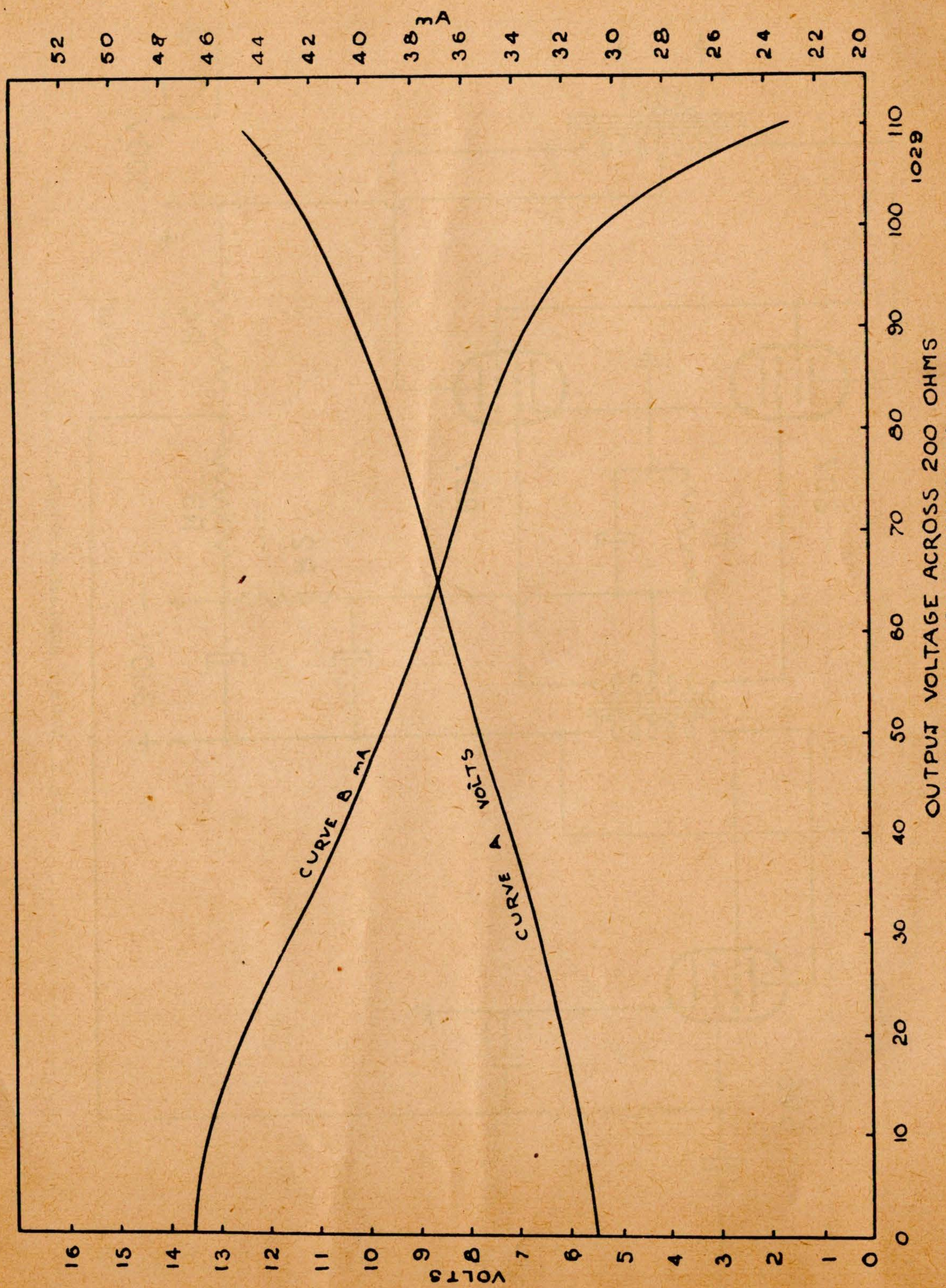

Fig. 6. Curve A, 6V6 Bias voltage. Curve B, 6V6 Cathode current.

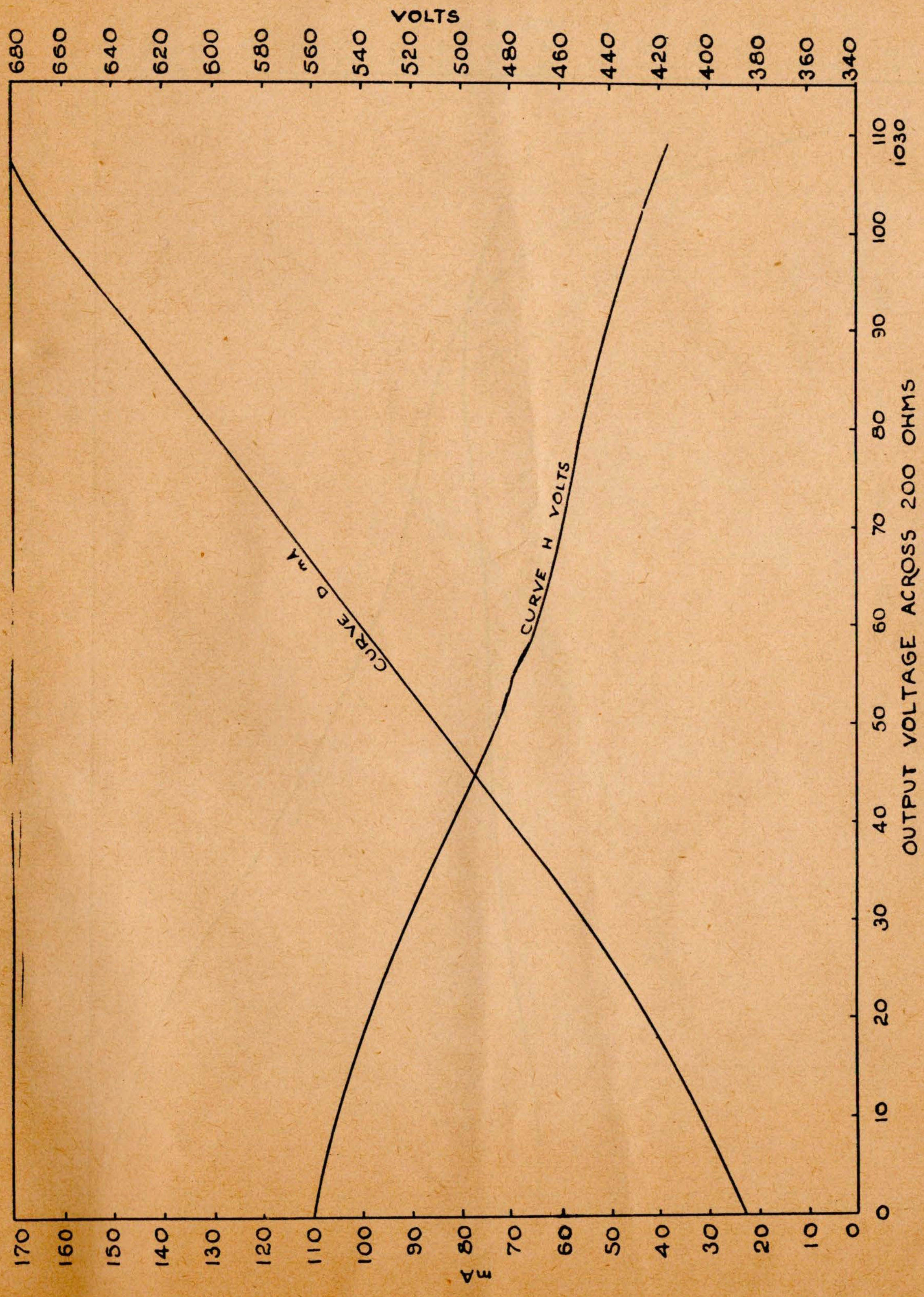

Fig. 7. Curve D, 807 Plate current. Curve H, 807 Plate voltage.

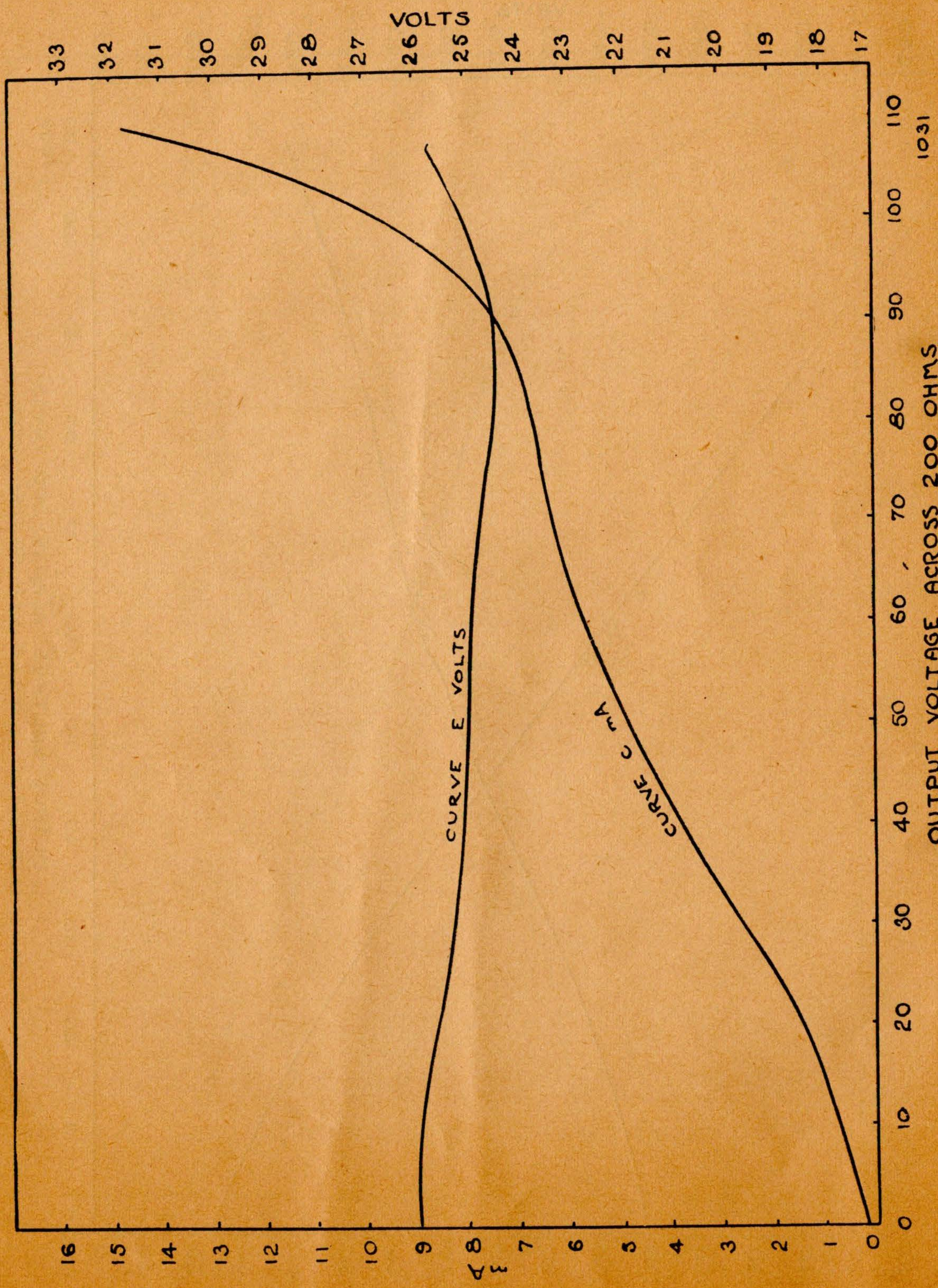

Fig. 8. Curve C, 807 Screen current. Curve E, 807 Bias voltage.

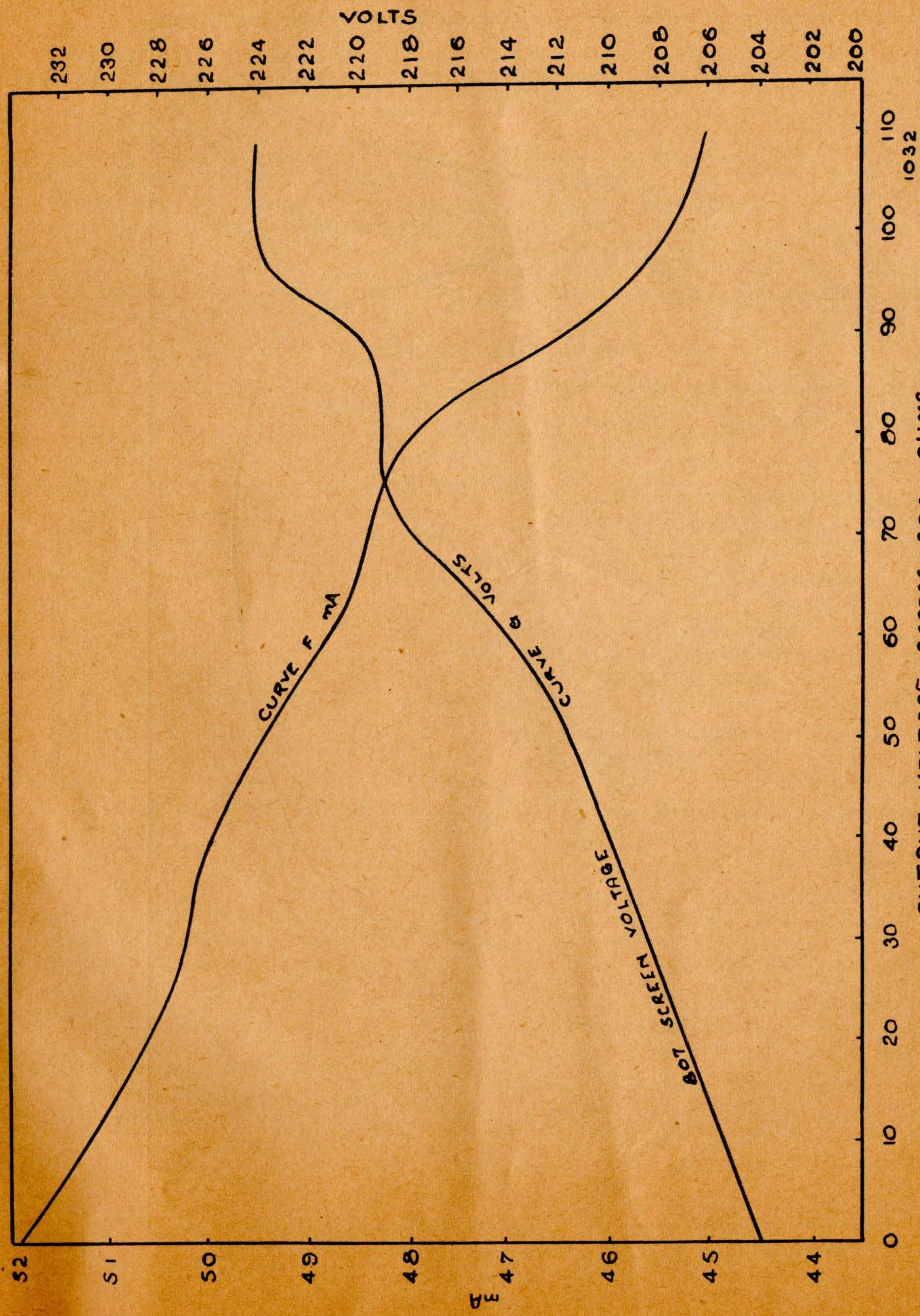

Fig. 9. Curve F, Current from low voltage supply. Curve G, Voltage of low voltage supply.

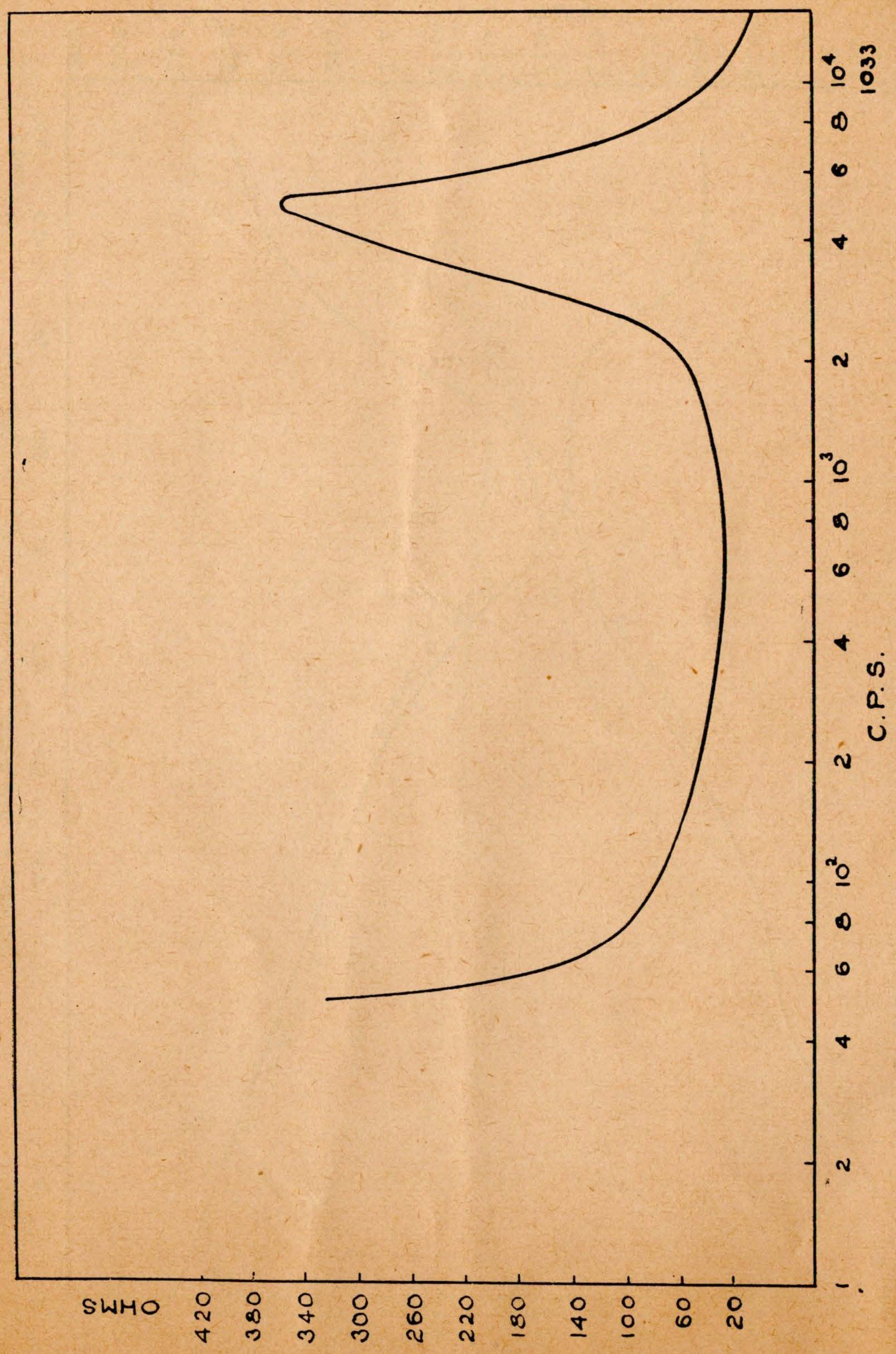

Fig. 10. Output Impedance.

A VERSATILE TRANSMITTING PENTODE AND ITS APPLICATION

In response to numerous requests a list of circuit values for Figure 1 printed on page 15 of the December issue of Technical Communication is given hereunder:

R1 = 50,000 Ohms 1 Watt
2 = 10,000 Ohms 10 Watts
3 = 25,000 Ohms 20 Watts
4 = 50,000 Ohms 1 Watt
5 = 500 Ohms Wire wound
6 = 50 Ohms 1 Watt Non-inductive
C1 = 250 pF
2 = 50 pF
3 = 0—30 pF Air trimmer
4 = 250 pF
5 = .005 Mica
6 = 100 pF
7 = .005 μF Mica
8 = 50 pF
9 = .01 μF, 600 Volts
10 = .005 μF Mica
11 = 100 pF
12 = .005 μF Mica
L1, L2 and L3—Adjust for frequency employed.

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CONTENTS

Articles to be published in future issues.

A contribution to the Theory of Acoustic Radiation *by C. J. Bouwkamp.*

Measurement and Calculation of Unsharpness Combinations in X-ray Photography *by H. A. Klasens.*

Theory of Grounded Grid Amplifiers *by A. van der Ziel.*

On the Excitation Temperature, the Gas Temperature and the Electron Temperature in the High Pressure Mercury Discharge *by W. Elenbaas.*

The Control of the Current Distribution in Electron Tubes *by J. L. H. Jonker.*

An Improved Method for Coupling Valves at Ultra Short Waves *by A. van Weel.*

Radiation and Heat Conduction in Light Scattering Material I. Reflection and Transmission *by H. C. Hamaker.*

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