This article collates design information on the 1949 'new' Williamson amplifier circuit. By detailing design considerations of the original circuit, assessment of altered operating conditions or part selection or circuit changes can be made.

A listing and commentary of changes proposed over decades by magazine articles and manufactured clones is provided.

The aim of this article is not to propose substantial changes to the original circuit, but rather to appreciate the original circuit's design outcomes, and why some have made changes to the design over time.

## Contents



## 1. Preamble

The 1949 Williamson 'new' amplifier circuit with 6SN7 and KT66 valves, and output transformer with 0.95Ω secondary windings configured as 8.5Ω (3 secondary sections in series), is the default circuit assessed here (807 related parameters given in {brackets}). Williamson used an output transformer with 1.7Ω windings configured as 15.3Ω (also 3 secondary sections in series). The circuit schematic shows idle condition voltages and currents (the first stage power supply voltage is ~305V, not 320V, and the driver stage power supply rail is ~430V). At 15W output in to 8.5Ω, the output voltage is 11.3Vrms.

Williamson related articles and information are collated at [http://dalmura.com.au/projects/Williamson.php](http://dalmura.com.au/projects/Williamson.php). Valve related parameters identified in this article may differ between datasheets, and hence various values may not exactly align with all datasheets or reference articles.

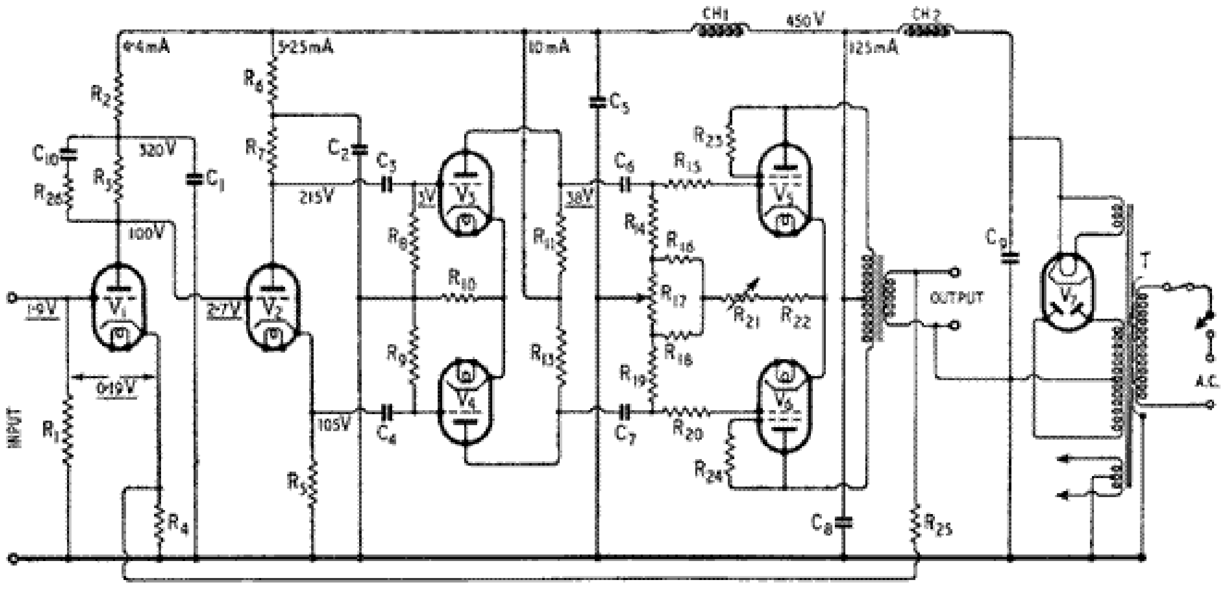



Fig. 1. Circuit diagram of complete amplifier. Voltages underlined are peak signal voltages at 15 watts output.

| | | |
|---|---|---|
| $R_1$ | 1MΩ | ¼ watt ± 20% |
| $R_2$ | 33,000Ω | 1 watt ± 20% |
| $R_3$ | 47,000Ω | 1 watt ± 20% |
| $R_4$ | 470Ω | ¼ watt ± 10% |
| $R_5$, $R_7$ | 22,000Ω | 1 watt ± 5% (or matched) |
| $R_6$ | 22,000Ω | 1 watt ± 20% |
| $R_8$, $R_9$ | 0.47MΩ | ¼ watt ± 20% |
| $R_{10}$ | 390Ω | ¼ watt ± 10% |
| $R_{11}$, $R_{13}$ | 47,000Ω | 2 watt ± 5% (or matched) |
| $R_{14}$, $R_{19}$ | 0.1MΩ | ¼ watt ± 10% |
| $R_{15}$, $R_{20}$ | 1,000Ω | ¼ watt ± 20% |
| $R_{16}$, $R_{18}$ | 100Ω | 1 watt ± 20% |
| $R_{17}$, $R_{21}$ | 100Ω | 2 watt wirewound variable |
| $R_{22}$ | 150Ω | 3 watt ± 20% |
| $R_{23}$, $R_{24}$ | 100Ω | ¼ watt ± 20% |
| $R_{25}$ | $1{,}200\sqrt{\text{speech coil impedance}}$ | ¼ watt (see table) |
| $R_{26}$ | 4,700Ω | ¼ watt ± 20% |
| $C_1$, $C_2$, $C_5$, $C_8$ | 8μF | 500V wkg. |
| $C_3$, $C_4$ | 0.05μF | 350V wkg. |
| $C_6$, $C_7$ | 0.25μF | 350V wkg. |
| $C_9$ | 8μF | 600V wkg. |
| $C_{10}$ | 200pF | 350V wkg. |
| $CH_1$ | 30H at 20mA | |
| $CH_2$ | 10H at 150mA | |
| T | Power transformer | Secondary 425-0-425V 150 mA, 5V. 3A, 6.3V 4A, centre-tapped |
| $V_1$, $V_2$ | 2×L63 or 6J5, 6SN7 or B65 | |
| $V_3$, $V_4$ | do. | do. |
| $V_5$, $V_6$ | KT66 | |
| $V_7$ | Cossor 53KU, 5V4 | |

**Figure 1. 1949 'new' Williamson amplifier schematic circuit**

## 2. Mid-band behaviour

A feedback voltage Vfb is applied to the first stage by R4-R25 divider from the amplifier output, with a feedback parameter β = R4/(R4+R25). R4 also carries the V1 anode signal current, which provides some current feedback, and although this is significant (R4 ~ 1/gm of V1), it is often ignored for expediency. R4 and R25 also set the V1 bias – see later section. The cursory equations that define the feedback are:

| | |
|---|---|
| Vgk = Vin – Vfb = Vin – β.Vout | signal input Vin is feedback voltage added to V1 grid-cathode voltage |
| Vout = Ao.Vgk | Ao is the open loop gain from input stage Vgk to Vout (with R4 bypassed to ground) for AC signals. |
| Av = Vout / Vin = Ao / (1+ β.Ao) | Av is the closed loop total amplifier gain with feedback applied. |

The first stage includes the global negative feedback connection, with the stage gain dependant on whether the feedback loop is closed or not.

- 6SN7 @ Ip=4.4mA, Eb=100V: Ri~9.2k, gm~2.2ms, μ=20
- The shelf network C10-R26 is considered an open circuit.
  - Only R3 loads the stage
- Stage gain A1 = Vo1 / Vgk
  = u.R3/ (R3 + Ri) = 16.7 (24.5dB)
  - Gain with R4 signal bypassed to ground
- Stage gain A1' = Vo1 / Vin = A1 / 10 = 1.67 (4.5 dB) = μ.R3/ (R3 + Ri')
  - Gain with 20dB feedback connected
  - Hence Ri' = R3 . (10.μ/A1 - 1) = 11 x 47k = 517k
- Stage gain A1'' = Vo1 / Vin (R25 disconnected, and not grounded)
  = μ.R3 / (R3 + Ri + (μ+1).R4) = 20 x 47k /(47k + 9k2 + 21x470)
  = 940k/66k = 14.3 (23dB)
  - Vin includes voltage across R4, but no feedback from output
    - eg. low or high frequency operation
  - Hence Ri'' ~ Ri + (μ+1).R4 = 9.2k + 9.9k = 19.1k
  - Quasi-open-loop circuit configuration

The phase splitter stage

- 6SN7 @ Ip=5.2mA, Eb=110V: Ri~9k, gm~2.3ms, μ=20
- Total stage output loading R = R7//R8 = 21kΩ
- Stage gain A2 = μ.R / (Ri + R(μ+2)) ~ 0.89 (-1 dB)

The driver stage

- 6SN7 @ Ip=5mA, Eb=215V: Ri~9k, gm~2.3ms
- Stage gain A3 = gm.(Ri//R11//R14) = 2ms x 9k//47k//100k = 2x7.0= 14 (23 dB)
- Fig.1 schematic shows driver stage gain A3 = 38/3 = 12.7

Output stage including output transformer

- 26.9Vrms on KT66 grid (38Vpk at 15W output with 40VDC cathode bias from [1])
- 11.3Vrms on speaker load (15W output at 8.5Ω load)
- Transformer turns ratio Vout/Vplate = A5 = 2 x √(8.5/10k) = (√Zo)/50 = 1/17 (-24.6dB)
- Vplate = 11.3V x 17 = 192Vrms
  - See Figure 12 for loadline showing ~192V x √2 = 270Vpk swing from bias point to Vg~0V.
- Stage gain A4 = 192V/26.9V = 7.14 (17 dB)

The open-loop amplifier gain Ao = A1.A2.A3.A4.A5 = 16.7*0.89*14*7.14 / 17 = 87.4 (38.8 dB).

- Based on R4 in circuit but bypassed, and with R25 disconnected.
- [2] identifies drooping open-loop gain above about 8W output with no feedback.
  - Likely due to 807 grid conduction (Vgk<0 but non-zero current), or driver stage approaching swing limits.

From [1], the measurement of loop gain is made by disconnecting the feedback resistor R25 from first stage cathode, and connecting it to ground via a 470Ω resistance, and measuring the signal voltage across that 470Ω. For 20dB feedback, the gain of the open-loop amplifier plus feedback is set to 10 (20 dB) = Ao.R4/(R4+R25) = β.Ao, and the closed loop total amplifier gain Av with feedback applied is then Ao/11. It is assumed that the input stage has R4 bypassed for this setup procedure (page 40 [50]), but that is not explicitly stated in [1].

Figure 1 assessment [23] of AC peak signal levels identifies the following:

- V1 plate current AC signal: 2.7Vpk/47kΩ = 0.057mApk
- V1 AC resistance loading the cathode: Rk = (R3+Ri)/(μ +1) = (47k+9.2k)/21 = 2.68kΩ
- Feedback voltage across R4 and Rk: 1.9V – 0.19V = 1.71Vpk.
- Signal current through Rk: 1.71V/2.68kΩ = 0.638 mApk in Rk.
- Signal current in to R4-R25 node: 0.638mA – 0.057mA = 0.581 mApk.
- R4-R25 node currents: 0.581 mApk = (√2 x 11.3V – 1.71V)/R25 – 1.71V/R4
- Hence R25 = (16V – 1.7V) / (0.581mA + 3.64mA) = 14.3V / 4.22mA = 3.4kΩ
- The calculated value of R25 is the same as the parts list value of 1200 x √(8.5Ω) = 3.5k.

Alternative approximation:

- The input signal sensitivity for 15W output is 11.3V/7.1 = 1.6Vrms.
- The schematic indicates the closed-loop amplifier gain Av is 11.3Vrms/(1.9Vpk/√2) = 8.4 (18.5dB).
- Feedback parameter β = R4/(R4+R25) = 10/Ao = 0.114 (-18.8 dB) from [1]
    - Ao = 87.4 (38.8 dB).
- The closed-loop amplifier gain Av = Ao/(1+β.Ao) = 87.4/(1+0.114x87.4) = 7.95 (18dB).
    - So the circuit has about Ao – Av = 38.8dB – 18dB = 20.8 dB of global midband feedback.
- R25 = (87.4*470/10) – 470 = 3.64kΩ.
    - The calculated value of R25 is the same as the parts list value of 1200 x √(8.5Ω) = 3.5k.
    - Parts list value appears to be an approximation:
        - R25 = Ao.R4/10 – R4
          ~ 47.A1''.A2.A3.A4.(√Zo) /50
          = 1195.√Zo

Measured 500Hz mid-band gain for R25=3k4Ω and 8.0Ω load, and 0.20Vrms input:

- Closed loop output +15.8dB
- Open loop output +35.8dB (no change for R25 connected to ground through 470Ω)
    - +17.5dB across 470Ω loading R25
    - 20dB increase from closed loop
- Closed loop output +17.1dB with R4 bypassed
- Open loop output +37.1dB with R4 bypassed
    - +18.6dB across 470Ω loading R25
    - 20dB increase from closed loop

## 3. Low-frequency behaviour

Low frequency behaviour of the Williamson amp is dominated by the CR high-pass filter couplings between stages, and the output stage RL high-pass filter due to output transformer inductance, with the addition of other CR power supply decoupling effects. The close proximity of the two highest pole frequencies possibly indicates poor phase margin, with Williamson purposefully incorporating the R2-C1 zero to assist the phase margin, given the transformer pole was likely to be moving around with signal level. Poor output stage static balance could easily raise the transformer pole frequency to equal or exceed the two dominant poles. (807 related parameters given in {brackets}).

Corner frequency summary:

| | | | |
|---|---|---|---|
| a) | Phase splitter stage CR coupling | 6.8 Hz pole | |
| b) | Driver stage CR coupling | 6.4 Hz pole | |
| c) | Output stage with transformer | 3.9 Hz pole (nominal) | {4.7Hz} |
| d) | Phase splitter stage B+ decoupling | 0.9 Hz zero | |
| e) | First stage B+ decoupling | 0.6 Hz zero | +18dB gain lift |

Williamson reported a low frequency phase shift of 90deg and -3dB at ~10 Hz for open-loop operation. For 20dB of feedback, Figure 2 indicates a phase margin of 20-30 deg at 2.9Hz, and gain margin of 8dB at 2Hz.

a) Phase splitter stage output CR coupling: 50nF-470kΩ = 6.8Hz HPF

b) Driver stage output CR coupling: 250nF-100kΩ= 6.4Hz HPF

c) Output transformer response:

- transformer PP inductance varies with:
  - idle DC current imbalance.
  - signal level excitation.
  - signal level valve conductance imbalance (causing a DC offset).
  - transformer core characteristics, including clamping and T-U lamination edge butting.
  - frequency due to core permeability variation above about 5Hz, and phase shifting to zero as reactance falls towards DC.
- KT66 triode plate resistance Ra ~ 1.45kΩ [Va=400V, Vg1=-38V, Ia~71mA]
  - Constant cathode voltage assumed.
  - {~ 1.9kΩ}
- corner frequency is at most (10k//(2x(1.45k+202))/(2π100H) = 3.9Hz HPF {4.7Hz}
  - 5VAC excitation rated [1] minimum 100H PP inductance used, for small signal condition.
  - At higher signal level, L will be higher and corner frequency will reduce.
  - At lower signal level, or when some DC imbalance exists, L will be lower and corner frequency will increase.
  - Primary winding DCR included.
- 10kΩ PP reflected loading from speaker side assumed.
  - 10kΩ from secondary loading is about 3x the primary side source resistance ( 2x(1.45k+202) = 3.3kΩ), which defines the nominal roll off rate and Q about the corner frequency. The roll-off rate of magnitude is significantly faster if secondary loading increases. [18]

d) Phase splitter stage supply R6-C2 decoupling: 22kΩ-8uF= 0.9Hz LPF

- At DC, the phase splitter anode loading increases from 22k to 44k, introducing phase splitter output gain imbalance.

e) First stage supply R2-C1 decoupling: 33kΩ-8uF= 0.6Hz LPF

- Stage gain at mid-band A1 = µ.R3 / (Ri' + R3) = 1.67 (4.5 dB)
- Stage gain at low frequency A1'' = µ.R3 / (Ri'' + R3) = 14.3 (23 dB)
- Stage gain at DC A1dc = µ.(R3+R2) / (Ri'' + R3 + R2) = 20 x 88k /(19k1 + 88k) = 16.4 (24.3 dB)
- At DC, the first stage gain is increased by ~20dB

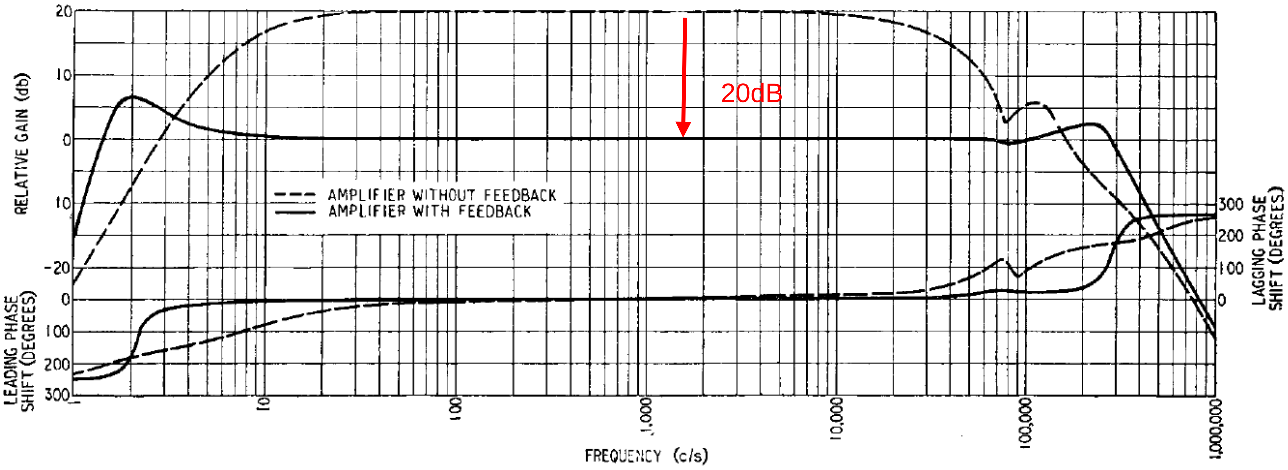



**Figure 2. Loop gain and phase-shift characteristics of amplifier**

Note that as the frequency reduces below 1Hz, the first stage and phase splitter stage lose the benefit of local supply decoupling due to the rising impedance of C1 and C2. The impedance of C5 is similarly rising, and so the stages share additional loading from CH1. The signal currents through V1 and V2 are in antiphase and of similar magnitude, so effectively cancel each other at the C5 node.

The onset of LF instability, such as with motorboating, is likely dominated by the output transformer's LF response, and excess phase shift from the CR coupling filters at the phase margin frequency. Preferred methods to alleviate LF instability include raising C3, C4 value or using a shelf network, as described in 8 c).

## 4. High-frequency behaviour

High frequency behaviour of the Williamson amp is dominated by the output transformer response, with the addition of multiple RC low-pass filter couplings due to inter-stage connections. Additional stray capacitance could significantly lower an RC corner frequency, such that phase margin is reduced. Parts and layout for the first stage to phase inverter stage, and the driver stage to output stage, are most susceptible. (807 related parameters given in {brackets}).

Corner frequency summary:

| | | | |
|---|---|---|---|
| a) | Transformer RLC | ~60 kHz | (2 pole roll-off) |
| b) | First stage output RC | 450 kHz max | |
| c) | Driver stage output RC | 450 kHz max | |
| d) | Phase splitter stage output RC | 5.7 MHz max | |

Williamson reported a high frequency -3dB at ~30 kHz for open-loop operation, and phase shift of 90deg at ~50kHz, with a resonance amplitude dip at about 60kHz when using an output transformer made by Vortexion to his specification of leakage inductance (Vortexion determined the sectioning of secondaries to suit that spec. The gain and phase curves in Figure 2 are for the 'improved' amplifier including the shelf network on the input stage, with a phase margin of 20-30deg at 150kHz, and a gain margin of 8dB at 300kHz.

a) The output transformer causes the dominant low-pass filter response that dictates the high-frequency bandwidth limit and stability margins of the amplifier. The output stage valves represent a voltage source with series resistance that drives signal current through the OT primary winding, which can be simplistically represented by series leakage inductance and winding DC resistance, and then a parallel connection of the primary half-winding presenting the reflected load impedance, and a shunt lumped capacitance across the winding. The source resistance, and the shunt capacitance, leakage inductance and dc winding resistance introduce an RLC two pole roll-off with a damped response. In reality, the interleaving and sectioning required to minimise leakage inductance for both class A and AB operation, and even UL operation, make it difficult to present anything but a simplistic representation.

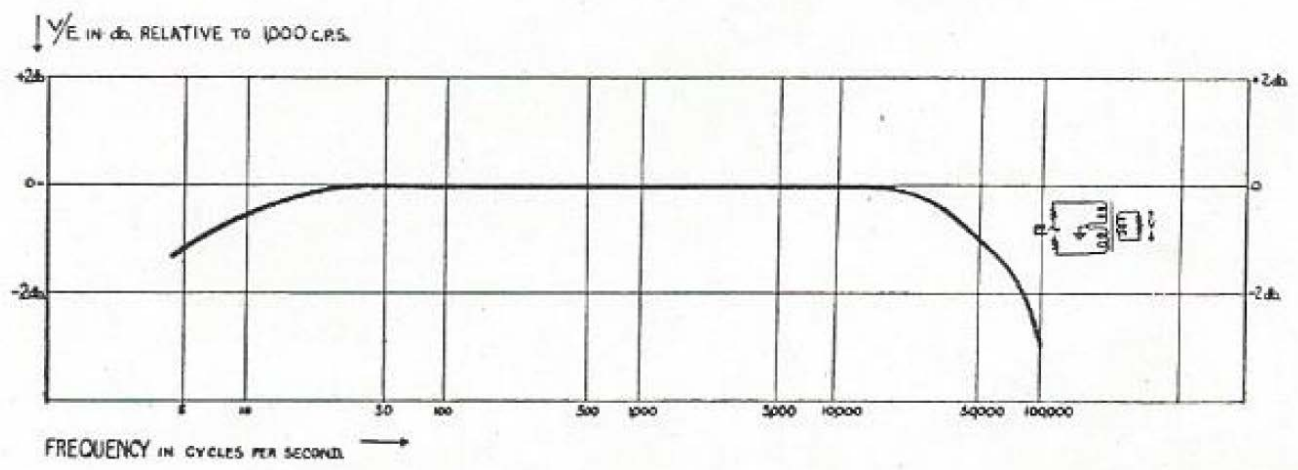



**Figure 3. Partridge WWFB frequency response**

The Partridge WWFB response [27] in Figure 3 is with a 3k3Ω source resistance (equivalent to the internal resistance of two series KT66 triodes) {3k8Ω} and a 10kΩ primary loading, and with 10W output. Lee [18] assesses the high frequency response for a step down audio transformer and provides response curves (see Figure 4) for when source resistance R1 is half the primary loading R2, which can be compared to the WWFB response.

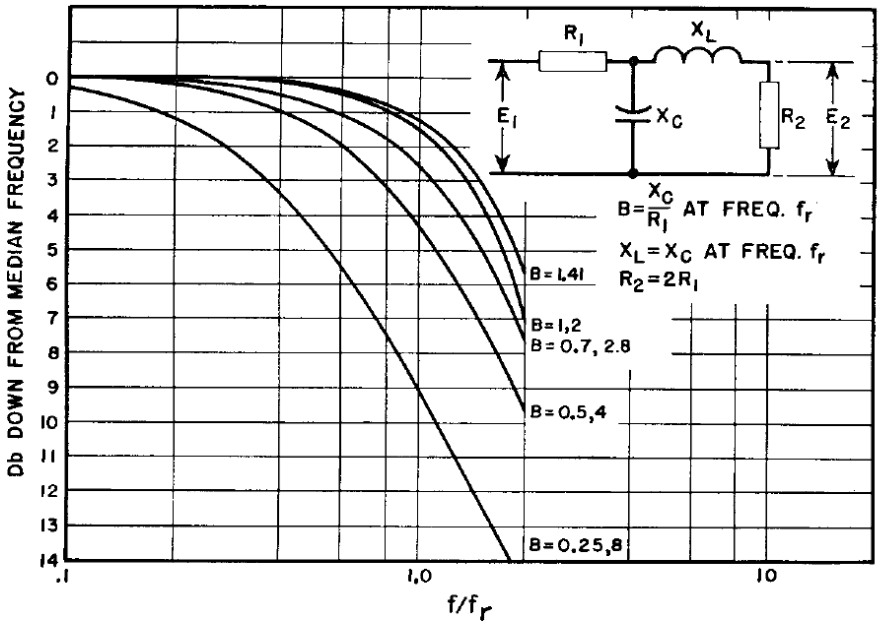



**Figure 4 High frequency response for step-down transformer for R2 = 2.R1 [18]**

The WWFB datasheet provides a self capacity value of 500-580pF ‘measured between either anode connection and the centre point of the primary commoned to the core and one point on the secondary’.

See [53] for one technique to measure self-capacity which represents the Xc reactance in Lee’s model.

- B = Xc/R1 ~ 1/(2π.60kHz.500pF.3k3Ω) = 1.6, assuming Fr = 60kHz
- From Figure 4 High frequency response for step-down transformer for R2 = 2.R1 [18], F-3dB ~ 1.6 f/Fr. So F-3dB ~ 1.6 x 60kHz = 96kHz. Which compares with datasheet frequency response.

Williamson targeted a primary winding leakage inductance of at most 30mH, and achieved 22mH at 1kHz. The WWFB datasheet provides a leakage inductance “measured as a series element in the primary” of 15-20mH.

- WWFB leakage inductance at 1kHz was measured as 14/16mH for 3.8/8.5Ω secondary [53].

- Gilson's W01796A output transformer datasheet [31] indicates that the primary winding leakage inductance should aim to be less than 25% of the P-P load of 10kΩ at the edge of audio bandwidth (eg. 25kHz), which would require leakage inductance to be < 16mH.

The lumped shunt capacitance and leakage inductance should have a calculated resonance frequency Fr = 1/[2π√(Lp.Cp)] similar to a measured OT resonance frequency. Williamson identified a resonant frequency of about 60kHz. The WWFB calculated frequency is ~ 60kHz (500pF, 15mH), and measured frequency was 48kHz (620pF, 14mH), the CFB calculated frequency is ~ 65kHz (600pF, 10mH).

The two-pole LC roll off reduces the open-loop gain by about 30dB over the decade from 60kHz to 600kHz, as indicated in Figure 2.

b) First stage RC roll-off due to effective output resistance and anode loading capacitance.

- First stage triode internal resistance Ri is increased by feedback signal at cathode. The effective first stage internal resistance is Ri'.
    - Ri' = 517k see mid-band section
    - R = Ra.Ri'/(Ra+Ri') [Ra=47k]
      = 43k
- First stage output loading capacitance.
    - First stage triode anode capacitance to cathode and heater
        - 0.7pf for 6SN7
        - Some valve holder parasitic capacitance as plate pin is adjacent to cathode
    - Parasitic capacitance should be low for dual triode 6SN7, due to direct anode to grid link.
    - Capacitance from shelf network across plate 47k not included here.
    - Phase splitter stage triode grid capacitance.
        - Cgk = 2.6pF
            - Cgk' = Ggk (1-A2)/A2 = 0.3pF due to cathode follower bootstrapping
        - Cgp = 3.8pF [6SN7 pin 4 grid]
            - Cgp' = Cgp (1+A2) = 7.2pF due to miller effect
    - Total loading capacitance ~ 0.7+0.3+7.2 = 8.2pF (neglecting the step network)
- RC corner frequency from ~43kΩ and ~8.2pF is ~ 450kHz max.
- The 'new' amplifier [1] includes a shelf network across plate 47k, comprising (200pF and 4k7Ω). Including the network was recommended if instability was experienced. The network steps down the loop gain from about 17kHz. [19] [20]
    - Stage gain at midband with feedback is A1' = 1.67 (4.5 dB) = µ.R3/ (Ri' + R3)
        - Ri' = R3 . (10.µ/A1 - 1) = 47k x 11= 517k (A1~16.7)
    - Shelf start corner frequency ~ 1/(2π.47k.200pF) = 17kHz (-3dB nominal corner frequency)
        - Ri' >> R3 [517k//47k=43k, f=18.5kHz]
    - Ri' will significantly reduce above nominal 17kHz. At high frequency above shelf stop corner f.
        - R3' = 47k//4k7 = 4.3k
        - A1 = u.R3'/ (Ri + R3') = 6.4 (16.1dB)
        - giving Ri'' = R3' . (10.µ/A1 - 1) = 4.3k x 30 = 130k (130k//47k//4k7 = 4.1k)
        - A1' = A1 / 10 = 0.64 (-3.9dB) = µ.R3'/ (Ri'' + R3')
        - So a drop of -3.9-4.5= -8.4dB from +4.5dB. (so n~2 from [19], rather than n=47k/4k7=10 with 20dB drop)
    - Shelf stop corner frequency ~ 1/(2π.4k3.200pF) = 185kHz (-8dB flattening-out corner)
        - Ri'' >> R3'
        - Stop to start frequency ratio is 11:1.
        - Max phase increase of ~30 deg at about 3.3x start corner frequency ~ 56kHz, then dropping back to zero.
        - The open-loop gain will start falling above 17kHz due to the shelf network at a max rate of ~ -8dB/decade, and level out above 185kHz.
        - The loop gain plot in Fig 2 of the 'improved' amplifier [1], shows the open-loop gain slowly falling from about 10kHz (as expected from the shelf network), although that fall then includes a contribution from the OT above about 40-50kHz, including a resonant dip around 80kHz.
    - Harmonic distortion components above about 17kHz will be attenuated less by negative feedback as the forward loop gain falls by up to 8dB to the stop frequency of 185kHz (not accounting for any loop gain reduction from the OPT). As such, THD likely increases for test frequencies above about 6-9kHz when the shelf network is included. In addition, the phase shift variation introduced

by the shelf network starts to occur well within the audio range, making distortion attenuation by feedback less effective.

c) Driver stage RC roll-off due to effective output resistance and anode loading capacitance.

- Ri ~ 9kΩ (although cathode is unbypassed, there is no cathode voltage change)
- R = (Ri // R11 // R14) + R15 = 7k + 1k = 8k [R11=47k, R14=100k, R15=1k]
- C= Cg of output stage valve plus Ca of driver plus parasitics.
  - Ca ~ 0.8-1.2pF
  - Cgk ~ 7pF (estimate) {~6pF}
  - Cg(a+s) ~ (1.1 + 3) ~ 4pF for KT66 {~3pF}
    - The grid to all except anode ~ 14.5pF {~11pF}
    - Cga ~1.1pF {0.2pF}
    - C grid to screen estimated at 3pF {~3pF}
    - grid to all other except screen and anode estimated at 7pF
  - Voltage gain A4 ~ 7 {A515 results are ~6}
  - C ~ 1.2 + 7 + (7+1)x4 + ~4 = 44pF {~25pF}
- RC corner frequency from ~8kΩ and ~44pF is ~ 450kHz max {400kHz max}

d) Although the phase splitter triode plate and cathode present different independent output resistances to the following stage, there is no high frequency imbalance as long as the following stage presents balanced loading (ie. no overdrive where grid conduction is not negligible). The output capacitance seen by each side of the splitter comprises the input capacitance from the driver stage, and also contribution from the splitter stage. A <1pF padder capacitance from driver 6SN7 pin 4 grid to anode may be needed to compensate for lower effective capacitance of that triode.

The phase splitter stage RC roll off is calculated from [4]:

- R' = effective plate or cathode load resistance [22k//470k=21kΩ]
- Ri = 9k2 [6SN7 @ Ip=4.8mA, Eb=110-115V]
- R= R'//Ri/(μ+2) = (R'.Ri/(μ+2))/(R'+Ri/(μ+2)) = 401Ω [μ=20.5 @ Ip=4.8mA, Eb=110-115V]
  - where Ri/(μ+2) = 409Ω and dominates the calculation of R.
- C1 = differential output capacitance of phase splitter (considered to be <<C2)
- C2 = Cg of driver stage triode.
  - Cgk = 3pF
  - Cgp = 3.8pF for 6SN7 pin 4 grid, and 4pF for other side
    - This imbalance may be trimmable
  - Triode gain of ~gm.R = 2.3m x 7k = 16
    - 6SN7 @ Ip=5mA, Eb=215V: Ri~9k. gm~2.3ms
    - R ~ Ri//47k//100k = 7kΩ
  - Cg ~ 3 + 16x4 ~ 67pF
- RC corner frequency from ~400Ω and ~70pF is ~ 5.7MHz max
- Merlin confirms the calculation method with a 12AT7 cathodyne circuit with 22k anode and cathode arm resistors, when operating at 2.6mA where the 12AT7 datasheet indicates μ ~ 46 and Ri ~ 22kΩ, using a lumped C=470pF and measuring a -3dB corner frequency of 595kHz.

e) Input to first stage.

- This roll off is outside of feedback loop.
- Grid stopper, or volume pot will introduce source resistance R
  - Contemporary volume pots were commonly from 250k to 1MΩ.
- First stage triode grid capacitance
  - Cgk = 2.6pF
  - Cgp = 3.8pF for 6SN7 pin 4 grid
  - Miller capacitance ~ 14x3.8=53pF
- RC corner frequency from 250kΩ and ~56pF is ~ 12kHz.

In summary, Williamson didn't design the amp during 1945-6 for unconditional stability (ie. no matter what the loading), as that aspect of performance was essentially new to designers and users. By 1949 when the 'new' version articles were presented, Williamson only presented a step network modification to mitigate cloned amps that showed HF instability. As such, the topic of ensuring stability has been up to the clone maker/user, and many techniques now exist to address this topic (see section 8).

## 5. Power Supply

The 'original' circuit used a Marconi U52 diode rectifier with directly heated cathode requiring 5V at 2.25A. The 'improved' circuit identified the more common 5V4 diode rectifier, with indirectly heated cathode that needs only a 2A heater current. The slower heat-up of the 5V4 indirectly heated cathode would have reduced the voltage stress on the filter and coupling capacitors, as the KT66's would be starting to load the supply at the time of power supply voltage rise. However, the 5V4 has a lower voltage rating than the U52, with some datasheets showing a 375VAC design max level, but the Sylvania datasheet shows design curves at 400VAC.

Almost all Williamson variants use a power transformer with HT voltage no more than 425-0-425V. The only commercial variant with a significantly higher HT voltage was the Heathkit W-5M with 465-0-465V. Higher voltage rated 5V 2A indirectly heated dual diodes include the GZ34 (introduced 1954) and the 5AR4 (introduced 1956-8) – see below for further assessment.

Contemporary parts can identify transformer and choke winding resistances. A Partridge TD2183 power transformer has primary 250VAC winding DCR=6Ω resistance and secondary 470V-0-470V winding DCR=91+96Ω resistance, with effective source resistance of about 120Ω. A Partridge TD2185 choke has 12H 200mA, and 158Ω DCR. A Partridge S15 choke has 30H 15mA, and 1.1kΩ DCR, and for nominal 20mAdc current of input, PI and driver stages would drop 22Vdc.

The PSUD2 simulation in Figure 5 shows a repetitive peak diode current of 0.5A which is just under the 5V4G max rating, and a hot turn-on causes a short-duration peak current of about 1A which is well under the max rating. The simulated hot turn on event indicates a ~2Hz ring on the output stage, and a slower damped response on the preamp stage rails, and an initial overshoot of nearly 600V. The simulation needs to be treated with caution as PSUD2 circuit loadings are modelled as either a constant current source or an end resistance, and PSUD2 is not being able to model non-linear choke inductance as DC current level changes.

CH1 and CH2 resistance provides substantial dampening to any transient disturbance. The PSUD2 simulation in Figure 6 shows a 5mA step change to the output stage loading, where some damped resonance is observable. For normal operation below clipping, the Williamson amp presents no transient loading.

Simulation shows that changing the diodes from valve to ss increases supply voltage rails by ~35VDC, but otherwise shows little effect to transient performance due to the well damped chokes.

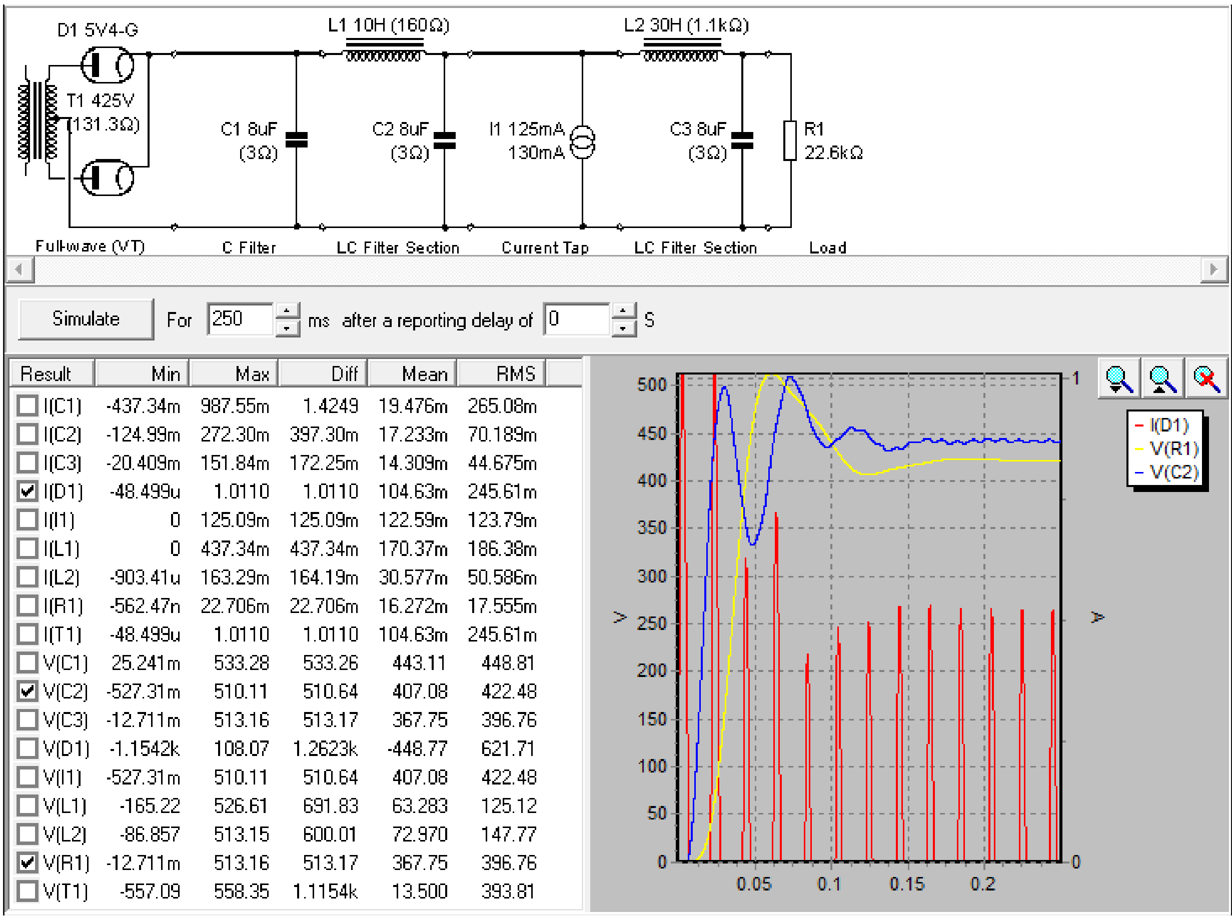


**Figure 5. PSUD2 hot turn-on simulation result.**

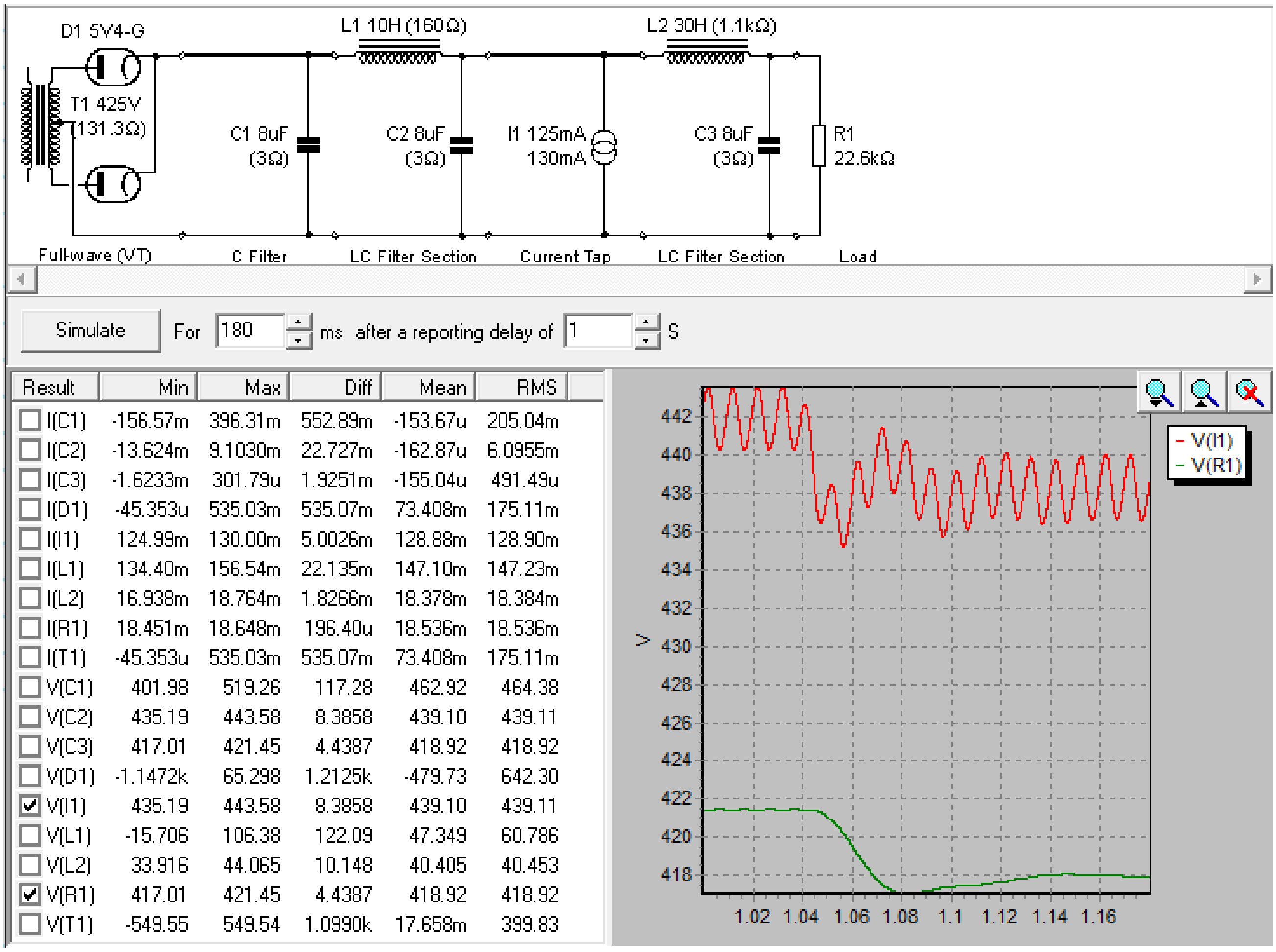



**Figure 6. PSUD2 step load change simulation result.**

The addition of power transformer secondary fuse protection is recommended. Simulation results in Table 1 for a fuse in the CT, and in Table 2 for a fuse in each winding arm, indicate an IEC 60127-2/3 quick acting F type fuse is appropriate [8]. Note that UL248-14 compliant fuses are not directly comparable.

| Simulate period in PSUD2 | 10ms | 50ms | continuous |
|---|---|---|---|
| Simulated RMS current | 0.61A | 0.53A | 0.25A |
| Multiplier (for 0.315A fuse rating) | 1.9 | 1.7 | 0.8 |
| Multiplier (for 0.25A fuse rating) | 2.4 | 2.1 | 1.0 |
| IEC 60127-2 **F** min limit multiplier | 4 | 2.75 | 1 |

**Table 1. CT fuse position assessment**

| Simulate period in PSUD2 | 10ms | 50ms | continuous |
|---|---|---|---|
| Simulated RMS current | 0.61A | 0.41A | 0.18A |
| Multiplier (for 0.2A fuse rating) | 3.1 | 2.1 | 0.9 |
| IEC 60127-2 **F** min limit multiplier | 4 | 2.75 | 1 |

**Table 2. Diode arm fuse position assessment**

A 315mA F 250VAC fuse in the CT connection suitably keeps operating current below the minimum limit ratings of the fuse. For a fuse in each winding arm, a 200mA F 250VAC fuse is appropriate. The continuous rms current level at idle is effectively the worst-case condition, as power supply current will reduce with signal level. Even though the secondary winding voltage of 425Vac exceeds the nominal 250Vac fuse rating, a 250Vac rated fuse is acceptable in this secondary side application.

Although a mains primary fuse is shown, no fuse rating is given. The power transformer is about 150-200VA with about 105W of secondary loading, indicating that a 0.8-1A Slow Blow fuse for 230VAC supply should manage the transformer and heater in-rush current.

In addition to appropriate fusing of the primary and secondary sides of the power transformer, adding two series connected 1N4007 in series with each anode of any valve diode reduces the risk of failure [46], [47].

Fusing beyond CH2 would need to cope with ~150mA DC, and the likely worst-case fault current may reach ~1A (ie. 5x multiplier for 200mA fuse). If the fault is through the OPT then at most the fault current may reach 0.8A. If the fault current is through the OPT and common cathode resistance then at most it may reach 0.5A.

Fusing beyond CH1 would need to cope with ~20mA DC, and the smallest practical fuse value would be 63mA, and the likely worst-case fault current may only reach 260mA (ie. 5x multiplier).

Fuse protection beyond just the mains primary fuse may seem excessive and difficult to retrofit, however original vintage chokes, transformers and output stage valves can be very expensive to replace if they fail first, or get stressed, and so fusing can be quite cheap insurance.

Two alternative indirectly heated diodes are compared to the 5V4G in Table 3. Comments on diode choice:

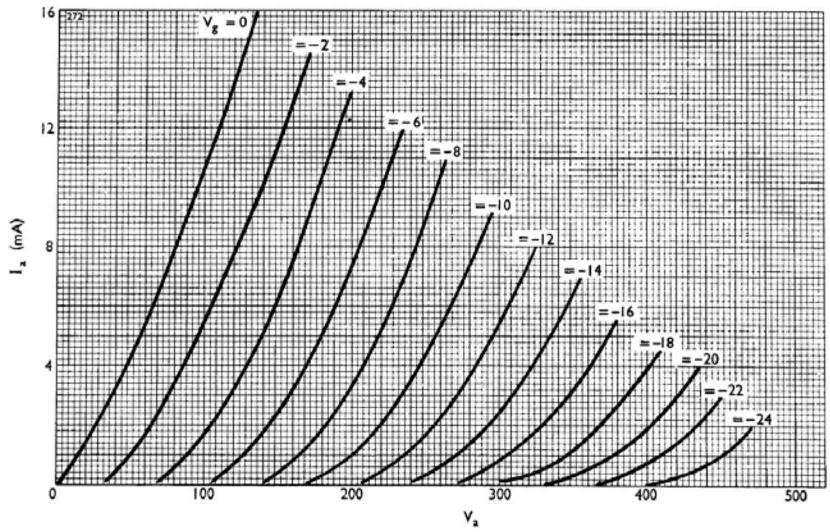
- Partridge TD2183 has sufficient supply impedance (~120Ω) to suit any diode in Table 3.
- The GZ34 and 5AR4 allow an increase in C9 capacitance, and additional PIV margin, but will raise the B+ supply voltage somewhat.
- The GZ32 allows an increase in C9 capacitance, although peak current capability is interpreted [47].

| | **U52** | **5V4G** | **GZ32** | **GZ34** | **5AR4** |
|---|---|---|---|---|---|
| Heater supply requirement | 5V 2.25A | 5V 2A | 5V 2.3A | 5V 1.9A | 5V 1.9A |
| Peak Inverse Voltage, PIV | 1430 V | 1400 V | 1400 V | 1500 V | 1700 V |
| Peak continuous current, $I_{ap}$ | 770 mA | 525 mA | ~700mA [47] | 750 mA | 825 mA |
| Peak transient current, $I_{apt}$ | 2.5 A | 3.5 A | ~2.0A [47] | 3.0 A [47] | 3.7 A |
| Max AC volts for capacitor input | 500 V | 375 V | 500 V | 400 / 550 V | 450 / 550 V |
| Min plate supply resistance, Rt | ~180 Ω | 100 Ω | 50 / 100 / 150 Ω | 100 / 175 Ω | 160 / 200 Ω |
| Max capacitor input capacitance | ~16 uF | 10 uF | 16u / 32u / 60u | 60 uF | 40 uF |
| Max DC output current | 250 mA | 175 mA | 250 mA, 350Vac<br>125mA, 500Vac | 250 mA /<br>160 mA | 225 mA /<br>160 mA |
| Plate voltage drop at 175mA | | ~ 25V | ~ 25V | ~ 13V | ~ 15V |

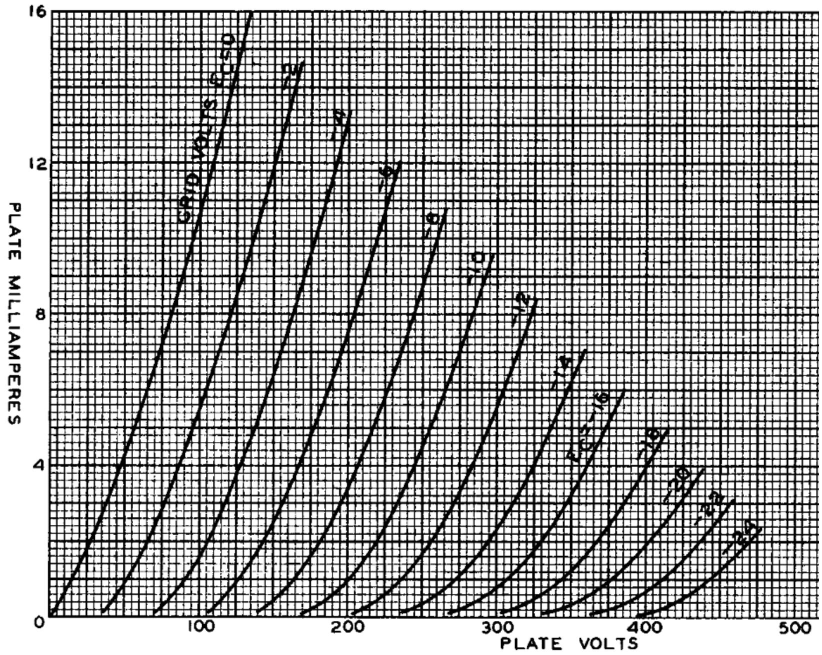
**Table 3. Valve diode option assessment**

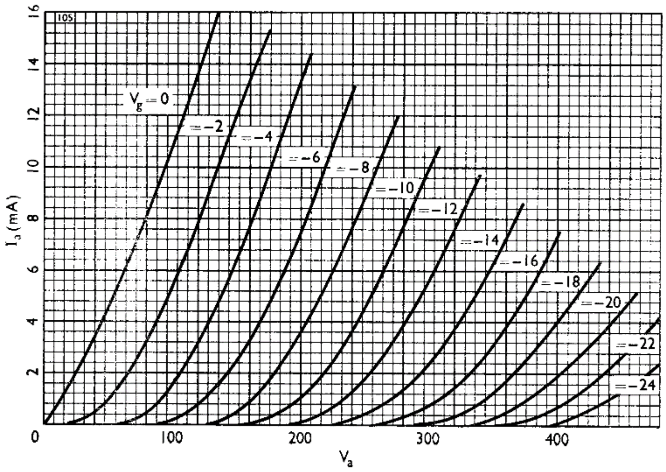
## 6. Signal stage valve types and bias conditions

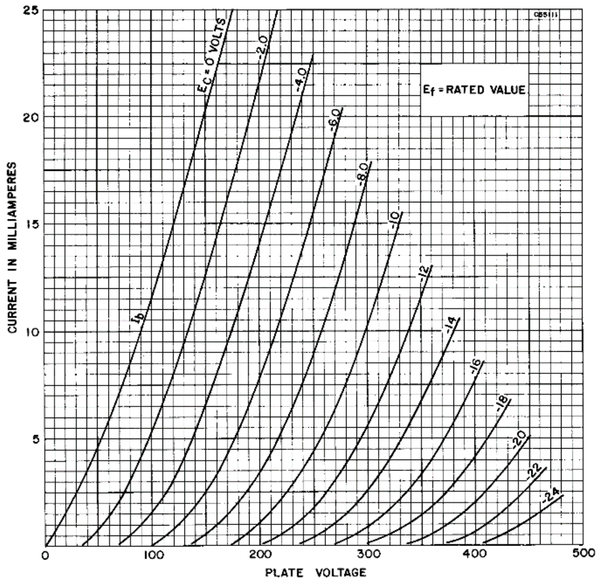
B65 plate curves




6J5 plate curves




L63 plate curves

6SN7 plate curves

**Figure 7. B65, L63, 6J5 and 6SN7 comparison**

Williamson appears to have originally used Marconi L63 single-triode valves in the amplifier's development in the early-mid 1940's, and then indicated that 6SN7 dual-triode valves were an alternative in the late 1940's. Figure 7 indicates that the L63, 6J5, B65 and 6SN7 have the same plate characteristic curves. The L63 triode and B65 dual triode have slightly higher grid to cathode and grid to anode capacitances than a 6SN7 dual triode, but the total heater requirement is the same. L63 and 6J5 have the same base, as do B65 and 6SN7.

Figure 8 to Figure 10 show the loadline and idle bias point for each stage in red using just the 6SN7 curves. The mid-band gain of the first stage is quite low at circa 1.7x based on 20dB feedback, with up to a 4Vpk signal requirement. Outside of mid-band, at the low and high frequency edges, the first stage gain increases to the non-feedback level of circa 17x, and so up to a 40Vpk signal requirement. Williamson biased the first stage and phase splitter to both accommodate up to 40Vpk operation depending on bogey tube and loadline.

In 1961, Wright [9] described changes to the bias points of the first, phase splitter and driver stages, and provided measured results on intermodulation distortion that indicated a significant improvement, although results were in general sparse and subjectively presented. Wright makes the ambiguous assertion that the US 6SN7 does not work well in the Williamson design, however a quick review of US 6SN7 datasheets and the Marconi L63 or 6J5 show no difference in plate characteristics! In summary, Wright's changes are:

- First stage – increase bias voltage from ~2.1V to ~3.0V, and maintain anode voltage at ~100V.
  - R2: 33kΩ → 47kΩ, R4: 470Ω → 910Ω
    - Supply voltage falls from ~305V to ~250V.
    - Idle current falls from ~4.5mA to ~3.25mA
    - Changing R4 requires a feedback network change. An alternative is to retain the 470Ω as the feedback point, but insert another bypassed 470Ω in between the 470Ω feedback point and the cathode – the bypass would need to be circa 470uF to maintain a corner frequency below 1Hz.
  - 6SN7 @ Ip=3.25mA, Eb=100V: Ri~9.2k, gm~2.2ms, μ=20 {likely minor changes}
  - Stage gain A1 = Vo1 / Vgk
    = u.R3/ (Ri + R3) = 16.7 {no change}
  - Gain with R4 signal bypassed to ground
  - Stage gain A1' = Vo1 / Vin = A1 / 10 = 1.67 (4.5 dB) = μ.R3/ (Ri' + R3)
  - Gain with 20dB feedback connected
  - Hence Ri' = R3 . (10.μ/A1 - 1) = 11 x 47k = 517k {no change}
  - Stage gain A1'' = Vo1 / Vin (R25 disconnected, and not grounded)
    = μ.R3 / (R3 + Ri + (μ+1).R4) = 20 x 47k /(47k + 9k2 + 21x910)
    = 940k/75k = 14.3 (22dB) {was 23dB}
  - Vin includes voltage across R4, but no feedback from output
  - eg. low or high frequency operation
  - Ri'' ~ Ri + (μ+1).R4 = 9.2k + 19.1k = 28.3k {was 19.1k}
  - Quasi-open-loop circuit configuration
- Phase splitter stage – increase supply voltage from 325V to 410V, and maintain anode current at ~5mA to increase idle Vak from ~110V to ~190V.
  - R6: 22kΩ → 3k9Ω.
  - Bias voltage increases from ~2.2V to ~6.2V.
- Driver stage – increase plate voltage from 160V to 230V, with Vgk increased from 4.5V to 8.5V.
  - R10: 390Ω → 1kΩ.
  - Stage current reduced from ~11mA to ~9mA.

Some comments can be made about these changes:

- Each of the modified stages has an increased signal swing capability that could reduce gross distortion at high signal levels if those stages were starting to limit signal swing.
  - For mid-band signals, only the driver stage approaches its voltage swing limits, and driver stage low order harmonic distortion is slightly higher with the Wright change.

- For low/high frequency signals, the first stage gain increases significantly, and depending on the coupling corner frequencies, the driver and output stages will see much higher levels due to their gain.

- The idle current requirements are not changed significantly, nor are the valve operating parameters of µ, gm or Ri. The first stage gain is only marginally reduced. The phase splitter stage cathode voltage doesn't change. The driver stage change is independent of the changes made to the input and phase splitter stages. Stage gain changes to be confirmed by test.
- The operating currents of the first stage and phase splitter stage are now significantly different, which may affect the original low-frequency compensation from the cascaded stages.
- The very low frequency response may change below the C1, C2 decoupling cap corner frequencies.
- Increasing the driver stage common cathode resistance should marginally improve gain matching of the two triodes. Even though the driver stage cathode current is reduced by about 20%, the anode idle dissipation is increased by about 15%, so service life is likely similar.

**Figure 8. First Stage**

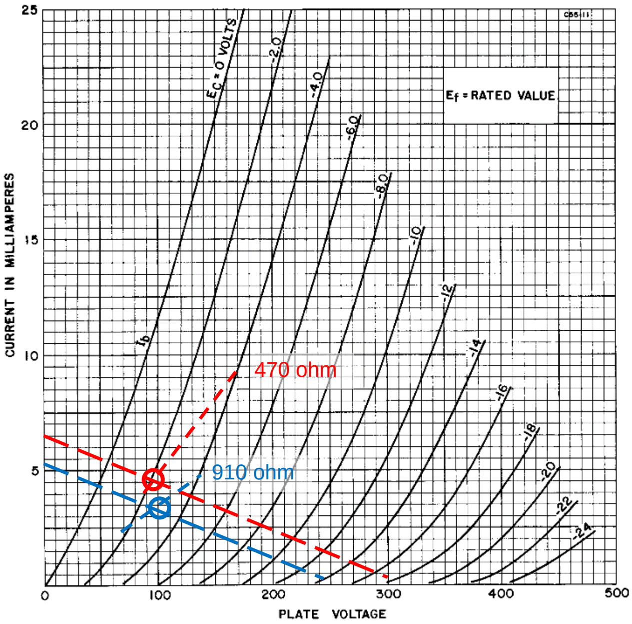



**Figure 9. Phase Splitter Stage**

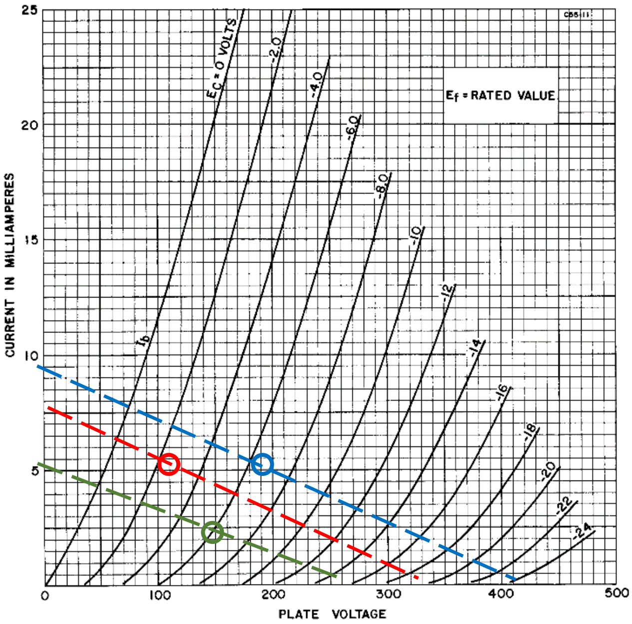
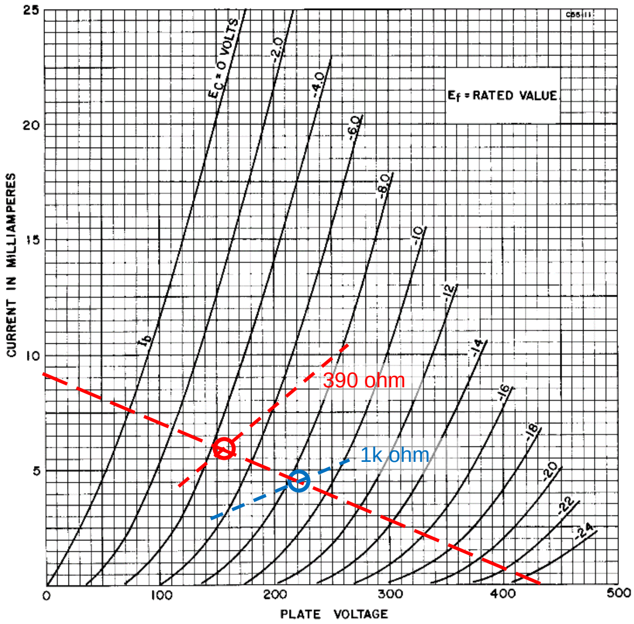


**Figure 10. Driver Stage**




The 6CG7 / 6FQ7 is the noval equivalent to the 6SN7 octal. The 6CG7 typically has an electrostatic shield between plates, whereas the 6FQ7 does not.

The 12AU7 has commonly been used as a noval alternative to the 6SN7, given that it has similar mu, anode impedance, and lower capacitances, and the benefit of half the heater current requirement. Bernard compared distortion levels of 6SN7 and 12AU7 in 1953 [34]. Barbour presents historical details on the 6SN7, along with harmonic test results, however the test circuit and signal level conditions are not given [36] so the reported results may not relate to signal conditions in the Williamson. Morgan Jones [23] compared distortion levels for 6SN7 and 12AU7 in a mu-follower circuit at moderately high signal voltage levels (20Vrms), which indicate the 12AU7 has significantly higher harmonic levels, whereas the blackened glass 6SN7 has lower levels. My testing of thirteen 6SN7 and five 12AU7 as the driver stage valve in the same Williamson amplifier indicated the 12AU7 had typically more than twice the dominant second harmonic distortion than 6SN7.

## 7. Output stage bias conditions

Williamson's intent was to operate V5-V6 at max design rated dissipation level, to maximise plate voltage swing under class A operation of the output stage. The output stage operates a KT66 triode at an idle point of ~400V and 62mA (25W). A 10kΩ PP class A load represents a 5kΩ loadline. The idle grid bias is about -38V. The maximum voltage swing is about 275Vpk, or 390Vrms PP, with close to symmetric swing to grid conduction or cut-off operation. The turns ratio PP 10k:8.5 is 34:1, so max design output voltage is 11.5Vrms, or 15.7W.

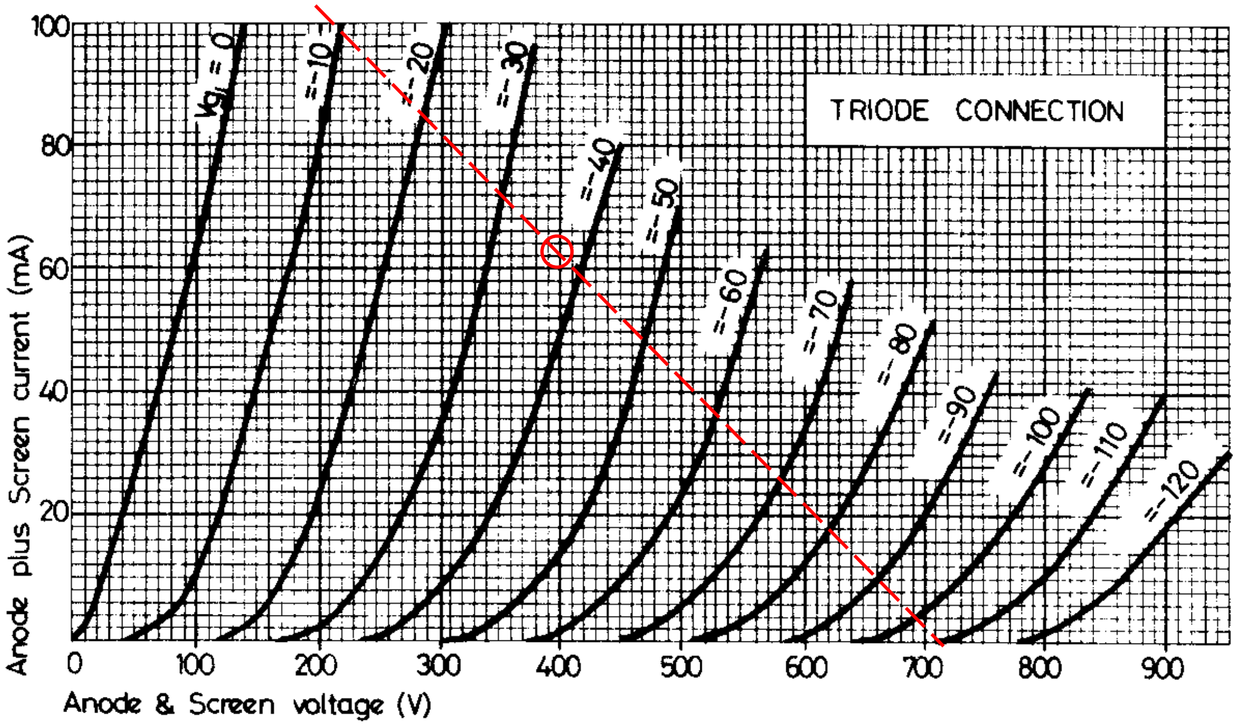



**Figure 11. KT66 triode characteristics.**

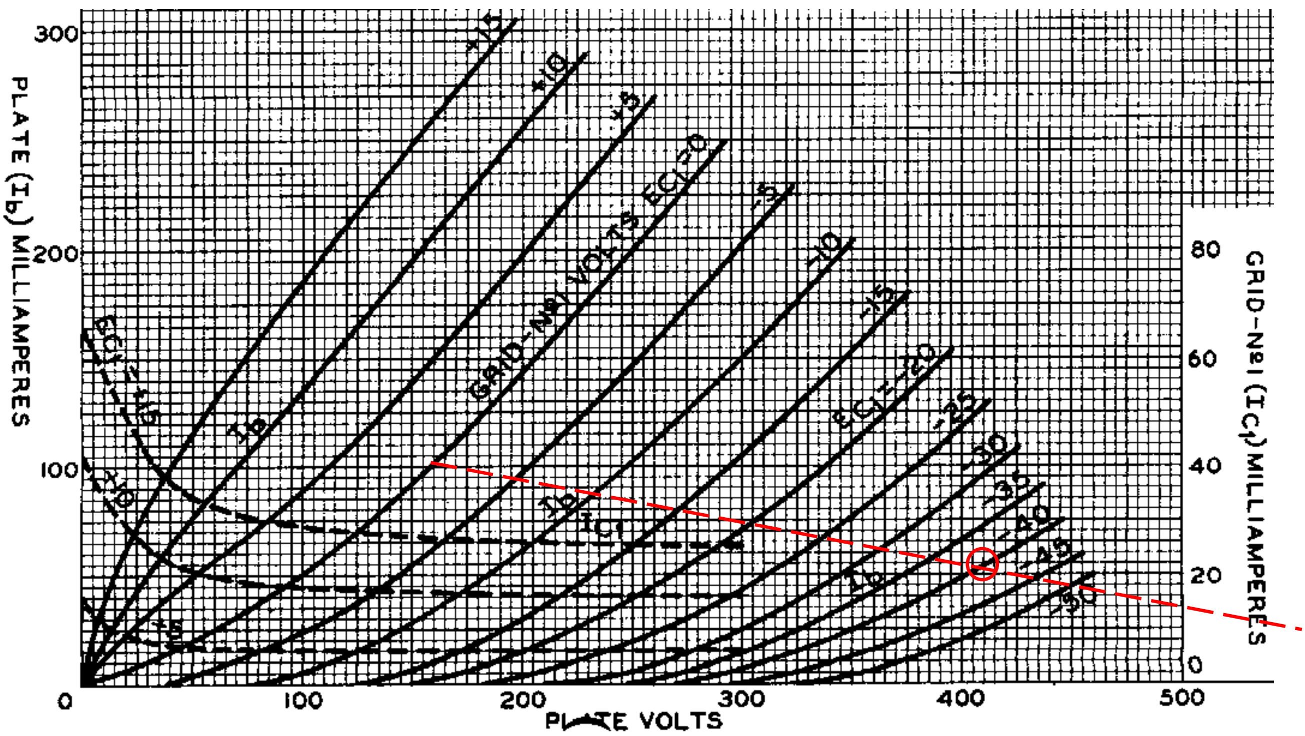



**Figure 12. 807 triode characteristics.**

The output stage operates an 807 at an idle point of ~425V and 55mA (25W) [1]. A 10kΩ PP class A load represents a 5kΩ loadline. The maximum voltage swing is a bit lower at about 225Vpk to grid-conduction. The 807 was not originally rated for screen operation at 400V, but was subsequently re-rated.

The RCA 6L6 beam pentode became more common than MOV's KT66, and RCA improvements out to the 6L6GC provide up to 30W plate dissipation, and 450V screen ratings. This allows a 6L6GC to operate at 80% plate power with the same bias current level as the KT66 at 100% plate power, for a longer service life.

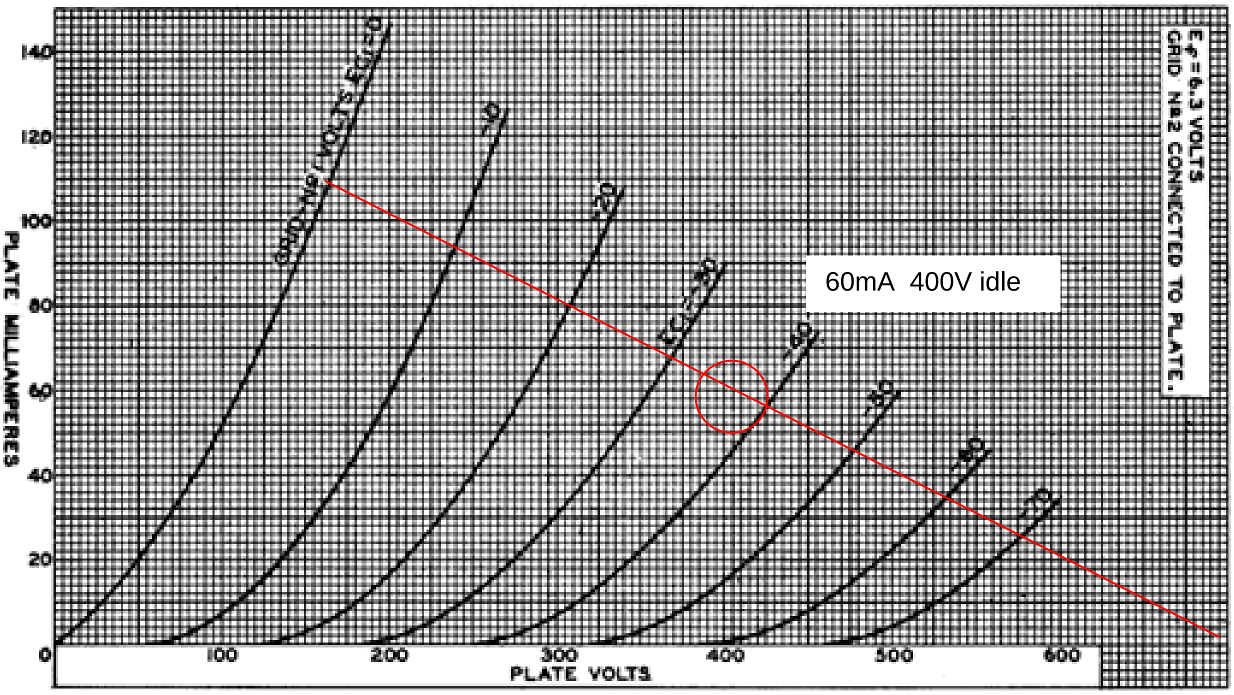



**Figure 13. 6L6GC triode characteristics.**

The balance between the two valves in the output stage has a significant effect on low-frequency distortion, as the symmetry of flux swing in the output transformer core, and the inductance drop-off due to a DC bias [40], become more important. Tube rolling, static and dynamic bias adjustment may change distortion performance markedly. Figure 14 is from the Heathkit W-5M instruction manual [10], with the x-axis representing a static bias current difference of 0.1V/30Ω = 3.4mA (or 3% of 62mA) per division away from balanced 0V, and so a net DC current of 3.4mA. Different output transformers may vary with their inductance roll-off with DC bias.

When setting the dc balance using just voltage measurement across cathode sense resistors it is worthwhile appreciating tolerances. A modern sense resistor may have a 1% tolerance, such that use of a +1% tolerance resistor on one side, and a -1% tolerance resistor on the other side, would result in a worstcase 2% bias current difference (eg. 2/3 of a div in Figure 14). The Heathkit W-5M uses 0.5% tolerance resistors for this application, but given 50 years of age these parts may have drifted in value. Even when using a meter to match the resistance value of each cathode current sense resistor, the resolution of the meter may not allow much lower than 1% deviation in resistance value to be discerned.

A further subtlety is that a UL output stage (as per the W-5M) may introduce some imbalance due to different screen currents, even though the cathode current is the sum of anode and screen currents, as the primary turns for screen and anode are different, and hence the magnetic bias of the core may be different.

For the autobias class A application, the 100kΩ grid leak is well below the datasheet recommended max value of 500kΩ for the KT66, and for the 807. Increasing the grid leak to 220kΩ would allow a 120nF coupling cap for the same 6.4Hz low-frequency corner, with likely reduction from over-drive blocking distortion and a 0.3dB higher mid-band loop gain.

The grid stopper value should not be significantly increased, as that would increase HF phase shift [4c)].

[1] The Radiotronics A515 [2] operated each 807 at 400V and 59mA anode current at idle.

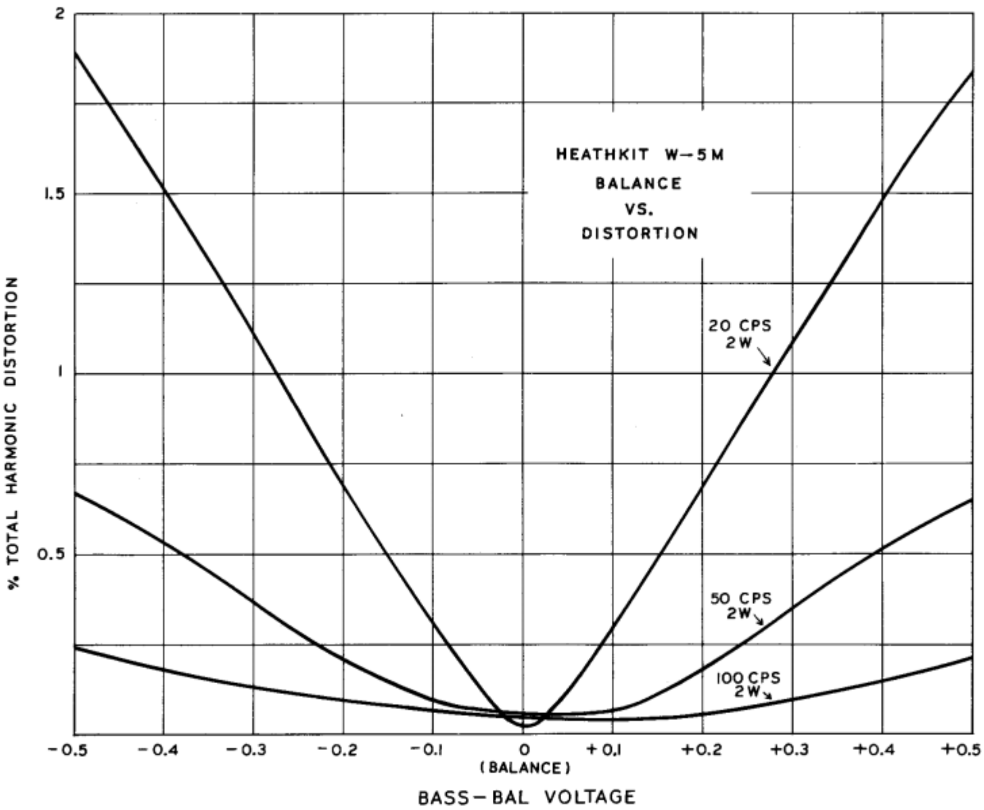



**Figure 14. W-5M output stage balance effect on low frequency distortion [10]**

The output stage was not normally operated with the valves in beam tetrode mode [2]. The following comments apply to using the beam tetrode mode:

- Grid characteristic curves not as linear as triode mode.
- Higher output stage gain.
- Screen current still included in common cathode voltage signal, but not in output signal.
- Screen voltage regulation influences gain.
- Lower miller capacitance due to decoupled screen.
- A beam tetrode has a higher Ra internal resistance
  - KT66 has Ra ~ 22.5k
  - 807 has Ra ~ 30kΩ (24k at 300V, 39k at 500V)
  - the output stage PP R-C high frequency roll off is likely to reduce by about 1.45kΩ/30kΩ ~5%. A triode KT66 has a closed loop roll-off of ~300kHz, and an 807 regulated screen has a closed loop roll-off of ~60-70kHz.
  - reduces speaker output damping factor
  - increases the low frequency RL to become the dominant pole.
  - accentuates high frequency transformer resonances and can make unconditional global negative feedback stability harder to achieve.
- Output stage B+ can be increased beyond ~440V, as screen voltage is held lower.
- The 807 pentode curves for Vg2 regulated at 300V in Figure 15 show the idle cathode voltage bias is lower (approx. -34V), and that voltage swing to grid-conduction is increased to about 370Vpk, although linear operation would significantly degrade beyond about 300Vpk. The STC application report extract below indicates lower cathode bias voltage (perhaps as Vg2 = Screen voltage – bias).

The June 1954 STC 807 application report provides results for 807 in pentode mode in class A PP, with the 807's idle anode dissipation at 24W.

[2] Author has a Miller Organ amplifier from 1960 using a Williamson circuit, but with output stage 807's operating in pentode mode.

**Class 'A' Amplifier** (Push-Pull):

TETRODE CONNECTION:

| | | | | | | | | |
|---|---|---|---|---|---|---|---|---|
| Plate Voltage | 250 | 270 | 270 | 500 | 500 | 600 | 600 | volts |
| Screen Voltage | 250 | 270 | 270 | 300 | 300 | 300 | 300 | volts |
| Grid Voltage | —16 | —17·5 | — | —27 | — | —29·5 | — | volts |
| Autobias Resistor | — | — | 125 | — | 270 | — | 360 | ohms |
| Peak AF Grid-Grid Voltage | 32 | 35 | 40 | 54 | 72 | 59 | 81 | volts |
| Plate Current (no signal) | 120 | 134 | 134 | 100 | 100 | 80 | 80 | mA |
| Plate Current (max. signal) | 140 | 155 | 145 | 154 | 119 | 150 | 97 | mA |
| Screen Current (no signal) | 10 | 11 | 11 | 2·5 | 2·5 | 1·5 | 1·5 | mA |
| Screen Current (max. signal) | 16 | 17 | 17 | 20 | 16·5 | 17·5 | 17·5 | mA |
| Output Load Impedance (plate – plate) | 5000 | 5000 | 5000 | 8000 | 9000 | 10,000 | 10,000 | ohms |
| Total Harmonic Distortion | 2 | 2 | 4 | 2·6 | 2·7 | 2·2 | 4·1 | % |
| Power Output | 14·5 | 17·5 | 18·5 | 38 | 32·5 | 47·5 | 36·5 | watts |

–Values are given for two valves.

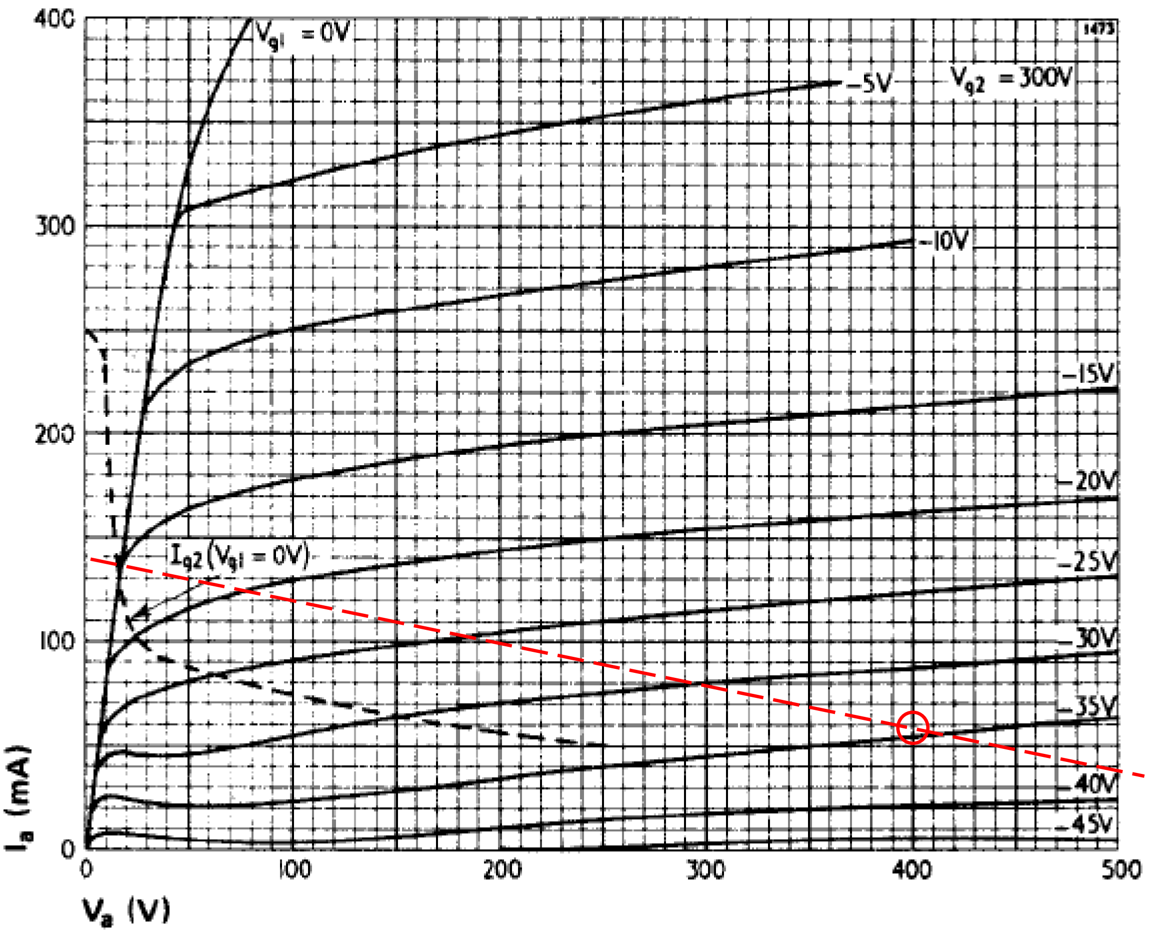



**Figure 15. 807 pentode characteristics for 300V screen and 10kΩ PP loadline.**

## 8. Changes

The original Williamson amplifier circuit was developed and demonstrated mainly in 1944, with the 1947 WW article written in late 1944 as preparation for an audio presentation to the Marconi-Osram Board [43]. Williamson used the services of Vortexion who were nearby to make his output transformer prototypes [43]. The 'new' circuit from WW 1949 only had the high frequency shelf circuit added to the input stage, and alternative valves identified, with the articles mainly adding preamplifier and tuner designs, and practical advice to cover reader feedback.

The following is a list of changes known to have been made to the Williamson circuit, with reference to the first use of the change, and any relevant comments. Some changes are practical and appropriate to improve performance, and make the amp bullet proof, although further assessment and testing may well be required.

a) Including an adjustment pot in the driver stage to adjust imbalance.

- Williamson originally included a 25kΩ adjustment pot in the common supply rail for the driver stage to allow adjustment of gain symmetry of each driver and output stage cascade circuit. The pot was removed in the 1949 'improved' circuit (with benefits of lower parts count and increased reliability, and safer operation given the pot was at B+) based on only minor variation in performance using typical new valves and matched parts.
- Nowadays, a pot may be useful to allow distortion minimisation, given the typical use of aged parts and the ease of distortion measurement. Adjustment at on-set of clipping seems appropriate.
- To maintain reliable and safe operation, inserting a 20kΩ multiturn 0.75W trimpot in each anode load leg (with wiper connected to one end for fail safe operation) is appropriate. Modern 19mm trimpots have a 0.75W rating, and an insulation rating of about 1kVDC.
- Dave Gillespie identified an imbalance in Chicago output transformers in the Heathkit W4M that was simply adjusted by paralleling 220k with R13 (47k) of the driver stage.
- Tube rolling can show up distortion performance changes.
- Varying the heater power of output stage valves can adjust their balance.

b) Adding an output stage cathode bypass capacitor to lower distortion.

- First referenced in Dec 1950 [11]. Claimed reduction in distortion at higher output signal levels.
- Kiebert [44] in April 1955 measured lowest IM when bypass capacitor was at a particular taping of the cathode bias resistance.
- A contemporary assessment by Mitchell [39] in Nov 1955 used the UTC W-10 Williamson amplifier to measure THD and IM with/without a bypass capacitor for KT66, 5881, and 1614 output valves, and for varying degrees of circuit unbalance. In general, results showed that THD could rise or fall, depending on the valve. CCIF IM results in general showed a slight improvement at higher output power levels with the bypass capacitor.
- Gillespie describes a turn-off stress on the OPT and output stage valves in the Heathkit W5M due to the bypass cap.
- Author measured THD for sample Williamson with feedback removed for (a) standard output stage, (b) capacitor bypass, and (c) CCS (LM317) in cathode. THD was the same for all options, except that standard circuit had lower THD below about 9W.
- During max output and overload conditions, the cathode voltage generally falls, so there is no benefit in deploying circuitry such as a Paul Ruby zener clipper (or simpler zener plus steering diodes) on the output stage coupling CR nodes [32].
- Author measured significant reduction in harmonic and intermodulation distortion in an 807 pentode mode output stage Williamson by adding bypass capacitor.

c) Modifying the CR coupling network corner frequencies for better low frequency stability.

- First referenced in Feb 1952 [12]. Subsequently the change was typically to increase C3, C4 from

50nF to 250nF.

- Phase splitter stage CR coupling frequency lowered (from 6.8Hz to 1.4Hz) to separate it from the now dominant driver stage pole frequency, and reduce its phase shift contribution at the phase margin frequency to help suppress any tendency to motorboat.
- Values as high as 2.2uF have been used to suppress 1Hz motorboating in a Heathkit W-2M, although such large capacitance parts would not have been practical in the 1950's, and any such large parts need to avoid adding stray capacitance to ground. Gillespie also shows a large C3/C4 in the W5M sufficiently lowers their phase shift contribution to obtain LF stability.

- The Heathkit W-5M [10] halved the phase splitter corner frequency (100nF), and quartered the driver stage corner frequency (1uF) to both separate the stage corner frequencies and lower them relative to the output transformer roll-off.
  - Raising the driver stage output coupling cap values may adversely degrade recovery from any transient blocking distortion event occurring from output stage grid conduction.
  - The nominal OT corner frequency is ~6.4Hz for UL mode KT66, assuming 100H, and so becomes the highest corner frequency.
- HMV 3051 raised C6/C7 to 0.5uF and raised R14/R19 to 470k, to lower the corner frequency to 0.7Hz.
- Driver stage CR coupling frequency negated by paralleling C6, C7 with high value resistance.
  - First referenced in Nov 1955 [13].
  - The change used a parallel 470k-1MΩ, with a zero corner frequency ~0.63Hz (~ a decade below pole frequency), to improve phase margin.
  - The added resistor changes the output stage bias level, which needs to be readjusted. Also there is risk introduced that removal or failure of V3 or V4 would lower V5/V6 bias leading to red-plating/damage. Increasing the parallel resistor by up to about 3x should still provide some phase margin benefit and alleviate bias shift.
- Phase splitter stage output CR step/shelving network.
  - Additional parallel Cx-Rx network in series with output coupling C (C3,C4), as detailed by Roddam [20] in 1951, with examples including Partridge (1947), and GEC [38] in 1957. Similar very low DC frequency pole response, but as coupling C impedance falls then a shelf attenuation is introduced by Rx and the driver stage grid leak. Then at higher frequency, Rx is bypassed by Cx, and normal mid-band gain is attained.
  - Similar to the first-stage HF shelf network, the aim is to reduce open-loop gain once outside of the nominal audio band, but still within the amplifiers bandwidth, so that stability margin is improved at the ends of the amplifier bandwidth, and to reduce the likelihood of overdriving the output stage in to grid-conduction from low frequency transients. The aim is for the shelf attenuation to at least offset the gain peak (typically around +4 to +6dB at 2-3Hz), so for the design parameters in [19] then at least n=2 (-9dB shelf), up to about n=5 (-15dB shelf) as typically used by Patrick Turner in his amps to improve LF stability margin [37].
  - With a 250nF coupling cap C, and 470k driver stage grid-leak R, and a 10:1 ratio of C:Cx, and 2.1:1 ratio of Rx:R (where Rx=1MΩ, and Cx = 27nF), the shelf attenuation has a mid-frequency of about 10Hz (27nF-470kΩ), with attenuation starting about 3x 10 = 30Hz, and plateauing at ~ -9dB at about 0.3x 10Hz = 3Hz [19]. Phase shift will start falling to -10deg at about 6 x 10 = 60Hz, and fall to a minimum of about -30 deg at ~10Hz, and then rise back to about -10 deg at 0.15 x 10 = 1.5Hz. The final low pass filter response from C and Rx has a 0.6Hz corner, with a minor influence on phase shift at about 2Hz.
  - Similar shelving response achieved by using a series RC network in parallel with a C, as used by Kiebert [44].
  - Shelving the forward loop gain below some bass frequency like 30Hz allows harmonic distortion components above 30Hz to be attenuated by the mid-band level of negative feedback.

d) Altering the feedback

- Including circuitry to add phase lead at very high frequencies.
    - A capacitor in parallel with R25 was first described by Williamson in 1949 in the 'Capacitive Loads' sub-section, and subsequently referenced in Feb 1952 [12], and by Brittain. Used to provide some 'free' phase margin improvement by placing corner frequency sufficiently above both the gain margin and phase margin frequencies, and to not degrade stability margin for no load or capacitive load conditions. 30 degree advance shift is provided at 60% of the RC corner frequency. As the capacitor tuning method is typically based on a squarewave response for a resistive load, non-ideal speaker impedance may adversely affect a real outcome in which case a zobel network may be appropriate to add (see 8n).
    - Adding a 500pF capacitance from each output stage screen to ground was used in the Radio & Hobbies March 1948 'Triode connected 807 Amplifier' that was closely related to Radiotronics A515 amplifier.
    - An inductor in series with R4 and ground that conducts speaker output current to provide output current feedback. Used to provide some phase advance to offset OPT leakage inductance phase lag [Turner Audio and Turner Audio – note schematic error]. Inductance of circa 1uH adds some influence above 100kHz. Author has used this to allow removal of step network and phase compensation cap, and achieve unconditional stability with no load or capacitance only loads, but note that neg speaker terminal is not a ground for probes. Technique is akin to that used for solid-state amplifiers for unconditional stability into capacitor loads.
- Lowering the feedback level from 20dB when stability margin improvement is needed.
    - The Heathkit W-5M used 18.1dB feedback and suggested that could be lowered should the stability be marginal.
- Including additional output to inner feedback loop.
    - Marshall [35] describes a circuit using a centre-tapped speaker output to provide feedback direct to the driver stage cathodes. No use of this technique has been identified in the literature so far. Typical interleaved secondary windings can usually be configured to allow a CT connection to ground, although that would likely restrict the speaker impedance options.
    - Gillespie describes using a centre-tapped speaker output to provide balanced cathode coupled winding feedback in a Heathkit W-5M.
- Including an additional inner feedback loop.
    - Hafler [22] introduced a 100pF capacitor from driver V4 anode to R4 to achieve additional 12dB margin, with bandwidth reduced from 200kHz to 80kHz and some minor increase in 20kHz distortion. The added capacitor imbalances the driver stage output, and lowers the driver stage RC roll-off from ~550kHz to ~170kHz.
        - A padding capacitor was not added to the other driver anode.
    - Duerdoth patented a subsidiary feedback technique circa 1950 that applies local feedback around the first stage(s) that is frequency restricted to outside the main signal bandwidth, such that the amplifier roll-off beyond the main bandwidth has a slower roll off [28], [29]. The technique requires the first stage(s) to have a very wide bandwidth, and the feedback signal is filtered for appropriate frequency response.
        - This technique appears to be what Hafler has introduced. The driver stage is used for feedback take-off, as the closed loop gain of first stage and phase splitter is only ~ 1.5, whereas the driver stage signal has a gain A1'.A2.A3 = 21. Also, loading the first stage anode would likely lower its corner frequency too much.
        - The first stage has a roll-off ~370kHz, which would likely work with an inner loop feedback corner of 230kHz to provide a subsidiary loop response described by Duerdoth, but not specifically using the patented means.
    - Although not applied to the Williamson amp, Kauder [49] applies feedback from an output stage anode to the global feedback point to control the high frequency peak in a very similar amp circuit.

e) Altering the output stage from triode PP, to 'ultra-linear' PP.

- First referenced in June 1952 [14]. Within 3 months, Williamson responded with an article [15] in Wireless World with co-author Walker from 'QUAD', concluding that QUAD's use of an output stage cathode feedback winding was a better outcome than the UL screen feedback tapping.
- The internal resistance of a beam pentode in UL configuration is higher than in triode mode.
    - KT66 increases from 1.45k to 3.1k at 20% impedance tap [21].
    - For low frequency response, UL mode increases RL corner frequency from 3.9Hz to 6.4Hz, which may degrade low frequency stability margins.
    - For high frequency response, UL mode would significantly lower B=Xc/Ri, and Ro/Ri. The net result should reduce response peakiness, especially for no load conditions, along with some minor reduction in -3dB frequency.
    - The driver stage RC roll-off frequency should increase due to lower V5-V6 miller capacitance (from lower screen voltage turns ratio).
- The UL primary winding segment should have a tight coupling to the remaining primary winding to avoid high-frequency resonance effects from degrading stability [Heath and Woodville, 1957].

f) Fixed bias instead of cathode bias

- Not a practical option for Williamson, as selenium diodes only became available in the late 1940's.
- Sparser & Sprinkle 'Maestro' 1952 amp used fixed bias.
- Hust in Radio & TV News Sept 1953 used fixed bias.
- Hafler in 1956 article [22] on modernising the Williamson.
- Max grid leak resistance needs to be lowered from 500k to 100k, but Williamson circuit uses 100k.
- Easier technique to balance output stage bias currents, and removes need for power pots, but bias pot variation changes loading on driver stage so best to add a suitable bypass cap on each wiper.
- A lower B+ level (~40V) can be operated.
- An alternative is to use a voltage regulator like the LM337 to replace R21 and R22. The EFB$^{TM}$ appears appropriate (with extra protection) although the 40V max in-out limit is approached.
- Fixed bias has been observed to reduce THD and IMD at higher frequencies.

g) DC elevation of heater supply.

- First referenced in Feb 1952 [12].
- Reduces maximum heater-cathode voltage Vk-h for 6SN7 (datasheet recommended limit of 200V).
    - Phase splitter cathode has Vk-h ~ +100V, plus a minor level of signal voltage swing.
- Reduces heater-cathode resistive AC hum transfer when |Vk-h| > ~ 20V.
    - Elevated supply ~ +40Vdc (from a divider, or output stage cathode bypass capacitor if used, or via an RC filter from the output stage cathode).
        - V1 Vk-h ~ -38V ; V2 Vk-h ~ +60V ; V3,4 Vk-h ~ -36V

h) Humdinger pot to minimise heater hum from residual AC coupling to grids.

- First referenced in Nov 1955 [16].
- Setup should be done open-loop, as closed loop feedback will attenuate the hum level.

i) Modifying bias points of V1, V2, V3, V4 to improve signal swing headroom.

- Pilot Radio's Pilotone AA-901 model from 1952 modified the bias settings for V1 and V2.
    - V1 bias voltage was lowered to 1.1V by reducing the supply rail from 305V to 185V. The anode voltage is lowered to 50V, from 100V, to adjust bias for the phase splitter. Given the low gain from the input stage, this appears to have a negligible negative effect.
    - V2 bias is increased to 6.0V by reducing supply rail from 325V to 261V, and increasing arm resistance from 22k to 27k – see Figure 9. The higher bias may help avoid stage clipping.
- Referenced in June 1961 [9]. See section 6.

j) Modifying driver stage

- Add individual cathode degeneration resistors to modify the stage gain and high frequency phase response. Keroes [13] indicated that the addition of 220Ω improved the stability margin.
- Replace the common cathode resistor to ground to improve cathode constant current operation by using a higher value resistor connected to a negative rail, or to a CCS circuit, to improve the balance and 2$^{nd}$ harmonic cancellation and CMRR, which may become significant at high drive signals.

k) Care with component selection, matching, and layout.

- Williamson indicated that certain layout aspects were significant:
    - Coupling between output and power transformers and first choke can degrade hum level.
    - High level and output level signal cables should be well separated from input wiring. An under-chassis metal screen between V1-2 and V5-6 was recommended to isolate V5-6 anode/screen voltages.
- Stray coupling capacitance from wiring and parts to either ground, or other circuitry, can noticeably affect high frequency roll-off, especially for the first stage and driver stage.
    - Coupling caps C3, C4, C6, C7 should not be placed on or near the chassis or other wiring, and the outer foil end should connect to the anode side of circuitry. The layout of capacitor pairs (eg. C3 & C4) should have similar parasitic capacitance to ground to maintain balanced high frequency performance. Avoid the use of metal can enclosed capacitors, where the can is mounted to chassis, due to stray capacitance to can (ie. to chassis).
    - Avoid the use of shielded cable due to its shunt capacitance to ground. If shielded cable is used for the feedback signal then use low capacitance cable and keep it as short as possible and keep any RC corner frequency above 1MHz.
- Kiebert was the first reference in Aug 1952 [17] to highlight the need for careful parts selection so as not to degrade distortion or noise performance.
    - The original parts list used mainly ¼, ½, and 1W carbon composition resistors. Vintage CC resistors all typically drift > +20% over the decades, especially if they have DC voltage across them. Modern metal oxide 2W resistors are small, low noise, and stable with 500V rating.
    - Ready access to resistance and capacitance meters, and 1% tolerance resistors, can now achieve excellent matching.
    - Free PC distortion measurement software can be used to minimise harmonic and intermodulation distortion, when feedback is disconnected, by tube rolling in all stages (eg. to match driver and output stage dynamic balance). A typical simple soundcard allows distortion assessment to well below 0.01%, and noise floor confirmation to < 90dB below rated output.
    - The input, phase splitter and driver stages could use 12AU7 with no need for any circuitry changes, except for the valve holder and heater. An example is the Heathkit W-5M. However harmonic distortion is likely to be higher, especially for a 12AU7 in the driver stage which operates at higher signal levels – see section 6.
    - Kiebert (1952) recommended changing the first 6SN7 for a 12AY7 to lower hum, however that result may have been achieved by using an elevated heater or as a result of circuit resistor changes needed due to significantly different parameters. Kiebert later (1955) identified high IM from the 12AY7 in phase splitter stage.

- Kiebert recommended changing the driver stage 6SN7 for a 5687 due to better triode matching. The 5687 has significantly different parameters and needs circuit resistor changes.

- Parallel each driver stage grid leak resistor (R8, R9) with 100pF to lower inter-stage RC corner frequency from ~5MHz to ~2MHz. The added capacitance dominates the driver stage capacitances [23], but need to be matched to maintain high frequency symmetric loading on the PI stage.
- Different output stage valves may allow a different idle power dissipation level and B+ voltage rating for triode operation to be chosen, but redesign of power supply and common cathode and grid leak circuitry is likely required as well as an awareness of electrode capacitance and plate resistance differences (eg. EL34 has slightly higher input and screen capacitance), and whether the existing driver stage can provide a sufficient peak signal level. In particular, the 807 can be susceptible to local resonances out past 1MHz, with an early R&H March 1948 clone based on Radiotronics A515 adding 500pF from screen to ground.

l) Increasing power supply decoupling values.

- The value of C9 can't be increased due to rectifier peak continuous current limits (see 5). Changing the rectifier type may allow an increase in C9 (see 5).
- Increasing C8 from 8u to 22u dampens any ringing on the output stage rail due to a step load change, but also increases the hot turn-on overshoot and diode peak current levels by a small amount.
- Increasing C5 as well to 22u dampens any ringing on the driver rail supply.
- Increasing C1, C2 power supply decoupling cap values lowers the corner frequency of the decoupling zeroes. Increasing C1 may reduce the phase shift benefit on low frequency stability margin. Increasing C2 would reduce phase splitter asymmetry but may influence total response.
- Many modern electrolytics are not rated to the 500V working level, with power-on levels to 600V as the diode valve typically starts to conduct prior to the remaining amplifier valves (the amplifier has no bleed resistor on the power supply), even with the indirectly heated cathode diode. When caps are used in series, a 47uF 350V radial or axial leaded electrolytic is relatively small in size. Radial leaded electrolytic capacitors of small physical size have negligible self-inductance, and their impedance is low for frequencies out to 1MHz. There is likely to be negligible benefit in bypassing a radial electrolytic cap with a lower valued poly cap, given the voltage and capacitance ratings used in the Williamson amp.
- Metallised polypropylene film caps could be used in place of electrolytics when layout and space is available, as they can allow adequate capacitance voltage rating in a single capacitor. Capacitors like ICEL's MAB A02 400VAC class B motor run cap have a 600VDC continuous, and 750VDC peak spec, and come in sizes up to 15uF. However, they can exhibit a very significant impedance notch around their first self-resonance (~1MHz) which may not be as benign as an electrolytic capacitor's flat (but higher value) impedance response.

m) Using a different output transformer

- Williamson described the detailed design of the output transformer in his April and May 1947 WW articles. During circa 1945-6 when the amplifier was developed and demonstrated, Williamson used an output transformer made by Vortexion (who were located nearby) – Williamson specified the leakage inductance for Vortexion. Commercial parts were first advertised in July 1947 in WW. Bert van der Kerk describes modern-day construction of the transformer and an improved version.
- Any transformer that varies in any way from that design could affect stability performance at the low and high frequency ends of the frequency spectrum. The measurement of output transformer inductance and impedance and frequency response for the Partridge WWFB, Ferguson OP25 and Red Line AF8 are presented in [53].
- The 1953 Naval test report [41] is indicative of the range of performance that some have experienced. The report indicates that the feedback circuitry was optimised for each transformer in an amp. The Williamson amp was constructed from parts and used 807's, a 250k input pot with 10k grid stopper for input, and had increased C1, C2 and C5 values with a regulated 400VDC supply. The amp was tested with the following transformers:

- Peerless S-265-Q
- Partridge WWFB/1.7
- Partridge CFB/1.7
- Audio Development 314E
- Stancor A-8054 (tested default, and with optimised feedback and input)

- The phase shift introduced by the output transformer, both at very low frequency due to the primary inductance, and at very high frequency due to leakage inductance, may require additional stabilisation measures to be applied [38]. Williamson introduced a high frequency shelf network in the 'new improved' 1949 WW articles to alleviate the possibility of high-frequency instability. GEC [38] illustrate the use of a similar low frequency shelf network, and discussed in thread in [37].
- Some people have experienced widely different perceived performance when comparing different output transformers. This may be entirely due to lack of care in checking that each transformer was appropriately set up in a test amplifier, or in the speaker loading used.
  - Some amplifiers operate the output stage into class AB. The output transformer leakage inductance between primary half-windings then becomes an issue as one half-winding stops and starts conducting. In addition, the effective shunt winding capacitance doubles when the other primary half winding stops conducting. Both effects can change the high frequency response. Note the difference in coupling between half primaries for Partridge WWFB and CFB transformers [27], and the discussion in [31].
  - A UL configuration can cause additional couplings that may degrade stability and response, as indicated by [30].
- Known output transformers with 10kΩ P-P primary impedance, and designed for or used in a cloned Williamson amp, or a close variant (note that multiple parallel valves or UL may be better with a different primary impedance) are:

| | | | | |
|---|---|---|---|---|
| Vortexion, UK | Series I | | | |
| Vortexion, UK | Series II | from June 1951 | [1] | |
| Partridge, UK | ? | from Sept 1947 | | |
| Partridge, UK | WWFB | from Aug 1949 | [1] , [27] | |
| Partridge, UK | CFB | from 1951 | [27] | |
| Partridge, UK | T/CFB | from 1954 | [45] | |
| Savage, UK | 2B36 | from Dec 1949 | Ref | |
| Savage, UK | 3C67A | | [1] | |
| Elstone, UK | MR/W | | Ref | |
| Woden, UK | WOT.25 / WOT.26 | from Jun 1949 | Ref | |
| Gilson, UK | WO.1796A | | Ref | |
| Gardners Radio, UK | O.P.735 & O.P.736 | | Ref | |
| Red Line, AUS | AF series | from Dec 1947 | Ref | |
| Ferguson, AUS | OP25 | from Feb 1948 | Ref | |
| Bramco , AUS | HF-4 | | | |
| Beacon Radio, NZ | 48S06 | from July 1948 | Ref1, Ref2 | |
| Wiseman Electric, NZ | 'Williamson' | from Oct 1953 | Ref | |
| Stancor, USA | A-8054 | | | |
| UTC, USA | LS-63 | | Ref | |
| UTC, USA | LS-60A | | Ref | |
| TRIAD, USA | HSM-81 | | | (8kΩ P-P) |

| | | | | |
|---|---|---|---|---|
| o | Freed, USA | F-1959, KA-10 | | |
| o | Hammond, Canada | 1770, 1772 | | Ref |
| o | Jorgen Schou, DK | Type 350 | | Ref |
| o | Unitran, Holland | 10U72 | early 1950's | Ref, Ref, Ref |
| o | Sansui, Japan | HW-731, HW-733 | from April 1954 | |

n) Adding a speaker output zobel network.

- The Heathkit W-5M from 1955 used a 0.1uF-47Ω speaker output zobel network, with 34kHz corner frequency, to maintain output loading. Heathkit referred to the circuit as the 'tweeter-saver', which allowed un-loaded operation of the amplifier as well as flattening the frequency response and smoothing out the phase variation when loaded with a speaker (compared to a test resistive load).
- An option to control the high frequency response of the transformer when the amplifier output is unloaded or capacitively loaded or loaded by a speaker. As feedback level falls, the default output response can include a peaked region (where gain and phase margins could diminish leading to degraded stability), depending on loading impedance at the frequency of the peak – see [41] and [42].
- A well designed multi-driver speaker system can provide a smooth impedance-frequency characteristic throughout the audio range and beyond. In contrast caution would be required for speakers with uncompensated, high-rate crossover networks.

o) Tuning the first choke CH2 to alleviate mains 2nd harmonic ripple.

- Although not an uncommon technique, this change is not known to have been applied to the Williamson amp in the 1950-60's. www.keith-snook.info
- The choke is bypassed with a capacitor to reduce the 2nd order harmonic output level, but causes an increased level of higher order mains harmonics. Given the likelihood of choke inductance being higher than its rated value when DC current is below the rated level, and given the increase in higher order harmonics with increasing capacitance, it is recommended that a lower capacitor value is used than what would be expected to tune the rated inductance at 2f – perhaps at least 20% lower. A 180nF 630V capacitor has a 118Hz self-resonance with a 10H choke. A 330Ω dampening resistance in series with the capacitor (ie. about twice the choke DCR) is recommended.
- If applied, then the C8 filter capacitor value should be increased to attenuate the higher order mains harmonic levels.
- Placing a common-mode choke prior to C8, with >100mH differential inductance, may be beneficial in maintaining high series impedance above 10kHz. The choke winding would need suitable insulation. Hum current may bypass the choke due to transformer winding stray capacitance to chassis.

p) Additional protection items

- Power transformer HT secondary over-current protection, such as fuse the CT [8] – see section 5.
- Power supply diode failure protection.
    - o SS diode in series with each valve diode anode to reduce PIV stress and arcing due to age, especially if a 5V4 is used (see section 5). Given the 425VAC secondary rating, at least two 1N4007 in series would be needed for each valve diode anode [47].
- V2 grid-cathode over-voltage protection to reduce stress at power on.
    - o Not a commonly reported failure mode, but Vgk is stressed at turn-on, in excess of +450V and up to +600V, until V1 and V2 start conducting.
    - o A neon tube (eg. NE2) across grid-cathode clamps voltage to ~100V, and has ~1pF shunt capacitance. Using a 1N4007 may be acceptable, as leakage current is negligible, and junction capacitance rises to about 20pF at zero bias, and so may have no influence at -2V.

- Output transformer primary winding over-voltage protection.
  - Not a commonly reported failure mode. Topic discussed in [46].
  - Windings with low leakage inductance and class A operation alleviate over-voltages.
  - Secondary winding has a minor amount of loading from feedback network, and any added zobel network.
  - A MOV across each primary half-winding with a minimum DC 1mA voltage rating of at least about 600V, and no more than about 800V. A resistor in series with the MOV (or series MOVs) could provide an RC filter corner frequency for some additional phase margin. For example, using two 7mm disk 330VDC 90pF part in series would add about 45pF of additional shunt capacitance across each half-primary (or use with a series resistor), which is about an 8% increase in half-primary winding lumped shunt capacitance for a WWFB OT.
- Output stage bias protection.
  - The variable resistors R17 and R21 pose a risk from an open wiper. R21 should have the wiper also connected to the unused end terminal. R17 should include an extra resistor from pot wiper to each end terminal.
- Power supply bleeder.
  - It is always advisable to include a power supply bleeder resistance, to reduce any chance that power supply or coupling capacitors remain charged after the mains supply is disconnected. Voltage balancing resistors across a series connection of electrolytics would achieve that function, as would a resistor divider for an elevated heater supply.
- Mains AC turn-on stress.
  - An ageing power transformer may be more prone to stress from primary winding in-rush current (ie. when heater filaments are cold). An NTC thermistor could alleviate that stress.
  - The turn-on timing of the 5V4G may be sufficiently before KT66 conduction starts, causing all B+ rail filter capacitors to be stressed. An NTC thermistor added in series with the 5V4G heater, and designed for 2Arms continuous current with a suitably low voltage drop, can slow the rise of B+ till after KT66 conduction starts. Eg. a CL-21 with a heat-shrink cover. Alternatively, a relay contact can short out the influence of the NTC as described in [47].

q) Additional items

- The 100kΩ grid leaks for the output stage were increased to 220kΩ, and the coupling cap lowered from 0.25uF to 0.1uF in one commercial amplifier – see 7 for discussion.

Note that some commonly applied changes to vintage amplifiers should not be applied without careful assessment. For example, changing from valve to ss diodes in the power supply may add substantial turn-on stress to the output stage valves, due to the rapid rise in B+ in a shorter time than the CR coupling time-constant, or bypassed cathode time-constant if bypassing is applied.

Some common changes are eminently worthwhile, although diy assessment is needed, and include:

- Power supply secondary fusing and valve diode protection ss diodes.
- Output stage bias adjustment protection.
- Output stage individual valve bias current measurement.
- Power supply bleeder.
- Improved low frequency stability by increasing C3/C4 values, and reducing C6/C7 values.
- Improved high frequency stability by assessment using modern gain-phase measurement and application of known compensation techniques to provide high-frequency stability assurance for no load, and substantial capacitance loading.

## 9. Setup, Testing and Restoration

The original articles [1] identify the following procedures to set up correct operating conditions:

- Balanced output stage DC idle currents set by measuring for zero DCV between KT66 plates as R17 is adjusted (assumes output transformer primaries have the same DCR).
- Minimum output stage signal current imbalance with a 400Hz test signal when driving a medium signal level into a resistive load.
    - Imbalance of gain between V3 x V5, and V4 x V6, will cause asymmetric signal voltages on each primary half-winding of the output transformer. Each valve's voltage gain can be expected to vary by up to about 10%. Even with matched triodes for V3/4, and V5-V6, the distortion can be expected to be tweaked to a lower level.
    - Tested using a sensing transformer primary winding inserted in the B+ lead to the output transformer [48] and listening to a headphone connected to the sensing transformer secondary. Alternatively, measure the AC residual voltage at the common cathode node.
    - Minimising signal imbalance can be achieved using a trimmer resistance across R11 or R13 (similar to pot R12 in the original schematic of the Williamson articles), and possibly a trimmer capacitor across V3 or V4 anode to cathode (when using a high frequency test signal).
- 20dB feedback level confirmed by disconnecting R24 and terminating it to ground via 470Ω and measuring signal voltages. Some ambiguity remains about whether R4 should be bypassed or not.
- Output transformer primary inductance value measurement. Measure AC current with 5VAC mains frequency excitation across total primary winding.

DIY test equipment at the time was typically an AVO Model 7 multimeter, and an oscillator. The meter would allow part resistance and capacitance checking and matching, and circuit operating voltages (at best with 1kΩ/volt loading on the circuit, requiring some care in choosing what to measure, especially around V1, V2).

It is surmised that Williamson used an oscilloscope such as a Cossor 339 to make the signal level measurements for 15W output, as the readings are given as peak voltage [3]. Only an oscilloscope or a VTVM voltage meter would have had sufficiently high impedance for measuring V2 and V3 grid voltage levels. Few people had access to an oscilloscope in the mid 1940's, and oscilloscope technology was very rudimentary, with no calibrated graticules, or constant impedance inputs yet available, and no probes were provided or were available, so a direct connection appears to impose about a 1MΩ load.

Nowadays, an oscilloscope and waveform generator are typically available for observation of test waveforms, and with sufficient bandwidth in to the 100's of kHz, and fast rise-time square-waves. Typical test techniques include:

- Square and sinewave response assessments with:
    - Changing feedback level
    - Changing load (resistive matched load, no load and capacitance loading)
    - Changing stability margin circuits
- Low frequency roll-off measurement:
    - A meter may not be able to adequately measure low frequency response (eg. below 10Hz).
    - Direct oscilloscope measurement of a square-wave with a 10% droop of the stepped level at a frequency F indicates a -3dB corner frequency of F/30 for a single-pole amplifier response.
    - Direct oscilloscope X-Y measurement of output-to-input gain and phase when feedback is disconnected can confirm amplifier response down to below 1Hz.
- High frequency roll-off measurement:
    - A meter may not be able to adequately measure high frequency response (eg. above 50kHz).
    - Direct oscilloscope measurement of a sinewave when feedback is disconnected can confirm

[3] Alternatively, a VTVM would provide an equivalent high AC impedance load of at least 1MΩ, however an AVO 7 on a low AC voltage range would heavily load the circuit.

amplifier response.

- Direct oscilloscope X-Y measurement of output-to-input gain and phase when feedback is disconnected can confirm amplifier response.
- Direct oscilloscope measurement of a square-wave when feedback is applied can indicate amplifier resonance response.

Testing for low and high frequency stability margins is highly recommended, even though it can be arduous due to the very low and high frequency ranges involved, and the need to confirm across a range of loading (and even secondary winding configurations). A good understanding of gain and phase margins [50],[51],[52], and the influence of remedial networks such as shelf and zobel networks, and the roll-off and resonance behaviour of an output transformer, is required to appreciate what if any circuit modifications may be needed.

Square-wave testing is likely the simplest method available to indicate what high frequencies resonances may be at issue, and how well they are controlled [54], including the common additions of a step network, a compensation capacitor, an output zobel network, and not-so-common inductor in series with R4. Testing does require an oscilloscope and probes with flat response to well beyond 1MHz, and a squarewave generator with similarly fast risetimes, so as to adequately excite and view output transformer resonances that can occur out at a few hundred kHz where gain margin and phase margin frequencies are likely occurring. Testing with ‘difficult’ loads such as an open-circuit, and just capacitance, and confirming the square wave is still stable (albeit quite resonant) is then a valid means of high-frequency stability assurance.

Another indirect method for testing stability margin is to reduce feedback resistor R25 value, thereby increasing feedback level beyond 20dB, and identify when instability starts [55]. This method avoids the need for some test equipment, such as an oscillator with a very low or high frequency range, but may need care to avoid causing HF instability when focussing on LF stability performance.

Nowadays, PC soundcard or USB interface equipment, as well as custom interfaces, are readily available, and can include and automate detailed and complex audio measurements [53]. Spectrum analysis, harmonic and IM distortion, and noise floor level measurements can be easily achieved with low-cost soundcard and free software. However common commercial soundcards are frequency band-limited to perhaps down to 5-10Hz, and up to 48-96kHz, and care is required to acquire accurate measurement detail out to those frequency limits. As such, the soundcard measurement approach is unable to acquire extended low and high frequency response detail relevant to stability margins.

Recent advances in automated measurement of Bode plots for switchmode power supplies allows practical frequency measurement from 0.1Hz to 1MHz with sufficient dynamic range to indicate gain and phase margins, such as with the PicoScope 2206B or 4224A using FRA4PicoScope software [53]. The author has used this method to confirm stability related response and achieve (practical) unconditional stability, as well as measuring the frequency response of impedance of small inductors.

Also note that as measurements extend to low levels of distortion, or frequencies above the audio range, even ‘simple’ test rig parts like a load resistor need careful selection to avoid parasitic non-linear resistance and inductance from contributing to distortion and instability.

As such, a “modern” Williamson, using matched, stable, modern passive parts and selected valves can be readily tested and should achieve the same performance as Williamson reported.

As a last comment, a hi-fi speaker with 90dB/W/m sensitivity can provide an 80dB average listening SPL at 3 meters with 12dB peak-to-average recorded material without exceeding a 15W power capability where waveform clipping starts. For comparison, a very loud cinema type experience may operate up to 85dB average SPL with up to 20dB peak-to-average recorded material.

**Restoration**

It is often difficult or impossible to restore a vintage Williamson amp and retain the aesthetic look of vintage parts and chassis, especially under a chassis. Of most concern is to use parts with adequate voltage rating to cover start-up surge voltage, ie. >550V, that can stress all parts before valves start loading down voltage rails when directly heated diodes are used.

Vintage capacitors, whether they be for power supply filtering or coupling, will require testing for capacitance, ESR and leakage current if the aim is to retain parts, and some resort to part stuffing. Few modern e-caps have voltage ratings above 500V, so two series caps with balancing resistors is often a practical replacement.

Vintage carbon resistors nearly always drift high, and typically by more than +20%. One option is to place a small modern metal-film resistor (with suitable voltage rating) surreptitiously behind a larger carbon part, in parallel, to return the part value back to nominal. General restoration of valve amps article link is [56].

## 10. References

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