This article collates measurement details on output transformers designed for the Williamson amplifier circuit.

The aim of this article is to identify a convenient measurement setup, along with test results for contemporary output transformers made for the Williamson amp, including the Partridge WWFB/0.95, Ferguson OP25/15 and Red Line AF8/2.

Tests include primary inductance as a function of excitation voltage at 50Hz, and impedance as a function of frequency at a fixed voltage for both open and shorted winding configurations. Frequency response of each transformer was tested, which included harmonic distortion measurement. Frequency response of each transformer in a Williamson amp was also tested.

The difference in measured values between commercial product, as well as from different speaker impedance settings are discussed.

Low and high frequency stability performance of an OPT in the Williamson amplifier is discussed using Williamson's amp gain and phase results and the 1950 technical assessment by Cooper. The squarewave and stability response of the Ferguson OP25/15 was measured in a Williamson amp, and showed that only additional phase compensation was needed for stability and good squarewave response.

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1. Preamble

The 1947 Williamson amplifier [1] was designed around an output transformer that could achieve:

- flat frequency response from 10Hz to 20kHz with phase shift limited to 20deg at 10Hz.
- -3dB frequency bandwidth of 10Hz/3 = 3.3Hz, and 3*20kHz = 60kHz.
- some closed-loop gain margin (with 20dB feedback when phase shift reaches 180dg).

This article assesses the measured primary winding impedance of 3 contemporary Williamson output transformers (commercially available from early 1948) with respect to inductance and phase shift, leakage inductance, and high frequency resonance. Measurement of frequency response is also made, which includes harmonic distortion. Transformer performance at both low and high frequency is compared with what Williamson reported. This article only presents discussion and results related to push-pull class A operation of the output transformer (ie. discussion and results of half-primary performance for class AB are avoided).

Section 2 details the measurement setups used:

- Primary inductance with stepped excitation voltage using mains frequency.
- Primary impedance from REW Impedance measurement setup.
- Frequency response from soundcard and balanced line driver.
- Frequency response from signal generator and scope with floating inputs.
- Frequency response from a Williamson amplifier over a 0.5Hz to 1MHz range.

Section 3 details the primary winding impedance when secondary windings are open circuit.

Section 4 details the primary winding impedance when secondary windings are short circuited.

Section 5 details frequency response results, including harmonic distortion, from 2Hz to 95kHz.

Section 6 details frequency response results in a Williamson amplifier, from 0.5Hz to 1MHz.

Section 7 details high frequency stability results in a Williamson amplifier.

Section 8 provides a summary of test results and some discussion.

Primary inductance target

Williamson identified that a -3dB low frequency first order filter response at 3.3Hz required an incremental primary winding inductance Lpp of at least 96H based on the filter response:

2π x 3.3Hz x Lpp > Rpp // Rload

- Rpp is the effective driving resistance of the KT66 push-pull circuit, and Williamson uses a value of 2.5kΩ (or Ra=1250Ω per KT66).
  - Williamson likely measured Ra or had internal MO data, as the earliest KT66 datasheet is from 1954 and shows Ra as 1450Ω at the nominal operating conditions of the amp.
  - It is not known if Williamson included the DCR of the output transformer primary winding.
  - Partridge used a 3.3kΩ driver resistance when measuring WWFB frequency response.
- Rload = 10kΩ is the reflected secondary side loading.
- Rpp // Rload = 2.5k x 10k / (2.5k + 10k) = 2kΩ
- Lpp > 2kΩ / (2π x 3.3Hz) = 96H

The ‘incremental’ inductance relates to the AC inductance when a DC current is flowing. In a push-pull configuration the core experiences a net zero DC bias and so inductance measurement with no DC bias is appropriate.

Williamson reported a measured Lpp=100H at 5V, 50Hz. If the KT66 datasheet Ra of 1.45kΩ is used, along with primary winding DCR, then the required Lpp > 120H.

Leakage inductance target

Williamson identified that a target -3dB high frequency first order filter response at 60kHz required a primary winding leakage inductance Llk of at most 33mH based on the filter response:

2π x 60kHz x Llk < Rpp + Rload

- Llk < (2.5kΩ + 10kΩ) / (2π x 60kHz) = 33mH

Williamson reported a measured Llk = 22mH at 1kHz. If the KT66 datasheet Ra of 1.45kΩ is used, along with primary winding DCR, then the required Llk < 35mH.

Tested output transformers

Manufacturer ratings for output transformers tested in this article are presented in Table 1. The Ferguson OP25 (early version) and the Red Line AF8 output transformers were commercially available from early 1948, and the Partridge WWFB from Aug 1949. It is likely that the Radiotronics No.130 (March 1948) article "Intermodulation Measurements on Radiotron Amplifier A515" used the early version of the Ferguson OP25.

Output transformer	Partridge WWFB/0/0.95	Ferguson OP25/15 [1]	Red Line AF8/2 [2]
Push-pull impedance	10kΩ	10kΩ	10kΩ
Primary inductance	130H @ 30Hz, 5V 100-130H @ 50Hz, 4V	110H @ 50Hz, 5V	>100H @ 5V >125H @ ?
Primary leakage inductance	15-20mH @1kHz	14mH	17mH @ 5V 1kHz
Primary DC resistance	220+220Ω	200+215Ω	-
Continuous power rating	16W	20W	34 D.B. [3]
Maximum DC current	80mA	-	-
Permissible DC unbalance	20%	0%	NIL [4]
Primary self-capacitance	500-580pF per half	-	-
Secondary winding sections	8 (0.95,3.8, 8.5 or 15.2Ω)	2 (3.75Ω or 15Ω)	2 (2Ω or 8Ω)
Alternating lamination arrangement (lam thickness)		1+1 (0.025”) new 2+2 (0.015”) old	3+3 (0.015”)

Table 1. Output transformer manufacturer ratings

In Section 6, the frequency response of additional output transformers was also tested, including the:

- Ferguson OP25/8 original (old) in 8Ω configuration
- Trimax 996A in 16Ω configuration
- Partridge CFB-1.7 in 6.8Ω configuration

[1] The original (old) version of the OP25 has a black hammertone finish, slightly smaller core stack thickness and changed primary terminal separation than a newer version, but it is not sure when the new version came out.

[2] Earliest 1948 part was labelled AF8.

[3] 34 dBm relates to 2.5W. It is unclear if 34DB relates to another scale, or is an error (ie. 43dBm is 20W).

[4] A rating plate with 5mA has been seen.

Figure 1. Photos of tested output transformers (including old and new OP25)

2. Measurement setups

Primary inductance measurement at mains frequency

This setup measures the mains frequency (50Hz) AC current through the transformer primary winding and calculates inductance based on the AC voltage across the winding. The secondary windings are open circuit.

Compared to the 1940's, modern true-rms meters can accurately measure the low AC current level either as a current through the meter, or as a voltage across a current sense resistor. The DC winding resistance is about 1% of the inductive impedance (2π.f.L) and so has little effect on measurement accuracy. The parallel loading of a modern meter when measuring winding voltage also has little effect on measurement accuracy.

Primary impedance measurement using a soundcard interface and REW software

A soundcard interface with two input channels and one output channel can be setup to accurately measure device impedance (magnitude and phase) over the frequency range of the soundcard, as shown in Figure 2. REW software [4] provides a 3-part calibration procedure to identify an open-circuit output, a shorted output, and then a reference resistance value to achieve accurate measurement performance. In effect, reference capacitors or inductors are not needed for accurate calibration, only a single high-tolerance resistor [5].

Preferably the soundcard has a headphone output with an internal amplifier that can drive low impedance loads, and has at least a 96kHz upper limit (192k sample rate), as OPT resonance between primary leakage inductance and self-capacitance is likely above 48kHz. A 1% test resistor of a few kΩ is used to confirm that the spectrum plot of resistance and phase over the frequency span is valid (including that the Controls setup has 'Use a log axis for impedance' checked).

Bandwidth limitations of the input and output channels of the soundcard are compensated (both for magnitude and phase) by REW during the calibration procedure. However, REW can generate signals from 0.1Hz (View setting) to the maximum sampling rate limited frequency of the soundcard, and so if the soundcard channels introduce significant signal drop off and phase shift at the measurement frequency extremes then compensated results may become suspect or noisy out towards those extremes (eg. < 1Hz, and > 80kHz).

For measurement of primary inductance, primary self-capacitance, and primary resonant frequency, the OPT terminals should be configured with the primary CT linked to the secondary gnd terminal and to the core – the secondary windings should be linked as per the speaker impedance requirement but the speaker output terminals left open-circuit. Note that different secondary winding configurations for different speaker impedances can noticeably affect the measured levels.

For measurement of primary leakage inductance and primary resonant frequency, the OPT terminals can be configured with no links from primary CT to secondary terminals or to the core – the secondary windings should be linked as per the speaker impedance requirement, and the speaker output terminals short-circuited.

As an example, an EMU 0404 USB soundcard was configured for impedance measurement using the headphone output, and a 100Ω 1% sense resistor (allows measurement of devices with negligible impedance without loading the headphone output). The headphone open-circuit output voltage level droops 10% at 3Hz and 90kHz, and 30% at 1.5Hz and 94kHz. A loopback test shows -11dB drop off at 2Hz and 94.5kHz, with 140deg phase shift at 2Hz, indicating the input channel requires more compensation than the headphone output, but overall the drop off is not so severe as to significantly limit calibrated measurement bandwidth.

The REW impedance plots shown in this article use 1 repetition of a 4M sweep length with the noise filter off. The output level is the same as used for impedance calibration at about 1.5Vrms. An ASIO driver is used for the soundcard and set for 192kHz. The plots often show minor glitches at mains frequency harmonics as the soundcard and PC were both mains power (those glitches can be mitigated by using a laptop and battery powered soundcard). There can also be some glitches at specific high frequencies due to one input channel of the soundcard exhibiting internal noise (a known issue with the EMU).

The soundcard signal ground is common to both the headphone output and the two input channels, and so connects to one of the primary winding plate terminals – this has no consequence for impedance measurements where the transformer under test (DUT) is kept distant from any metalwork and has negligible capacitance coupling to the soundcard, or for any combination of links from the primary CT to the secondary.

[5] I used a 10Ω HOLCO H8 Y (0.05% 15ppm TCR) resistor as the reference resistor for calibration, and that achieves accurate measurement of 0.1% tolerance resistors up to 47kΩ, and a 99uH reference grade inductor.

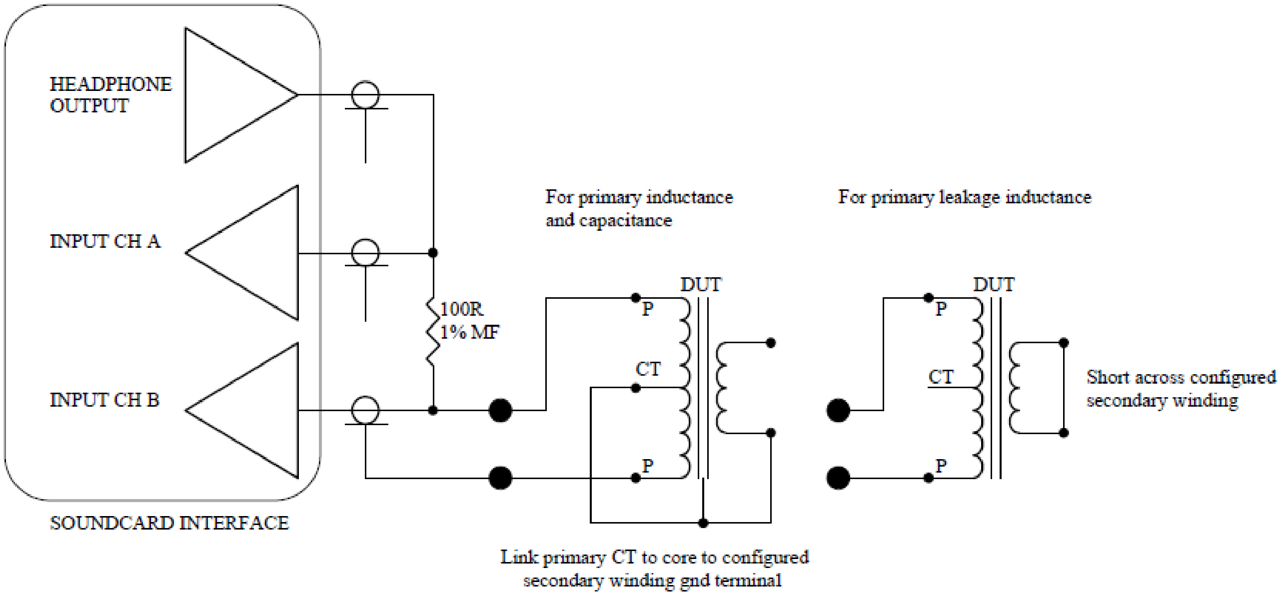


Figure 2. Soundcard connections for Impedance measurement

Frequency response measurement using a soundcard interface, balanced line driver and REW software

Similar to the impedance measurement setup, the soundcard interface headphone output provided the signal source, but a balanced line driver was then used to amplify the soundcard output, and to provide two balanced outputs to drive the PP terminals of the DUT through a total drive resistance of 2.5kΩ as shown in Figure 3. The DUT secondary winding was configured for a rated load, and the DUT output terminals were loaded with a resistor of the same value. The soundcard output and input grounds are connected internal to the soundcard interface. The ground of the soundcard unbalanced input was extended to the secondary winding gnd terminal, which was linked to the primary CT terminal to provide a balanced PP drive signal with all relevant DUT terminals at signal ground to provide consistent capacitance loading within the DUT. The mid-band line driver output level was 4.1Vrms, with excitation down -2.2dB at 2Hz and -0.9dB at 90kHz due to the soundcard. The drive level at the DUT PP terminals is attenuated by the total drive resistance of 2.5kΩ.

The balanced line driver comprised an NE5534A inverting input gain stage (x6.7), followed by two NE5534A unity gain stages – one inverting and one non-inverting. The unity gain stages were capacitor compensated (33pF), and the input and inverting stages were trimmed for offset, and the inverting stage trimmed for common-mode output <-70dB using the signal at the middle of two 1.62k 0.1% resistors loading the output. A loopback calibration file that includes the line driver (one side of driver output connected to input channel) provided compensation for the measurement system gain and phase deviations at low and high frequencies, and the same test levels used for the calibration were also used for test results.

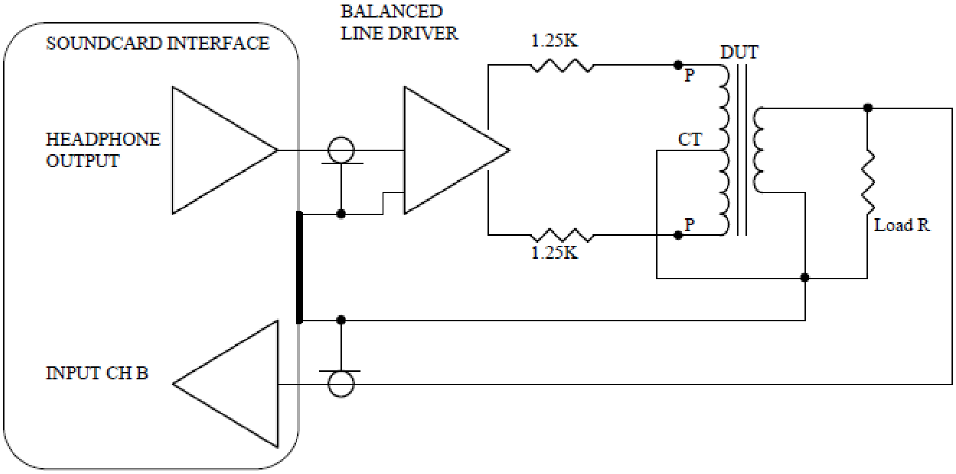


Figure 3. Soundcard and driver connections for Frequency Response measurement

Frequency response measurement using a signal generator and digital scope with floating inputs

Similar to the frequency response measurement setup with soundcard and balanced line driver, the signal generator output provides a sinewave to the PP terminals of the DUT through a total drive resistance of 2.45kΩ as shown in Figure 4. The DUT secondary winding was configured for a rated load, and the DUT output terminals were loaded with a resistor of the same value. The channel 1 and channel 2 inputs to the digital scope are floating, which allows the DUT primary winding CT to be connected to the secondary winding to allow a balanced PP drive signal when the DUT is floating. The mid-band DUT PP drive level was 10.0Vrms constant from 0.1Hz to 1MHz.

Amplitude response was calculated from the peak level shown on Channel 1, relative to the midband peak level. The waveform at very low frequencies (<1Hz) was quite distorted so the peak voltage is not directly related to a sinewave response.

Phase response was determined by inspection of the X-Y plot. The digital scope shows no internal phase shift when displaying signals to 1MHz. A persistence control allowed easier viewing of waveforms for frequencies below about 4Hz. No effort was made to calculate phase shift, and only visual identification of 90deg and 180deg conditions were recorded.

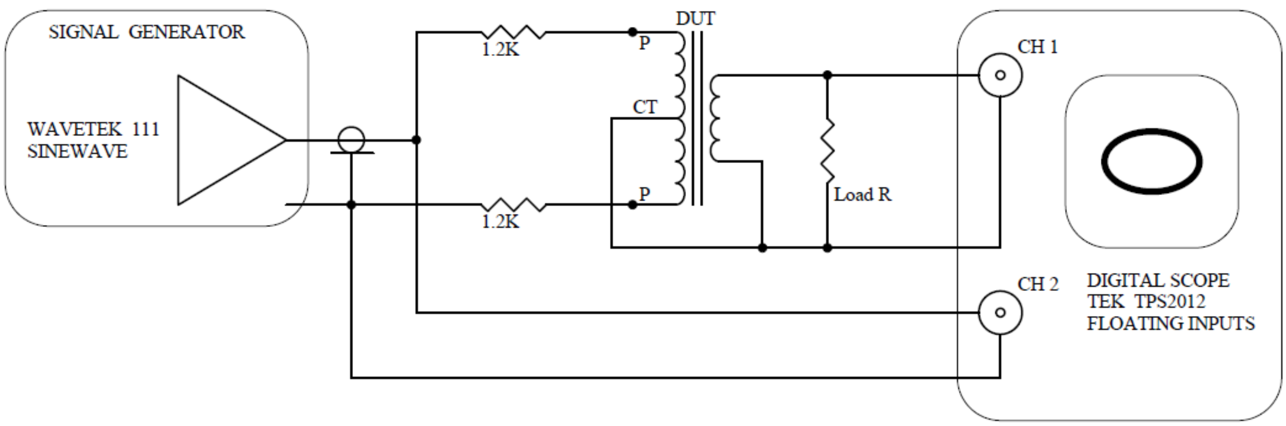


Figure 4. Signal generator and digital scope connections for Frequency Response measurements

Frequency response measurement of Williamson amplifier from 0.5Hz to 1MHz

The gain and phase responses were measured on a Williamson amplifier, for each output transformer. Either a nominal 8Ω resistive load or a 15Ω resistive load was used on the amplifier, and the feedback resistor (R25) was set at 3k4 or 4k7 respectively to generally provide about 20dB of feedback for the particular output transformer in use – otherwise the operating conditions and the amplifier itself were common for all testing. The WWFB transformer has a nominal 8.5Ω secondary, but was loaded with 8.0Ω, and 3k4Ω feedback, so the achieved feedback was about 18dB. For the 8Ω secondary configurations, the amplifier provided about +17dB of midband gain with feedback applied, and about +37dB with feedback disconnected. The 15Ω secondary configurations provided about an additional +2dB of mid-band gain.

A PicoScope 4224A used with the FRA4PicoScope application provided a frequency response sweep and resulting plot of gain and phase. To simplify plot comparisons, a separate low frequency sweep from 0.5Hz to 1kHz was made, along with a separate 1kHz to 1MHz sweep, even though a single sweep from 0.5Hz to 1MHz can be made. The low end 0.5Hz could have been made lower, but there was insignificant additional information to be identified for the standard Williamson circuit. The high end 1MHz was sufficient to observe gain reductions out to about -20dB (ie. sufficiently past the frequencies associated with gain margin and phase margin measurement). The scope used 10:1 probes (ie. 10MΩ loading) compensated for flat gain and phase response out to 1MHz, and the 12-bit measurement resolution allowed a 70dB dynamic range to cover the highest resonance peaking of nearly +50dB, down to below -20dB attenuation out towards 1MHz.

The 4224A provides up to 1.4Vrms sinewave output, but testing used an attenuated input signal to the amplifier such that typical amplifier mid-band output signal level was about 0.5Vrms. The input signal level is not the signal level at the feedback summing junction, so formal gain and phase margins cannot be directly taken from the open-loop plots (when feedback is not connected).

3. Open circuit secondary winding tests

Primary inductance measurement

In the 1940's Williamson was likely aware that the AVO Model 7 [6] would be a common meter used to confirm the primary PP inductance of an output transformer. To meet the target of 100H, winding current needed to be less than 150uA using a 5VAC 50Hz supply across the winding, which was a very low reading (1-2% of FS end) on the 10mA FS setting (5mA FS with the x2 switch) of the Model 7 dial. So 5V would have been the lowest practical excitation level appropriate to indicating the primary inductance.

Inductance falls below 5V although the reduction is somewhat minor, as indicated by Figure 5 showing measurement results for the Partridge WWFB based on mains frequency (using the meter setup) where 132H was measured at 5Vac 50Hz. Inductance increases with excitation voltage, being 3x at 115V, but would level out as core saturation is experienced. With any DC bias of the core, incremental inductance falls for unbalanced idle half-winding currents due to DC offset [8].

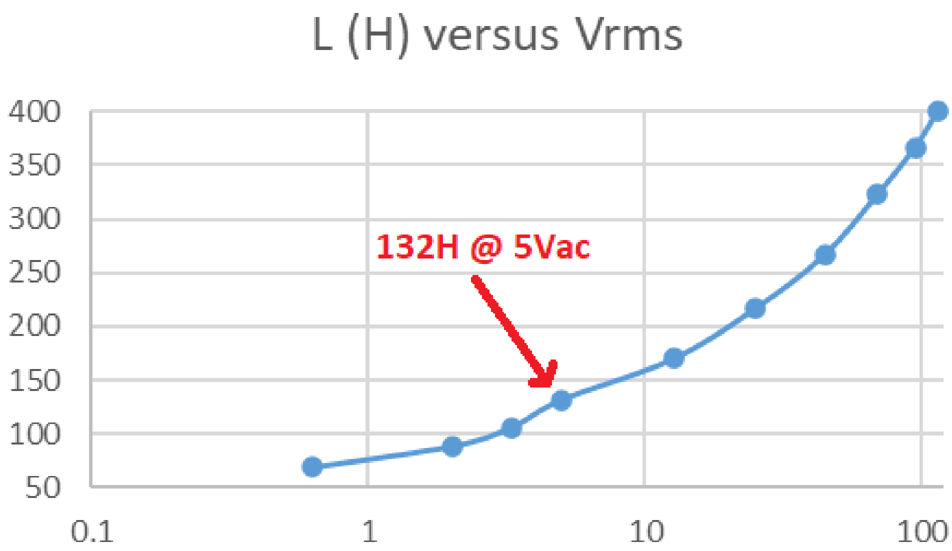


Figure 5. Primary winding inductance versus excitation voltage for WWFB.

The primary winding impedance versus frequency plot in Figure 6 (using the soundcard/REW setup) shows the impedance magnitude as the red trace (with a log axis from 200Ω to 300kΩ) and the impedance phase as the green trace (with a linear axis from -100 to +100 deg).

The measured inductance as calculated by REW [7] is 88H at 50Hz for the Partridge WWFB for a 3.8Ω secondary configuration [5]. That measurement value at 50Hz aligns with the values measured in Figure 5, as the excitation voltage is lower at about 1.5Vrms.

The REW measured inductance increases as frequency reduces, being double at 5Hz, and similarly the measured inductance decreases as frequency increases towards the self-resonance with the shunt capacitance.

The measured inductance starts to fall off below about 1Hz as indicated more directly by the phase change below about 4Hz, however the excitation voltage is dropping off noticeably below about 5Hz (-10% at 3Hz).

The REW measured winding capacitance is about 615-630pF in the frequency range 3kHz to 22kHz, which is sufficiently after the main resonance at circa 850Hz and before the next resonance at circa 100kHz (as noted by the rising phase above about 65kHz.

The self-resonance at circa 850Hz equates to a capacitance of 630pF and 55H.

[6] AVO Model 7 introduced in 1936.

[7] REW calculates the series R + L model components (shown in the bottom left of the plot) from the impedance magnitude and phase at the marker frequency.

Figure 6. Primary winding impedance versus frequency at about 1.5Vrms for WWFB

For the measurement in Figure 6, an expanded plot showing just the phase with linear Y axis is given in Figure 7. The phase shift is within 20 deg of 90 deg from 3.6 Hz to 130 Hz, and at 10Hz the phase shift is 16.5 deg and meets Williamson's 20 deg shift requirement.

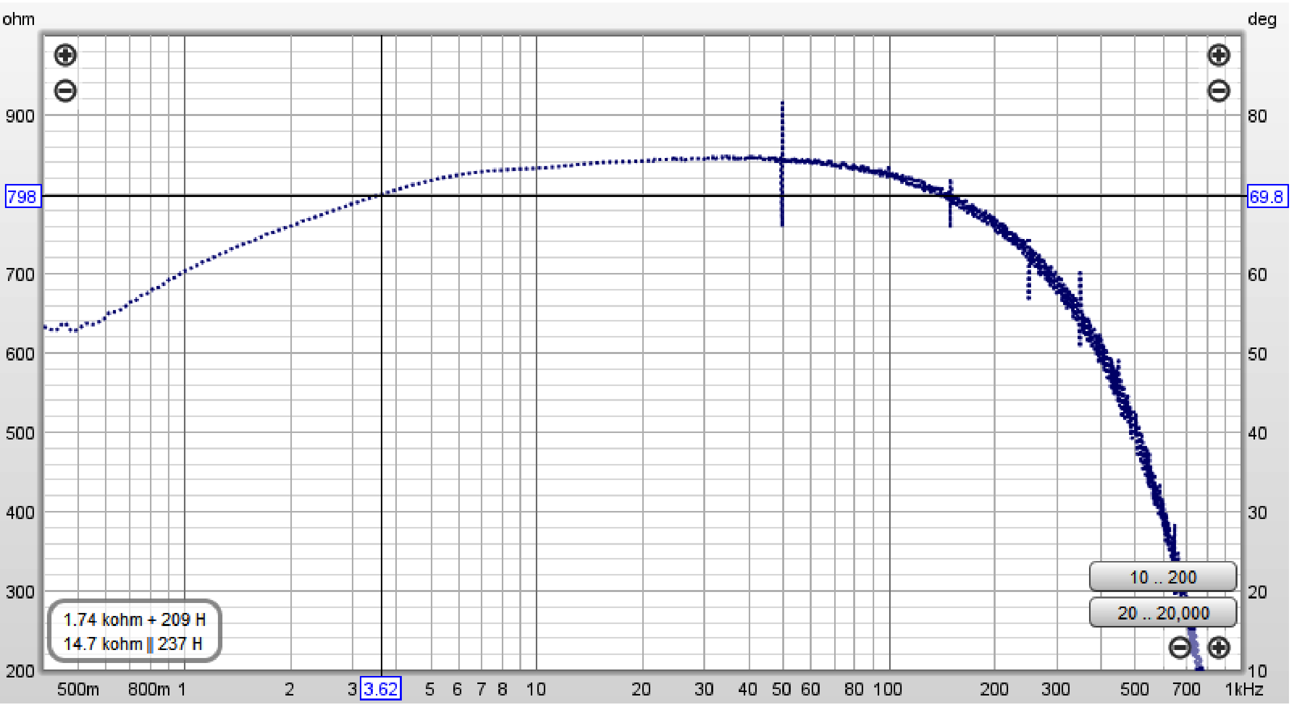


Figure 7. Primary winding impedance phase response for WWFB with 3.8Ω configuration

The WWFB has 8 secondary windings that are typically connected into parallel and series sections [5]. The response in Figure 7 is for 2 sections (3.8Ω configuration for WFFB/0.95). The impedance response is influenced by the interconnection arrangement of secondary windings, and notably for the 8.5Ω configuration the phase shift at low frequency is not as good, and the primary inductance is lower.

Figure 8. Primary winding impedance of WWFB with 8.5Ω secondary configuration.

For comparison, Figure 9 shows the measured inductance of 80H at 50Hz for the Ferguson OP25 (new) in a 4Ω secondary configuration. The main resonance is circa 900Hz and the next resonance is circa 95kHz.

Figure 10 shows the measured inductance of 73H at 50Hz for the Red Line AF8/2 in an 8Ω secondary configuration. Although the main resonance is at a similar 1kHz, the next resonance is at about 49kHz.

Figure 9. Primary winding impedance of Ferguson OP25 (new) with 4Ω secondary configuration.

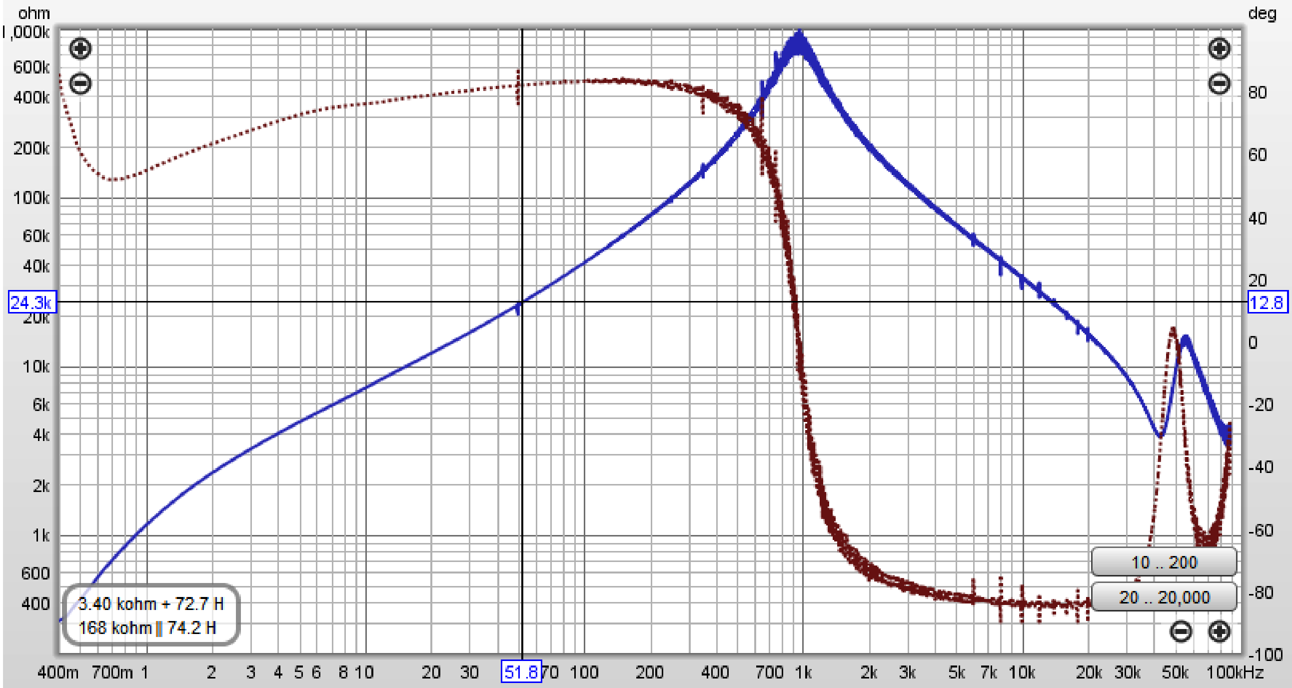


Figure 10. Primary winding impedance of Red Line AF8/2 with 8Ω secondary configuration

4. Short circuit secondary winding tests

Primary leakage inductance measurement

The primary winding impedance versus frequency plot in Figure 11 (using the soundcard/REW setup) shows a measured leakage inductance of 14mH at 1kHz for the Partridge WWFB in a 3.8Ω secondary configuration [5], where the impedance magnitude trace is green, and phase is purple.

The measured leakage inductance increases as frequency reduces, rising rapidly below about 50Hz where it was double. The measured leakage inductance was constant for frequency up to about 20kHz, above which the winding capacitance starts to become significant towards the self-resonance with the shunt capacitance.

The winding capacitance is about 630pF at 65kHz just before the minimum phase shift. The main resonance at circa 48kHz equates to a capacitance of 630pF and 17mH. The next resonance is at about 100kHz, given the measurement sweep stops at 90kHz.

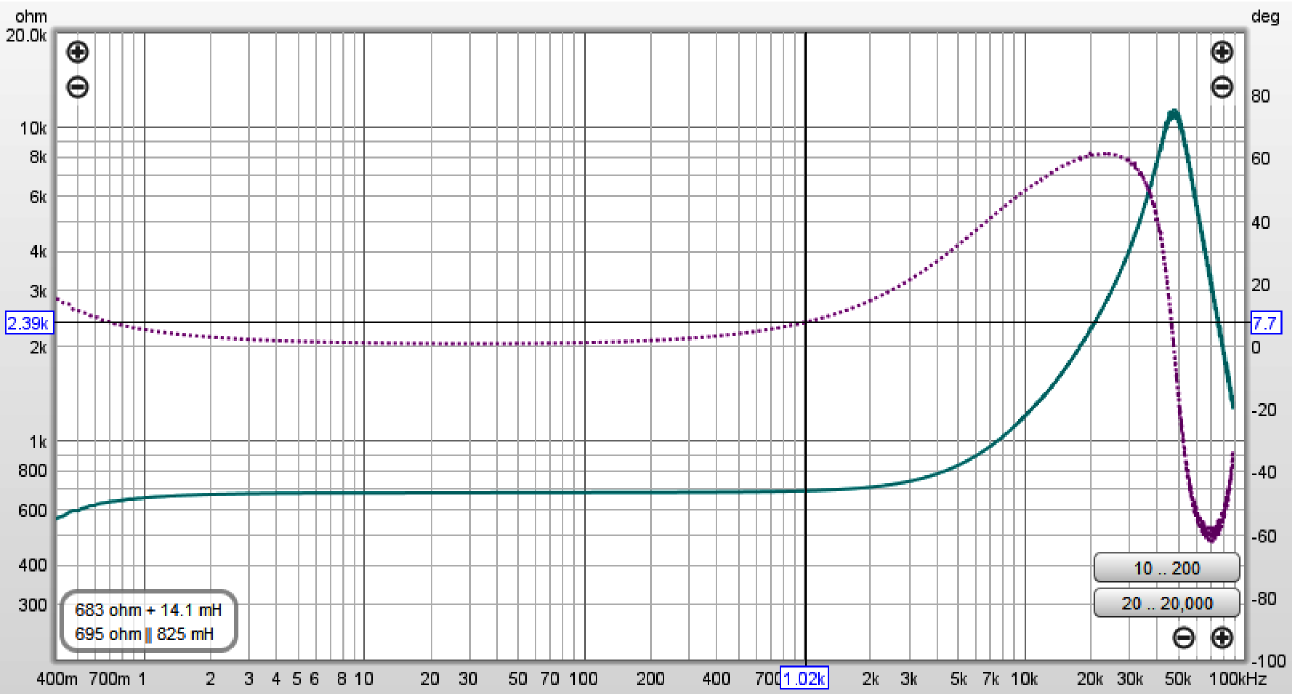


Figure 11. Shorted secondary impedance for WWFB in 3.8Ω configuration

For comparison, Figure 12 shows the measured leakage inductance of 16mH at 1kHz for the Ferguson OP25 (new) in a 4Ω secondary configuration, where the impedance magnitude trace is blue, and phase is purple. A minor resonance is just observable at 32kHz although the main resonance is out at 70kHz.

Figure 13 shows the measured leakage inductance of 38mH at 1kHz for the Red Line AF8/2 in an 8Ω secondary configuration, where the impedance magnitude trace is green, and phase is blue. The main resonance is at a significantly lower frequency of 26kHz, and the next resonance is at about 75kHz.

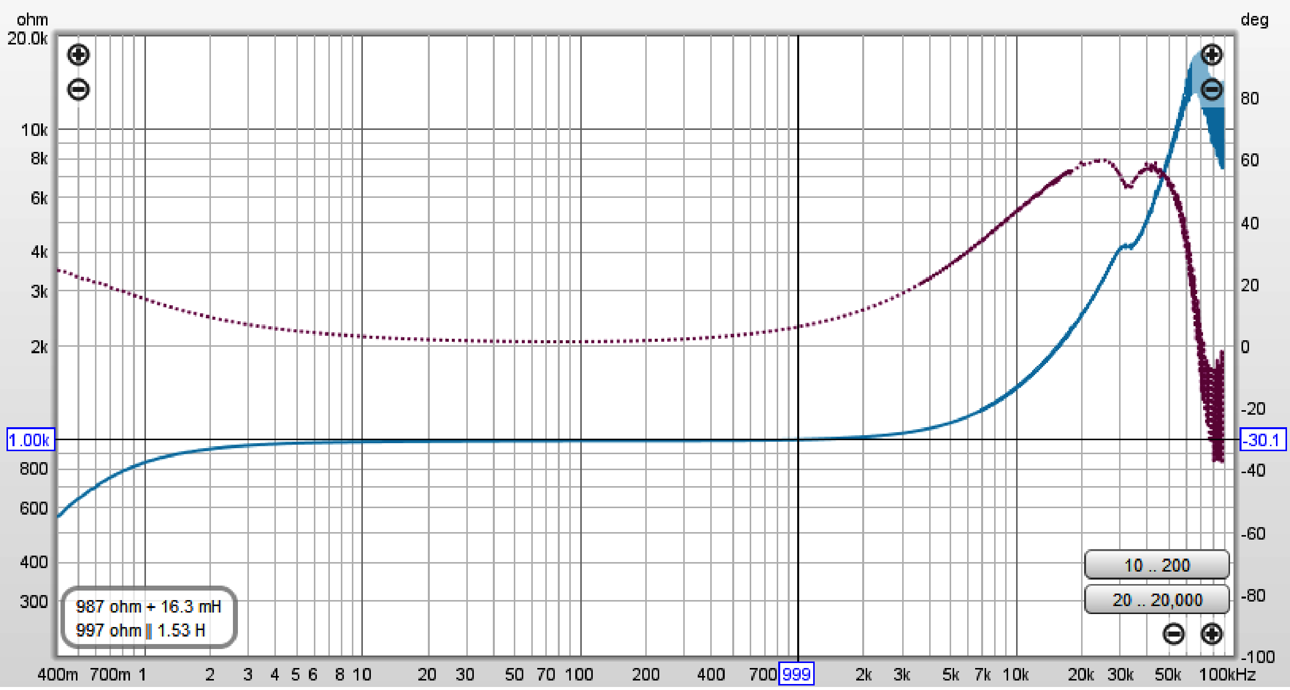


Figure 12. Shorted secondary impedance for Ferguson OP25 (new) in 4Ω configuration

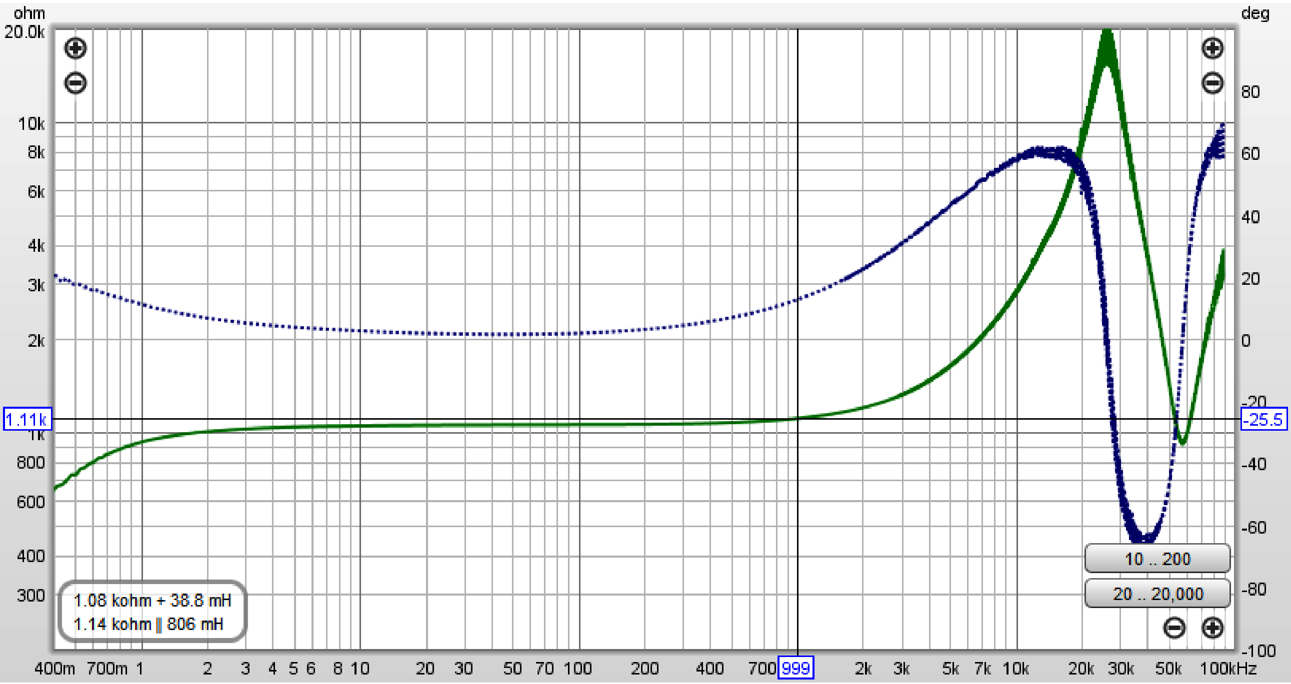


Figure 13. Shorted secondary impedance for Red Line AF8/2 in 8R configuration

Shorted primary to shorted secondary capacitance

This test indicates the total effective capacitance between all sections of the primary winding and the secondary winding by shorting the primary winding and shorting the secondary winding and measuring the impedance between the primary and secondary sides.

The capacitance value is typically at a minimum around 10kHz, given that phase approaches 90deg at lower frequencies, and then a higher frequency resonance starts to force phase shift back through zero at higher frequencies. The high frequency resonance at 41kHz for the OP25/15 (old) indicates the 4.6nF resonates with a 3.6mH leakage inductance. For the WWFB resonance of about 100kHz, the 8nF resonates with a 0.3mH leakage inductance. For the AF8/2 resonance at 70kHz, the 6.8nF resonates with a 0.75mH leakage inductance. Although the lower effective leakage inductance value would indicate a better coupling between primary and secondary windings, the AF8/2 has a noticeably high primary leakage inductance (as measured in the first part of this section).

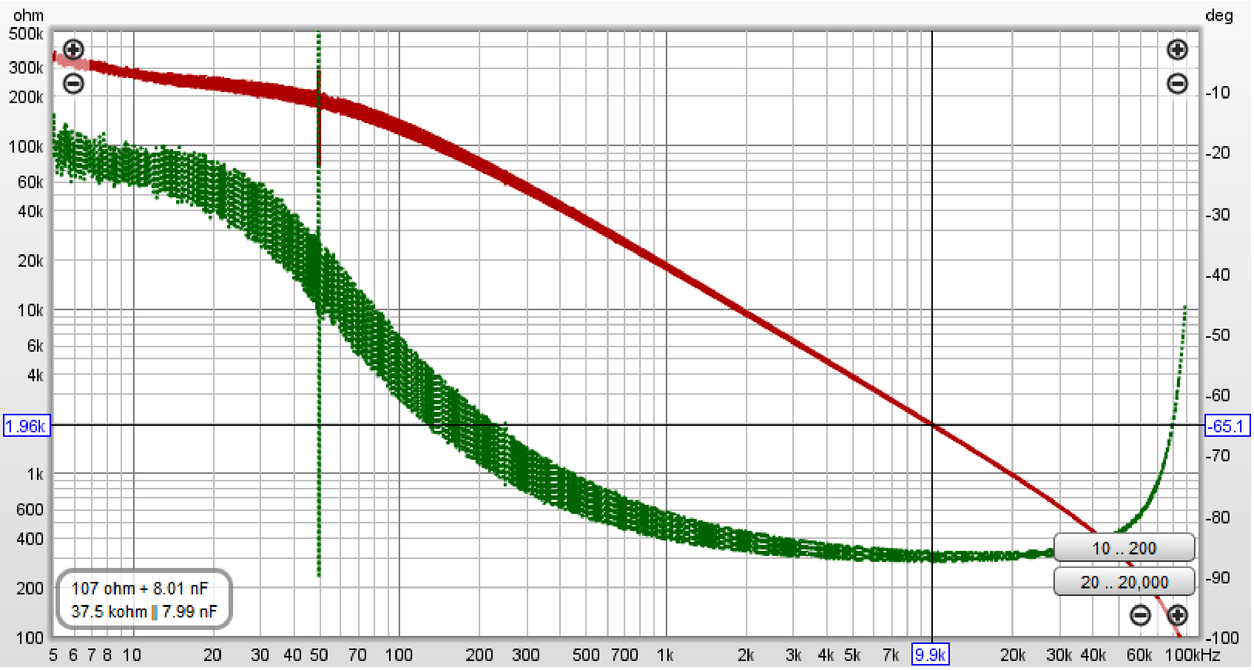


Figure 14. Shorted primary to shorted secondary impedance for WWFB

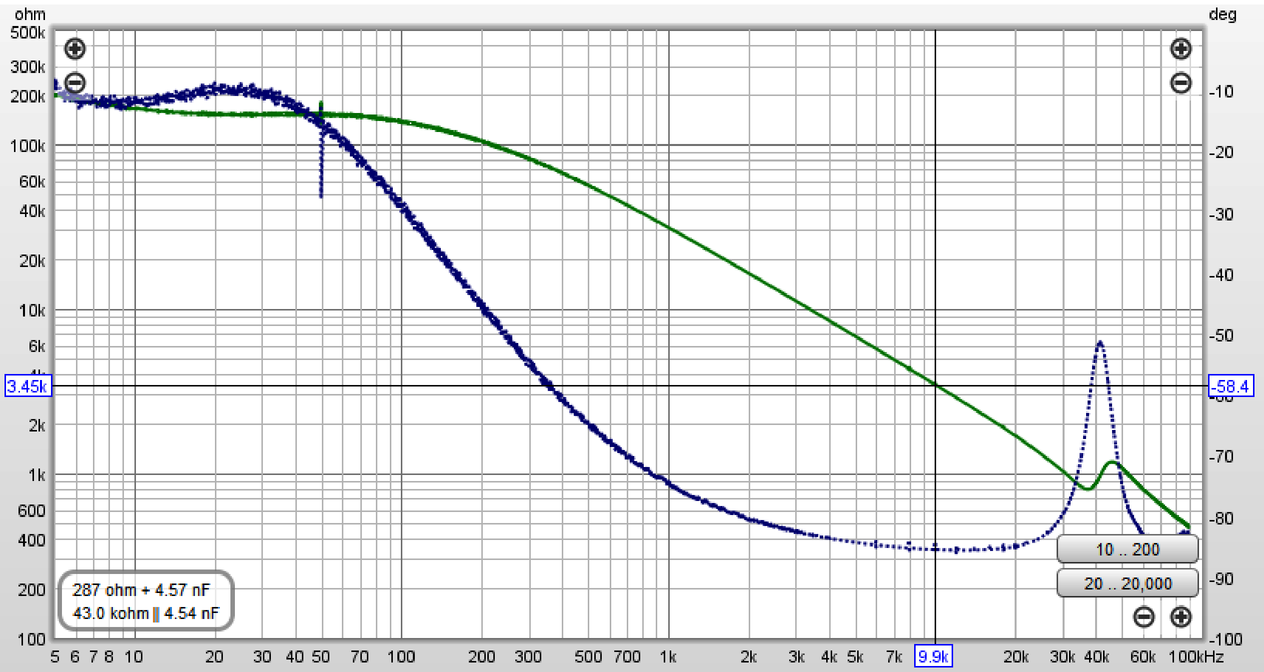


Figure 15. Shorted primary to shorted secondary impedance for Ferguson OP25/15 (old)

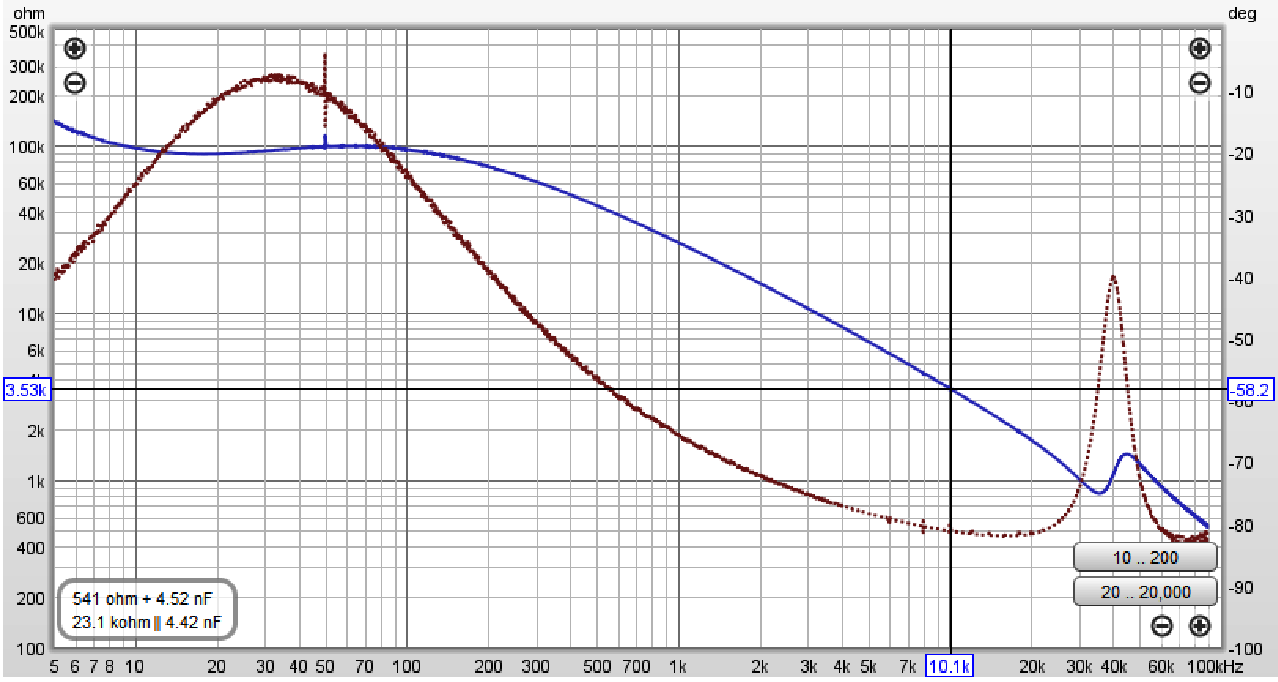


Figure 16. Shorted primary to shorted secondary impedance for Ferguson OP25/15 (new)

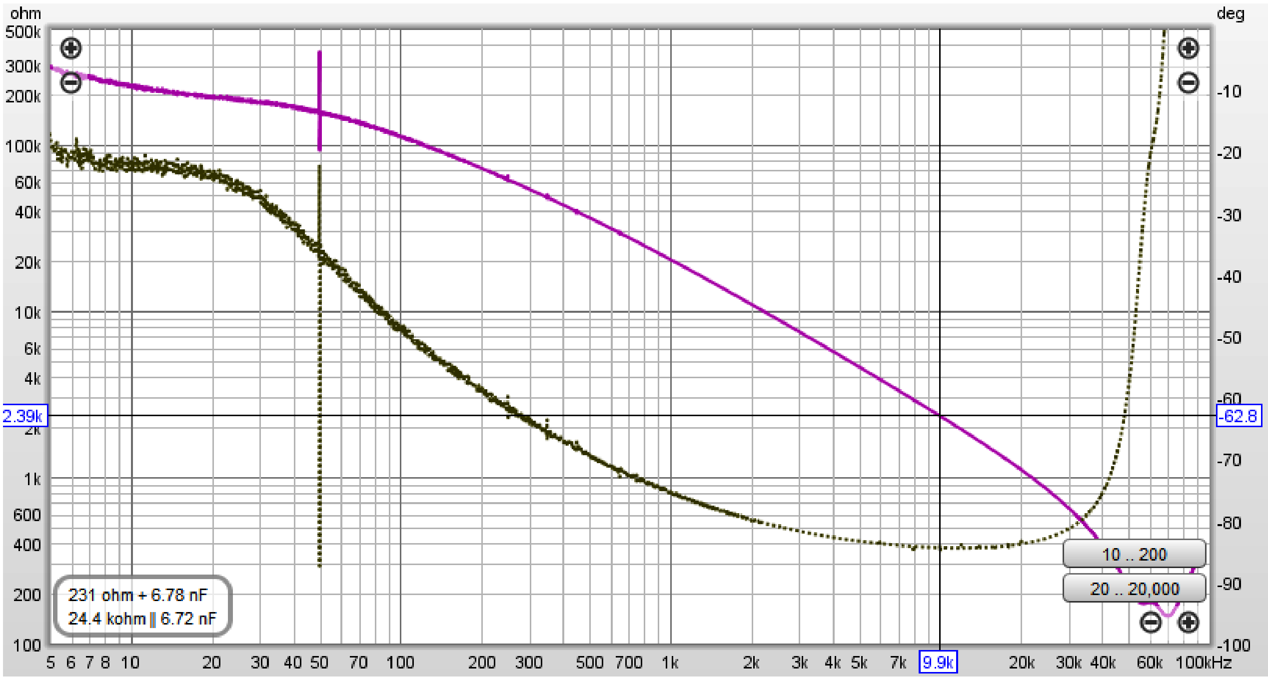


Figure 17. Shorted primary to shorted secondary impedance for Red Line AF8/2

5. Frequency response tests – 2Hz to 95kHz

The frequency response of each transformer was measured from 2Hz to 95kHz with the secondary winding configured and loaded for its rated matching resistance to present a 10k PP primary impedance and with a 4.1Vrms driver voltage. The primary winding CT was connected to the secondary ground terminal, and the primary PP terminals driven by a balanced line driver. This same measurement also acquires data to present the harmonic distortion contributed by the OPT (as the signal loop has negligible distortion and noise relative to the harmonics generated by the OPT).

Figure 18 shows the amplitude (red) and phase (green) frequency response traces for the Partridge WWFB with a 3.8Ω secondary winding configuration and 3.9Ω resistive load. Figure 19 shows the WWFB frequency response with an 8.5Ω secondary winding configuration and 8.5Ω resistive load. The scales span 6dB and +/-30deg.

Below each figure are the related manual measurements of gain and phase at the low and high frequency extremes. A higher driver voltage of 10Vrms was used for the manual measurements. Very low frequency waveforms are quite distorted.

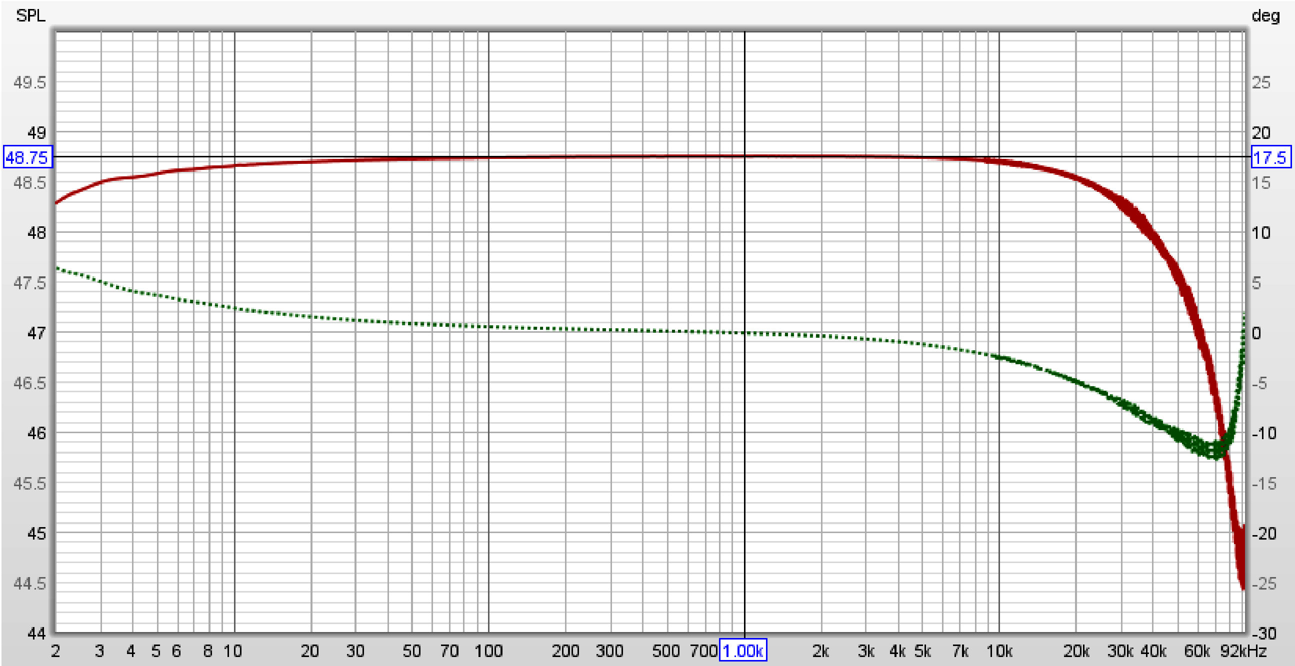


Figure 18. Frequency response of WWFB with 3.8Ω secondary configuration and load

10Vrms source; 10Hz -0.2dB; 2Hz -1.3dB; 1Hz -2dB; 0.5Hz -3.2dB; 0.2Hz -5.6dB;

-10dB dip 93kHz; -5dB pk 130kHz; -12dB dip 220kHz; -9dB flat 280-380kHz; -16dB pk 430kHz;

-26dB flat 610-760kHz

Phase: <90deg 0.1Hz; 90deg 610kHz; .

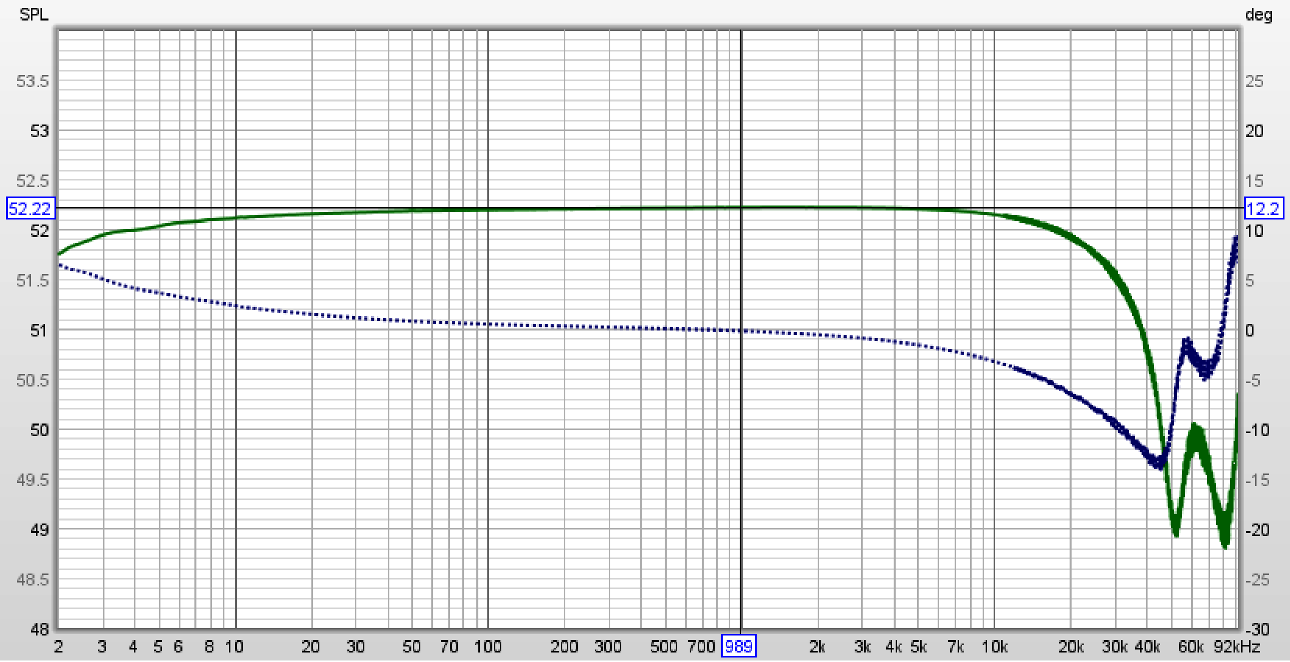


Figure 19. Frequency response of WWFB with 8.5Ω secondary configuration and load

10Vrms source; 10Hz -0.4dB; 2Hz -1.5dB; 1Hz -2.1dB; 0.5Hz -3.2dB; 0.2Hz -5.7dB;

-0.8dB dip 25kHz; -0.4dB pk 30kHz; -12dB dip 85kHz; -3dB pk 133kHz; -8-10dB flat 210-480kHz; -25dB flat 600-750kHz.

Phase: <90deg 0.1Hz; dip then 0deg 100kHz; 90deg 600kHz.

The harmonic distortion generated by the WWFB at the 4Vrms driver level is shown in Figure 20. The 3rd harmonic level was 0.12% at 20Hz, reducing to 0.03% at 100Hz. The higher order odd harmonics are consistently lower, and the even order harmonics are negligible. See Table 2 summary results for all OPTs.

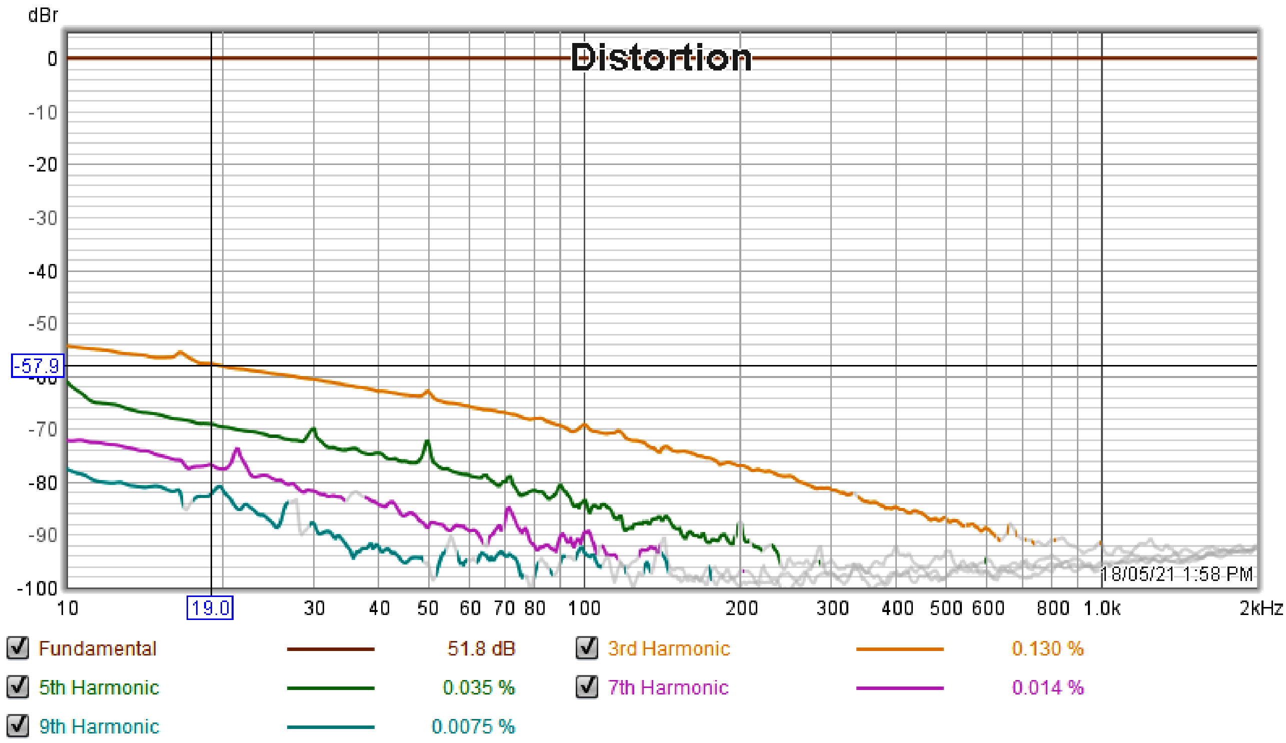


Figure 20. Harmonic distortion of WWFB with 8.5Ω secondary configuration and load

Figure 21 shows the amplitude (blue) and phase (brown) frequency response traces for the Ferguson OP25/15 (old) with a 3.75Ω secondary winding configuration and 3.75Ω resistive load. Figure 22 shows the OP25/15 (old) frequency response with a 15Ω secondary winding configuration and 15Ω resistive load. The scales span 10dB and +/-50deg.

Below each figure are the related manual measurements of gain and phase at the low and high frequency extremes. A higher excitation voltage was used for the manual measurements. Very low frequency waveforms are quite distorted.

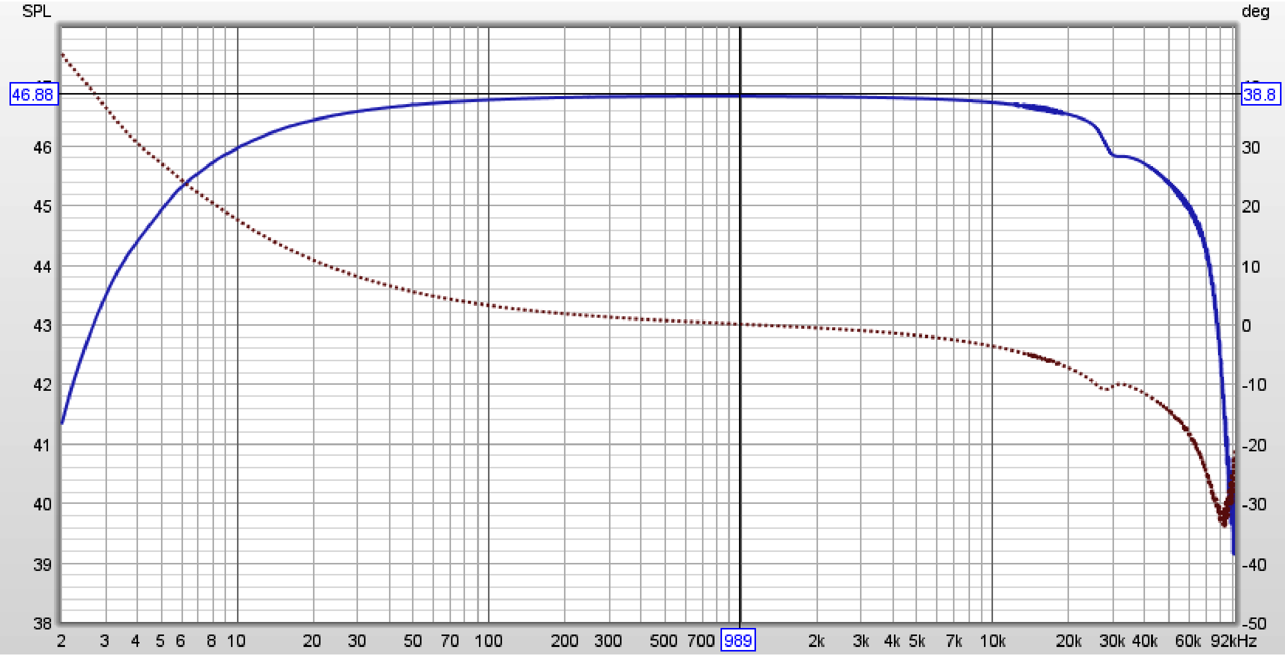


Figure 21. Frequency response of OP25 (old) with 3.75Ω secondary configuration and load

10Vrms source; 10Hz -0.4dB; 2Hz -2.1dB; 1Hz -3.3dB; 0.5Hz -4.7dB; 0.2Hz -8dB;

-10dB flat 105-120kHz; -14dB dip 160kHz; -12.5dB pk 210kHz; -17dB dip 350kHz;

-16dB pk 430kHz; -26dB flat 610-760kHz

Phase: ~90deg 0.1Hz; 90deg 81kHz; 90deg 120kHz; 90 deg 160kHz; 180deg 450kHz; 270deg 650kHz.

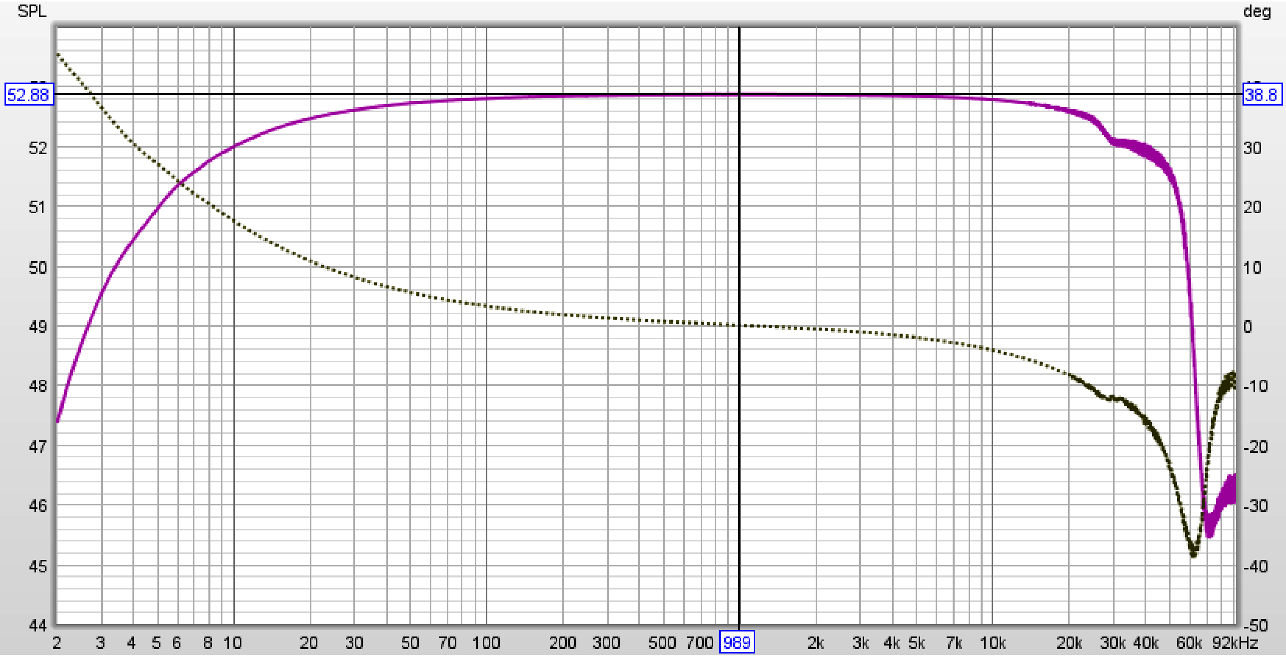


Figure 22. Frequency response of OP25 (old) with 15Ω secondary configuration and load

10Vrms source; 10Hz -0.4dB; 2Hz -2.0dB; 1Hz -3.2dB; 0.5Hz -4.6dB; 0.2Hz -8dB;

> -6dB flat 71-84kHz; -14.7dB dip 165kHz; -13dB pk 210kHz; -18dB dip 365kHz; -16.6dB pk 440kHz; -25dB flat 600-750kHz.

Phase: <90deg 0.1Hz; 90deg 105kHz; 90deg 160kHz; 90deg 205kHz; 180deg 440kHz; 90 deg 640kHz

The same nominal high frequency amplitude response was measured with the OPT in a Williamson amp without a first stage step network, however the phase response showed only a negligible dip below 0 deg and rising to 40deg at 90kHz.

Figure 23 shows the amplitude (green) and phase (purple) frequency response traces for the Ferguson OP25/15 (new) with a 3.75Ω secondary winding configuration and 3.75Ω resistive load. Figure 24 shows the OP25/15 (new) frequency response with a 15Ω secondary winding configuration and 15Ω resistive load. The scales span 10dB and +/-50deg.

Below each figure are the related manual measurements of gain and phase at the low and high frequency extremes. A higher excitation voltage was used for the manual measurements. Very low frequency waveforms are quite distorted.

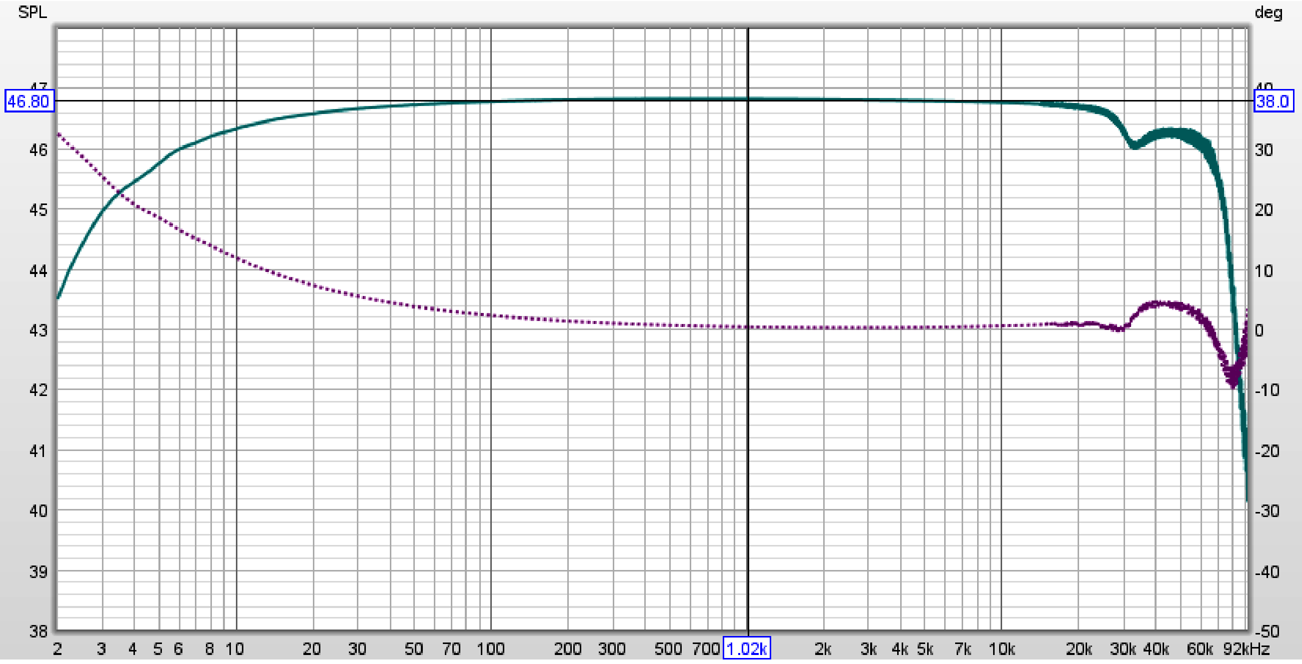


Figure 23. Frequency response of OP25 (new) with 3.75Ω secondary configuration and load

10Vrms source; 10Hz -0.4dB; 2Hz -1.9dB; 1Hz -2.7dB; 0.5Hz -4dB; 0.2Hz -7.1dB;

-7dB flat 105-175kHz; flat -11dB 450k-550kHz.

Phase: <90deg 0.1Hz; dip ~100kHz; pk 110kHz; 90deg 210kHz; 180deg 750kHz.

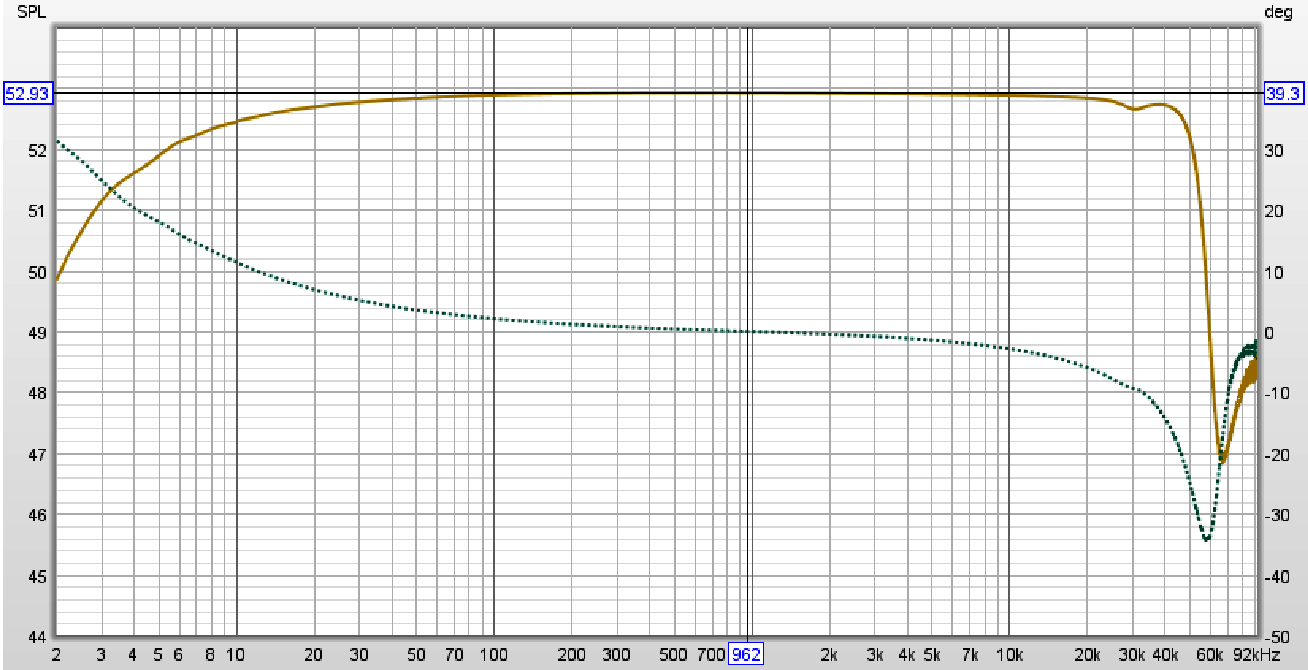


Figure 24. Frequency response of OP25 (new) with 15Ω secondary configuration and load

10Vrms source; 10Hz -0.3dB; 2Hz -1.7dB; 1Hz -2.6dB; 0.5Hz -4dB; 0.2Hz -7.5dB;

-6dB dip 68kHz; -4.4dB pk 108kHz; -20dB flat 450-540kHz; -34dB flat 850k-1MHz.

Phase: <90deg 0.1Hz; dip ~55kHz; pk 75kHz; 90deg 215kHz; 180deg 600kHz.

The same nominal high frequency amplitude and phase response was measured with the OPT in a Williamson amp without a first stage step network.

Figure 25 shows the amplitude (green) and phase (purple) frequency response traces for the Red Line AF8/2 with a 2Ω secondary winding configuration and 1.95Ω resistive load. Figure 26 shows the AF8/2 frequency response with an 8Ω secondary winding configuration and 7.8Ω resistive load. The scales span 10dB and +/-50deg.

Below each figure are the related manual measurements of gain and phase at the low and high frequency extremes. A higher excitation voltage was used for the manual measurements. Very low frequency waveforms are quite distorted.

The 8Ω secondary winding can be connected in two configurations. The first configuration bridges the two middle terminals and the load connects to the two outer terminals. The second configuration bridges the two outer terminals and the load connects to the two middle terminals. The response of both configurations differ from above a few kHz mainly in phase shift, with little difference in magnitude, with the second configuration showing less phase shift before resonance

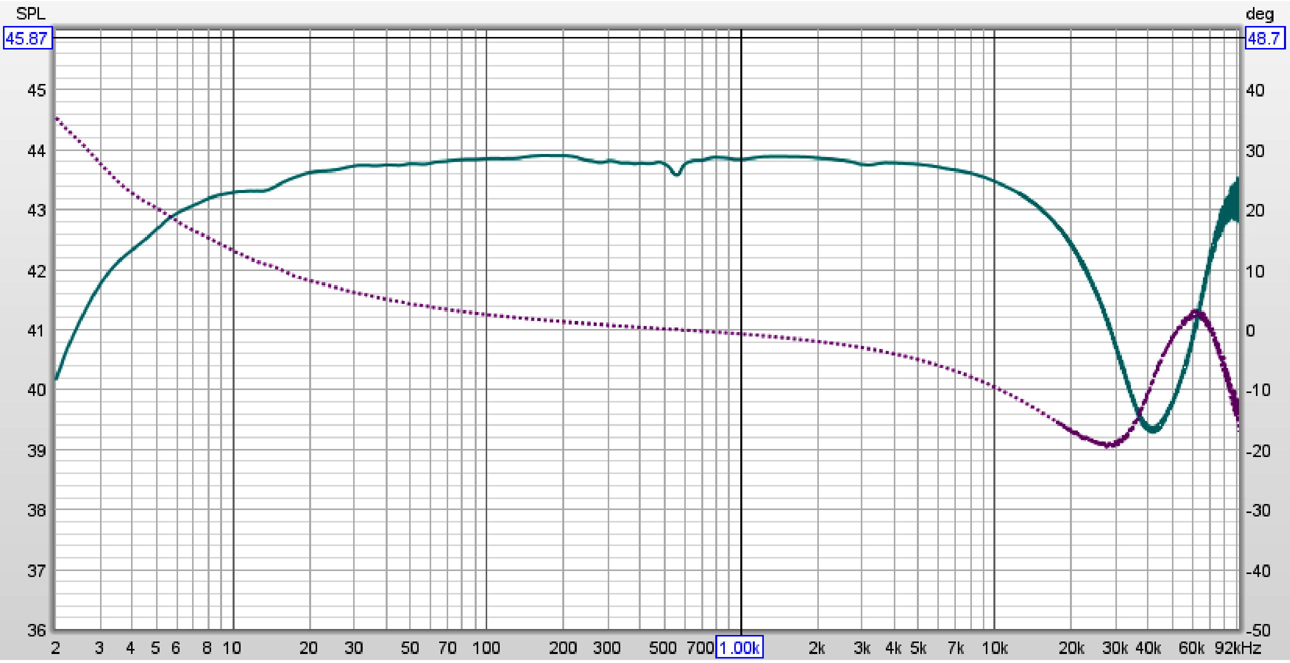


Figure 25. Frequency response of AF8/2 with 2Ω secondary configuration and load

10Vrms source; 10Hz -0.6dB; 2Hz -2.3dB; 1Hz -3.5dB; 0.5Hz -5dB; 0.2Hz -8dB;

-5dB dip 24kHz; -1.2dB pk 33kHz; -6dB dip 55kHz; -0.6dB flat 85k-105kHz; -11dB flat 180-250kHz

Phase: ~90deg 0.1Hz; dip then ~0deg 37kHz; dip then ~0deg 61kHz; 90deg 150kHz; rise then 90deg 390kHz.

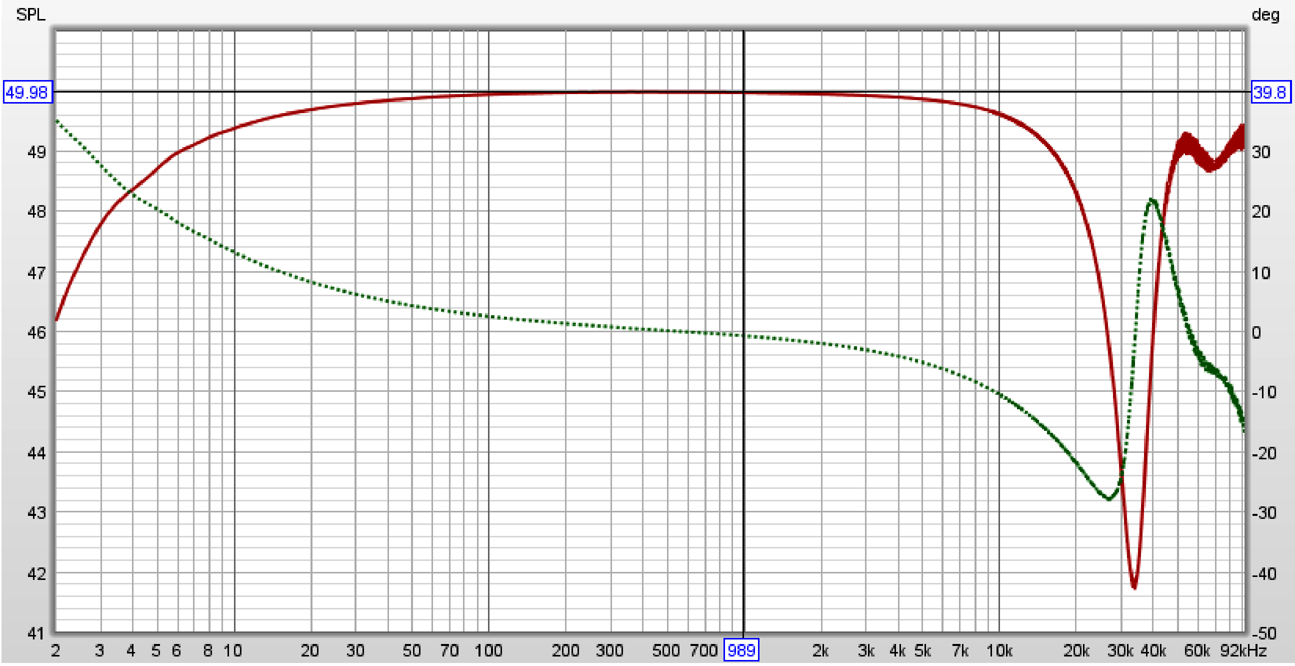


Figure 26. Frequency response of AF8/2 with 8Ω secondary configuration and load

Configuration with two middle terminals bridged and the load connected to the two outer terminals.

10Vrms source; 10Hz -0.3dB; 2Hz -2.1dB; 1Hz -3.5dB; 0.5Hz -5dB; 0.2Hz -8dB;

-9dB dip 36kHz; 0dB pk 68kHz; -1.6dB flat 100-120kHz; -10dB flat 175k-200kHz.

Phase: <90deg 0.1Hz; dip then 0dg 36kHz; pk then 0deg 54kHz; ~90deg 150-400kHz; 90deg 900kHz.

The same nominal high frequency amplitude response was measured with the OPT in a Williamson amp without a first stage step network, however the phase response showed a more negative phase shift although the characteristic was the same.

6. Amp frequency response tests – 0.5Hz to 1MHz

The frequency response of each transformer fitted in the same Williamson amplifier was measured from 0.5Hz to 1MHz with the secondary winding configured and loaded for its rated matching resistance to present a 10k PP primary impedance. The amplifier used for this testing was an AWA Williamson, without the first stage shelf network or any other form of stability compensation, and although some power supply differences exist compared with the original Williamson design, the amplifier is a close copy.

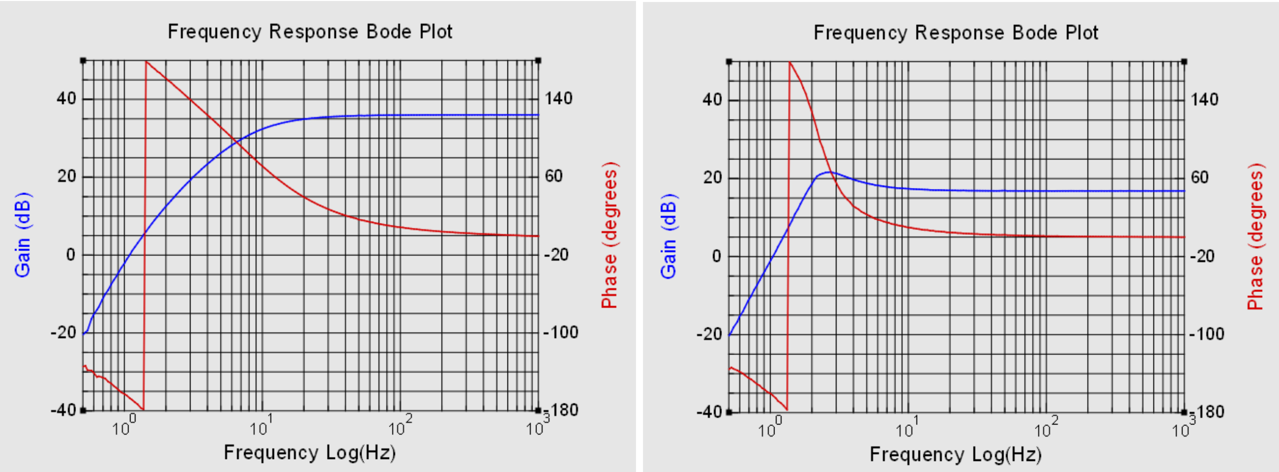


Figure 27. WWFB-0.95/8.5Ω low frequency gain and phase responses without and with feedback.

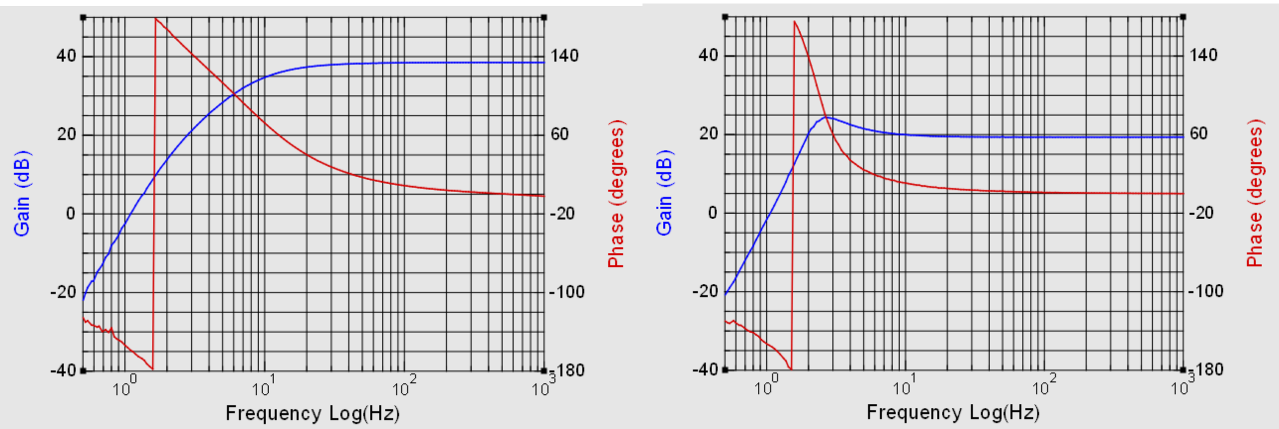


Figure 28. Old OP25/15Ω low frequency gain and phase responses without and with feedback.

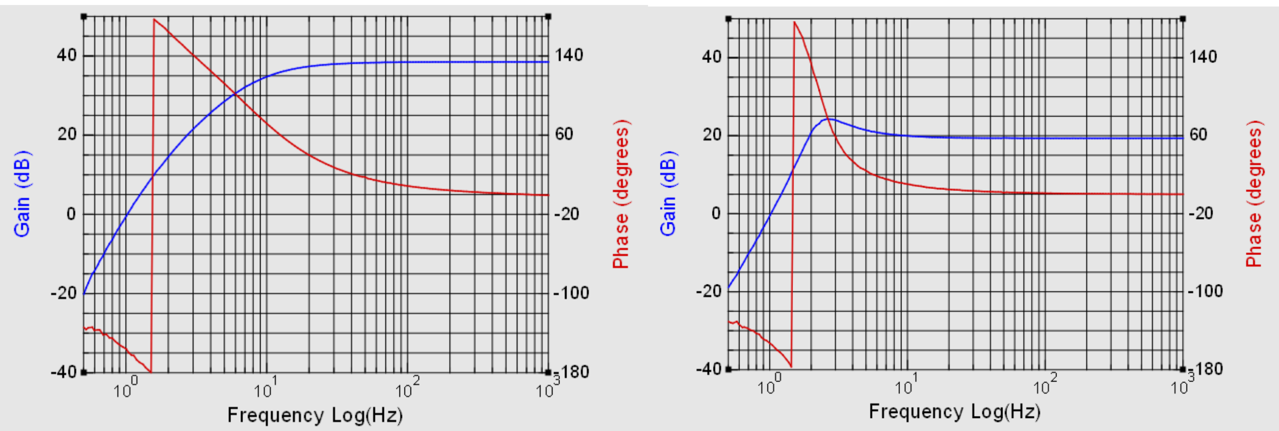


Figure 29. New OP25/15Ω low frequency gain and phase responses without and with feedback.

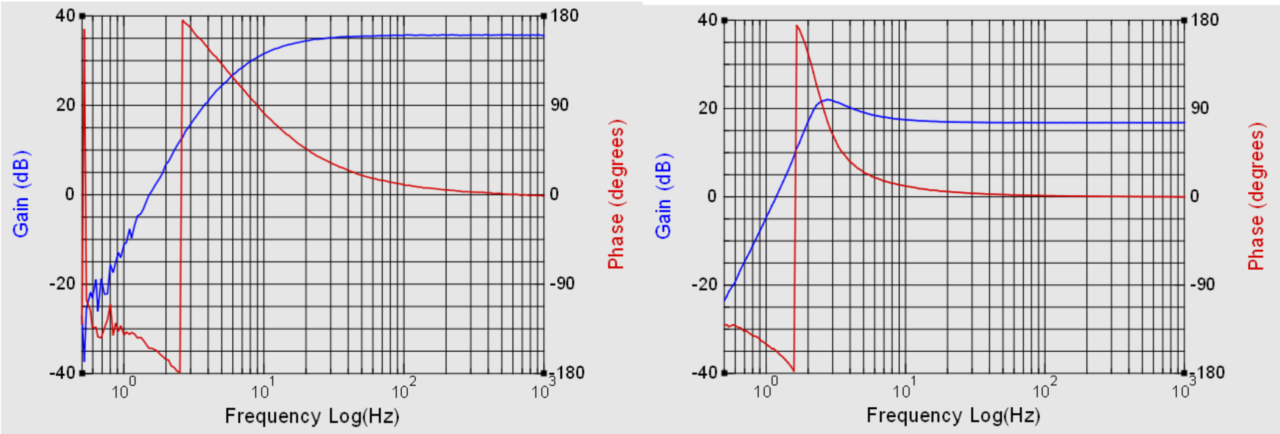


Figure 30. AF8/2 - 8Ω low frequency gain and phase responses without and with feedback.

Figure 27 to Figure 30 present low frequency gain and phase responses, both without and with feedback, from 0.5Hz to 1kHz. The mid-band gain with feedback is nominally 20dB lower than the response without feedback. With feedback, each transformer introduces a low frequency response peak of about +5dB at about 2.8Hz, and phase reaches 180 deg by between 1.3 and 1.8Hz. Increasing the input signal level causes the response peak's frequency to reduce, and vice-versa for a lower input signal level.

Figure 31 to Figure 34 present high frequency gain and phase responses, both with and without feedback, from 1kHz to 1MHz. The mid-band gain with feedback is nominally 20dB lower than the response without feedback.

The WWFB exhibits the highest frequency where phase shift reaches 180 deg, and with feedback the gain at that frequency is the lowest. All responses with feedback exhibit high frequency gain peaking, ranging from +10dB to +28dB above mid-band gain.

Without feedback, the noticeable dips in gain at 85kHz for the WWFB, and at 35kHz for the AF8/2, are significantly suppressed when feedback is applied.

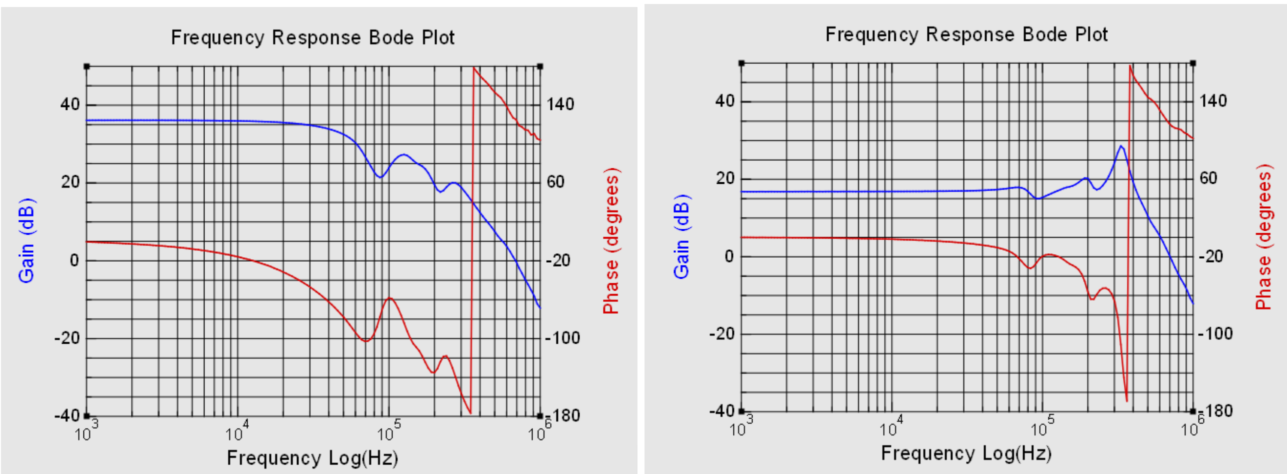


Figure 31. WWFB-0.95/8.5Ω high frequency gain and phase responses without and with feedback.

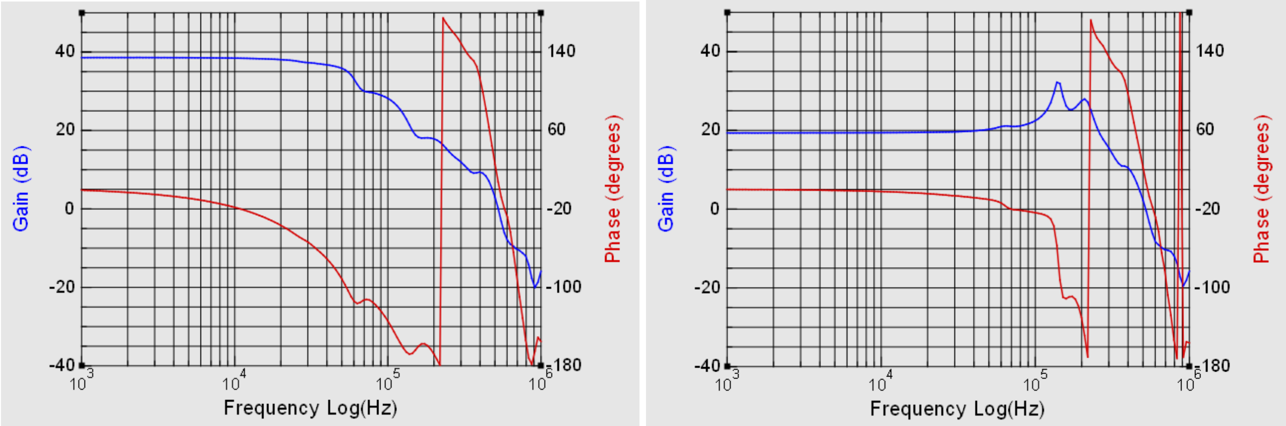


Figure 32. Old OP25/15Ω high frequency gain and phase responses without and with feedback.

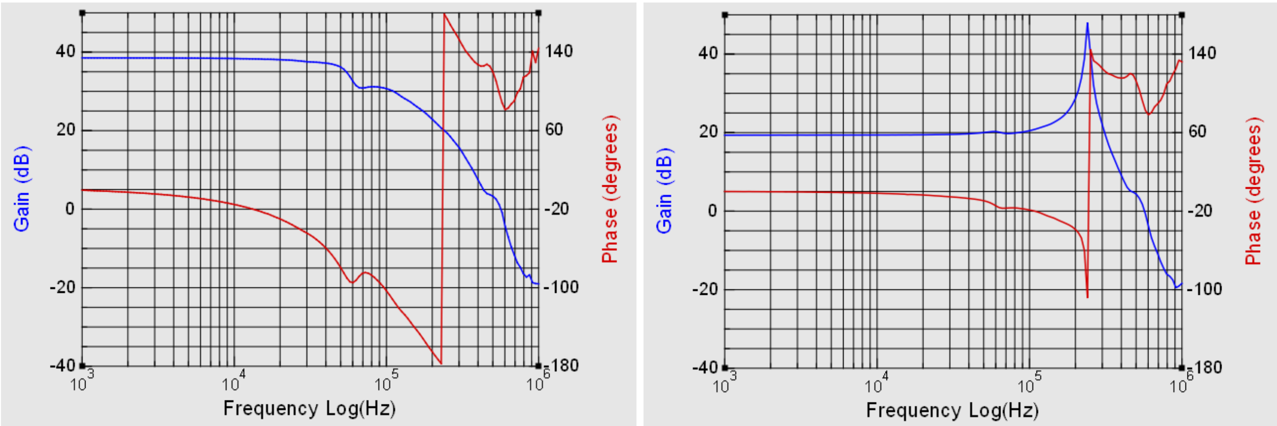


Figure 33. New OP25/15Ω high frequency gain and phase responses without and with feedback.

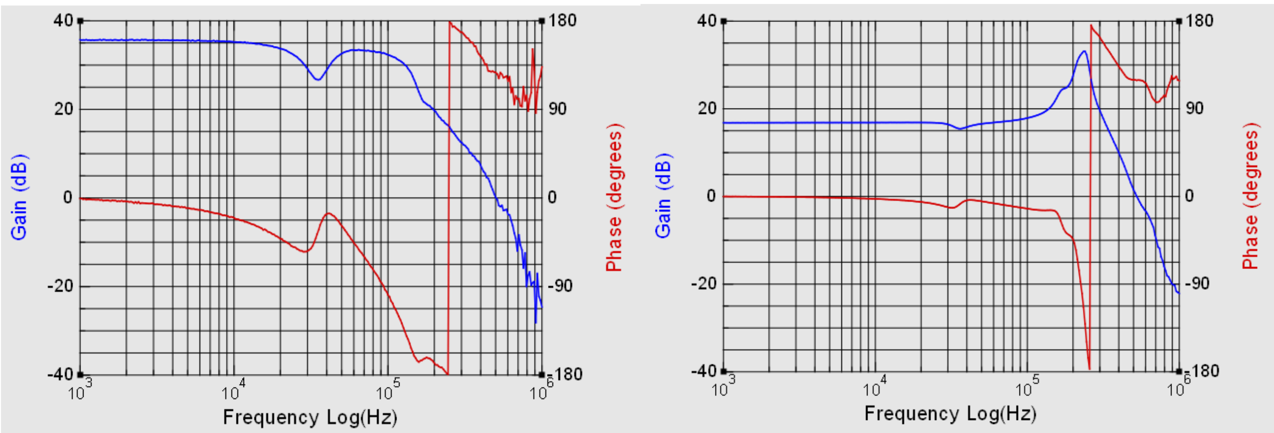


Figure 34. AF8/2 - 8Ω high frequency gain and phase responses without and with feedback.

Figure 35 to Figure 37 also present high frequency gain and phase responses for three additional vintage output transformers. The Partridge CFB stands out with a much higher frequency where phase shift reaches 180 deg, and with feedback the gain at that frequency is by far the lowest.

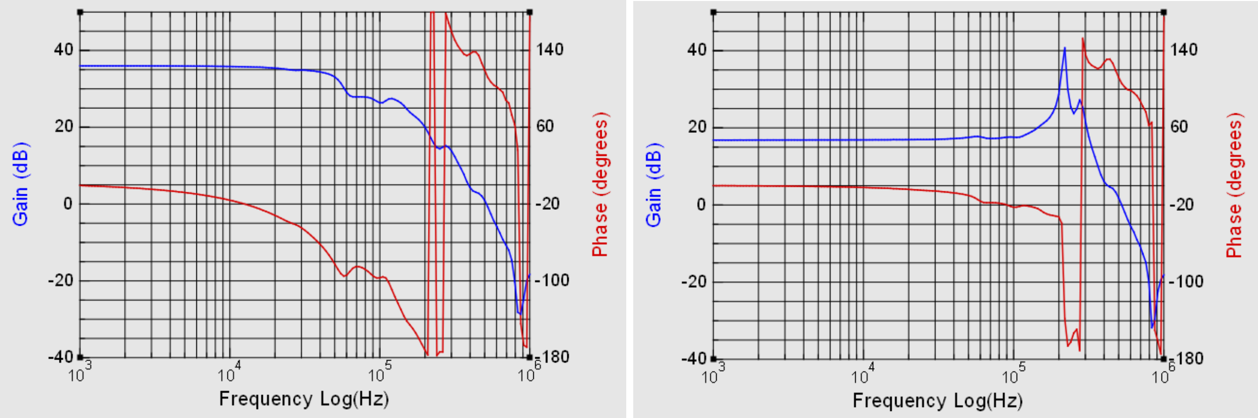


Figure 35. Old OP25/8Ω high frequency gain and phase responses without and with feedback.

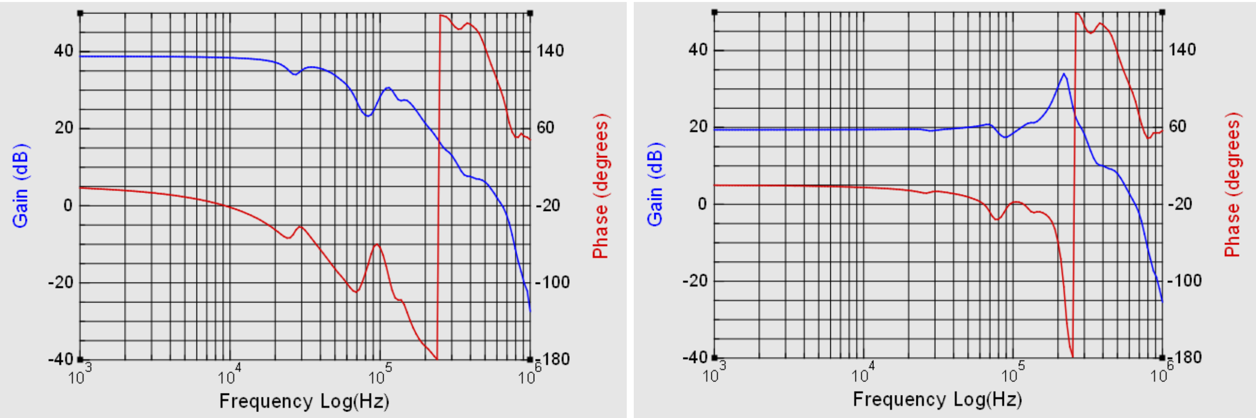


Figure 36. Trimax 996A - 16Ω high frequency gain and phase responses without and with feedback.

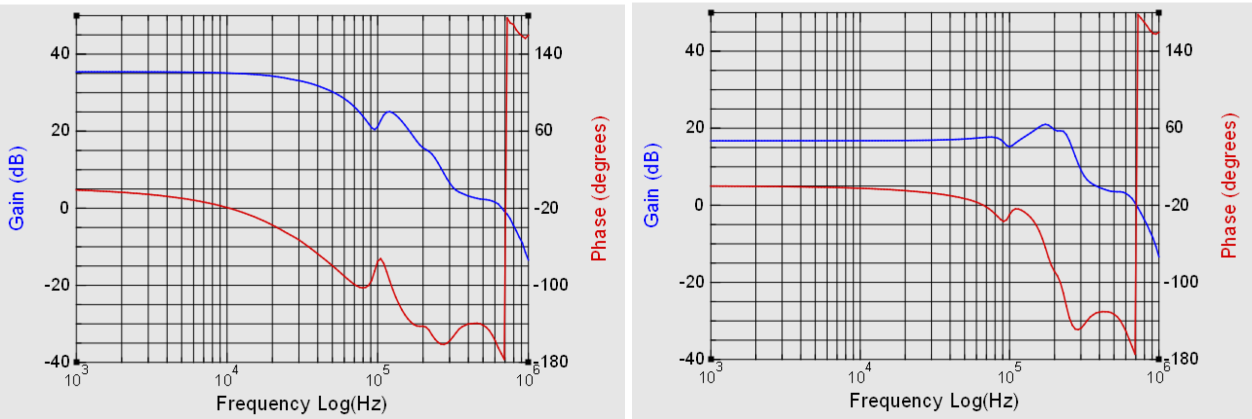


Figure 37. Partridge CFB-1.7 – 6.8Ω high frequency gain and phase responses without and with feedback.

7. HF stability with feedback and compensation

Ferguson OP25 (new)

This was assessed for stability with feedback in a Williamson amplifier, initially using squarewave response as an indicator of resonances well above 100kHz, and then using gain-phase measurements to 1MHz.

The AWA Williamson test amp did not include a first stage step network, and the output transformer secondary was configured for a standard 15Ω load, and a 15Ω load was applied. The squarewave supply was from a Wavetek 154 with a suitably fast rise time, and the reported waveforms are from a TEK TPS2012 with a suitably wide bandwidth.

Without global feedback the squarewave response at 1W shows a risetime to 100% of about 15us with minor ringing of ~17us period (60kHz). The resonance frequency is consistent with the first major resonance shown in Figure 24 and Figure 33. The same behaviour was observed at a 5W level.

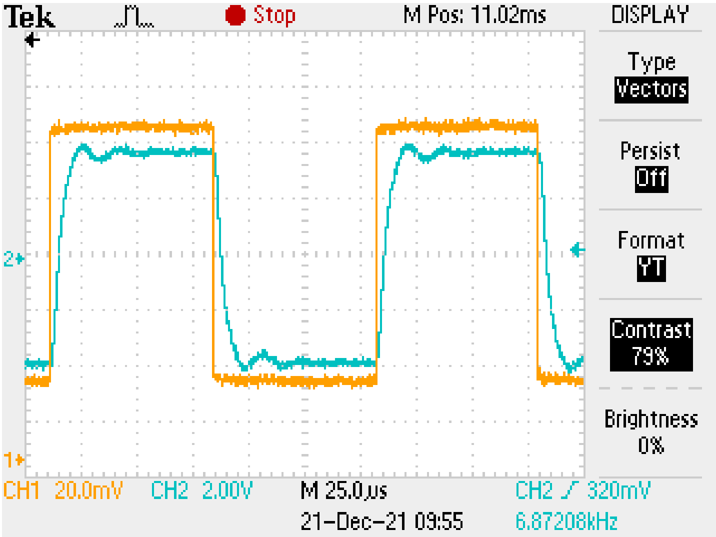


Figure 38. Squarewave response with no feedback.

With global feedback applied using R25=5k6, and output level reduced by 18dB, the squarewave response shows a strong but damped resonance at ~200kHz.

Feedback level is comparable to -20dB as R4 was not bypassed for the initial level measurement.

X-Y scope assessment of phase shift indicated a 90deg shift at 170kHz, and a 180deg shift at 200kHz, with 270deg shift out at 600kHz.

The gain-phase plot in Figure 33 was made with R25=4k7 but shows the strong resonance just above 200kHz.

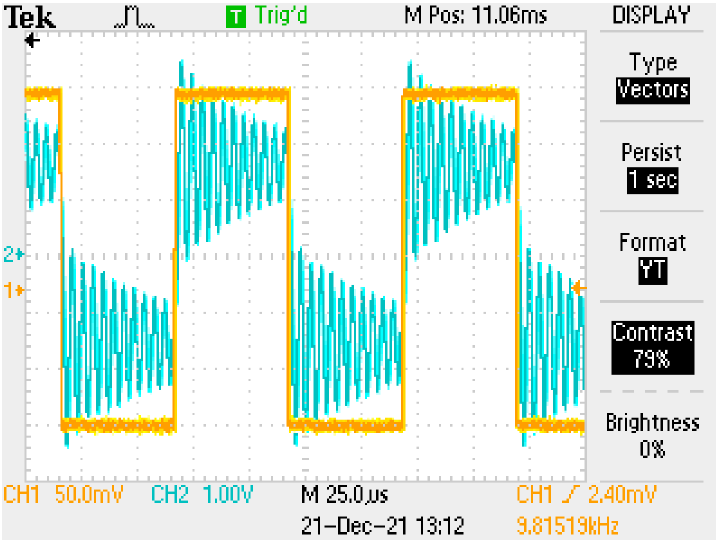


Figure 39. Squarewave response with feedback.

The resonance was able to be suppressed using a compensation capacitor across the feedback resistor R25. The plot on the right is for an added 150pF (190kHz corner), showing a small level of damped resonance of about 280kHz.

A lower value of compensation capacitance gave increased ringing, and a higher value caused the initial rising edge to peak at a lower level.

A gain-phase plot for nominally this compensation is shown in Figure 41.

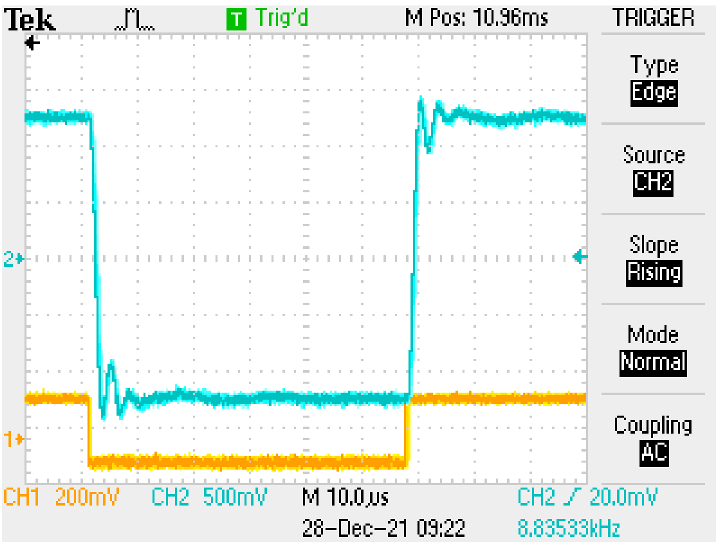


Figure 40. Squarewave response with 150pF compensation.

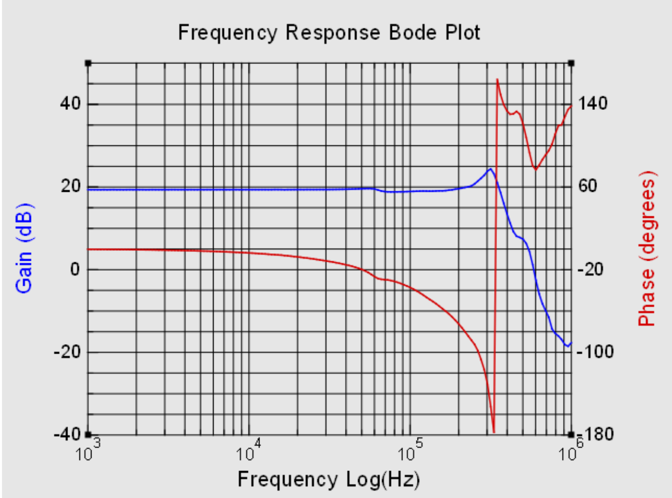


Figure 41. Gain-phase plot with 210pF compensation across 4k7 feedback.

However, for no-load or capacitance-only load conditions the amplifier is unstable even with a compensation capacitor, or the standard shelf filter in the first stage, so additional compensation is needed.

The resonance was also able to be suppressed using a series inductor in the load ground path [11]. The plot on the right is for L1=2.0uH [8], and shows a small level of damped resonance of about 280kHz.

A lower value of L1 inductance gave increased ringing, and a higher value caused the initial rising edge to peak at a lower level.

A gain-phase plot for nominally this compensation is shown in Figure 43.

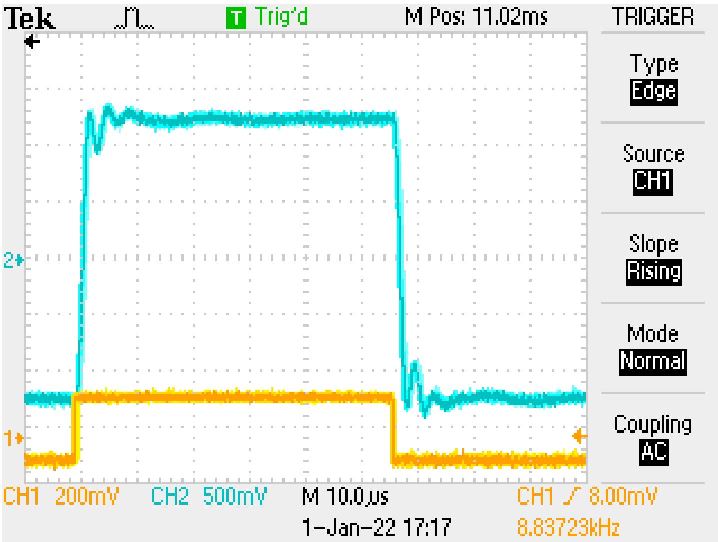


Figure 42. Squarewave response with series 2.0uH.

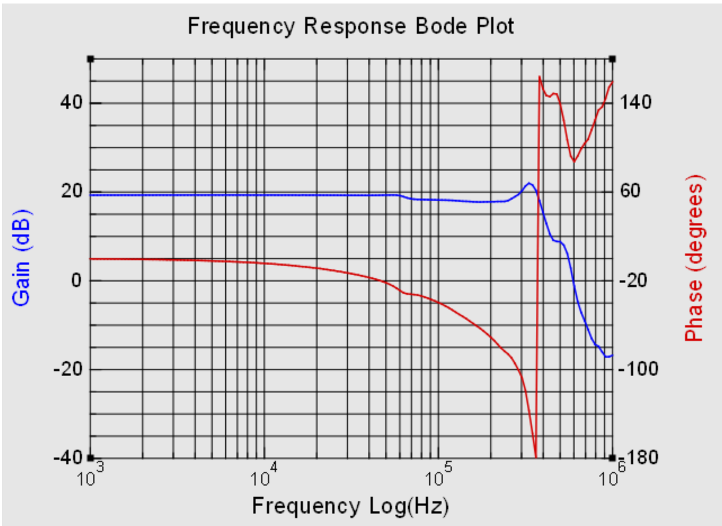


Figure 43. Gain-phase plot with series 2.0uH.

[8] Various small inductors based on ferrite cores and beads were prepared and tested up to 96kHz using the impedance measurement tool from REW software, where accurate measurement to about 0.1uH was obtainable. The impedance of the inductor beyond 100kHz,where feedback stability is influenced, depends on the type of ferrite used, and was assessed using the PicoScope 4224A and FRA4PicoScope software.

A combination of a bypass capacitance across R25, and adding a series inductor in the load ground path was able to further supress the small level of damped ringing. The plot on the right is for L1=2.0uH and 33pF of compensation capacitance. Including both forms of stability compensation reduces the value needed for the compensation parts.

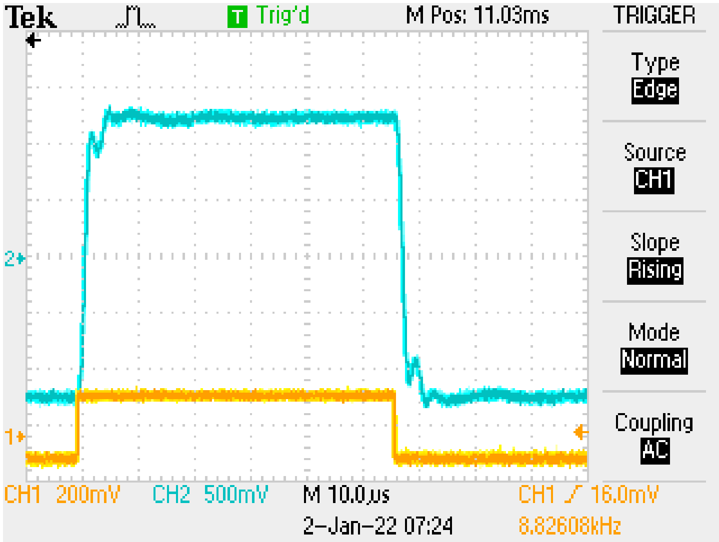


Figure 44. Squarewave response with series 2uH and 33pF compensation.

With the two forms of compensation applied as in the above test, but without a load connected, the squarewave response shows some overshoot but with similar resonance frequency and damping.

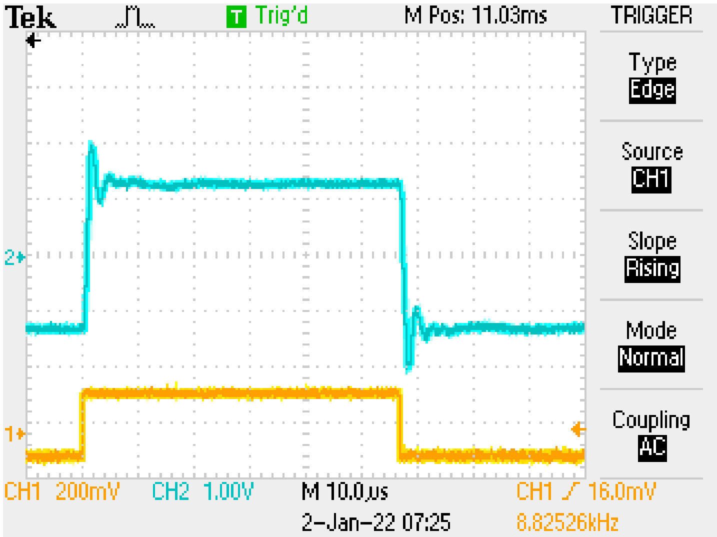


Figure 45. Squarewave response with series 2uH and 33pF compensation for no load condition.

When capacitance-only loading was applied (4.7nF, 22nF, 100nF) the response ringing was noticeably increased but was still damped as per the initial feedback plot above with no compensation.

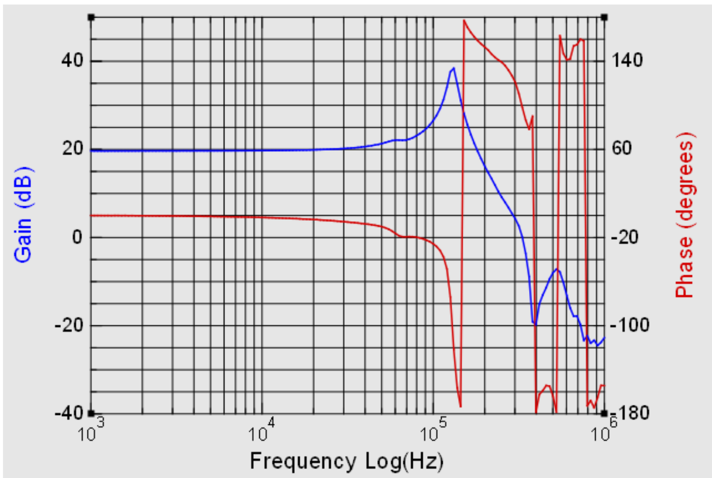


Figure 46. Gain-phase plot with series 2uH and 33pF compensation for 100nF load condition.

For this output transformer, the application of phase correction only compensation for HF stability provides a HF stable Williamson amplifier without the need to include a first stage step network.

Note that probing the output terminals should not ground the negative terminal when the series inductor compensation technique is used.

Ferguson OP25 (old)

This output transformer was assessed in an Audio Engineers Williamson amp for 15Ω output configuration. To achieve stable operation at no load, and up to 100nF loading, a 10Ω 220nF zobel network was needed on the output, along with a 150pF-10kΩ step network and 220pF parallel compensation across the feedback resistor. A strong 55kHz resonance with capacitance only loading was too low in frequency to manage with an inductor based approach, or a heavier step network (without compromising open loop gain bandwidth).

Trimax TA996A

This output transformer was assessed in diy Williamson amp[9] for 16Ω output configuration, and a 3k3 feedback resistor for 19dB feedback. Stable operation at no load, and up to 47nF loading, required a 220pF-4k7Ω step network and an 8T inductor and 690pF shunting each 220k grid leak. Any added compensation cap across the feedback resistor caused hf noise on the leading edge of a square-wave (asymmetric response).

HF stability options

Williamson appreciated that the amplifier could become unstable for capacitive loads like long speaker leads [WW 1949], and suggested the step R-C filter on the input stage anode load would be appropriate to reduce forward path loop gain above 20kHz.

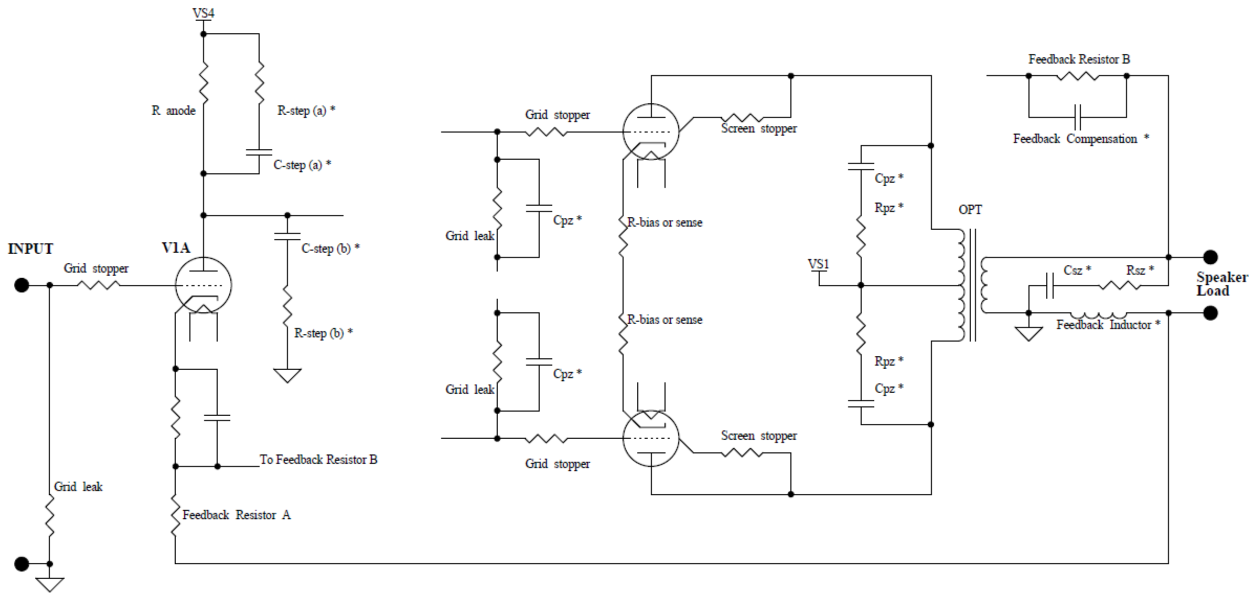
Forward path gain can be stepped or ramped down using an R-C or C across first stage R-anode or to ground, and similarly across the grid-leak resistance of the output stage, or as zobel networks across each half-primary winding, or the output secondary winding, as shown in schematic below. An advantage of ramping down gain closer to the input stage is that subsequent stages don't have to amplify as large a HF signal, especially as feedback is significantly reduced at HF signal frequencies. A concern with bypassing the feedback resistor with a capacitor is that resonances from capacitor only loading may couple through.

The feedback inductor has no effect on stability with speaker load disconnected, but does have an influence for capacitor only loading, and for normal audio bandwidth and roll-off.

Example speaker impedance measurements extending to 100kHz [12] indicated that one sample speaker, the Sony APM-X250, represents 100nF at 100kHz, with an ESR of about 7Ω, after a resonance at about 50kHz. Portions of a cross-over network could present LCR loading, or CR loading equivalent to a zobel network of large C (>1uF) and low R of about the speaker nominal resistance (eg. SEAS D014). The added series resistance of speaker leads are likely significantly lower than the speaker nominal resistance.

If the speaker was disconnected, then the speaker lead may be connected and may present a significant capacitance load. Measurement of a 'worst-case' speaker lead [10] of 6.5m length showed 11nF capacitance loading, with a self-resonance at 18Hz, and a likely impedance minimum out past 100kHz due to capcitance impedance reaching DCR.

Apart from testing amplifier stability for a 'no load' condition, it would likely be worthwhile testing with capacitance only loading of say 10nF, 47nF and 100nF (with a series 6.8Ω). Tests on a few OPTs indicate this level of stability often requires a feedback inductor, and RC ramp-down of forward gain, such that open-loop gain starts rolling off around 10kHz.



[9] Amp with 2x 6SN7 and simple RC supply droppers, and 6CM5 output stage. Split power supply providing 420Vdc to driver and input/PI stages, and 350Vdc to output stage, and fixed bias with separate bias adjust. 470Ω 6CM5 screen stopper, with 1k8 grid stopper, 220k grid leak. 0.47uF and then 47nF coupling caps.

[10] Tocord SHF-SP and later rebranded as Polk Cobra – a litz type cable product.

8. Results and discussion

Table 2 provides a summary of test results (excluding frequency responses) for the three OPT products. Two sets of results are given for the Ferguson OP25 (for old and new).

Output transformer	Partridge WWFB/0/0.95	Ferguson OP25/15	Red Line AF8/2
Primary inductance (50Hz)	88/80H for 3.8/8.5Ω	56/60H for 4/15Ω (old) 80/75H for 4/15Ω (new)	66/73H for 2/8Ω
Primary phase shift 20deg	3.5/4.5Hz for 3.8/8.5Ω	4Hz for 4Ω & 15Ω (old) 3Hz for 4Ω & 15Ω (new)	4Hz for 2Ω & 8Ω
Primary leakage inductance (1kHz)	14/16mH for 3.8/8.5Ω	30mH for 4Ω & 15Ω (old) 16mH for 4Ω & 15Ω (new)	38mH for 2Ω & 8Ω
Primary DC resistance	220+220Ω	207 + 224Ω (old) 215 + 232Ω (new)	277+276Ω
Primary self-capacitance	620/630pF for 3.8/8.5Ω	460pF for 4/15Ω (old) 450pF for 4/15Ω (new)	550/470pF for 2/8Ω
Primary resonance (sec open)	>95kHz for 3.8/8.5Ω	95/65kHz for 4/15Ω (old) 95/62kHz for 4/15Ω (new)	70/49kHz for 2/8Ω
Primary resonance (sec shorted)	48/46kHz for 3.8/8.5Ω	50kHz for 4/15Ω (old) 70/74kHz for 4/15Ω (new)	25/26kHz for 2/8Ω
3rd harmonic distortion level at 20Hz	0.11/0.12% for 3.8/8.5Ω	0.41/0.38% for 4/15Ω (old) 0.27/0.35% for 4/15Ω (new)	0.36/0.40% for 2/8Ω

Table 2. Output transformer measured performance

The primary winding inductance varies with excitation voltage, frequency, and any DC imbalance. The fall in inductance from circa 5Hz to 1kHz is likely due to a fall in permeability of the core as frequency increases [11]. As such, the primary inductance in the 5Hz region is about double the inductance measured at 50Hz, although this is counteracted by the measured reduction of inductance at low excitation levels when comparing to Williamson's requirement of 100H at 5Vac at 50Hz.

In Table 2, the primary winding leakage inductance is noticeably higher than specification for the Red Line AF8/2 and the original version of Ferguson OP25/15. An obvious change in design occurred for the OP25/15.

The primary self-capacitance of the WWFB is marginally higher than the specification, which may be a consequence of the measurement configuration and technique. The WWFB also has noticeably higher measured capacitance than the other two OPTs, but as they only extend two secondary sections it is unclear how much interleaving occurs compared to the WWFB [12].

[11] Heck's 1974 book 'Magnetic Materials and their applications' identifies an almost constant permeability below a critical frequency, where the critical frequency reduces for high permeability material and thicker sheet. Lee's 1955 book 'Electronic transformers and circuits' Figure 162 on page 217 shows a 0.014" GOSS example that is applicable to OPT lamination thickness and material and shows permeability falling rapidly above 1kHz, indicating the critical frequency is well below 1kHz. Macfadyen in J.I.E.E. 1947, p.407, plots the modulus and the angle of complex permeability over the frequency range from 50Hz for a range of lamination material and thickness. OPT's with a greater effective air-gap from different configurations of lamination interleaving should have less inductance variation due to a change in excitation voltage, frequency and DC imbalance as the air gap becomes a larger influence on the effective permeability of the core.

[12] An AF15 with open-circuited primaries is available for a tear down.

In general, manufacturer specification of OPT parameters is seen to be minimal, which may be just a marketing decision or related to what has survived in the public domain. The Partridge datasheets appear to provide the most detail but even they do not identify resonance frequency or primary inductance with DC current imbalance. Modern OPT specs appear to be even less detailed. The 1967 Ferguson datasheet below appears to be the most informative example identified so far.

FERGUSON

JANUARY 1967 OP 387

GRAIN ORIENTED
ULTRA-LINEAR OUTPUT TRANSFORMER
TYPE OP387

Ferguson Ultra-Linear Output transformer Type OP387 is a high fidelity type employing grain-oriented steel laminations. It is intended for use with Type EL84 or similar output valves. Type OP387 is thoroughly wax impregnated.

SPECIFICATIONS

PRIMARY IMPEDANCE....................8000 ohms, Plate to Plate
SCREEN TAP IMPEDANCE.................18.5% of plate impedance
SECONDARY IMPEDANCES...............OP387-15, 15 or 3.75 ohms
OP387-8.4;8.4 or 2.1 ohms
NOMINAL POWER RATING....................................12 watts
FREQ. RESPONSE ATTAINABLE(@ 1 watt)..20 cps-100,000 cps ± 1db
POWER RESPONSE ATTAINABLE(@10 watts).30 cps- 50,000 cps ± 1db
PRIMARY D.C. RESISTANCE (EACH HALF)...................90 ohms
MAX. D.C. PRIMARY CURRENT (EACH HALF).........80 Milliamperes
SECONDARY D.C. RESISTANCE...........Less than 6% of sec. imp.
PRIMARY INDUCTANCE
(At 20V 50 cps. 5 m.a. unbalance).............70 Henries
LEAKAGE INDUCTANCE............................7 Millihenries
PRIMARY FUNDAMENTAL RESONANCE........................140 KC.
INSERTION LOSS................................Less than 1db

Frequency response and stability

The following extract (Fig.7) from the May 1947 WW article indicates the frequency response of Williamson's OPT using the 1.7Ω secondary configuration. The WW article text identifies a 2.6dB resonance dip at 60kHz. The later 1949 WW article shows Figure 47 below, with the 60kHz resonance identifiable in the open-loop amp responses for gain and phase, with a gain dip of about 3dB and a phase shift swing of about 50deg.

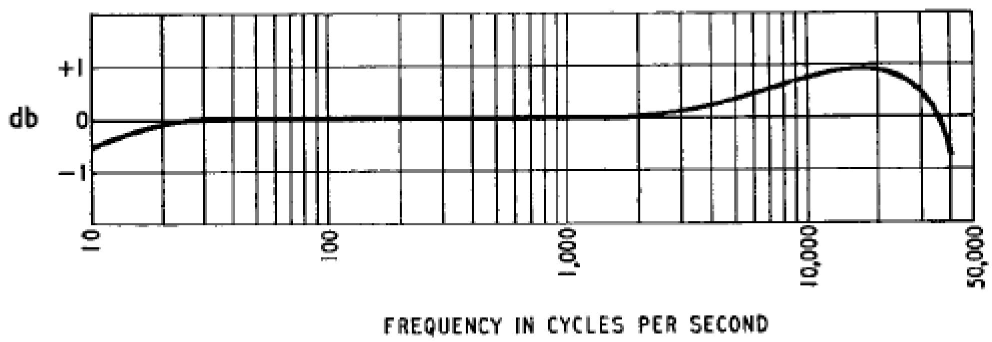
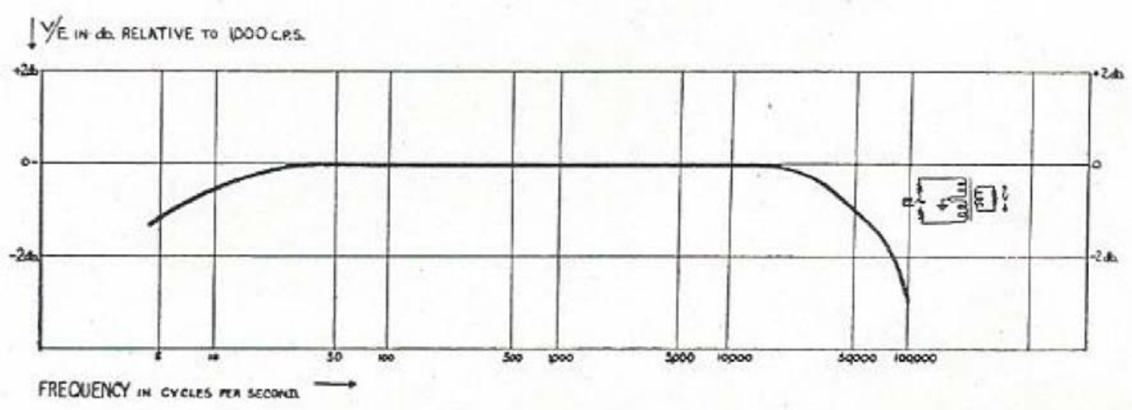


Fig. 7. Frequency response (without feedback) of 20 watt output transformer described in appendix. Generator resistance 2,500Ω load resistance 1.7Ω. Measured with 5V r.m.s. on primary. At higher excitations the bass response improves progressively up to saturation.

The WWFB frequency response presented in Partridge's Technical Datasheet No.2 is shown below and used a 3k3 resistive source for PP drive and a matched secondary load. The datasheet indicates a 10W signal level, which is assumed to be at the PP terminals (ie. 316Vrms), and there is no mention of the secondary configuration.



The WWFB frequency response in Figure 18 with 3.8Ω configuration does not exhibit any positive peaking, but does show a smooth high-frequency roll-off in magnitude with similar -1dB response out at 45kHz dipping to -4dB at 96kHz, and the start of a phase swing (only the initial part of the swing is observed below 96kHz in the plot). In contrast, the response in Figure 19 for 8.5Ω secondary exhibits more noticeable gain and phase changes beyond 40kHz, with the same gain dip from 60-90kHz. One could surmise that the excellent Partridge datasheet plot above is for the base 0.95Ω secondary configuration, or the base connection configurations of the WWFB 1.7 or 3.6Ω versions.

The manual frequency response results are made with a higher excitation voltage (10Vrms versus 4V), and the low frequency drop off is noticeably less for all OPTs when compared to the soundcard/REW plots. Manual results below about 1Hz contain significant distortion, making magnitude and phase measurement values quite suspect.

The manual frequency response results show equivalent gain dip levels to the plotted results for dips occurring below 96kHz, except for the WWFB OPT where plotted dip levels were significantly smaller than manual results (-3 to -4dB compared to -10 to -12dB). Gain-phase plot in Figure 31 for no feedback also shows a -12dB dip at 80kHz. It is uncertain how this discrepancy arises, but maybe related to the measurement technique and the WWFB location in a large amplifier chassis.

The OP25/15 low frequency droop in magnitude, and substantial phase shift at 2Hz is considerably more than the WWFB, with the new version being somewhat better than the old version. The poorer low frequency performance is not indicated by the primary inductance measurements or impedance responses.

The OP25/15 high frequency responses (for old and new) show a barely noticeable dip at about 30kHz which is just detectable in the impedance response. The extended high frequency response and reduced phase shift for the 3.75Ω compared to 15Ω configuration is consistent with the impedance response resonances. The new OP25/15 in a 3.75Ω secondary configuration shows a slightly extended high frequency response behaviour to the WWFB, but certainly not as smooth. High frequency results in a Williamson amp exhibited the same amplitude and phase change character to support the test results here.

The AF8/2 low frequency responses show magnitude and phase performance somewhere between the OP25/15 old and new versions. The AF8/2 high frequency responses show a significant resonance near 40kHz, especially for the 8Ω secondary configuration with an 8dB magnitude dip and 50 deg phase shift. The 8Ω secondary can be configured in two ways, and there may be some benefit in choosing a particular configuration to improve stability margins at high frequency. High frequency results in a Williamson amp exhibited the same amplitude and phase change character to support the test results here.

The impedance results reported in this article use an excitation voltage falling below about 5Hz due to the headphone output capability (-10% at 3Hz), so interpreting droop in inductance below 3-5Hz and especially below 1Hz is uncertain. This measurement uncertainty with inductance and change in phase shift is in a frequency range just below the phase splitter and the driver stage RC networks in the forward path of the Williamson amp, with two pole frequencies between 6-7Hz. Cooper [10] graphs the gain and phase responses of those two poles [13], along with the anticipated response of the OPT with 100H primary inductance.

Williamson's plot below indicates about 20-30deg phase margin, and 8dB gain margin and with the zero-crossing gain and phase occurring in the region of 2-3Hz [1], [9], and Cooper estimates a 10deg phase margin

[13] Cooper uses 150kΩ for the KT66 grid leaks, instead of 100kΩ - perhaps to provide some visual separation of the response plots for those poles.

and 4dB gain margin. Cooper shows the two power supply zeroes contribute about 25deg of phase correction at 2-3Hz, indicating that an OPT with less phase shift than 60 deg at 2-3Hz (or similarly less than 20 deg at 10Hz as was Williamson's original target) should benefit stability. All OPT's measured phase shift below 20deg at 10Hz, and below 60deg at 2Hz, although the WWFB was clearly the better performer and could likely provide about 50deg improved low frequency phase response at 2-3Hz [14] than the original Williamson OPT.

Also of note is the closed loop peak of +6dB at 2Hz, which indicates a need for caution when assessing measured impedance plots with actual amplifier response in the sub 5Hz region.

In the later August 1949 WW article, Williamson notes that it is not practical (without very special equipment at that time) to make direct measurements of forward path gain and phase at frequencies below 10Hz, and hence identify the performance and influence of primary inductance.

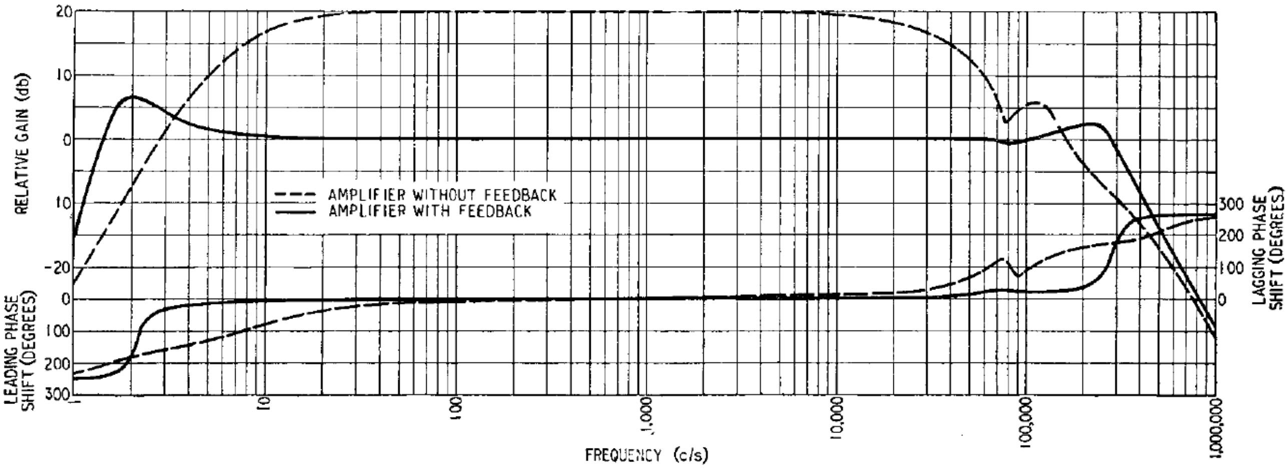


Figure 47. Williamson amp open and closed loop gain and phase responses

Cooper [10] assesses the high frequency stability by firstly assuming an OPT leakage inductance of 30mH and an effective driving impedance of 10k+2.5k=12.5kΩ causing a dominant pole at 63kHz, with the next significant contribution from the cathodyne stage coupling pole at 320kHz, and then the first stage and driver stage poles out further at 800kHz and 950kHz. From those poles, Cooper estimates the 180deg phase crossing at 410kHz, when gain is -22.5dB, giving a gain margin of 2.5dB and phase margin of about 15deg. My estimated pole frequencies for stage couplings [9] are 450kHz for the first stage and driver stage, and a 2-pole roll-off from 60kHz for the OPT.

The gain and phase curves in Figure 47 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.

Cooper estimates the inclusion of the 200pF/4k7 shelf network to change the gain-phase at 300kHz of the first stage from -1dB and 20deg, to -7.5dB and 32deg. And based on total phase reaching 180deg at 300kHz, and a gain drop from -18dB to -26dB, the gain margin is then improved to 6dB. Cooper estimates the 0dB gain crossing at 220kHz where the phase margin is about 20deg. Cooper's reverse-engineered estimates approximately align with Williamson's measured response in Figure 47.

With respect to the gain and phase response of the OPT, Cooper estimates the OPT has a phase shift of 75deg at 220kHz and -14dB gain at 300kHz. The WWFB manual measurements related to Figure 18 and Figure 19 indicate about 45deg and -9dB at those frequencies, with no nearby dips or bumps appearing significant enough to encroach on the stability margins. The other tested OPTs have measured phase shift of at least 90deg at 220kHz, although gain at 300kHz is typically not too different than -14dB, which indicates that they may be unstable due to excessive phase shift.

The gain-phase plots for the WWFB in the test AWA Williamson are shown in Figure 31 for no shelf network, and in Figure 48 with the shelf network. Without feedback, the shelf network starts to noticeably drop gain from an earlier frequency of about 17kHz, compared to about 30kHz, and a 20deg phase shift starts earlier at 7kHz, compared to 12kHz. However the shelf network does improve the indicative gain margin by about 12dB

[14] The WWFB would push the 0dB crossing frequencies down below 2-3Hz to about 1-1.5Hz but it is uncertain how much improvement in the phase and gain margins would occur.

(-15dB at 350kHz to -2dB at 480kHz), and improves the indicative phase margin by about 50 degrees, but noting that these gain-phase plots do not directly allow margin levels to be determined.

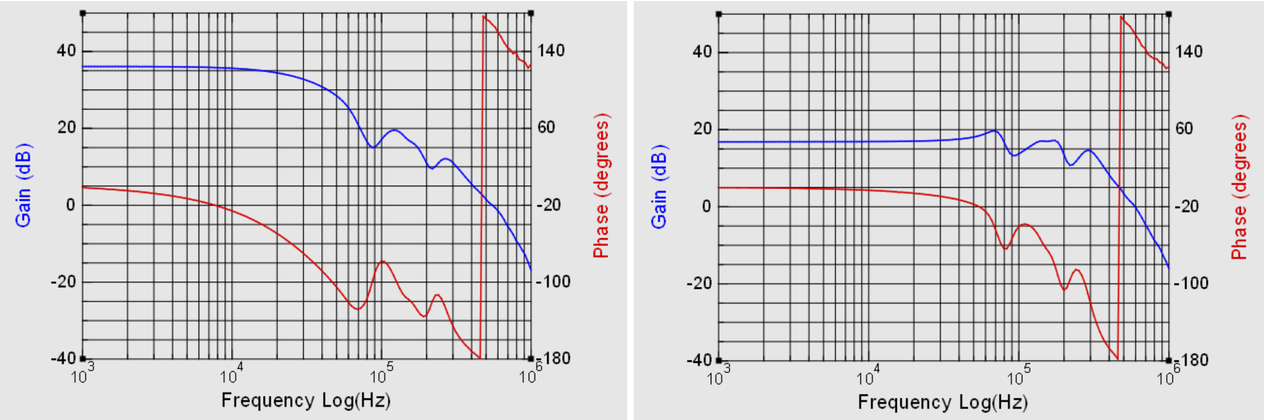


Figure 48. Gain-phase plots for WWFB and shelf network, without and with feedback.

With respect to high frequency feedback stability margins in an amplifier, the frequency response plots show an OPT can exhibit a poorer gain/phase characteristic for the higher speaker impedance configuration – presumably due to poorer interleaving – however the manual gain/phase measurements made at test frequencies higher than 96kHz indicate that (in general) there is little gain/phase difference in the frequency range where gain and phase margins are determined.

Testing of the new Ferguson OP25/15 in a Williamson amp in Section 7 showed that good squarewave response, and stable operation under worse-case loading, could be achieved using only phase correction compensation and without the need for the first-stage shelf network, with the only noticeable (but minor) resonance pushed out to about 300kHz.

Testing of the old Ferguson OP25/15 in a different Williamson amp showed stable operation with no load or up to 100nF capacitance loading needed a combination of output zobel network, phase correction compensation, and a minor first-stage shelf network, in order to retain a wide bandwidth. 10W power bandwidth extended to only 50Hz, likely due to the relatively small core and lower primary inductance.

Similar vintage Trimax 996A exhibited a first resonance around 80kHz.

The gain-phase plot of the Partridge CFB output transformer in Figure 37, compared to the WWFB in Figure 31, shows that high-frequency performance improved somewhat over the next year or so, given the CFB was introduced from about January 1951. Partridge presented the CFB improvements in their Technical Data Sheet No. 3.

Latter Partridge P2502 and 2503 output transformers pushed first resonance out to nearly 200kHz.

For capacitor only loading, a resonance peak often develops beyond 100kHz with 100nF, and out to 500kHz with 10nF, and that peak can instigate unstable operation. The resonant frequency is somewhat related to the capacitor value used (testing done with 10nF, 47nF and 100nF), and indicates a secondary winding leakage inductance typically in the range 10uH to 40uH for a 16Ω configuration (typically with secondary segments wired in series). There may be advantage to a low ohm speaker configuration that parallels secondary segments, if that lowers the secondary winding leakage inductance, and raises the resonance frequency enough.

Harmonic Distortion

The harmonic distortion result summary in Table 2 shows the WWFB having significantly lower 3rd harmonic distortion at 20Hz than the other OPTs. The frequency characteristic of the harmonic distortion components was similar with all OPTs.

Partridge is the only manufacturer known to publish a distortion curve for their OPT, as shown in Figure 49. The reported distortion dips below 0.5% THD for power below 0.01W. The testing in this article is down around 0.001W due to the relatively low excitation voltage, and hence core flux level, and really needs measurement data at higher voltages to show the distortion curve characteristic approaching the design limit of the OPT as presented by Partridge.

Although the soundcard/REW setup used for this article allows impedance and frequency response to be measured from about 1Hz to 95kHz [15], it would be informative to be able to extend that frequency range (using a 384k/s sampling rate soundcard) to better characterise the OPT at the frequency limits, especially the high frequency end where the Williamson amp's HF gain and phase margins are determined. However, most opamps made for audio do not have bandwidths much above about 120-140kHz, so even the latest interface devices with 384 and 768kb/s sampling rates do not provide bandwidths above about 140kHz. So automation of results would likely require a script based control approach using a controllable signal generator and high frequency voltmeter. Manual measurement of magnitude dips/bumps/flats at high frequency is likely to be acceptable for base level stability assessment.

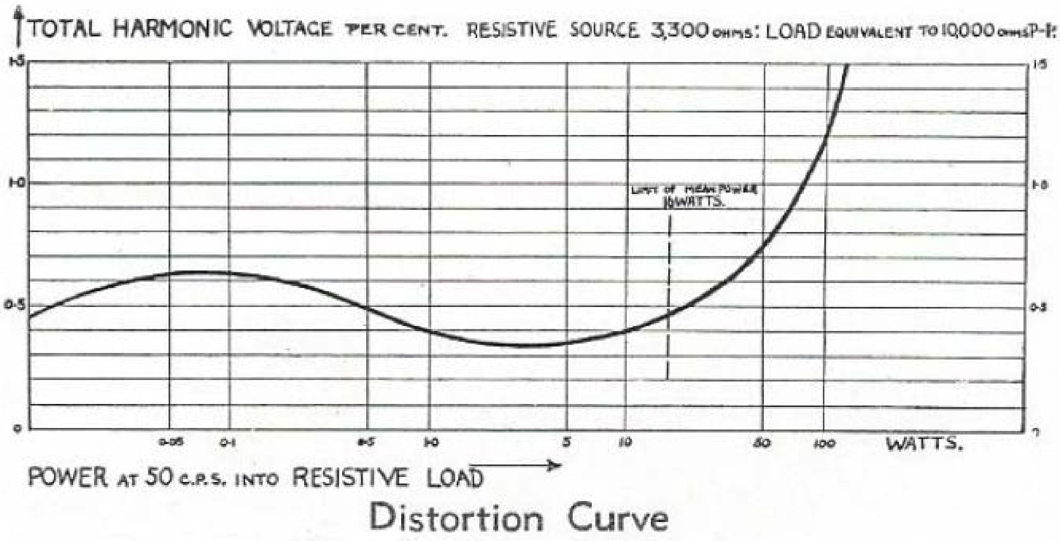


Figure 49. WWFB datasheet distortion curve

[15] The EMU0404 USB has an input channel response roll off that can be extended by lowering the sampling rate. At 192kHz sample rate and 2Hz, the roll off is -11dB and 140deg phase shift. At 44kHz sample rate and 2Hz, the roll off is -5dB and 100deg phase shift. However, lowering the sampling rate then limits the high frequency response. The output response is not affected by sampling rate, or which output port is used.

9. References

[1] 1952 The Williamson Amplifier WW compendium.

[2] Partridge WWFB and CFB datasheets

[3] R.F. Gilson technical specification for WO 1796 and 1796A output transformers.

[4] Room EQ Wizard software and Impedance Measurement.

[5] Partridge information with connection details.

[6] High Fidelity Circuit Design, N.H.Crowhurst & G.F.Cooper, 1956.

[7] Audio Measurements, N.H.Crowhurst, 1958.

[8] Power Supply Choke Measurement,

[9] Williamson design info

[10] Audio feedback design Part 3, G. Cooper, Radio Electronics, Dec 1950.

[11] Turner Audio discussion of L1=0.8uH for HF stability

[12] 'Speaker Impedance' by Max Robinson, circa 2014.

Known output transformers with 10kΩ P-P primary impedance, and designed for or used in a cloned Williamson amp, or a close variant are:

o	Vortexion, UK	Series I			
o	Vortexion, UK	Series II	from June 1951	[1]	
o	Partridge, UK	?	from Sept 1947		
o	Partridge, UK	WWFB	from Aug 1949	[1]	
o	Partridge, UK	CFB	from 1951		
o	Partridge, UK	T/CFB	from 1954		
o	Savage, UK	2B36	from Dec 1949	Ref	
o	Savage, UK	3C67A		[1]	
o	Elstone, UK	MR/W		Ref	
o	Woden, UK	WOT.25 / WOT.26	from Jun 1949	Ref	
o	Gilson, UK	WO.1796A		Ref	
o	Gardners Radio, UK	O.P.735 & O.P.736		Ref	
o	Red Line, AUS	AF series	from Dec 1947	Ref	
o	Ferguson, AUS	OP25	from Feb 1948	Ref	
o	Trimax, AUS	TA996A			
o	Bramco , AUS	HF-4			
o	Beacon Radio, NZ	48S06	from July 1948	Ref1, Ref2	
o	Wiseman Electric, NZ	'Williamson'	from Oct 1953	Ref	
o	Stancor, USA	A-8054			
o	UTC, USA	LS-63		Ref	
o	UTC, USA	LS-60A		Ref	
o	TRIAD, USA	HSM-81			(8kΩ P-P)
o	Freed, USA	F-1959, KA-10		Ref	
o	Hammond, Canada	1770, 1772		Ref	

o Jorgen Schou, DK	Type 350		Ref
o Sansui, Japan	HW-731, HW-733	from April 1954	