Towards a reconciliation of measurements with listening tests

1. Introduction

A vexing question in audio is how come we can connect amplifiers giving THD figures of 0.001% to loudspeakers giving THD figures of 1% *without* compromising quality? Similarly, how come tape recorders decades ago with 1% distortion sounded really great, and 16-bit CD players that replaced tape recorders with 0.003% THD were later found inadequate [1,2] and were superseded by DVD audio players with at least an order of magnitude lower distortion? Is there any logic in all this?

The reason for this apparent nonsense lies in the inadequacy of the Total Harmonic Distortion (THD) measurement that we use to assess distortion in audio equipment. THD is the ratio of the total power in distortion harmonics divided by the total power of the signal. It is popular because it is based on the RMS value of the distortion which can be relatively easily measured. Unfortunately, THD is an inappropriate indicator of how well or poor an amplifier actually *sounds*.

This has been known since the beginning of audio engineering. D. Masa around 1930 said, “The total harmonic distortion (THD) is not a measure of the degree of distastefulness to the listener and it is recommended that its use should be discontinued.” [3,4]

THD does not take into account the way our ear hears the range harmonics produced by nonlinear distortion generated while passing through audio equipment. THD assumes all harmonics sound the same, so, the obvious 'fix' would be to weight each harmonic by a suitable factor to reflect our increased sensitivity to the higher harmonics Quite a few harmonic weighting ideas have been looked at [5-10] but no method has been become popular with power amplifier designers and reviewers so far.

A diachronic disconnect between subjective listening tests and objective distortion tests using THD arose after transistor amplifiers appeared in the 1960's [11,12]. Before that time valve amps produced almost no high-order distortion that transistor amps produce and before that time valve amps THD measurements correlated well with how good a valve amp sounded. So at that time designers and reviewers had no need to replace THD and Masa was ignored. But when transistor amps appeared with their high-order distortion, THD measurements no longer correlated with how good an amplifier sounded.

Transistor amplifiers were able to use very high levels of negative feedback to reduce THD to much lower than was possible with valve amplifiers and so by the mid '80s levels of 0.01% THD became common place for transistor amplifiers. But even that level was found to be not low enough for some listeners, so by the mid '90s THD figures of 0.003% became common for transistor amps [2,13].

Nearly every designer of transistor power amplifiers now aims for the lowest THD that the technology can provide, hoping that high-order crossover distortion can't possibly be audible. My gripe with this brute-force paradigm is that it deters designers from investigating low feedback and other non-conventional topologies with relatively high THD figures which, surprisingly, often sound very good.

In hind sight, to avoid this fiasco, we should have taken Masa's advice and used a weighted distortion measurement method when transistor amps first appeared. If this had been done then the relentless downward trend of THD figures over the decades would probably have looked different and the problem of crossover distortion would probably have been dealt with long before now.

There are quite a few non-conventional topologies that have been largely ignored due to our faulty THD distortion metric. My low feedback Square-law Class-A amplifier [14] is an example. Someone commented that its distortion is "quite high" at 0.1%. Nevertheless, I found that it does sound very good. There are no high-order distortion products so it should sound good despite its 0.1% THD.

Surely there must be a better tool than brute-force THD measurements for assessing how good an amplifier is likely to sound. This article presents a distortion measurement methodology that gives a much better indication of the true, perceived audible quality of an amplifier's distortion than regular THD measurements.

2. A Better Approach

For a number of years I have been using harmonic weighting filter in distortion simulations to scale the distortions harmonics. I add a filter after the power amplifier and before THD analysis to weight all distortion harmonics during the simulation. This can also be applied to work with bench tests as well.

This article first shows how to use published auditory curves to specify a weighting filters response. I then weight distortion from a simple Class-AB power output stage with several filters and compare their results with a spreadsheet using accurate coefficients. Having found a good filter I then simulate a novel minimal feedback bipolar Class-AB topology using weighted distortion analysis. Some bench tests are also included.

3. Hearing sensitivity curves

Our hearing works a bit like a 24 channel real-time spectrum analyser [1,15] where the dynamic range 'floor' is determined by our auditory threshold curve [16] and band-widths vary from 1-octave at low frequencies to 1/5th of an octave in the 1-3kHz region [1]. The skirts from each band overlap and this prevents a faint tone of one frequency from being heard in the presence of a louder tone of another's slightly different frequency. This effect is called masking because when the loud tone is removed the faint tone at a different frequency becomes audible. For example, our threshold for 2nd harmonic distortion is about 2% of the fundamental level which is high relative to the high-number harmonics – masking prevents us from hearing significant second harmonic distortion since the 2nd is so close to the fundamental.

For a 19th harmonic the threshold is about 0.01%, about 100 times more sensitive than the 2nd. In this case masking is not the determining factor because the frequencies are so far apart and it is determined by our auditory threshold curve. Our hearing has a peak sensitivity in the 1-3 kHz range so we can hear very faint harmonics from distortion if they fall in this sensitive frequency range (with negligible background noises). For example, the 19th harmonic needs a fundamental of 50Hz to 150Hz so it appears in the sensitive 1-3 kHz range

Figure 1 shows some hearing curves for a 50Hz sinewave to give an idea of what percentage distortion is audible at three widely different sound pressure levels [17]. The lines delineate the lowest "just audible" level for any harmonic. If the harmonic level lies under this curve then it can be considered "inaudible". Harmonics that lie above the line are considered "audible".

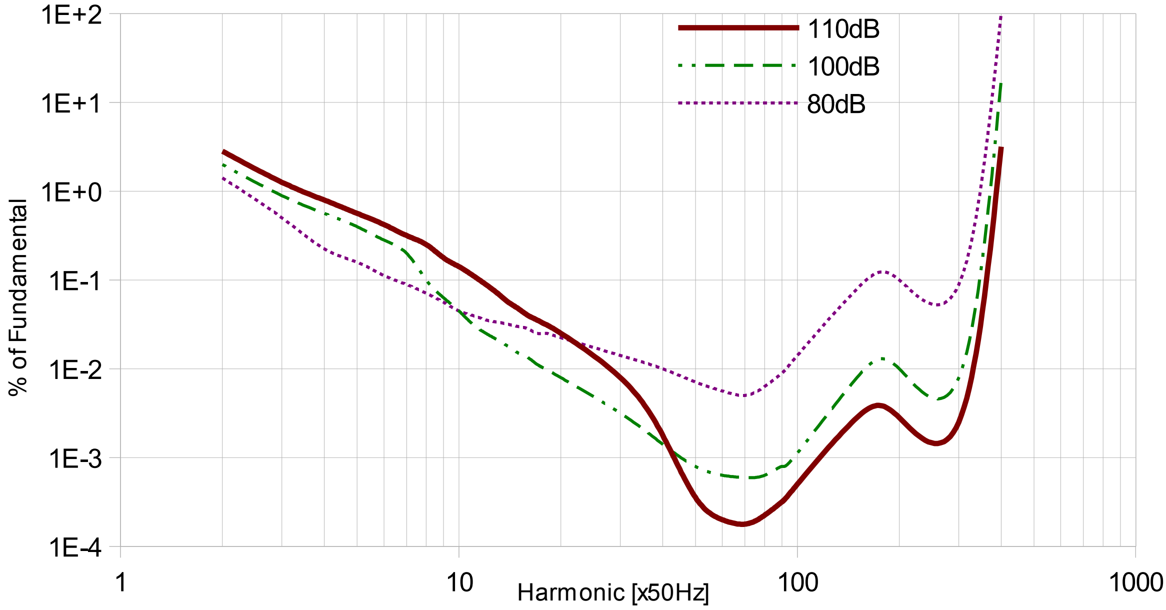

Figure 1. Average listener distortion thresholds with a 50Hz sinewave for 80dB, 100dB and 110dB levels. From JAES, Fielder, 1988 [17].

What about multiple harmonics and noise? If two or more harmonics are within a hearing 'band' (about 1/3rd of an octave from 500Hz to 15kHz) then their levels can be RMS summed. Collectively, multiple harmonics can push the level into audibility even though each may be just below the threshold. But for this to occur they need to fall within the same band. This condition also applies to noise components [1].

The question of the accumulation of a very large number of harmonics is an interesting one. Some claim that large numbers of intermodulation products *all* sum to become audible in complex signal such as music, and if true then extremely low THD figures are required in audio [13,18]. But it seems unlikely due to the band requirement for accumulation. Also not all amplifiers produce high-order distortion that gives rise to a very large number of harmonics.

Class-A amps are an example. Their so called “inherent linearity” where reducing power results in progressively more linearity means that at the normal low average power levels with music the linearity in Class-A is so good that few harmonics are generated and significant accumulation of harmonics doesn't happen. You can have relatively high levels of distortion at full power but at normal lower average power levels with music the distortion falls well below the audibility thresholds.

But with Class-AB the linearity does not improve at normal lower power levels with music and so accumulation of high-order distortion harmonics is a concern with Class-AB. Class-D and 16 bit audio coding with noise shaping was inadequate for the same reason [1].

The effect of noise levels (dynamic range) on threshold levels is also an interesting one. Noise is added by the processing equipment and from the listening environment and biophysical effects in our ears. Different levels of added noise can change listening test conclusions and this may be one reason why contradictory results are reported for the same amplifier. Noise levels are also a factor in tests where distortion is below the threshold audibility and listener fatigue is being evaluated [11,19,20].

The listening room's background noise is the hardest to control and Fielder found that this type of noise typically dominates and masks recording environment noise and microphone noise by about 10dB over the 50Hz to 5kHz range [21]. Interestingly, our hearing mechanism relies on some noise to set its thresholds, since it is a self-tuning dynamic system and our ultimate sensitivity is set by Brownian noise in absence of external noise [22]. Strangely, we do not hear Brownian noise as a constant sound (apart from tinnitus sufferers) and the reason is not yet understood. This mechanism may explain how we can hear faint tones buried in audible noise to around -10dB below the noise. THD+N measurements (which are usually used) ignore this hearing ability. FFT residual analysis using a DSO after the distortion meter can filter out the noise giving just the THD part and as a bonus it also increases the resolution to 2 ppm or 0.0002% [23,24].

The question of whether thresholds in audio actually exist is also debated. But most agree that there are physical limits for *all* sensors, due to accepted noise and information-theory limits [25]. The ultimate hearing limit is given later.

Table 1 summarises these thresholds up to the 20th harmonic for three tones, 20Hz, 50Hz and 100Hz for 3 sound levels of 80dB, 100dB and 110dB.

Fund [Hz]	2nd [%]	3rd	4th	5th	6th	7th	8th	9th	10th	11th	12th	13th	14th	15th	16th	17th	18th	19th	20th
	110dB level																		
20	5.0	1.6	0.7	0.3	0.2	0.1	0.08	0.06	0.05	0.04	0.032	0.025	0.022	0.019	0.016	0.014	0.013	0.011	0.010
50	2.8	1.3	0.8	0.6	0.4	0.3	0.25	0.18	0.14	0.11	0.089	0.071	0.056	0.047	0.040	0.035	0.032	0.028	0.025
100	2.5	1.5	1.1	0.9	0.7	0.6	0.50	0.45	0.40	0.36	0.316	0.266	0.237	0.188	0.158	0.112	0.089	0.071	0.056
	100dB level																		
20	4.0	1.3	0.5	0.3	0.2	0.1	0.08	0.06	0.05	0.04	0.03	0.03	0.02	0.019	0.016	0.014	0.013	0.011	0.010
50	2.0	0.9	0.6	0.4	0.3	0.2	0.10	**0.06**	**0.05**	**0.03**	**0.03**	**0.02**	**0.02**	**0.015**	**0.013**	**0.011**	**0.010**	**0.009**	**0.008**
100	3.2	1.8	1.3	0.9	0.7	0.6	0.45	0.36	0.28	0.22	0.18	0.14	0.10	0.075	0.056	0.040	0.032	0.022	0.018
	80dB level																		
20	6.3	1.6	0.7	0.4	0.3	0.2	0.18	0.14	0.11	0.09	0.08	0.07	0.06	0.056	0.050	0.047	0.045	0.042	0.040
50	**1.4**	**0.5**	**0.2**	**0.2**	**0.1**	**0.1**	**0.07**	**0.06**	**0.05**	0.04	0.04	0.03	0.03	0.030	0.028	0.025	0.025	0.024	0.022
100	1.6	0.7	0.4	0.2	0.2	0.1	0.09	0.06	0.05	0.04	0.03	0.03	0.02	0.020	0.018	0.016	0.014	0.013	0.011

Table 1 Single-harmonic perceptual distortion limits to the 20th harmonic.

Notice the **bold** values in Table 1, they identify the most sensitive conditions for the first 20 harmonics which shows that a 50Hz fundamental gives the highest possible hearing sensitivity for any harmonic up to the 20th. The first 10 harmonic are most sensitive at a quieter 80dB SPL. The remaining harmonics from the 11th to 70th harmonic require a loud 100dB or 110dB tone (Fig. 1).

For example, for the 13th harmonic of 50Hz (650Hz) the threshold is 0.02% and is just audible at 100dB SPL. Class-AB crossover distortion produces high-order harmonics in this harmonic range so the design of Class-AB amplifiers with THD distortion levels of under 0.01% THD can be justified by this data.

No data was available for higher fundamental frequencies since the source data was intended for designing a subwoofer operating up to 100Hz. But earlier tests on a Class-AB MOSFET output stage [26] showed a 300Hz tone at a quiet level of 80dB was best for hearing the main 3rd, 5th and 7th harmonics since the higher-order harmonics were not audible.

Another important observation is that our ultimate hearing sensitivity from Figures 1 is -110dB or 0.0003% with a sound level in the range of 100dB to 110dB. For example, any harmonic from the 30th harmonic to the 60th of 100Hz will be just audible given a very quiet environment. This is usually outside the harmonic order for crossover distortion. Some output stage types generate very sharp gain changes, such as the Sziklai compound feedback transistor pair (CFP) when operated in Class-AB (but not in Class-AB) and CFP crossover distortion is expected to be heard at lower THD levels than the Darlington pair.

The data above are for averaged hearing sensitivity. Hearing sensitivity varies from individual to individual and some listeners will be able to hear harmonic distortion and noise at slightly lower levels than predicted by these curves. But how good are the best of the 'golden eared' listeners? As far as I can discover [27], one standard deviation (1-sd) represents a sensitivity change of +-3dB, and thus 3-sd listeners are about 10dB more sensitive than the averaged data used here. When masking is not happening and background noise is low enough then a 10dB advantage can be added to cover most of the golden-eared listeners as well and their ultimate threshold is 0.0001% (-120dB or 1 ppm) with caveats.

Music does not reveal distortion as easily as a low frequency single tone because of masking since the main tones are often within an octave of middle-C. But there is likely to be some music that can reveal distortion just as well as low-frequency single-tone tests, so thresholds should not be relaxed for *all* music.

For peace of mind designers like to apply safety margins to cover unknown variables that can arise in production. For audio power amplifiers applying a safety margin of 10 lower distortion is probably a good level. Robert Stuart [28] suggested the basic distortion specification should be changed to a weighted-THD of 0.1% which is a factor of 20 lower than the just-audible threshold for 100dB and 50Hz.

4. Specifying a weighting filter

It is a tedious process to individually measure each harmonic and manually compare them with thresholds in Table 1. A way to automate the process would be very much appreciated.

For circuit simulations it is convenient to pass all the harmonics through a filter having an inverse transmission function to our hearing curves. **Figure 2** shows an upside down Figure 1 and normalised to the second harmonic for a 50Hz fundamental. This gives three bandpass functions that depend on the sound level, which is determined by both the loudspeaker sensitivity and the amplifiers output power.

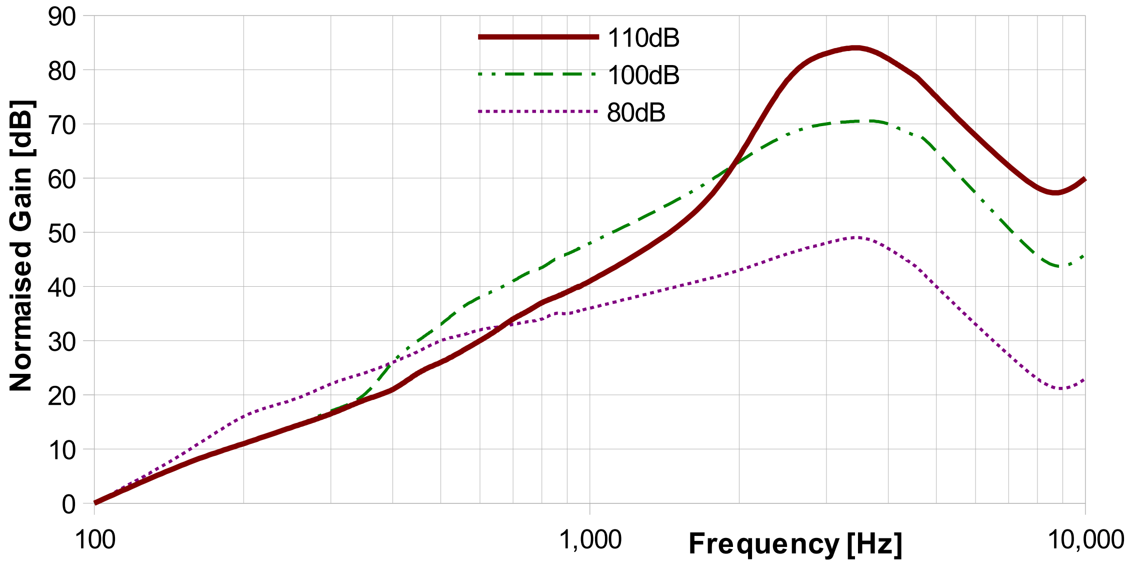

Figure 2. Weighting gains for a 50Hz fundamental at 3 sound levels.

It would be convenient if just one filter could be chosen that covers the most of the likely listening conditions. The highest sensitivity to high-order distortion is with 100dB SPL and for this level the filter roll-up slope is around 50dB/decade. Crossover distortion's high-order harmonics are considered here to be the 7th and above. A Band-Pass Filter (BPF) with slope of 50dB/decade, a centre frequency (f_0) of 3.5kHz, and a low Q appears to be required. Interestingly, the filter's gain at 2kHz or the 40th harmonic of 50Hz is 60dB (1000 times) relative to the 2nd harmonic. It shows just how sensitive our hearing can be to high-order harmonic distortion.

The filters roll-off slope is around 5th order. But in simulations the roll-off slope is not very important since the FFT command line can be arranged to truncate all harmonics above the pass-band effectively acting as a high slope digital filter.

The weighting function used by Shorter [5] used ¼ n^2 or the harmonic number squared which is equivalent to a 2nd order filter and he found that this is still not enough weighting to make high-order harmonics score as high as listener tests rated them. Therefore a filter with a 3rd order roll-up slope may give a better correlation with the perceived audible quality of an amplifier's distortion. Shorter used the mathematical weighting of ¼ n^2 where N is the harmonic number (n>2). The ¼ term preserves the 2nd harmonic percentage level while higher harmonics are progressively boosted. For a 3rd order type weighting function we would use 1/8 n^3 for n>2. At high frequencies the weighting function should stop increasing around 3kHz and remain constant or reduce (when n exceeds about 50 with a 50Hz fundamental).

5. A-weighting filters

Standard A-weighting filters are bandpass filters that attenuate low frequencies and very high frequencies but pass mid-high frequencies in the 1-3kHz range. They are normally used to weight wideband noise in the audio frequency range in a way similar to our hearing with quiet signals.

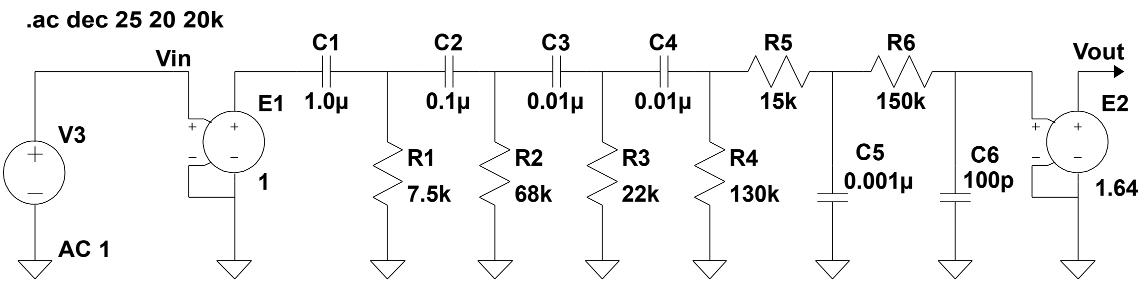

Figure 3. Single A-Weighted filter circuit.

Figure 3 shows a single A-weighted filter circuit courtesy of Bob Cordell's AWF subcircuit [29] in the LTSpice circuit simulator [30] and plot data was imported into a Calc spreadsheet for plotting. See the section at the end to download and include the AWF subcircuit in your circuits.

The frequency response for a single filter is shown in **Figure 4** (fine line). Figure 4 also shows two (dotted) and three filters in cascade (bold). The dashed line show a 3rd-order slope passing through 50Hz. Each filter stage includes a gain stage of 1.64 on the output which gives a gain of 0dB at 1kHz. The passband frequency is 2.5kHz where the gain peaks at 1.5dB (x1.2) for the single filter. The desired centre frequency above is 3.5kHz (Figure 2) and 2.5kHz is close enough. The question now is how many stages are needed to give a closest weighting at 100dB SPL?

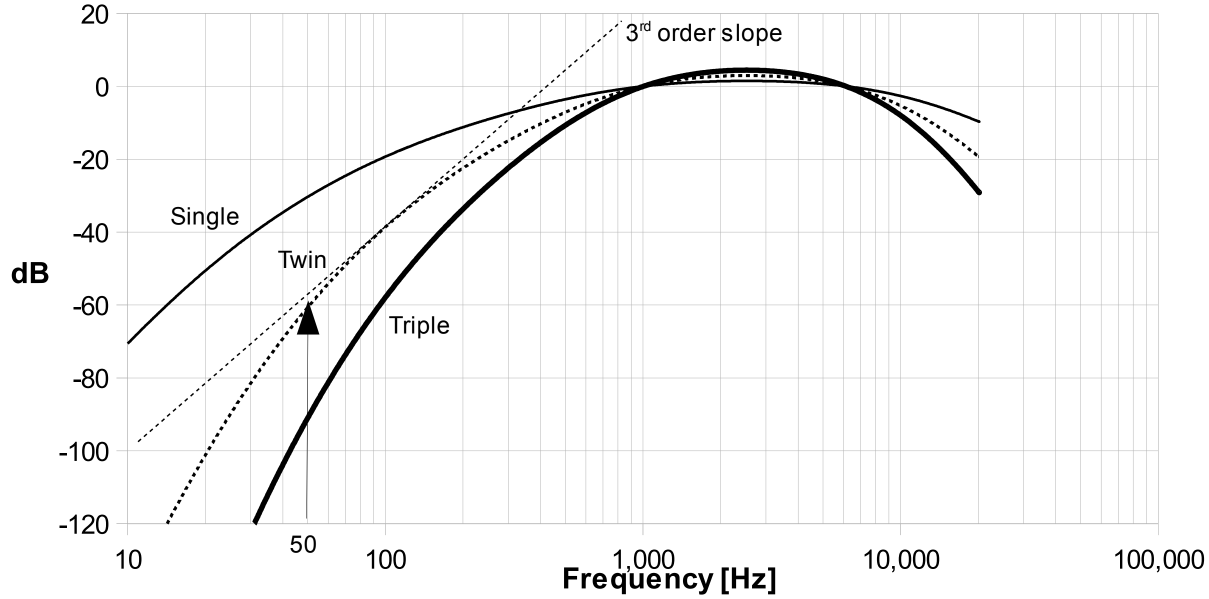

Figure 4. A-Weighting filter responses for single (upper), twin (dotted) and triple-filter (lower). Dashed line show a 3rd order slope passing through 50Hz.

The twin A-weighting filter appears to provide the required 3rd order slope. The roll-off slope with a twin filter is only about 20dB per decade. As mentioned, a very fast roll-off is not necessary for simulations.

A simulations is provided later to determine how many A-weighting stages should be used to obtain the best weighting for a 50Hz tone and 100dB SPL.

6. CCIR-468 noise weighting

This is an alternative noise weighting curve to A-weighting. Dolby Laboratories used the CCIR-468 weighting curve for measuring their noise reduction systems and this has been renamed to the ITU-R 468 weighting curve [9]. It provides about the same average roll-up slope as A-weighting. The 468 filter peaks higher at +12 dB at 6.3 kHz to better capture the sensitive 1-3kHz range and provides a very fast rolloff above 10kHz (like our hearing). It's gain at 1kHz is the same 0dB as A-weighting. Lindos Electronics' website mentions using the '486 curve for weighting distortion using a 1kHz tone [10]. **Figure 5** shows a simulation 468 filter circuit [31]. **Figure 6** shows the responses for single, twin and triple filters. The slopes are 20, 40, and 60dB/dec.

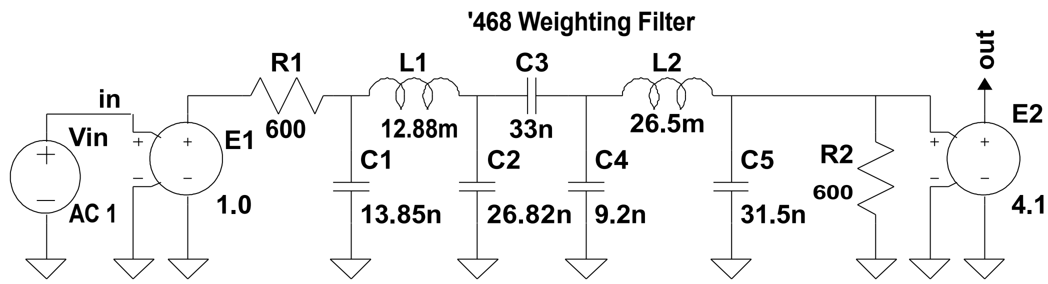

Figure 5. CCIR-468 weighting filter circuit.[1]

1 An was error spotted by Burkhard Vogel in Fig 5 circuit: L1 and C2 were incorrect, now show the correct values.

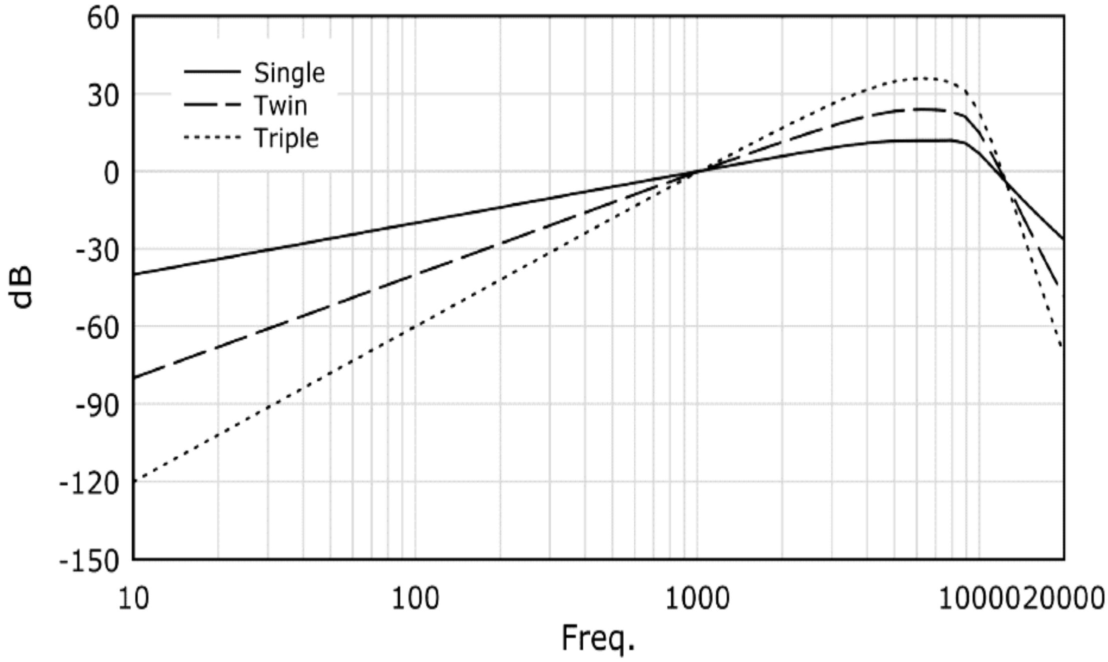

Figure 6. CCIR-468-Weighting filter responses for Single (Bold), Twin (dash) and Triple-filter (dotted). Roll-up slopes are 20, 40, 60dB/dec respectively.[2]

7. Distortion simulation

I will use a simple two transistor output stage that can deliver 100dB SPL using a typical 89dB/W/m loudspeakers. This requires 80W rms [32]. For 80W rms we need 36V peak into 8 ohms, so with 0.33 ohm emitter resistors an extra 2 volts is required so a 40V split-rail supply is used to prevent clipping with 80W rms into 8 ohms.

The transistor parameters have been chosen such that the distortion is only from the crossover nonlinearity with the bipolar transistors in Class-AB when optimally biased. The parameter for the Early effect is disabled by making the VAF parameter very large. Also the transistors are matched. This allows the inherent nonlinearity from crossover distortion to be simulated. The other disabled secondary effects can be introduced later to see how they alter the underlying minimum crossover distortion for a bipolar Class-AB stage.

Figure 7 shows the Voltage Follower (VF) circuit with twin A-weighting filters as AWF subcircuits.

Figure 8 shows twin and triple filters. The amplifiers gain of around 0.96 is mirrored by R3, R4. Two mirrored banks are used to make it easy to obtain the distortion residual without manual trimming. The mirror bank provides the right amount of fundamental to cancel the fundamental in the distortion signal. E2 and E3 sources act as differential amplifiers and their gain is chosen such that the 2nd harmonic is unaffected by the weighting filter. An undistorted signal is reintroduced to the amplified residual at the negative side of E2 and E3 outputs for FFT analysis and THD percentage figures can be read directly without further scaling.

2 Now show the correct plot [ref Linear Audio, Letters, Dec 2012 view here.

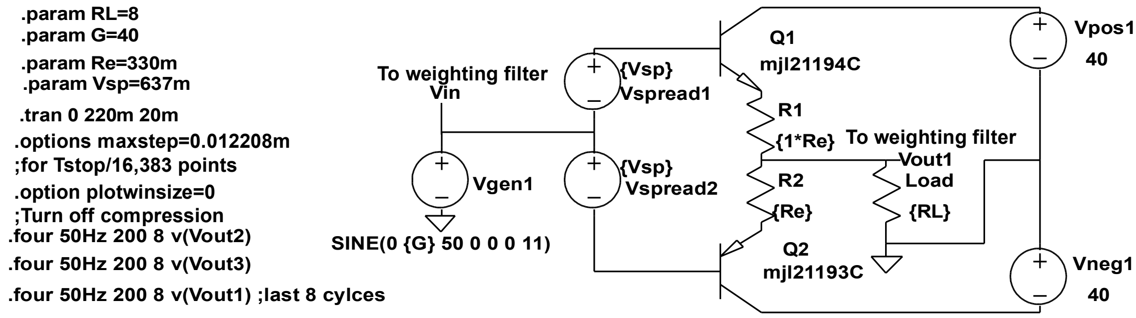

Figure 7. Simulation of a Voltage Follower for 100dB 80W into 8 ohms.

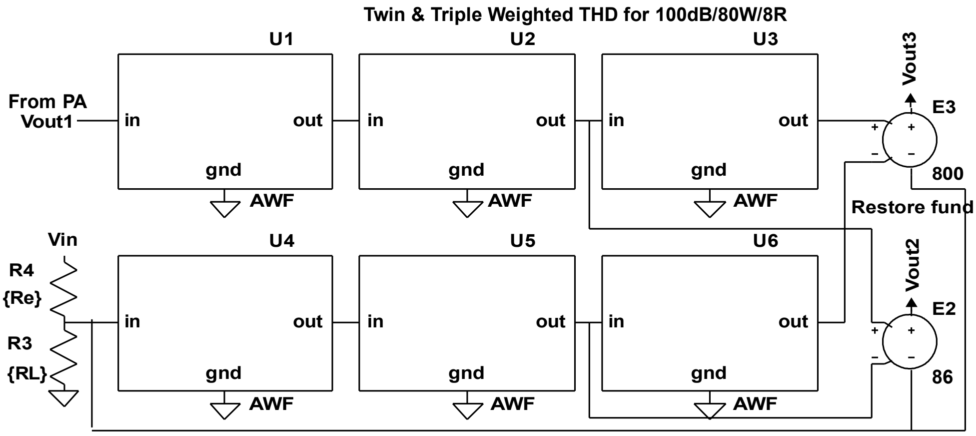

Fig 8. Filter banks for Twin and Triple A-weighting using the Total Difference method.

The command `.four 50Hz 200 8 v(Vout2)` generates FFT results using the last 8 cycles for the first 200 harmonics of 50Hz, in this case for the twin filter output.

When the 2nd harmonic distortion passes through the filters Figure 4 shows it is attenuated by -38.6dB (x1/85.1) for twin filters and -58.0dB (x1/794) for triple filters so E2 and E3 gains can be set to these figures so the 2nd harmonic is unaffected. The values for E2 and E3 can be checked by unbalancing the emitter resistors R1, R2 to give some 2nd harmonic and E2 was changed to make the 2nd harmonic in the Error Log for Vout2 the same reading as the 2nd harmonic Vout1. This required a gain of 86 for E2 and a gain of 800 for E3 so these values were used in Figure 7.

Figure 9 shows the simulated distortion spectrum for 5 cases; the unweighed harmonic spectrum (lowest), twin and triple A-weighting filters (middle) and the desired filter response (top two) based on Figure 2 100dB SPL curve. The threshold for 100dB and 50Hz is shown at 2% (Table 1).

One of two desired responses uses the weighting function $(n/2)^m$ where *m* is the power law, and *m* was varied in the spreadsheet until the weighting function gave the closest fit to the weighting response in **Table 2** (100dB). An upper limit was used limit at 1.5kHz (30th harmonic of 50Hz) and n remained constant above 1.5kHz. Interestingly, the power law for a good fit is 2.2, slightly more than Shorter's power law of 2.0 but as expected from Figure 2 is less than 3 (a 3rd order slope or 60dB/decade). The weighting function is therefore $(n/2)^{2.2}$ (n≥2 and n<2.5kHz/f1).

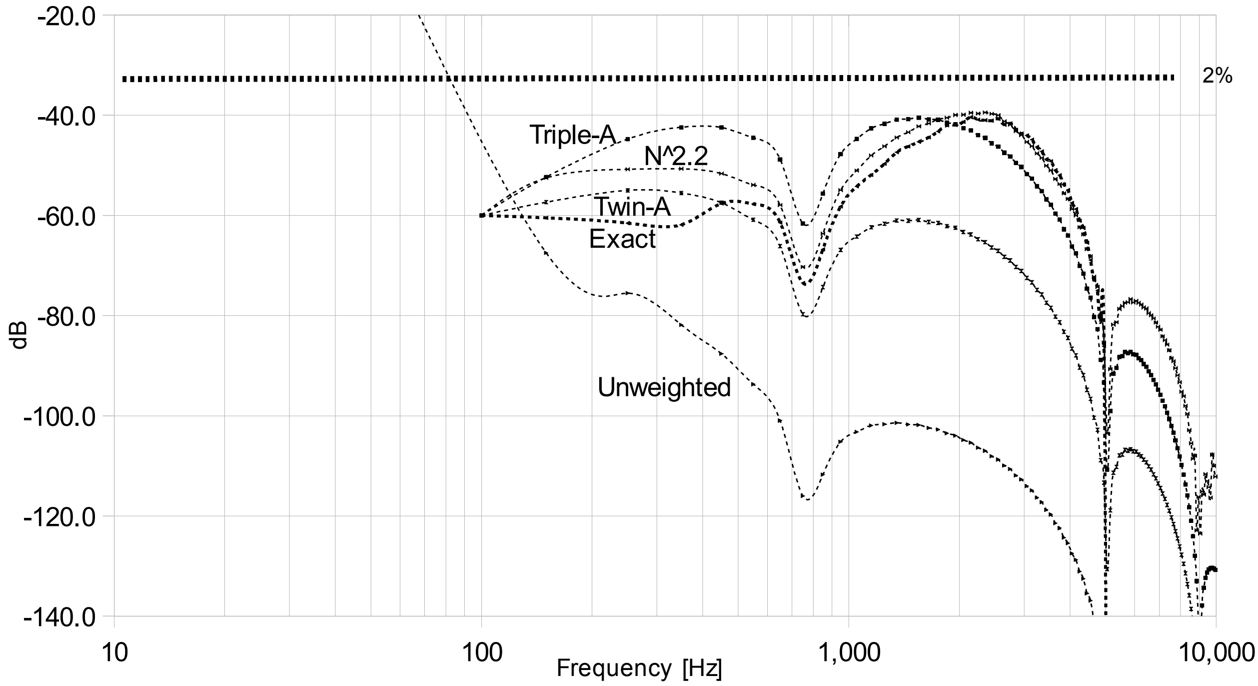

Figure 9. Voltage Follower distortion spectra for 80W and 50Hz; unweighed (lower), twin-A (mid), exact weighting (bold) and $(n/2)^{2.2}$ (fine) and triple A-weighting filters (upper).

The triple A-weighting filter gives a reasonably good approximation to the ideal weighting above the 15th (750Hz) but over-weighs the 5th to 11th harmonics by about 10dB. The RMS values for the triple filter is 3.53% and the exact weighting is 3.46%. The highest exact weighted harmonic is 0.95% at 2.15kHz, the 49th harmonic of 50Hz.

Notice how the ideal and triple-A boost the high-order harmonics in the 20th to 60th range by 10-20dB above the lower order harmonics. Previously we assumed these harmonics disappearing from view and could be ignored since they were below 0.001% (-100dB). But here they are prominent and close to the 2% threshold so they can accumulate beyond the 2% threshold and make the distortion audible! We knew crossover distortion was pernicious but now with weighting we can see how incredibly sensitive our hearing actually is to high-order distortion such as crossover distortion.

The high-order harmonics in the 20th to 60th range can be grouped into 1/3rd octave bands. Between the 20th and 40th there are 3 bands and with 10 odd harmonics there are about 3 harmonics per band. When RMS summed the accumulated value roughly doubles and this could push the bands above the threshold. Here 2 times the highest exact weighted harmonic is 1.9%.

The simpler RMS summing of *all* the weighted harmonics for the ideal filter accumulates to 3.46% or 3.6 times more than the largest weighted harmonic of 0.96%. An accumulation factor of 3.6 compared to the estimated 2 suggests the RMS sum of all harmonics is still good enough for simulations and bench tests. True band accumulation requires a complicated post processing algorithm whereas the weighted RMS for all the harmonics is a much easier and more convenient measure.

Figure 10 shows Voltage Follower distortion spectra for 80W for a twin 468 filter compared to a triple A-weighting filters and the exact weightings. The gain settings in Figure 7 for E2 and E3 for normalisation are 400 and 7500 respectively.

The low order harmonics are better weighted than a triple A filter but under weighted for the high-order harmonics. A twin-A filter weights low-order harmonics as well as a twin-468 but the twin-

468 weights the high-order harmonics better than a triple A filter. The twin 468 THD is 0.92%[3] which is close enough to 1.9% to still be useful. The triple-A appears to be a slightly better overall than a twin-468 filter for high-order distortion such as crossover distortion in bipolar output stages but appears to overweight most of the harmonics by a factor of 2 to 3 (6-10dB).

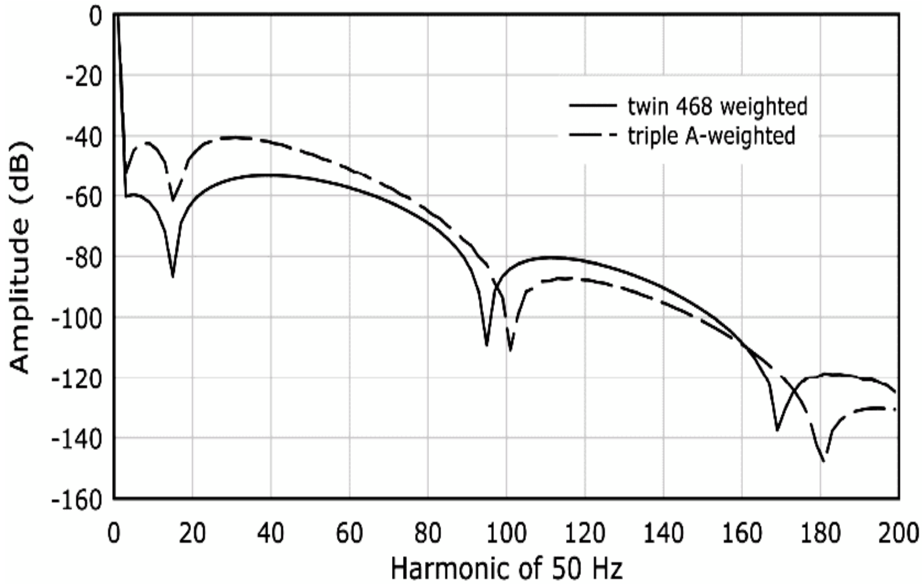

Figure 10. Voltage Follower distortion spectra for 80W for 5 case; unweighed (lower), Twin-486 (mid), Exact weighting (bold), and Triple A-weighting filters (upper).[4]

In this Class-AB example the unweighed THD is 0.046%. Similar unweighed and simulated THD values have been reported by Douglas Self's EF output stage of 0.03% with optimum biasing [33]. But the correctly weighted value is about 40 times higher and at 1.9% we are at the threshold for audibility for the average listener.

Time-domain waveforms in the crossover region where the distortion becomes visible are shown in **Figure 11** . The crossover region shows a main distortion component in the 1-2kHz range – harmonics around the 20th to 40th range. This is seen in Figure 9 with weighted harmonics peaking around 1.5kHz. Someone reported that low-order distortion is just-visible in an oscilloscope waveform when it is just audible. However, high-order type distortion can be clearly audible yet invisible on an oscilloscope. Weighting now fixes this, making any harmonic structure just-visible when it is just-audible.

3 LA was 1.3% and corrected to 0.92% after an was error spotted by Burkhard Vogel in Fig 5 circuit.
4 Now show the correct plot [ref Linear Audio, Letters, Dec 2012 view here.

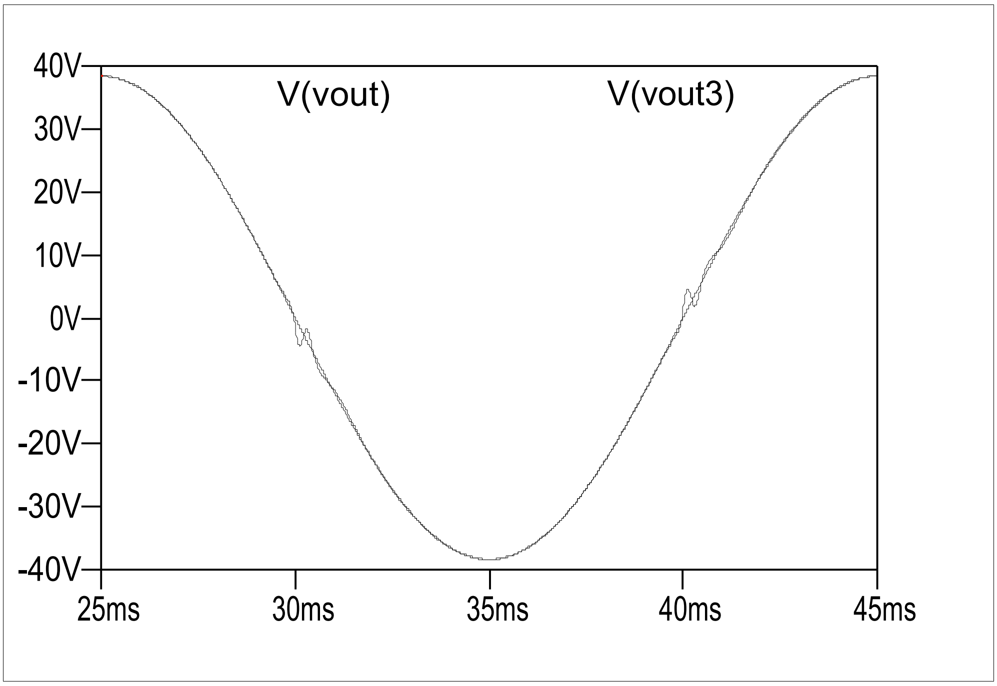

Figure 11. Voltage Follower output (0.046% THD) crossover distortion becomes visible with triple A-weighting (Vout3), here it is 1.6%.

8. Class-AB with lower-order distortion

Figure 12 shows a bipolar output stage similar to Figure 7 for 80W into 8 ohms but configured as a Common Emitter (CE) amplifier. This allows us to look at the inherent distortion of an output stage before local negative feedback is applied. The aim is to see the effect of reducing the emitter resistors on the internal distortion by using weighted filters and FFT spectral analysis.

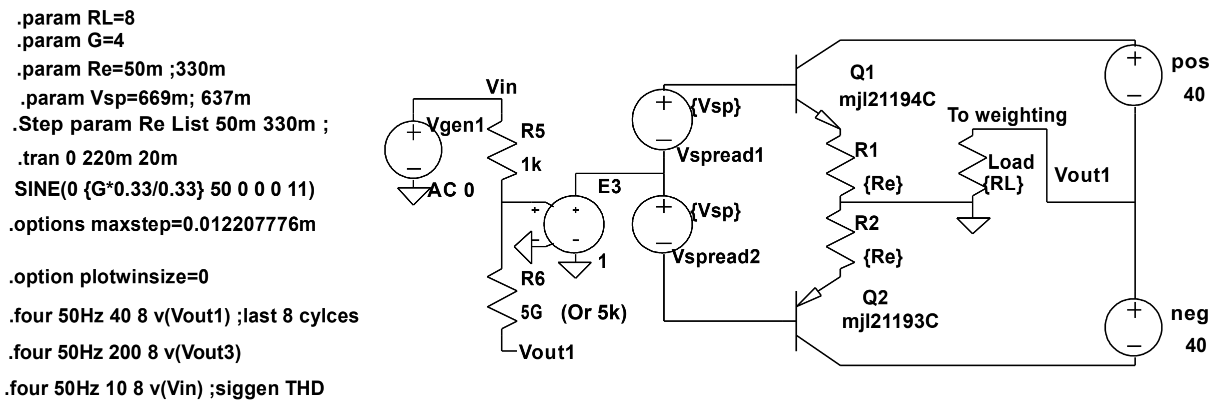

Fig 12. Common emitter output stage to examine weighted distortion with low value Re's.

I noticed someone on the DIY audio forum asked how a Common Emitter floating supply output stage provides voltage gain [38]. Douglas Self's article 'Common-emitter power amplifiers: a different perception?' [39] provides an explanation, or you can build the simple Voltage Follower stage presented below and experiment with it as a Common Emitter stage by moving the common from one side of the load to the other.

Basically the Common Emitter floating supply output stage is the push-pull form of the simple single transistor common emitter Class-A amplifier which provides voltage gain. In the single stage version one side of the load is usually connected to the collector and the other to the supply rail, but in the push-pull version split supplies are connected to the collector and the load is moved to the centre connection of the power supply. The total voltage gain is now the sum of the two transistors individual gains.

In Figure 12 the gain is inverting so shunt feedback can be applied using R5 and R6. R6 is effectively removed for my first part (set very high) so the open loop distortion can be viewed but later R6 is reduced until the weighted THD is less than 1.5%. The weighting filters use voltage source E2 to mirror the gain of the output stage, being approximately -Rl/Re (inverting) with no feedback (R6 large) and Re is the emitter resistance for each transistor.

Three values for Re are stepped in the first simulation run using 330mΩ, 100mΩ and 50mΩ emitter resistances. We want about the same output swing of 36 volts peak but the gain varies with Re so the signal generator voltage is also stepped using a function {G*(Re+10m)/(0.33+10m)} for the Amplitude value, where G is the baseline amplitude for 330mR emitter resistors where G~Vpk/(Rl/Re) or 2 in this circuit for full output swing. The quiescent current needs slightly higher V_{spread} when Re is reduced so the variable {Vsp+1.6m/Re-5m} is used to vary each V_{spread} voltage. This notation allows the input level to be stepped during a run while keeping the output power constant and quiescent current optimum.

Figure 13 shows the harmonic spectra for the unweighed 50mΩ case and the weighted harmonics with both 330mR and 50mR emitter resistors and at full power. Notice all the high-order harmonics above 1kHz are eliminated with a low emitter resistance. This improvement is obtained without using any negative feedback. The THD with 50mΩ and 330mΩ are 2.1% and 1.2% respectively, and w-THD with 50mR and 330mR are 22% and 77% respectively. Quiescent currents are 76mA with 330mΩ and 312mA with 50mΩ.

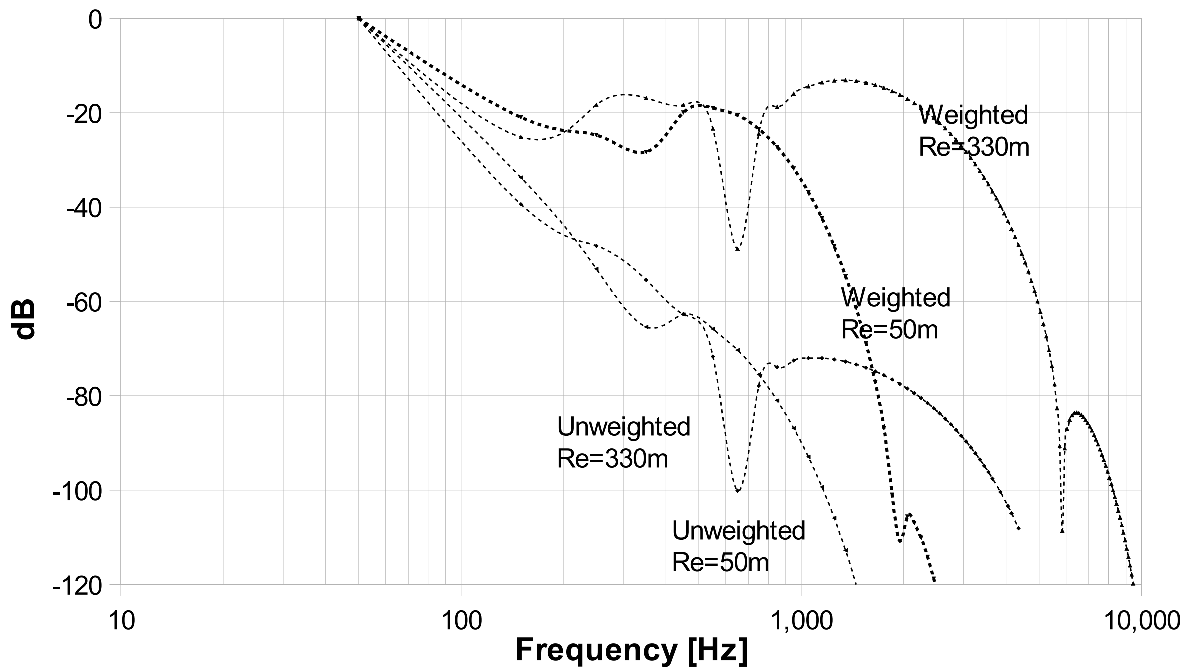

Figure 13. Open loop Common Emitter harmonic spectra for unweighed and Weighted harmonics for Re 330mΩ and 50mΩ and all at full power. Note a low emitter resistance eliminates the high-order harmonics above 1kHz.

The next step is to add some shunt voltage negative feedback to reduce the weighted THD figure to say 1.5% at full power when using 50mΩ emitter resistance. **Figure 14** shows the weighted distortion at full power and 1/10th full power with 13dB of negative feedback to achieve a target of 1.5% closed loop weighted distortion figure. The accumulated weighted THD is 1.6% at full power and the largest harmonic is the 9th at 0.4% and falls to 0.1% at 1/10th full power. The unweighed THD at full power is 0.1%. The output resistance is a satisfactory 0.3 ohms giving a damping factor of around 25.

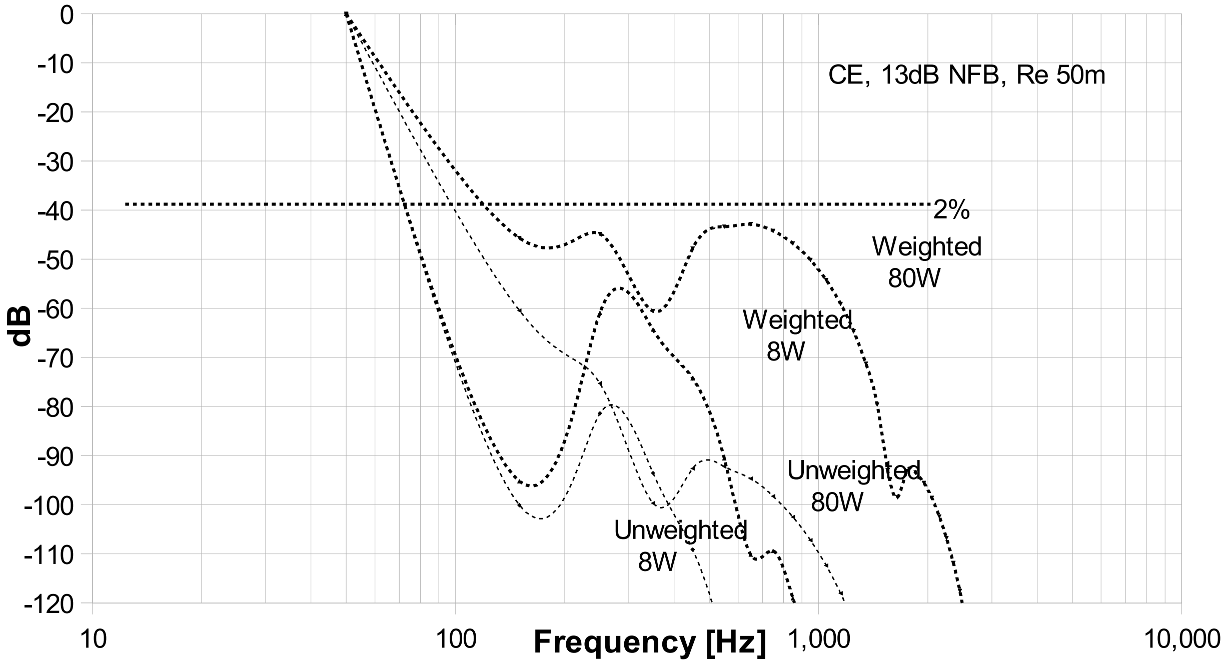

Figure 14. Closed loop Common Emitter distortion using 13dB of negative feedback (full and 1/10th power). Weighted distortion figures (thick lines) are below 2%.

This approach appears to solve the long-standing problem of high-order crossover distortion in Class-AB output stages while using low levels of negative feedback and without compromising loudspeaker damping. If a practical amplifier can be made using this topology then it would supersede my Square-law Class-A design since distortion at half full power is similar but power dissipation is a factor of 4 lower and the damping factor is very good.

To make a practical amplifier based on Figure 12 the output stage needs to be driven by a preamplifier. Since the output stage requires 8V peak (5V rms) a standard opamp preamplifier can be used. The power transistors need additional current gain such as Darlington pairs or a high current opamp buffer for E2.

An interesting option with this topology is R6 can be a linear potentiometer for the volume control. Series resistor R5 ensures there is enough feedback at full volume setting. It has the advantage of reducing the hum and noise at low volume settings but requires attention to loop stability at minimum volume since the output is connected directly to the input. This arrangement was bench tested with excellent results but it is not covered in this article.

9a. My A-weighting bench test jig [footnote[5]]

Tests were carried out on a two transistor Voltage Follower output stage shown in **Figure 15a**. ThermalTrak™ transistors NJL3281D (NPN) and NJL1302D (PNP) [Onsemi, Brusier – see Further reading] are used because I wanted to check whether the integral diodes eliminates thermal problem as claimed. I avoided using Darlington driver transistors so distortion is only from the power transistor.

Bootstrapping is used to increase the input impedance and reduce the amount of power needed to

5 This section was omitted from the Linear Audio article due to safety concerns with mains voltages. Take care!

drive the circuit. It requires 21V rms input at 24mA or 500mW. The input resistance is 880 ohms with betas of 100 and 110. Clipping started at 20.6V rms input with 19.6V rms across the 8 ohm load.

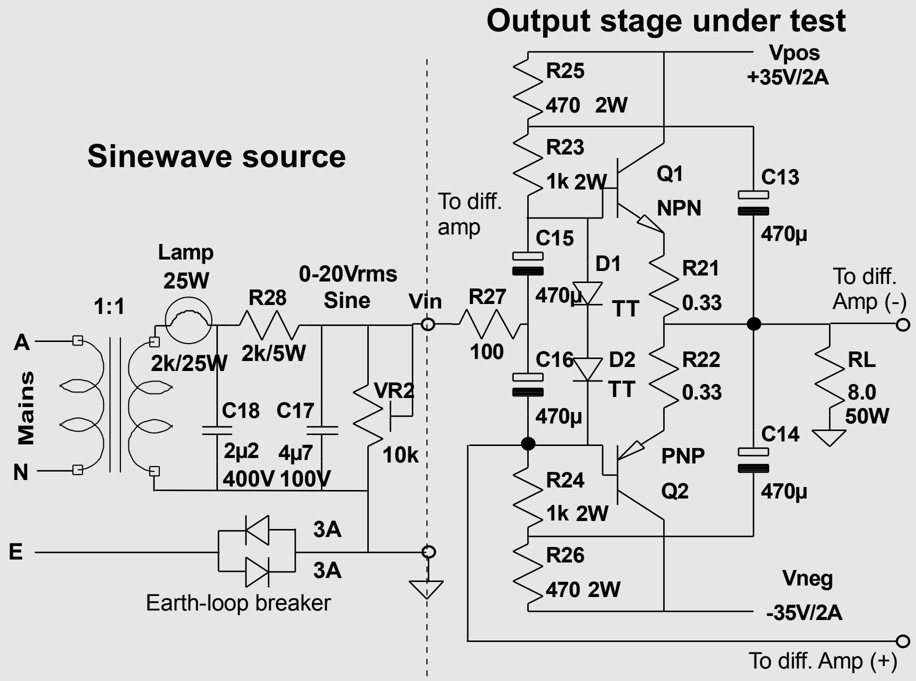

Figure 15a. A ThermalTrak™ Voltage Follower jig using a mains derived 20V sinewave.

I derived a sufficiently low distortion signal from 240V 50Hz mains using a passive low-pass filter. At the time of these tests I did not have a low distortion signal generator or a power amp ready for at least 20Vrms. A simulation of this filter showed the mains distortion (with a typical chopped off top) is reduced to around 0.1% and with the Total Difference method this is adequate for measuring the distortion of this output stage which gives distortion down to 0.05%. An isolating transformer is used for safety reasons and this can be easily constructed using two identical step-down transformers back-to-back of at least 20VA rating. The lamp could be replaced by a 25 watt 2k2 resistor.

There are lots of other options for deriving a low distortion 50Hz mains source such as a 4th order active filter fed by a low voltage transformer winding and this should give 0.01% THD for feed a standard power amplifier with gain.

The input level is varied by VR2 which should be rated for 0.5 watts. A cement trim pot is adequate. A 100 ohm input resistor is added to stop oscillations when the wiper is at minimum and the usual Zobel RC output network is not needed with a dummy load. A diode earth loop breaker is also included.

Biasing resistors R23-R26 provide 25mA through the ThermalTrak™ diodes D1,2. The quiescent current trimmed downward by adding a 500 ohm trimpot across one of the bias diodes to reduce the quiescent current (not shown). To increase the current a 10A Schottky diode is added in series with D1,2 and a 100 ohm trimpot across the Schottky diode. This combination allowed the optimum quiescent current to be found. The untrimmed quiescent current is 83mA after warm up with 0.33 ohm emitter resistors. It is close the optimum quiescent current of 73mA which required 100 ohms to reduce the quiescent current.

Interestingly, the ThermalTrak™ diodes do not completely cancel the +2.1 mV/ºC temperature coefficient of the transistors [40] so the quiescent current starts from 78mA at turn on and rises to

83mA at idle after 1 hour. When operated at 11Vrms for 1 hour the heatsink reaches its maximum temperature and the quiescent current starts to fall from 103mA to 97mA after 5 seconds and eventually returns to 83mA. During these tests the ambient temperature was 18ºC, the idle heatsink temperature was 22ºC and the hottest was 44ºC.

The temperature coefficient of ThermalTrak™ diodes is -1.7 mV/ºC and the transistor is +2.1 mV/ºC so there is a shortfall of 0.4mV/ºC if the quiescent current is to remain constant. However, this residual temperature coefficient did not lead to thermal runaway with 0.33Ω emitter resistors (or 0.1Ω resistors). Insufficient compensation was only noticed after these tests were completed that the OnSemi application note [Brusier – see Further reading] makes use of a string of 6 tracking diodes with a Darlington output stage with 4 base-emitter junctions, or 1½ tracking diodes per transistor to give the required temperature compensation.

9. Distortion tests

Figure 16 shows a setup to obtain the distortion residue using total difference method [23,36-39] with a differential opamp stage. The unweighed THD is calculated from the residual voltage divided by the voltage when the link is opened. When the link is open the reading is given by R4/R3 or 30% of Vout in this case.

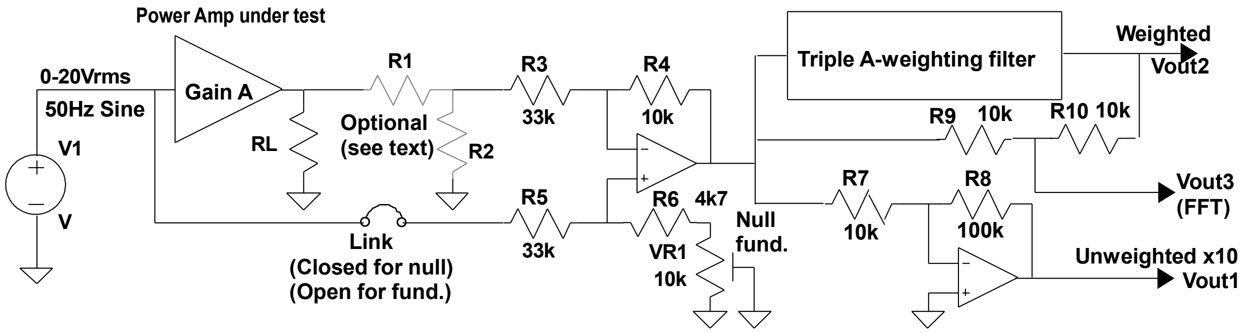

Figure 15. This A-weighting filter can measure unweighed and weighted THD's.

The residual distortion is then filtered and the output measured. The weighted-THD value is the ratio of the filter output voltage divided by the output voltage and multiplied by a factor to normalise the 2nd harmonic. If you want to spectrum analyse the weighted distortion then the fundamental is restored using summing resistors R9, R10 (with the link open). With two resistors and a high impedance load the output is -3dB but the harmonics percentages are unchanged. For low impedance loads an I-V (transresistance) opamp can be used as a buffer.

Both twin-A and triple A-weighting filters were used. The circuit values are the same as Figure 3 except that a non-inverting x11 opamp buffer is used between each stage [31,40]. I used a 1k feedback resistor and 10k divider resistor for each buffer and a TL074 quad opamp. This gives enough voltage gain at the triple-A output to be measured using an ordinary 200mV AC voltmeter. For the twin-A output the x10 opamp can be used. I used both a true RMS meter and a peak averaging meter and found that on average the twin-A and triple-A residual RMS voltages were 1.8 to 2.2 times higher than the averaging meter which is due to very peaky residual waveforms. If you have a standard averaging meter then multiply the readings by a factor of 2. The unweighed residual does not need scaling.

The scale factors for twin-A weighting is $SF_{2A} = 86*1.64^{\wedge 2}/11^{\wedge 2} = 1.9$. For the triple-A weighting it is $SF_{3A} = 800*1.64^{\wedge 3}/11^{\wedge 3} = 2.66$. These derive from the relative gains of simulations using Figure 3 (x1.64) and bench tests with Figure 16 (x11) opamps. The factors 86 and 400 are the inverse of the filters attenuation at 100Hz (Figure 4).

If you want to measure amplifiers with more than 20V rms output (30V peak) then increase R3, R5 because the differential opamp cannot have more than ±15V at the input terminals with 15V supply rails. An easy option is to add another 33k resistor in series with each of R3 and R5. Another option if you only want to measure the triple-A weighted distortion (no unweighed THD) then you can omit the differential amp stage and feed the power amplifier output directly into the first A-weighting filter stage. Figure 4 shows the triple-A output is about -90dB at 50Hz or 0.5mV with 20V input and residual readings were mostly over 30mV. But the null is needed for accurate twin-A readings.

The simplified triple A-weighting version without the differential amp can be made even simpler if you incorporate the scale factor (2.66) into an opamp's gain. Without the diff-amp the weighting filters gain is 3.3 times higher so you can drop the 2.66 scale factor if you also reduce the gain of one opamp by 20%. Now you simply divide the filter output by the amp output and times 100 to get the weighed-THD as a percentage. There's no nulling now. You can change the last opamp resistors from 1k/10k to 1k2/10k to achieve this.

It is helpful to view the distortion waveform on an oscilloscope. With an oscilloscope you get a better idea of what generates the majority of the distortion and if there is too much hum getting into the residual signal path. For lower output levels the hum and noise were RMS subtracted from my readings. This is measured with no input signal.

Another helpful tip is to measure the output stage voltage gain when the emitter resistors are shorted. This is safe for ThermalTrak™ transistors but may not be possible with standard transistors. This allows the parasitic base and emitter resistances and the bias diode string dynamic resistances to be found and this is important for simulations. Initially my simulations used 330mR resistors and when I ran my tests I found significant differences particularly when I reduced the emitter resistors to 100mR. When I measured and added the parasitic resistance the simulations results were closer to bench tests.

The parasitic resistance is calculated from the full swing voltage gain using $Rep=(1-Av)/R_L$ for a Voltage Follower. In my case the gain is 0.985 with no emitter resistors giving 120mR of parasitic resistance. For standard transistors where it is probably not safe to short the emitter resistors then plot the gain against 3 safe emitter resistances and the y-intercept is the zero resistance gain.

Figure 17 shows the Voltage Follower output into 8 ohms for 19V rms output and the standard unweighed distortion waveform residual. The output waveform is provided so the zero cross region can be seen. The figure of 19V is used because this is slightly below clipping. **Figure 18** shows the triple A-weighted residual distortion. **Figure 19** Shows the restored weighted signal that can be used for FFT. This is similar to the simulated waveform (Figure 11).

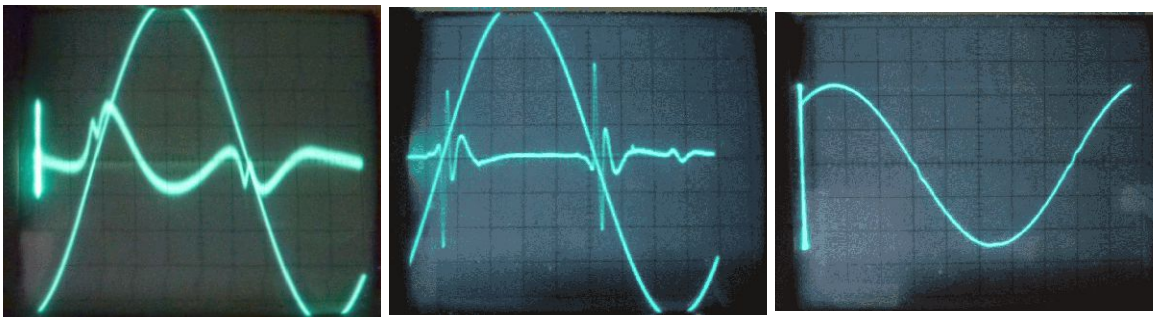

Figure 17. VF output 28Vpk and the unweighed residual (Left, 5mV/div). …

Figure 18. triple-A-Weighted distortion residual (200mV/div), output (5V/div).
Figure 19. Weighted signal with restored fundamental (right): note the visible crossover distortion. Voltage is halved by the summing resistors (now 10V/div)

Table 2 shows the distortion test result at full power and 1/10th power. The unweighed THD increases 1.6 times at the lower power level which is normally observed due to the same distortion in the crossover range and less power in the large signal region.

The triple A-Weighted distortions at full power and 1/10th full power shows the weighted THD stays the same or reduces slightly when the power is reduced to 1/10th (depending on the emitter resistance) whereas unweighed THD increases. The increasing THD with reducing power can be regarded as plainly wrong as an indicator of what we hear. The weighted figures give a better indication of what we hear when masking is not occurring which is the case with a 50Hz tone here.

Notice the triple A-Weighted distortion is lowered significantly when the emitter resistance is lowered. The weighted THD for the simulation was 3.5% at full power and with 330mR emitter resistors. The corresponding test is with 0.22Ω resistors since there is 120mR parasitic resistance and the measured value is 4.5%. Overall the tests results are slightly higher than the simulations but the same trends can be seen. The bench test used lower supply rails and only allowed 45W rms just before clipping where the simulation was done at 80W just before clipping. Also the bench results included the transistors second order effects which normally increase distortion figures.

Notice the triple-A-weighted distortion reduces as the emitter resistance falls. This confirms the usefulness of aiming for as low as possible emitter resistance values. The parasitic resistance (120mΩ here) limits the weighted distortion to around 1% per pair, but paralleling several stages can reduce the distortion further. The number of parallel pairs is limited either by cost or by the quiescent dissipation.

	Unweighted 0.33Ω	**Twin-A 0.33Ω**	**Triple-A 0.33Ω**	**Triple-A 0.22Ω**	**Triple-A 0.10Ω**
Residual 19Vout	2.6mV	20mV	123mV	97mV	37mV
Residual 6.5Vout	1.3mV	-	42mV	28mV	8mV
THD 19Vout	**0.045%**	**0.66%**	**5.6%**	**4.5%**	**1.7%**
THD 6.5Vout	**0.072%**	**-**	**5.6%**	**3.8%**	**1.1%**

Table 2. Distortion test results

10 Listening tests

I was interested to know whether I could hear the distortion for the Voltage Follower. According to the simulations above the largest triple weighted harmonics are 0.9% (49th) and 0.7% (15th) which shouldn't be audible being slightly below the 2% threshold after applying the estimated factor of 2 for accumulation. However, the simulated RMS accumulated value is 3.5% which is over the 2% threshold and this suggests distortion from the Voltage Follower should be just audible. Can it be heard? It seems not according to my ears.

I used a pair of sealing earphones with a series resistor to give a sound level of 100dB for 19V output. I first checked for optimum bias at the current heatsink temperature which required a slightly lower current of 51mA. With optimum bias I could not hear any particular harmonic tones. When I reduced the quiescent current from 51mA to 44mA so that I could just notice a change in the sound, the triple weighted THD increased from 4.5% at optimum bias to 6.0%, an increase of 1.3 times. The frequency now heard was 750Hz, the 15th harmonic. This shows that the 6% RMS accumulated reading is slightly over-weighted or over-accumulated.

Simulations also suggested that 1.8 times over-accumulation occurs with RMS summing, so removing this still leaves an over-weighting by a factor of about 2 for my hearing tests. I checked my hearing threshold for the second harmonic of 50Hz at 100dB and found it is within the normal hearing range. Part of the remaining factor of 2 is due to second-order effects present in the test circuit and other differences that are not included in the simulations. A simulation to check the Early effect suggests distortion does not increase by more than 1/5th with this included. It will be interesting to hear whether others find that there is an overweighting occurring with triple filters.

The twin A-weighting gives under-weighting by a factor of about 2-3 and could be used with a suitable correction factor. The correction factor should increase the filter's 'boost', the ratio w-THD/u-THD and not just scaling the w-THD. This ensures the 2nd harmonic remains the same. For example, using the figures in Table 2 (for 340mR net) the filter's 'boost' is 0.63%/0.045% or 14 times, so applying a correction factor of 2.4 to the twin-A readings will give the required 1.9% value found above. (It is also possible to apply a correction factor to a single A-weighting filter but the error can be quite large though it may still be useful for ballpark tests if you only have one A-weighting filter available).

When I reduced the emitter resistors to 100m ohms and checked the quiescent current was optimum, the main harmonic heard was the 7th (after under-biasing to make distortion just audible again). This is the same trend shown in simulations (Figure 13) where reducing the emitter resistance moves the dominant high-order harmonics down close to the less objectionable low-order harmonics.

My test also showed that the Voltage Follower distortion at optimum bias was slightly below audibility for my hearing to detect any outstanding harmonic. I had no way to check whether the accumulated harmonics just below the audibility were still detectable. Detecting subthreshold distortion will affect how much lower than the 2% threshold we need to go to make distortion undetectable. Is a factor of 20, for 0.1% weighed distortion, sufficient as Robert Stuart suggested?

Summing up

What's new? Using three A-weighting filters in cascade to evaluate amplifier distortion can generate distortion numbers that correlate much better with listening tests than a regular all-in-one THD measurement. The method is easy and can be used in simulations and bench tests. A 50Hz tone seems to be the best choice for detecting the presence of crossover distortion. Designers can now see how good an amplifiers is likely to sound before they do bench tests and listening tests. One interesting example given where weighting predicts that crossover distortion can be reduced is the Common Emitter output stage by employing very low value emitter resistors and paralleled output transistors. It is hoped that many designers will evaluate weighted-THD figures alongside conventional THD and alternative weighting approaches for a wide variety of amplifier types to confirm which measurement gives the most consistently relevant results.

Acknowledgements

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Didden's perseverance in getting our material in a form all can understand and diagrams we can read is very much appreciated☺.

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- Douglas Self, 'Night Thoughts', Electronics World, Nov 1996 p858-863. Fig. 7 shows Voltage Follower gain plots as Re is varied from 0.47Ω to 0.10Ω.
- OnSemi datasheet for NJL3281D, NJL1302D, 2006.
- Mark Busier, 'AND8196/D ThermalTrak™ Audio Output Transistors', OnSemi App Note for NJL3281D & NJL1302D, 2005. http://www.onsemi.com/pub_link/Collateral/AND8196-D.PDF

How to download the subcircuits for LTspice simulations

Go to http://www.cordellaudio.com/book/tutorial_simulations.shtml and download the zip file SPICE_Tutorial_Examples.zip and unzip and the AWF.asy and AWF.sub files are found in the directory: SPICE Tutorial Examples\Figure 19_2\A Weighting Filter\. Copy these files to your LTSpice working directory. To check them Copy the files A Filter Subcircuit.asc and A Filter Subcircuit.plt to your working directory and run "A Filter Subcircuit.asc" and you should see a plot of the filter V(vout). You can now copy the Filter into your circuit. Open your circuit and then click on the Filter (in the other circuit), Duplicate (F6), click on the filter icon, move the mouse to your circuit area, zoom out if necessary, click on a clear space to place the filter. You can now use the AWF filter in your circuit. You can edit an AWF file by double clicking the AWF.sub file and LTSpice will open an editor. Save the changes and keep an original copy of the AWF.sub somewhere. Bob Cordell's book provides helpful information on creating subcircuits.

To triple A-weight the output signal form an amplifier insert three AWF's in series, like Figure 8. If you only want the triple weighted output then you do not need the mirror bank. Add a controlled voltage source ("E" source) with a gain of 800 and the divider R3, R4 (if you want the weighted 'restored' signal to be the same amplitude as the output signal). For the Common Emitter circuit (Figure 12) you need to boost Vin using a controlled voltage source with the same gain as the output stage and inverting. Add the FFT directive to obtain a table in the Error Log, eg, using .four 50Hz 200 8 v(Vout3). You can view an FFT plot without this directive. To get a low 'floor' include `.options maxstep=0.012208m` where this value is calculated from Tstop/16,383 and Tstop is the "Stop Time" set in the .tran directive. Also set `.option plotwinsize=0` to turn off compression and check in the FFT setting that "Number of Data Point Samples" is sufficient (eg 262,144). A 'floor' of -150 to -180dB was obtained with these tweaks.

[Edit: Since 2018 it can be downloaded from www.paklaunchsite.jimdofree.com]

Equation solution for the CE output stage

When I did my course, lecturers said that it is not possible to obtain an exact analytical solution for the current Id and the voltage Vd of the exponential model for the diode in our circuits. And this included all bipolar transistor circuits because the bipolar transistor is considered to operate as an amplified diode! So until recently, only small signal solutions or numerical simulations could be used to design analogue circuits.

But thanks to Thomas Banwell [41] and others we now have an exact analytical solution for diode and transistors in analogue circuits using the little know Lambert W-function. I have applied this function to find equations for push-pull bipolar output stages at low frequencies. The simple Class-AB output stage with voltage drive is the easiest and allows an exact solution if all second-order effects are neglected (Early effect, Beta-fall, base spreading resistance and capacitances).

The Voltage Follower is treated as a Common Emitter stage with 100% voltage feedback. The Common Emitter stage is shown in **Figure 12**. The input voltage is fed by E3 to the bias spreader voltage sources and to Q1 and Q2 bases. The emitters of Q1 and Q2 are connected to common via emitter resistors R1 and R2 (with values of Re ohms). The collectors of Q1 and Q2 deliver current to the load via the power supply rails using a floating power supply. The load current is Ic1+(-Ic2) where Ic1 and Ic2 are currents through Q1 (NPN) and Q2 (PNP). The current through the load gives a negative voltage when Vin is positive, meaning the circuit is inverting.

Using Banwell's approach the currents I1 and I2 can be found for a symmetrical circuit. This gives:

$Ic1=\frac{Vt}{ReTot}\,W\left(\exp\left(\frac{Vsp-Vk+Vin}{Vt}\right)\right)$ where $Vk=Vt\,\ln\left(\frac{Vt}{Is\cdot \mathrm{Re}Tot}\right)$ and

$W(x)\simeq\ln(1+x)\left(1-\frac{\ln(1+\ln(1+x))}{2+\ln(1+x)}\right)$ and $ReTot=Re+Rep+\frac{Rbp}{Beta}$ where Rep and Rbp are Q1's parasitic emitter and base resistances respectively. The solution for Ic2 (PNP) is similar but with Vin is inverted:

$Ic2=\frac{-Vt}{Re_{Tot}}\,W\left(\exp\left(\frac{Vsp-Vk-Vin}{Vt}\right)\right)$. Vt is the thermal voltage of 26mV at 27°C or 300K. The function W(x) is the Lambert W-function and the equation given above is an approximation that gives about 1% accuracy.

Figure 19 shows the wingspread gain plots generated by the above equations when run concurrent with a symmetrical basic CE circuit. The LTspice circuit gm is plotted as d(I(RL1)) and the other equation solution I(B(21)) for the gain is very close, to within 1% and is limited only by the Lambert W(x) approximation used here. Vspreads are stepped from 480mV to 880mV in 100mV steps and for these the quiescent current varies from 0.46mA to 638mA. An intermediate value of 630mV is included to show 'optimum bias' (Iq=65mA). Settings were Beta's = 100, Is = 4x10^-12 and Re = 330mR (IKF and ISE were disabled using IKF=4k, ISE=1.2x10^-20).

The Lambert W-function approach to circuits also provides explicit gain equations for the first derivative as well as all the other derivatives. This allows the THD to be estimated. For the 2nd harmonic and 3rd harmonics HD2=1/6·(G1+G2) and HD3=1/24·(G1-G2) where G1 and G2 are the gains (gm) at the signal peaks [42,43].

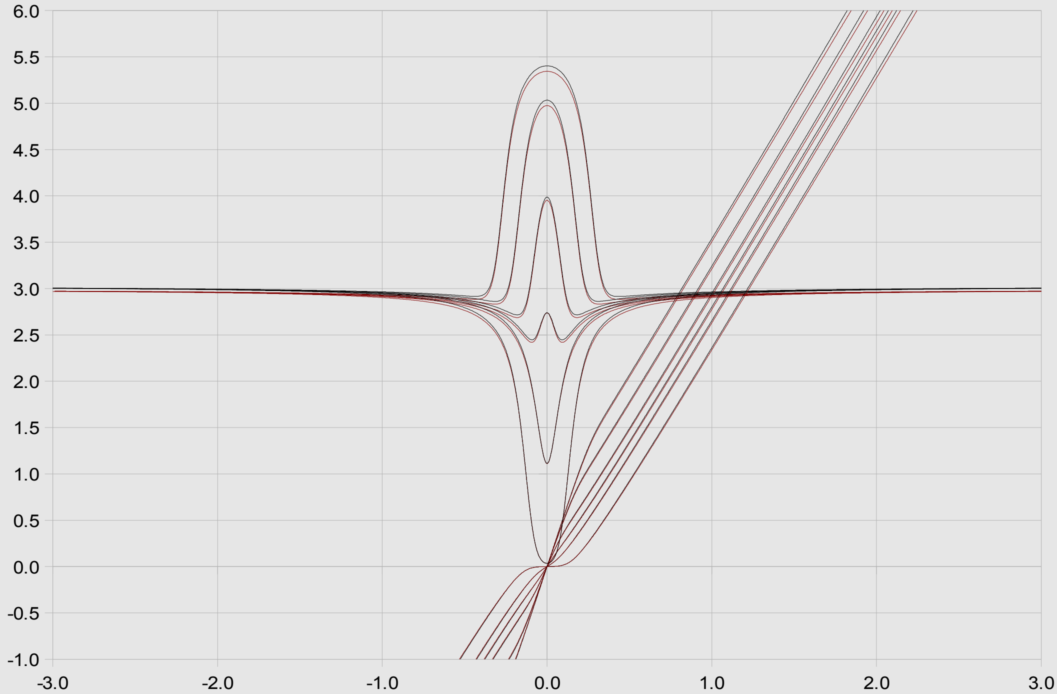


Figure 20. Wingspread gain plots generated by equations run concurrent with a CE circuit simulation. Small differences can be just seen. Load currents are also shown.

Crossover distortion in a symmetrical circuit for any bias condition can be estimated using

$$THD \approx SF \cdot \left| 1 - \frac{Gm_{pk} \cdot Vin_{pk}}{I_{pk}} \right|$$

where SF is a proportionality factor found from a simulation, Gm_{Pk} is the gain at Vin_{Pk} and I_{pk} is the load current at Vin_{Pk}. This method is based on the difference in current from the Gm curve and the current curve; if the gm curve is a flat line then Gm_{Pk} times Vin_{Pk} gives the same value as I_{pk}, but if the gm curve is not flat then there is a difference and the difference in energy is proportional to the THD value. This approximation goes to zero at optimum bias, where the area above the gm curve cancels with the area below the curve. The approximation can be improved by adding an offset and RMS summing to give a small finite THD at optimum bias.

The THD as a percentage at optimum bias is found to be around 1% independent of the emitter resistances (Re) [35]. The solution for Ic1 and Ic2 are given by I=k·W(Vin) where k varies with 1/Re and Lambert W-function W(Vin) gives the same family of curves when Re is varied and optimum bias is maintained. There is a small change in the quiescent current point (due to Vk) when Re is varied but for optimum bias we change Vspread to maintain an optimum bias condition. The same family of curves with optimum bias as Re is varied means that changing Re cannot significantly change the percentage of distortion in a Common Emitter push-pull stage at optimum bias. But since lower Re means more gain (k is larger) the Voltage Follower distortion reduces when Re is reduced and this is entirely due to local negative feedback of the combined currents I1-I2.

A spreadsheet for calculating power amplifier distortion

A spreadsheet, PAK (Power Amplifier Kalculator) has been put together to allow distortion calculations for various power amplifier circuits to be run and to plot the curves like a SPICE simulator would. unweighed and weighted-THD's can be estimated reasonably accurately using available SPICE parameters.

The PAK spreadsheet is being launched as an open source project to allow others to add their own circuit solutions. PAK version 1.0 contains solutions for the Emitter Follower and CFP output stages, as well as the differential input stage, current sources, current mirrors, Vbe multiplier and the Common Emitter stage (for a VAS). With all these solutions it is now possible to find explicit low frequency equations for a complete 3-stage power amplifier, for example. Second order effects such as the Early effect and beta-fall can be included and is left for later versions as are HF solutions.

PAK 1.0 is open source, anyone can use it or modify it and distribute it under the Creative Commons licence. A User Guide for PAK Ver 1.0 explains how bipolar circuits are solved.

[Edit: Since 2018 it can be downloaded from www.paklaunchsite.jimdofree.com]

Letter to the Editor

Burkhard Vogel writes:

Dear editor,

I've studied Hans van Maanen's Guest Editorial and Ian Hegglun's interesting article in Linear Audio Volume 4 and very quickly stumbled over Ian's statement that "…. no method has become popular with power amplifier designers and reviewers so far".

His statement seemed to be true on the face of it. Generally, I like to study competing theories à la Kant's advice that "there is nothing more practical than a good theory" – and then Occam's Razor will take over. It motivated me to write this Letter.

I'm not an expert in distortion matters at all and I know only a tiny part of the world-wide existing literature on that specific issue in depth, mainly the British view on it. However, I also know that 27 years ago Mr Johannes Maier from Stuttgart, Germany, published an article about his theory on the correlation between listening test results and amplifier measurement results. Not until 24 years later would this theory lead to practical applications because first the corresponding measurement approach and equipment had to be developed and tested.

During the past 3 years it was Mr Peter Schüller, the head of Stuttgart, Germany, based TESTfactory, who invented the final test approach that allows qualifying an amplifier in a way that gives a clear signal to the user whether it is free of audible add-ons (tendency: short piece of wire with amplification) or less free (tendency: creates a specific sound that adds additional information to the original signal). This would be valid for any kind of output load - purely resistive as well as complex (like a loudspeaker simulation or a real loudspeaker). Or, in other words: the measurement approach allows qualifying whether an amplifier measurement result will fully correlate with the listening test result or only part wise or not.

Shortly and in my words, the most important content of the MST (Maier-Schüller-Theory) could thus be described as follows: In addition to some other and minor important points the structure of sequence, order, and level of the nonlinear harmonics of **all** frequencies in the whole audio band play a major role in the reconciliation of listening test results and amplifier measurement results. These results are displayed by a specific measurement tool and thus can easily be assessed.

This MST was firstly presented at the Munich 2012 High-End Fair's expert forum (in German), and some weeks later on the Burosch website (also in German). Burosch is a world-wide operating company that offers a broad range of test signals for all kinds of video and audio codec purposes; this company also runs many websites that offer an additional selection of white papers on various kinds of video and audio issues.

Unfortunately, the whole lecture cannot be presented in this Letter because the many slides are all in colour and cannot be changed into the Linear Audio black and white format. However, I could convince the two inventors to formulate the MST in a way that could be published in Linear Audio in the near future – including additional findings that could not be shown at the fair.

For readers who want to acquaint themselves immediately with MST I've arranged with Mr. Schüller to translate the text of his lecture into English; this can be found at:

http://burosch.de/audio-technik/509-high-end-2012-klang-2-english.html

One of the many advantages of MST is the fact that all measurements can be performed rather elegantly by application of a top measurement instrument like the AP 2722.

I also think that Hans van Maanen's concerns and doubts about the selection of a listening team can be scattered by the two inventor's experience (together with a well known group of other long-time experienced helping colleagues). A work life long they've performed hundreds of loud-speaker and amplifier listening tests and measurements. I guess these engineers clearly know what they are talking about and there is no need for them to learn how to listen in an anechoic chamber (they have one) or in many different kinds of living rooms, nor how to listen as a scientist or as an average person.

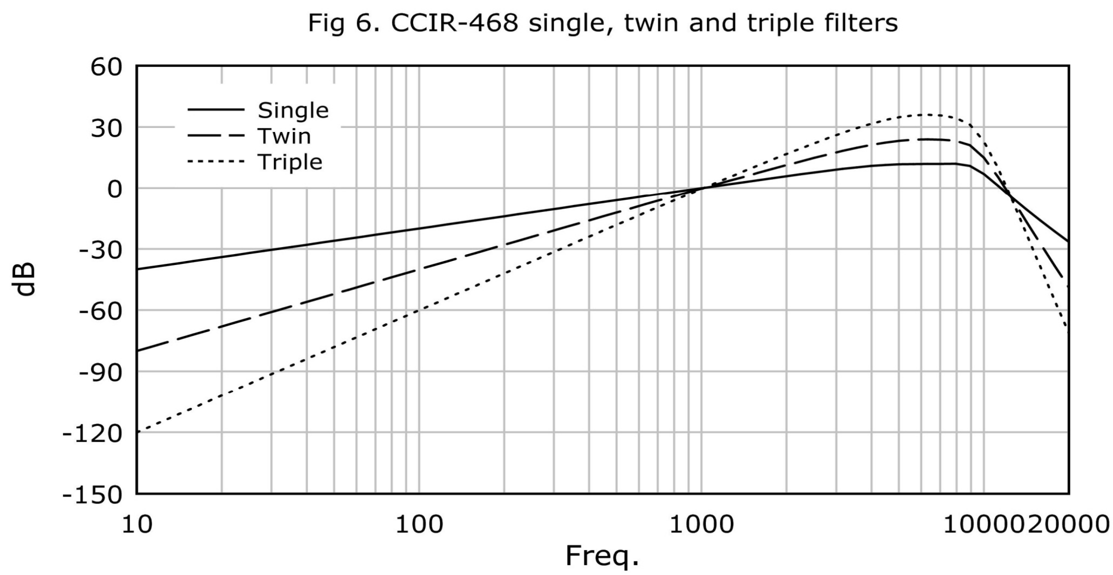
It is very rare to find audio equipment manufacturing-independent people in the audio test community that did not change their test company during 25 years, and as a result this company has collected a huge amount of knowledge about all kinds of audio issues. By diving very deep into the loudspeaker/amplifier problems, TESTfactory is highly recognized as independent inventor of new and challenging test methods - obviously only known in Germany, as proven by Ian's literature list.

Burkhard Vogel
Stuttgart, Germany

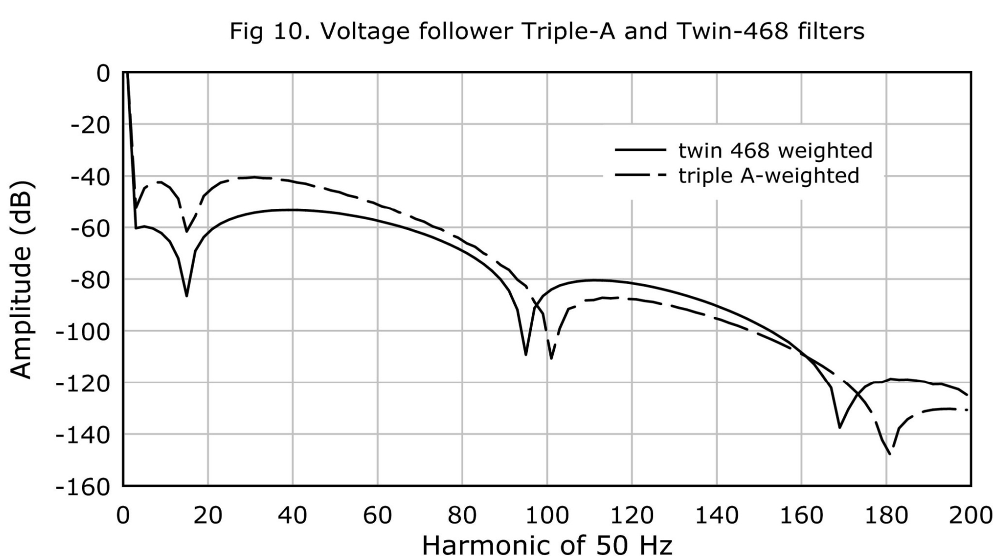
PS. I know very well the many traps of the sub-editing process. That's why I kindly ask Mr Hegglun to check again his Fig. 6. I think it's wrong because Fig. 5 is wrong. His source [31] will give the right CCIR Standard based component values. Another source is the ITU-T J.16 paper.

Ian Hegglun replies:

Thanks Burkhard for spotting my typo in Figure 5 circuit where L1 should be 12.88mH and C2 26.82nF. My apology. The corrected Figure 6 appears below.



This changes Figure 10 slightly. The corrected Figure 10 plot appears below:



One sentence on p43 needs a slight change: "...The twin 468 THD is 1.3% which is close enough to 1.9% to still be useful." should read "The twin 468 THD is 0.92% ..."

I consider the same conclusion can be made using the correct figure of 0.92% and the article is not significantly affected by this typo.

After pondering on whether it is better to use Twin-468 filters or Triple-A the Twin-468 filters consistently underweight all the harmonics by a factor of two whereas the Triple-A overweights only the first 12 harmonics (see added plot). This would allow a scale factor of two to be applied to the THD reading with Twin-468 filters to get close to an exact weighting and1/3rd less components to boot. Scaling the Triple-A reading is not recommended because harmonics above the 12th become underweighted.

On the body of Burkhard's letter - I was interested to read the translation of Peter Schüller's talk first presented at the Munich 2012 High-End Fair on the Johannes Maier and Peter Schüller Theory (MST) on the special distortion impact on amplifiers (sublink http://burosch.de/images/Schueller-lecture-03.pdf).

This raises the interesting question of what is the best harmonic structure for distortion if we cannot make it inaudible. The topic of euphonic distortion, which I did not mention in my article or references, is also a commonly raised.

An underlying assumption of my article is that a threshold exists, a level where harmonic distortion can no longer be heard. So a designer can use weighted THD measurements to try get the figure down to where it does not matter what the harmonic distortion spectra looks like. BTW Section 8 of my article showed how that can be done for a novel Class-AB topology.

I suggested a factor of safety be applied to ensure harmonic distortion is inaudible. We can still detect subtle differences even if the harmonics cannot be heard. Dr. Hans van Maanen in the guest editorial of Vol.4 mentions the masking thresholds used for MP3 are insufficient for many listeners. Robert Stuart had suggested a safety factor of around 20 and I was hoping more research will be carried out to arrive at a practical value for this for the audio industry.

How does this relate to MST? My approach and MST are mutually exclusive. If harmonic distortion can no longer be heard or detected then we do not need to be concerned with the harmonic structure of a power amplifier. Applying MST to amplifiers with distortion below the threshold of audibility may give us false negatives. It would be good to see more work in this area with alternative approaches.

The MST approach is mainly relevant to amplifiers where the distortion is high enough to be audible such as low or no negative feedback power amplifiers, to select or design for minimum unpleasant effects.

Interestingly, my article's reference 4 by Rupert Neve suggests that the 7th harmonic is the first harmonic number that needs to be made inaudible for a good amplifier because it is the first harmonic that does not form a standard musical interval. If Neve and MST are both correct then the 2nd, 3rd, 4th, 5th and 6th can exist at small levels with *minimal* deleterious effects provided these harmonics reduce monotonically and the 7th is below the threshold of detectability, say 20 times below the audible threshold. Maybe we can call this Neve+MST?

Neve+MST could explain why "...a sizeable segment of the audiophile population“ like SETs (Nelson Pass p77 Vol.4) – because the harmonic structure decreases monotonically and the 7^{th} harmonic is well below audibility with typical music material.

On euphonic distortion. The original work by Jean Hiraga on the harmonic structure of power amps was presented in English in the Hi-Fi News March and April 1977 and a critical review can be found in Stereophile "Euphonic Distortion: Naughty but Nice?" by Keith Howard (at http://www.stereophile.com/reference/406howard/index.html). Keith Howard points out that Jean Hiraga is claiming that particular patterns of distortion actually enhance fidelity, now called euphonic distortion. But is there any hard evidence available to support this claim?

Keith Howard provided various synthesized distorted files for listener tests on a web site that aimed to meet Hiraga's criteria. But positive test results have not been reported, at least that I am aware of. This does not prove that it can't be done so we are still left waiting for a positive result.

A positive result has profound implications. Some suggested that if a recipe can be found then euphonic distortion can be added using DSP to any amplifier and possibly at any stage of the reproduction chain. If euphonic distortion exists then amplifiers that are made 'bland,' because they generate no audible distortion for example, then they too can be brought back to 'life' again by adding euphonic distortion.

But in the mean time many designers including myself aim to design amplifiers that add no audible distortion, and as a fallback, to choose the best harmonic structure for the distortion that we cannot make inaudible.

Finally, I noticed in the Peter Schüller's talk that the AP analyzer 2722 appears to now remove the noise component from the THD plots so the distortion flattens out at low signal levels, eg p28 slide #27. Previously 'THD+noise' plots were generated that would rise at very low power giving a misleading indication of distortion at very low power. At last we can have distortion-only AP plots at low power levels in publications and reviews. Fantastic!

BTW in the Peter Schüller's talk p28, the text appears to have a typo in the English translation, where "k5 (red)" I think should read "k5 (blue)".

Ian Hegglun,
Australia

Hans van Maanen replies:

I have read the comments of Burkhard Vogel with great interest and I think that I have to clarify one of the statements in my Guest Editorial (Linear Audio Vol 4). When I made critical comments about the listening teams, I tried to elucidate that -in my view- gathering more or less randomly a number of people is insufficient to base far-reaching conclusions on e.g. the ability of human hearing. This did not exclude the possibility that very experienced listening teams do exist. But at e.g. AES conferences I often hear at presentations that the listening team was just a number of students and although such young people will mostly have good ears (although I fear the I-pod!), they usually lack experience. It took me tens of years with frequent (around 30 classical live concerts / year) visits to the Amsterdam Concertgebouw to get (I hope) a useful data-base of listening experiences. Including the problem that no existing audio system meets the same quality criteria as human hearing, I take the results of “scientific” listening test as described above with a large number of grains of salt. The MP-3 disaster, in which lossy compression techniques are used which were “scientifically proven” to be inaudible, is in my view a clear example of the incorrect conclusions, drawn from such listening tests.

I also read the presentation of Peter Schüller, referred to in the letter of Burkhard Vogel. Although I welcome any development to further improve our understanding of the relations between measurements and listening tests, I think the presented theory is incomplete and does not explain related problems. First of all, harmonic distortion does not come alone. As soon as a system generates harmonic distortion, it also produces e.g. intermodulation distortion.
And it is unclear (at least to the best of my knowledge) which is the worst of the two when it comes to audibility. I can imagine that when you play “simple” music (like the infamous jazz-trio), intermodulation distortion is not very disturbing or annoying compared to the harmonic distortion. But when it comes to reproducing the classical symphony orchestra, intermodulation easily leads to a “grey, misty” sound, in which details drown in the intermodulation distortion products. So the situation may then be completely opposite. This would require a lot more investigation before any final conclusions can be drawn. As far as my own experiences go, any further reduction of the amplifier's

distortion increases the audible presence of fine details in the reproduced sound. But I often listen to the classical symphony orchestra, which might be more susceptible to intermodulation distortion.

The statement that an amplifier sounds good “if and when it shows a regular and harmonic decline of its distortion spectrum” comes out of the blue. In general, such a decline is a pretty common property of distortion components: even a square wave shows this behavior. So I think this statements needs more detailed specifications to understand and to be underpinned. It also implies that the ADDITION of distortion components to acquire this property (so with a higher overall distortion figure!) would improve the sonic quality of an amplifier.
This contradicts my own experience: more distortion masks the reproduction of details as mentioned above. If the addition of distortion components would improve the sonic quality of an amplifier, it would be quite easy to build “good” amplifiers.

As a side remark: the theory of (negative) feedback stems from the twenties of the 20th century, long before the transistor was invented, so the suggestion that it was invented to improve the quality of transistor amplifiers is incorrect. It can be used to reduce the distortion products in (transistor) amplifiers, but it is by no means a “miracle cure” for all deficiencies of amplifiers.

Moving the feedback pick-up point “upstream” of the power amplifier transistors results in a signal strength dependent output impedance when the power transistors are not operating in class A. This will have an effect on the dynamic behavior of the amplifier (see also below).

The discussion about the harmonics is interesting, but it is unclear to me why the author stops at the fifth harmonic when “integral” distortions are compared (from sheet 31). Of course, I understand that the fifth harmonic of 20 kHz is outside the range of interest, but harmonics of low frequencies tend to end up in the most sensitive frequency range of human hearing, so they might be quite disturbing, especially when we are talking about harmonics which do not occur naturally in instruments. In my own experience, the presence of harmonics above the fifth, even in small amounts, often sounds annoying and tiring. Also, harmonics of harmonics may play a role (N.B. In some instruments, the upper harmonics are stronger than the first! And they get distorted too). So the actual distortion products can reach a lot further than one would expect at first sight when we look at the signals from real instruments.

Secondly, there seems to be a difference between “harmonic distortion” and “harmonic distortion”. Looking at the figures for harmonic distortion of solid-state amplifiers, valve amplifiers and loudspeaker units, one could easily conclude that the distortion of solid-state amplifiers is so low that it is completely inaudible when loudspeakers are used (but I don't know of a way to avoid these at this moment). Yet, it is not very hard to hear the effect of amplifier distortions which are several orders of magnitude smaller than those of loudspeakers. A similar remark can be made about valve vs. semiconductor amplifiers.
So there are still a lot of aspects which still are not understood and need further investigation as I don't think the theory as presented by Peter Schüller explains this.

We could start to make distortion measurements of amplifiers under more realistic conditions. The -in more than one sense- complex behavior of the loudspeaker impedance has its influence on the sonic behavior of the amplifier. A point the author does not address is the reaction of the feedback loop to the phase difference between the voltage and current when complex loads are used. The easiest way to understand this problem is to look at a zero-crossing of the output voltage. When

there is a phase difference between voltage and current, the feedback loop is forced to generate an error voltage in order to open one of the power transistors to provide the required current.
This is a distortion, which is not present when a pure Ohmic load would be used. This is the main reason why I use impedance compensation already for tens of years (see the reference below, also showing that the impedance can be a lot more constant than is shown in slide 41 of Peter Schüller's lecture). Therefore, I also support the call for loudspeakers with resistive impedance behavior, but I think the underpinning, given in the presentation of Peter Schüller is incomplete.

Another problem is the amplifier dynamics. Systems which measure nicely on steady signals can become pretty bad with dynamic signals (like you find in music). Remember that all the theory of Fourier c.s. is based on linear systems. And we are studying non-linear systems! So in order to get a good insight in the limitations of our system, we need to add tests which challenge the non-linearities. I have encountered several cases which looked pretty good with steady signals and broke down completely when some serious dynamics were used for testing.

Related to this is the statement "We all know that each power output measurement heavily depends on the exact mains voltage". This is only true for amplifiers with non-regulated power supplies. If an amplifier has good functioning regulated power supplies, the amplifier supply voltage should be independent of the mains voltage (within, of course, reasonable limits) and thus the output power should be independent of the mains voltage. If this is not the case, the dynamic behavior of the amplifier is flawed, so the first test of an amplifier would be to verify that its output power does NOT depend on the mains voltage!

I would welcome further elucidation and explanation from both Burkhard Vogel and Peter Schüller and I am willing to elucidate my points of view to them if necessary.

Dr. Hans R.E. van Maanen,
Bleiswijk, The Netherlands

Reference:

Hans R.E. van Maanen and E.T. Zonneveld, "An Extended Model for the Impedance and Compensation of Electro-Dynamic Units and their Determination", paper no. 3823 (P8.1), presented at the 96th AES convention, February 26 - March 01 1994, Amsterdam (The Netherlands)