

Feedback & Distortion in SE amps

A 300B amplifier design project including theoretical considerations



Author: M. Paanakker
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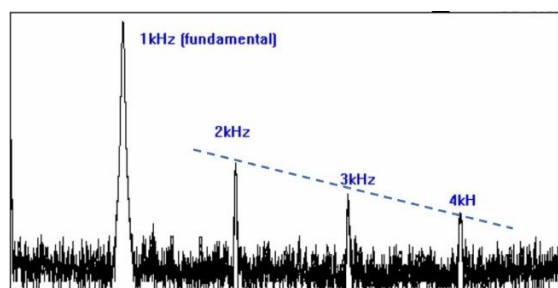
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Introduction, theory and results

Many Single Ended (SE) tube amplifier fans and audiophiles will tell you that negative feedback sucks the life out of music, leaving a dried husk, devoid of soul sound. At the same time, they are most likely listening to, and highly appreciating, their new DAC with negative feedback in its' output stage, as well as their preamp that may have multiple feedback loops. Things are never black or white...

Not everybody appreciates the sound of SE amps equally. Some people have a preference for classic Class-A FET amps, while others prefer push-pull tube amplifiers, etc. Different types of amplifiers offer a different harmonic profile. So, what constitutes the specific harmonic profile of SE amps?



A typical SE amp will produce harmonics that resemble the illustration on the left. Both odd and even harmonics will be present and a straight (dotted) line may be drawn through the harmonics' peaks, as long as the Y-axis is logarithmic (dB) while the X-axis is linear. All of the distortion is low order. Ideally it should be just a 2nd, 3rd and potentially 4th harmonic.

Figure 1 - The harmonic signature of SE amps

According to [Nelson Pass](#), the preference for harmonics breaks out roughly into 1/3 of people liking 2nd order harmonics, 1/3 liking 3rd order harmonics, and the remainder liking neither or both. This would explain why different people have different amplifier topology preferences.

The 2nd harmonic (2 kHz in the illustration above) is reported to provide a “fuller” and “warmer” sound, while the 3rd is reported to add “dynamic contrast”. Higher order harmonics, so 4th and 5th order or above, are likely perceived as actual distortion and less pleasing to our ears. Single Ended amplifiers seem to provide a mix of low order harmonics that is simply appreciated by SE fans. **The character of Single Ended amplifiers seems to stem from the fact that the production of odd and even harmonics adheres to the dotted line**, the amplitude ratios between 2nd, 3rd and 4th order harmonics being the key factor.

Simulation of the 300B amplifier design replicated the dotted line perfectly, despite the use of local feedback in the driver stage (see “300B Amplifier incl. driver – THD and IMD”).

As expected, connecting our signal source directly to the grid of our 300B power stage, simulations confirmed that **the dotted-line harmonic profile is entirely created in the output stage** (300B and OPT) and that the driver stage does not contribute to this harmonic profile.

Negative Feedback (NFB)

Feedback offers many advantages and only a single disadvantage. Let's start with the advantages;

- NFB stabilizes gain
- Increases bandwidth
- Lowers output impedance
- Lowers linear and non-linear distortion

A known disadvantage of NFB is that it potentially increases higher harmonics (see graph below).

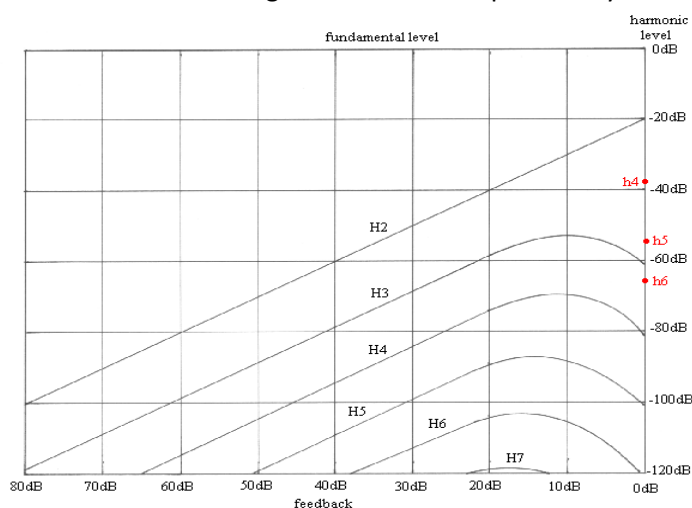


Figure 2 - Graph on increased harmonics as a function of feedback | Source: John Linsley-Hood

The conclusion can only be that if you're going to use feedback, you should use a lot. The reason is that, as Figure 2 shows clearly, the problem of increased higher order harmonics decreases above 15 dB, and is strongly reduced above 20 dB feedback. This implies that, in order to use feedback without increasing higher order harmonics, an open-loop gain of at least 20 dB above your target gain will be required.

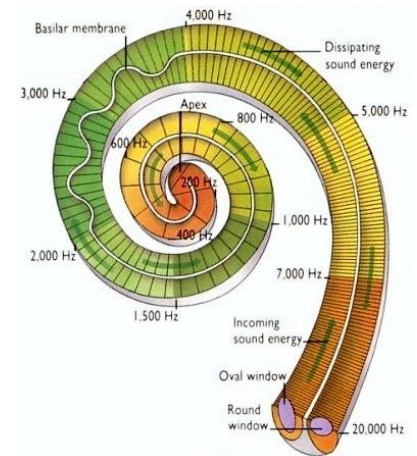
*The design of the driver stage in this project has an **open-loop gain of 59 dB**, while our **closed-loop gain is 33 dB**, so **26 dB feedback** was applied. That's more than enough, as confirmed by a very low THD (0.0076%) as well as simulated FFT results (see chapter "Driver-stage THD and IMD").*

The fact that NFB lowers linear and non-linear distortion has been listed above as an advantage. Generally speaking, this is one of the key reasons for using it. As previously discussed, in the case of an SE output stage, we actually like the harmonic distortion it creates, and so suppressing that distortion using NFB is a bad thing to do if you like the sound of SE amps. In our design, we **keep the SE output stage outside the NFB loop** (contrary to most 300B designs that use NFB).

Anecdotally, "Keeping the SE output stage outside the NFB loop" has frequently been generalized to "Feedback in SE amps is bad". This may explain why there are no Single Ended amp designs out there, or at least none that I could find, that make use of NFB in the driver stage. There are however many euphoric statements about "zero-feedback systems" to be found on-line. But given that probably every audio system is in fact using feedback somewhere or at multiple points in the chain (so from source to speaker), zero feedback systems don't really exist, and devices like a DAC or pre-amplifiers will have multiple sorts of feedback present for good reasons.

Studies via instrumenting sets of outer hair cell neurons have verified the creation of harmonics within the cochlea. Those studies (documented already in 1924) show that the human ear produces such harmonics by itself. The ear creates significant levels of 2nd order harmonics of nearly 10% of the fundamental (SPL > 90dBA). Some harmonics, at even 40dB and well above the average human threshold of hearing remain masked and will not be heard.

The ear and brain appear to be able to completely suppress certain harmonics, as long as these conform to a specific pattern. This pattern is called the **aural harmonic envelope**. It stands to reason that harmonic distortion in sound systems will be masked if they follow the same pattern.



Because THD and IMD measurements do not take the above into account, the correlation between THD or IMD measurements and perceived sound quality is poor. Earl Geddes and Lidia Lee state that;

- There is only a small correlation between THD and IMD measurements and the subjective impression of the sound quality of that system, and shockingly, **this correlation was negative**
- Measurements need to be based on the actual non-linear characteristic of the system and scaled (weighted) to account for the human hearing system (as in Gm measurements)

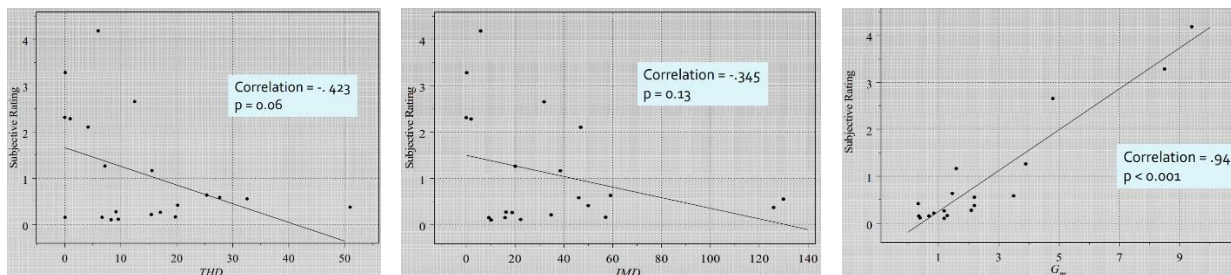


Figure 3 - Correlation between (from left to right) THD, IMD, Gm and perceived sound quality

Not disputing the above, it's likely that the perceived quality of SE amplifiers is (mostly) achieved by its' output stage, because the output stage is (almost) solely accountable for the harmonic profile.

Harmonic and Intermodulation distortion

Harmonic and Intermodulation distortion are a very broad and extensive area to cover. This chapter will be limited to highlighting most relevant characteristics of both. For further and in-depth theory and fundamentals, the following sources are hyperlinked and are recommended reads;

- [Understanding Distortion: A Look at Electronics, Part 2](#)
- [Auditory Perception of Nonlinear Distortion – Theory \(AES I\)](#)
- [Auditory Perception of Nonlinear Distortion \(AES II\)](#)
- [Audio Measurements Handbook by Bob Metzler](#)

Harmonic distortion (THD) occurs when a signal passes through a non-linear signal processor. In the illustration below this is visualized for a signal that is a single and distortion free sine wave with frequency “F”;

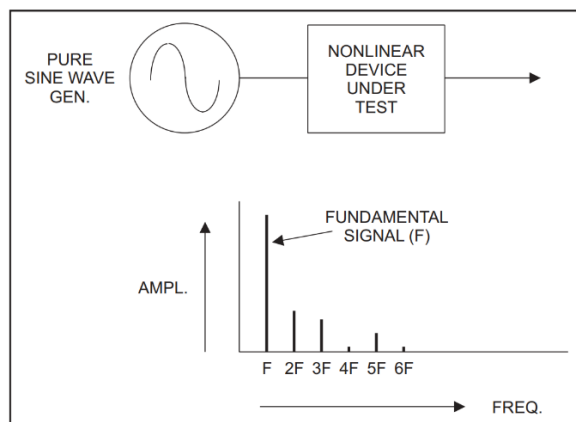


Figure 4 - Harmonic distortion (THD) of a single tone

In THD% measurements, the level of distortion is expressed as the ratio between the energy in created harmonics, excluding the fundamental, and the energy in the fundamental frequency (the clean input signal “F”). It’s important to understand that the outcome will hinge on the number of harmonics included in the measurement. Throughout this document, 19 harmonics above the 1 kHz fundamental have been included to ensure that we cover the audible spectrum (up to 20 kHz) in THD measurements. In practice, this only changes outcomes to a very limited extent, as the higher order harmonics are on a much lower level compared to lower order harmonics.

Intermodulation Distortion (IMD, see Figure 5) is the distortion that results from signals with different frequencies passing through a single non-linear signal processor. IMD unlike THD often has no harmonical relationship to the original signal at all. This results in distortion products that, even at low levels, may sound objectionable to human ears.

IM distortion products occur at frequencies that are sums and differences of the original input signals with different frequencies. For example, if the input is 500 Hz and 2200 Hz, then the strongest distortion products will occur at 1700 Hz and 2700 Hz. These frequencies (mostly) have no musical relationship to the original frequencies at all, so the result will sound discordant.

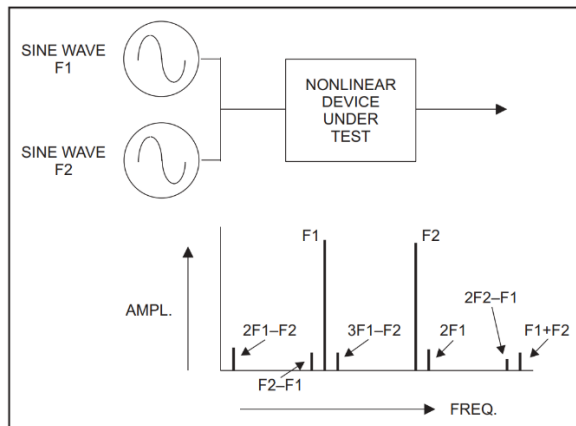


Figure 5 - Intermodulation Distortion (IMD) with two tone input

It's easy to understand how the number intermodulation products will add up, specifically when listening to complex music passages that hold many different frequencies. In addition, these products will generate new intermodulation products between themselves as well as with the fundamental, and so on, and so forth. The result is a whole spectrum of distortion products that weren't present in the original signal (referred to as "a complex noise floor" by Nelson Pass). This also explains why IM distortion may get aggravated when passing through multiple non-linear amplifier stages that are, to some extent, introducing different distortion products and mixing with ones already present.

In order to control IMD in the audio chain, it seems important to minimize IMD at each stage, with the possible exception of the output stage. Starting with a signal that has IMD can't be recovered from, and the effects get aggravated in downstream stages that introduce further non-linear distortion.

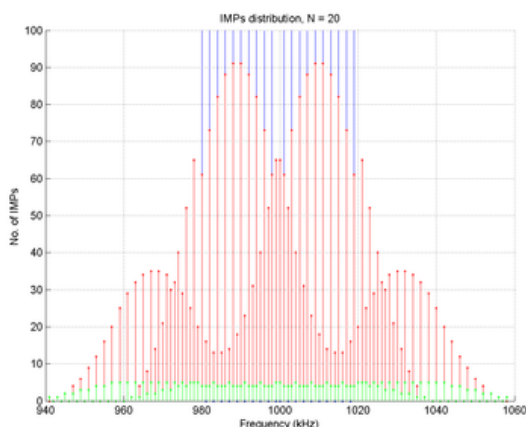


Figure 6 - Illustration of the "complex noise floor" (green) as a result of IMD and 20 input frequencies (blue)

Design philosophy

Nelson Pass makes a distinction between simple and complex distortion. He states that “IM distortion creates a complex noise floor which masks musical detail and already at lower levels takes the life out of music and makes it uninteresting, even irritating”. He also argues that distortion complexity increases when passing a signal through successive (non-linear) gain stages, each adding and mixing with newly incurred IM artifacts.

This may lead to the conclusion that, given that people like distortion as suggested by Geddes and Lee, **the desired distortion is best incurred within a single stage; The Single Ended output stage**, in order to keep the “complex noise floor” as low as possible.

Pass also argues that NFB increases distortion complexity. We also know that, at the same time, it lowers IMD and THD. Using NFB to reduce IMD in the driver therefore may have pros and cons. The question now becomes... which of the following negative effects results in more distortion complexity?

1. Increased complexity due to NFB
2. Increased complexity due to sending a signal with IMD present into a highly non-linear output stage

It has been demonstrated that IMD incurred in the driver is aggravated in the output stage when evaluating the “complex noise floor” that stems from IMD. **The “complex noise floor” is significantly reduced** by using a driver stage with NFB (see chapter “The ‘complex noise floor’ compared”).

Given these results, it can only be concluded that negative effect #2 by far outweighs negative effect #1 in this scenario. Given the significant reduction of the ‘complex noise floor’, there is likely (in line with statements by Pass and others) a positive effect on musical detail and liveliness.



“Why are we looking to reduce a subjective experience to objective criteria? The subtleties of music and audio reproduction are for those who appreciate it. Differentiation by numbers is for those who do not.”

– Nelson Pass –



Limitations of the simulation

There are some limitations to the accuracy of provided simulations. These relate to the simulation of the power supply as well as the OPT, and therefore the power-stage as a whole. The power supply was simulated as a 478.6 DC voltage with a 30V_{PEAK} sine ripple on top and 220 Ω of internal resistance. This has been done deliberately to ensure a robust driver design with high Power Supply Rejection Ratio (PSRR). The output-stage simulations are providing a **qualitative assessment only** as a result.

Specifications & measurements

The 300B design created as of part of this project (see chapter “Final schematic & operating points”) shows the following simulated specifications:

300B Amplifier E2E	
Input sensitivity	1 V _{RMS}
Power @ 0 dBV input	7.2 Watt
THD Amp 1 kHz / -9 dBV input (1W)	1.75 %*
Driver stage	
Gain	33 dB
THD driver-stage 1 kHz / 0 dB input	0.0076%
THD driver-stage 1 kHz / -9 dB input	0.0096%

*) Real world value depending on the actual OPT and Power supply, so the provided value is indicative

Normal (home audio) line level back in the days was 500mV_{RMS} (-6 dBV). Today, modern CD players and DACs typically output 2V_{RMS} (+6 dBV). As a consequence, a CD player or DAC with attenuator may be directly attached without the use of a preamp, and even leaves 6 dB headroom.

Initially the design of this amp intended to make do with just the 6SN7 stage, accepting a lower input sensitivity of 2 V_{RMS}. It was only when deciding to use an additional amplifier stage (the 12AX7 stage) that the idea of applying local negative feedback (NFB) in the driver stage came to mind, as well as the potential benefits that come with it. The remainder of this chapter will list the effects, as well as share the simulated results of the added stage in conjunction with NFB.

Bandwidth & phase shift	Lower -3dB point	Upper -3dB point	10 Hz Phase-shift	20 kHz Phase-shift
6SN7 stage only, without NFB	1.01 Hz	> 1 MHz	-5.8°	+1.0°
6SN7 + 12AX7, with NFB	0.98 Hz	300 kHz	-6.2°	+4.0°
300B + 6SN7 + 12AX7, with NFB	6.03 Hz	26 kHz	-50°	+45°

These results may give the impression that bandwidth and phase-shifting suffered from the application of NFB, but this is not the case. The “weakest link” within the NFB loop is actually the 12AX7 SRPP stage, which introduces significant phase shifting and has a pretty poor bandwidth. This is a consequence of the fact that the 12AX7 stage has been optimized for use **within** the NFB loop, in order to produce the lowest possible THD in the driver stage (see chapter “Driver-stage THD optimization”).

Bandwidth & phase shift	Lower -3dB point	Upper -3dB point	10 Hz Phase-shift	20 kHz Phase-shift
12AX7 stage only	1.2 Hz	14.7 kHz	-11°	+52°

As can be seen when taking the 12AX7 stage characteristics above into account, the NFB loop was able to compensate most of the 12AX7 stage weaknesses, while significantly improving overall THD and IMD results. The penalty for this in the end is a worsened phase response, when compared to the stand-alone 6SN7 stage. Phase shifts are a form of linear distortion and therefore may be considered benign.

THD & IMD	THD 1 kHz -9 dB input	2 nd order IM product CCIF 19 kHz + 20 kHz -9 dB input	3 rd order IM product CCIF 19 kHz + 20 kHz -9 dB input
6SN7 stage only, without NFB	0.0760%	-84 dB	-94 dB
6SN7 + 12AX7, with NFB	0.0096%	-103 dB	-108 dB
300B + 6SN7 + 12AX7, with NFB	1.75%	-51 dB	-81 dB

An actual IMD% figure hasn’t been established, but given the data above, we see an improvement of 19 dB for the 2nd order IM product, and a 14 dB for the 3rd order IM product. Assuming that the overall IMD across IM products would improve > 14 dB, **IMD%** of the driver is likely to be a **factor of > 5** smaller than the IMD% of the 6SN7 stage by itself. **THD%** improved by approx. a **factor of 8**.

Output impedance	Z _{OUT} @ 10 kHz
6SN7 stage only, without NFB	900 Ω
6SN7 + 12AX7, with NFB	90 Ω

Although not quantified, a visual of the ‘complex noise floor’ in the 300B amplifiers output (with 11 input frequencies), comparing the 6SN7 driver without NFB to the 6SN7+ 12AX7 with NFB, shows a staggering difference between the two. Results clearly favor the use of NFB in the driver stage (see chapter “The ‘complex noise floor’ compared”).

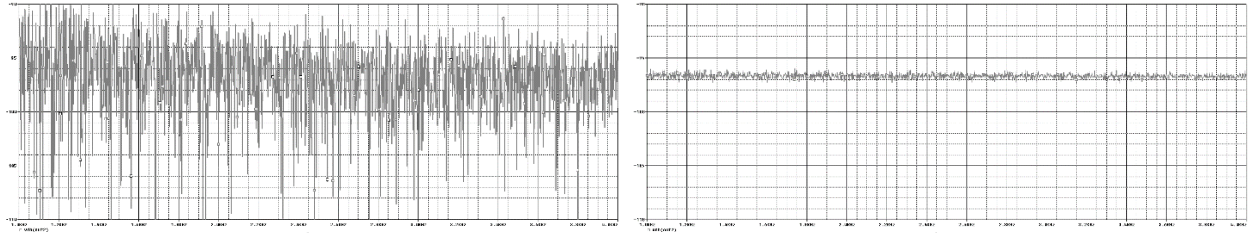
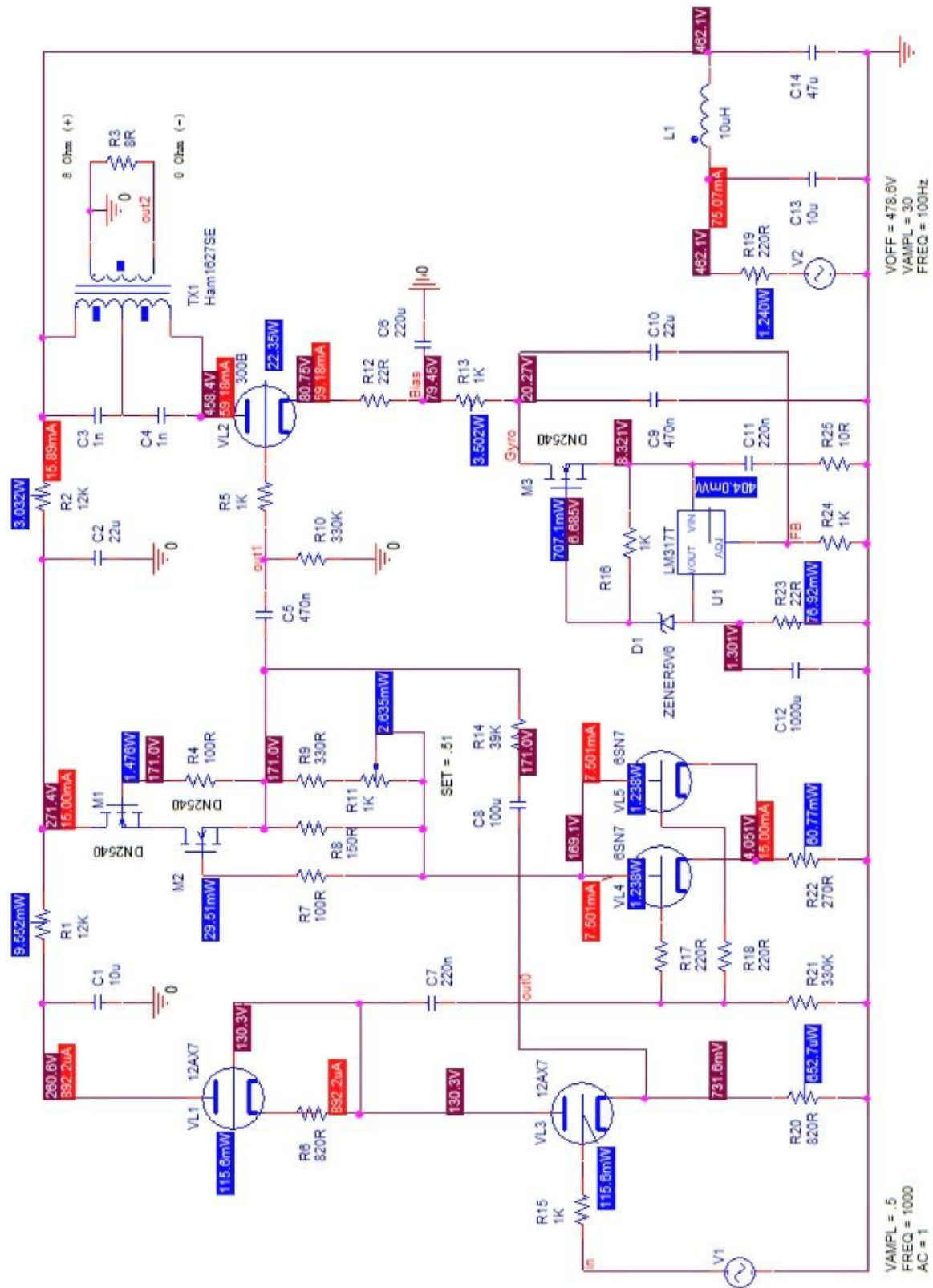


Figure 7 - The “complex noise floor” without NFB (Left) vs. with NFB (Right) with 11 input frequencies (comparable view)

Listening impressions

Given that the amp is yet to be built, this section is currently empty.

Final schematic & operating points



The driver-stage design

The 12AX7 SRPP stage design

The input-stage uses the 12AX7 (or equivalent ECC83S) tube. Given the roughly 260V power on the input-stage, each 12AX7 triode will have a U_a of approximately 130V. As can be seen in the ECC83S datasheet below, the curve of this tube becomes somewhat more linear at 1 mA. I chose to bias the tube at 900uA, the typical bias current according to the datasheet. The corresponding $-U_g$ therefore needed to be in the order of 0.6V (see plate characteristics below). The cathode resistor required is $U_g/I_a = 0.7V/900\mu = 778R$. The simulation using an 820R cathode resistor shows a current of 892uA and a $-U_g$ of 732 mV at a supply voltage of 260.6V, so 130.3V per 12AX7 triode. The tube now dissipates $2 \times 116mW$, which is well below the indicated maximum of 1W.

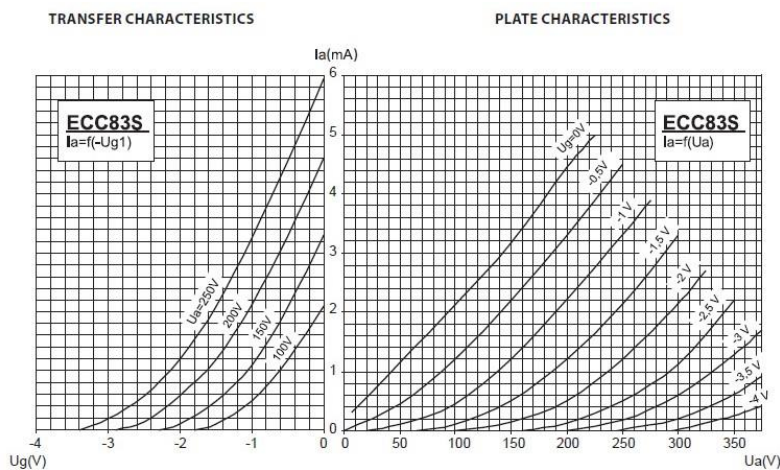


Figure 8 - Datasheet 12AX7 or ECC83A

The SRPP-stage by itself (without the feedback, so R14 or C8 removed) shows a simulated **THD of 0.194% at 1 kHz** and an input level of **0.1V_{PEAK}**. Without R14 or C8, the amplification of this stage is quite high and rapidly causes clipping. At the given input level, output V_{out0} shows a level of 4V_{PEAK}. The gain of the 12AX7-stage without feedback is therefore $4/0.1 = 40$ or **32 dB**.

Bandwidth is poor as I_a is quite low (900uA) and shows -3 dB points at **1.2 Hz** and **14.7 kHz**.

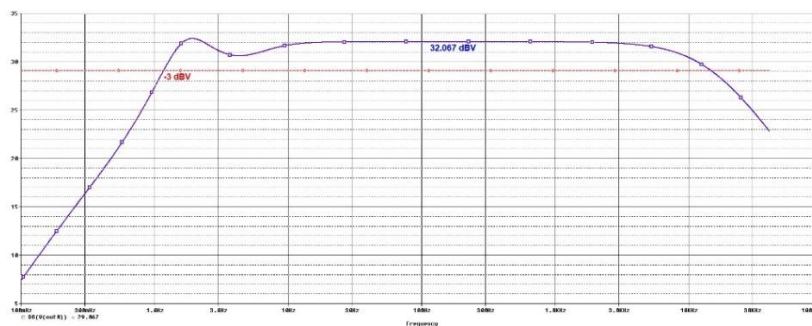


Figure 9 - 12AX7 SRPP Frequency Response

The CCS design

The Constant Current Source (CCS) design uses DN2540N5 devices. These are depletion NMOS FETs, used in a cascaded design. It was copied from “Bartola” on-line sources with no alterations and has been used by many. It will therefore not be explained in great detail, as more information can be found at;

<https://www.bartola.co.uk/valves/2015/08/31/ccs-not-everything-that-glitters-is-gold-part-i/>

In essence R6, R7, R8 and R10 combined set the required current, which is calculated as;

$$I_{BIAS} = U_{DS} / R_{SET} \approx 2V / R_{SET}$$

Please note that U_{DS} may vary between 1.5V and 3.0V between devices in the real world. It very much depends on the vendor and batch, so making R_{SET} adjustable is recommended and allows accurate matching of the bias current in the left and right channel.

In this case, to obtain a current of 15 mA, the combined resistance of R6, R7, R8 and R10 should be in the order of $2V/15mA = 133\Omega$. Our design allows us to vary between 103Ω and 135Ω . Simulation shows us that the voltage across $R_{SET} = 1.9V$, which matches exactly my actual batch of devices (because the PSpice model was adjusted slightly to reflect reality).

Note: The values for these resistors in the schematic match a specific sample of DN2540N5 devices, but may need to be adjusted slightly to replicate results with devices from a different batch. The important thing is to ensure that the left and right channel are biased equally. Whether that's at 14mA or 16mA is of less importance and hardly influences THD or gain levels (determined by simulated experimentation).

The 6SN7 stage design

The 6SN7 stage uses 2 triodes in parallel in order to lower the circuits output impedance. This also increases the combined bias current (now 15mA), which in turn improves the CCS performance at higher frequencies as it reduces the impact of the NMOS leakage capacities.

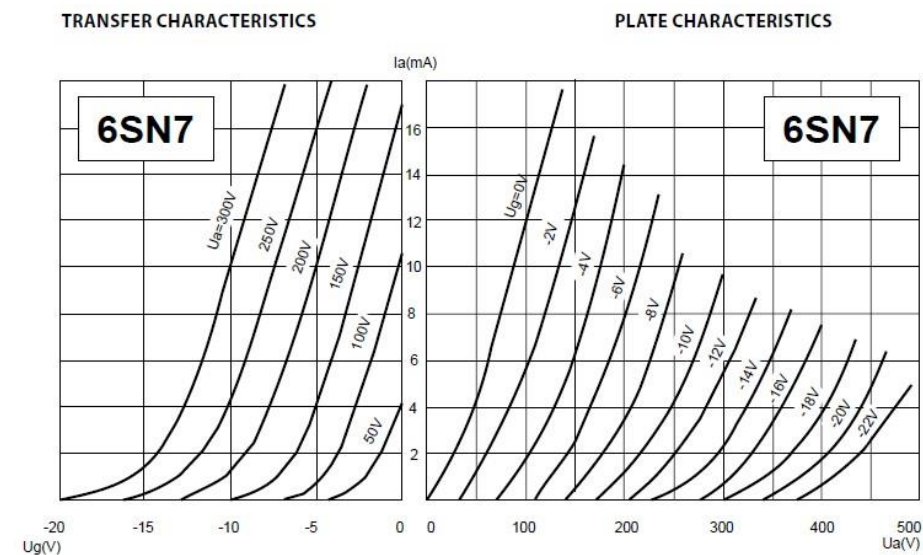


Figure 10 - 6SN7 Datasheet (JJ)

To reduce dissipation in the CCS, the voltage across the CCS is lower than the voltage across the 6SN7. If the voltage was spread evenly, then the dissipation in the CCS would be at the level of 2W ($135V * 15mA$), which leads to significant cooling requirements. This implies that the voltage across the 6SN7 should increase to 170V and so the 6SN7 dissipation will need to increase to 2.5W ($2 * 166V * 7.5mA$), which is well within limits (7.5W according to the datasheet). The voltage across the CCS will now be roughly 100V, so the power dissipation of M1 dropped to approximately 1.5W. At 1.5W dissipation we can more easily suffice with an on-PCB TO-220 heatsink.

The 6SN7 plate characteristics show us that at a U_a of 166V and an I_a of 7.5mA, the corresponding $-U_{gs}$ will be approximately 4V. The cathode resistor should therefore be $4V/15mA = 266R$. A bias simulation with a cathode resistor of 270R confirms this (see Final schematic & operating points).

When disconnecting C7 (no FB and 12AX7 stage disconnected) and connecting the input source to the second stage, we can now evaluate the behavior of the 6SN7-stage with the CSS separately.

The 6SN7-stage by itself now shows a **THD of 0.076% at 1 kHz** at an input level of $0.5V_{PEAK}$, which is a pretty good THD level for a well-designed 6SN7-stage. At this input level, output V_{out1} shows a level of $10.8V_{PEAK}$. The gain of the 6SN7-stage without feedback is therefore $10.8/0.5 \approx 22$ or **27 dB**.

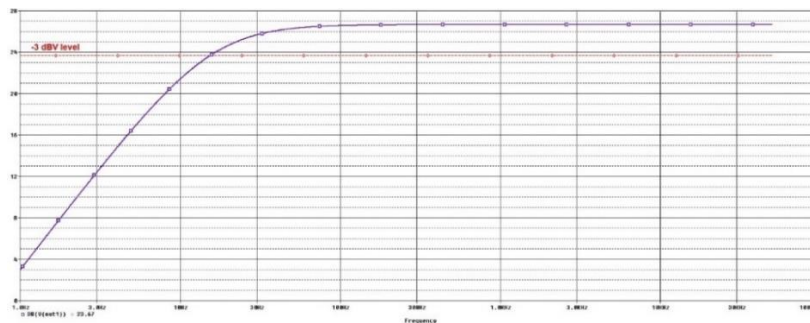


Figure 11 - 6SN7 stage Frequency Response

Bandwidth is more than sufficient and shows a -3 dB point at **15.4 Hz**, with the upper -3 dB point well outside the audible spectrum. The value of C5 needn't be increased, especially considering that the illustrated bandwidth was simulated without feedback.

Driver stage E2E (incl. NFB)

The open-loop gain of the driver-stage end-to-end without feedback is 32 dB (12AX7 stage) plus 27 dB (6SN7 stage), so 59 dB (a factor of **891**), way too much for our design. However, the open-loop headroom in gain is required in order to achieve enough closed-loop gain when negative feedback is applied. The reward will be that bandwidth increases and THD is reduced.

$$\text{GAIN}_{\text{CLOSED-LOOP}} = \text{GAIN}_{\text{OPEN-LOOP}} / (\text{GAIN}_{\text{OPEN-LOOP}} * K)$$

In this formula “K” represents the feedback factor. Given that the value of R20 is pretty low (820R), the plate resistance of the 12AX7 may be neglected. This means that K is solely determined by R20 and R14.

The required overall gain to achieve full power ($V_{\text{out1}} = 130V_{\text{PP}}$ @ 0 dBV input) is 65/1.41, so **46** or 33 dB. Therefore, K needs to be;

$$\text{GAIN}_{\text{CLOSED-LOOP}} = \text{GAIN}_{\text{OPEN-LOOP}} / (\text{GAIN}_{\text{OPEN-LOOP}} * K)$$

$$\Rightarrow K = \text{GAIN}_{\text{OPEN-LOOP}} / (\text{GAIN}_{\text{CLOSED-LOOP}} * \text{GAIN}_{\text{OPEN-LOOP}}) = 891 / (891 * 46) = \mathbf{0.022}$$

We also know that K is determined by the voltage divider R20 and R14, where;

$$K = R20 / (R20 + R14)$$

$$\Rightarrow R14 = (R20 / K) - R20 = (820 / 0.022) - 820 = \mathbf{36K5} \text{ (39K for R14 being the closest E24 value)}$$

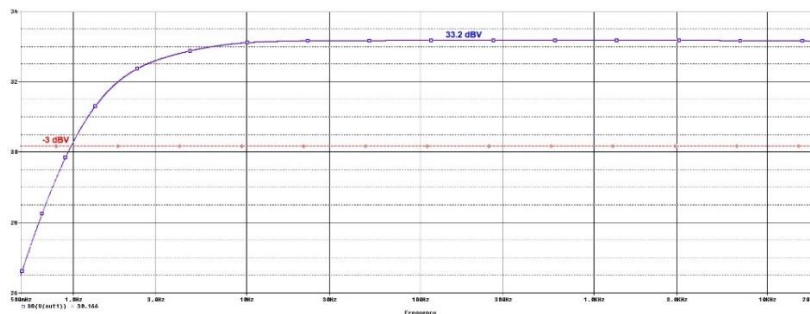
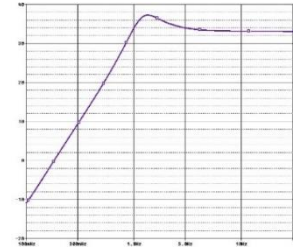


Figure 12 - Driver stage Frequency Response

With the use of a 39K value for R14, the simulation now shows an overall gain of 33.2 dB. We can also see that bandwidth increased, as the lower -3 dB point is at **0.98 Hz**, and the upper one well out of site at **300 kHz**.

The THD of the driver stage end-to-end at 0 dBV input (130 V_{PP} driver output) dropped to a staggering **0.0076%**, while at 0.5V_{PEAK} input (-9 dBV) it is on the level of **0.0096%**. At 0.2515V_{PEAK} input (-15 dBV) THD has increased to **0.015%**.

It has been considered to use the signal AFTER capacitor C5 instead of before C5, closer to the grid of the 300B for feedback. Although this showed improved THD results, it has a significant drawback. Due to capacitor C5, very low frequencies are filtered out, and therefore the negative feedback loop doesn't work for very low frequencies. This produces a peak at around 1 Hz. Although filtered out by the OPT, this will unnecessarily load the OPT and lead to saturation effects sooner and at lower power levels. A 2nd order filter in the feedback-loop solved the issue, but worsened THD results (as filters do). Using a choke at the grid of the 300B might solve our issues, yet would raise the cost of our design.



Driver-stage output impedance

The output impedance (Z_{OUT}) can be assessed by loading the output with a resistor to the point that output is lowered by 3 dBV (half the voltage). Experimentation shows us that this occurs when the circuit is loaded with a 90R value for R10 (V_{out1} @ 30.2 dBV at 0dBV input >10 kHz, instead of 33.2 dBV).

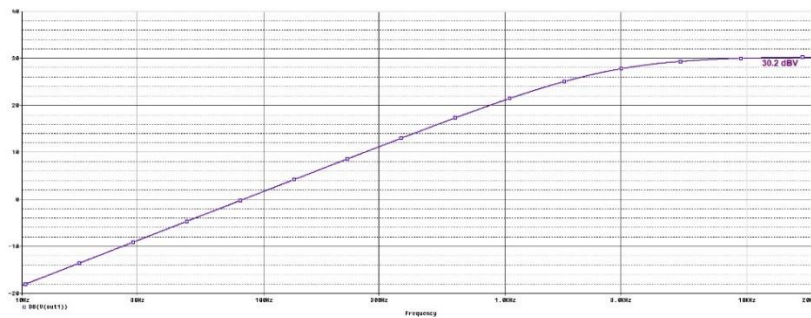


Figure 13 - Output impedance measurement

The driver-stage output impedance (Z_{OUT}) is therefore in the order of **90 Ω at 10 kHz**, yet much higher at lower frequencies due to C5. In case C5 is replaced with a 100uF capacitor, then Z_{OUT} remains on the level of 90 Ω down to 15 Hz, and so the driver-stage could easily be used as a headphone amp.

Note: Both increasing **and** lowering C5 gives us increased THD at -12 dBV input.

Driver-stage phase-shift

The phase shift that occurs is mostly caused by C5 and C7 at the bottom-end, as well as the 12AX7s internal capacities at the top-end.

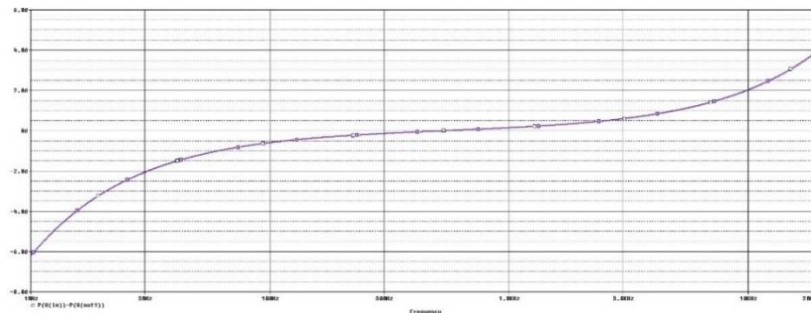


Figure 14 - Driver output phase shift vs. frequency

There is a phase shift of **-6.2° at 10 Hz** and a shift of **4.0° at 20 kHz**. The phase shift without feedback is much worse, with -16.5° at 10 Hz and 53.7° at 20 kHz.

Driver-stage THD and IMD

Driver-stage FFT view @ -15 dBV input (262mW output)

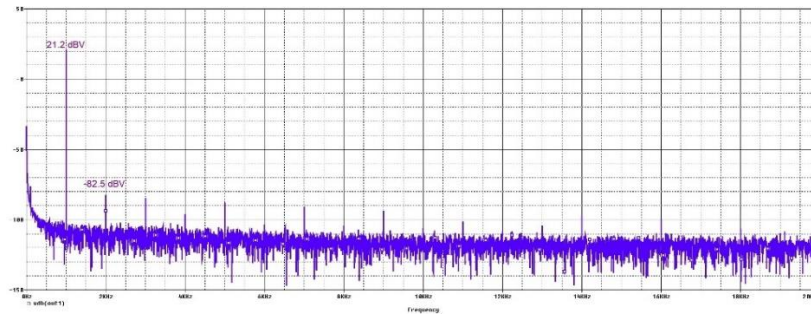


Figure 15 - FFT driver output | 1 kHz -15 dB input | THD% = 0.015%

Note1: The 1 kHz fundamental frequency measures 21.2 dBV, so the 2nd (2 kHz) harmonic has a relative level of minus (21.2+82.5) dB is -103.7 dB.
Note2: There is a mix of odd and even harmonics, with dampened even harmonics between 5 and 13 kHz. Odd harmonics die out above 13 kHz.

Driver-stage FFT view @ -9 dBV input (1W output)

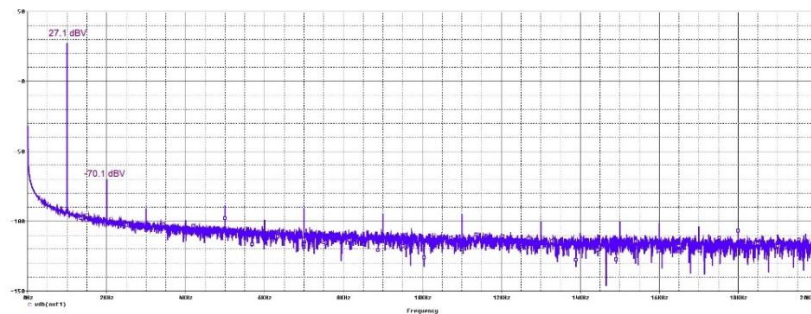


Figure 16 - FFT driver output | 1 kHz -9 dB input | THD% = 0.0096%

Note1: The 1 kHz fundamental frequency measures 27.1 dBV, so the 2nd (2 kHz) harmonic has a relative level of minus (27.1+70.1) dB is -92.2 dB.
Note2: There is a mix of odd and even harmonics, with dampened even harmonics between 7 and 15 kHz. Odd harmonics die out above 18 kHz.

Driver-stage FFT view @ 0 dBV input (7.2W output)

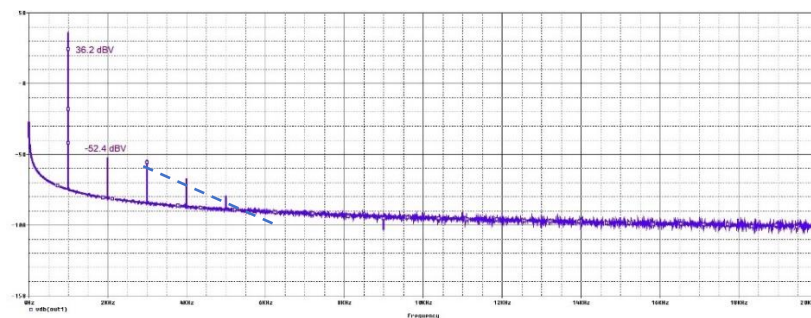


Figure 17 - FFT driver output | 1 kHz 0 dB input | THD% = 0.0076%

Note1: The 1 kHz fundamental frequency measures 36.2 dBV, so the 2nd (2 kHz) harmonic has a relative level of minus (36.2+52.4) dB is -88.6 dB.
Note2: There is a mix of odd and even harmonics that die out above 5 kHz, and that seem to decline linearly with the frequency (dotted line).

Unfortunately, IMD measurements can't easily be compared to other amps as they are rarely published in relation to audio amplifiers. There is however some level of standardization. Standards include CCIF and SMPTE. The choice here is to use CCIF (see Appendix C – CCIF (aka ITU-R) IMD measurement), as it's simpler and less tedious compared to SMPTE measurements (see Appendix B – SMPTE standard RP120-1994) and also easier to interpret. The CCIF measurement is selectively looking at **2nd and 3rd order intermodulation products** relative to the fundamentals. It provides a sufficiently accurate assessment.

In order to assess **the impact of Negative Feedback (NFB) on Intermodulation Distortion (IMD)**, let's compare IMD between a circuit with NFB (the driver as a whole) and one without NFB (the open loop 6SN7 stage only). The 6SN7 is selected for comparison, because it has low THD in itself and the 12AX7 stage hasn't been optimized for use without NFB. Using both tubes would also provide us with too much gain for a fair comparison.

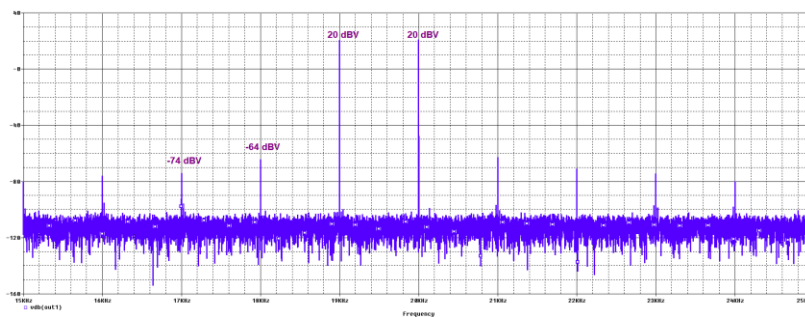


Figure 18 - CCIF IMD assessment – 6SN7 stage open-loop

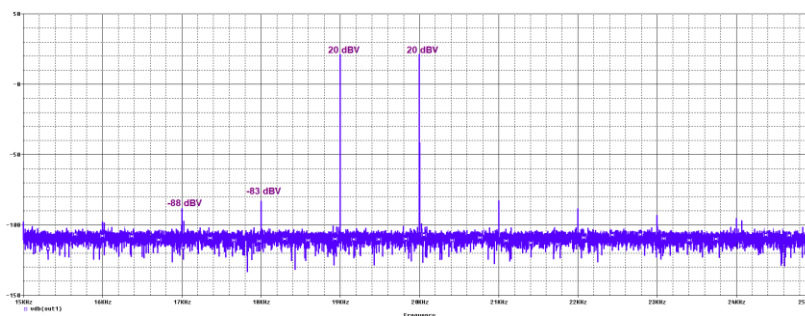


Figure 19 - CCIF IMD assessment / Driver-stage with 12AX7 + 6SN7 with NFB

The data shows a steep reduction in 2nd and 3rd order intermodulation products when comparing with and without NFB scenarios. It demonstrates a reduction of **19 dB of the 2nd order** intermodulation product, and a reduction of **14 dB of the 3rd order** intermodulation product when using NFB.

The 6SN7 stage in itself has a THD of 0.076% @ 1 kHz and -9 dBV input (16 dBV fundamental output level). This is probably better, but at least on par with most 300B driver stages. We can therefore conclude that using **local feedback in the driver-stage lowers the IMD of the driver-stage by roughly 15dB** or a factor of 5 at -9 dBV input.

Driver-stage THD optimization

Getting the THD to where it is now wasn't a first-time right exercise. It took multiple iterations and might still be improved upon e.g., by considering different tubes and topologies, like making use of a mu-follower for the 12AX7 stage which may increase gain and lower distortion even further (not yet investigated).

Some experimentation revealed that the following aspects allowed THD optimization and are listed below in the order in which they have been explored;

1. Adjustment of R1 and R2 to optimize supply voltages per stage and THD
 - Determining the resulting THD without feedback
 - At signal levels that are present with feedback at listening levels (33-59-12= -38 dBV input)
2. The bias current (I_a) of the 12AX7
 - Determining THD with feedback (closed loop) at different 12AX7 bias currents
 - Adjusting R6/R20 values in conjunction with R14 to meet the 33dB gain objective
 - All explored at listening level (-12 dBV input, 520 mW output)
3. Adjusting R22 to optimize THD at high voltage swings (130V_{PP}) at 0 dBV input
4. The value of C5 and C7
 - Determining THD with feedback (closed loop) with different C5 and C7 values
 - At a constant listening level (-12 dBV input, 520 mW output)

Going through the steps above three times resulted in a THD that is now over 5 times lower than the initial version. It's something that would be very hard to do in the real world, and where PSpice simulation really adds loads of value.

Driver-stage IMD optimization

According to Electra Print Audio (vendor of audio transformers), besides the obvious non-linearity of components, there is a very short list of causes of IMD that can be addressed (see Appendix A – Electra Print Audio IMD theory). For a driver-stage these are;

1. Selecting incorrect operating points (leading to increased non-linearity)
2. Too low values for bypass capacitors across self-bias resistors (as filtering aggregates IMD)
3. Too high value grid resistors (low frequency motion in U_g introducing IMD)
4. Lowering gain and/or the use of negative feedback

Fixed-bias (so a CSS or diodes to set the cathode voltage) as well as NFB to lower gain are mentioned as mitigations for #3, so well addressed in our design. Also, our closed-loop bandwidth is sufficiently high, thus avoiding increased IMD introduction due to filtering effects, so that point #2 is also covered. And as our THD and IMD measurements indicate, the linearity of our driver with NFB is excellent, indicating low non-linearity. In conclusion; There isn't much else we can do to further lower IMD in the driver.

The 300B output-stage design

Some output-stage design choices are based on the OPT and power transformer I have in my possession. For example, I have a power transformer that currently runs pretty hot at 75mA per channel. I therefore want to bring the current down a little and decided to aim for 60mA per 300B.

My current power supply provides me with a +B of 462V, resulting in a 300B plate voltage of 450V in conjunction with the OPT.

300B operating point

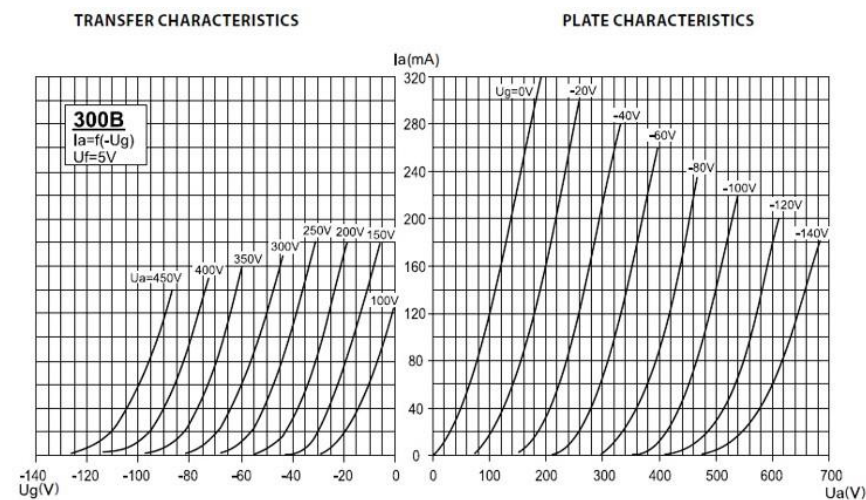


Figure 20 - 300B Datasheet

Due to the absence of a self-bias resistor, a load-line can't be drawn in the plate characteristics graph. The graph $-U_g$ as a function of U_a is not documented, but it can be derived from the datasheet for a fixed value of I_a , which we know will be 60mA.

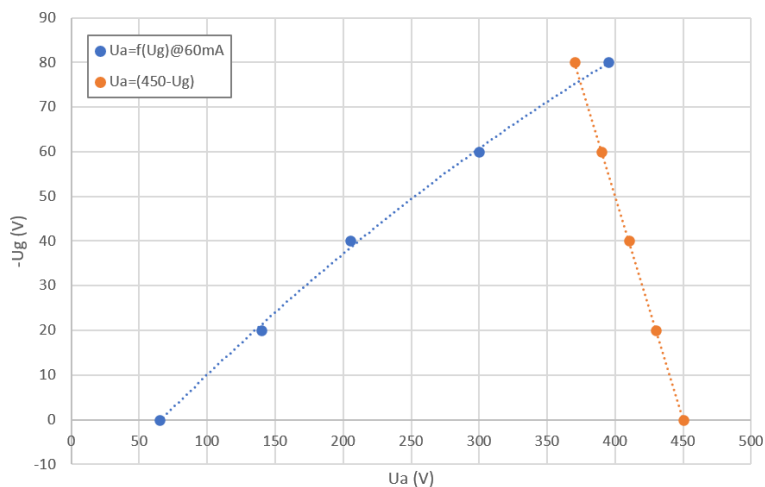


Figure 21 - Derived $U_a = f(U_g) @ 60mA$

The blue line in the graph above is derived from the datasheet, while the orange line is the formula $U_a = 450 - U_g$. We basically have 2 equations with 2 unknowns, thus proving the answer at the cross point.

Microsoft Excel allowed us to easily approximate the blue line as $-U_g = 0.2424 * U_a - 13.56V$

So, math now tells us that $0.2424 * U_a - 13.56V = U_a - 450V$

=> **$U_a = 373V$ and $-U_g = 77V$**

Simulation results divert slightly, as the simulated Hammond OPT has a different R_{DC} than my real-world OPT, and so the simulated plate voltage comes in 7 volts high. Also, the bias current is not exactly 60mA due to the need to select an E12 value for R23 (22R) that sets this current.

The auto-bias circuit design

The 300B auto-bias circuit is implemented using a Constant Current Source (CCS) with a basic LM317T. The schematic shows the LM317 configured as a current source, where the current is set by R23. In essence the LM317 compares the voltage across R23 to an internal V_{REF} of 1.25V. Hence the value of R23 to obtain 60mA is $1.25V/60mA = 20.8R$. Simulation using 22R gave us 59.18 mA.

The LM317 has some limits in order to function properly. The main ones are:

1. $V_{IN} - V_{OUT} < 40V$
2. $V_{IN} - V_{OUT} > 4V$

Because of limitation #1, the LM317T did not match circuit parameters ($U_g = -77V$ across the current source). There is a need to divert over-voltage and power-dissipation to another device. We chose (again) the DN2540 for this, because the combination of LM317 and the DN2540 has been reported to show excellent AC rejection across audio frequencies and may allow us to repurpose the schematic.

see https://www.pearl-hifi.com/06_Lit_Archive/14_Books_Tech_Papers/Jung_W/Current_Sources_101.pdf

Because of limitation #2, we have to add D1, a 5V6 Zener diode. This Zener ensures that $V_{IN} = V_{OUT} + \text{the Zener voltage} + V_{GS}$ of the DN2540.

=> $V_{IN} - V_{OUT} \approx 5V6 + 1V6 \approx 7.2V$ (may vary depending on the V_{GS} of the DN2540, in this case 1.6V)

We now have two issues left:

1. The DN2540 dissipating $(77V - 7.6V) * 60mA \approx 4.2W$, which requires a significant heatsink.
2. We now have a great CCS @ 59.2 mA with **high AC rejection**. But actually, we don't want any AC rejection as this implies a very high AC impedance, which means the 300B won't be driving any AC current through the OPT. The **OPT current will be kept constant, so there is no output**.

Issue #1 can be resolved by off-loading power dissipation to R13 (a 1K 25W resistor). Issue #2 can be resolved by adding C6, C10, C12 and R24, thus introducing AC hardening as discussed and explained by TubeCAD Journal (see <https://www.tubecad.com/2021/03/blog0531.htm>). The resulting auto-bias circuit will slowly regulate the bias current to 59.2 mA, while providing a short for AC signals and keeping V_{GYRO} at a fixed DC voltage (AC being rejected).

R13 lowers V_{GYRO} by 60V ($1K * 60mA$) and dissipates **3.5W**, so a heatsinking will be required.



Simulation shows a V_{GYRO} of 20.3V, with 12V across the DN2540, so it dissipates $12V * 59.2mA =$ **710mW**. It will be easy to cool with an on-PCB TO-220 heatsink.

Note: R12 only exists to simulate two 47R resistors or a potentiometer across filament connections (cathode) of the 300B. In a real build that includes a 300B heater, it should be replaced and a heater supply must be added.

Output frequency response & phase shift

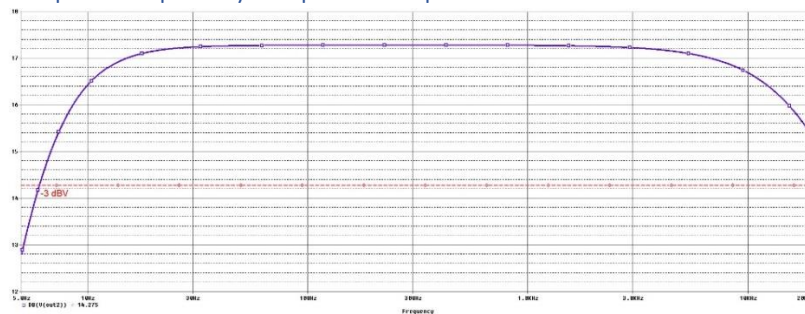


Figure 22 - 300B amplifier Frequency Response

The lower -3 dB point is at **14.3 Hz**, and the upper one out of site at **27 kHz**.

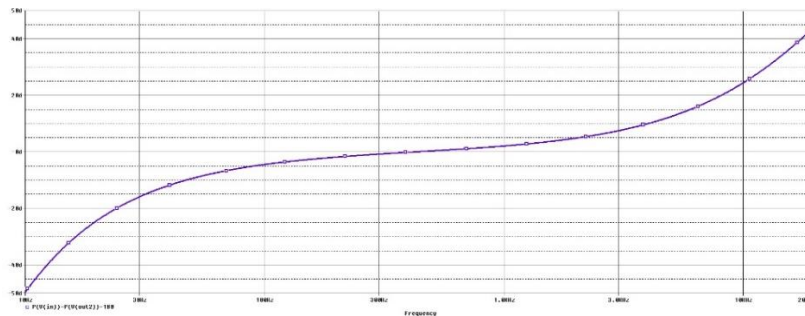


Figure 23 - 300B amplifier phase shift vs. frequency

We see a phase shift of **-50° at 10 Hz** and a shift of **45° at 20 kHz**.

Please note that in reality these results will (almost) entirely depend on the actual OPT and power supply. Therefore, these results should be considered qualitative.

300B Amplifier incl. driver – THD and IMD

Please note that the **30V/100Hz power supply ripple has been removed** for these simulations to make the FFT views more readable (otherwise showing 100 Hz modulation sidebands to the harmonics). All results should be considered **qualitative**, given the inaccuracy of the power supply and OPT simulations. The real world OPT and power supply will determine actual outcomes, and in fact play a critical role.

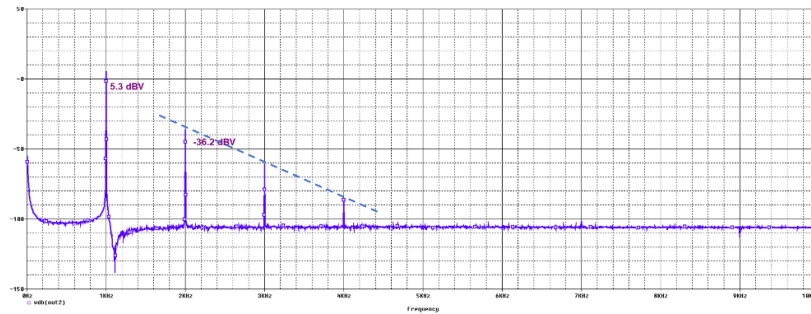


Figure 24 - FFT 300B Amplifier | 1 kHz -15 dBV input (262mW output) | THD = 0.819%

Note1: The 1 kHz fundamental frequency measures **5.3 dBV**, so the 2nd (2 kHz) harmonic has a relative level of minus (5.3+36.2) dB is **-41.5 dB**.

Note2: Harmonics from the driver stage are not showing (too small) and a 5th harmonic is absent (as it's below the dotted line).

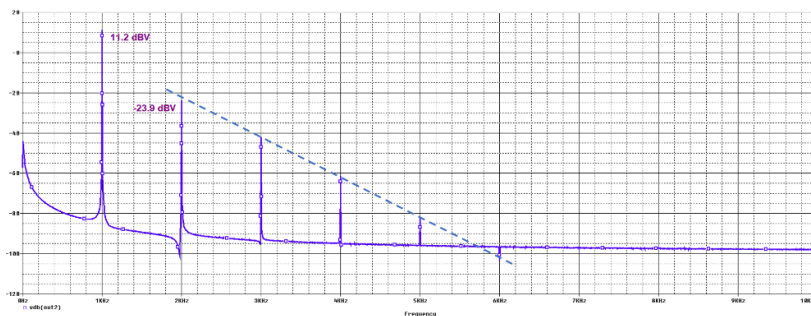


Figure 25 - FFT 300B Amplifier | 1 kHz -9 dBV input (1W output) | THD = 1.75%

Note1: The 1 kHz fundamental frequency measures **11.2 dBV**, so the 2nd (2 kHz) harmonic has a relative level of minus (11.2+23.9) dB is **-35.1 dB**.

Note2: Harmonics from the driver stage are not showing (too small), a 5th harmonic now appears and a 6th negative harmonic spike is showing.

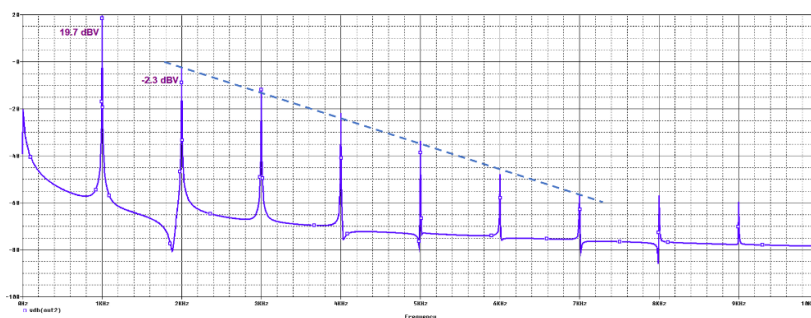


Figure 26 - FFT 300B Amplifier | 1 kHz 0 dBV input (7.2W output) | THD = 7.52%

Note1: The 1 kHz fundamental frequency measures **19.7 dBV**, so the 2nd (2 kHz) harmonic has a relative level of minus (19.7+2.3) dB is **-22.0 dB**.

Note2: Harmonics from the driver-stage are still not visible, but due to clipping (and in reality, OPT saturation) there is a deviation from the dotted line.

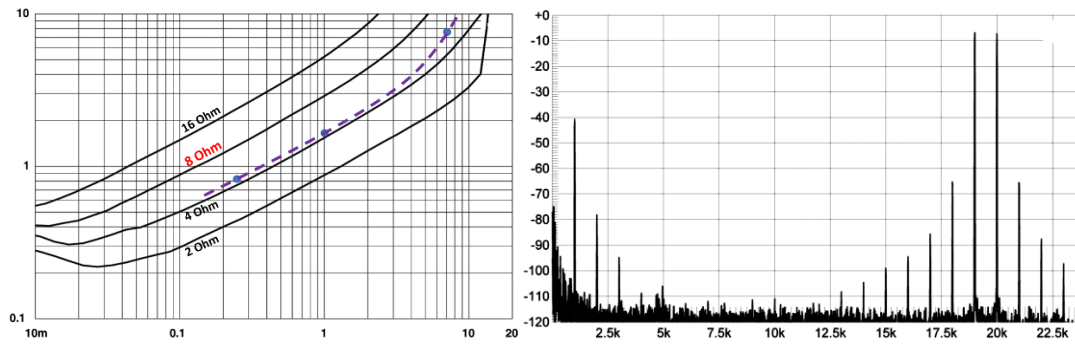


Figure 27 - (Left) $THD\% = f(Power)$, (Right) IMD measurement

Above are the results of a similar THD and IMD measurement, in fact a high-end 300B amplifier called the **Mastersound 300B SE**, as published by Stereophile.com in 2008. This provides a **reference point**, as long as we measure results in the same manner.

When we overlay the Mastersound 300B SE THD measurements with our own simulated measurements (purple dotted line), we can compare to real world measurements for an 8 Ω load. The simulated results show a lower THD% @ 8 Ω , but given the limitations of our power supply and OPT simulation, we can't draw any conclusions. It does however confirm a similar order of magnitude for THD% and increase with power, thus enabling a qualitative comparison in case we are making changes to our design.

The Mastersound 300B SE (CIFF) IMD measurement shows a 2nd order intermodulation product of -58 dBA at 18 kHz, which equals **-56 dB** (relative to the fundamental of ~2 dBV or -7 dBA at 19 kHz) and a 3rd order intermodulation product of -85 dBA at 17 kHz, which equals **-78 dB** (relative to the 1.7 dBV fundamental). However, this measurement was performed using an 8 Ω resistor while connecting to the 4 Ω OPT tap at 6W output. This provides a **higher primary OPT impedance**, and as Electra Print Audio points out (see Appendix A – Electra Print Audio IMD theory), that may lower IMD. Unfortunately, our simulated Hammond OPT doesn't have a secondary 4 Ω tap, and so results won't be fully comparable.

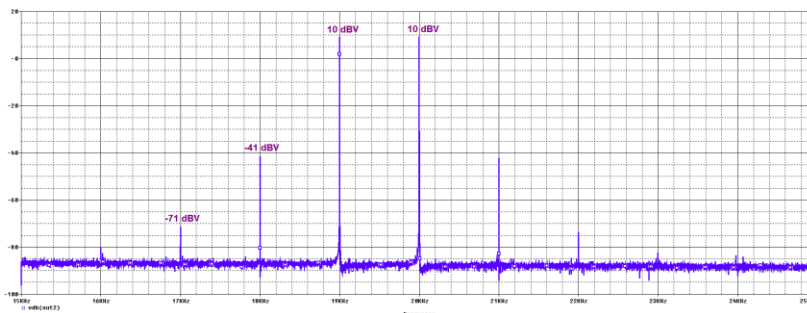


Figure 28 - IMD analysis (CCIF) at 1.6W output with 12AX7 + 6SN7 and NFB

We now know that the output-stage shows a 2nd order intermodulation product of **-51 dB** at 18 kHz, and a 3rd order intermodulation product of **-81 dB** at 17 kHz (relative to fundamental dBV level). This outcome is quite comparable to the measurements of the Mastersound 300B SE.

Now let's evaluate IMD in the output with and without NFB. To do that, we'll compare two scenarios, while keeping the fundamentals in V_{OUT1} at the same level (21.1 dBV). This V_{OUT1} level represents an "engaged listening" power level of 400 mW.

1. 300B Amplifier with just the 6SN7 stage without feedback (higher IMD into 300B)
2. 300B Amplifier with both the 12AX7 as well as the 6SN7 stage with NFB (lower IMD into 300B)

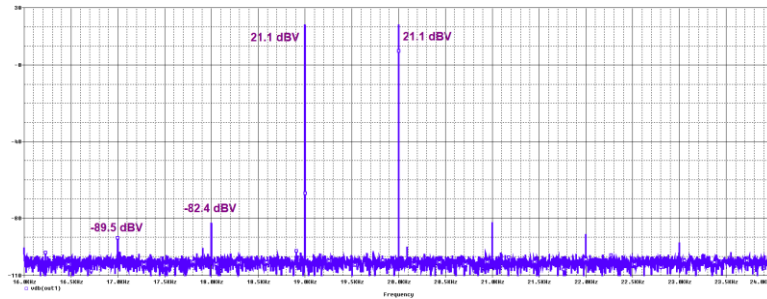


Figure 29 - 12AX7+6SN7 with NFB | CCIF (19 kHz + 20 kHz) V_{OUT1} FFT view

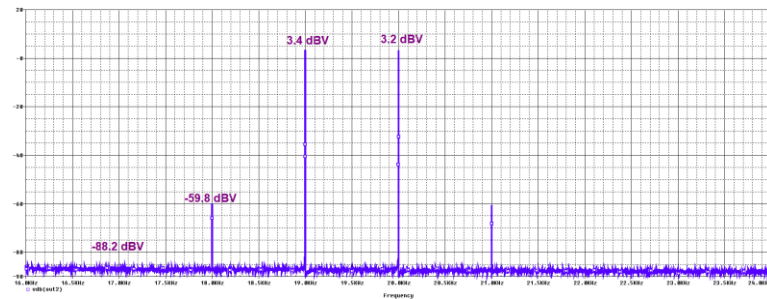


Figure 30 - 12AX7+6SN7 with NFB | CCIF (19 kHz + 20 kHz) V_{OUT2} FFT view [V_{OUT1} = Figure 29]

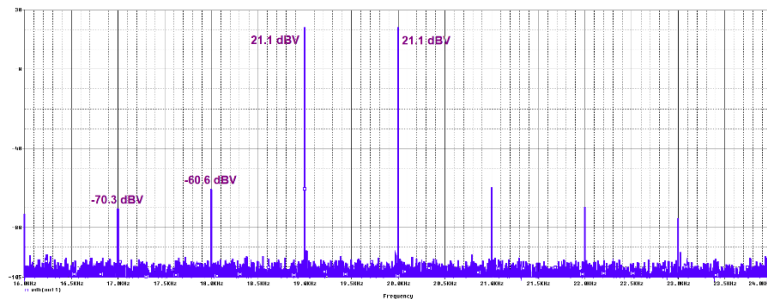


Figure 31 - 6SN7 without NFB | CCIF (19 kHz + 20 kHz) V_{OUT1} FFT view

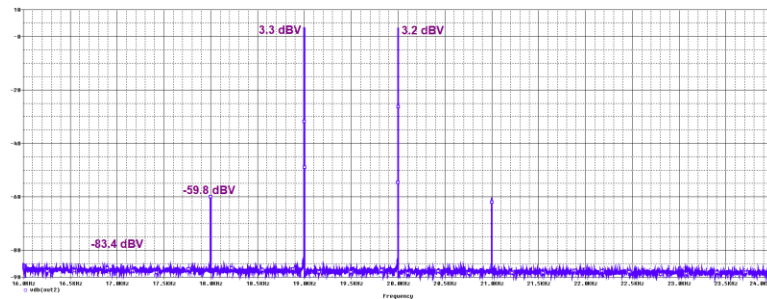


Figure 32 - 6SN7 without NFB | CCIF (19 kHz + 20 kHz) V_{OUT2} FFT view [V_{OUT1} = Figure 31]

Clearly, the CCIF measurements are unable to demonstrate any significant difference, or to provide conclusive proof of an aggravating effect between scenarios.

The 'complex noise floor' compared

Now let's evaluate the "complex noise floor" present in the output with and without NFB. To do that, we'll compare two scenarios, while keeping the fundamentals in V_{OUT1} at the same level (3.5 dBV). This V_{OUT1} level results in a V_{OUT2} level of -12.5 dBV (135 mW).

1. 300B Amplifier with just the 6SN7 stage without feedback (higher IMD into 300B)
2. 300B Amplifier with both the 12AX7 as well as the 6SN7 stage with NFB (lower IMD into 300B)

To create a 'complex noise floor' for comparison, we'll inject an input signal that consists of 11 different frequencies with no harmonic correlation between them. This signal is generated by the schematic illustrated in Figure 33. To obtain measurable results, and as was the case for measurements in the previous chapter, the power supply ripple (30V_{PEAK}/100 Hz) has been removed.

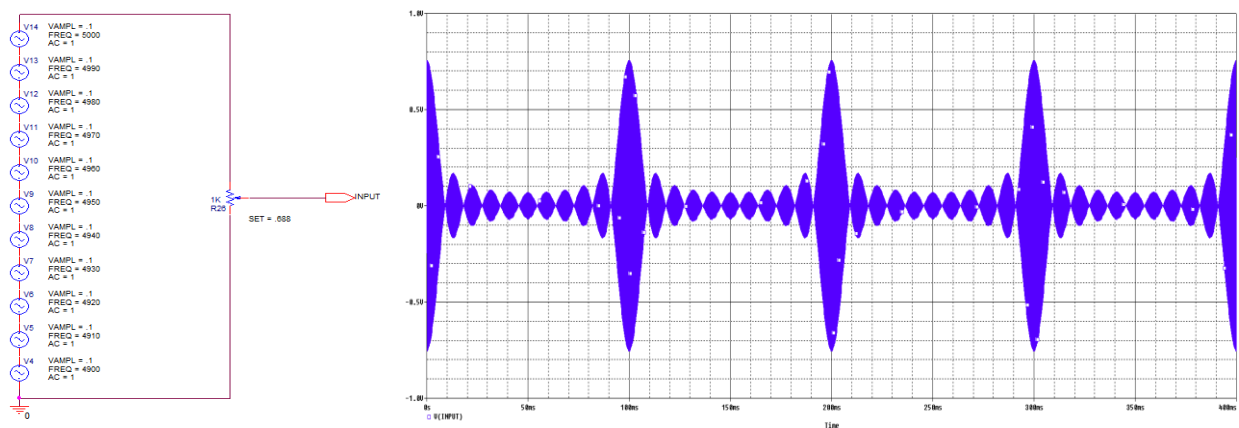


Figure 33 – Input source (11 frequencies) for noise floor evaluation schematic and waveform

As can be seen in Figure 34, there clearly is a complex noise pattern present in scenario #1, where we are only using the 6SN7 stage without NFB. Although the view zoomed in on the range 1 kHz - 4 kHz, where the 2nd, 3rd, 4th, 5th and 6th order IM products don't obscure our view, this pattern is present throughout the audible spectrum, up to well above 20 kHz.

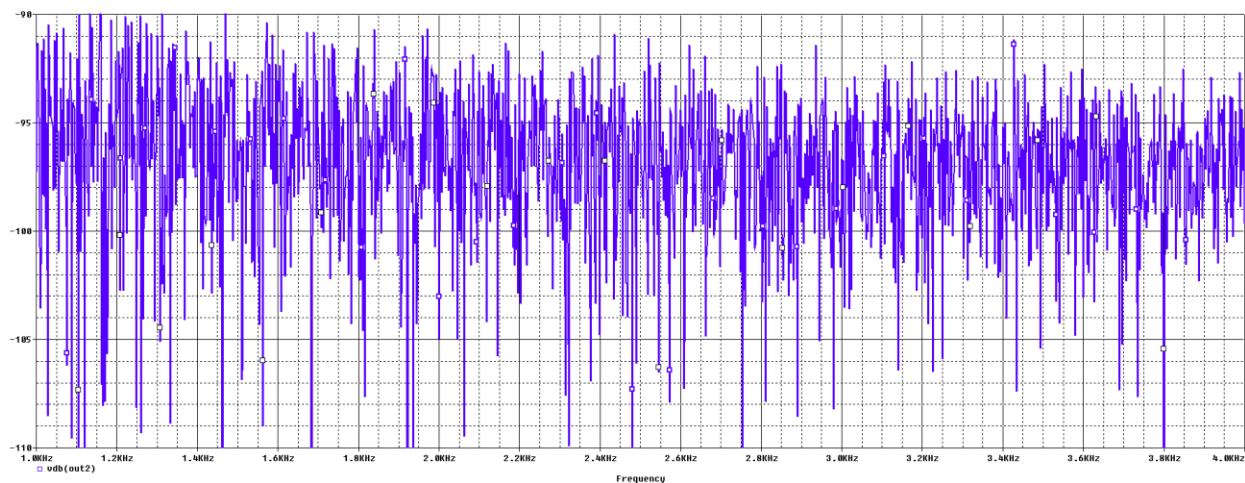


Figure 34 - The "complex noise floor", 6SN7 + 300B, without NFB (SET_{POT} = 0.688)

So does a lower IMD input signal for our 300B output stage (using NFB) impact the “complex noise floor”? We are now repeating the measurement using both the 12AX7 and 6SN7 in the driver with NFB, while ensuring that V_{OUT1} is on the same level of 3.5 dBV. This ensures that **Figure 34 and Figure 35 are fully comparable**. We also keep the scales of our graphs identical. The impact on “complex noise” is somewhat staggering, as can be seen in Figure 35.

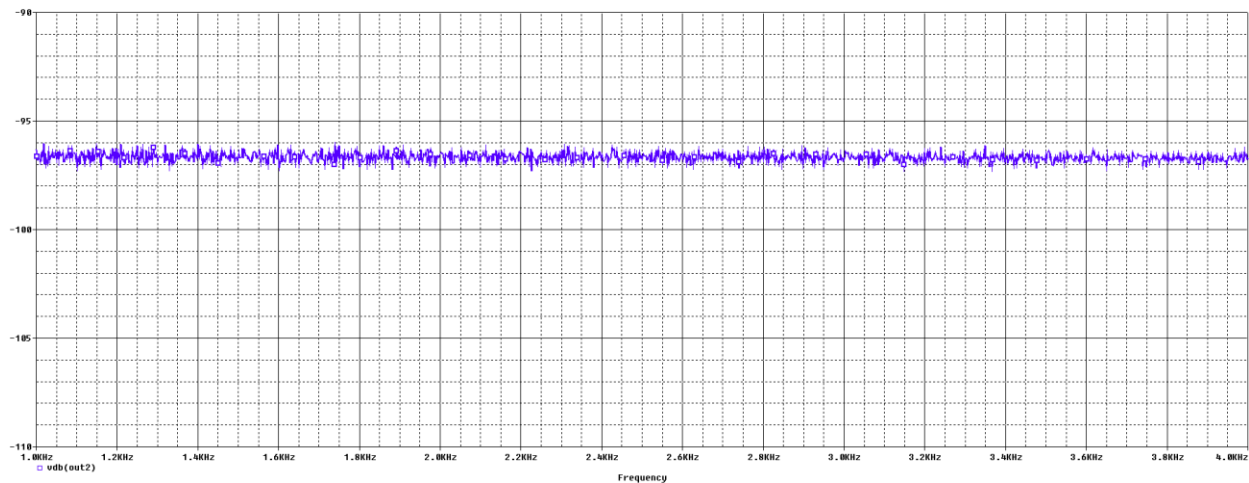


Figure 35 - The “complex noise floor”, 12AX7 + 6SN7 + 300B, with NFB ($SET_{POT} = 0.330$)

Conclusions

This comparison demonstrates that there is a significant positive effect on the “complex noise floor” in the amplifiers output signal, when using NFB in the driver stage. This is true, despite the fact that our signal travels through an extra high gain amplifier stage (12AX7). Clearly, the benefit of entering the 300B power stage with a lower IMD signal by far outweighs the IMD complexity added by using NFB (also see chapter “Design philosophy”).

When doing standard CCIF or THD measurement on these two 300B amplifier scenarios, the results as demonstrated are pretty much the same, and in fact not very impressive. This is because the CCIF and THD measurements are (almost exclusively) determined by the 300B output stage, power supply ripple and the OPT.

CCIF and THD measurements provide little to no insight into the “complex noise floor”, which in this scenario is mostly dependent on the IM distortion present in the driver stages’ output signal.

It is hereby confirmed that the presence of IMD in the input signal that feeds our 300B output stage (in other words, the IMD incurred in the driver stage), significantly aggravates the level of “complex noise” in the amplifiers output, while using NFB in the driver stage did not.

According to Pass, the most impactful consequence of IMD is the “complex noise floor”, as this is what takes the liveliness out of music and makes it sound uninteresting. The **level of complex noise** may therefore provide an **important metrics to evaluate quality** in Single Ended amplifiers.

Output stage IMD optimization

The previous chapter raises the question how IMD in the output stage may be further optimized. Electra Print Audio (see Appendix A – Electra Print Audio IMD theory) points at the following, specifically for SE power stages;

1. Reduce the OPT interwinding capacity, so that high frequency phase shift is minimized and increase the OPT core mass for less motion at the used power level. Summarized... **buy a better OPT.**
2. Load the output tube with a higher primary impedance so the tube is not subjected to its own nonlinearity below its optimized power operating point. **Using the 4 Ω instead of 8 Ω tap** of the OPT comes to mind in order to achieve higher primary impedance (many will recognize the benefit of doing this, as well as the downside that comes with it; reduced maximum power).

What stands out is that Electra Print Audio doesn't point to the use of interstage transformers (which they sell and produce) or the power supply. Interstage transformers are likely to have a poor effect on IMD, as each transformer is also a "mixer", hence the steep increase of IMD in the output stage, mostly due to the OPT.

The impact of the power supplies wasn't investigated as part of this project, but it's expected that it plays an important role in IMD generation as well.

PSpice simulation profiles & models

All THD simulations have been executed with following PSpice Time domain profile:

Run-to-time:	20ms
Start saving data after:	15ms
Maximum step size:	100ns
SKIPBP:	OFF
Center frequency:	1000 Hz
Number of harmonics:	19 (up to 20 kHz)

Note: Under these conditions the input signal simulates a 0.0094% THD distortion

All FFT simulations for harmonics have been generated with following PSpice Time domain profile:

Run-to-time:	200ms
Start saving data after:	0
Maximum step size:	1000ns
SKIPBP:	OFF

All FFT simulations for IMD have been generated with following PSpice Time domain profile:

Run-to-time:	1s
Start saving data after:	0
Maximum step size:	1000ns
SKIPBP:	OFF

```
.SUBCKT 300B 1 2 3 ; P G C; NEW MODEL
+ PARAMS: MU=3.95 EX=1.4 KG1=1550 KP=65 KVB=300 RGI=1000
+ CCG=2.3P CGP=2.2P CCP=1.0P ; ADD .7PF TO ADJACENT PINS; .5 TO OTHERS.
E1 7 0 VALUE=
+{V(1,3)/KP*LOG(1+EXP(KP*(1/MU+V(2,3)/SQRT(KVB+V(1,3)*V(1,3)))))}
RE1 7 0 1G
G1 1 3 VALUE={{PWR(V(7), EX)+PWRS(V(7),EX))/KG1}
RCP 1 3 1G ; TO AVOID FLOATING NODES IN MU-FOLLOWER
C1 2 3 {CCG} ; CATHODE-GRID; WAS 1.6P
C2 2 1 {CGP} ; GRID-PLATE; WAS 1.5P
C3 1 3 {CCP} ; CATHODE-PLATE; WAS 0.5P
D3 5 3 DX ; FOR GRID CURRENT
R1 2 5 {RGI} ; FOR GRID CURRENT
.MODEL DX D(IS=1N RS=1 CJO=10PF TT=1N)
.ENDS
```

```
.SUBCKT Ham1627SE P Sg B Sp1 Sp2
* Single Ended transformer, with taps at 50%
* 2500 to 8 ohms, 3db 10 to 40000 Hz
LP1 1 2 7h ; Primary
LS1 2 B 3.5h ; primary, screen grid tap portion
LSA 5 Sp2 0.07 ; secondary
KALL LP1 LS1 LSA 0.999342670070034
RP1 P 1 62
RP2 Sg 2 62
RS Sp1 5 0.1
.ENDS Ham1627SE
```

```
.SUBCKT 6SN7 1 2 3
+ PARAMS: MU=22.004 EX=1.2128 KG1=1213.7 KP=203.06 KVB=355.09
+ RGI=2000
+ CCG=2.4P CGP=3.9P CCP=0.7P ;
E1 7 0 VALUE=
+{V(1,3)/KP*LOG(1+EXP(KP*(1/MU+V(2,3)/SQRT(KVB+V(1,3)*V(1,3))))}
RE1 7 0 1G
G1 1 3 VALUE={ (PWR(V(7),EX)+PWRS(V(7), EX))/KG1 }
RCP 1 3 1G ;
C1 2 3 {CCG} ;
C2 2 1 {CGP} ;
C3 1 3 {CCP} ;
D3 5 3 DX ;
R1 2 5 {RGI} ;
.MODEL DX D(IS=1N RS=1 CJO=10PF TT=1N)
.ENDS
```

```
.MODEL DN2540 NMOS (LEVEL=3 RS=1.05 NSUB=5.0E14
+DELTA=0.1 KAPPA=0.20 TPG=1 CGDO=3.1716E-10
+RD=11 VTO=-2.12 VMAX=1.0E7 ETA=0.0223089
+NFS=6.6E10 TOX=725E-10 LD=1.698E-9 UO=862.425
+XJ=6.4666E-7 THETA=1.0E-5 CGSO=2.50E-9 L=4.0E-6
+W=59E-3)
.ENDS
*
```

```
.SUBCKT 12AX7 1 2 3 ; P G C; NEW MODEL
+ PARAMS: MU=100 EX=1.4 KG1=1060 KP=600 KVB=300 RGI=2000
+ CCG=2.3P CGP=2.4P CCP=.9P ; ADD .7PF TO ADJACENT PINS; .5 TO OTHERS.
E1 7 0 VALUE=
+{V(1,3)/KP*LOG(1+EXP(KP*(1/MU+V(2,3)/SQRT(KVB+V(1,3)*V(1,3))))}
RE1 7 0 1G
G1 1 3 VALUE={ (PWR(V(7),EX)+PWRS(V(7),EX))/KG1 }
RCP 1 3 1G ; TO AVOID FLOATING NODES IN MU-FOLLOWER
C1 2 3 {CCG} ; CATHODE-GRID
C2 2 1 {CGP} ; GRID=PLATE
C3 1 3 {CCP} ; CATHODE-PLATE
D3 5 3 DX ; FOR GRID CURRENT
R1 2 5 {RGI} ; FOR GRID CURRENT
.MODEL DX D(IS=1N RS=1 CJO=10PF TT=1N)
.ENDS
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Appendix A – Electra Print Audio IMD theory

AMPLIFIER DESIGNS FOR REDUCED IMD

We all would like our new amplifier to offer a clear display of the individual instruments, imaged perfectly and in place with distinct bass. Electra-Print Audio Company has developed new low IMD circuitry and reviewers say it sounds like twice the power, with better imaging, authoritative bass and does not show harshness at high volumes. Intermodulation Distortion (IMD), is the resulting sound heard of musical instruments of a lower frequency, modulating or mixing with, instrument(s) of higher frequencies during playback. This combination of mixed tones, depending on the music and level changes, results in a harsh dissonant screech as music increases in volume. This is commonly mistaken with Total Harmonic Distortion (THD), probably due to the fact that IMD is never indicated in the amplifier specs, but THD is.

Electra-Print Audio has been investigating the effects, causes and reduction of IMD in tube amplifiers. The following is a list of causes and cures. IMD being the modulation of high frequencies by the lower frequencies with a resulting sum and difference of many odd combinations of audible products. The standard measured value of IMD for very good performance of an SE amplifier is less than 2% at 1 watt and about 4% at the amplifier's mid-power level. For very low power amps, such as the 45-tube, 2% at .5 watt is used. All of our tests are done with, two tone, 60Hz/6kHz, 4:1, SMPTE standard signal. These amplifier designs do not use global negative feedback, common in push-pull amplifiers. Those amplifiers that use negative feedback show results with very low IMD as well but also generate artificial bandwidth and its resulting signal compression. This is at the expense of output power with a dependency on Nyquist phase analysis so that the amplifier will not oscillate.

Causes and Cures:

Cause #1 - Output transformers, SE design: The output transformer when combined with the output tube gives highest generation of IMD. The core mass of an SE output is in play with the lower end of the bandwidth. The core has a slight magnetic lag at these low frequencies and as the high frequencies are induced, this lag modulated these highs. The core is not in use above 2-3 kHz but does steer the field inwards for the coil and this keeps the efficiency high. The resulting IMD is a product of the individual cadence of core and induced high frequencies.

Cure #1 - is to reduce the interwinding capacity so high frequency phase shift is minimized and increase core mass for less motion at power level used. Also, loading the power output tube to a little higher primary impedance so tube is not subjected to its own nonlinearity below its optimized power operating point. The interwinding capacity will shunt the primary and lower the impedance above 10 kHz and output tube's nonlinearity will add to the transformer IMD generation.

Cause #2 - Output tubes will generate IMD in many ways. For example, when the plate load is too low, an incorrectly selected tube operating point, a low value bypass cap across self-bias resistor and grid bias voltage in motion with the signal due to a too high value grid resistor. Any combination of the above-mentioned problems will result in two or more different motions in a signal path. This increases IMD. Driver and voltage amp tubes with un-bypassed self-bias resistors used for decreasing THD and/or used for local feedback for lowering gain, will increase IMD due to the many different motions at their cathodes.

Cure #2 - Heavy bypassing of this resistor so all motion is at a very low frequency which will not be too noticeable. Bias source impedance is reduced as low as possible so it remains in control of the tubes linear operating point. Using a fixed bias or bias supply and grounding its cathode or filament, would be the best combination for this. With driver and voltage amp stages an algebraic summing resistor can be used to null IMD motion slightly but measurably, also to lower gain slightly (local NFB).

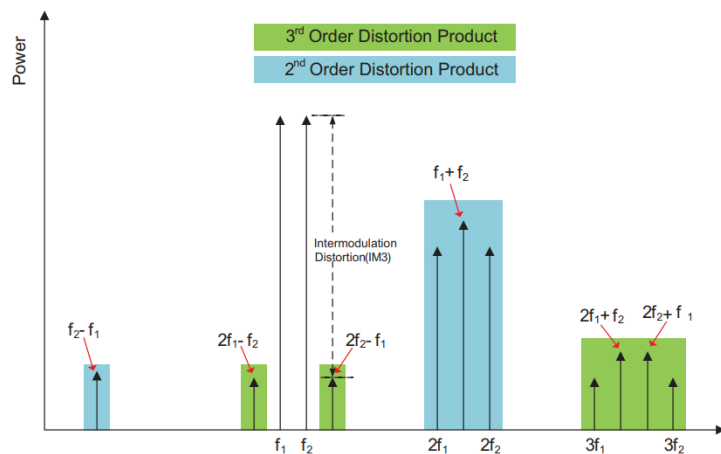
Very low THD will follow the lowering of IMD product generation, not the other way around. We have measured a very low THD with very high IMD measured at the same operating level.

Appendix B – SMPTE standard RP120-1994

Intermodulation distortion is a multi-tone distortion product that results when two or more signals are present at the input of a non-linear device. All semiconductors inherently exhibit a degree of non-linearity, even those which are biased for “linear” operation. The spurious products which are generated due to the non-linearity of a device are mathematically related to the original input signals. Analysis of several stimulus tones can become very complex so it is a common practice to limit the analysis to two tones. The frequencies of the two-tone intermodulation products can be computed by the equation:

$$M \cdot f_1 + N \cdot f_2$$

Where M & N are integers (can be positive or negative) The order of the distortion product is given by the sum of M + N. The second order intermodulation products of two signals at f1 and f2 would occur at f1 + f2, f2 – f1, 2f1 and 2f2 (see figure below).



It is common to talk about the third order intermodulation products as being $2f_1 \pm f_2$ and $2f_2 \pm f_1$. IMD measurement, then, describes the power ratio between the power level of the output fundamental tones (f2 and f1) and the third-order distortion products ($2f_1 - f_2$ and $2f_2 - f_1$).

How to measure

The stimulus is a strong LF interfering signal (f1) combined with a weaker HF signal of interest (f2). f1 is usually 60 Hz and f2 is usually 7 kHz, at a ratio of Amplitude(f1) : Amplitude(f2) = 4 : 1 (-12dB). The stimulus signals are simply superimposed. When non-linear distortion is present, this stimulus results in an AM (amplitude modulated) waveform, with f2 as the “carrier” and f1 as the modulation. In analysis, f1 needs to be removed, and the RMS level of the modulation products is expressed as a ratio to the RMS level of f2.

Appendix C – CCIF (aka ITU-R) IMD measurement

Another approach to standardization of IMD measurements is the CCIF standard. The CCIF IMD method is described in document no. 11 of the Commission Mixte, CCIF/UIR, March 1937, issued by the International Telephonic Consultative Committee (CCIF). CCIF no longer exists as an organization, having become the ITU-R division of the International Telecommunications Union (ITU). This method is also referred to as IMD (ITU-R).

The CCIF stimulus is two equal-level high-frequency tones f_1 and f_2 , centered around a frequency called the mean frequency, $(f_1+f_2)/2$. The tones are separated by a frequency offset called the difference frequency. The tones inter-modulate (in case non-linear distortion is present) to produce sum and difference frequencies.

How to measure

For analysis CCIF selectively measures the 2nd and 3rd order intermodulation products, combines their values arithmetically and provides a result that is the ratio of the sum of the products to a reference voltage defined as 2x the voltage of f_2 (effectively, the sum of f_1 and f_2).

Because the stimulus tones are high in frequency, CCIF is a useful measurement for observing distortion in devices that exhibit distortion that rises with frequency. Much of the energy contained in the distortion products will fall near or below the stimulus tones. This makes CCIF a good choice for measuring distortion at higher frequencies in band limited devices, where harmonic distortion products from high-frequency stimulus tones would fall out of band.

