

Simulation of Crossover Distortions in Class AB Audio Amplifiers: 0.00002% at 20kHz / 100W

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ABSTRACT

The following is an attempt to apply the theory of robust feedback control to the design of Class AB audio amplifiers. The potential designs are analysed by simulations using LTspice invoked from MATLAB in a batch mode. It is repeatedly demonstrated that application of a standard -20dB/dec feedback loop is broadly equivalent to taking a derivative of the pre-existing non-linear distortions which may lead to the lower distortions - or may not. The proposed analysis and visualization framework aids in better understanding of Class AB amplifiers nuts n' bolts and serves as an illustration of the non-trivial underlying complexity of feedback control over time-invariant static plants with hard non-linearities. Armed by this better understanding, we can apply nested feedback loops and 50kHz low-pass filtering to lower the total harmonic distortions from 0.5% for 3-stage Emitter Follower down to 0.00002% for frequencies $\leq 20\text{kHz}$ and power $\leq 100\text{W}$ for $V_{cc}-V_{ee}=67.2\text{V}$, $R_L=4\Omega$, according to simulations.

INTRODUCTION

The history and the modern state-of-art of Class AB amplifiers' development is well described in the books of Douglas Self¹. The approach is based on the Audio Precision (TM,C) measurement approach, which has become a de-facto standard quite a long time ago. This approach is based on relatively easily measurable data but it has an implicit assumption that there are only "soft" distortions. It is quite adequate for relevant systems, with compact² non-linear kernel³ and therefore quickly decaying spectrum such as old-fashioned tube (or Class A) amplifiers. Within this approach, developers can focus at a single number of THD, may look at the Fourier harmonics but do not need to bother looking into the details of the time-domain distortions.

¹ see <http://www.douglas-self.com/ampins/dipa/dipa.htm> and references therein

² And invariant to the basis's rotation, be it time/polynomial/frequency as far as the transform is orthonormal.

³ Apologies to EEs for using mathematical jargon here and there. There is no intent to show off or impress the readers. I tried to avoid it, as well as formulas, using only graphics to illustrate the discussed concepts.

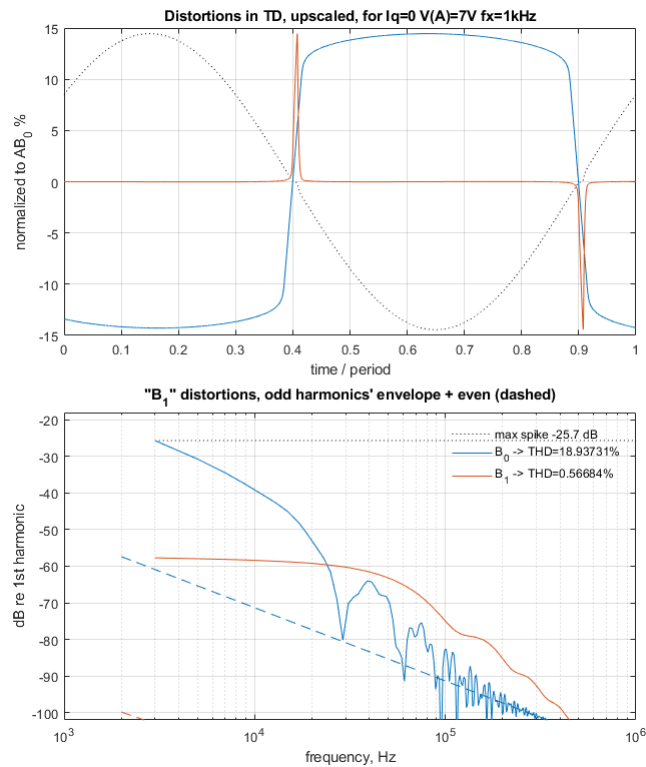
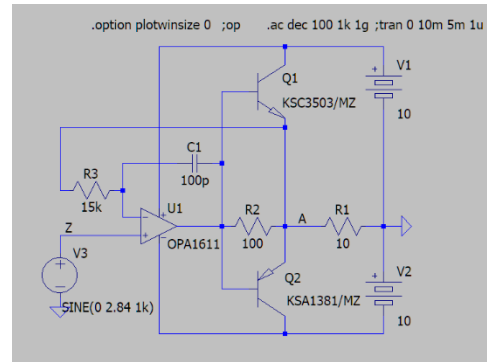
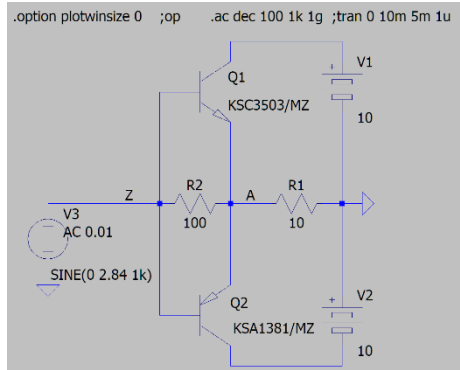
The amplifiers with “hard” non-linearities, such as Class AB, do not have a compact non-linear kernel i.e., polynomial approximations make no sense. Applying the same “soft” approach to their characterization is of a debatable merit; moreover, it does not agree with well-established theory and practice of robust feedback control.

Here we use a viewpoint that a feedback control loop is a sequence of inversely-matched differentiating and integrating operators⁴ because it helps to explain the observable phenomena easier (and without contradicting any known fundamental facts).

Let’s start with reiterating a few fairly obvious statements:

1. A feedback loop is incapable of fixing a “true” Class B amplifier with the output stage’s flat differential gain $dy/dx = 0$ at the $I_q=0$; regardless of the op-amp linear gain.
2. The feedback loop of the first order of astatism (-20dB/decade) is incapable of tracking a linear drift without a delay $\sim 1/f_T$, therefore missing an opportunity to fix the non-linear transient crossover distortion. Even in the ideal theoretical case, even with application of several nested loops, feedback-based control can not eliminate crossover distortions.
3. Adding a feedback loop with an integrator reduces non-linear distortion, crossover or “soft”, same as any other distortions, by the value of feedback loop depth at a given frequency, which is functionally equivalent to taking a derivative of the distortions. The absolute amplitude of distortion does not necessary decrease with the increase in Gain Bandwidth Product (GBP) f_T , but the width certainly changes. Note that the high-order distortions always increase, due to the high-pass nature of “inverted” integrator, which is a differentiator. See *test_b.m* for more details.

⁴ Michael Zrull (2020). Fast Subband Adaptive Filtering (FSAF)
(<https://www.mathworks.com/matlabcentral/fileexchange/<...>>), MATLAB Central File Exchange.



For the characterization of “hard” crossover distortions (like on the picture above) here, we explicitly check that we conform to band-limitedness i.e., Parseval’s theorem to ensure that the distortions’ energy measurements are invariant to the orthonormal transformations of the basis.

METHOD

As we've already seen, crossover distortions closely resemble a sequence of Dirac delta functions striking at each zero-crossing point (of current not voltage). Design of a real-life testbed that ensures reliable measurements of signals with so wide spectra, 20+ bit precision, and an ability to clearly separate the real signal from artifacts is considered a problem to be addressed later. Here, we resort to simulation only⁵.

Several pSPICE modelling software packages were tried. It seems that the modelling software vendors genuinely hate their customers. LTspice (TM,C) was found to be the lesser evil. Still, its user interface is horrible, support is insulting, documentation is misleading, etc. Saving grace, there are forums⁶ and user groups⁷ with knowledgeable, genuinely respectful and helping people.

Simulation Stages

The simulation is split onto the following stages:

1. The schematics are created in the LTspice graphical editor and saved as *.sp:
View -> SPICE Netlist -> right click -> Generate Expanded Listing. Any other editor could be used as far as the resulting netlist is the same.
2. The expanded netlist is imported into MATLAB as a text file.
3. Inside MATLAB, the netlist can be modified in a few ways:
 - a. Component (including voltage source) values, can be modified as desired
 - b. Options can be deleted, modified or added
 - c. Commands can be deleted, modified or added
4. The modified netlist file, containing either .OP, .AC or .TRAN command, is saved as a *tmp.net*.
5. The LTspice binary is called in a batch mode to execute that *tmp.net* and to save the results in a *tmp.raw* file.
6. *LTspice2MATLAB()*⁸ function is used to retrieve data from *tmp.raw* into MATLAB workspace.

⁵ With full understanding of the limitations of this method as being a pure speculation till real-life results are obtained and verified to a degree agreed as sufficient.

⁶ <https://www.diyaudio.com/>, <https://www.audiosciencereview.com/forum/index.php>

⁷ <https://groups.io/g/LTspice>

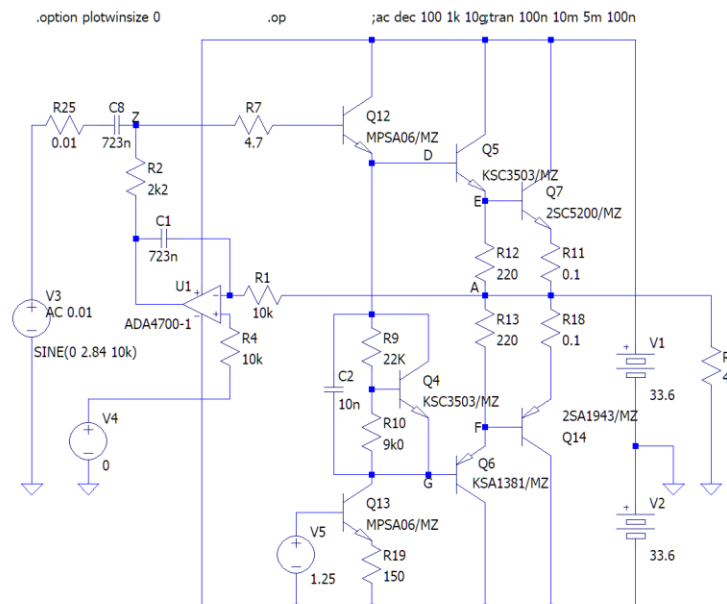
⁸ <https://github.com/PeterFeicht/ltspice2matlab>. Copyright (c) 2009, Paul Wagner. Copyright (c) 2019, Peter Feichtinger. All rights reserved by the authors. See license file for details.

7. Steps 2...6 can be repeated in loops to collect the necessary data, such as repeat simulation for different I_Q , different amplitudes of Input signal, different frequencies, different values of particular components, different schematics, etc.
8. The collected data is analysed. For each run, we collect at least 100 periods on $f(x)$, remove long-term trends if any, and visually check that all 100 periods are essentially the same. Then data plotted in a few of [hopefully] easy-to-understand ways in both time domain (TD) and frequency domain (FD).

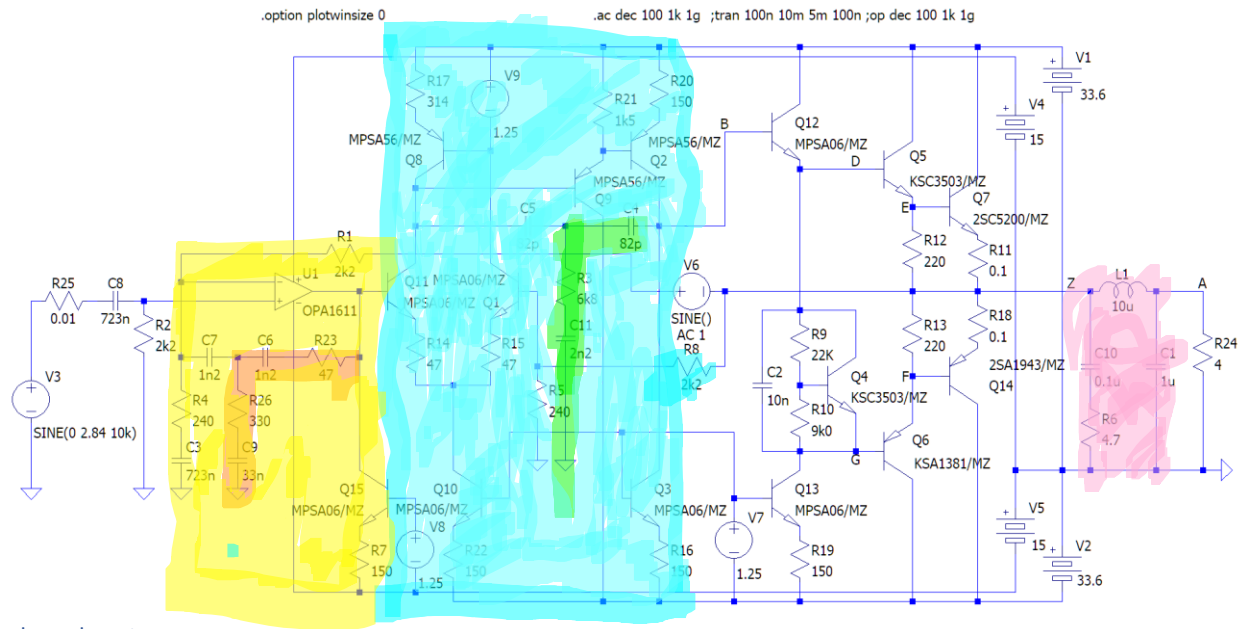
A “Six-Step” Design Template

...is used to analyse different schematics:

0. An output EF cascade is set as a reference point. R10 is used to control I_Q of the output cascade. V4 sets an arbitrary bias.



1. Adding the first layer of “standard” feedback control. The op-amp could be either monolithic or discrete. The target voltage gain is 20dB ($1 + 2k2/240$). Phase margin $\geq 60^\circ$.
2. Improving the standard feedback control with a second order astatism, gain plan restructuring, etc
3. Introducing nested (more precisely, nesting) feedback loop
4. Improving the nested feedback loop, usually by 2nd degree astatism with $C6 =$ doubled previous $C7$ and $R26 + C9$ chain. The $\tau_2 = (C_6^{-1} + C_7^{-1})^{-1} R_{26}$ shall be $\geq 5/\omega_T$ to maintain a safe 60° phase margin.
5. As it was noted, feedback can not eliminate crossover distortions but can push them towards higher and higher frequencies (like in noise shaping). Then, we can suppress the remnants of crossover distortions by adding an LC passive 50kHz low-pass filter, essentially the same as for Class D amplifiers, on output.



Class ltspace.m

... provides low-level generic functionality such as

- Read and write .sp / .net expanded netlist file
- A component value get /set
- An option get / set / delete
- A command set / delete / modify
- Process a netlist by LTspice: .OP, .AC or .TRAN
- Read a variable (such as node voltage or current) or time / frequency
- Bode plot for one or several variants
- Plot THD in both time and frequency domains for one or several variants
- Logging / self-test / utilities / etc

Class amp_ab.m

... provides higher-level functionality specific for Class AB audio amplifiers with a few assumptions, as $R10$ (8k...12k) is used to control I_q , certain node naming, etc:

- Find out which values of $R10$ correspond to the chosen set of varying $I_q = [5.0; 10.0; 20.0; 40.0; 80.0; 160.0]$ mA by inverse spline resampling.
- Plot linearized Z_{IN} for all stages of EF, various biases from Vee to Vcc and various I_q
- Linearized Bode plot for chosen nodes and conditions
- Plot detailed THD in time and frequency domains with intermediate results
- Plot THD in time and frequency domains, including 3D mesh plots whenever appropriate for
 - ✓ A set of varying component values,

- ✓ A set of varying I_Q ,
- ✓ A set of varying amplitudes of input signal,
- ✓ A set of varying frequencies of input signal
- ✓ A set of different schematics.

The scripts `test_ltspice.m` and `test_ab.m` contain [presumably typical] use cases for these classes.

Notes

1. General advices⁹ on how to get repeatable and reliable LTspice simulations are very helpful except that you start looking for them only after you realize you are up to ears in surreal troubles. It turns out that explicitly setting `“.option plotwinsize 0”` is the only way to prevent LTspice from saving data in a badly compressed format.
2. It also turns out that `“.option numdgt 10”` (or anything higher than 6), which shall force LTspice to save data with double precision, renders the `.raw` file unreadable by `LTspice2MATLAB()`.
3. The non-linear circuitry naturally has several DC solutions, and sometimes LTspice converges into a wrong or no solution. Unfortunately, we did not find a way to hint LTspice towards the operating point we care about. Alas, some otherwise valid circuits can not be simulated no matter what.
4. Usually, transistor models are ... despicable, with omitted confidence intervals. It takes quite an effort to find a suitable tested model and then clean it up so that the simulation results do not contradict the most basic university-level physics and do not exploit weird unrepeatable performance tricks. Such models were added to the *standard.bjt* as `*/MZ`. Vendor-provided models of TI op-amps are actually quite nice.
5. The LTspice computation engine itself is beyond reproach. Still, it's hard to believe that a “normal” EE without deep knowledge of numerical methods and signal processing nuances can figure out how to use it properly.
6. Both alternative and normal solvers with various thresholds and tolerances have been used, to avoid making conclusions based on numerical artifacts. However, that does not guarantee error-free results.
7. The `tstep` and `tmax` parameters of `.TRAN` must be manually set up and re-checked to agree with the actual minimal step used by the LTspice engine.
8. The raw signal with variable timing is resampled to the standard uniform PCM by `interp1(..., 'spline')` function which is wrong but simple. The correct way is to use Lennart Ljung's “System Identification Toolbox” kernel-based spline interpolation but such fine instruments can not be used by uninitiated.

⁹ Such as <https://gist.github.com/turingbirds/c90672c3b126d0d5f37f90494d5057cb>

9. Instead of plotting spectra with noisy & messy holes between harmonics, we plot only the envelopes of the spectra, taken on the $k*f_x$, separately for odd and even k , using solid and dashed lines respectively.
10. To shorten simulation time and eliminate possible computational artifacts, everything possible and impossible is simplified as much as possible and beyond:
 - a. Instead of a realistic voltage source like a couple of diodes or REF2912, a simple ideal 1.25V voltage source is used.
 - b. Batteries are assumed to be ideal.
 - c. The high-pass capacitors are intentionally chosen to be a way too low, with funny nominals like 723nF - to shorten transient time. The high-pass frequency grows to about 1kHz which is ok for us.
 - d. $R_6=40\Omega$ is a standard but also unrealistic purely active load.
11. We consider 100W as a sufficiently high full-band peak power needed for home audio. For average listening distance of 3m, stereo, average listening level of 80...85dB SPL, loudspeaker sensitivity of 92 dB SPL/W/1m, we need about 0.5W of sustained, thermal power. Peak to average ratio varies wildly for different styles of music, from 8dB for heavy metal to 26 for jazz and classical. The real peak-to-average dynamics at live concerts are even often higher – but they are usually compressed during mastering sessions due to psychoacoustics-based differences between at-home and at-concert listening. 100W correspond to SPL about 105dB.
12. The test signal is set as a single frequency sine wave at 20kHz, with amplitude varying from 0 all the way to 100W i.e., 28.3V. As we concentrate on Dirac-styled crossover distortions, there is little if any sense to use IMD or any other composite signal because they make the distortion analysis much more complicated with unclear if any benefits. Here, we do not care if an amplifier saturates properly or not.
13. The precision is hard-limited by floating point [23-bit mantissa] resolution in *.raw file to about $23*6.02 \cong 140\text{dB}$, or 10^{-7} , or 0.00001%. The low THD numbers listed below depend on the particular choice of components, their models, solver, thresholds and tolerances and therefore have a somewhat symbolic sense. However (again), we compute not for the sake of getting a number but for the sake of improving our understanding¹⁰ of what is going on inside a nested feedback control loop of a system with hard non-linearities.

Reproducibility

- ✓ LTspice version XVII. Unfortunately, I did not find a place where LTspice saves its configuration data.

¹⁰ Richard Hamming “Numerical Methods for Scientists and Engineers”, originally published: New York: McGraw-Hill, 1973.

- ✓ MATLAB R2019b¹¹ and Signal Processing Toolbox version 8.3.
- ✓ Schematics *.asc included, netlists *.sp are redundant but also included
- ✓ MATLAB scripts *.m included. The code does not use any tricks nor undocumented features and written in the most primitive way possible.
- ✓ A .bjt file with transistor models used included. It shall be manually copy-pasted to the C:\Users\your_name\Documents\LTspiceXVII\lib\cmp\standard.bjt
- ✓ The .asy files for op-amps included. They shall be manually copied into C:\Users\your_name\Documents\LTspiceXVII\lib\sym\OpAmps\.
- ✓ The copyrighted models of op-amps / components shall be obtained from their corresponding vendors and copied into C:\Users\your_name\Documents\LTspiceXVII\lib\sub\. Some are included here for convenience only and will be removed without a warning shall a vendor object.

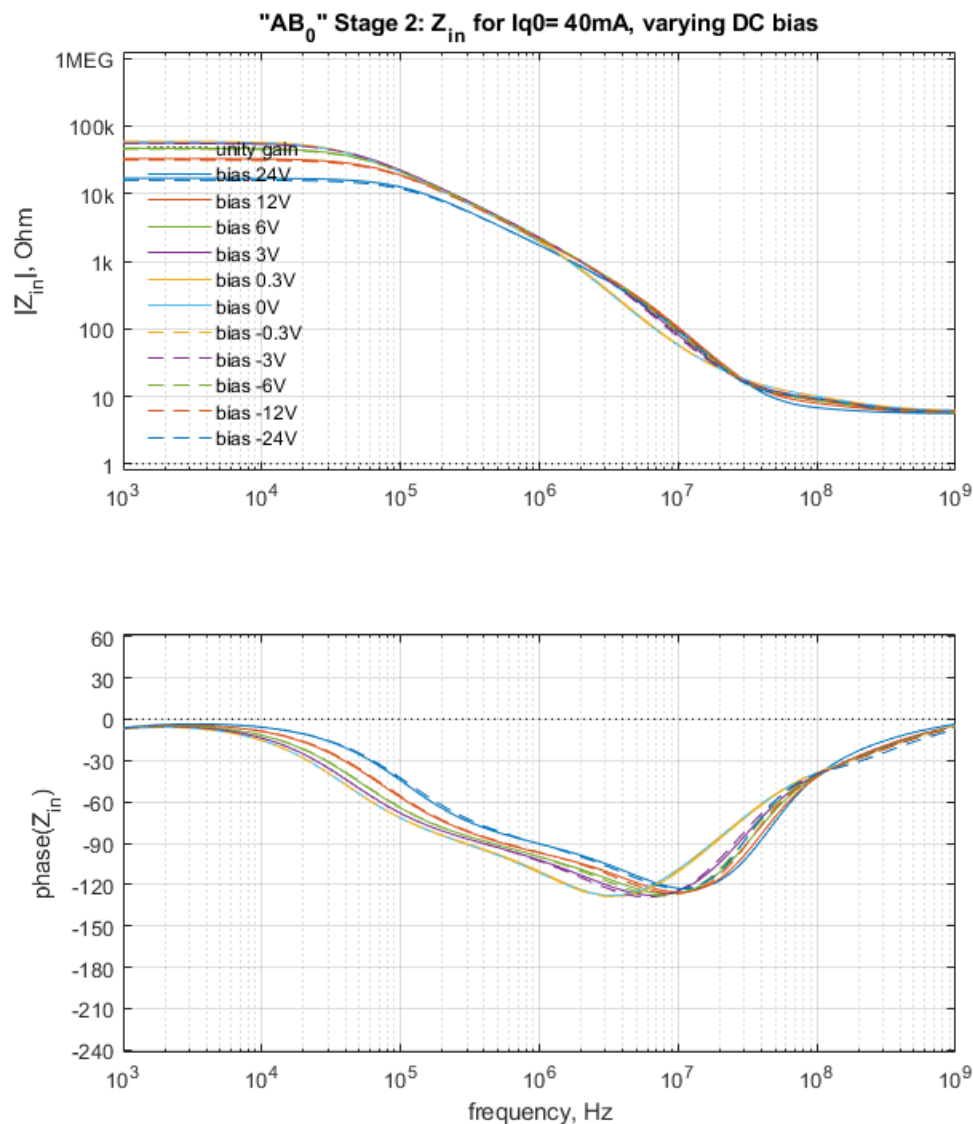
¹¹ DIYers: since 2016, MATLAB provides “home license” for non-professional use, which can NOT (!) be used for creating a commercial product, for a very low price. It’s a way better deal than any free clones.

RESULTS

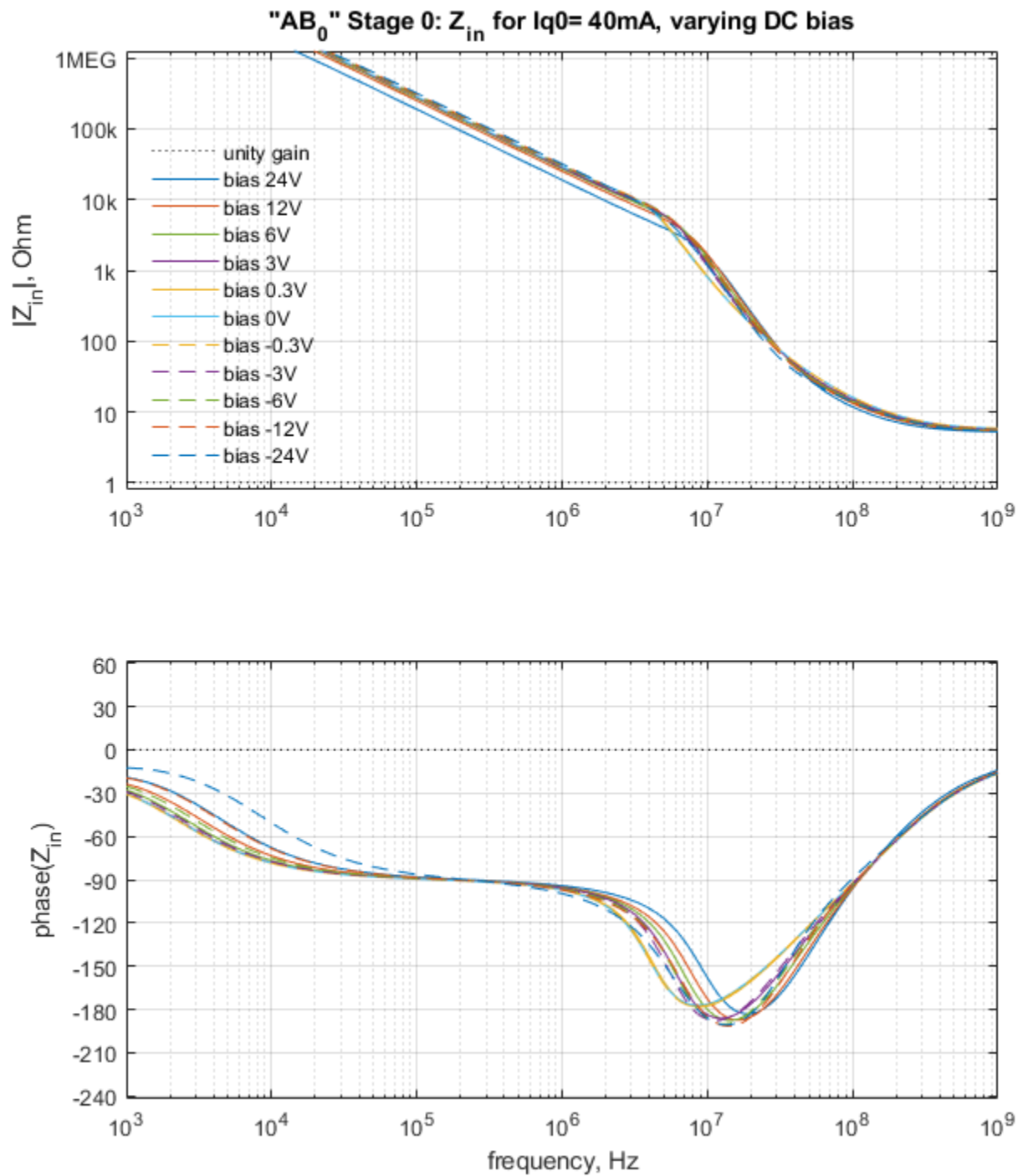
Three stage EF

EF Input Impedance

Input Impedance of the EF output 2nd stage varies a lot, from low $\sim 15\text{k}\Omega$ at rails to about $70\text{k}\Omega$ at zero, and input capacitance is about 100p (also non-linear). The output stage serves as the load for pre-amp (VAS = Voltage Amplification Stage in D. Self terminology), so the VAS gain varies too, introducing undesirable odd and even order distortions.

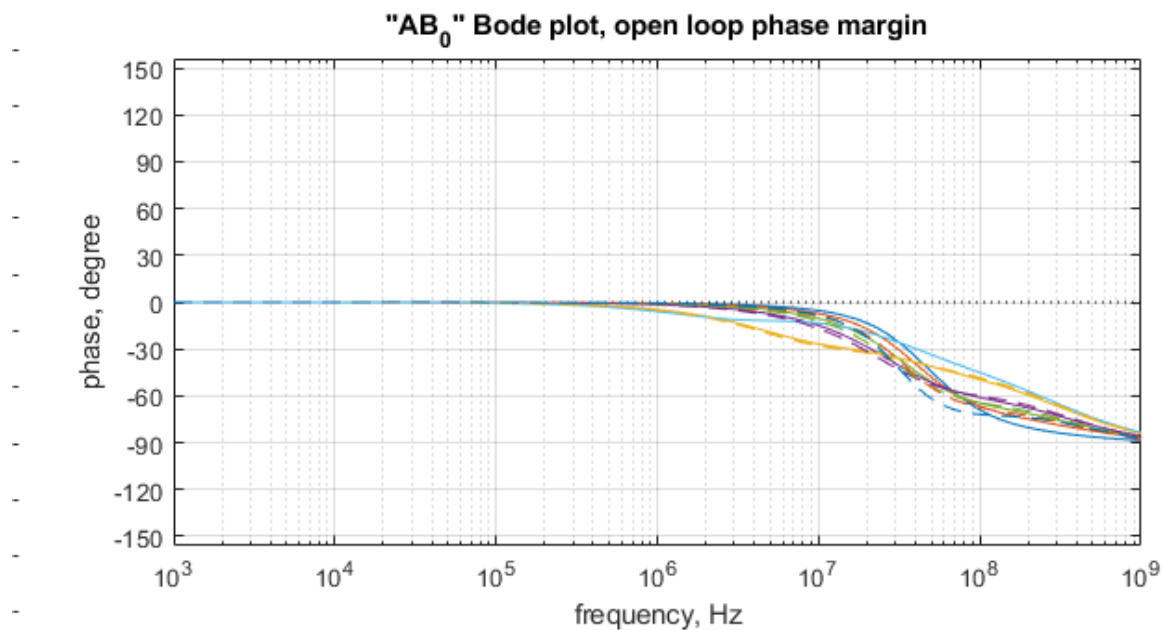
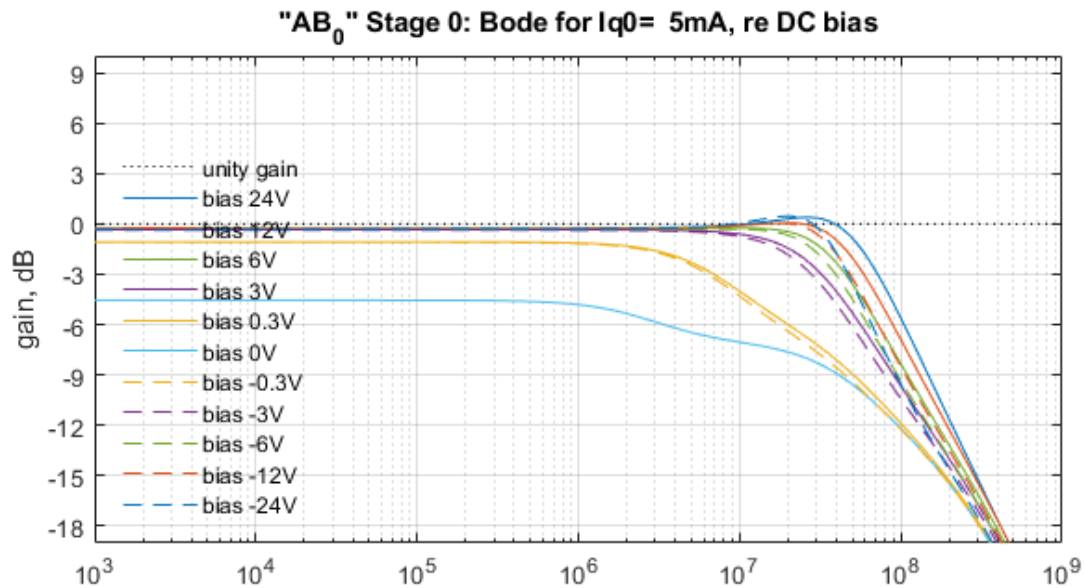


The same happens for all I_q . It's preferable to add a 3rd stage of EF to make Z_{in} higher (so that it matters less) and more invariant to bias. Then, the EF input impedance becomes essentially equal to the base-collector capacitance of Q_{12} and no longer determines the VAS pre-amp gain.



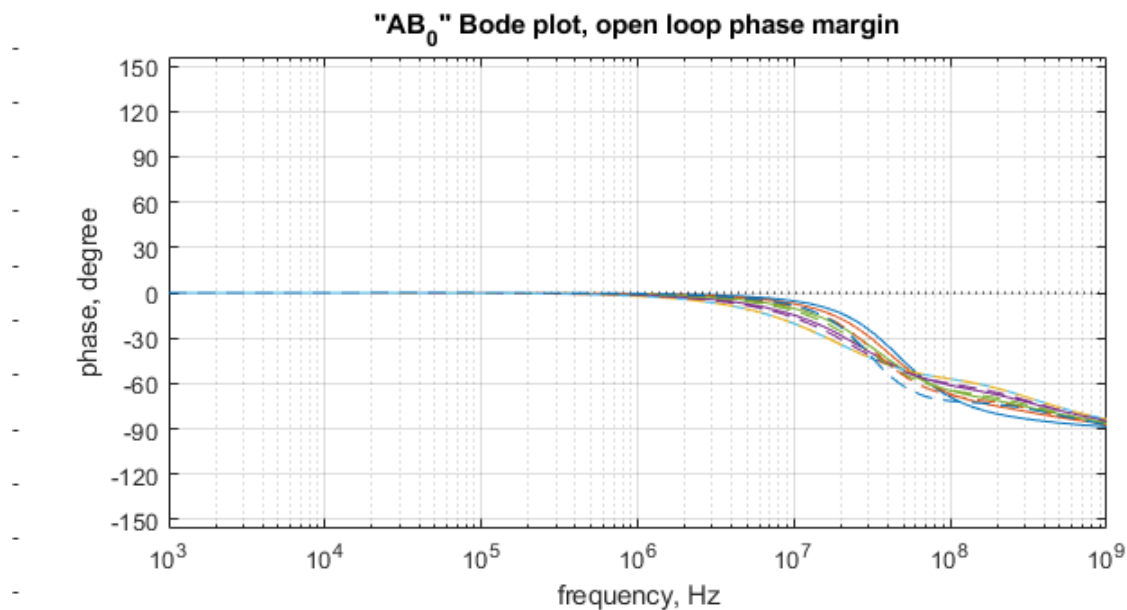
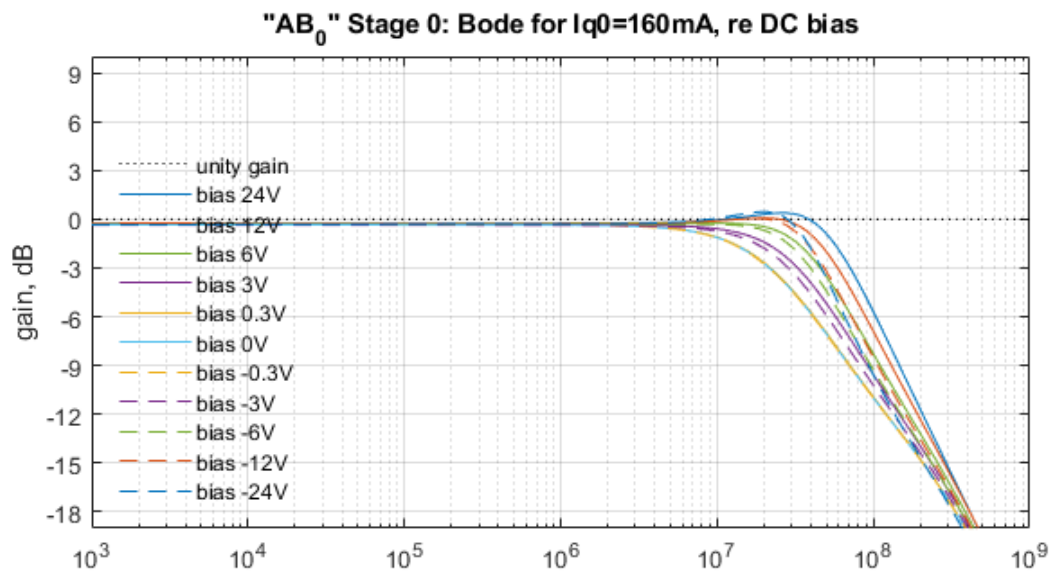
EF Bode

Transfer function of EF heavily depends on the bias, especially for low $I_{q0} = 5\text{mA}$, and higher frequencies.



On 10MHz, the phase varies between -30° and 0° - which forces the preamp to have GBP f_T to be safely lower than 10MHz, say, maximum 3...5MHz, to push that unpredictability and uncontrollability down below unity gain and ensure that the phase margin at f_T is at least 60° . The -4dB drop at zero bias shall

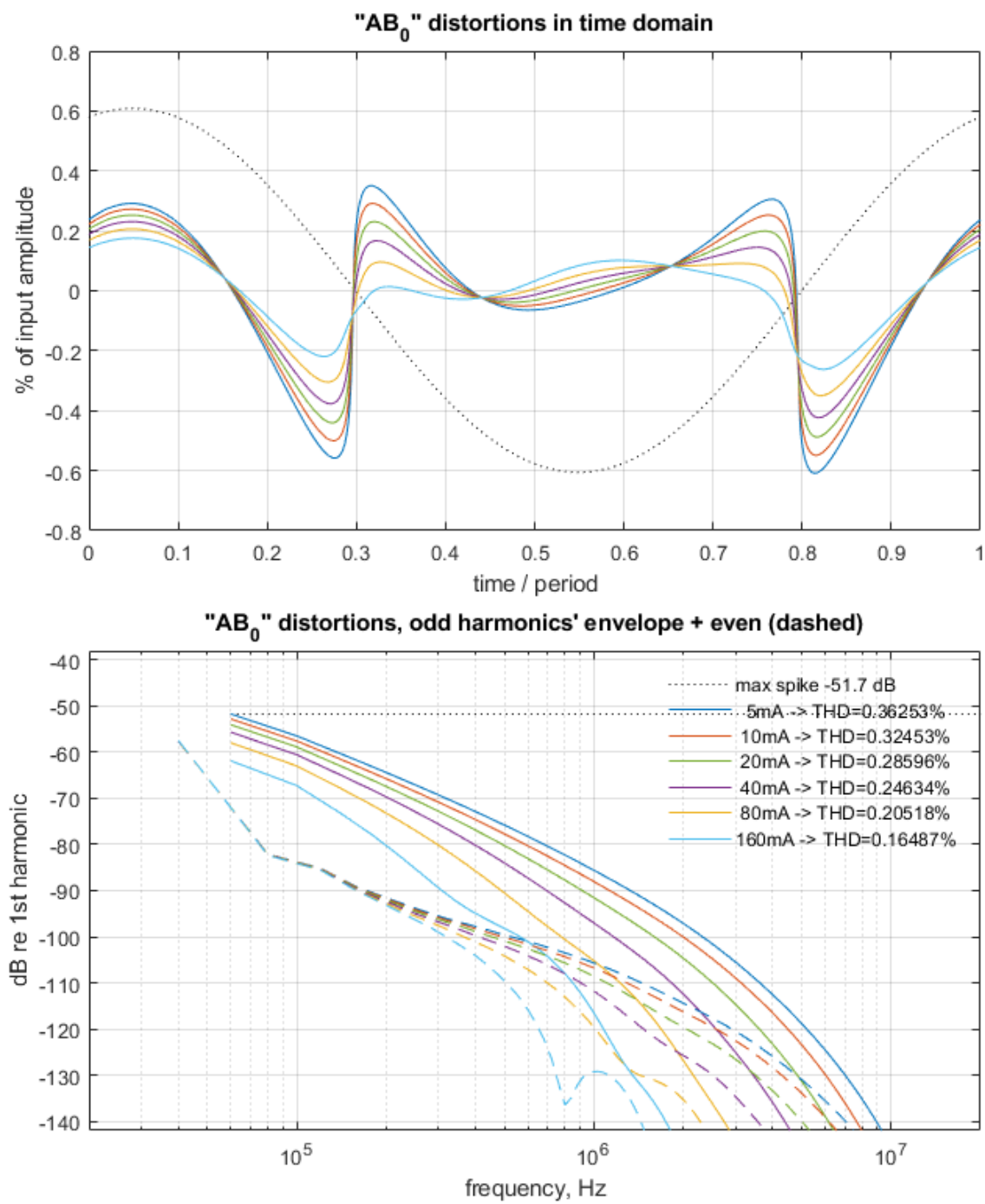
be properly accommodated when 2nd order astatism is used. Of course, it's much smoother for $I_q = 160\text{mA}$.



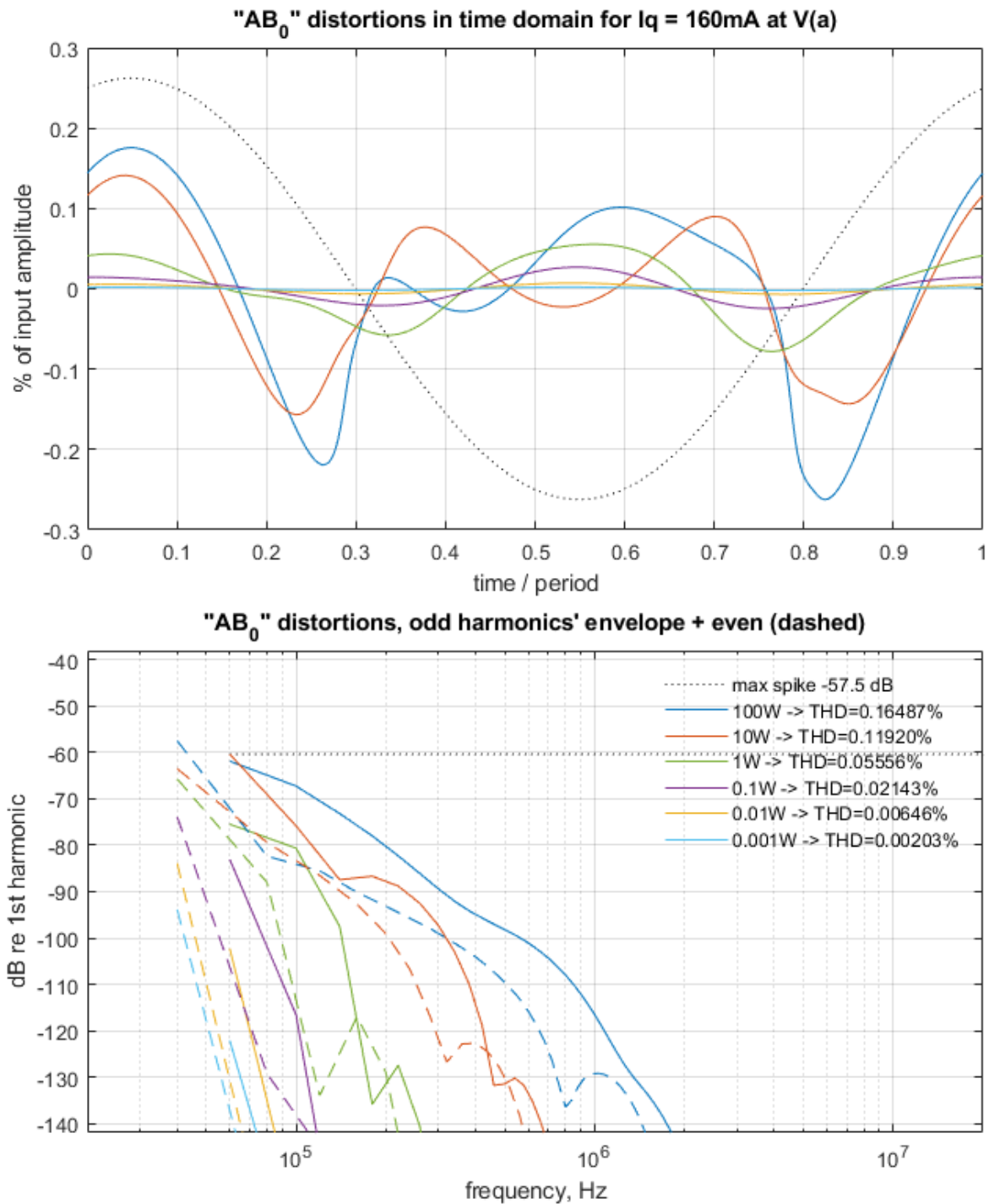
EF THD for various I_q , max amplitude

The distortions differ mainly by their spectra width ($\sim 8\times$) rather than by amplitude ($\sim 2\times$). The shape of distortions may look weird but they are essentially the same distortions as plotted before - with

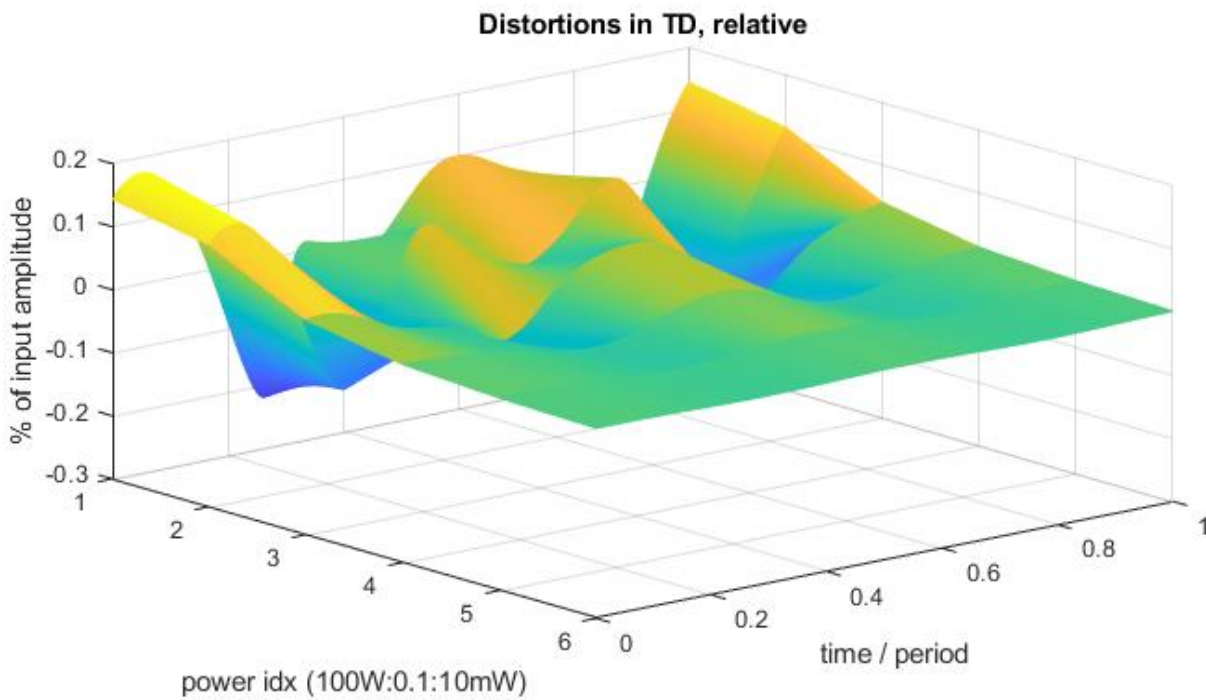
fundamental harmonic zeroed.



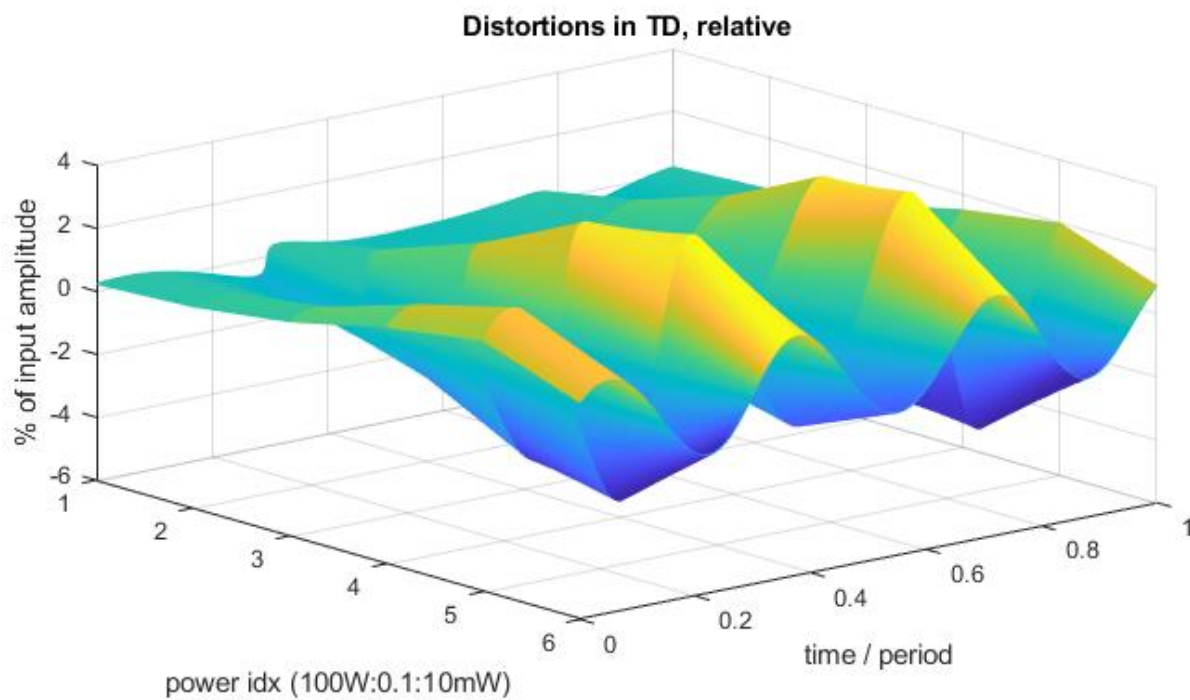
EF THD for varying amplitude, $I_q = 160\text{mA}$ (almost Class A)

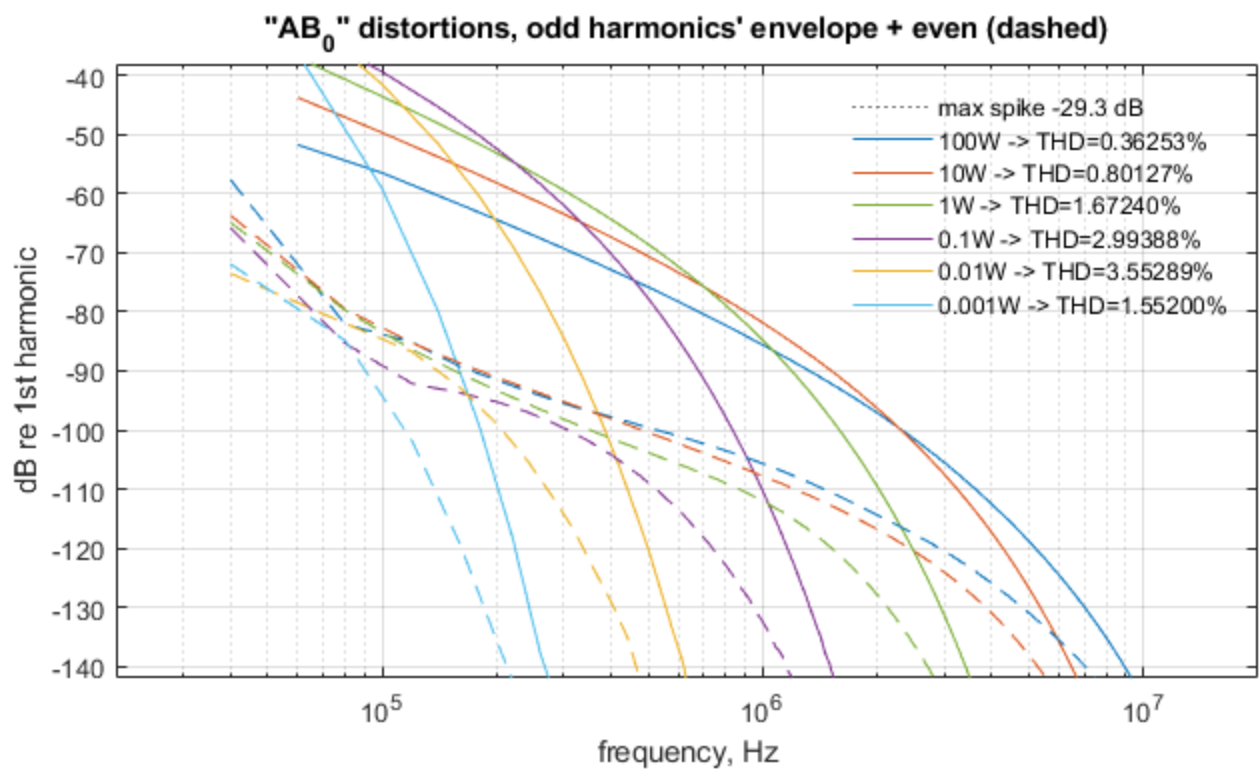
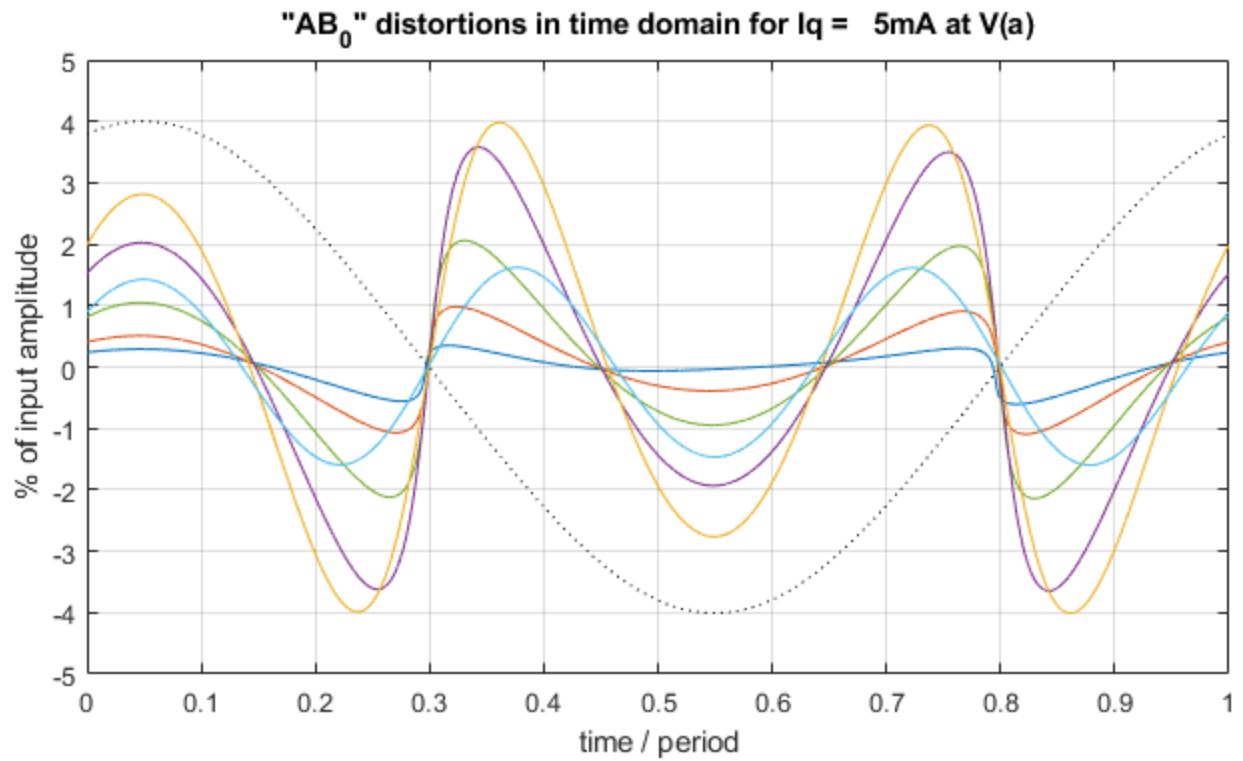


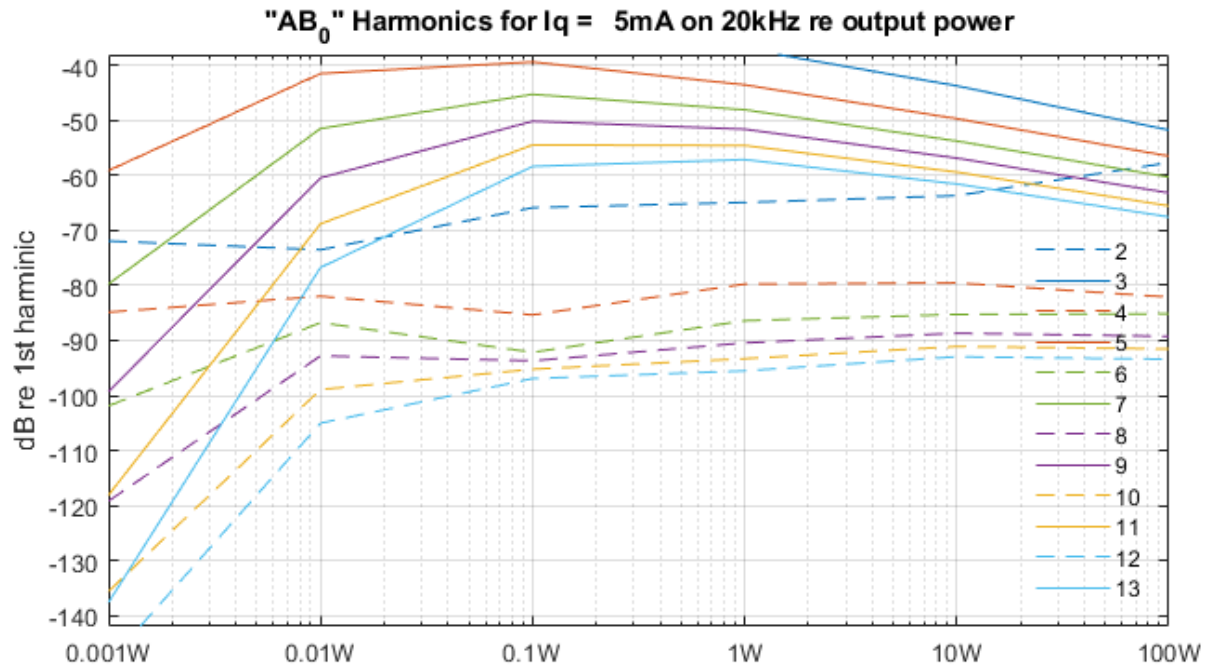
Again, amplitude affects the width of the distortions' spectra much more than its peak value.



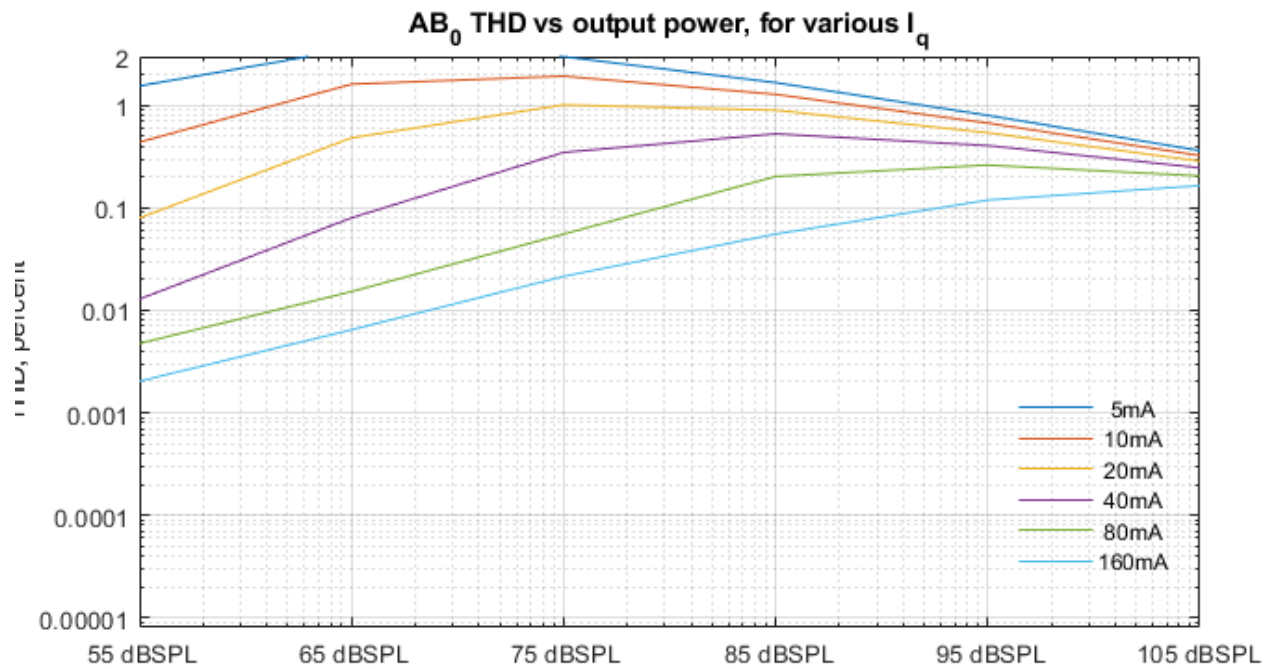
EF THD for varying amplitude, $I_q = 5\text{mA}$ (almost Class B)



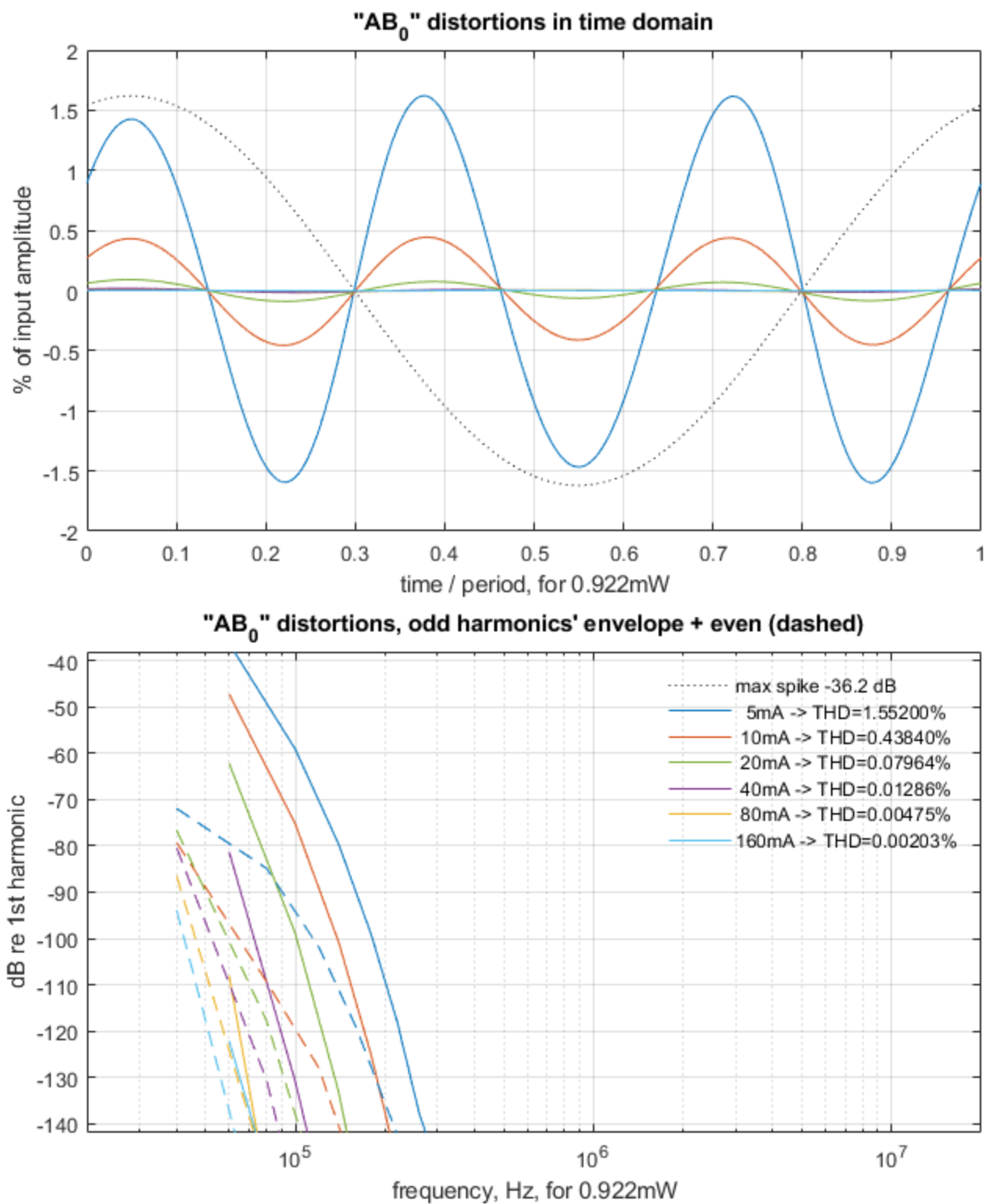




When output decreases down to 100mW (about 1V), the harmonics increase, the lower is I_q and the closer EF resembles Class B.

EF THD Amplitude – I_q Map

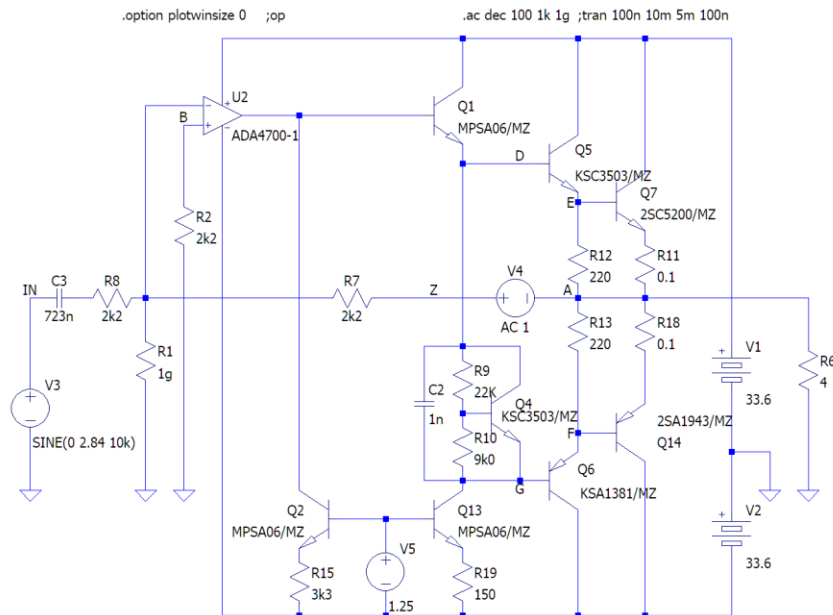
As expected, higher I_q lowers distortions. The main difference between distortions on 100W and on 1mW is not in the THD number but in the width of the distortion spectra and the decay speed of harmonics. On 1mW, it looks like a Class A for any I_q.



The effects of variations due to changes in component values for EF are not very interesting and left for the curious readers.

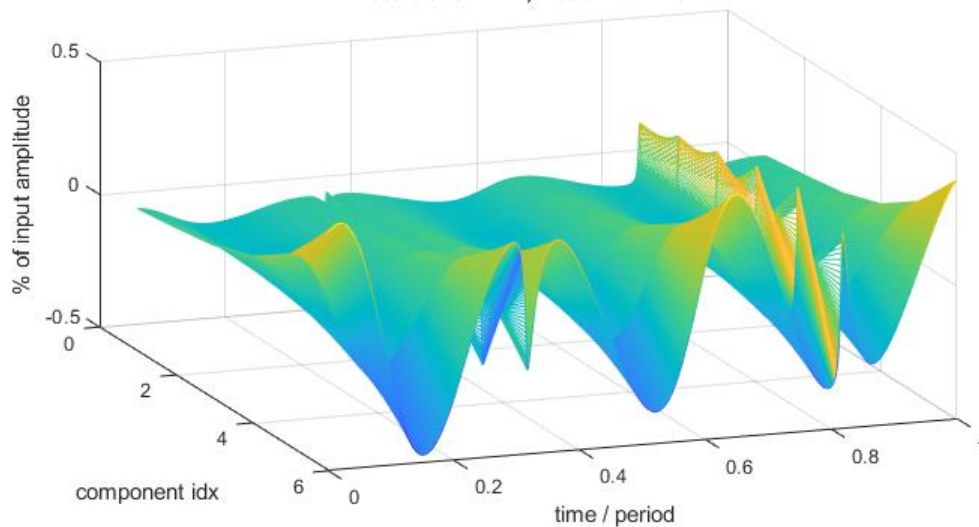
Total Harmonic Distortion “Measurements”

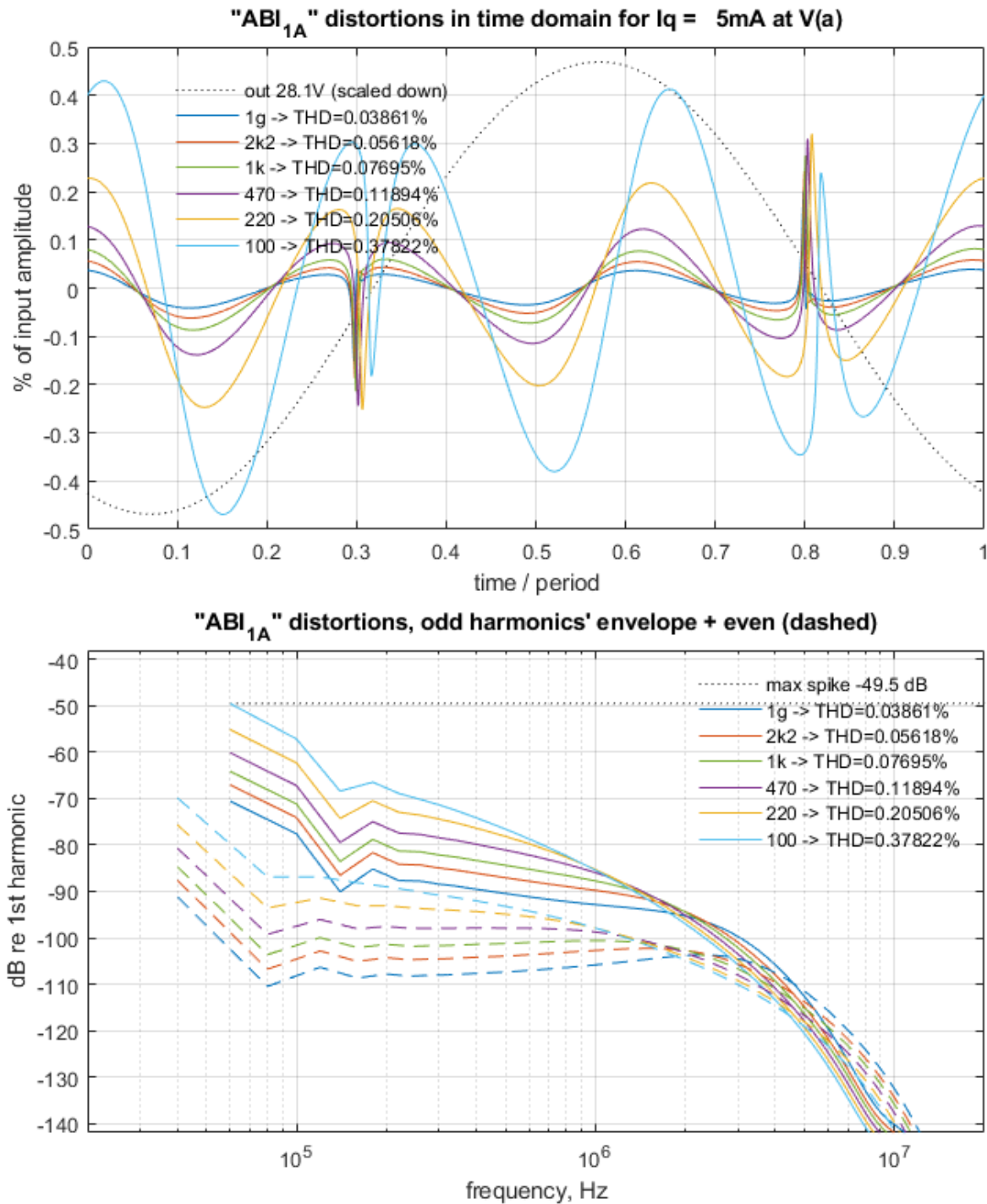
The wide spread belief that feedback reduces the distortions appears to be logical and mathematically sound. Many sources describe it like “...The closed-loop gain is unchanged, but the feedback available for error correction is reduced by the factor of $X=1+R_7/(R_8 \parallel R_1)$, thus reducing the resolution by X. Validity of this technique can be verified at high gain and/or high frequency...”. When you follow this line of thought you can’t find anything to object. However, let’s simulate an applicable schematic below for $R1 = \{ '1g', '2k2', '1k', '470', '220', '100' \}$; for $I_q=5mA$, 28.3V input on 20kHz.



As you have already guessed, the results do not look as nice as we could have expected... not because the ADA4700-1 is a bad op-amp (which is not true) or something like that.

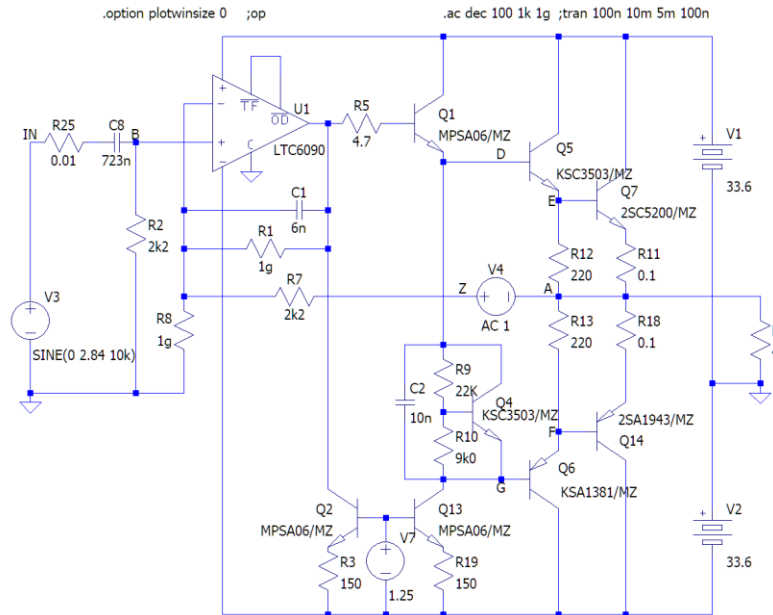
Distortions in TD, visualized in 3D



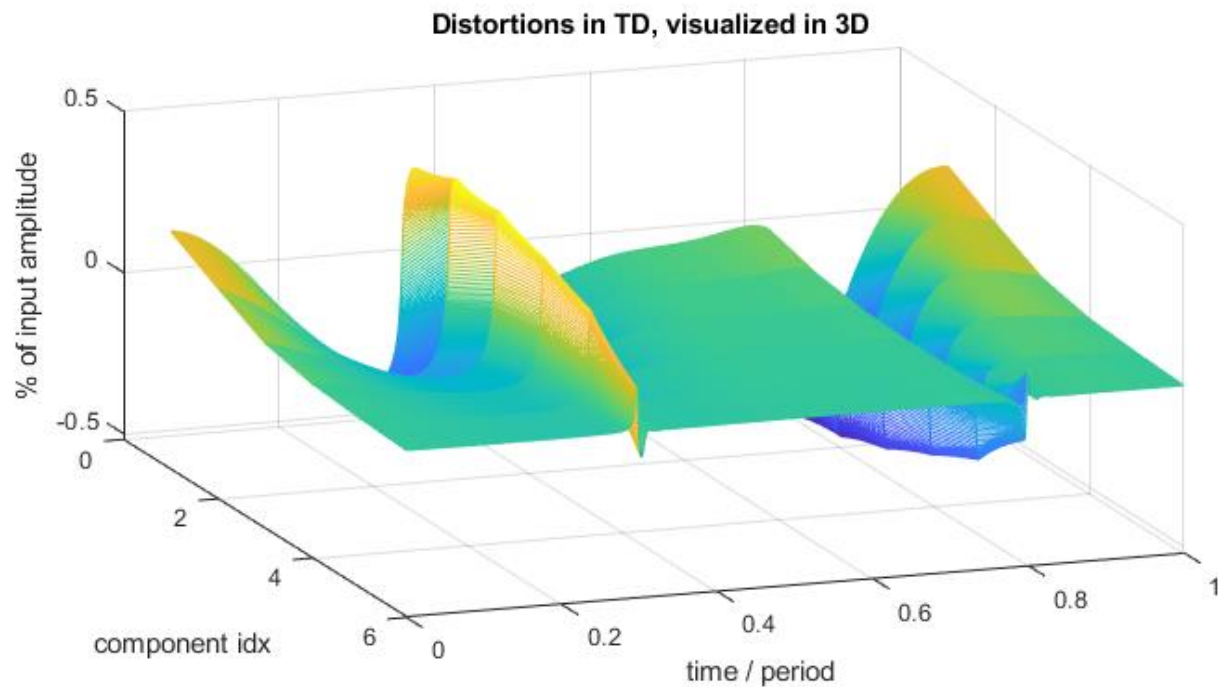


The problem is in that line of thought itself, which has an implicit assumption of a compact kernel, i.e., of "soft" distortions whose spectrum is narrow and contained completely within the amplifier passband. If that assumption were true, neither the shape of distortions nor the spectrum of distortions would change. Obviously, it's not true for "hard" wide-spectrum distortions. Well, it's nice to understand and examine our assumptions, both explicit and implicit, before jumping to any conclusions. Sometimes.

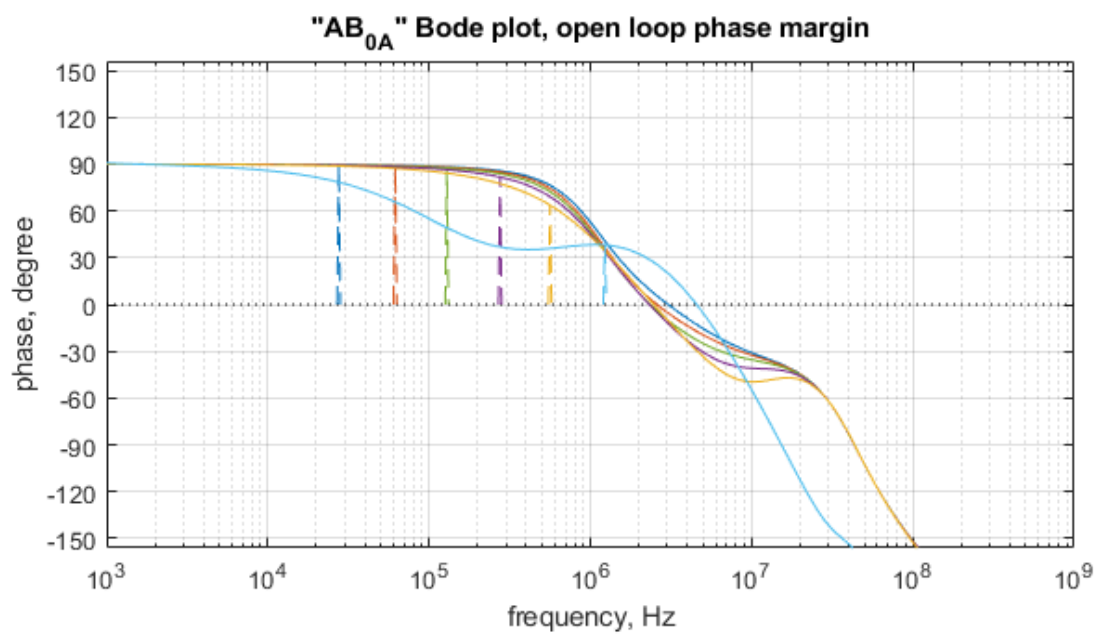
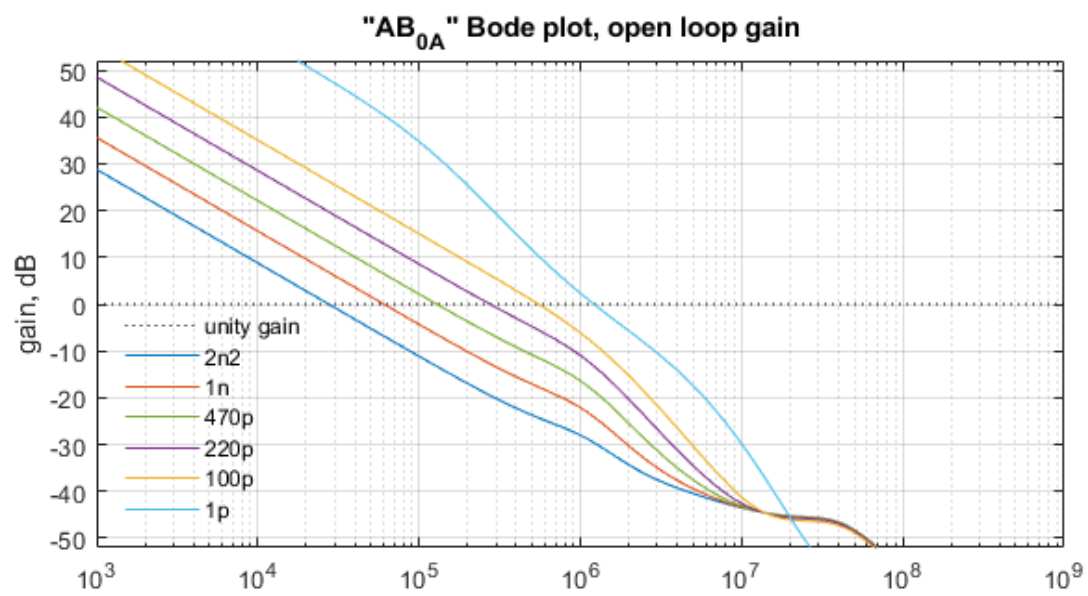
A simple feedback loop over EF (AB_{0A})



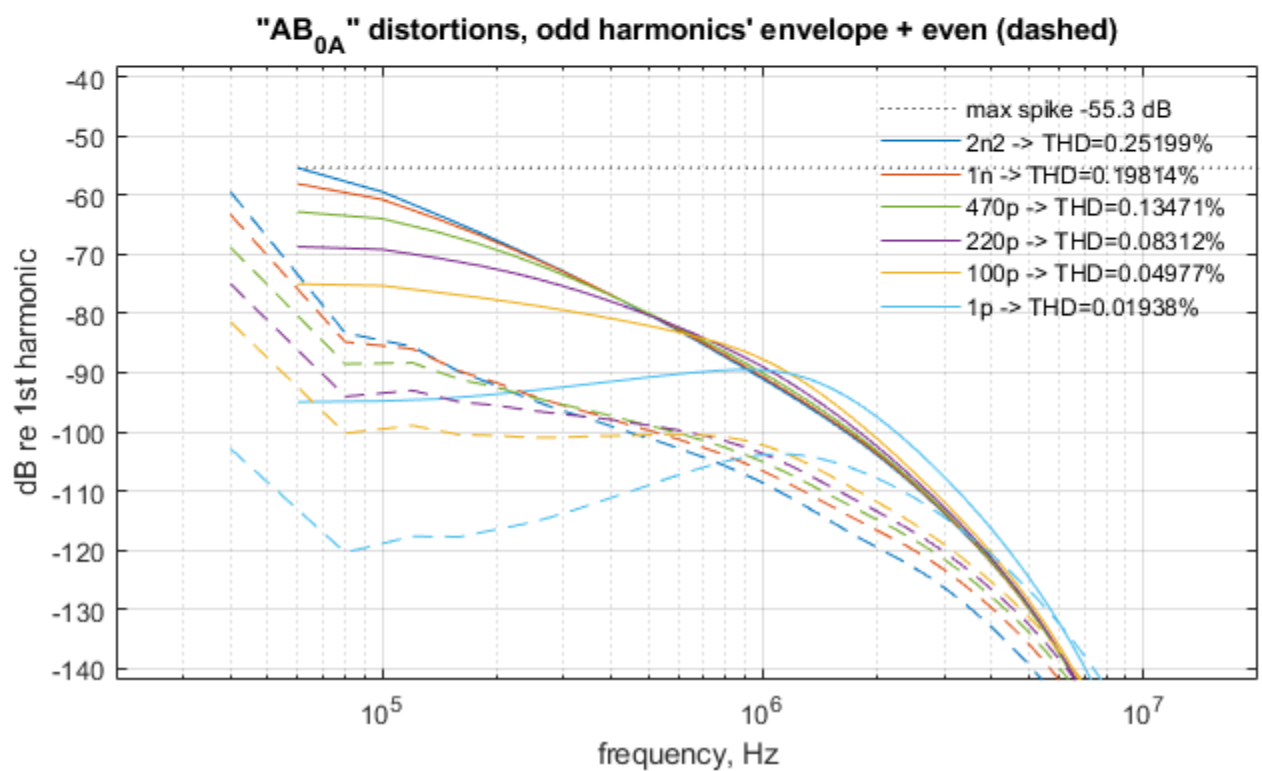
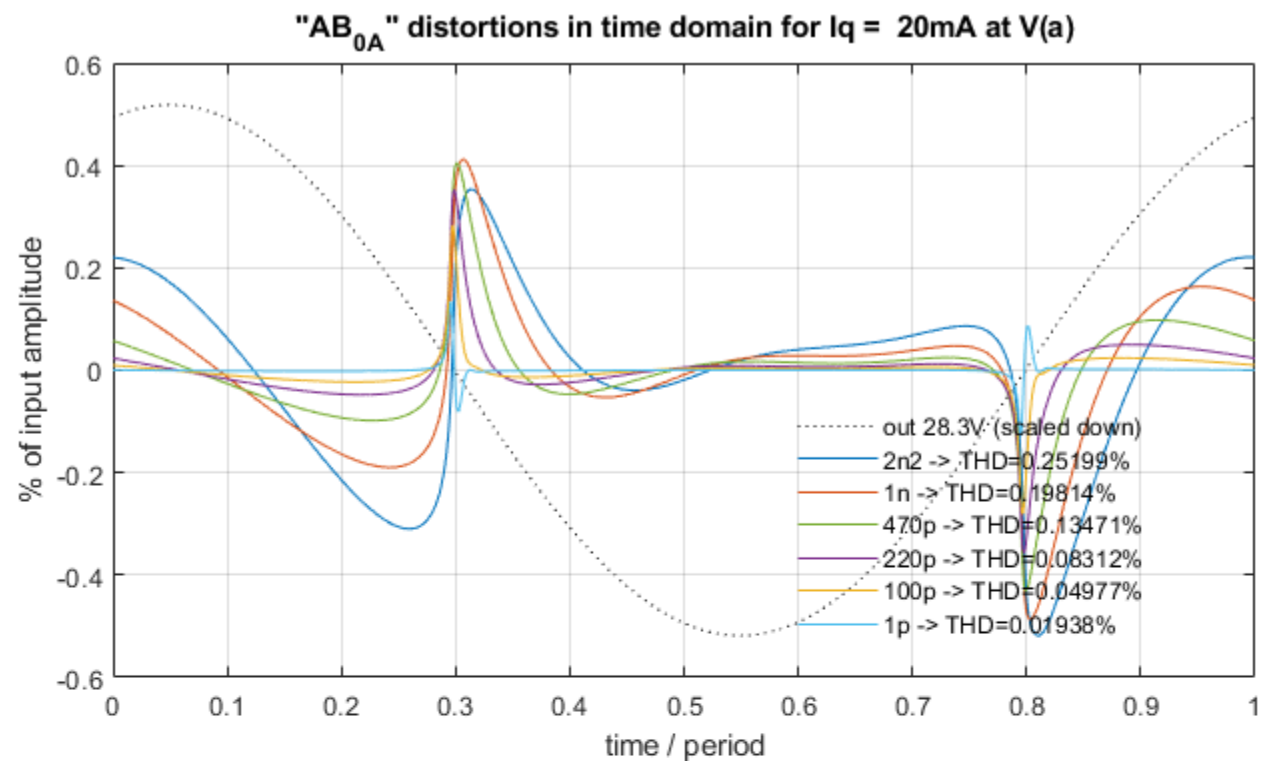
We can adjust feedback loop GBP by changing $C1 = \{ '2n2', '1n', '470p', '220p', '100p', '1p' \}$, and observe the corresponding changes in non-linear distortions.



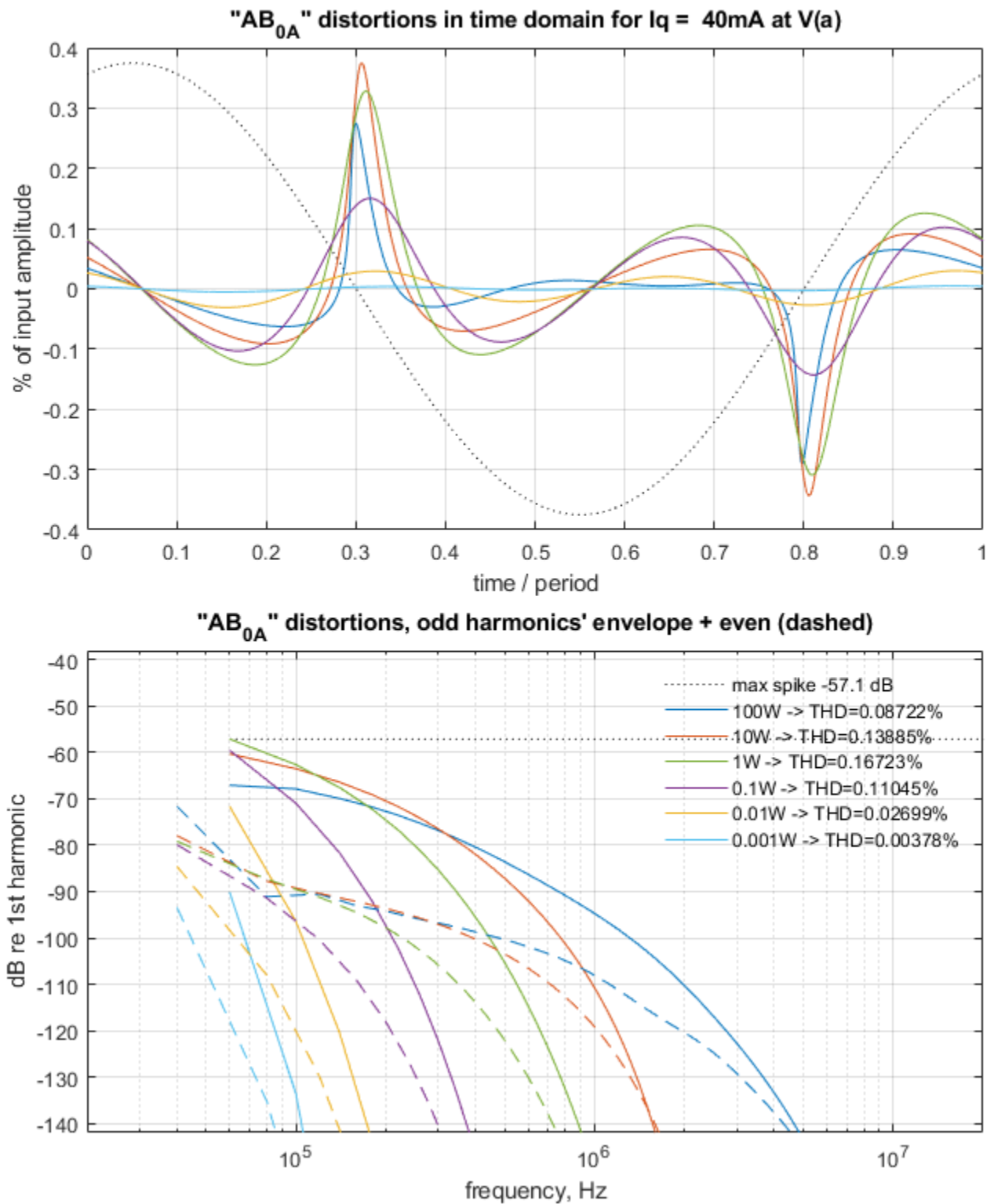
Bode Plots



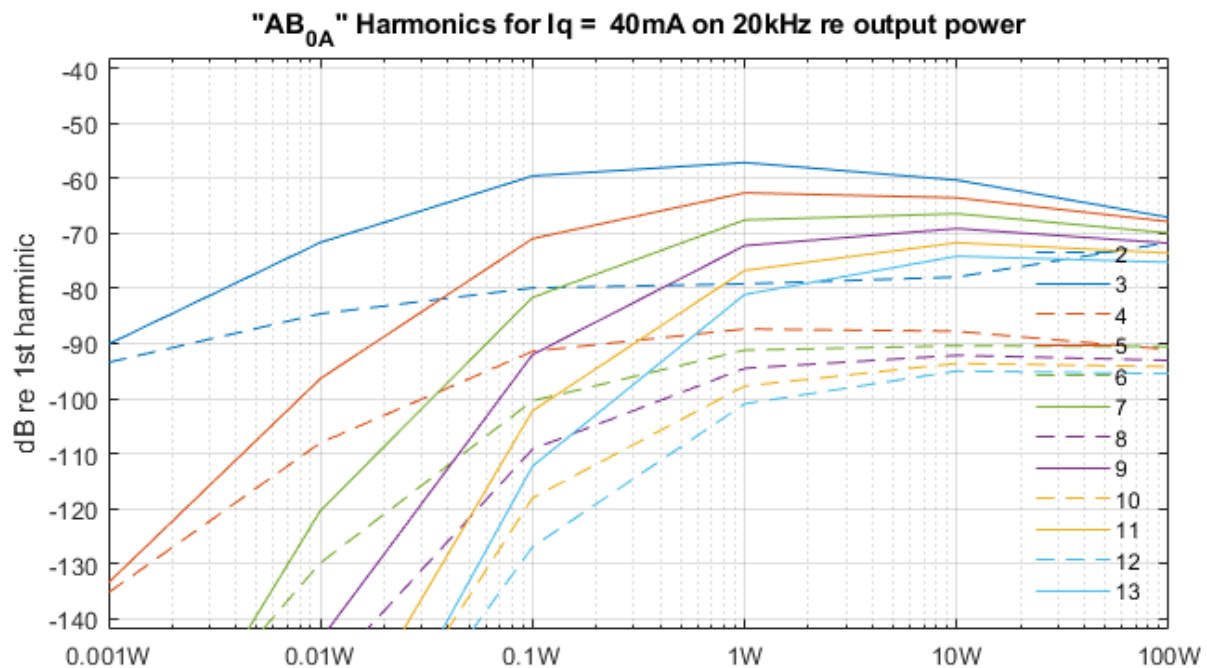
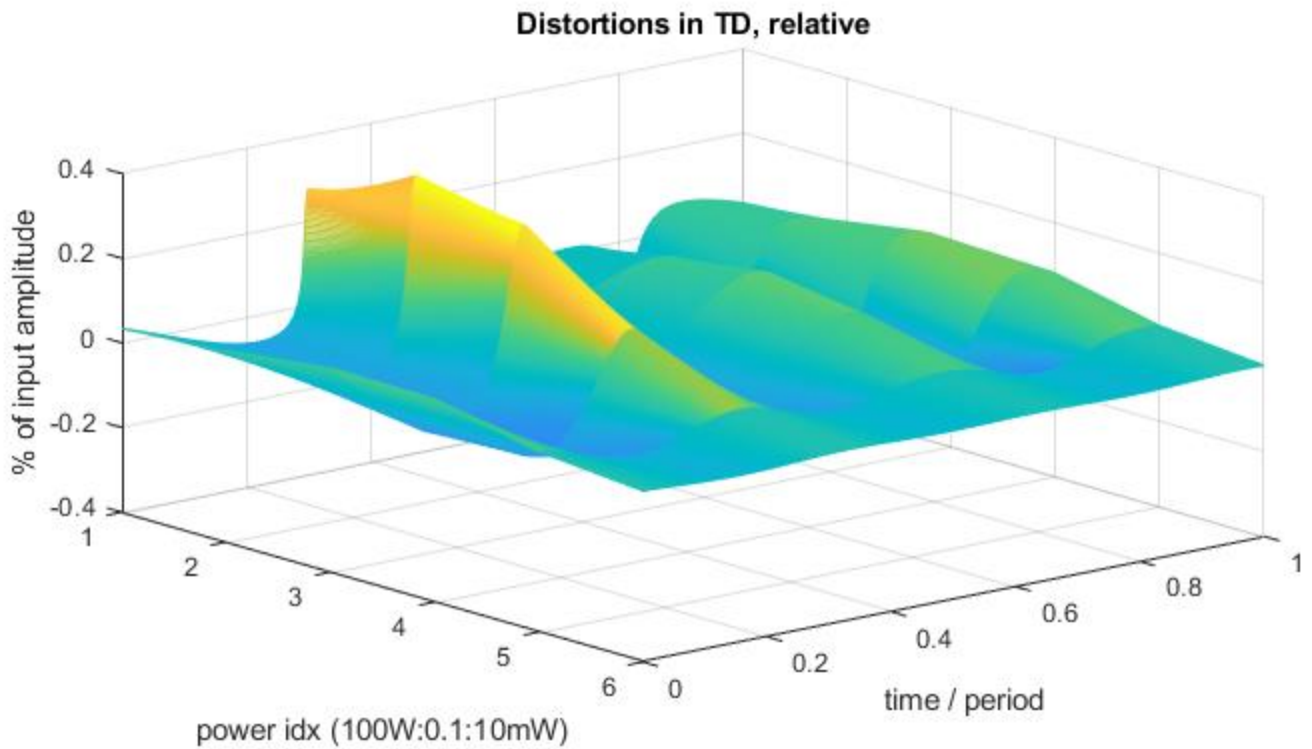
THD Plots



THD for varying amplitude, $I_q = 40\text{mA}$, $C_1 = 330\text{p}$ ($f_T \cong 200\text{kHz}$)

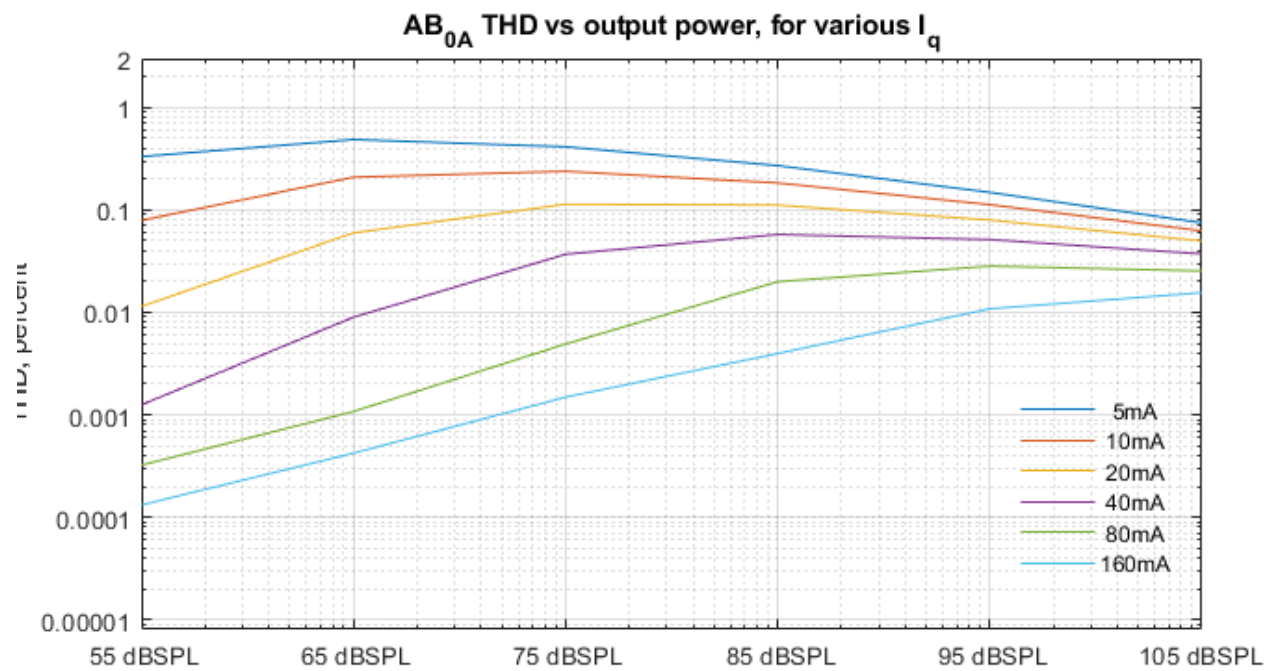


A typical monolithic Class AB amplifier, like LM3886. Of course, $\text{THD} \sim f(x_{\text{IN}})$ for 1st-order astatism.

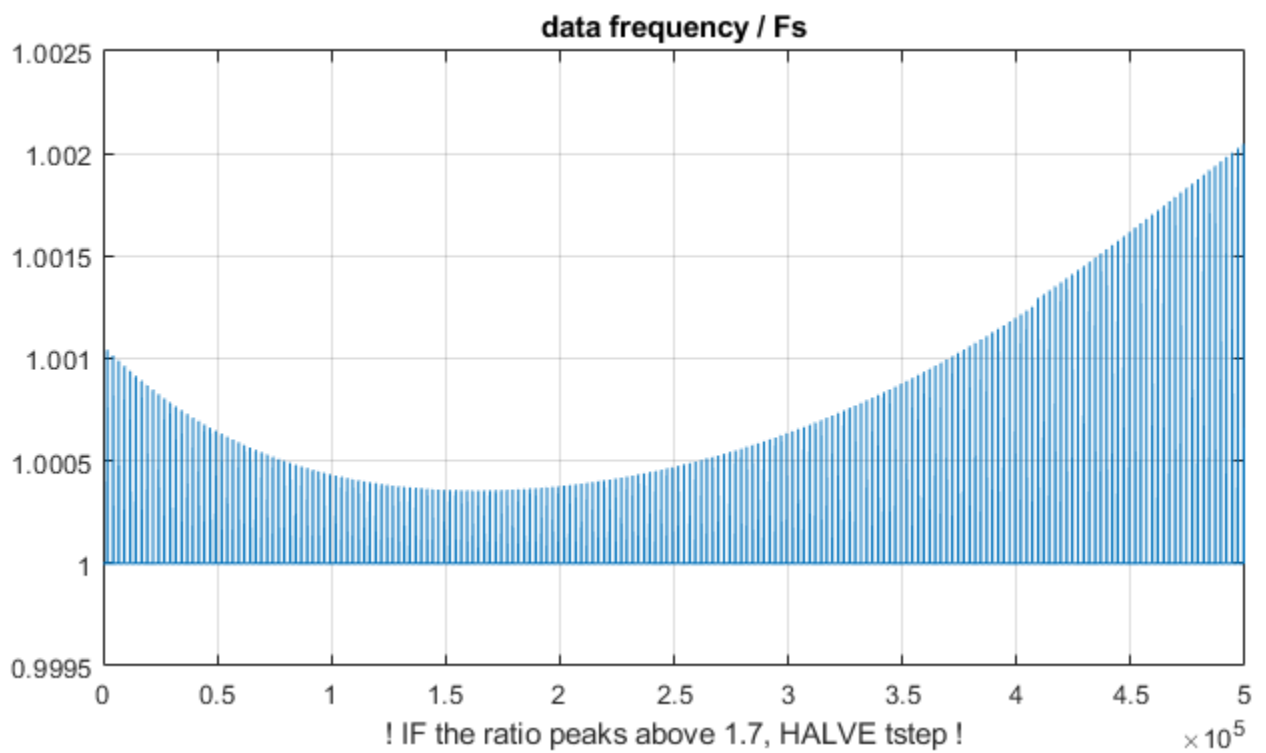


For “soft” non-linearity the amplitude of n^{th} harmonic of output for the $x_{IN}=a*\exp(-j\omega t)$ excitation shall in/decrease as a^{n-1} . Here we see that the amplitude of the first few harmonics does not change much with output amplitude. “Soft” rules come into play only for <10mW output power.

THD Amplitude – I_q Map for $C_1=100p$

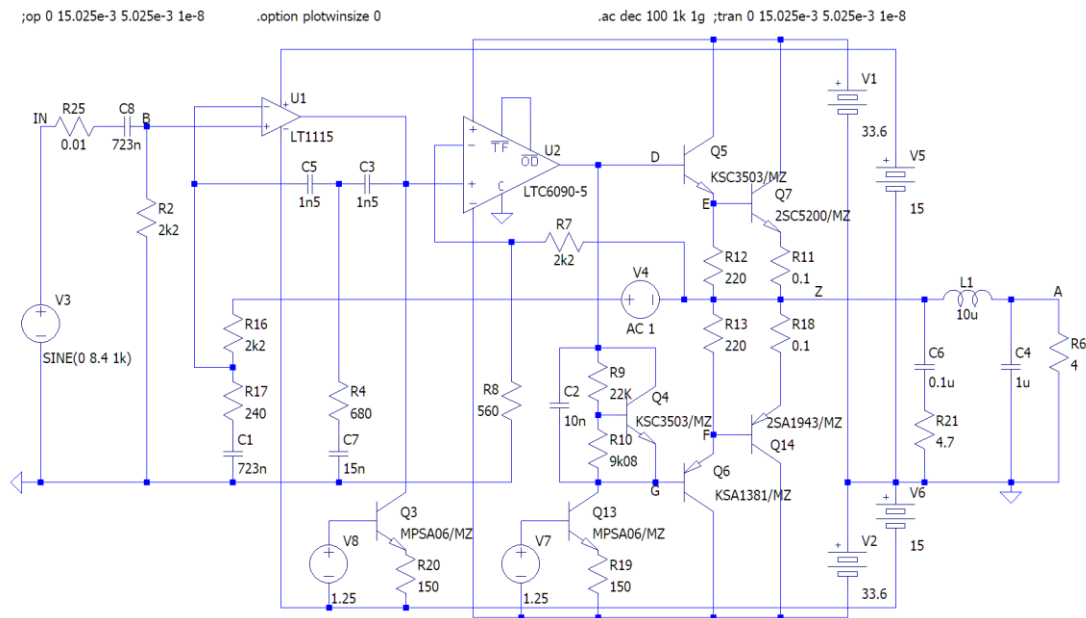


Timing Plot



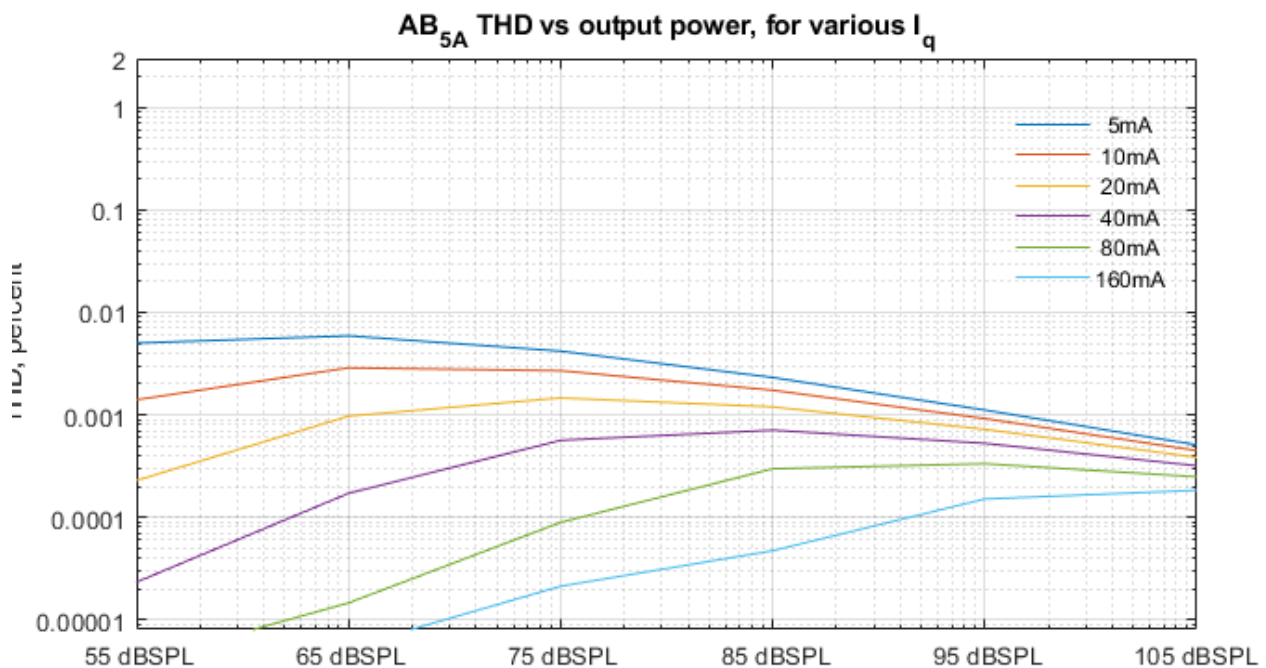
Here we used sampling rate of $f_s=50\text{MHz}$ for the $C1=100\text{p}$, $I_q=40\text{mA}$, and found that this frequency is marginal. Sometimes LTspice uses time step smaller than $1/f_s$. It would be prudent to rise f_s to 100MHz .

“Six Steps” with LTC6090-5 ($AB_0 \rightarrow AB_{1A} \rightarrow AB_{2A} \rightarrow AB_{3A} \rightarrow AB_{4A} \rightarrow AB_{5A}$)



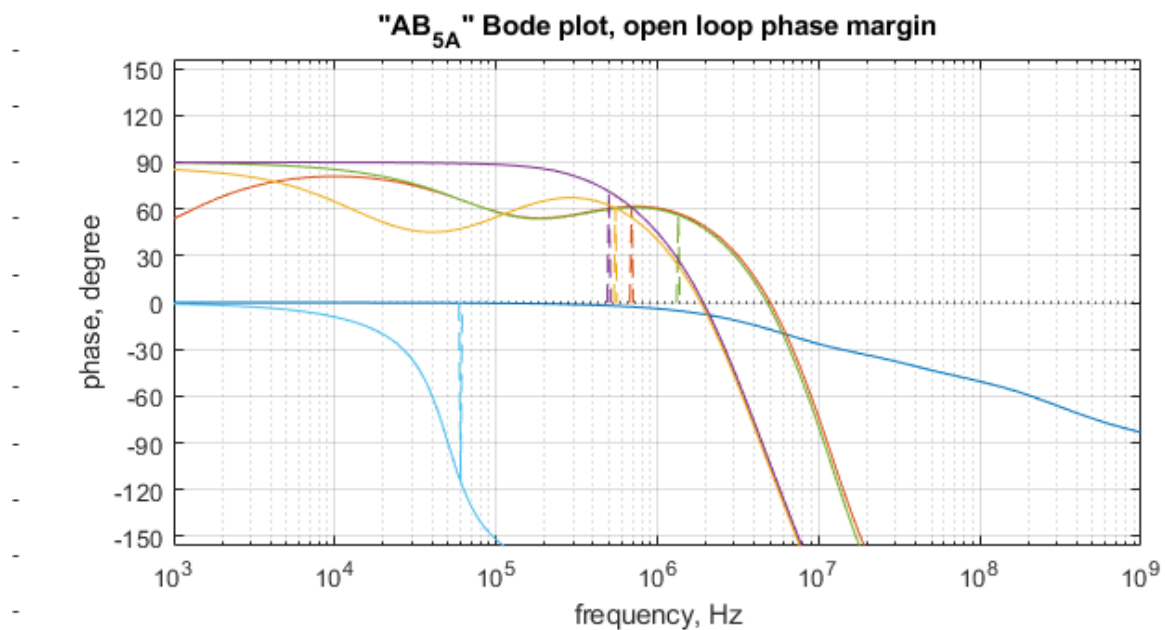
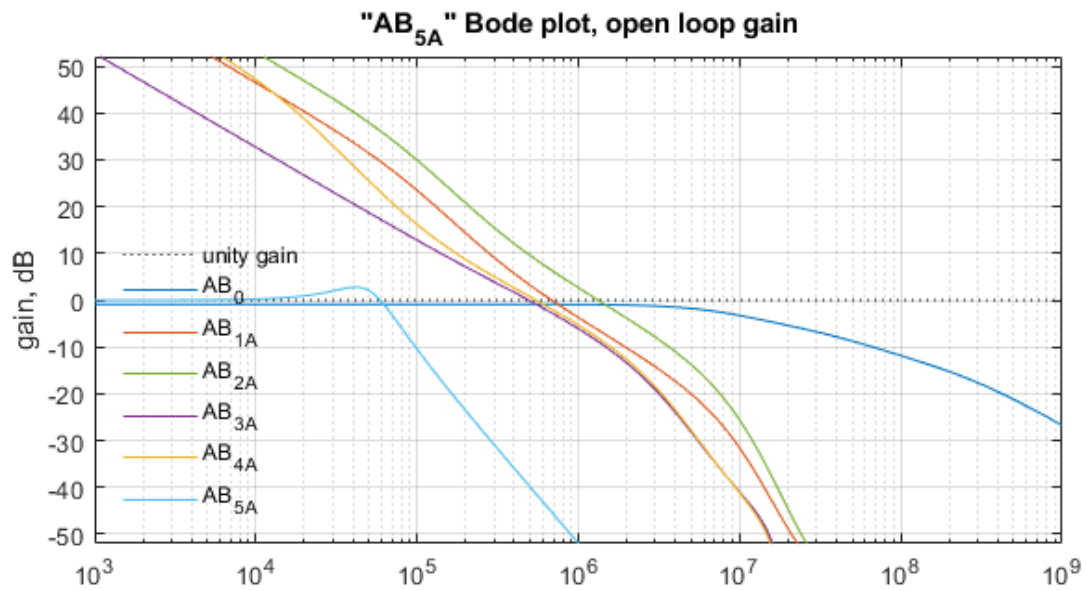
On step 1, LTC6090-5 is used with 20dB gain. On step 2, it is used with 14dB gain, and 6dB are added by LT1115. Everything else goes exactly as described in the template. It's important to note that the output of LT1115 must be biased into pure Class A mode to avoid otherwise inevitable ringing. It is not important how exactly it's done. Otherwise, the schematics are trivial.

To avoid overwhelming readers with excessive number of graphs on hundreds of pages, only the final comparison and summary plots are provided.

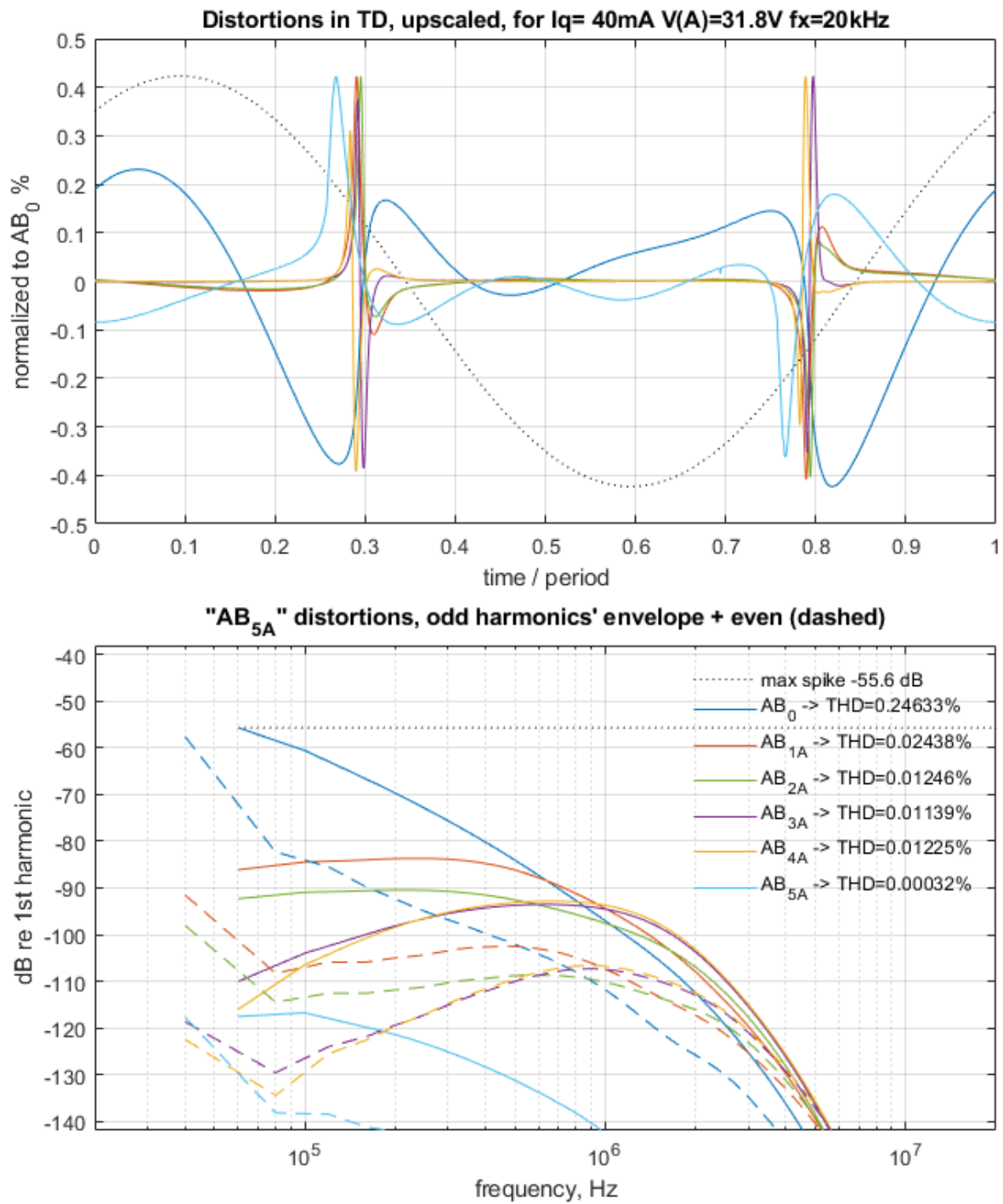


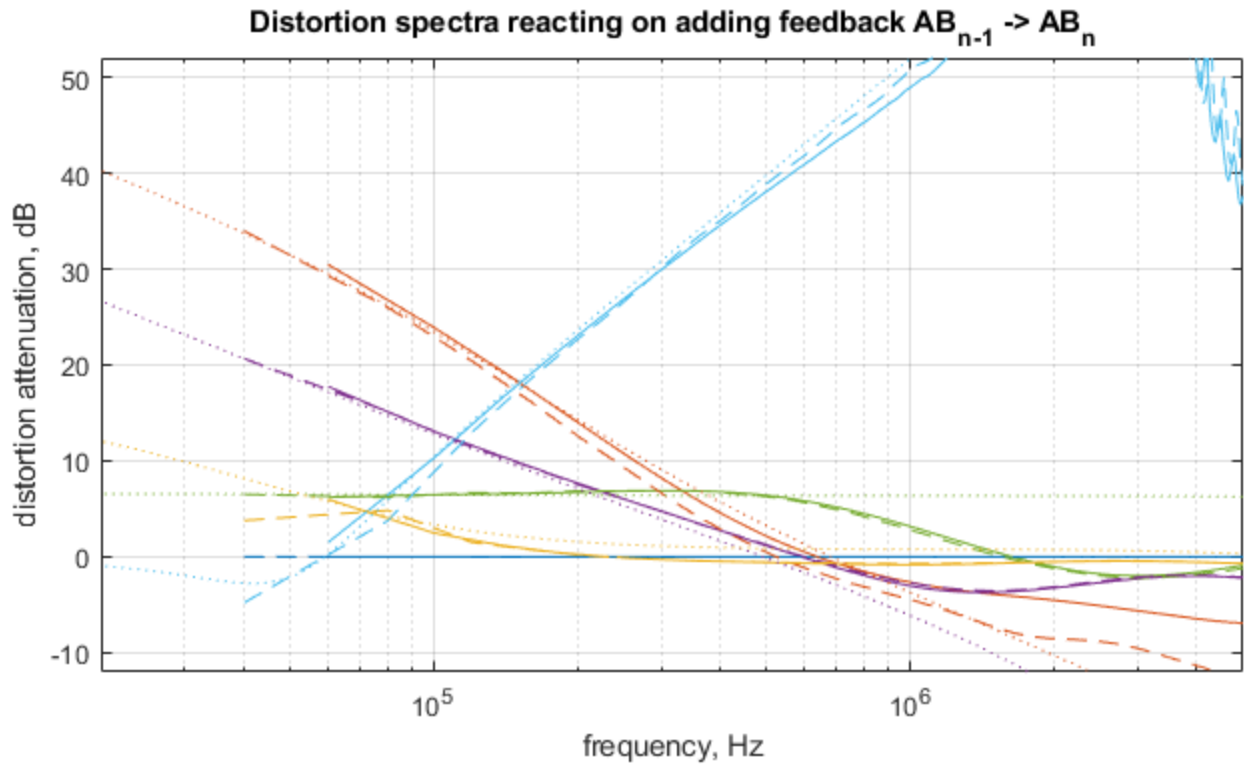
Bode for the steps, $I_q=40\text{mA}$

The f_T for each version was chosen so that the phase margin is at 60° or above. The choice of 50kHz for AB_{5A} is traditional.



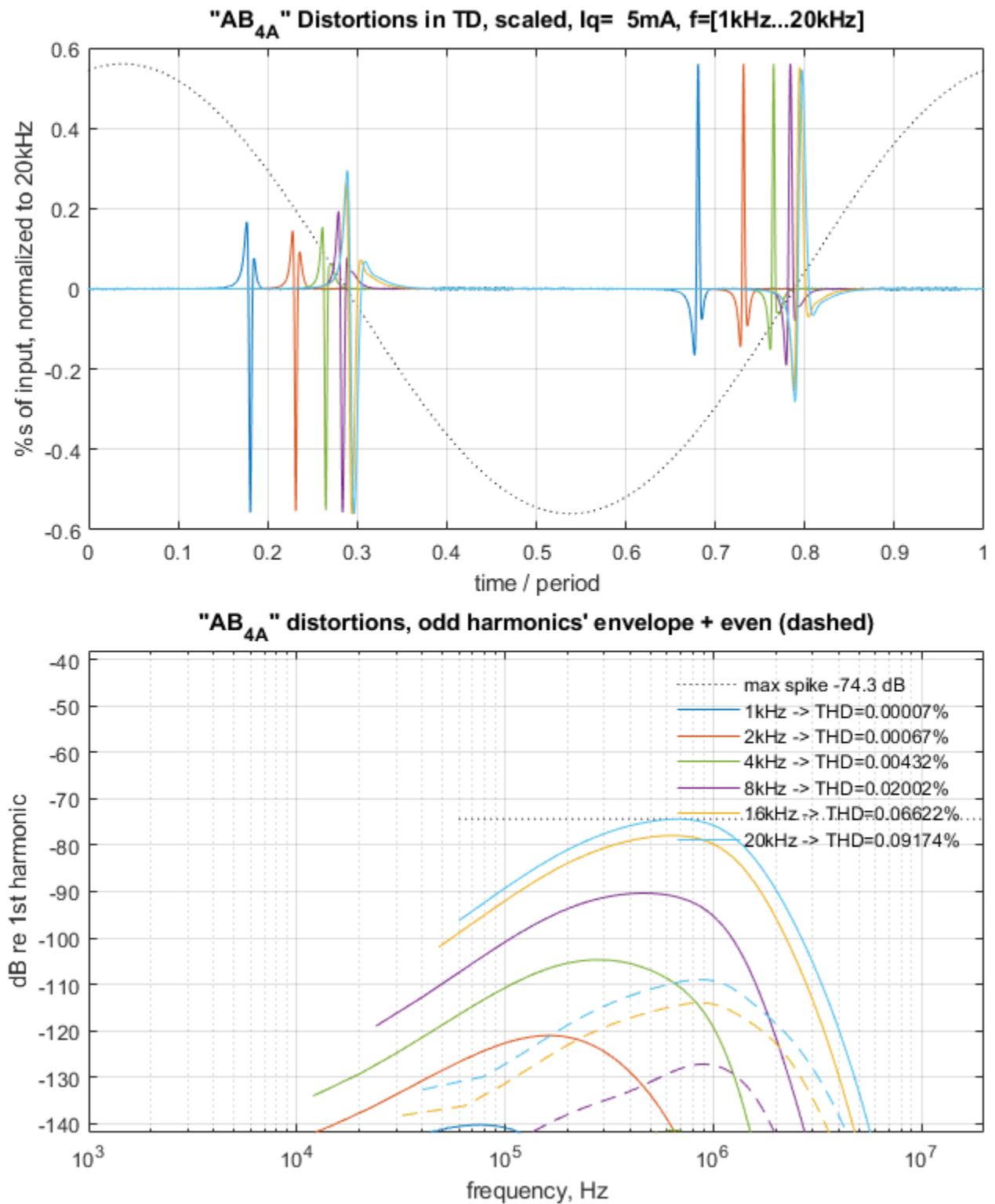
THD for the steps, same I_q





Here we see that the changes in the feedback's open-loop gain (dotted lines) and the changes in distortions' spectra (solid and dashed lines for odd and even harmonics) have one-to-one mapping, which also tells us that the op-amps' internal distortions are low.

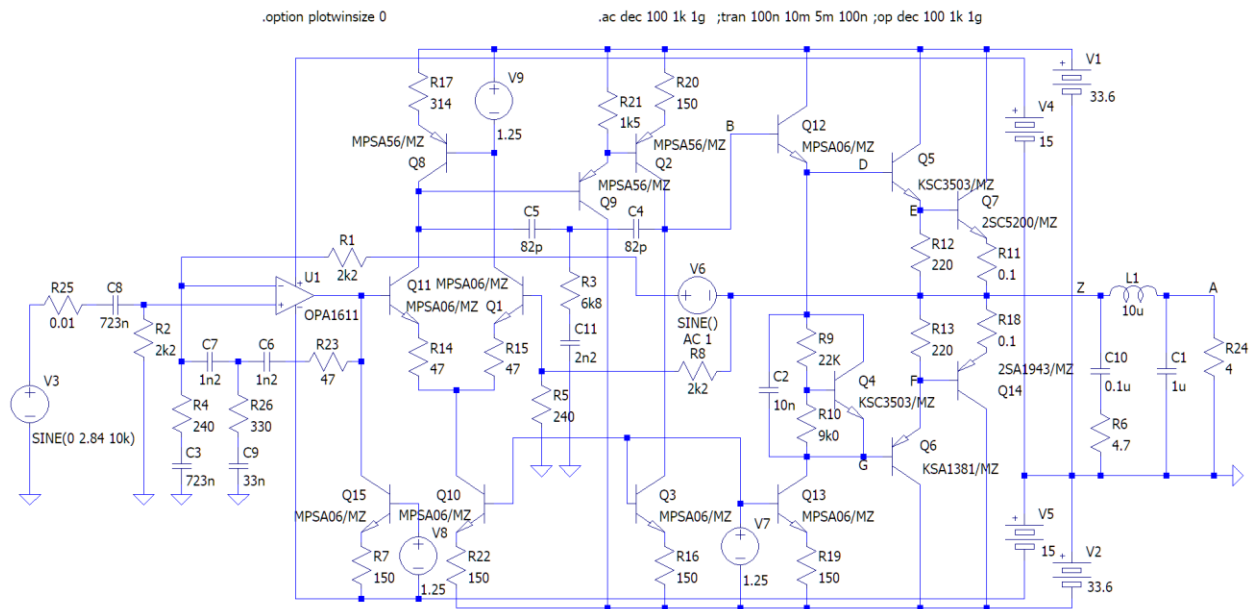
THD for varying frequency, $I_q=5\text{mA}$, 1W



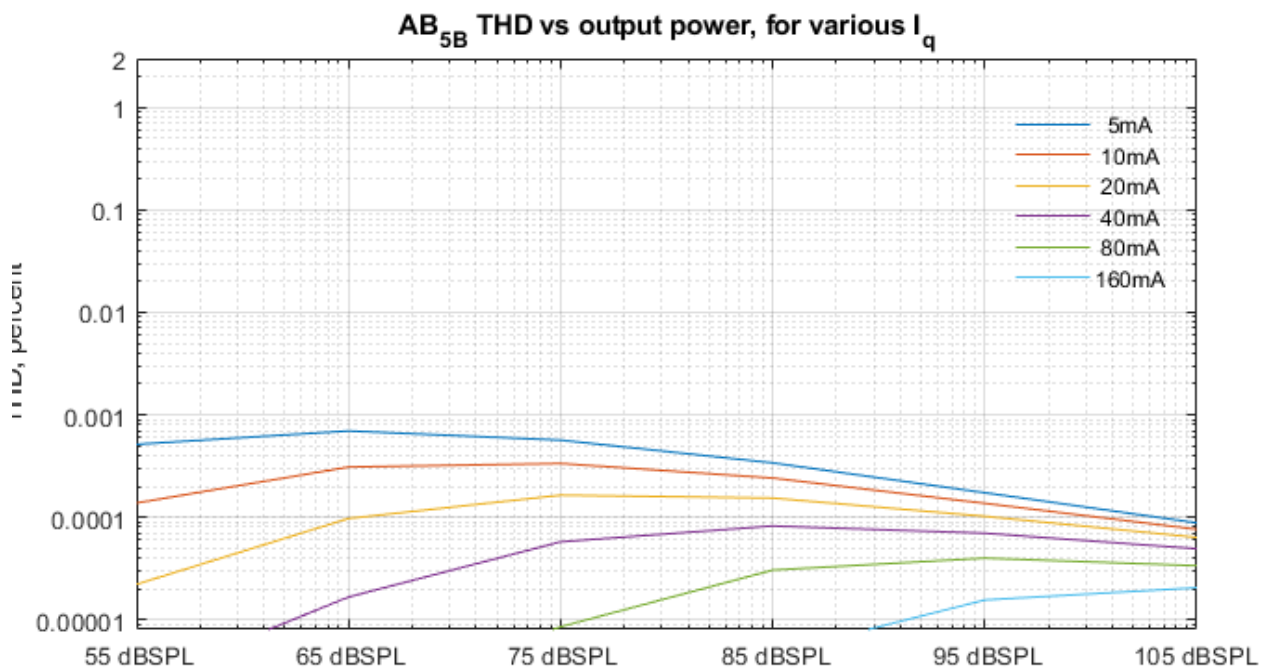
... fall/rise as f_3 because of a 1st degree of astatism loop nested inside a 2nd degree loop.

“Six Steps” with “Blameless”-like amp ($AB_0 \rightarrow AB_{1B} \rightarrow AB_{2B} \rightarrow AB_{3B} \rightarrow AB_{4B} \rightarrow AB_{5B}$)

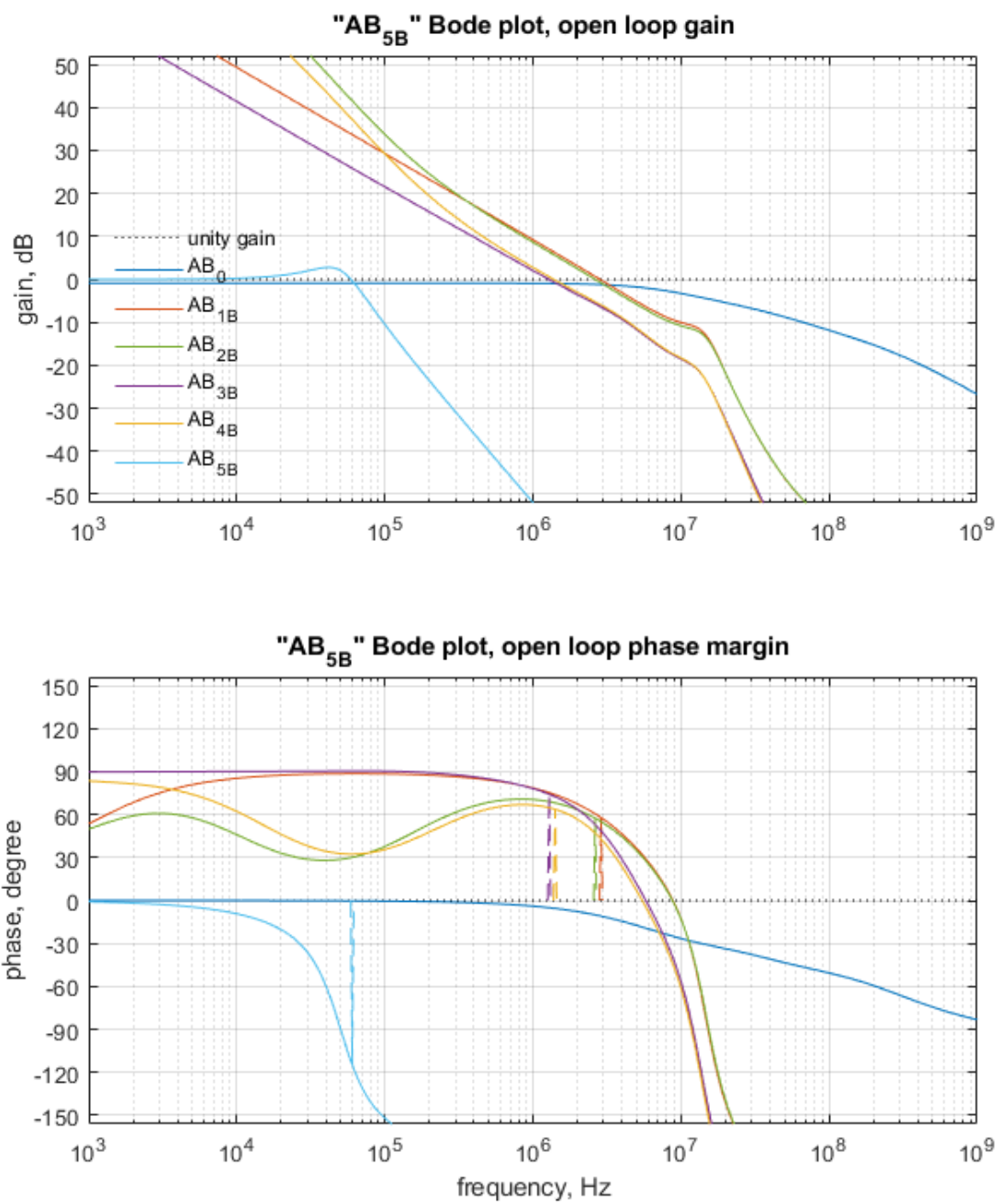
Significantly better results can be achieved with purposely-build discrete op-amps. The schematics below is (not an exact copy but) rooted in the famous “Blameless Amplifier” by D. Self whose impact on the development of audio amplifiers could not be overstated.



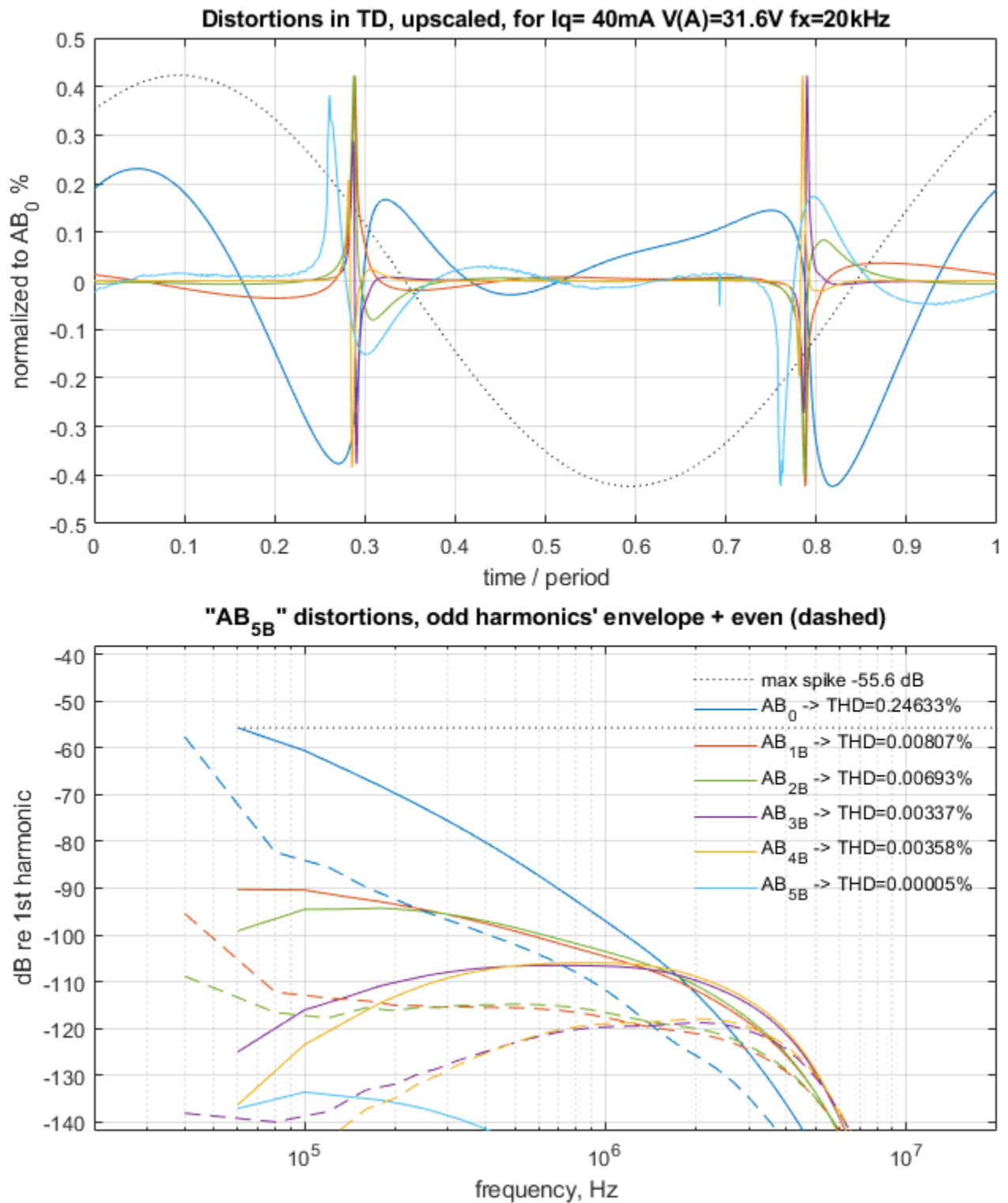
The R_{23} plays a fairly important role made possible by predictability of AB_{2B} 's f_{-3} because the amplifier's frequency response is well under control.



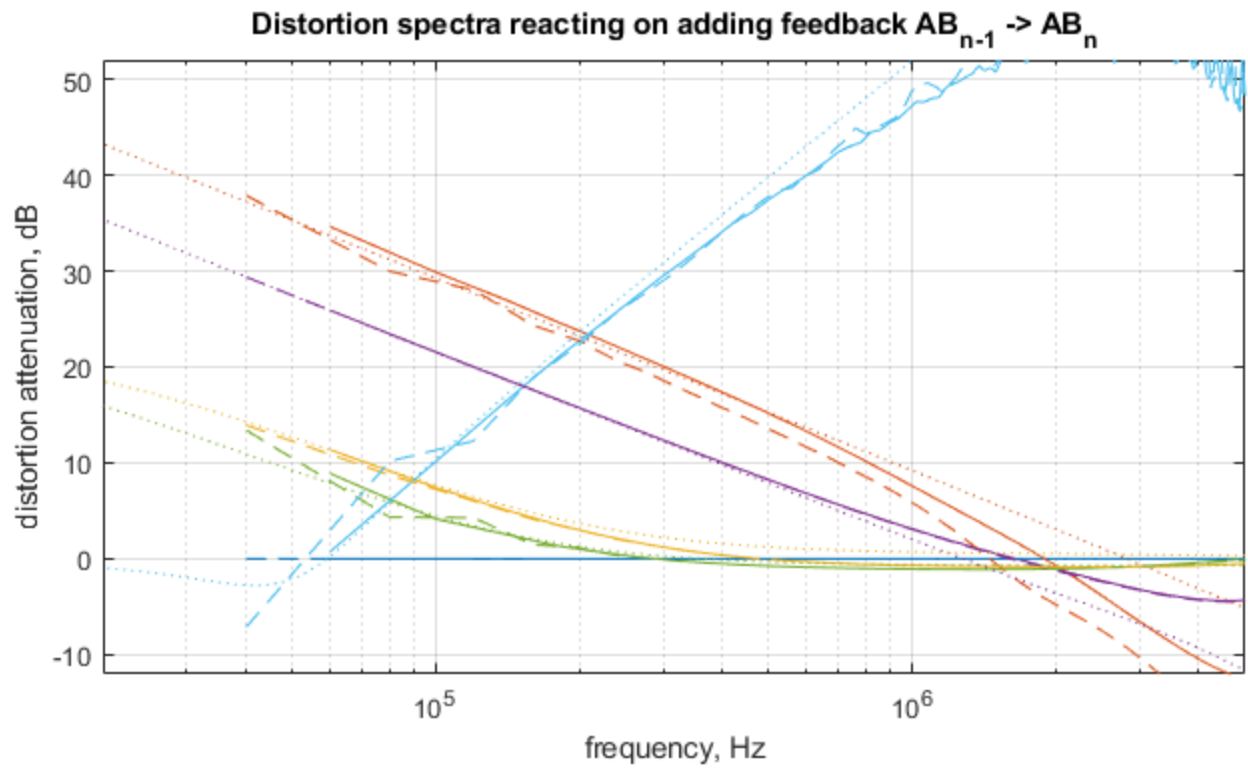
Bode for the steps, $I_q=40\text{mA}$



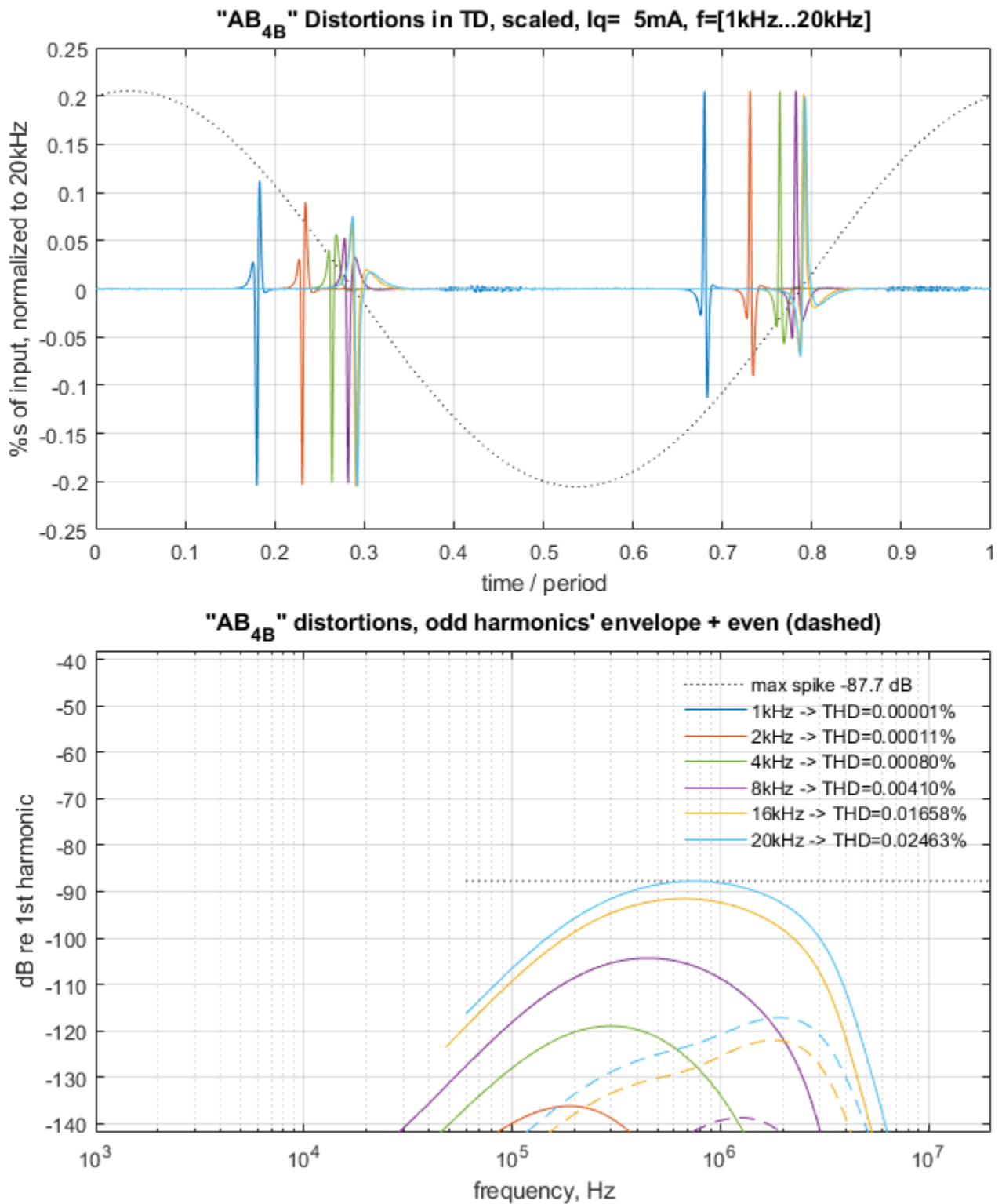
THD for the steps, same I_q



The AB_{5B} THD curve is buried in the floating point's discretization noise.



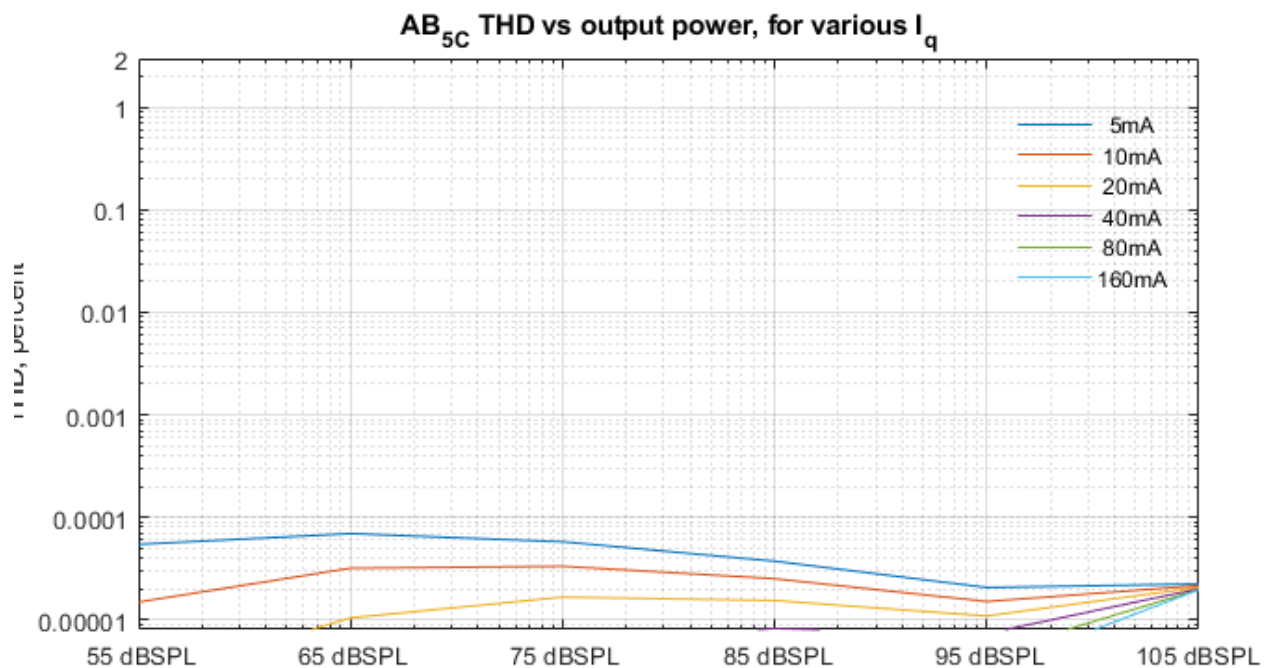
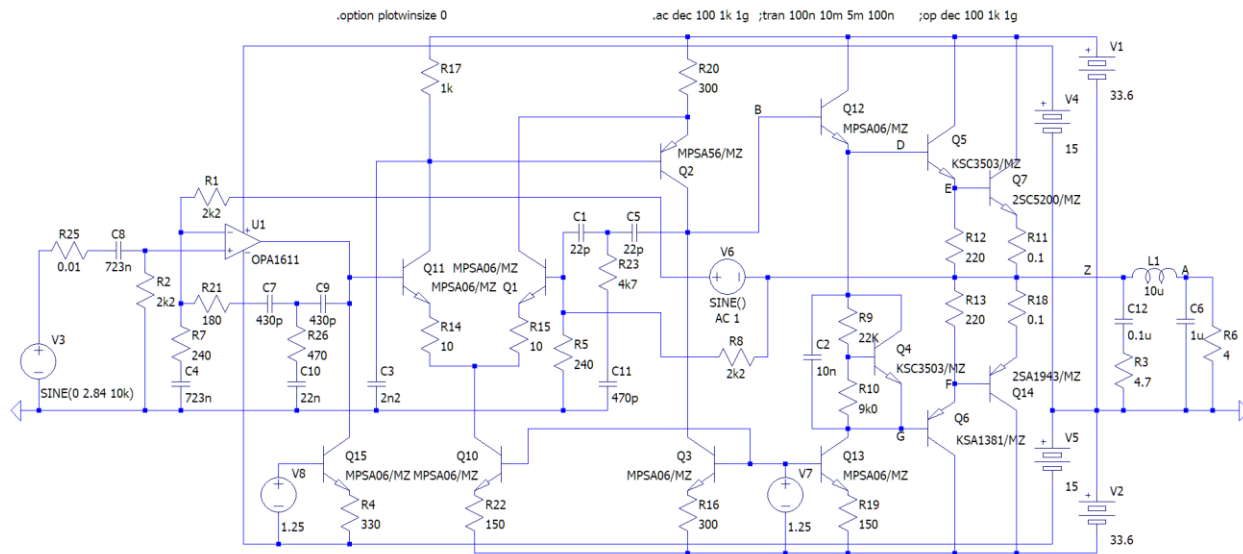
THD for varying frequency, $I_q=5\text{mA}$, 1W



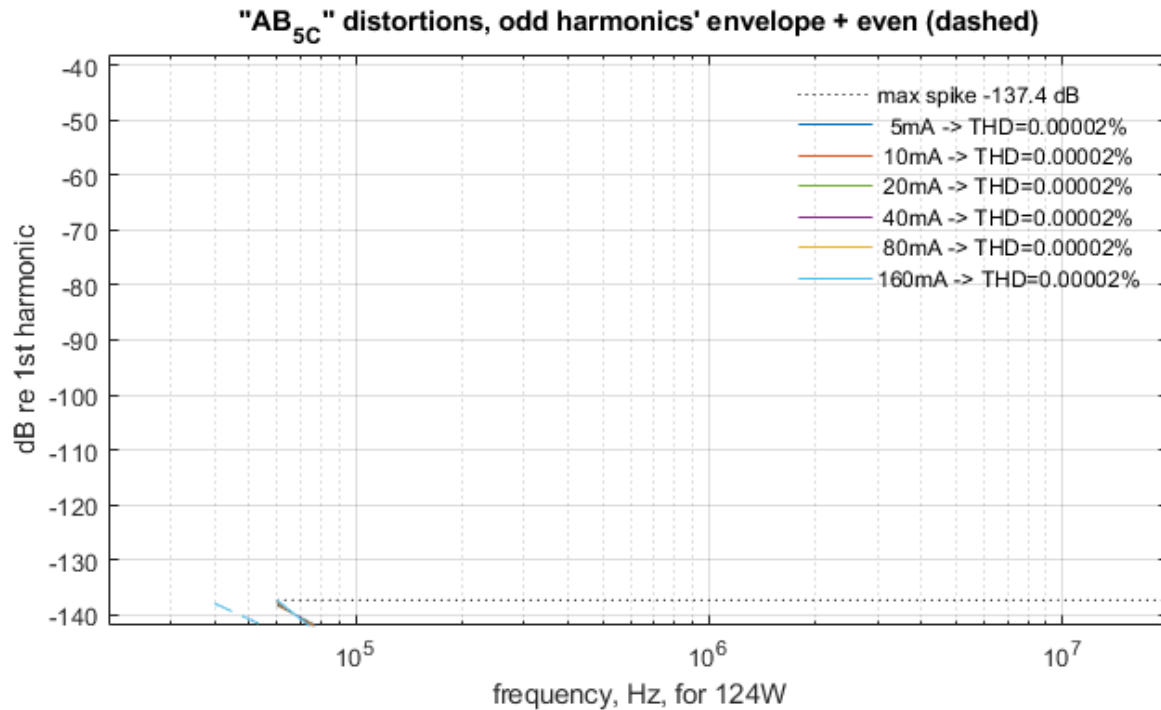
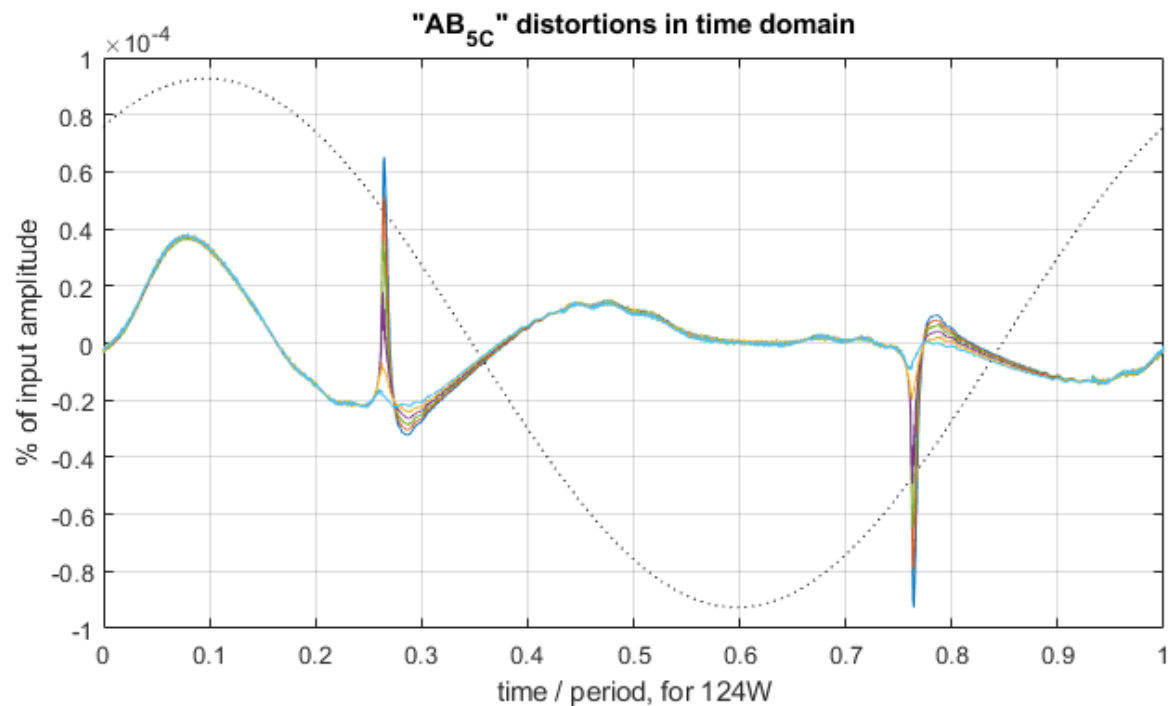
...fall/rise as f^4 because of two nested 2nd order astatism feedback loops. As soon as we get satisfactory THD at 20kHz, we are fine and do not need to about lower frequencies.

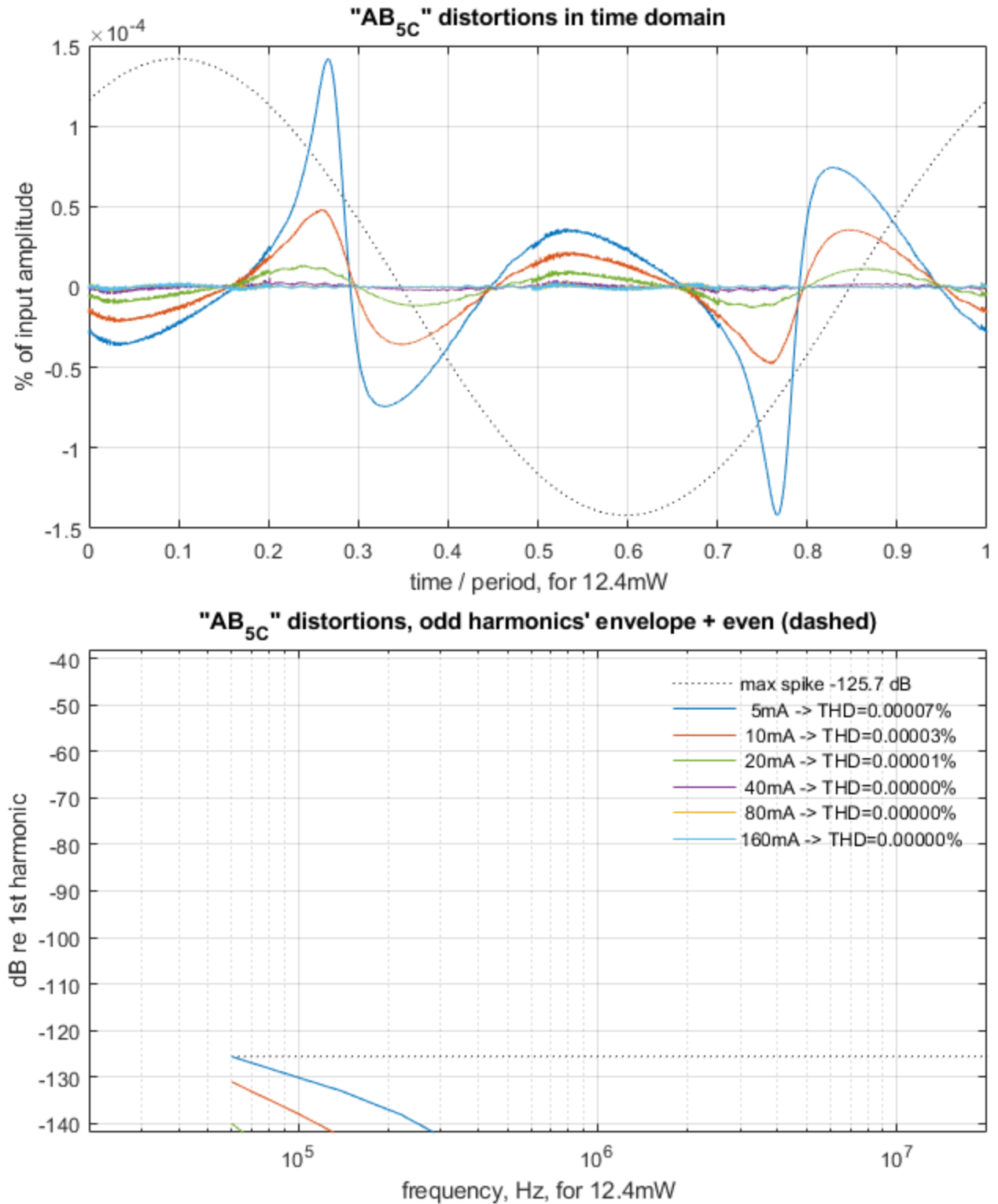
“Six Steps” with a fast amp ($AB_0 \rightarrow AB_{1C} \rightarrow AB_{2C} \rightarrow AB_{3C} \rightarrow AB_{4C} \rightarrow AB_{5C}$)

It appears that the usual 2-cascade pre-amps with Miller compensation are nice but their output impedance is inductive - which affects phase margin and limits the maximum f_{-3} frequencies. If we could avoid it and double the first f_{-3} , we double all other f_{-3} - thus increasing the total feedback loop gain by 24dB vs 6dB for non-nested simple feedback design. As an example of such design, a single stage pseudo-cascode is suggested. Surely, an experienced RF EE (which I am not) could design a much better pre-amp.



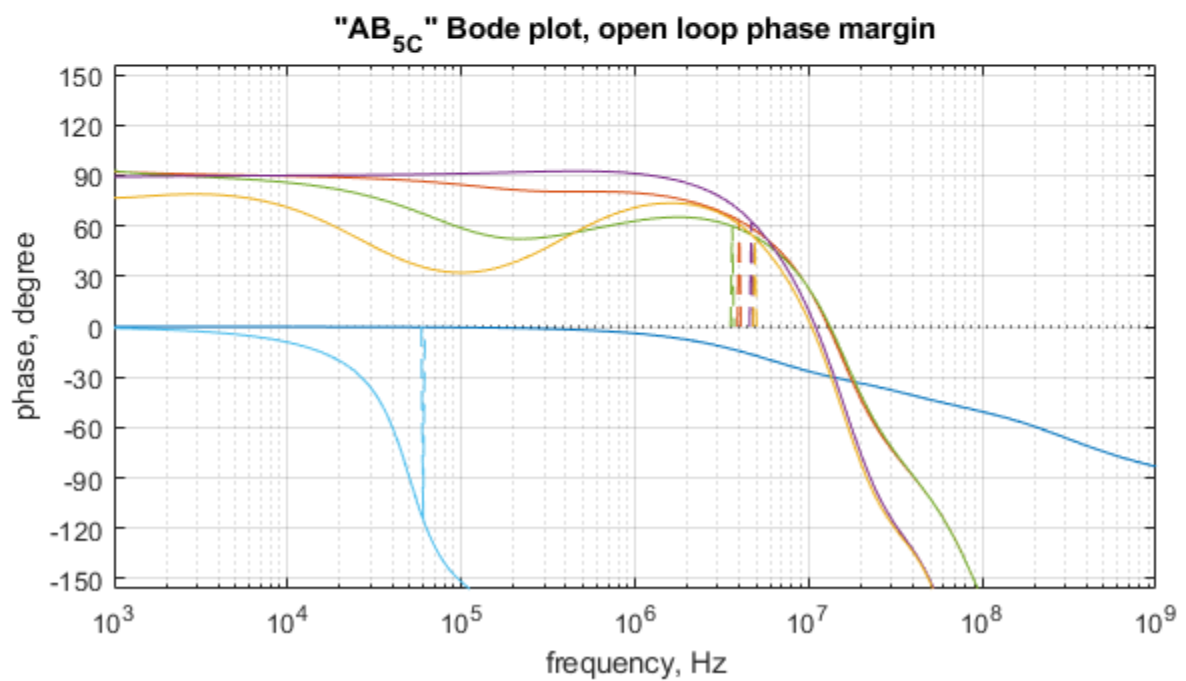
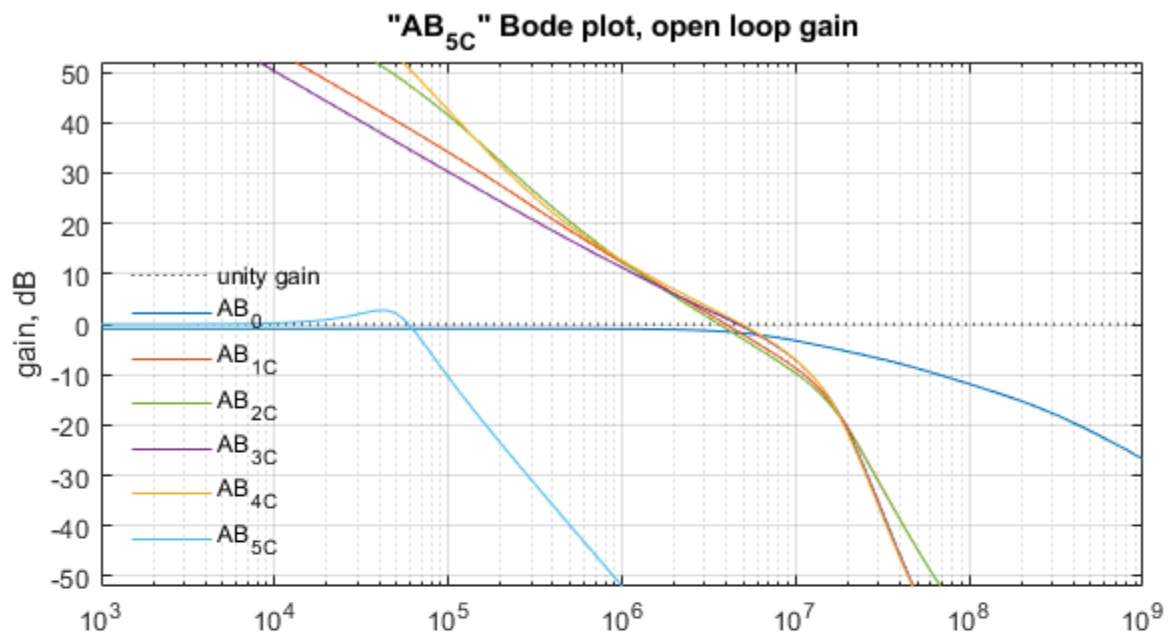
The 2nd degree astaticism on the AB_{2C} step has very little impact on the overall performance. R₂₁ in series with C₇+C₉ helps to compensate the roll-off of AB_{2C} on f₋₃ ("PI" mode of control), thus extending the depth of AB_{3C}-AB_{4C} feedback.



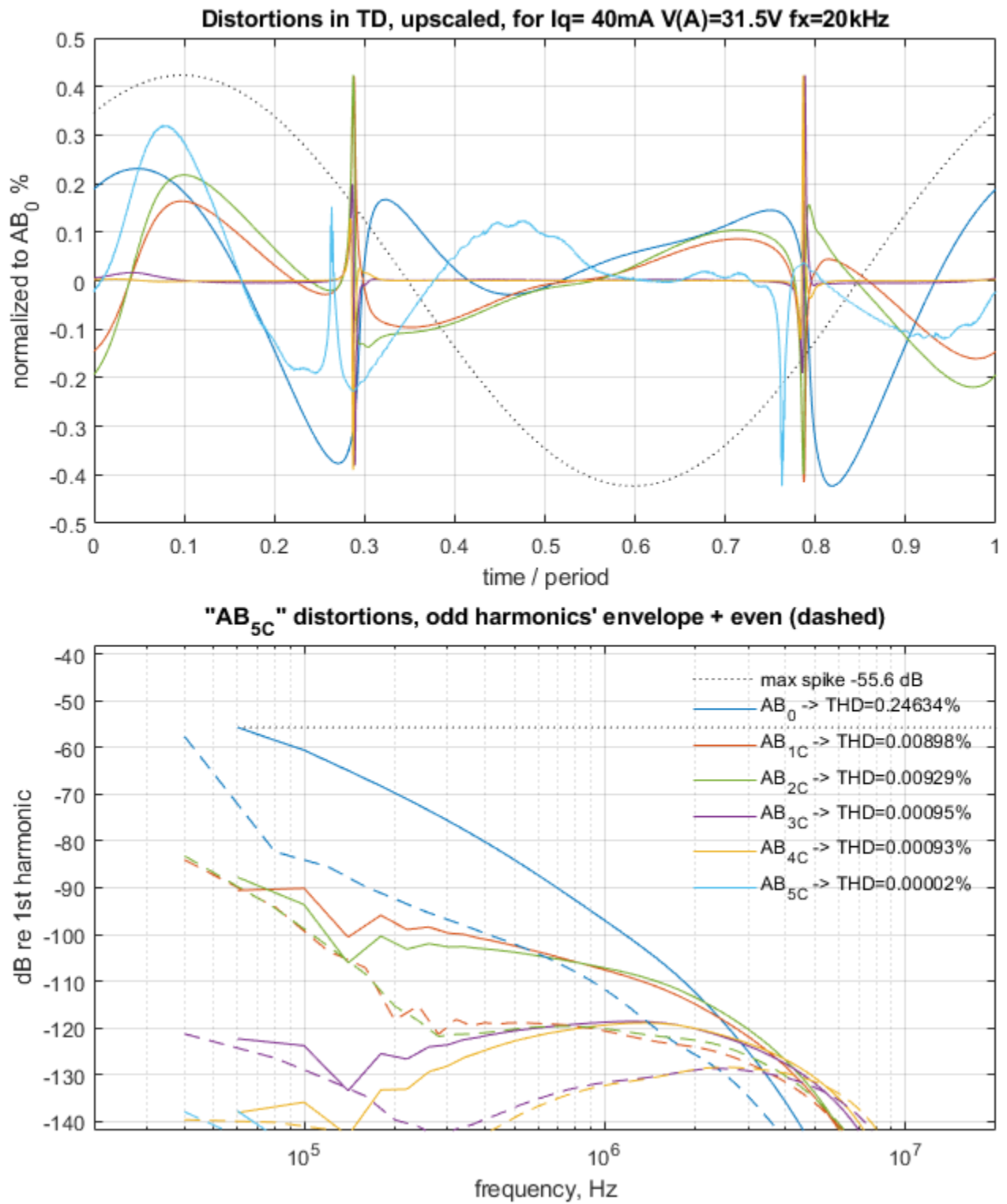


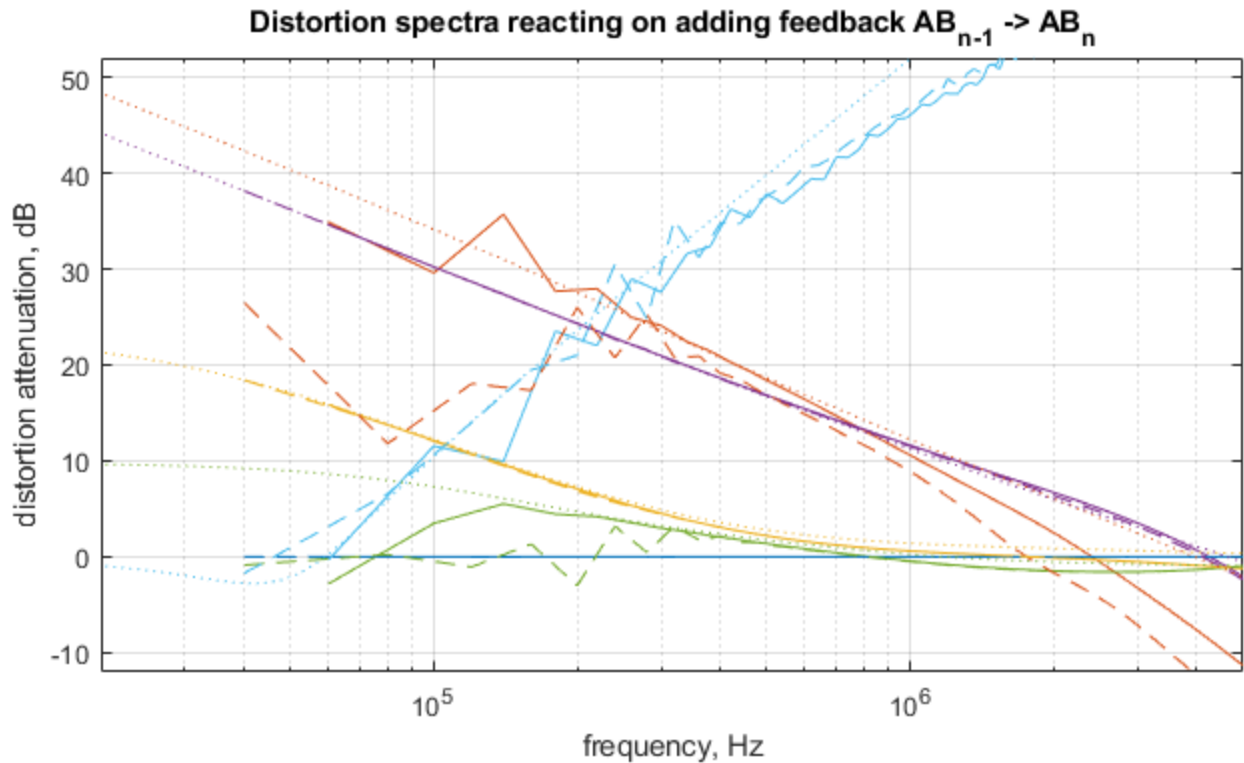
We can clearly see floating point discretization noise here.

Bode for the steps, $I_q=40\text{mA}$



THD for the steps, same I_q

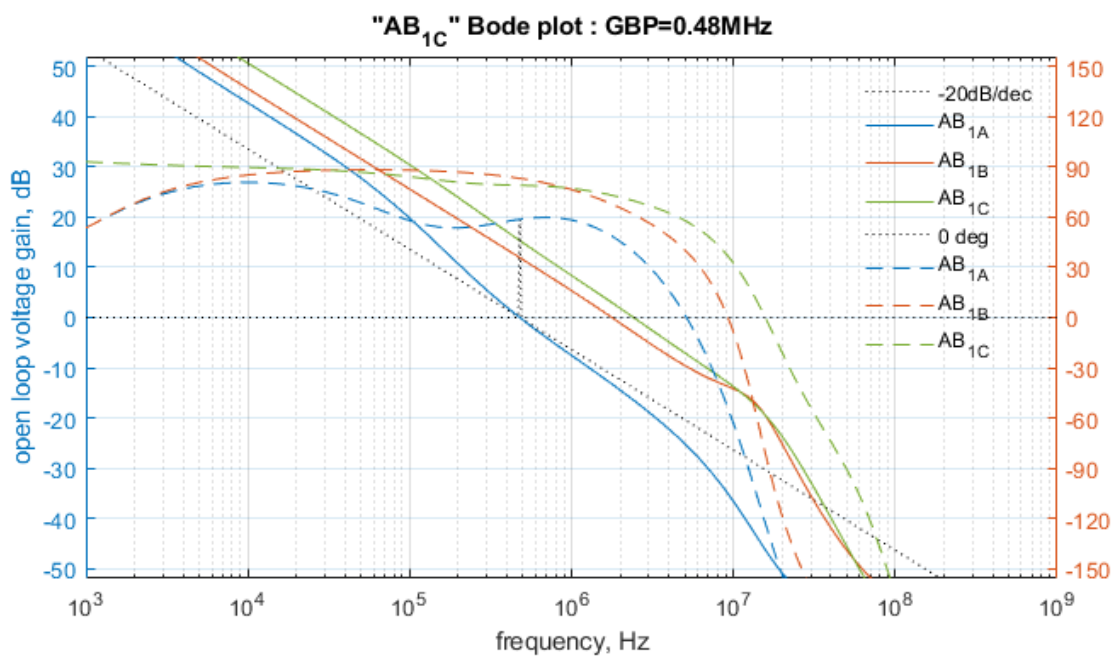


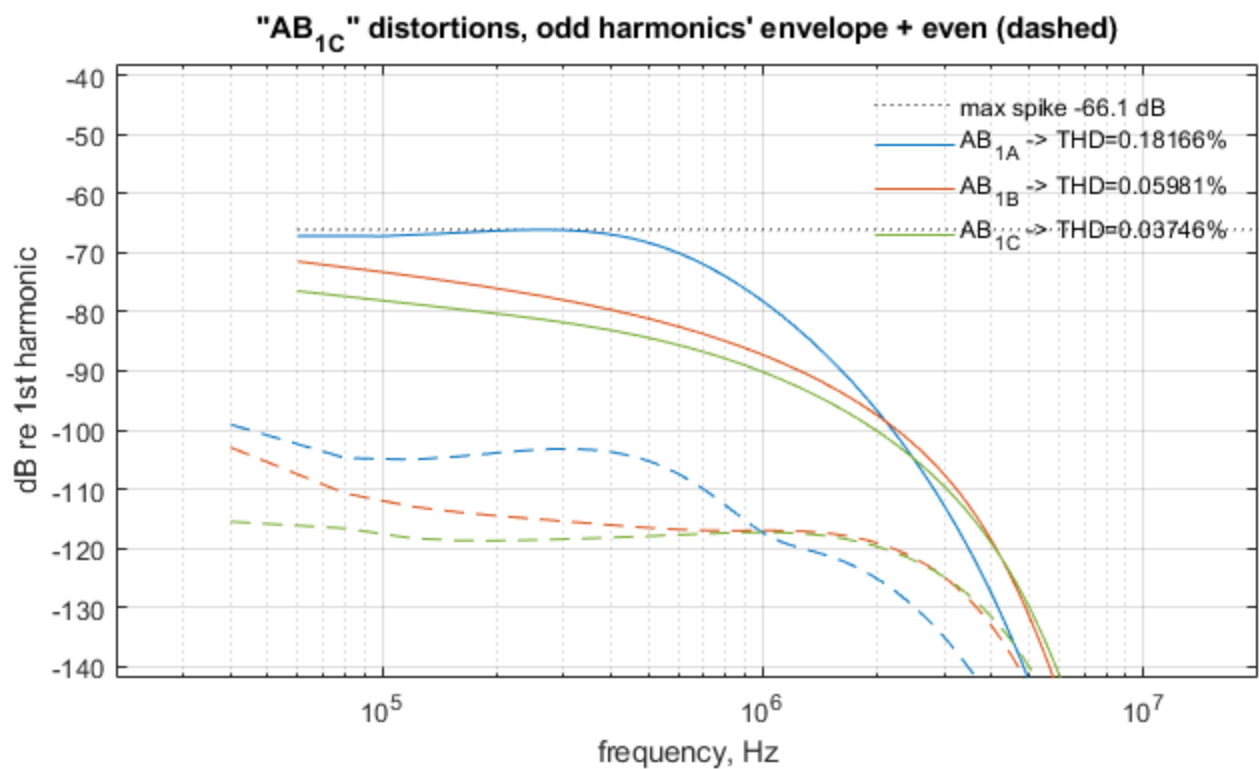
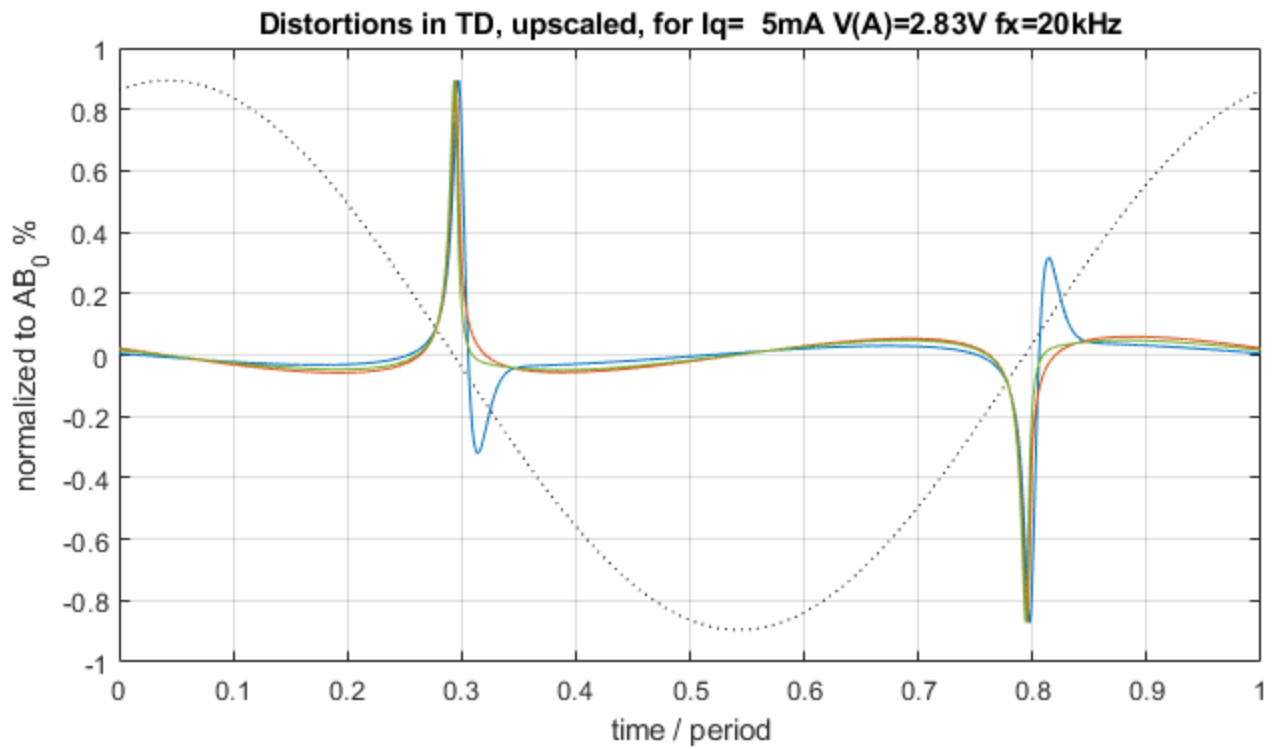


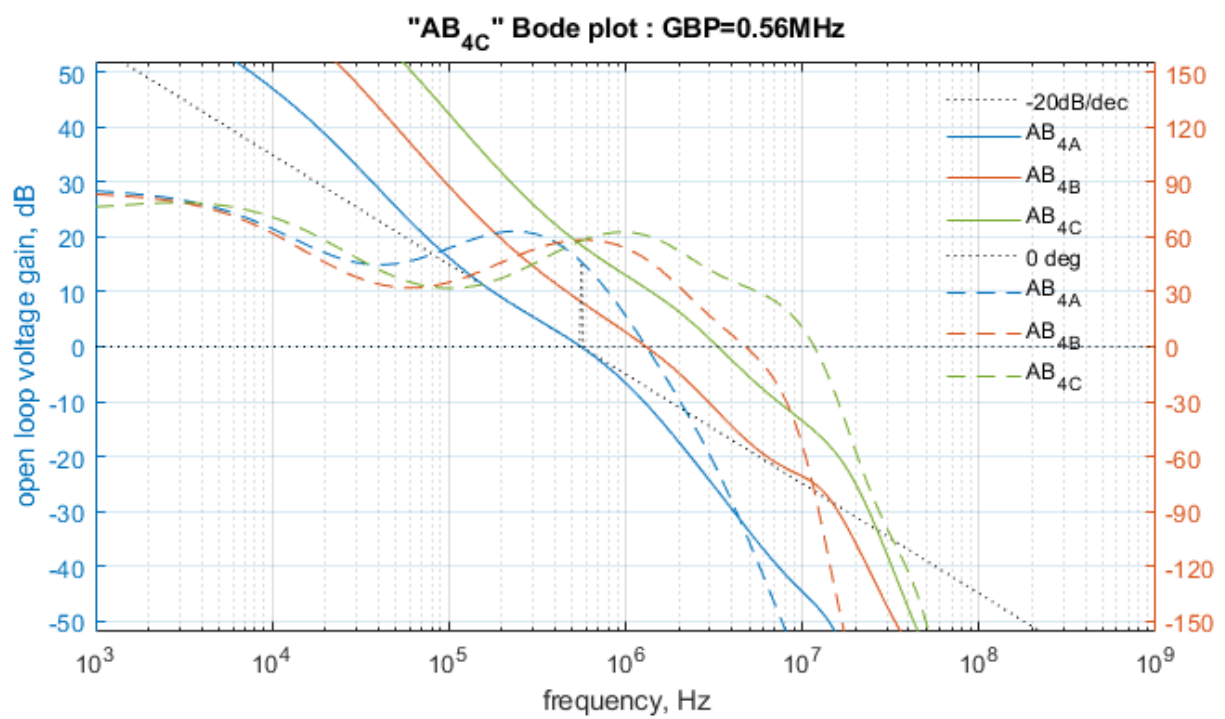
Note that this fast pre-amp has significant, clearly visible even-order “soft” harmonics that neither AB_{1C} nor AB_{2C} can manage with. However, the outer feedback loop suppresses them easily.

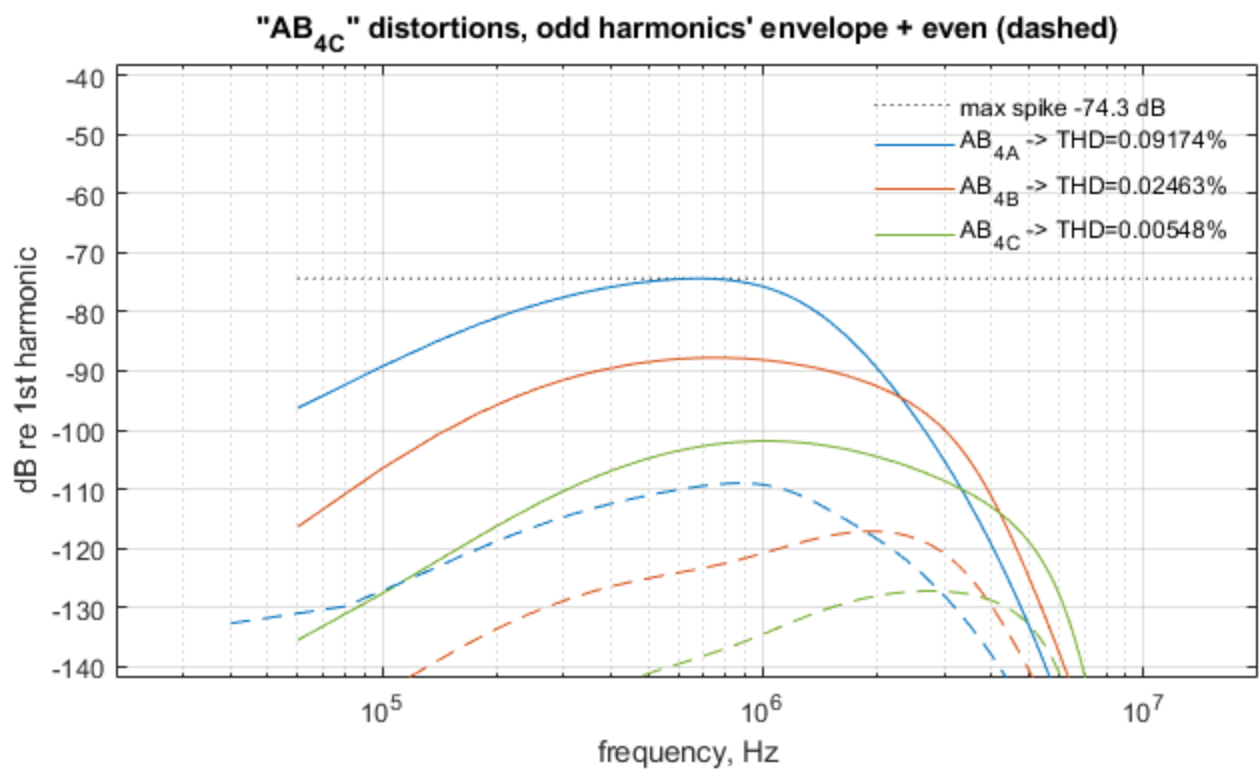
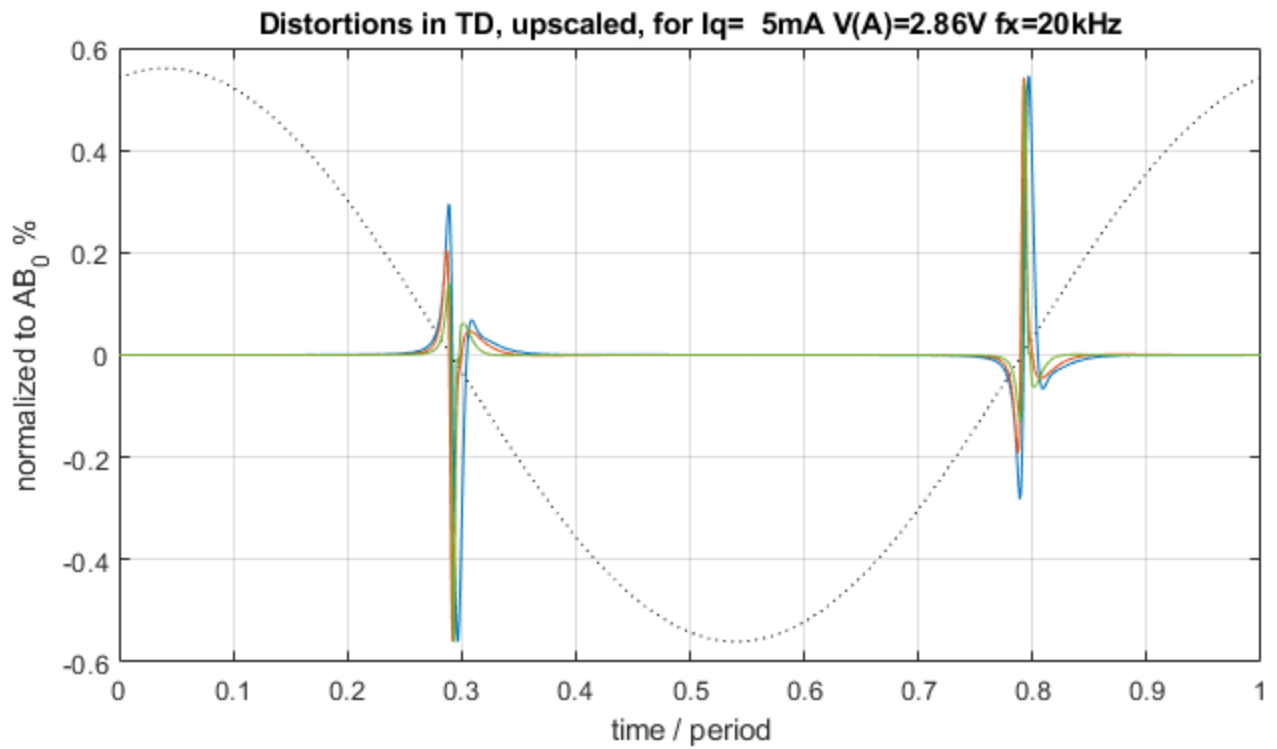
Graphs for varying frequency are trivial and omitted.

AB_{1A} vs AB_{1B} vs AB_{1C}





AB_{4A} vs AB_{4B} vs AB_{4C}



DISCUSSION

1. Speed of pre-amp f_T plays much more important role than its cleanliness. Generalizing, Bode plots contain more information (relevant for THD) than small details of the schematics' implementations.
2. "Soft" distortions of EF play very little if any role in the deep-nested amplifiers' output distortions which are practically all due to "hard" crossover distortions.
3. Nested feedback, taken to the extreme, may help to achieve very high degrees of linearity – if and only if applied properly. Good knowledge of both electronics and the theory of automatic control is a must, so it's a team work.

The role of adequate simulation in the design of nested feedback audio amplifiers is critical. The team of developers would need to start with the simplest schematics, measure it, compare with simulation, correct inevitable mistakes (both in simulation and in hardware), and so on till it performs sufficiently well. Then make a small step - add a thin layer of complexity, again measure the amp and compare with simulation, and so on, i.e., perform feedback-controlled design of a feedback-based amplifier. The major component of such feedback control procedure is differentiator – a device which can compare the measured output with the expected value. There are now high-speed precision ADC like PCMD3140, ADS127L01, ADS8900B, etc with development board available. These or similar devices should be capable of adequately sampling [high-pass/stop-band prefiltered] output waveforms containing wide-band crossover distortions for high excitation frequencies f_x (since distortion are most noticeable there).

The underlying processes inside even the simplest "hard" non-linear systems with feedback loops are complicated, counterintuitive and not easy to measure or simulate. The author does not claim to possess the complete understanding of the problems touched and believes there is more to discover. "While I have endeavored to describe things as they appeared to be, I am conscious of having been unable to avoid the usual proportion of errors, for which I beg indulgence, and which I leave for others who shall pursue the same path of investigations to correct"¹²

ACKNOWLEDGEMENTS & LICENSES

This hobby was carried out as "art for the art's sake" in unpaid time over the last 40 years with no expectations of any future rewards. It was not sponsored, directly or indirectly, by any commercial (or non-commercial) organization. Any biases or preferences towards any particular vendor are non-intentional.

¹² Joseph Leidy, "Fresh-water Rhizopods of North America", 1874.

LTspice2MATLAB

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Blameless Amplifier

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The rest of this “Crossover Distortions...” package

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REFERENCES

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