The Wavelet Transform

The well-known Fourier Transform (FT) shows what frequency components (spectral components) exist in the signal. That's all.

When the time localization of the spectral components is needed, a transform giving the time-frequency representation of the signal is needed. The Wavelet Transform (WT) is a transform which provides the time-frequency representation.

By examining the FT, we cannot tell what spectral component exists at any given time instant. The best we can do is to investigate Short Time Fourier Transform STFT-based spectrogram to determine what spectral components exist at any given interval of time. However, there is still an issue of resolution. The STFT gives a fixed resolution at all times, whereas WT gives a variable resolution. This has some advantages. Higher frequencies are better resolved in time, and lower frequencies are better resolved in frequency.

Like the FT, the Continuous Wavelet Transform (CWT) uses inner products to measure the similarity between a signal and an analyzing function. In the FT, the analyzing functions are complex exponentials, $\mathbf{e^{j\omega t}}$. The resulting transform is a function of a single variable, $\boldsymbol{\omega}$. In the STFT, the analyzing functions are windowed complex exponentials, $\boldsymbol{\omega(t)e^{j\omega t}}$, and the result in a function of two variables. The STFT coefficients, $\mathbf{F(\omega,\tau)}$ represent the match between the signal and a sinusoid with angular frequency $\boldsymbol{\omega}$ in an interval of a specified length centered at $\boldsymbol{\tau}$.

In the CWT, the analyzing function is a wavelet, $\boldsymbol{\psi(t)}$. The CWT compares the signal to shifted and compressed or stretched versions of a wavelet. Stretching or compressing a function is collectively referred to as *dilation* or *scaling* and corresponds to the physical notion of *scale*. By comparing the signal to the wavelet at various scales and positions, you obtain a function of two variables. The two-dimensional representation of a one-dimensional signal is redundant. If the wavelet is complex-valued, the CWT is a complex-valued function of scale and position. If the signal is real-valued, the CWT is a real-valued function of scale and position. For a scale parameter, a>0, and position, *b*, the CWT is:

$$C(a,b;f(t),\psi(t)) = \int_{-\infty}^{\infty} f(t)\frac{1}{\sqrt{a}}\psi^*\left(\frac{t-b}{a}\right)dt$$

The wavelet function used in the program is Morlet Wavelet:

$$cmor(x) = \frac{1}{\sqrt{\pi F_b}} e^{2i\pi F_c x} e^{-x^2/F_b}$$

It depends on two parameters: **Fb** is a bandwidth parameter, and **Fc** is a wavelet centre frequency.

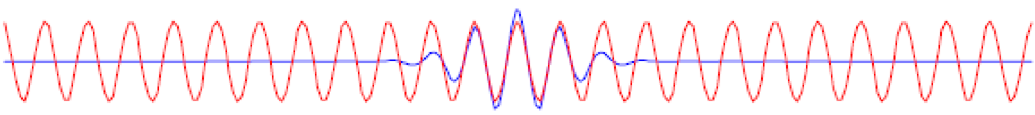


Figure 16.154. Centre frequency of complex Morlet wavelet – blue curve.

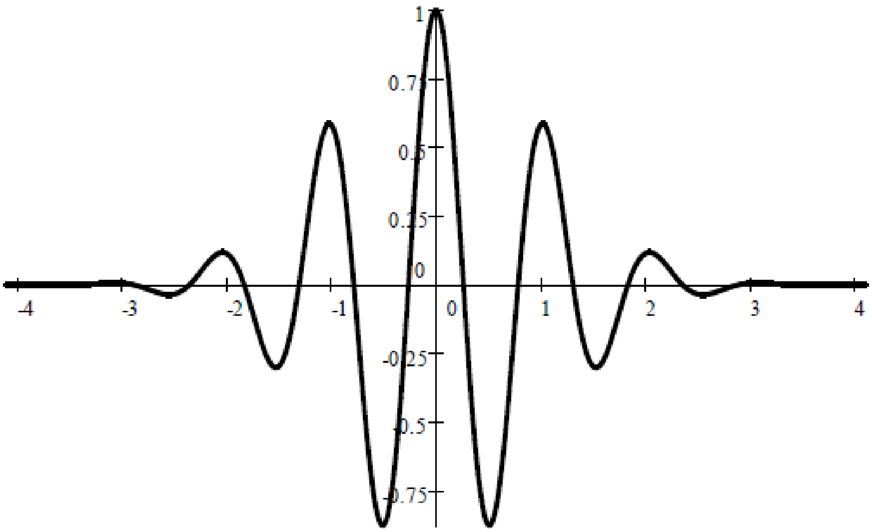


Figure 16.155. Typical Real Morlet wavelet

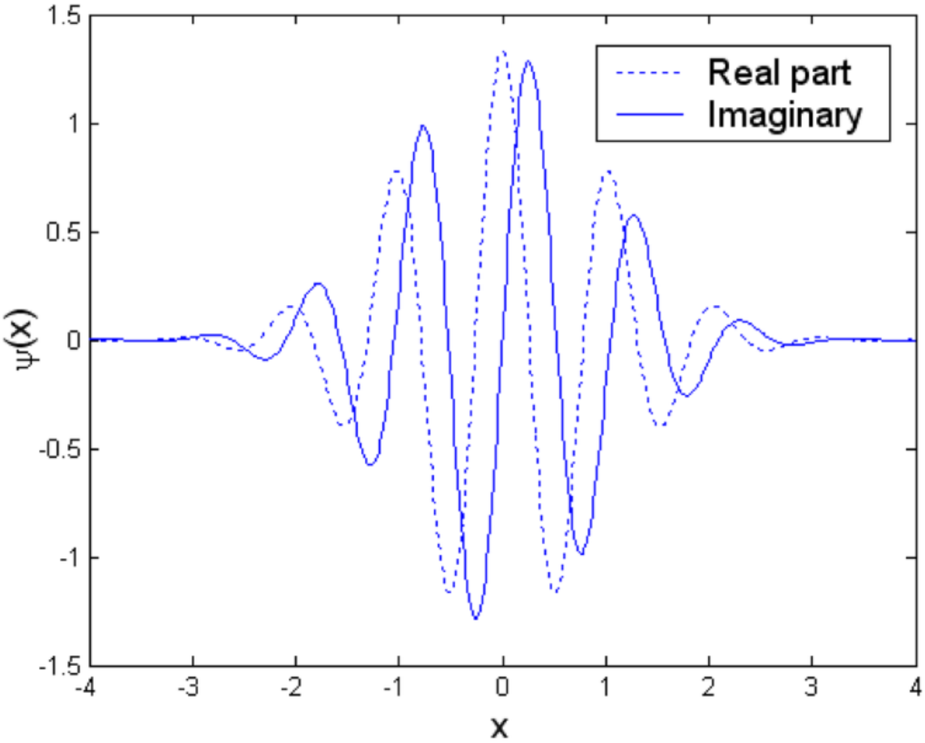


Figure 16.156. Typical Complex Morlet wavelet

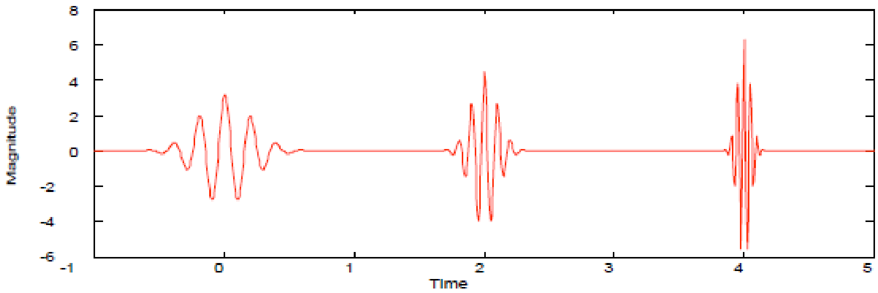


Figure 16.157. Instances of Morlet wavelet.

We can now start with a review of a very simple example of a 1kHz sine-wave represented on the frequency-time scale, using wavelets. It is observable, that the both; positive and negative peaks of the signal are being accounted for, and the signal is located exactly where it should be – at 1kHz.

Mother Wavelets Self-Tests

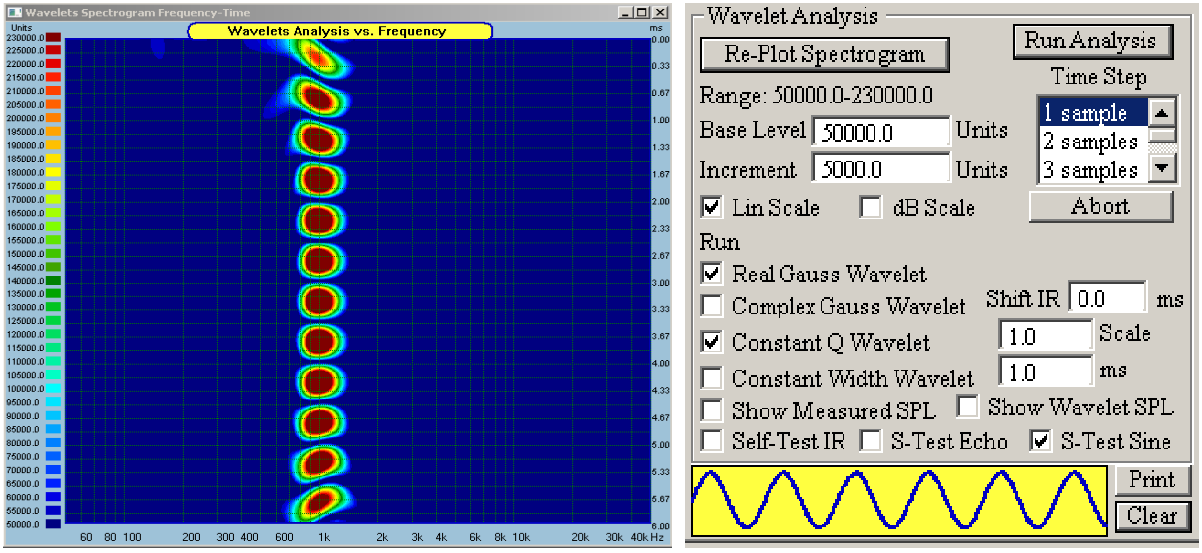


Figure 16.158. 1000Hz Sine-wave, 96kHz sampling, Constant Q Wavelet

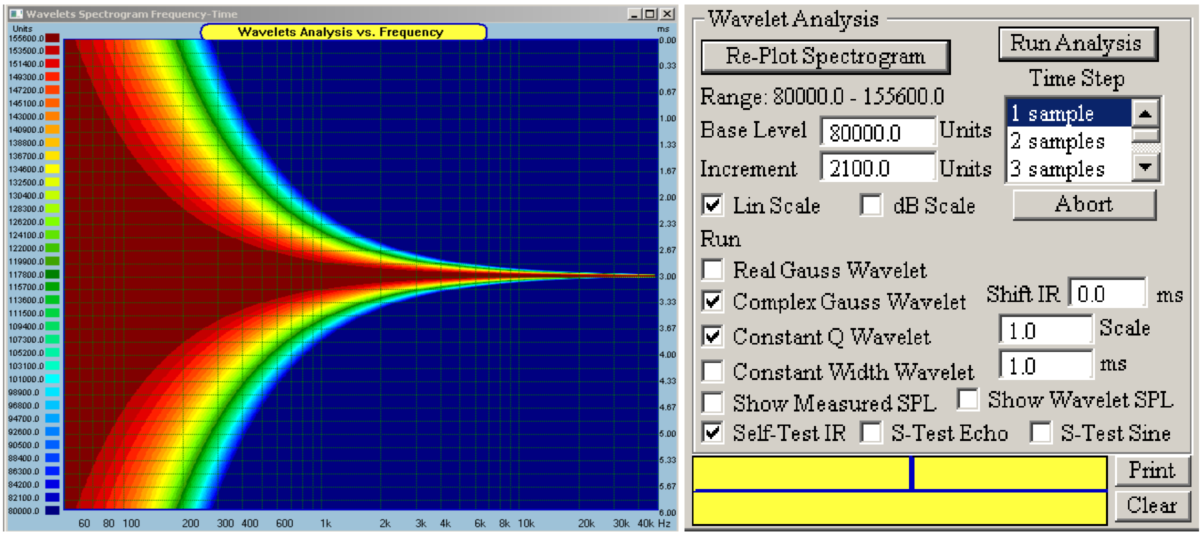


Figure 16.159. Ideal Impulse Response. 96kHz sampling, Constant Q Wavelet

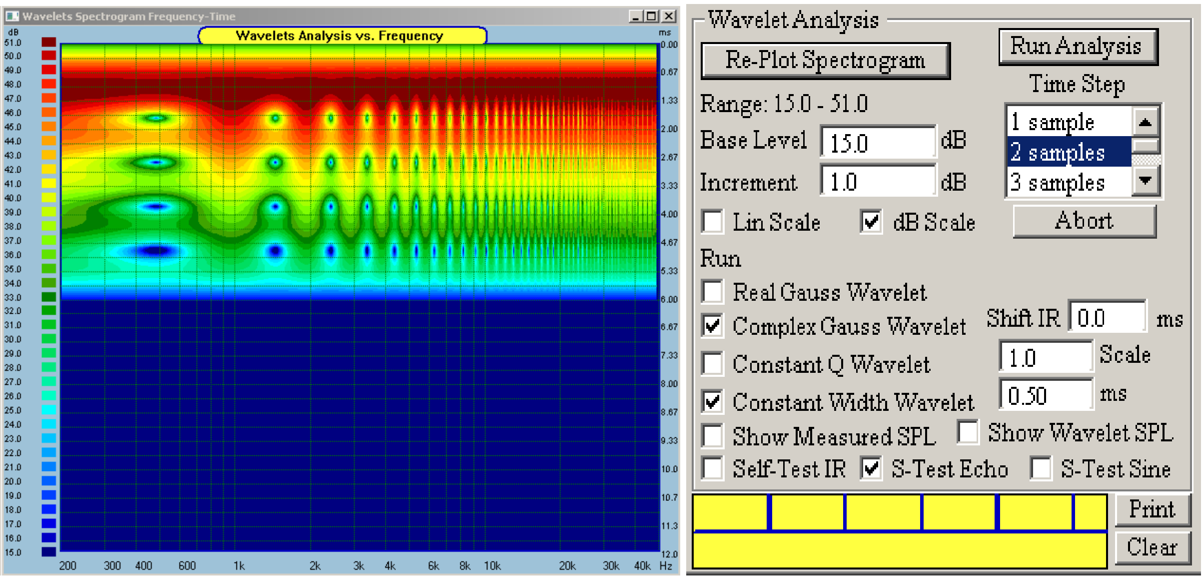


Figure 16.160. Echo. 96kHz sampling, Constant Width Wavelet, Log Scale, 2 Sample time resolution.

Rotating Wavelet Display

If you prefer to have the wavelet map displayed with time as the x-axis and frequency as the y-axis, then checking the **"Rotate"** checkbox will rotate (replot) the whole display, including the "Measured SPL" and "Wavelet SPL" curves.

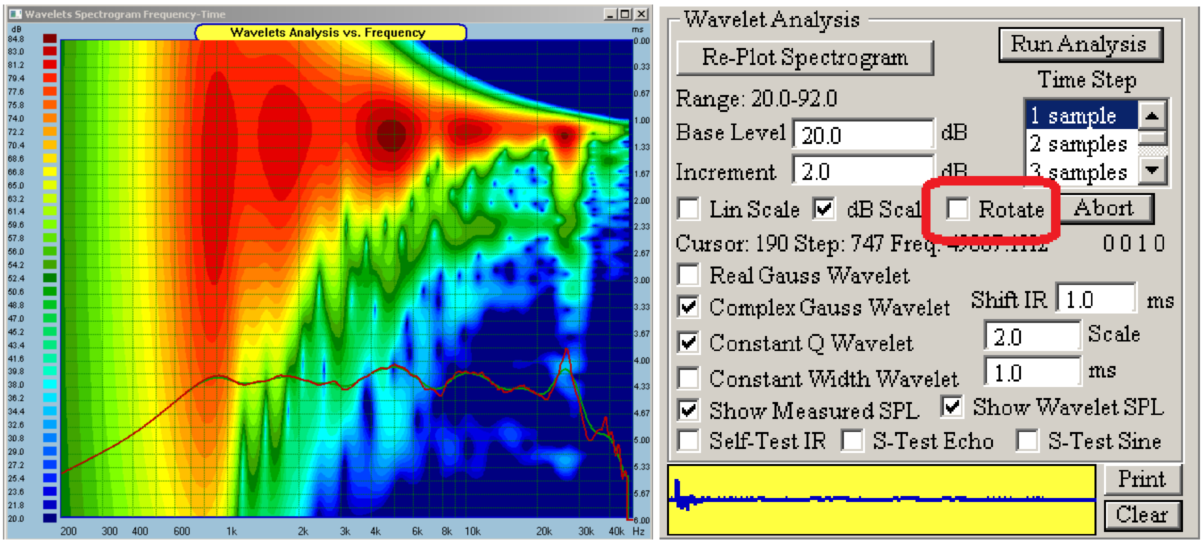


Figure 16.161. Standard Wavelet Display.

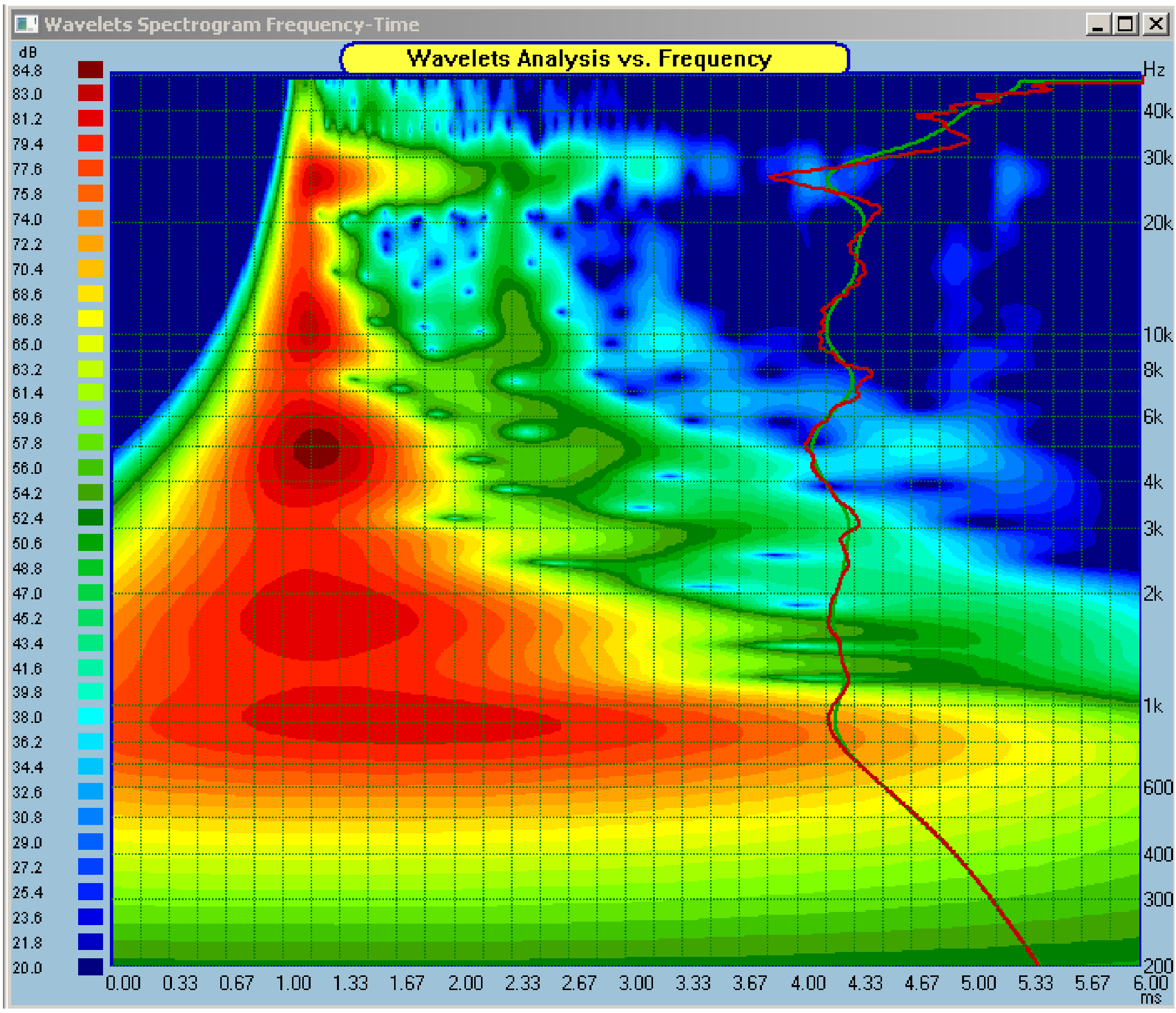


Figure 16.162. Rotated Wavelet Display – the display is re-drawn for this to happen.

Adjusting “Scale” Variable

The Scale variable can be used to optimize the wavelet mapping within the frequency range of most interest. Higher scale factors will move the focus into higher frequencies.

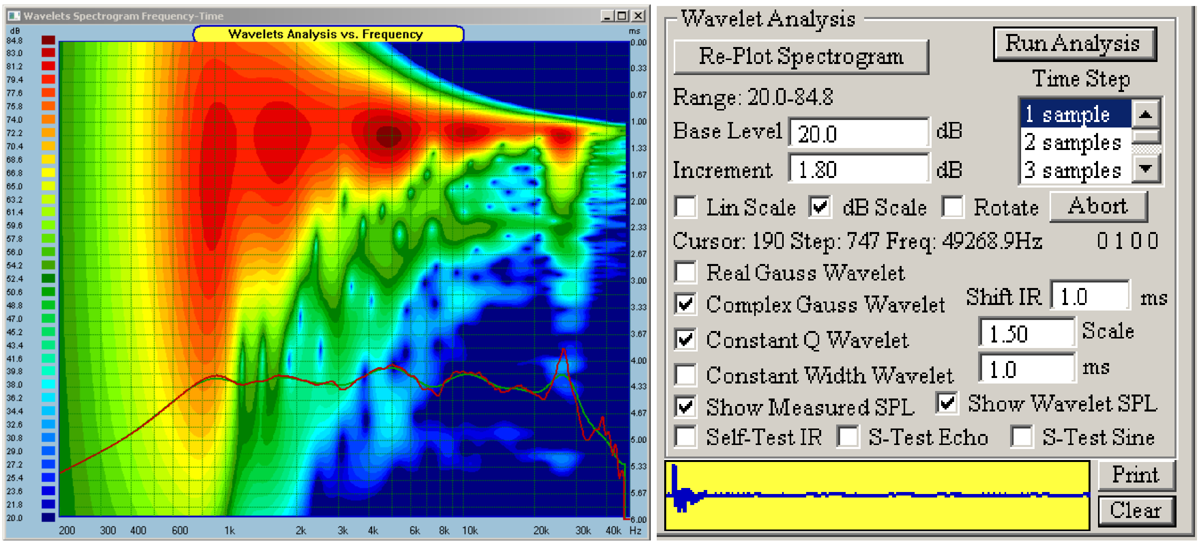


Figure 16.163. “Scale” factor changed from 2.0 to 1.5

While reducing the Scale factor, it is observable, that the SPL calculated via Wavelet Transform (green SPL curve) will get smoother at higher frequency range, thus reducing the amount of details projected by the wavelet mapping. Therefore, it is recommended to balance the Scale setting for the best detail resolution within the frequency range you are most interested.

Dissemination of Impulse Response

Information contained in an Impulse Response (IR), shown below as the red curve, can be disseminated in several ways.

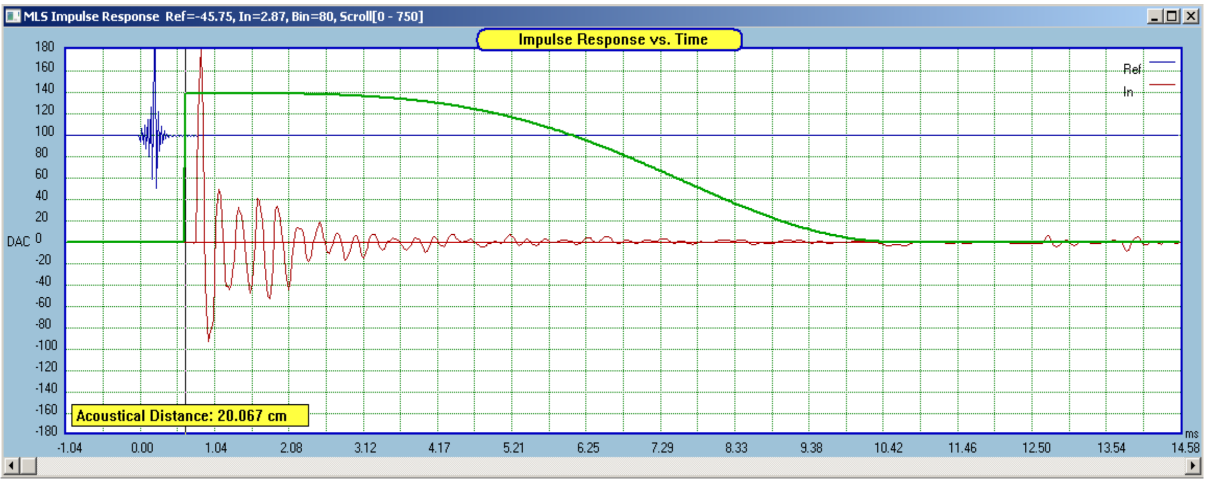


Figure 16.164. Example of a measured Impulse Response.

The most popular way is to transform the IR into frequency domain, so that frequency response can be obtained.

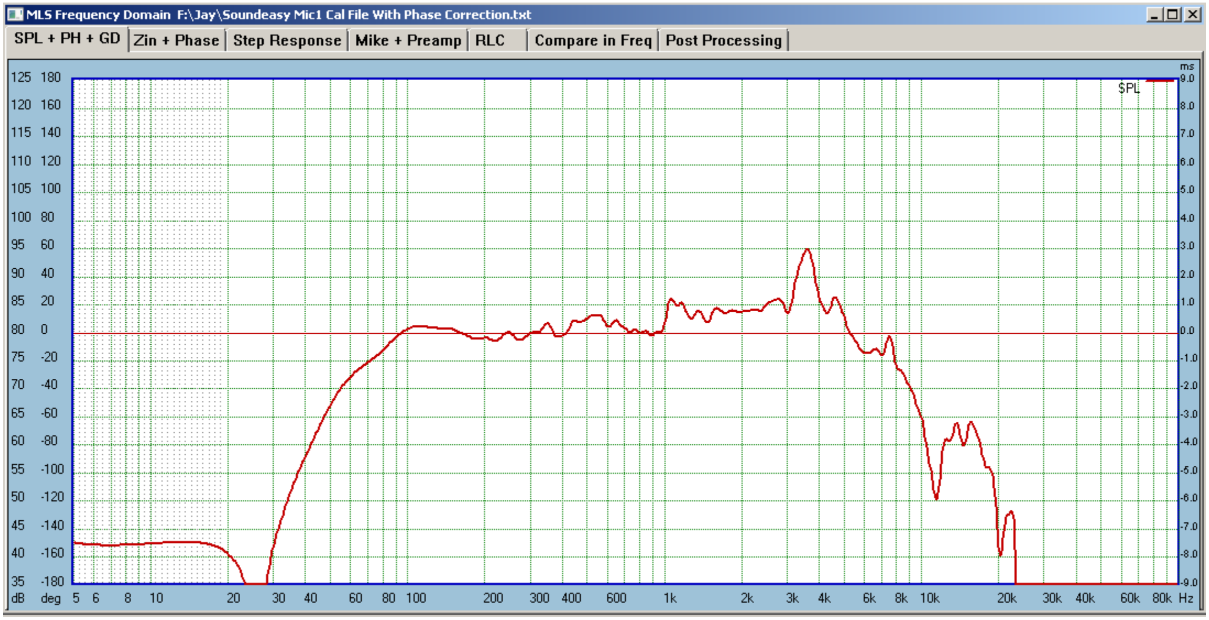


Figure 16.165. Frequency Response corresponding to IR from Figure 16.164.

However, as mentioned before, the well-known Fourier Transform (FT) shows what frequency components (spectral components) exist in the signal. That's all. It does not show at what time those frequency components were occurring.

Somewhat more information can be extracted from the IR by employing time-shifted Short Time Fourier Transform (STFT) and presenting the results as Cumulative Spectral Decay plots shown on figure below.

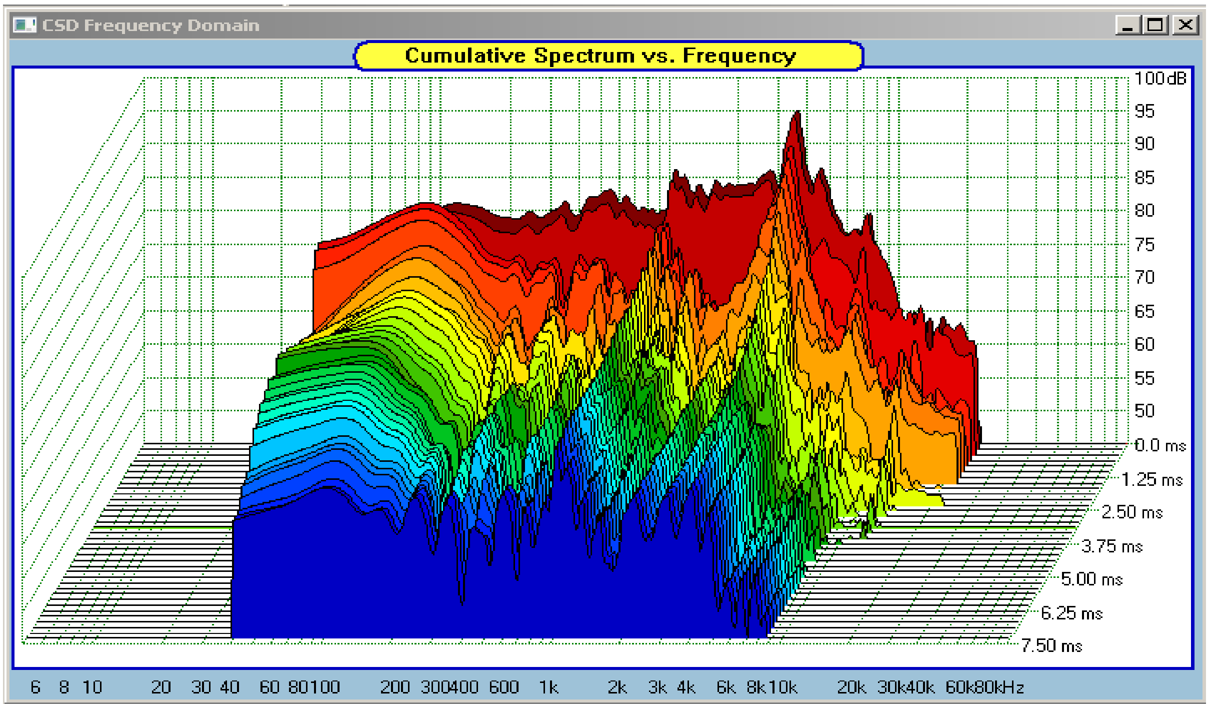


Figure 16.166. CSD plots corresponding to IR from Figure 16.164.

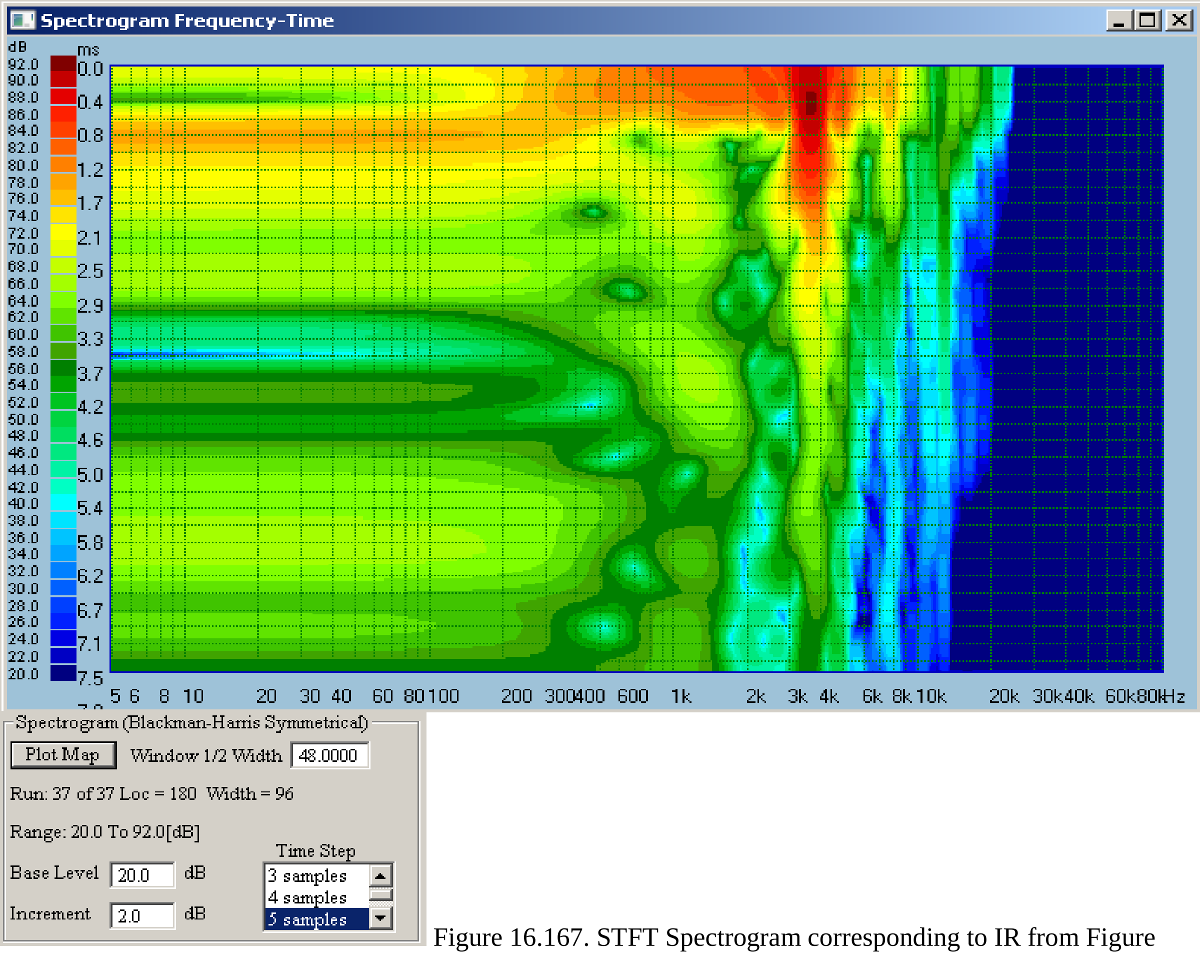


Figure 16.167. STFT Spectrogram corresponding to IR from Figure 16.164. More accurate time-frequency representation of IR can be accomplished via Wavelet Transform – see Figure 16.168 below. Please note, that WT vertical scale is in shown in decibels.

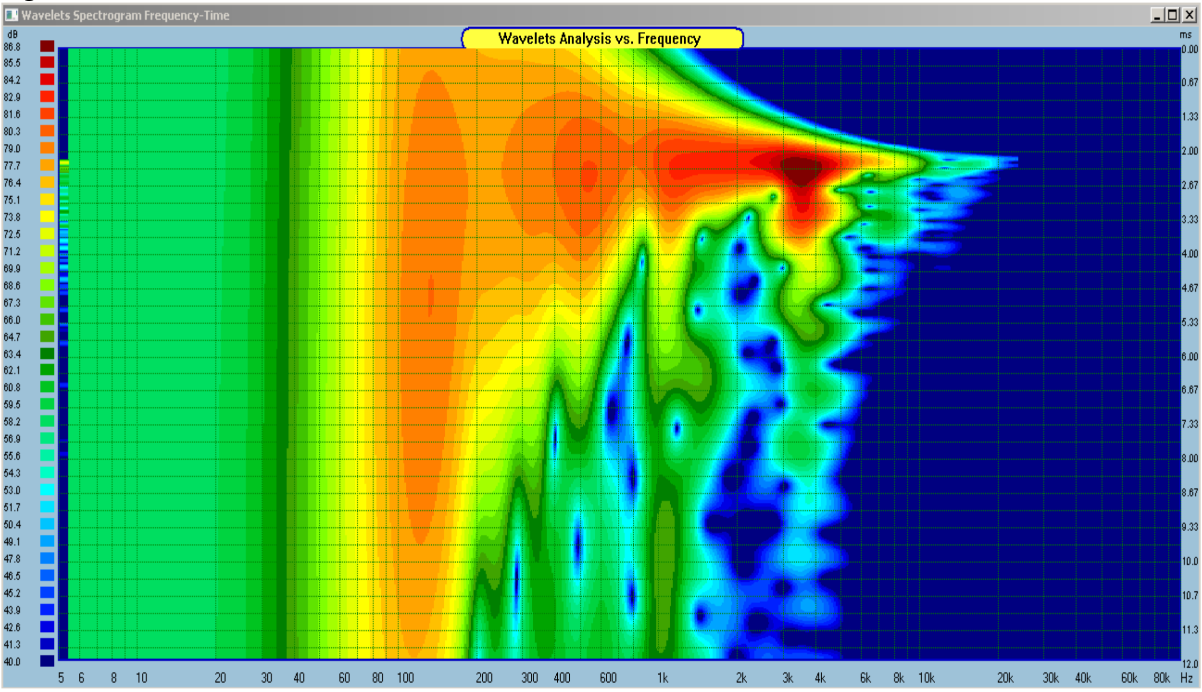


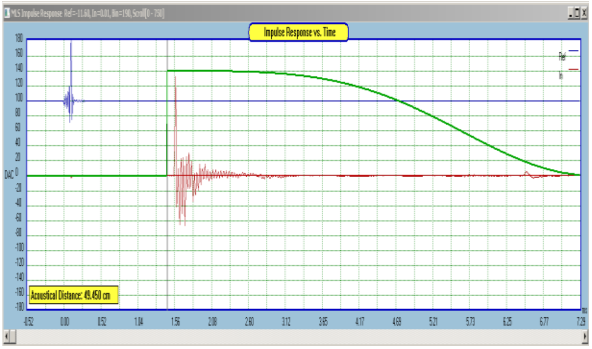
Figure 16.168. Wavelet Analysis plots corresponding to IR from Figure 16.164.

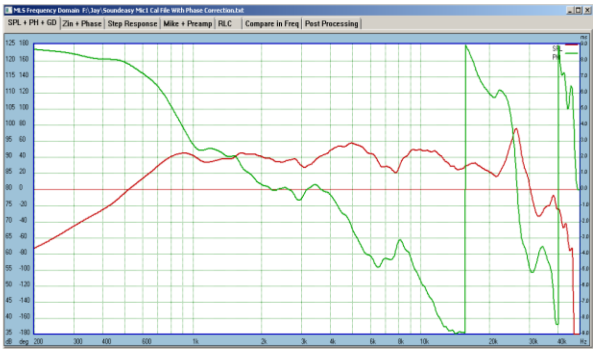
In summary, the IR can be represented by the following transformations:

(1) Frequency response, (2) CSD, (3) sliced CSD Spectrogram based on STFT and (4) Wavelet Transform

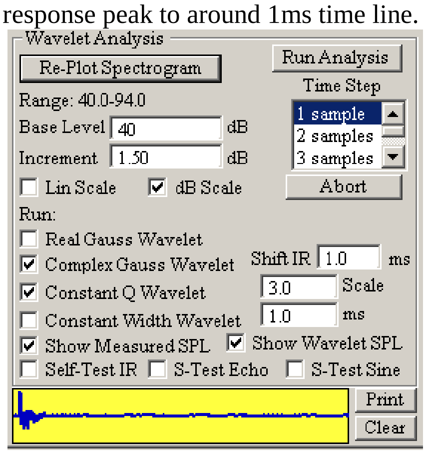
Usage With Minimum-Phase Systems

Perform typical MLS measurement, resulting in collected IR and select and position appropriate FFT window – see example below





Go to CSD tab and select appropriate parameters for your Wavelet analysis – example below. Please note 1ms Shift IR to move the impulse response peak to around 1ms time line.



Then press "Run Analysis" button. It will take some time to complete the analysis.

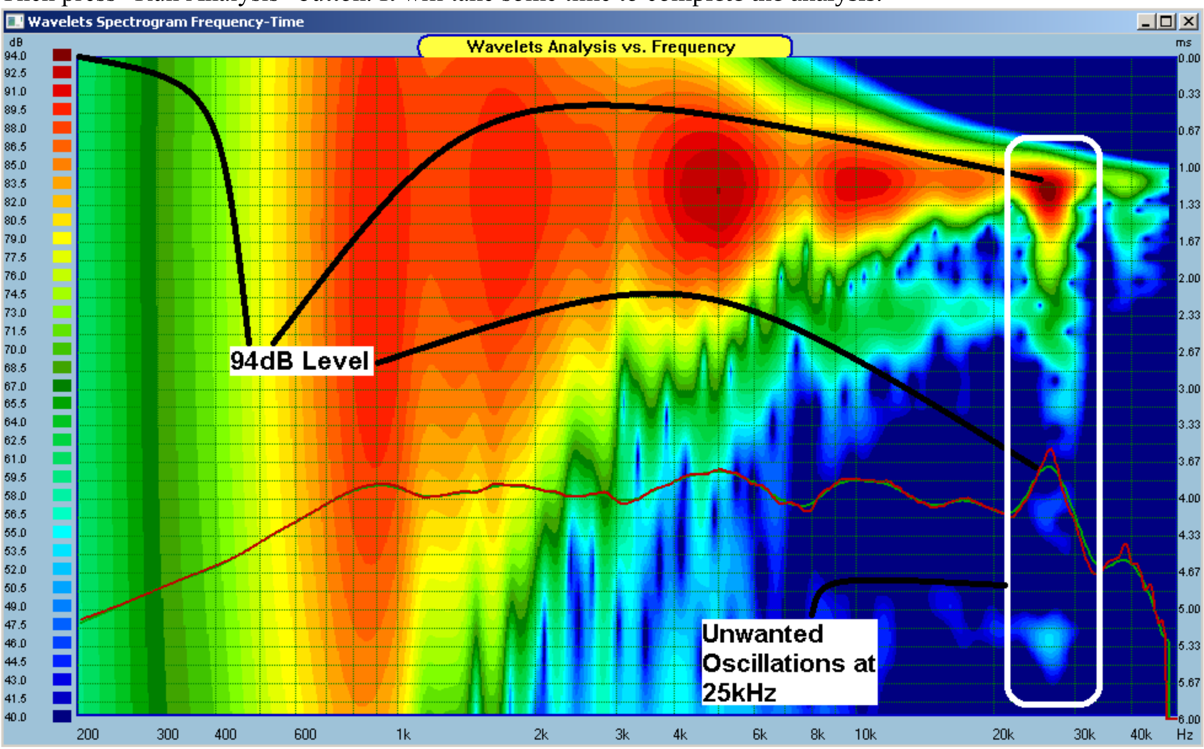
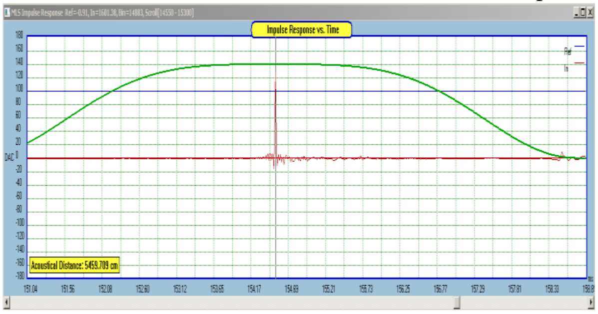


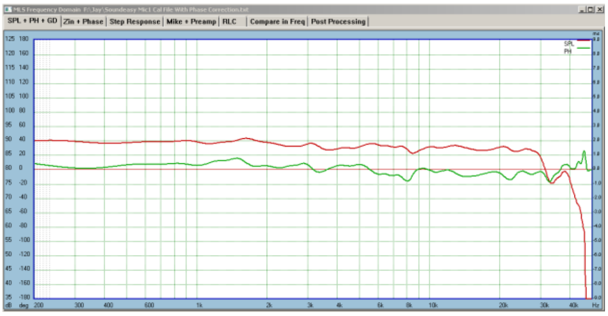
Figure 16.169. Top 50dB of the tweeter's performance

Centre frequency of the Morlet Wavelet is selected accordingly to the horizontal frequency axis frequency points. Vertical time scale is determined by selecting “Time Step” size.

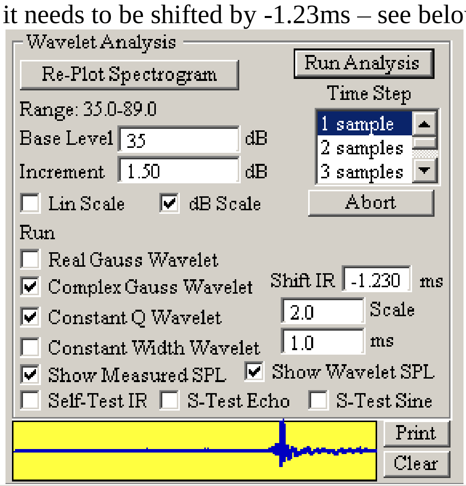
Usage With Linear-Phase Systems

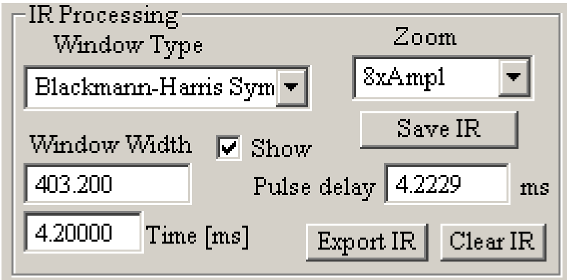
1. Perform typical MLS measurement, resulting in collected IR and select and position appropriate FFT window. Please note, that the actual width of the symmetrical window is always twice the “Window Width” shown – see example below.





The FFT window is symmetrical, and of 4.2229ms on each side of the peak. In order for the IR peak to appear in the centre of the Wavelet Analysis screen (3ms), it needs to be shifted by -1.23ms – see below.





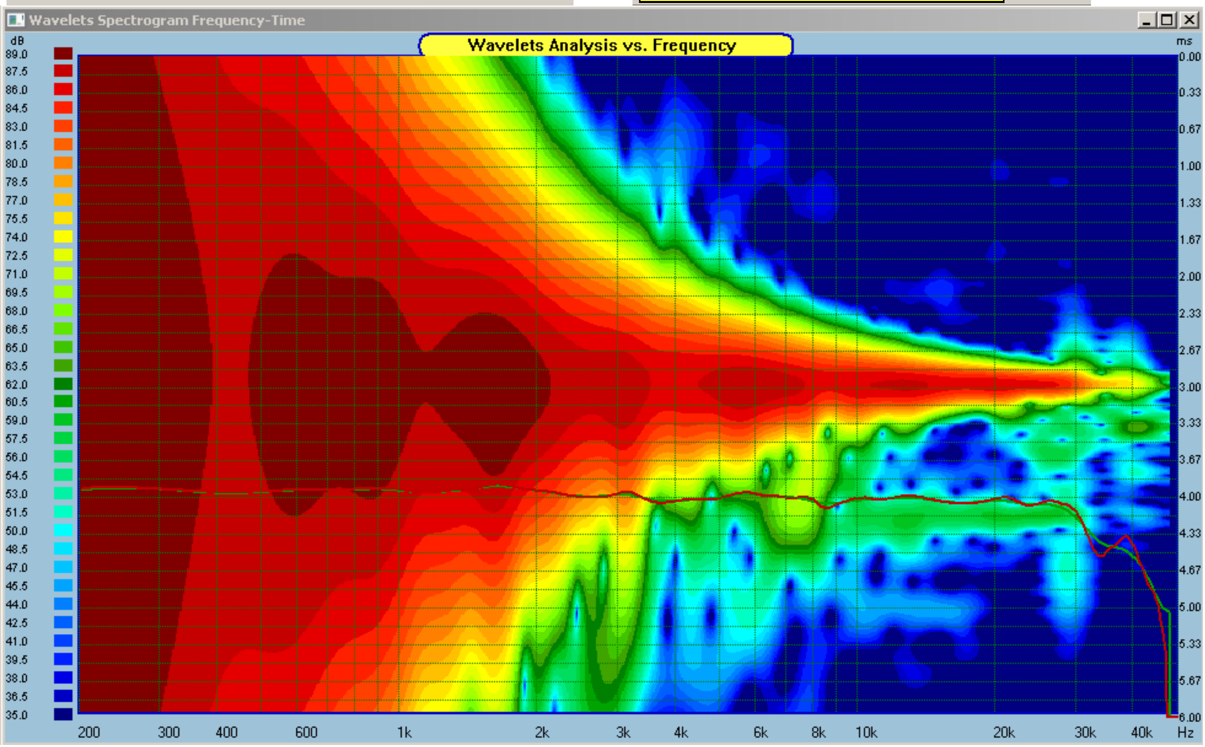


Figure 16.170. Top 50dB of the linear-phase system performance

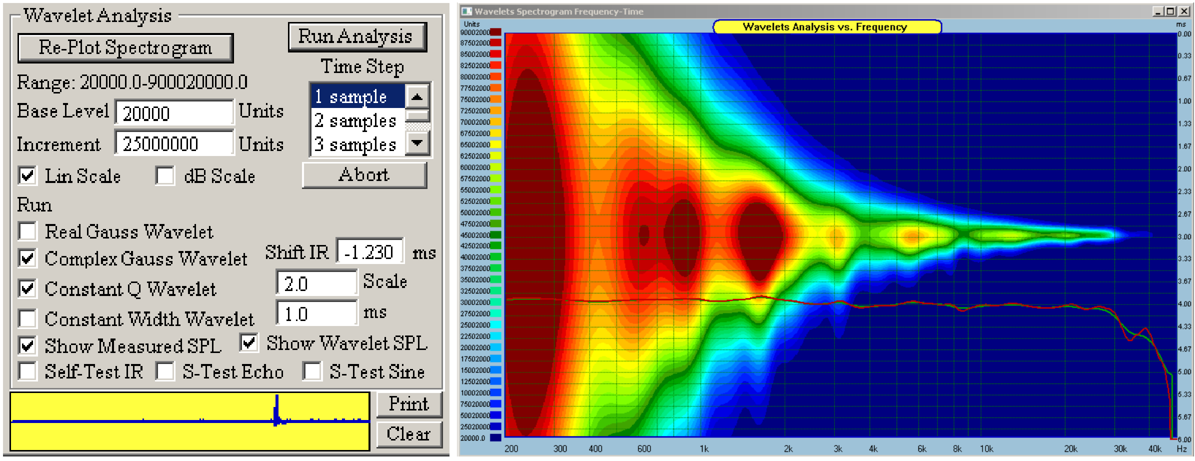


Figure 16.171. The same system shown in Linear Scale.

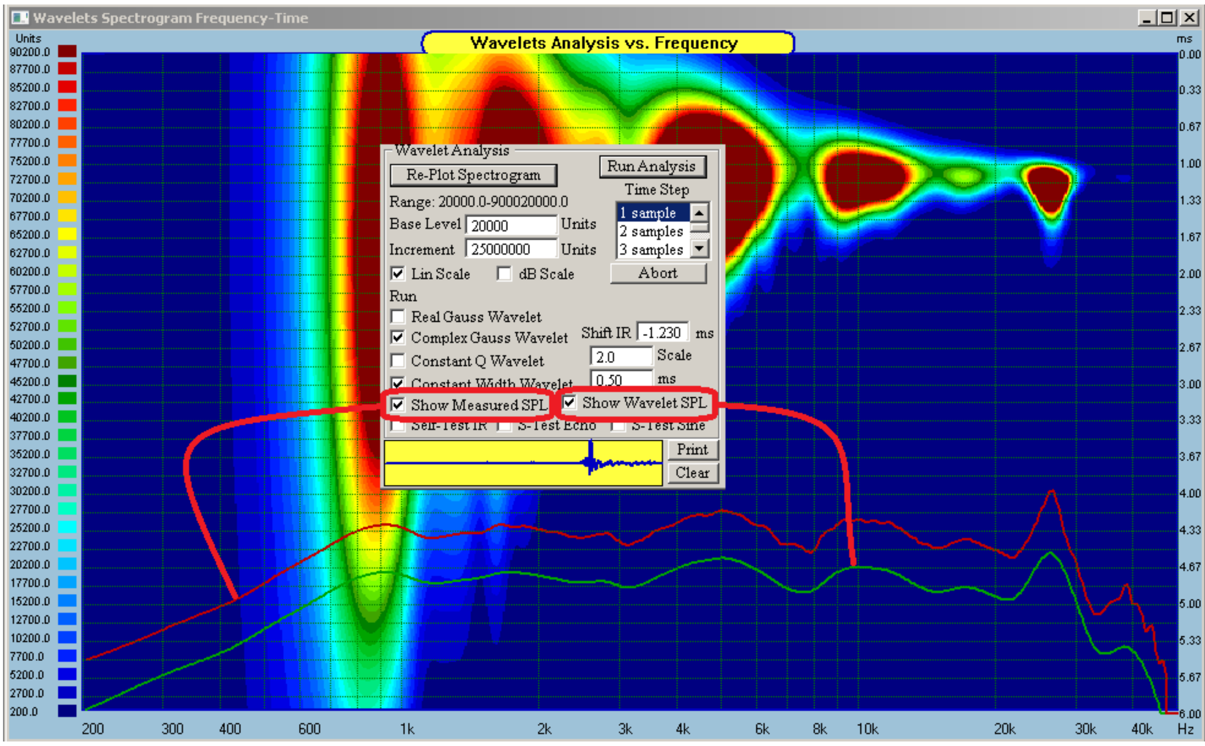
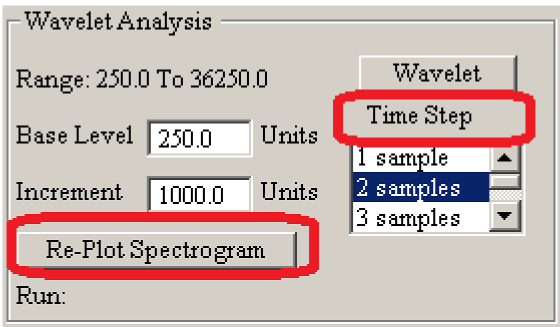


Figure 16.172. "Show Measured SPL" and "Show Wavelet SPL". Checkboxes control display of the corresponding curves.

There exists a relationship between the length of the vertical scale and the selectable "Time Step" parameter: Vertical scale = (74 / Sampling_Frequency) * Time_Step, for example: Vertical scale = (74 / 48000) * 2 = **3.0ms.** It is therefore important to select sufficient Time_Step value for the vertical scale to cover 2 x Windows Width (shown in MLS tab). The length of the vertical scale can be checked upfront by selecting Time Step value from the list box, and then pressing "Re-Plot Spectrogram" button – the vertical scale will be re-adjusted then.



Graphical Improvement to CSD plots

In the previous revisions of CSD plots, when the plotting window was enlarged to full-size screen, the graphics were somewhat disjoined – see Figure 11 below.

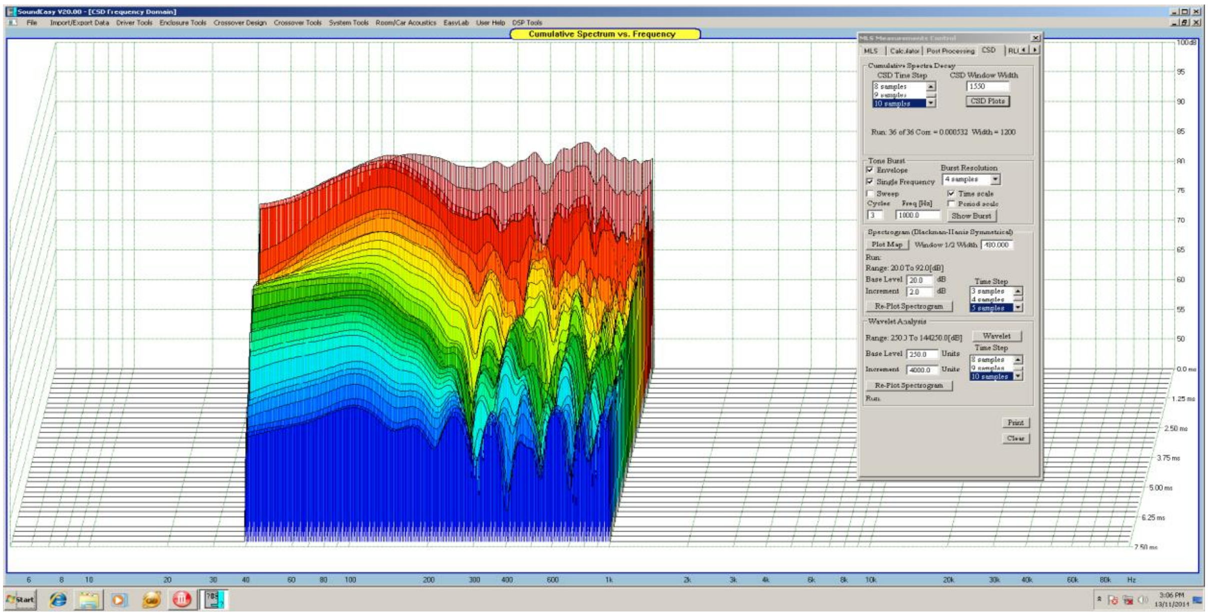


Figure 16.173. CSD plots, when the plotting window was enlarged to full-size screen

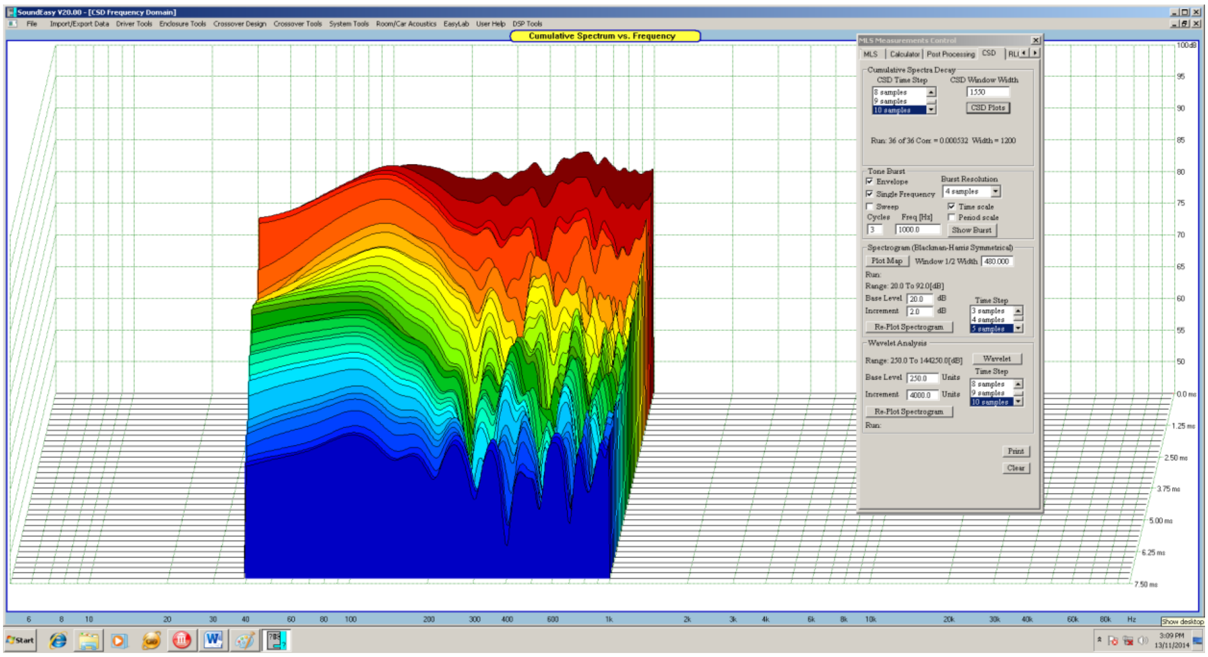
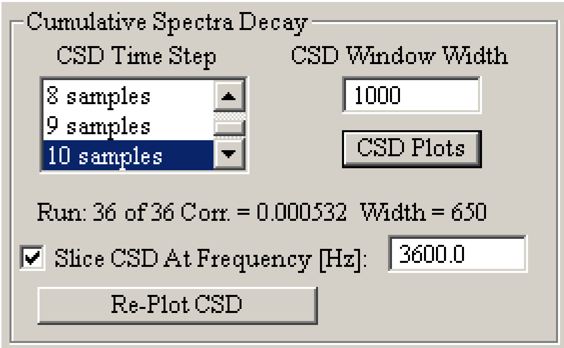


Figure 16.174. New CSD window - CSD plots, when the plotting window was enlarged to full-size screen.

Sliced CSD

In CSD analysis, it may be beneficial to trace some offending ridge along the time axis. This way one can better visualize it's irregular amplitude in the time domain. This process can be accomplished by "slicing" the CSD at any desired frequency. The example below, shows the CSD plot sliced at 3600Hz, which emphasizes the highly resonant peak on the frequency response at 3600Hz.



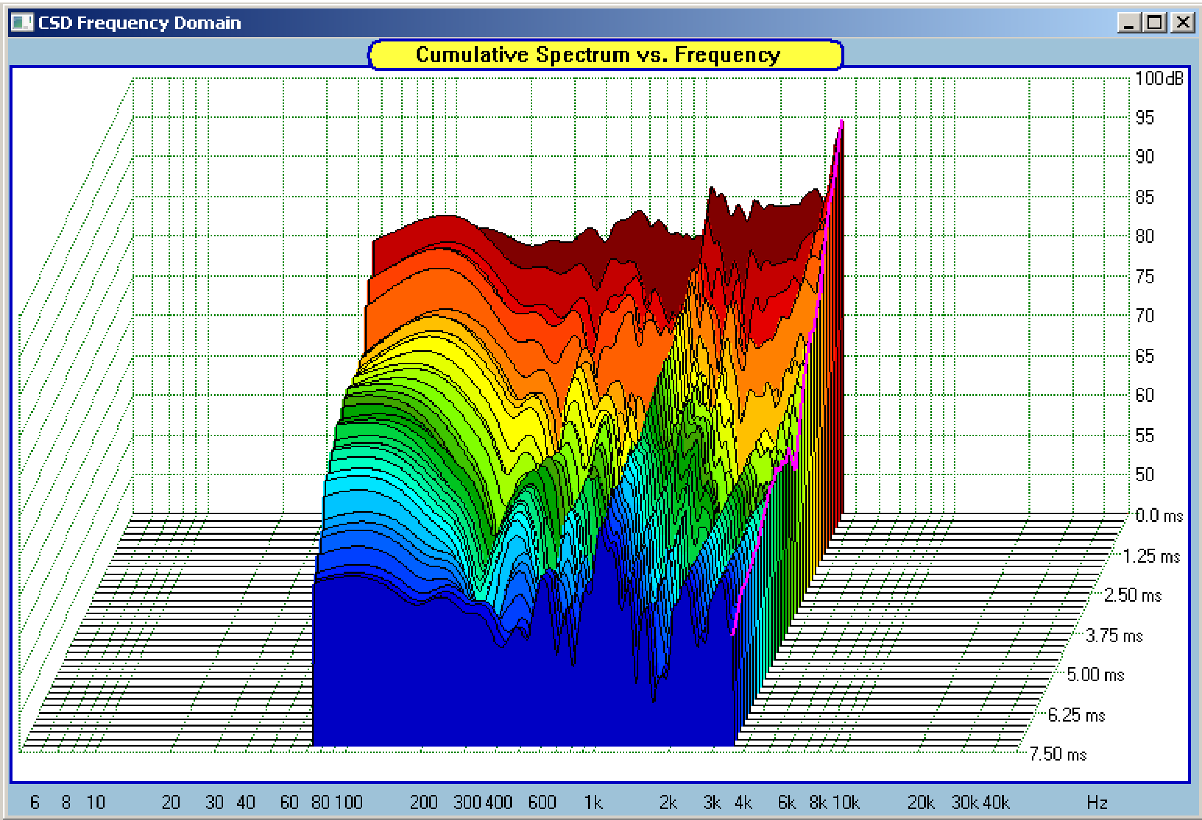
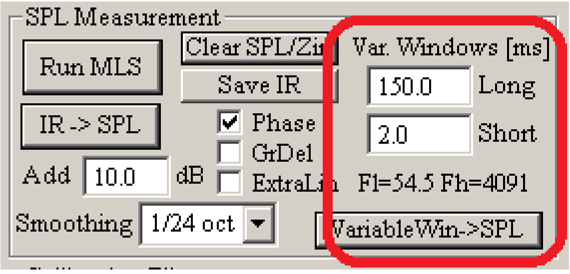


Figure 16.175. "Sliced" CSD with the ridge highlighted in pink.

Adaptive Windowing

With the Adaptive Window feature the length of the time window applied to the measured room impulse response varies, smoothly and continuously, between two pre-defined lengths.



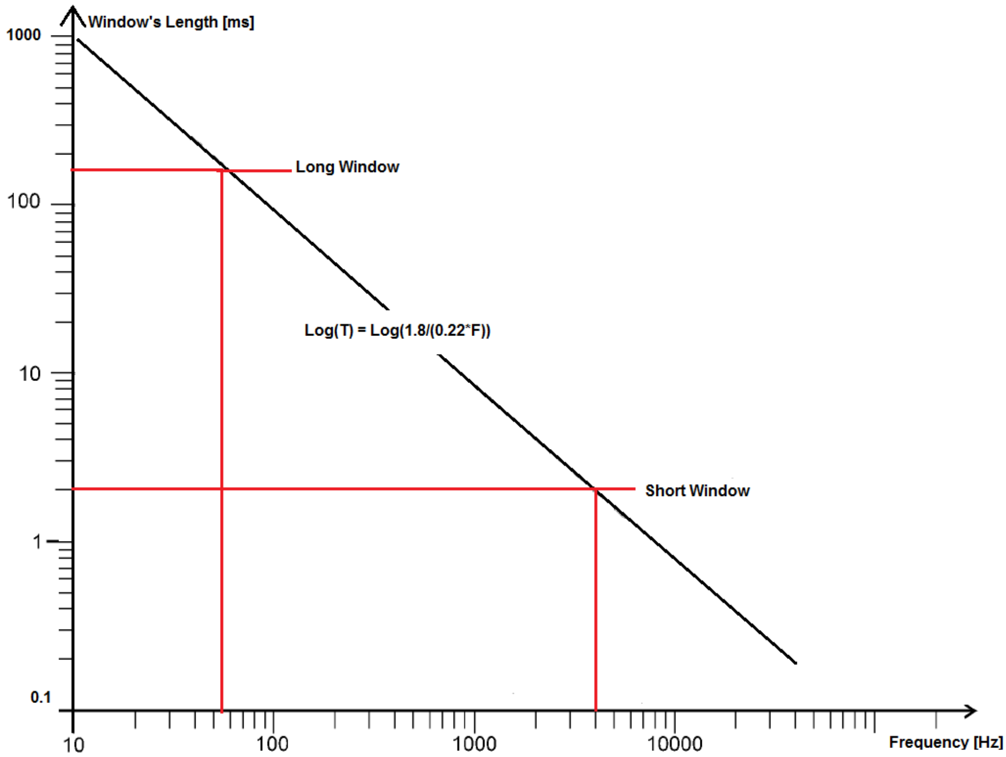


Figure 16.176. Defaults for Long Window and Short Window.

The lengths of both windows are set manually from the MLS dialogue box. The Short Window is set to exclude almost all reflections in the measurement location – the default value is set to 2ms. The long window (set to 150ms default) captures nearly all room reflections, including the late arrivals. The window starts at the same position as the short window but is more them 150ms long. In the frequency range between these two extremes, the algorithm automatically varies the time window length to yield a frequency resolution of 1/3 octave, which approximately corresponds to the width of the ear's critical bands.

Literature suggests, that one should not include reflections later than roughly 500ms because the ear tends to perceive such very late reflections as reverberation or ambience.

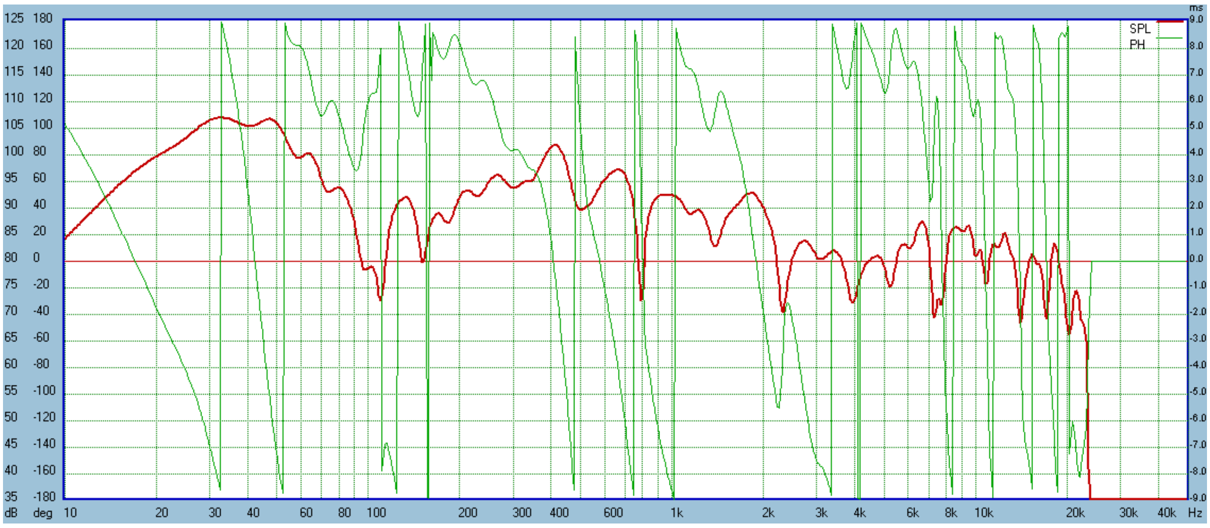


Figure 16.177. SPL calculated with Adaptive Windows: 2ms (Short Window) and 150ms (Long Window).

The adaptively windowed room frequency response therefore excludes all but the earliest room reflections at high frequencies, gradually includes more reflections in the midrange and finally includes nearly all room reflections in the bass region. The Adaptive Window algorithm also preserves phase response – see Figure 3.

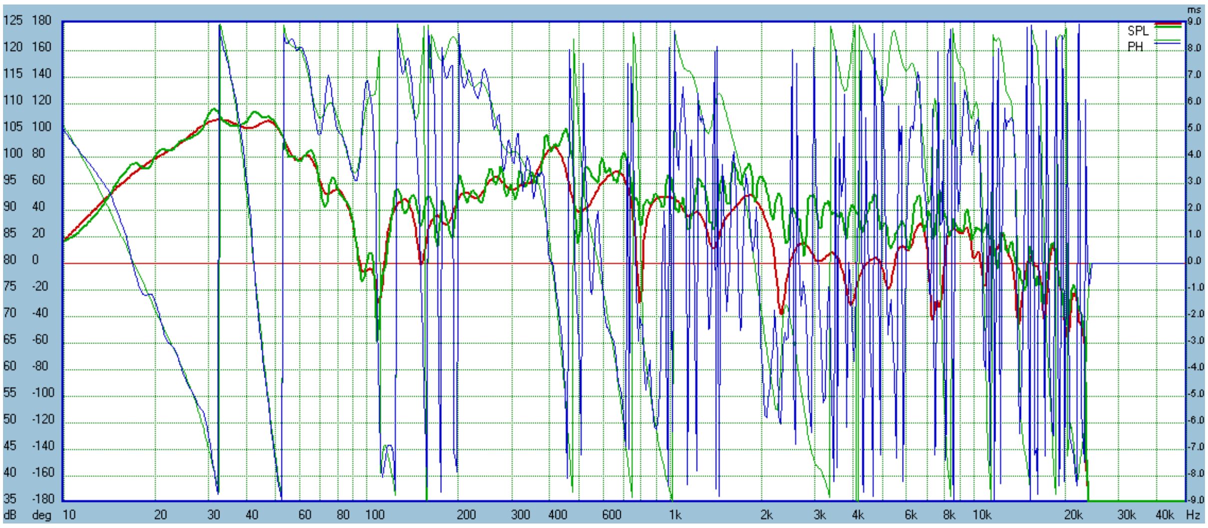


Figure 16.178. Adaptive Window algorithm preserves phase response.

Control Dialogue

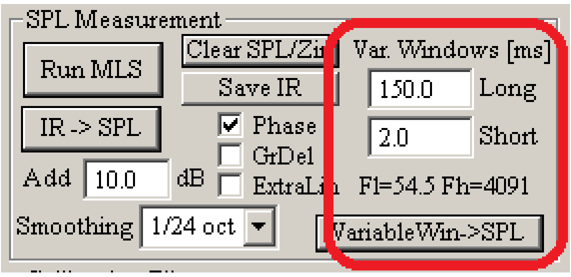


Figure 16.179. Controlling Adaptive Windowing algorithm.

The Adaptive Windowing algorithm can only be activated after the room impulse response has been measured. For more information, please consult appropriate sections of SoundEasy program manual.

Next, you would select the lengths of the Long Window and the Short Window from the MLS control dialogue box – see Figure 4.

Finally, you need to position the cursor marker on the Impulse Response window where you wish the windowing should start, and then press the “**VariableWin->SPL**” button to start the algorithm.

Now, for each of the 750 frequencies of the SPL curve, the algorithm selects different length of the FFT window – as per curve on Figure 1. The whole SPL is then re-calculated from the impulse response. This process can be quite lengthy. Particularly, when you select longer MLS sequences (and therefore FFTs) for the measurements. The current frequency and the intermediate results are shown on the SPL window, and will be cleared once the process is completed for all 750 frequency bins.

Comparison with standard measurements

Figure 5 shows comparison between SPL recovered via Adaptive Window SPL (red) with a standard 500ms “fixed window” (green).

It is observable, that at the very low-end and very high-end both SPL curves are very similar. However, they differ significantly everywhere else.

The differences are most pronounced in the midrange frequency range. Here, the SPL calculated via the Adaptive Windowing algorithm sits several dBs below the SPL measured with single, wide window.

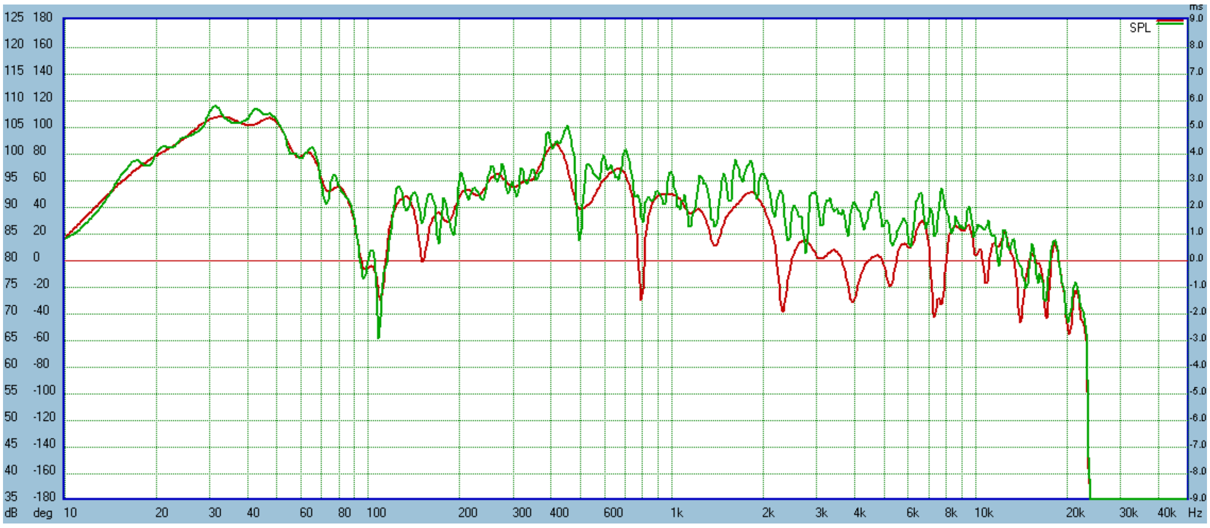


Figure 16.180. Comparison between Adaptive Window SPL (red) with a standard 500ms “fixed window” (green).

Similar conclusions can be drawn from MLSSA measurement system – see picture below.
http://www.mlssa.com/pdf/DRA-MLSSA-Manual-10WI-8.pdf

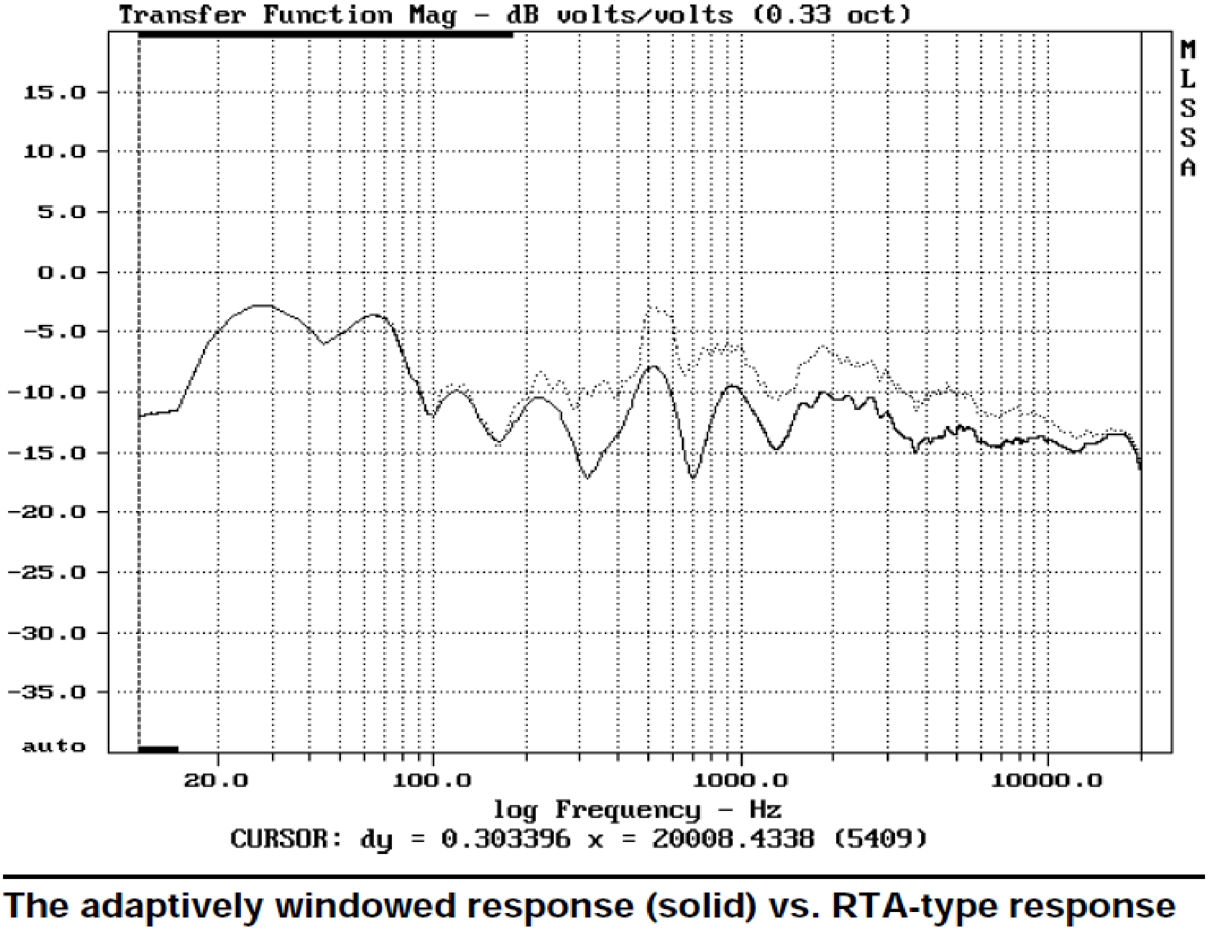


Figure 16.181. Example of Adaptive Windowed measurements vs. fixed window.

This offset creates an interesting dilemma for those, who would like to use the SPL calculated via the Adaptive Windowing algorithm as the input curve for calculating room correction filter as the minimum-phase inverse of the room response.

If the measured room response has a general, very broad dip in the midrange frequencies (as calculated by Adaptive Windowing algorithm), then the inverse correction filter will have a broad peak in the midrange frequencies. As a result, your equalized system may sound overly bright.

You can re-measured the equalized system, once again, using the Adaptive Windowing algorithm scheme, and it will show nice and flat room frequency response, but it may sound overly bright.

MLS SPL calibration using sine-wave

Maximum Length Sequences (MLS) are easy to generate digital sequences, and possess interesting properties, that make them really interesting in the field of acoustic measurements.

Being digital signals with binary +1/-1 levels, its root-means-square (RMS) level is, theoretically, equal to its peak level – meaning, that the crest factor is unity (CF=1). This is exactly what we would need for generating maximum power from our amplifiers and loudspeakers, used for Room Impulse Response (RIR) measurements in large concert halls and venues.

The measurements can be affected by:
1. Background noise
2. Nonlinearities
3. Time variance in the system under test.

Background Noise

Theoretically, and MLS provides the maximum excitation level for a given amplitude (low crest factor) and therefore the highest obtainable SNR. However, in practical applications, the digital MLS stream must be low-pass filtered before it's delivered to the loudspeaker, and this process will degrade the crest factor (CF). Some sources quote deterioration in CF as large as 10-12dB, however, I have measured the CF on my sound card, and the results are shown below. The CF is somewhat dependant on sampling frequency and the length of MLS signal. Here are some examples measured on Delta1010LT sound card:

MLS length	Sampling	Vpp	Vrms	Crest Factor
64k	48k	7.25V	1.91V	5.60dB
64k	96k	6.56V	1.84V	5.02dB
132k	48k	7.25V	1.74V	6.30dB

Please note that the crest factor for sine-wave is 3dB. Therefore, the deterioration of CF due to using MLS signal is only about 3dB. The available SNR will be reduced by the same amount. However, the common practice to improve the SNR is to average repeated measurements. The SNR improves by approximately 3dB if the number of averaged measurements is doubled. So, two consecutive measurements will improve the SNR by 3dB, four consecutive measurement will improve the SNR by 6dB and so on. It may be interesting to visualize the MLS and sine-wave with identical CF and RMS values – see figure below.

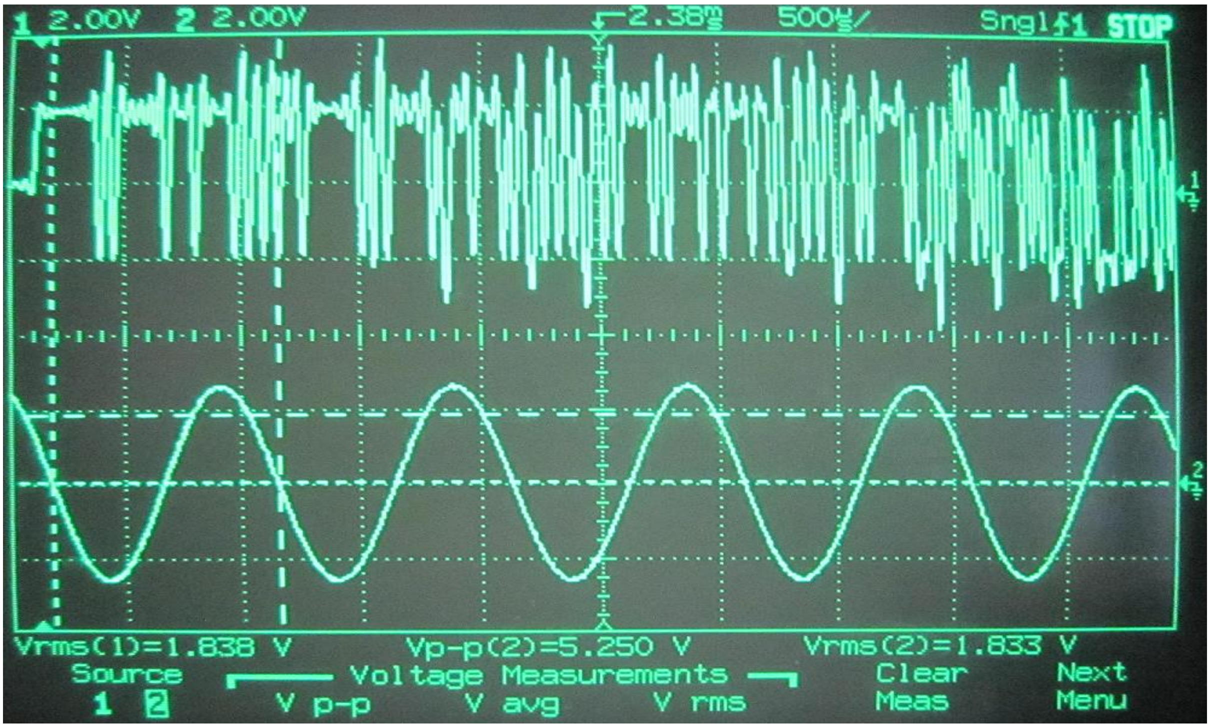


Figure 16.182. MLS and sine-wave (1000Hz) with almost identical RMS values.

Nonlinearities

If the system to be measured exhibits nonlinearities, the measured impulse response will be affected. The nonlinearities manifest themselves as small spikes along the decaying tail of the impulse response.

The position of these spikes can not be predicted, however, for a particular system, the appearance of these spikes is directly linked to a particular MLS sequence used, and repeated measurements will always see them in the same location- so they can not be reduced by simple averaging.

There exists quite effective way of dealing with nonlinearities in MLS systems. The idea is to perform repeated measurements using different, cyclically distinct, MLSs, so that the nonlinearity peaks will pop up in different locations. Then simply apply averaging. Each time the number of averages is doubled, the noise due to nonlinearities is reduced by approximately 6dB. Even a relatively small number of repetitions will push the nonlinearity peaks into the noise floor, so there is no need to be concerned about them any more.

It is possible to provide an MLS scheme wich is capable of largely eliminating nonlinearities. However, I have not encountered any instances, where such scheme would be needed.

Time variance

Time variance can be divided into:

1. Inter-periodic – effects that occur between two measurements, for instance, slow temperature changes.
2. Intra-periodic – effects that occur within an MLS measurement period, like wind and clock jitter.

Slow temperature changes are typically not a problem in loudspeakers driven at 1Watt/1meter measurement scenarios.

Wind can be a factor in outdoor measurements, so there is an obvious need to avoid windy conditions – this would be true for any type of measurements.

MLS Advantages

1. MLS- related signals have a SNR improvement proportional to their period length.
2. They can be used in noisy environment or at low excitation level.
3. Relative high noise immunity and an efficient correlation algorithm have significantly improved RIR measurements, acoustic impedance-tube measurements, absorption coefficients in reverberation chamber even with rotating diffuser, in situ measurements of sound insolation of noise barriers, acoustic surface reflection functions, both indoors and outdoors, measurement of head-related transfer functions and hearing aids, applications in underwater acoustics and landmine detection.
4. The interest in MLS measurement technology has also stimulated investigations on its advantages, such as high-transient noise immunity.
5. Spectrally, MLS signal more closely resembles real musical signal, while sine-wave is just a single spike on the frequency spectrum.
6. Analogue gated sweep is a single-channel, amplitude only measurement, so does not provide phase information. Phase must be recovered with high degree of probability using HBT. However, HBT only works for single driver measurements (because it's minimum-phase). Therefore you can not measure the phase response of a loudspeaker system - which is typically non-minimum phase.
7. On the other hand, the MLS system is dual-channel, will measure phase of anything you throw at it, and even has "coloured" MLS version to measure tweeters without feeding low-frequency spectrum into the driver (driver protection).

Can MLS system be calibrated for DIY?

Process discussed here sets out to determine “efficiency” of the driver. Typically, a 1Watt signal is fed into the driver, and a calibrated measurement microphone picks up the sound at 1 meter distance. The result is presented as some value in decibels. For instance, Efficiency = 95dB would indicated very efficient driver, while Efficiency of 80dB would indicate rather inefficient driver.

Before we bring the MLS into the discussion, its worth mentioning, that the Efficiency of a driver is one of the most basic parameters of a driver, and is typically always provided by the manufacturer.

Secondly, the Efficiency can be easily calculated from TS parameters of the driver. So if it's missing from the specs, then can be easily calculated – see formulas 2.34, 2.35, 2.36 and 2.37 in "Testing Loudspeakers" by Joseph D'Appolito.

Clearly, we are trying here to duplicate something that has already been done by the manufacturer.

OK, but the above process relies solely on sine waves used for measurements. Now, we can feed a sine wave into the MLS system and discover what happens.

Measurement System Set-up

Here is a simple idea: Assuming, that sine-wave and the MLS signal have the same RMS values - Is the MSL system going to respond with ***the same SPL level***, when a sine-wave OR MLS is fed into it?.

In order to check the MLS signals, I have used loopback mode. For sine wave comparisons, I have injected a 1000Hz sine wave from external generator right into the MLS input port of the sound card. Both signals were monitored by a digital storage CRO – see Figure 16.164. Smoothing was 1/12oct, MLS length = 131071 and sampling frequency was 48kHz. These parameters are important.

When the sine wave was used as an input signal, I have used symmetrical Blackman-Harris windows of 20ms length, centered at 10ms – see Figure 16.165 below.

My arbitrary voltage level was RMS = 0.0dB, or 1.8V.

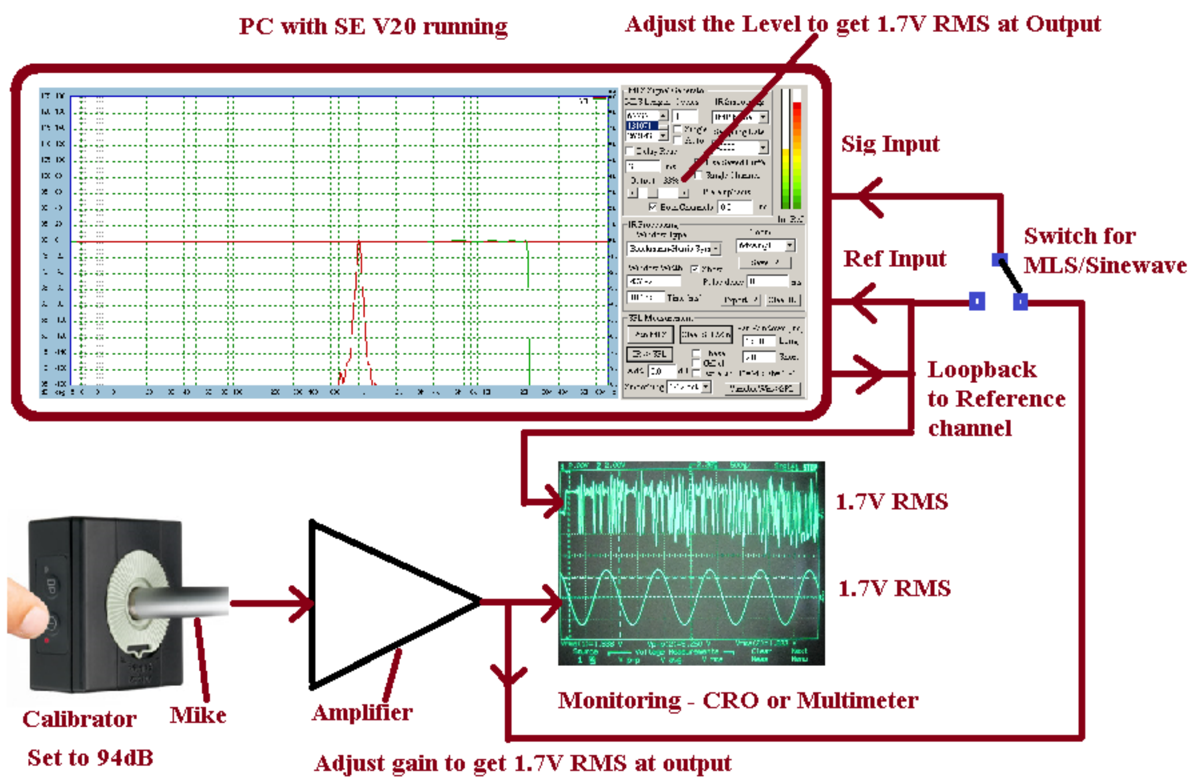


Figure 16.183. Measurement system diagram.

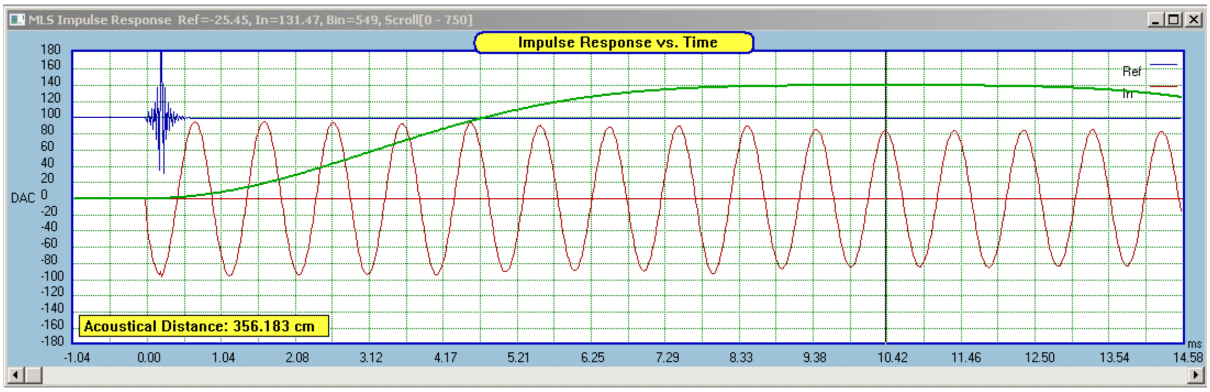


Figure 16.184. Sine wave RMS = 0.0dB (1.802V), Blackman-Harris 10ms long, set at 10ms, Amplitude Zoom=64

When the MLS was used as an input signal, I have used standard Blackman-Harris windows of 100ms length, located at 0ms – see Figure 16.166 below. My voltage level was RMS = 0.0dB, or 1.801V. (must be very similar as for sine wave).

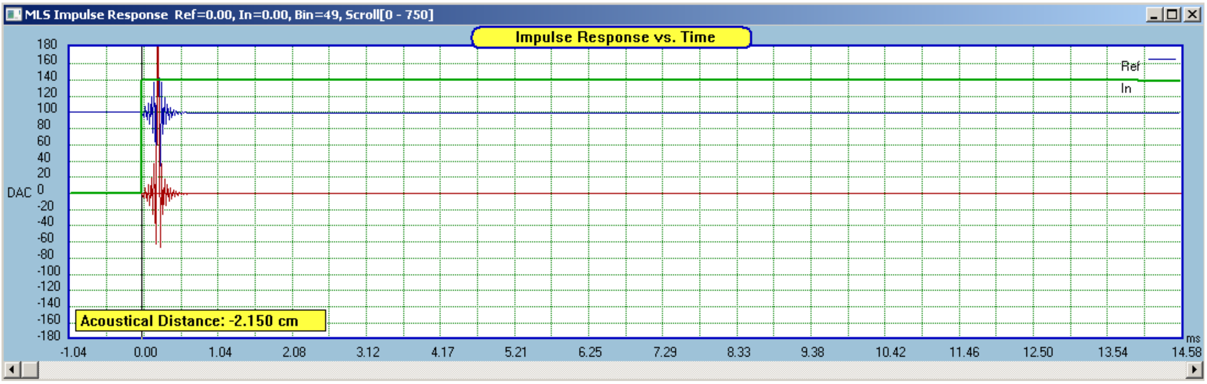


Figure 16.185. MLS generator at 33% (1.79V RMS). Blackman-Harris 100ms long, set at T=0. Amplitude Zoom=1

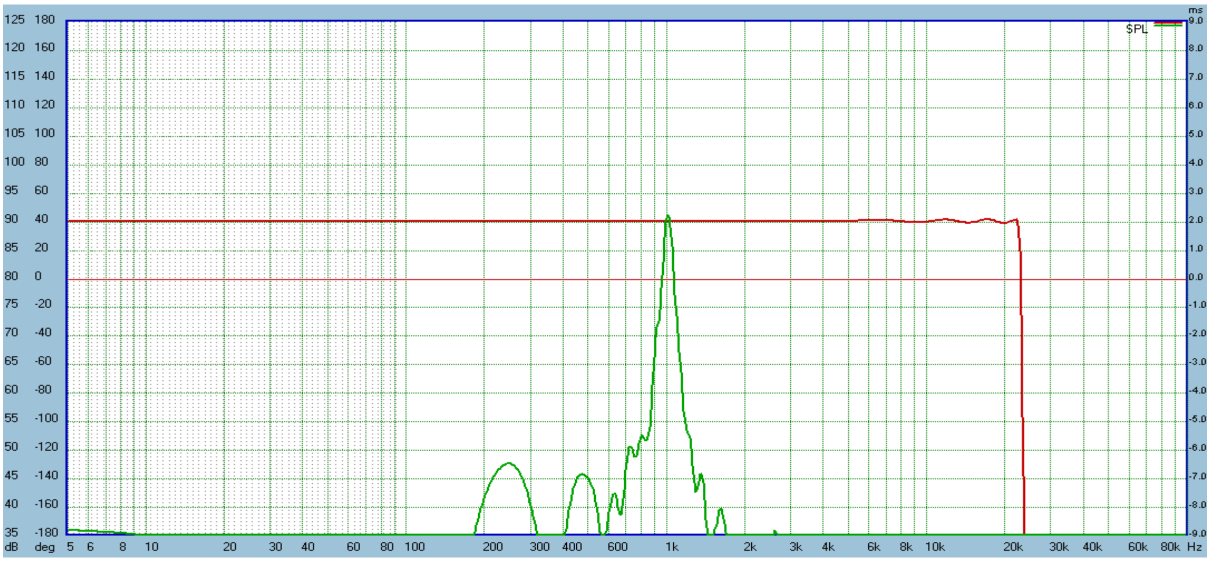


Figure 16.186. Green – sine wave response after FFT, Brown – MLS response after FFT.

Note, almost the same level of 90dB.

In the next step, the levels of both signals were reduced by 10dB, and the measurements were repeated.

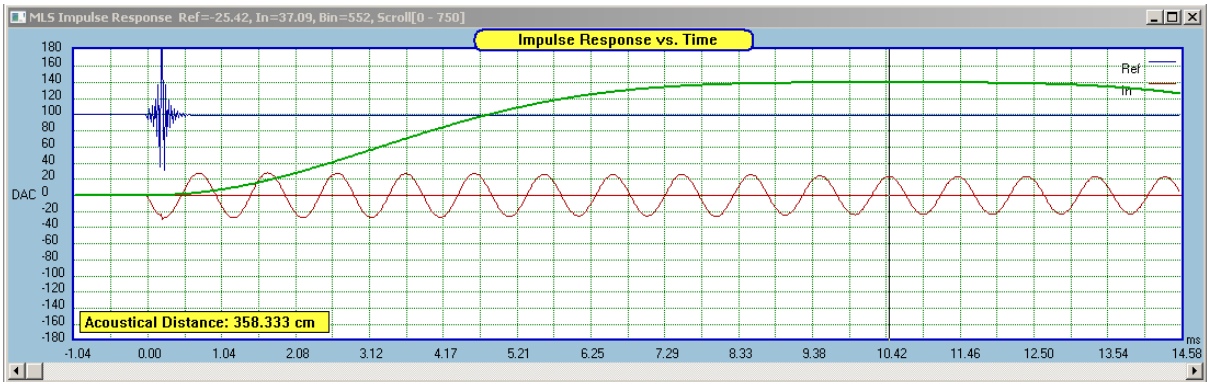


Figure 16.187. Sine wave RMS = -10dB (0.554V), Blackman-Harris 10ms long, set at 10ms, Amplitude Zoom=64

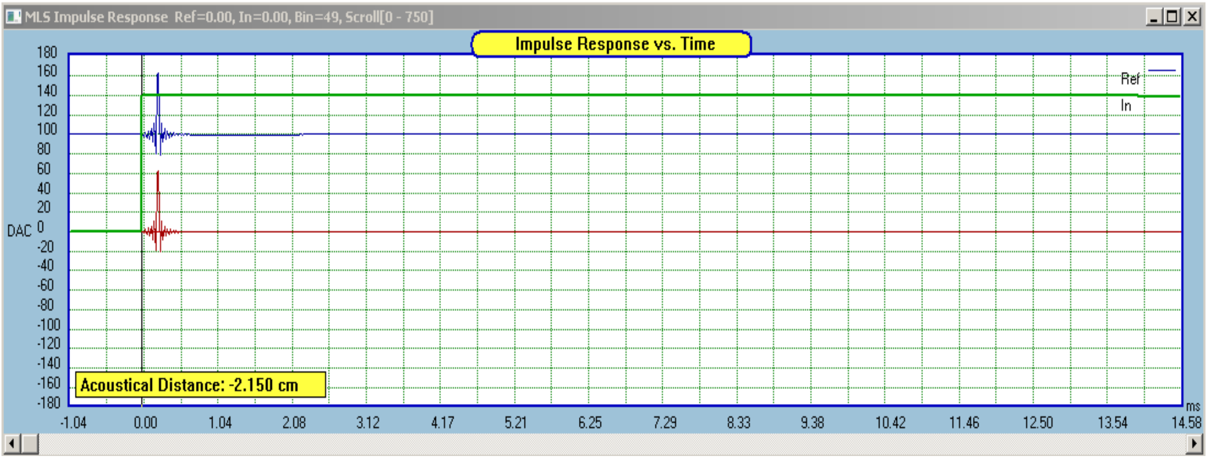


Figure 16.188. MLS generator = -10dB (or 11% or 0.556V) Blackman-Harris 100ms long set at T=0. Amplitude Zoom=1

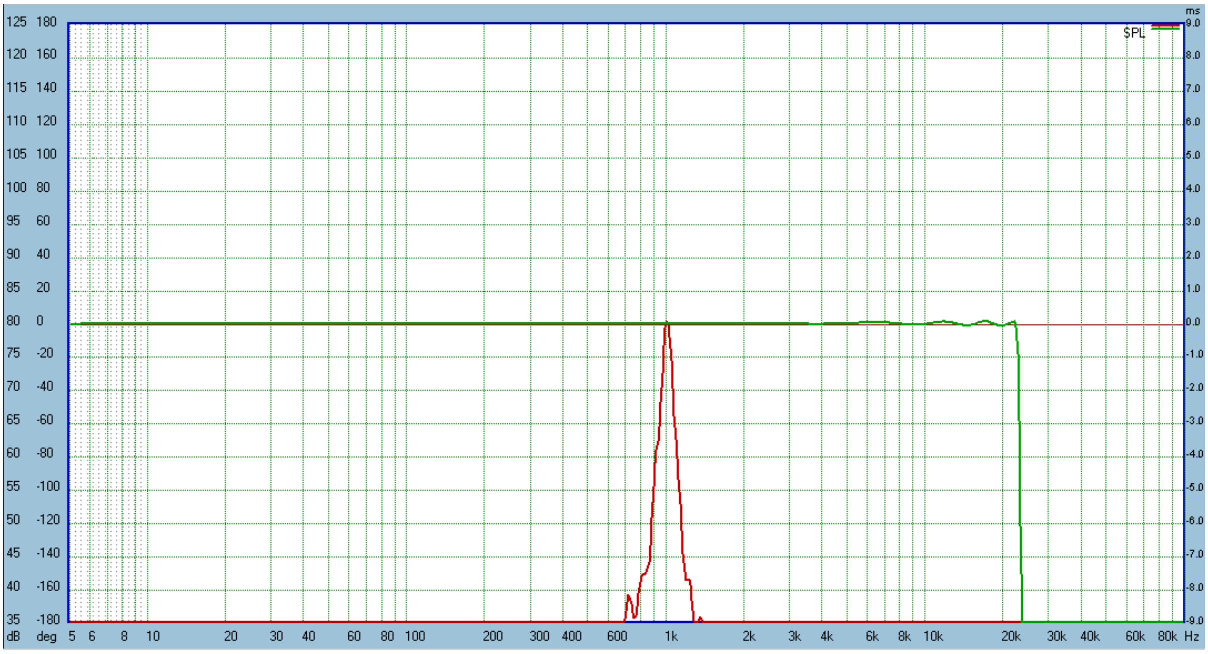


Figure 16.189. Brown – sine wave response after FFT, Green – MLS response after FFT. Note, almost the same level of 80dB.

Interestingly, I can feed an MLS signal with a given RMS level, OR a sinewave with the same RMS level, and BOTH signals will manifest themselves at the same level on the SPL plots – is it stable though?

The process is linear. I could reduce the level of both signals by -10dB, and obtain the same 10dB reduction in both SPL displays.

Further investigation

It strongly recommended to conduct some investigation of your PC+MLS system in order to confirm the findings shown below.

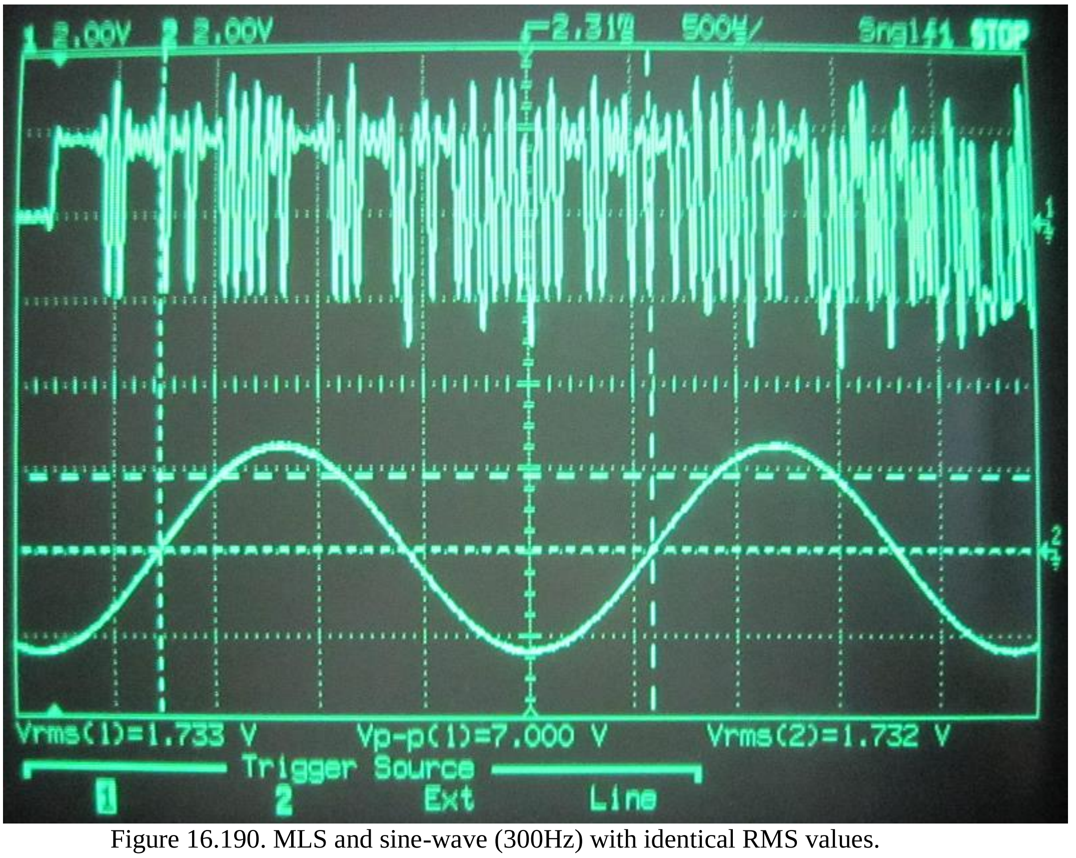


Figure 16.190. MLS and sine-wave (300Hz) with identical RMS values.

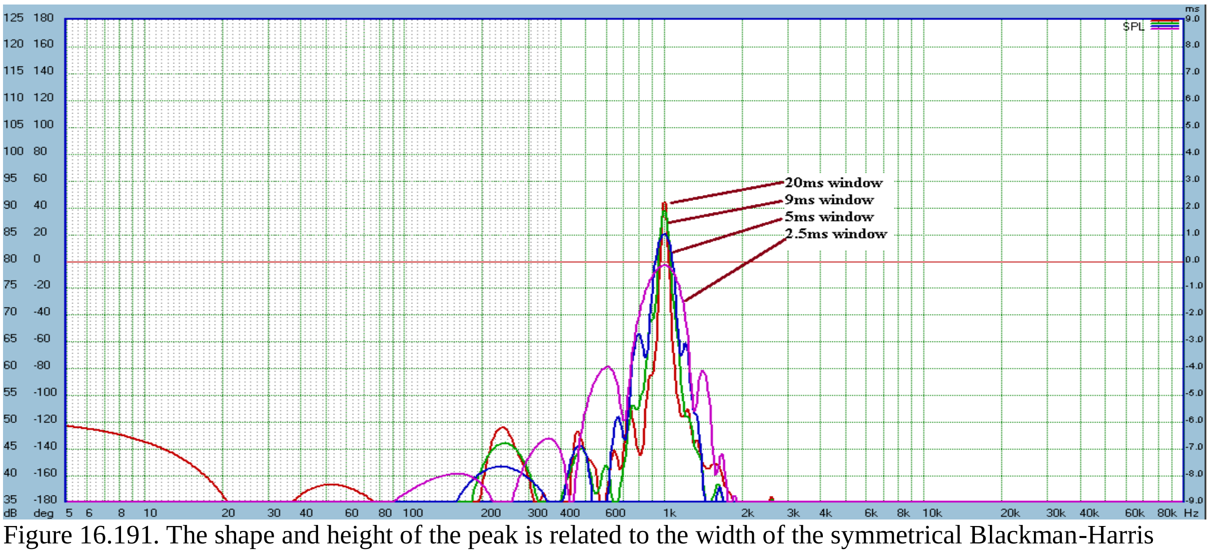


Figure 16.191. The shape and height of the peak is related to the width of the symmetrical Blackman-Harris window.

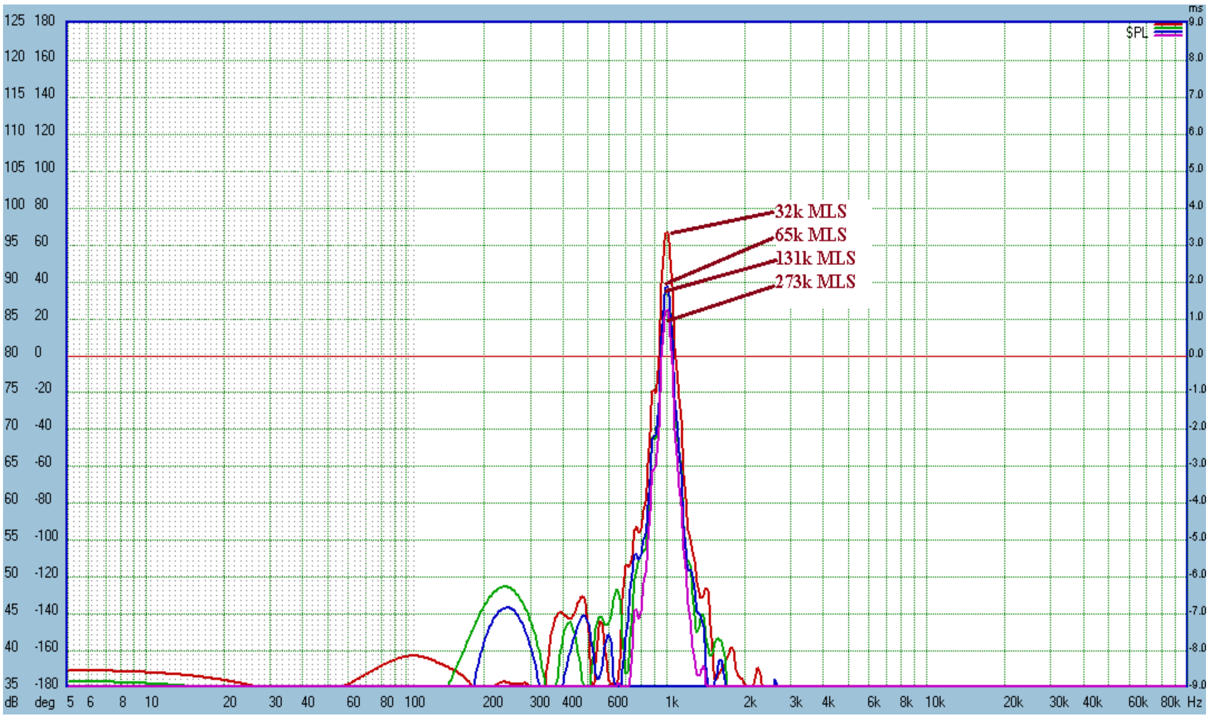


Figure 16.192. The shape and height of the peak is related to the width of the length of MLS. Recommended length is 65k or 131k. Window is 9ms Blackman-Harris window.

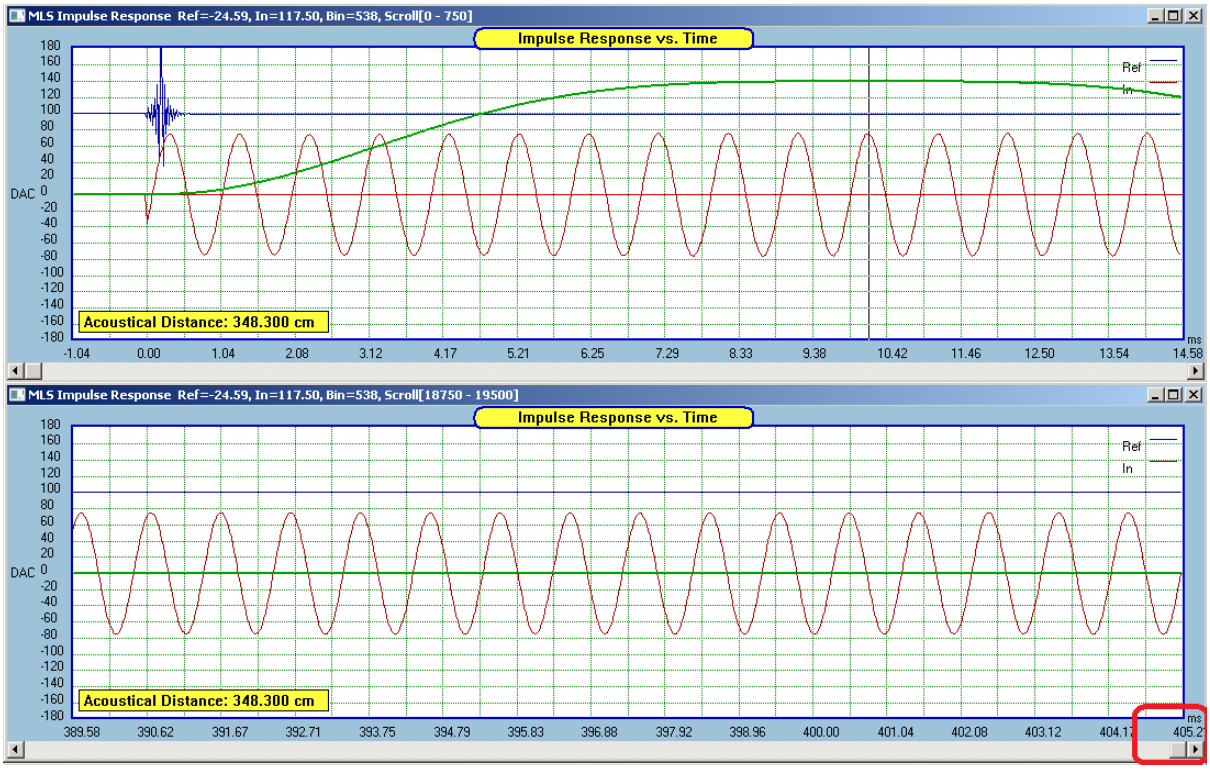


Figure 16.193. Captured sine-wave may fluctuate in amplitude after passing through Hadamard-Walsh Transform. Scroll through the captured Signal buffer to make sure, that the largest amplitude of the sine-wave is being windowed.

Fast Hadamard Transform

The Fast Hadamard Transform (FHT) is used in MLS system to reorder received MLS sequence back into the impulse response of the DUT.

However, the reordering process, when applied to the received sine-wave was causing unwanted amplitude fluctuations. It was then decided to replace the FHT by simple vector scaling process when sine-wave was fed into the MLS system.

As a result, a checkbox, “Calibration” was introduced in MLS system which allows the user to switch the MLS system into sine-wave calibration mode.

Proposed Calibration Sequence

Investigation into MLS system calibration conducted so far indicated, that regardless of the input signal, be it MLS sequence or sine-wave, the MLS measurement system can be modified to maintain it’s linearity. This is very important, as it allows us to propose a calibration method using an inexpensive, 1000Hz / 94dB microphone calibrator:

http://www.bksv.com/products/transducers/acoustic/calibrators/4231

Figure 16.194. 1000Hz / 94dB microphone calibrator

The calibration process described below is based on capabilities of the available hardware as follows:

1. Maximum undistorted input signal = 10Vpp (for Delta1010LT)
2. MLS crest factor = 6dB (for Delta1010LT)
3. Sinewave crest factor = 3dB
4. Calibration signal from microphone = 1000Hz at 1.0volt. This includes any microphone amplifier.

Let’s consider the sine-wave first, with the voltages delivered by the Calibrator.

1 Vrms = 1.42Vp = 2,84Vpp - This voltage will be delivered to the MLS system when 94dB calibrator is used. The system should be able to handle +6dB level as well, so:

2Vrms = 2.84Vp = 5.68Vpp - This voltage will be delivered to the MLS system when 94dB + 6dB level arrives, so it’s well within the 10Vpp input capabilities of the Delta1010LT sound card.

However, we will be using MLS pulse train ***with 6dB crest factor***, so let's have a look at the MLS situation: 1Vrms = 2Vp = 4Vpp - This voltage will be delivered to the MLS system when 94dB SPL level is received. The system should be able to handle +6dB level as well, so:

2Vrms = 4Vp = 8.0Vpp - This voltage will be delivered to the MLS system when 94dB + 6dB level arrives, so it's well within the 10Vpp input capabilities of the Delta1010LT sound card.

First, please enter 0.00 value as the "Calibration" factor into the Preferences screen.

Step 1 – Feed the 1.0Vrms signal from the Calibrator, with MLS system in Calibration Mode.

MLS Settings for sine-wave input – please note the "Calibration" checkbox is ON.

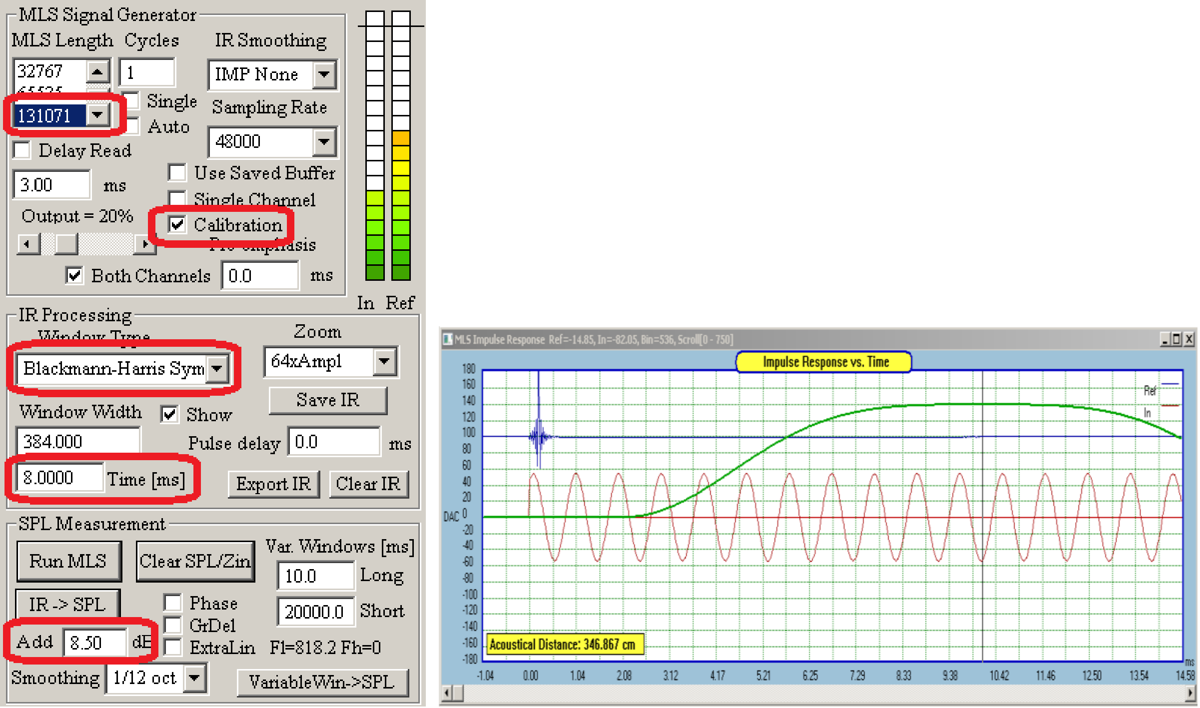


Figure 16.195. Major settings for sine-wave input – note the "Calibration" check box.

When the 1V rms sine wave from the calibrator is fed into the MLS system in "Calibration" mode, an 8.5dB is added to the SPL plot to bring it to the required 94dB level – see Figure 15 below.

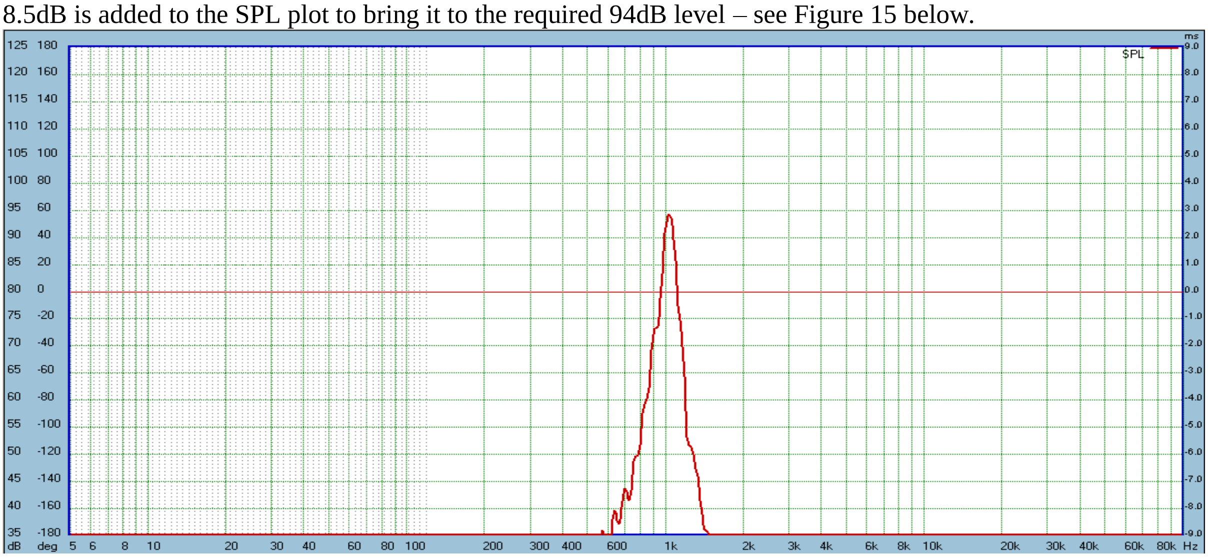
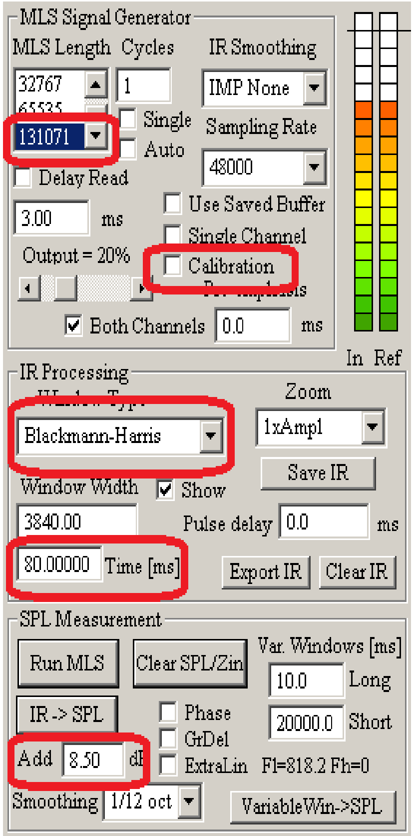


Figure 16.196. Modified MLS system response to sine-wave input.

Step 2 – Feed the 1.0Vrms signal from the Delta1010LT MLS system.

When the 1Vrms MLS signal from the Delta1010LT output is fed into the MLS system, the same 8.5dB is added to the SPL plot to bring it to the required 94dB level – see Figure 16.178 below.



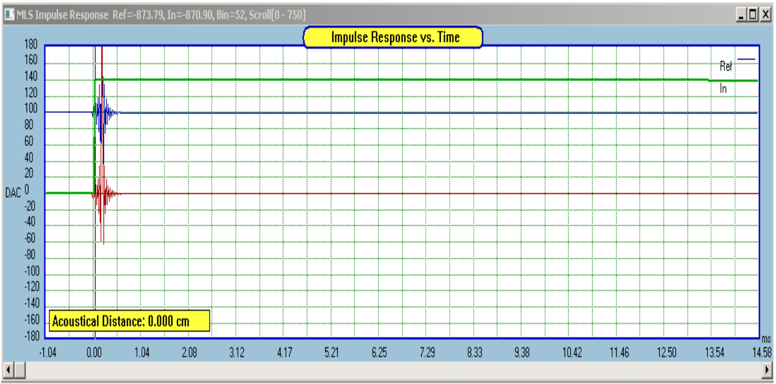


Figure 16.197 Major settings for MLS input – note the “Calibration” check box NOT checked.

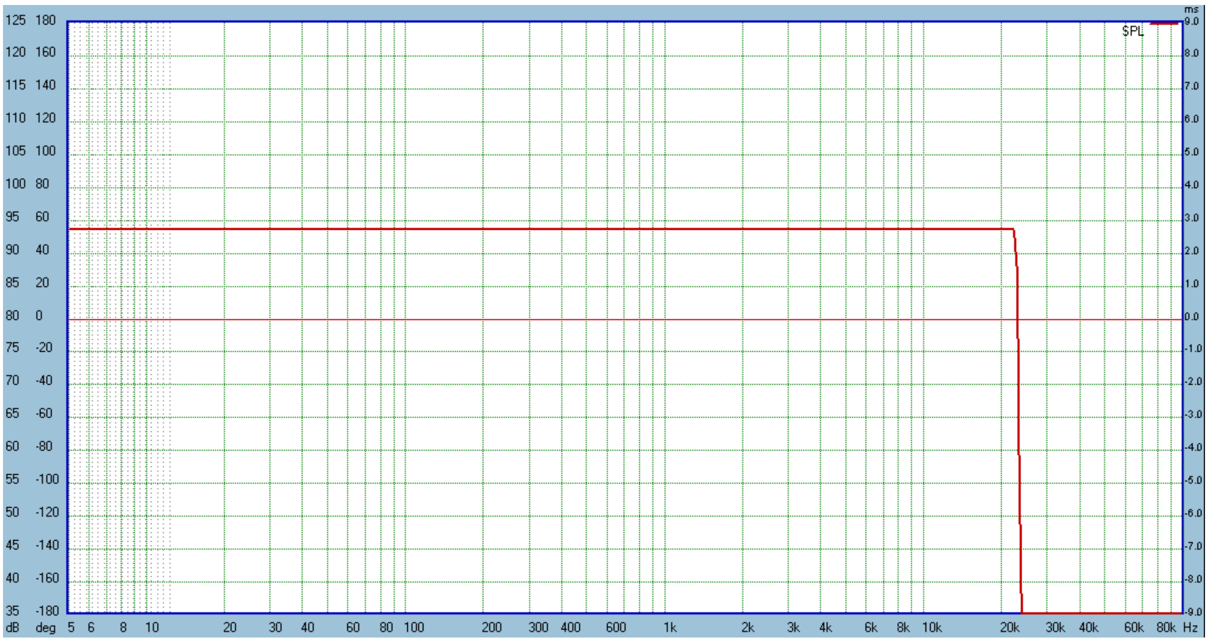


Figure 16.198. MLS system response to MLS signal input.

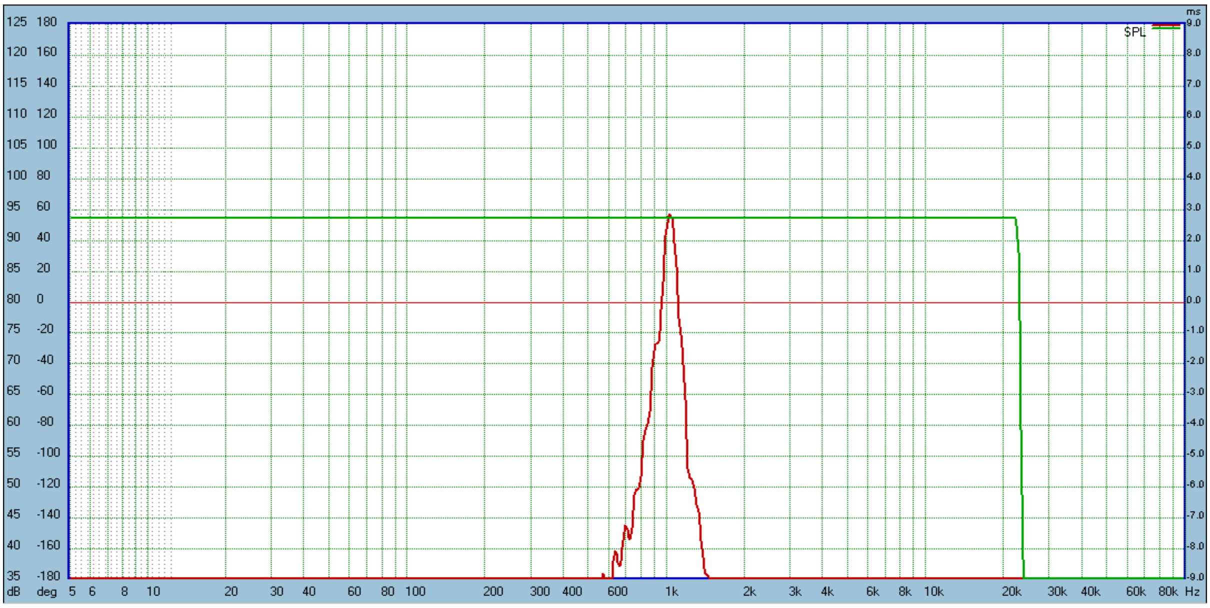
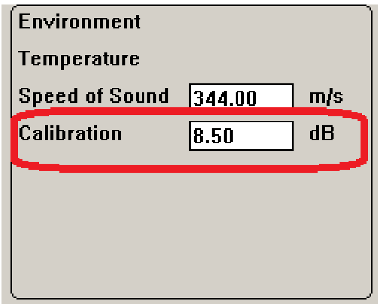


Figure 16.199. Both, MLS and sine-wave signals overlapped on the same plot.

As it can be observed from the Figure 16.180, the SPL plots are quite close to each other. It is estimated from the plots, that the accuracy of the sine results tracking the MLS results in the process described above is about +/- 1dB.

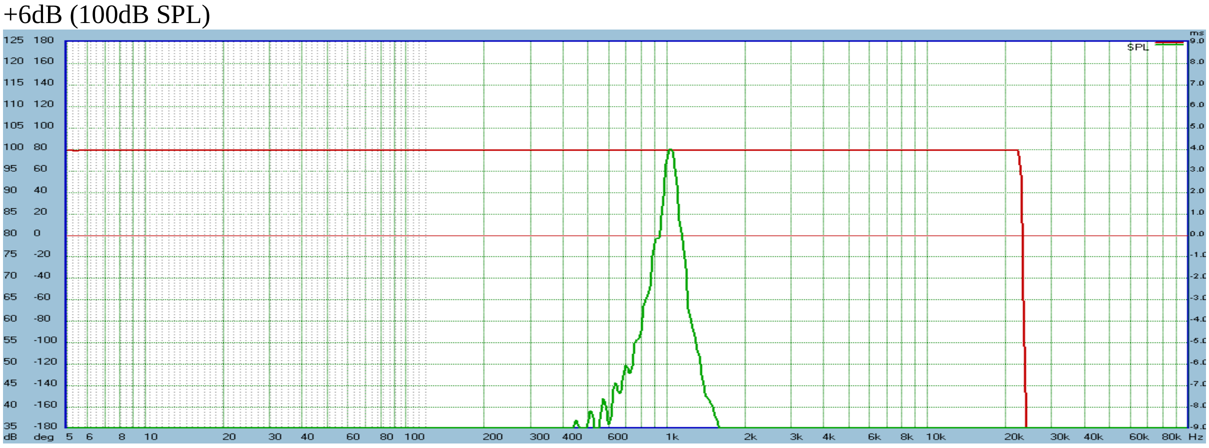


Finally, please transfer the added decibel value as the "Calibration" factor into the Preferences screen. This value is now saved into the Preferences file, and you are free to use the "Add" data field in the MLS TAB as usual – see above.

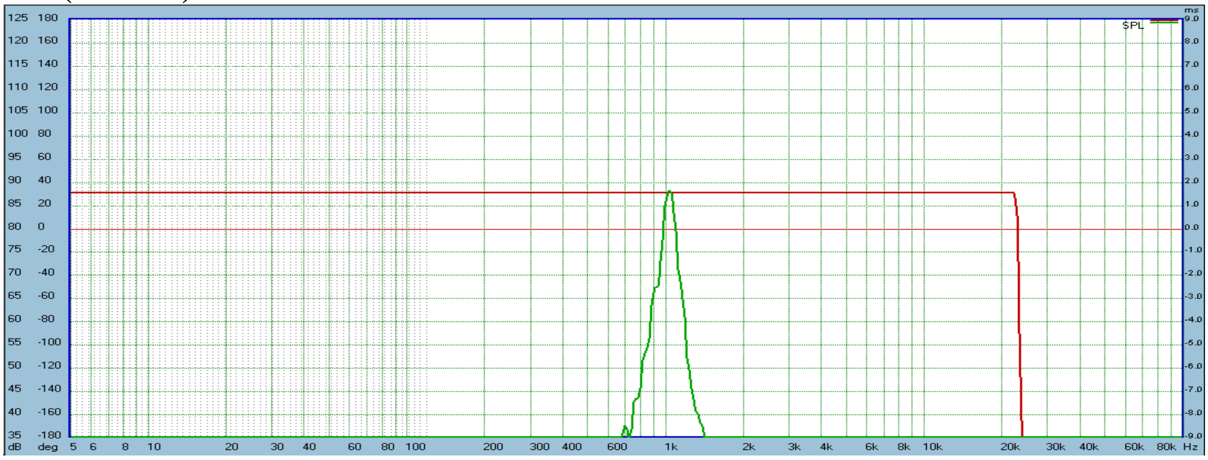
Linearity check

Final check for tracking accuracy between MLS and sine-wave inputs are shown below. Both signals were increased by +6dB, then decreased by -6dB and finally by -20dB. As observable from the figures below, excellent tracking has been achieved. Each time the system uses sine-waves, it must be switched to "Calibration" Mode.

+6dB (100dB SPL)



-6dB (88dB SPL)



-20dB (74dB SPL)

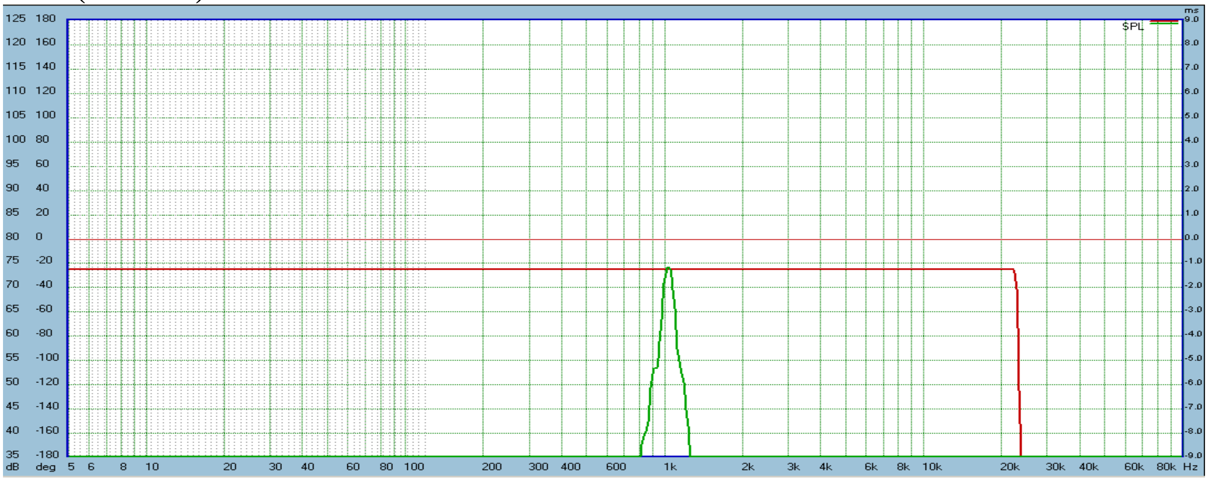


Figure 16.200. +6dB, -6dB and -20dB linearity check.

Step 3 – Set the MLS signal at 1Watt and feed it to the loudspeaker

Built-in “Reflection-Free Path” calculator may help you with locating the first surface reflection, and that should assist you in selecting FFT window length for supressing the reflections showing on impulse response.

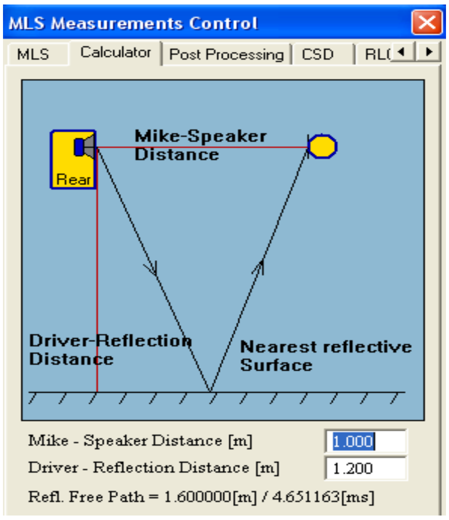


Figure 16.201. Calculator assisting with estimating FFT windowing.

RMS value of an AC current can be measured using a very rudimentary multimeter. I had a good success even with an old, analogue multimeter such as Micronta 22-204C

Figure 16.202. Multimeter used to measure 2.83Vrms for 1W/1m at 8ohm measurements

I set it to 5VAC range. It shows a very good agreement for RMS values with my digital CRO. The 1W RMS at 8ohm is 2.83V. So for instance, while my CRO is showing 2.85Vrms, the Micronta shows almost 2.9Vrms. The 2.9Vrms at 8ohms = 1.05Watt, so it's close enough for me.

Complex Smoothing in MLS System

Traditional smoothing applied to SPL and Phase curves involves “fractional octave smoothing”, like 1/12oct or 1/6oct and so on. The smoothing is performed in logarithmic frequency scale, and is easily applied to SPL curve and significantly more difficult to phase response curve.

However, now the **SPL/Phase data never makes it to the logarithmic scale for processing**, therefore the smoothing needs to be applied differently. The example below shows an impulse response of a tweeter driver.

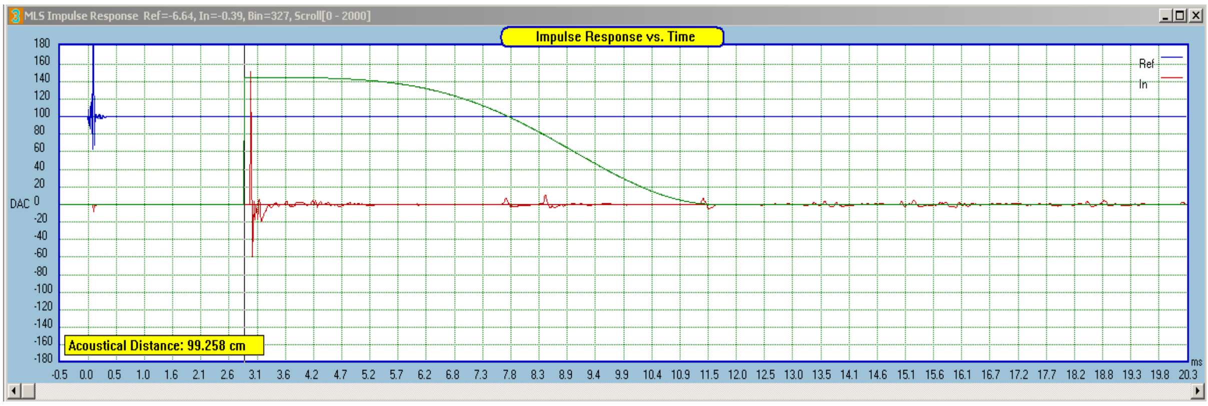


Figure 16.203. Impulse response of a tweeter driver

In the next step, FFT is applied to ultimately obtain SPL and phase responses. However, the immediate result of the FFT is the driver’s complex Transfer Function, expressed as TF(jw) = {Re(w)} + j{Im(w)}. The real and imaginary parts do not even resemble the final SPL and phase curves yet, but are quite good candidates for smoothing.

Such operation is depicted on the figure below, where all relevant variables are plotted in linear frequency scale. The smoothing algorithm must take into account, that at low frequencies, data is very sparse, but at high frequencies, there is almost too much data. So the frequency range over which the smoothing is calculated, has to be progressively expanded, as the frequency increases.

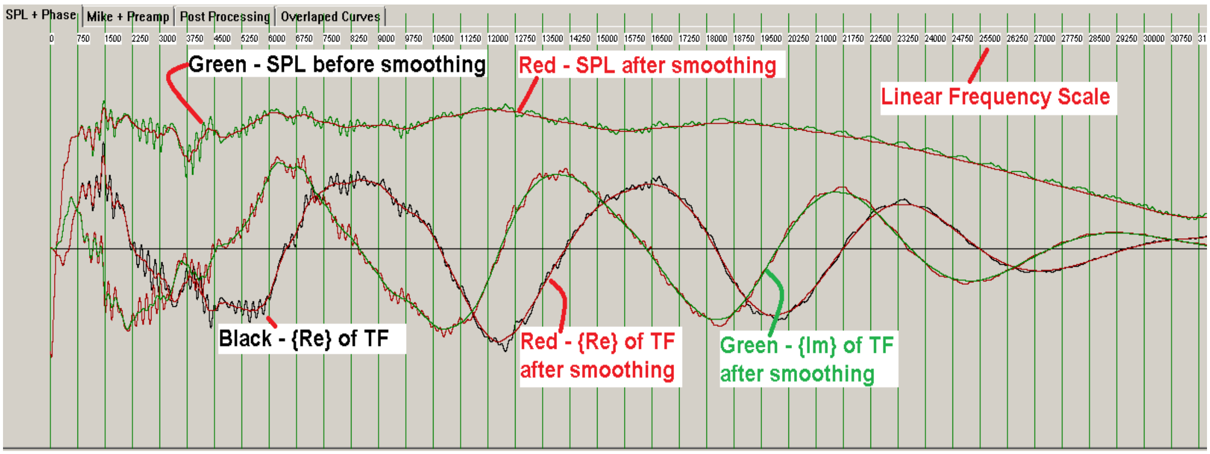


Figure 16.204. Complex Smoothing explained.

The degree of smoothing is no longer expressed in “dB/oct”, but it offered as “Level_1” to “Level_8” smoothing.

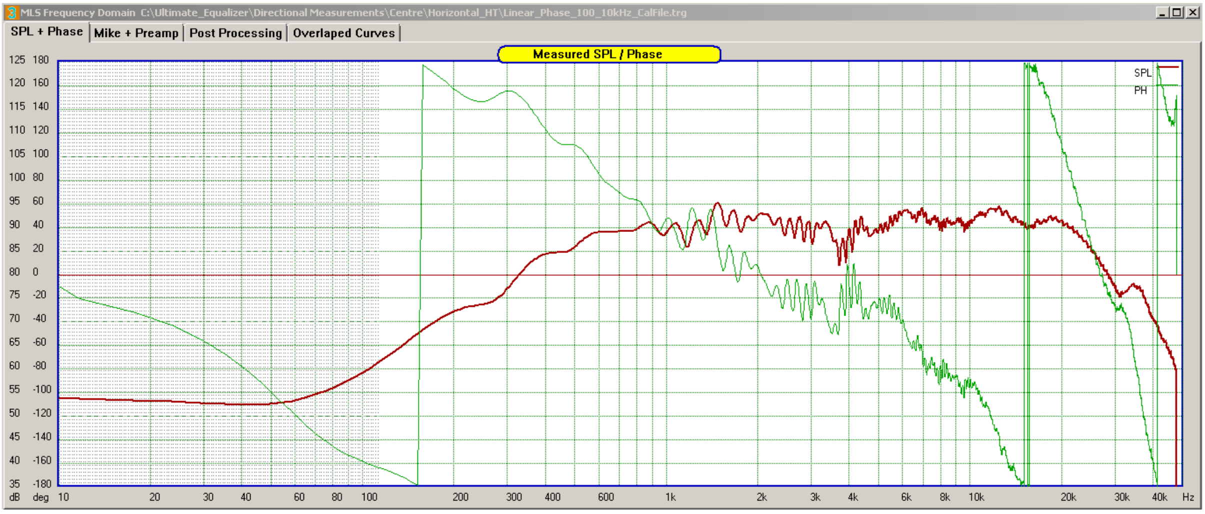


Figure 16.204. Final SPL/Phase with no smoothing

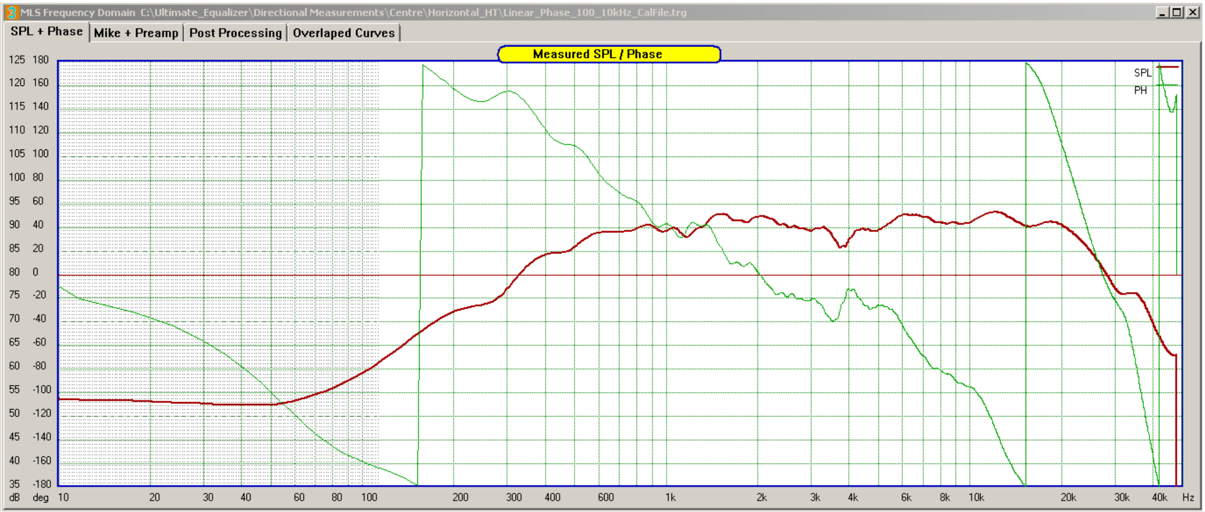


Figure 16.205 Final SPL with “Level 8” smoothing

As observable on the figures above, Complex Smoothing in linear frequency scale is very effective in obtaining smooth SPL and phase curves. It is also very effective in removing rapid phase fluctuations (+/-180deg), resulting in well-defined phase transitions, greatly assisting in extracting the minimum-phase phase response. Combination of FFT windowing and Complex Smoothing provides excellent results for SPL and phase smoothing.

It is recommended to apply FFT windowing to remove most of the time-of-flight from the measurements before applying Complex Smoothing. This can be easily done by correctly placing the FFT window.

Direct Measurements Compared With CAD simulation

At some point during the design process, you may wish to compare that last built prototype system with the current CAD simulations.

This can be easily accomplished by selecting “**Save As System**” checkbox option in “MLS Measurements Control” dialogue box. This needs to be performed before you start making measurements – see Figure 16.188.

You can now proceed and make SPL measurements of your system. The first SPL, Phase and Group Delay measurements will be saved into the project file data space. Then, you can proceed to measure input impedance (Zin) and impedance phase (ZinPhase) of the system, and these measurements will also be saved into the project file data space.

These measurements can be re-called into CAD and System design screens to compare the measurements results with the current CAD simulations – see Chapter 8, page 8.48..

Data buffer holding the 5 curves: SPL / Phase / Group Delay / Zin / Zin Phase cane be cleared by pressing "**Clear SPL/Zin**" button in MLS system – see Figure 16.188 below.

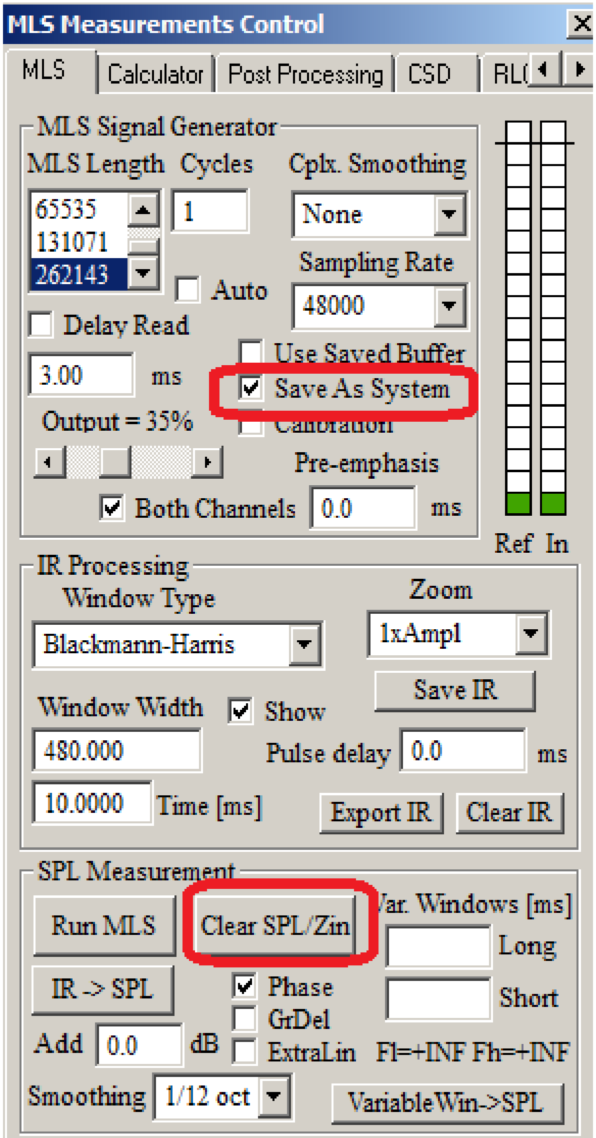


Figure 16.206 – Selection in MLS System

Non-Coherence Distortion

Before we introduce the Non-Coherence Distortion concept, it may be beneficial to highlight the essence of the traditional, Total Harmonic Distortion (THD) measurement. Let's consider the test circuit shown on Figure 1 below.

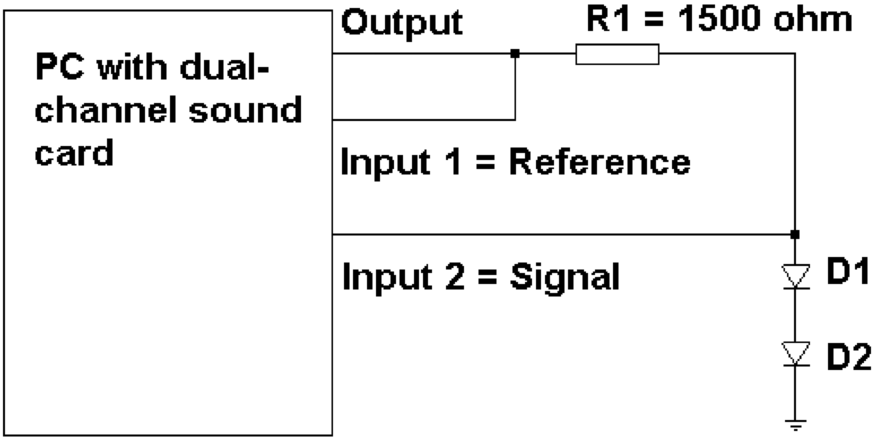


Figure 16.184. Distortion test circuit.

Sine wave generated by the Signal Generator (PC with dual-channel sound card) is applied to a diode clipping circuit, which clips the positive part of the sine wave, giving it the shape as shown on Figure 16.185. The positive voltage is only allowed to go up to 2 x 0.7V = 1.4V, as the result of the diodes' clipping action. Please note, that only the "Signal" input is necessary to calculate THD figure.

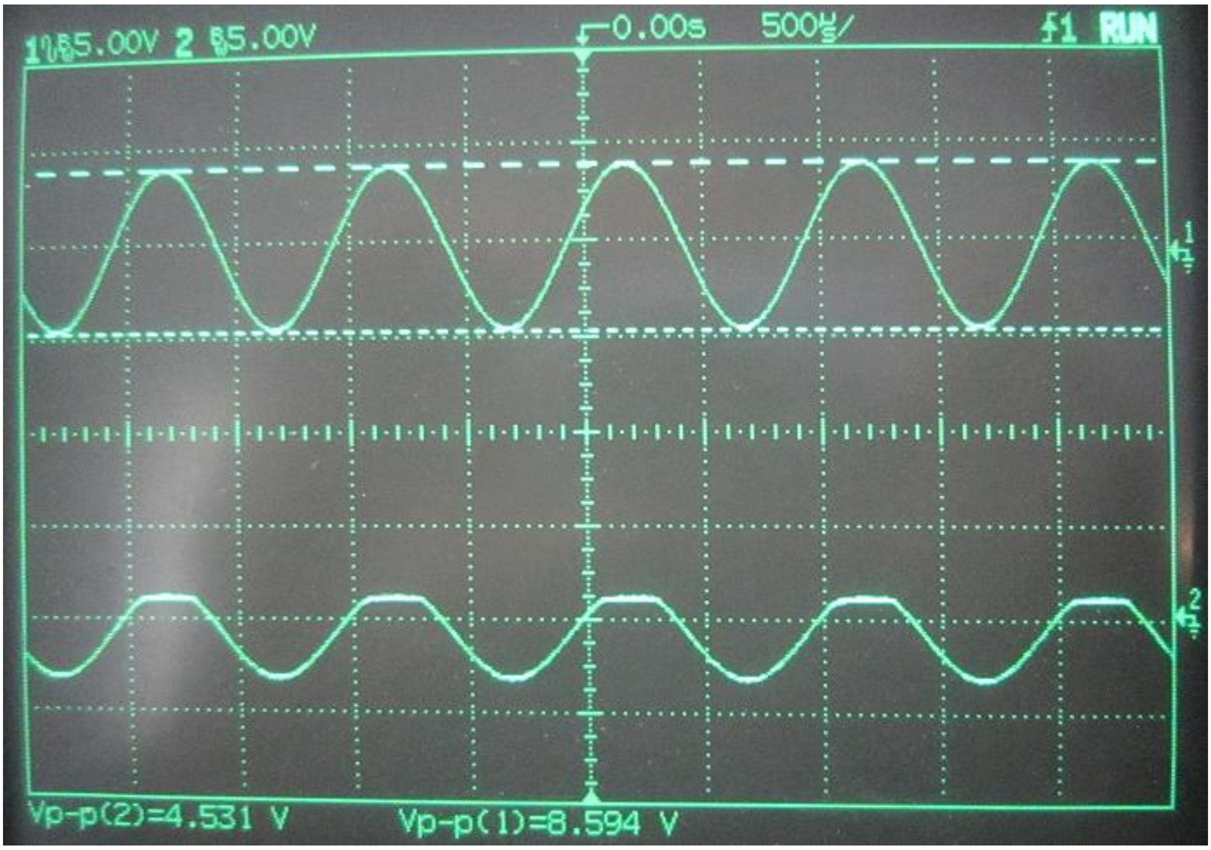


Figure 16.185. Reference = top trace, Signal = bottom trace.

When a standard THD test is applied to such signal, the result is 7.95% distortion – see Figure 16.186 below. It must be emphasized, that test signal is a steady-state in nature. It does not change in time, and the harmonics are easily distinguishable from each other up to the order of 9-10th. Clearly, the THD test is matched very well with the test signal.

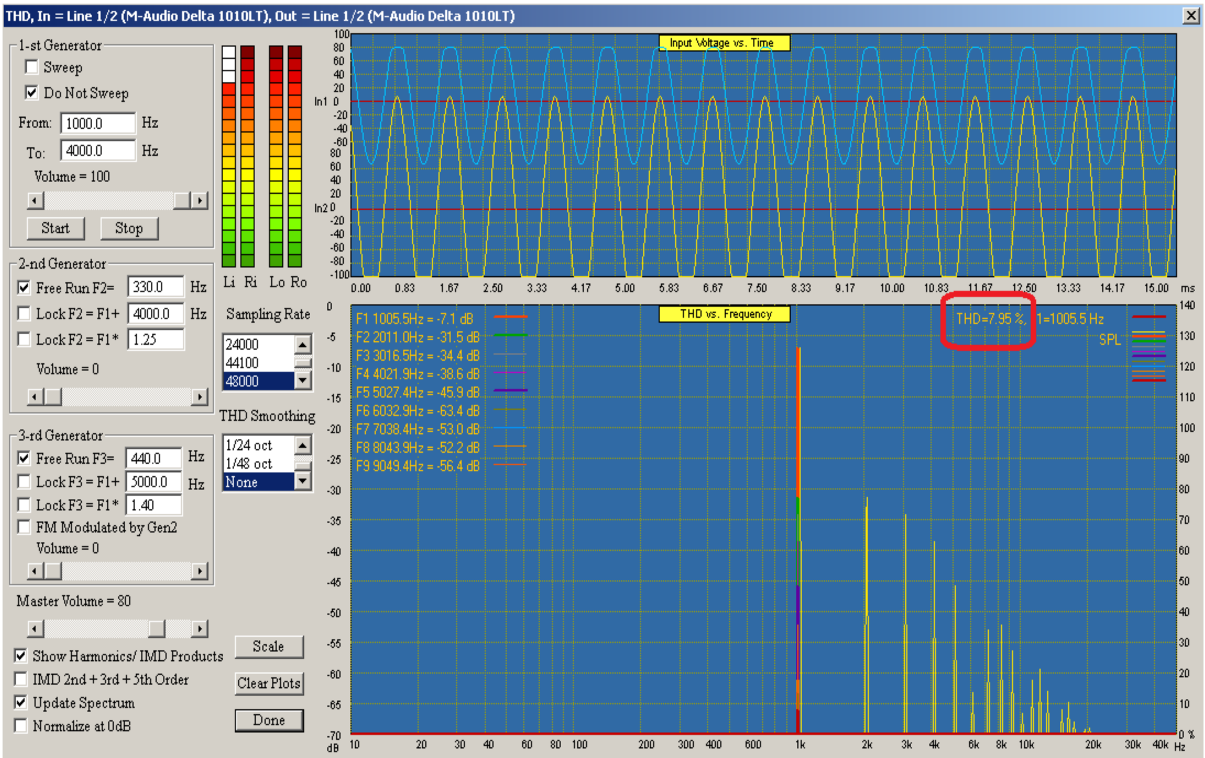


Figure 16.186. THD test reveals 7.95% harmonic distortion.

The THD is a single-channel test method, as all information necessary to extract the test result is embedded in the distorted channel (Input 2). It is an excellent laboratory test, that helps to uncover and subsequently correct undesirable signal distortions within the amplifier or loudspeaker driver unit.

With the above in mind, we can now introduce the **Non-Coherence Distortion method**. It is a **dual-channel method of analysis** of non-coherence between the reference signal (stimulus) and the response of the system under test. Interestingly, the NCD method, provides continuous distortion curve versus frequency and can provide single number, as the Total Non-Coherent Distortion figure. Most importantly, the preferred stimulus signal can be noise, music or speech. This is where the NCD method is most useful.

Non Coherence is defined as: Non-Coherence = $1 - \gamma(w)^2$

Where coherence:

$$\gamma(w)^2 = \frac{|Gxy(w)|^2}{Gxx(w) * Gyy(w)}$$

And:

$$Gxx(w) = \frac{1}{M}\sum_{i=0}^{M-1} |X(t,w)|^2$$

$$Gyy(w) = \frac{1}{M}\sum_{i=0}^{M-1} |Y(t,w)|^2$$

$$Gxy(w) = \frac{1}{M}\sum_{i=0}^{M-1} \overline{X(t,w)} * Y(t,w)$$

In the notation above, the Gxx is the input Auto-Spectrum, the Gyy is the output Auto-Spectrum and Gxy is the input-output Cross-Spectrum.

Loop test using white noise on both inputs connected together, indicates low **NCD = 0.034%** across 40Hz-20000Hz, as shown on the Figure 16.187 below.

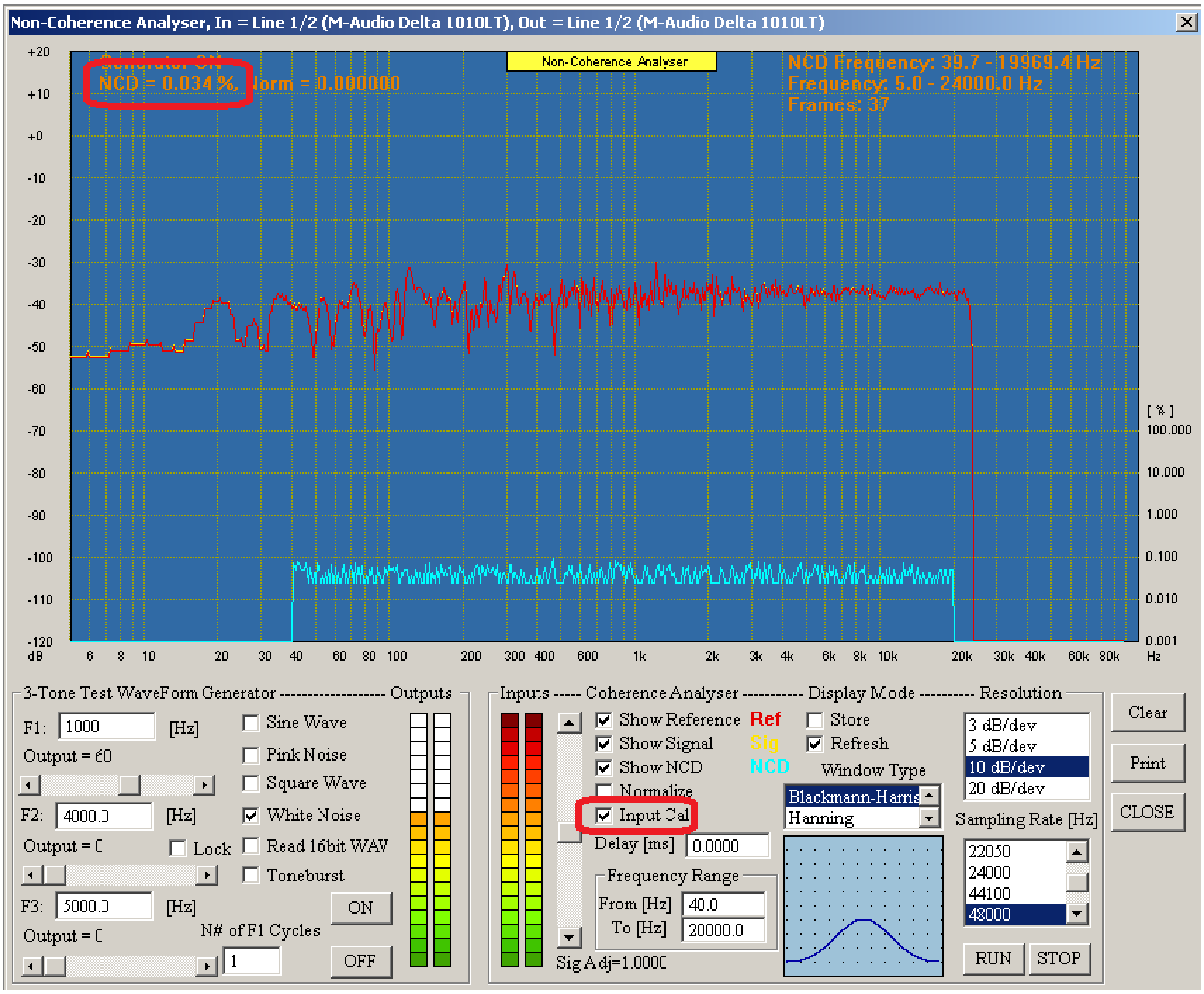


Figure 16.187. NCD loop test

This would be the accuracy limit in this particular testing setup.

Calibration

It may be beneficial to perform initial calibration of the NCD system by accounting of the small frequency response differences between two sound card inputs. This accomplished by running loop test with both inputs tied together, while using MLS or ESS systems. On the MLS or ESS tabs, press “IN1-In2” button after the loop test has been performed.

The process if fully automatic and will result in small calibration file “NCD_Calibration.ncd” stored on your HD. This file will be used anytime during NCD measurements when the “Input Cal” checkbox is ON – see Figure 16.187 above.

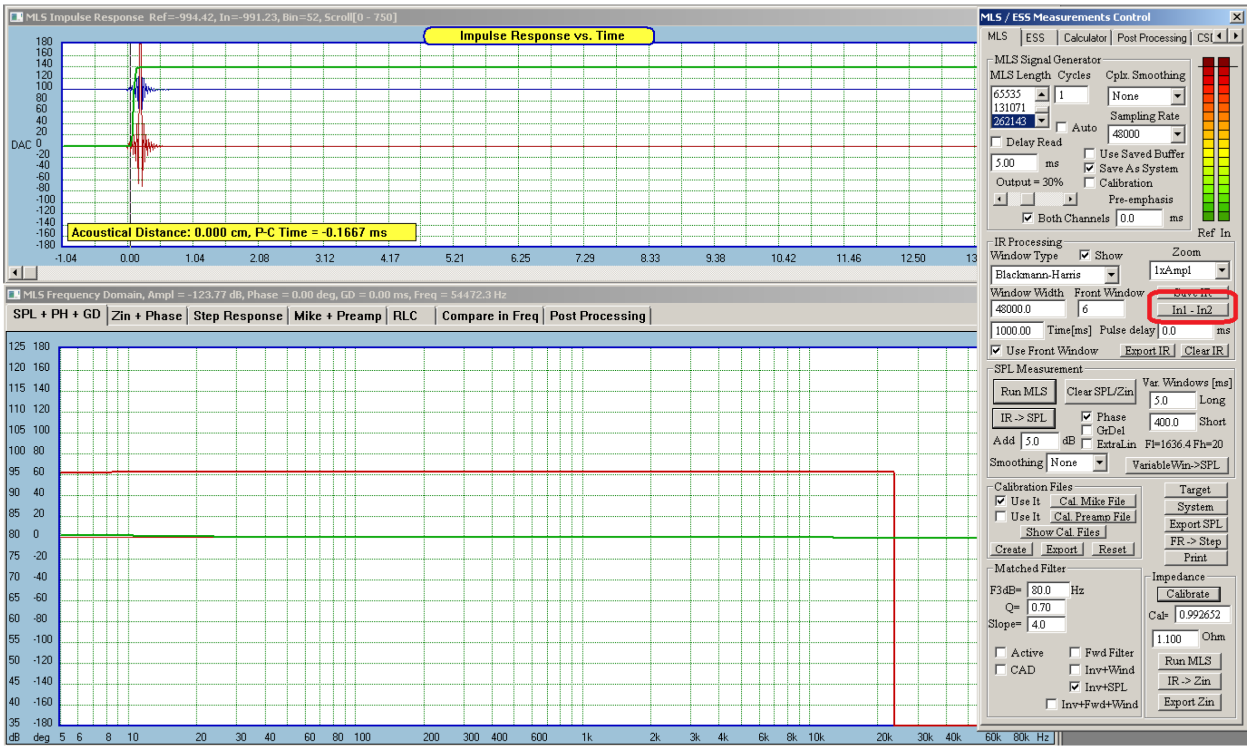


Figure 16.188. Generating Input Calibration File.

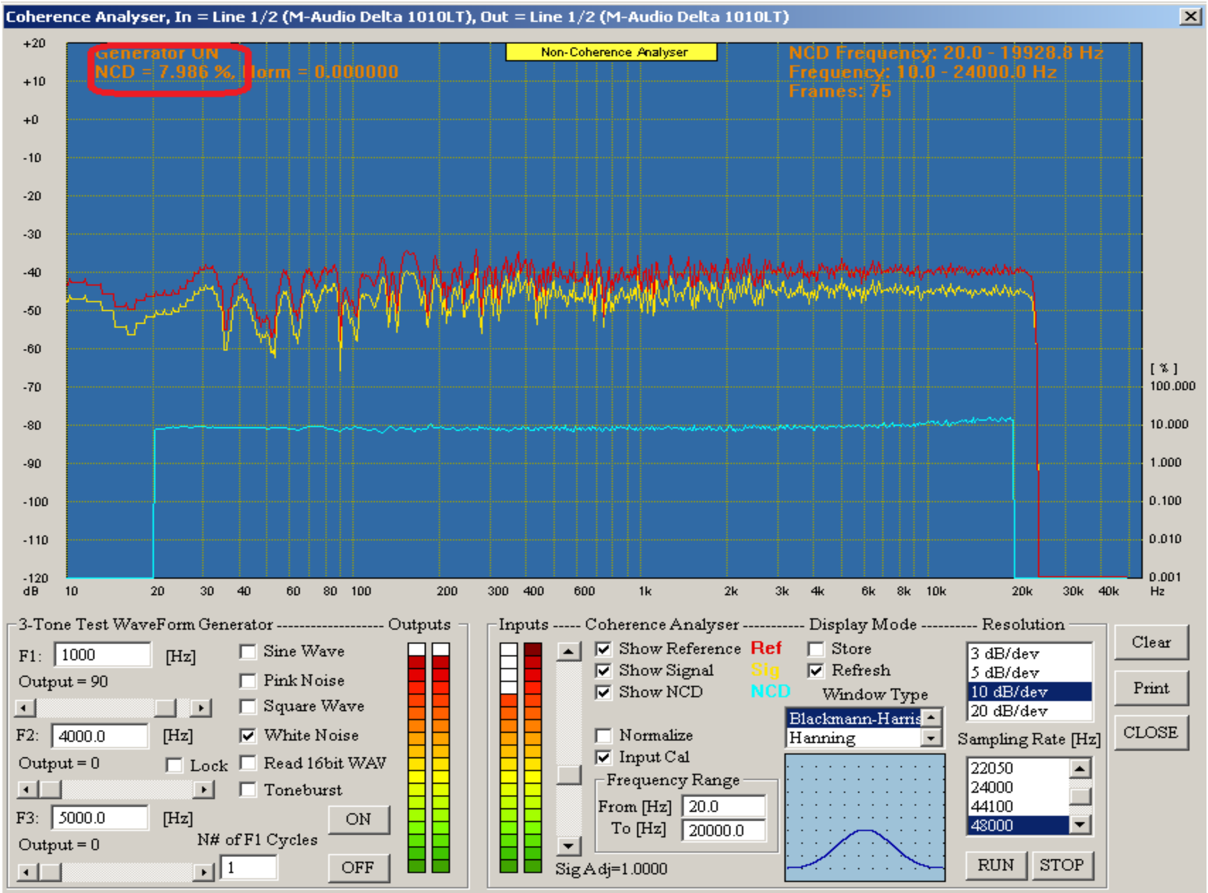


Figure 16.189. NCD diode clipping test reveals 7.986% NCD distortion

NCD formulas give above, clearly indicate, that NCD is not the same as THD. The former method measures non-coherence between stimulus and response, rather than level of harmonics in the output signal, normalized to the fundamental frequency.

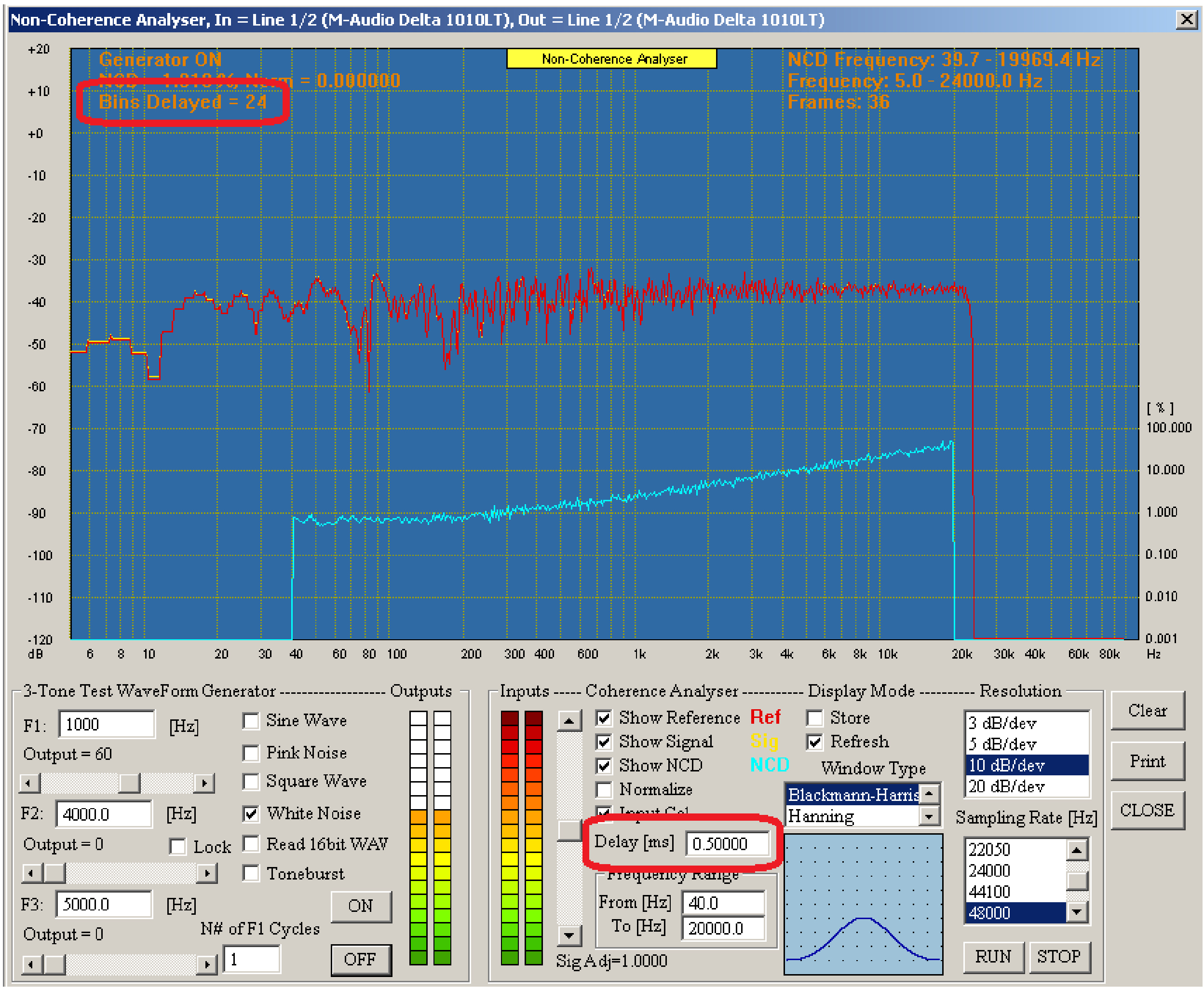


Figure 16.190. Example of time-domain errors.

The NCD is sensitive to time-based errors. Therefore, when evaluating loudspeakers, the time-of-flight must be accounted for. You can enter the delay due to time-of-flight in the “**Delay[ms]**” entry field. An example of time domain error is shown on Figure 16.190 above, where the loop-test was affected by adding 500usc delay into the Signal Channel. It is observable, that NCD errors increased with frequency, as the correlation between the non-delayed and delayed paths is steadily decreasing.

Loudspeaker Tests

THD = .63%

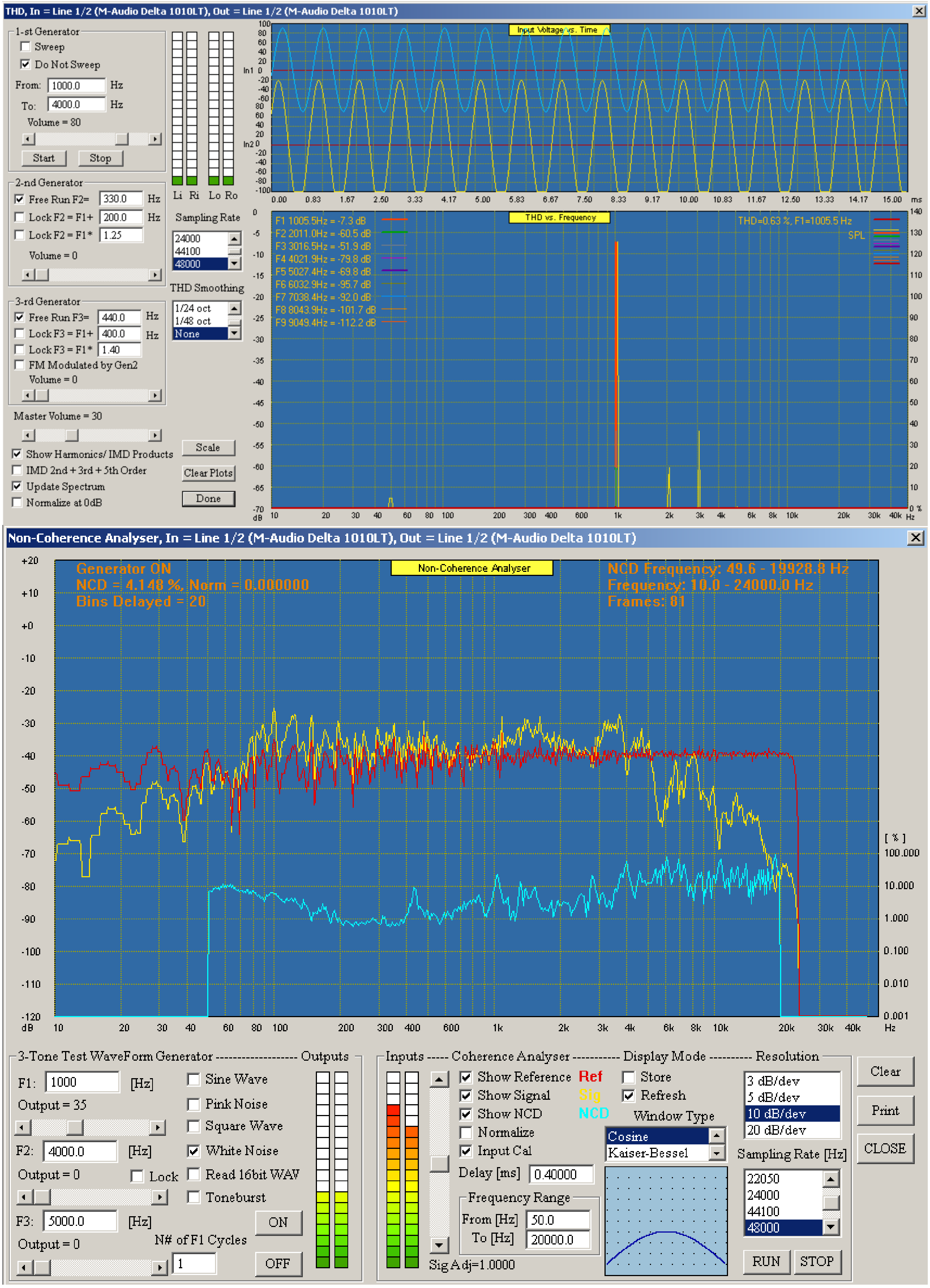


Figure 16.191. NCD Loudspeaker tests.

Minimum-Phase Development

Correct determination of the minimum-phase phase response of a loudspeaker driver has been often viewed as nearly impossible task. The developed method described below takes one step further in providing a solution to this dilemma. The method was developed to exemplify the following logic:

Loudspeakers are known to be minimum-phase devices within their operating frequency range. Therefore, their SPL and Phase response are mathematically locked via Hilbert-Bode Transform (HBT). This characteristics allow us to construct a simple band-pass model of the loudspeaker, using measured SPL as a template, from which we can instantly obtain minimum-phase phase response of the model. Since we are mainly interested in the asymptotic slopes of the SPL response, the actual model does not have to be very complex. Small SPL variations will only cause local phase variations, but will not affect asymptotic slopes. After optimization, the band-pass model provides a template for the HBT parameters, such as the asymptotic slopes of the SPL and at the same time guides the location of the FFT window for converting measured SPL into minimum-phase SPL/Phase of the measured driver.

The optimization process mentioned above is facilitated by employing a curve-fitting algorithm to optimize the fitting accuracy of the band-pass model to the measured SPL. The algorithm manipulates several parameters of the band-pass model to **minimize the error function objectively and therefore is independent of the operator's judgement or speculations**. If correctly executed, the band-pass model is the best fit to the measured SPL, thus providing the most accurate minimum-phase phase response possible. From here, the acoustic centre, "time-of-flight" and location of FFT window can be determined. Several examples of various measured loudspeakers have been provided and minimum-phase phase response has been developed.

Woofer Example

A 12" loudspeaker was placed in a vented box. Measured frequency response of the driver in a vented box with QB3 tuning is shown on Figure 16.192 below.

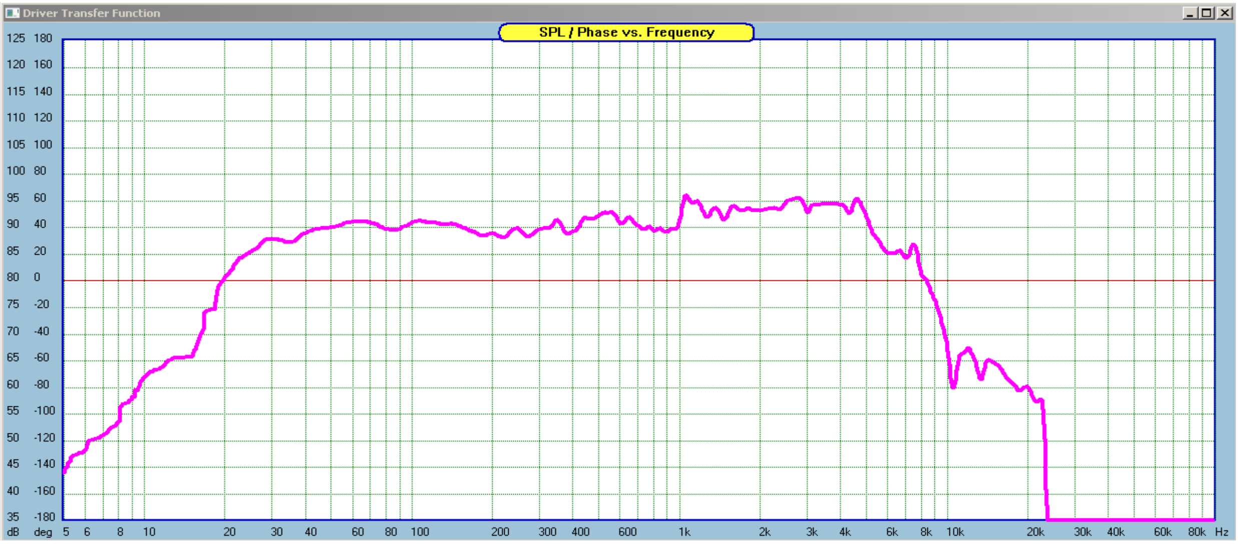
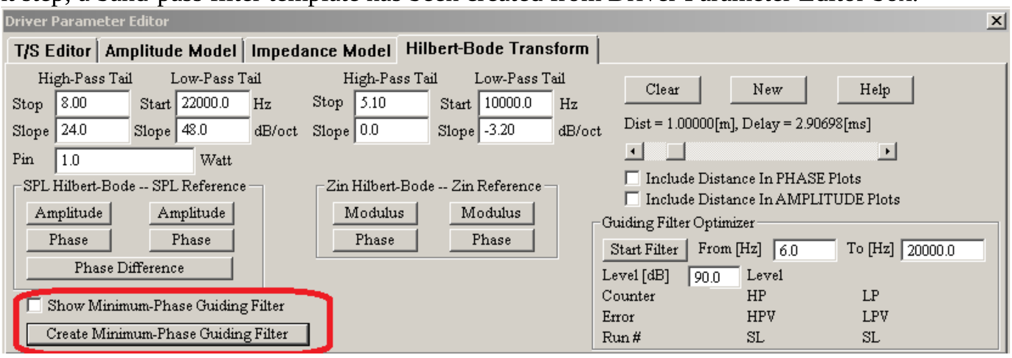


Figure 16.192. Measured frequency response of the driver.

In the next step, a band-pass filter template has been created from Driver Parameter Editor box.



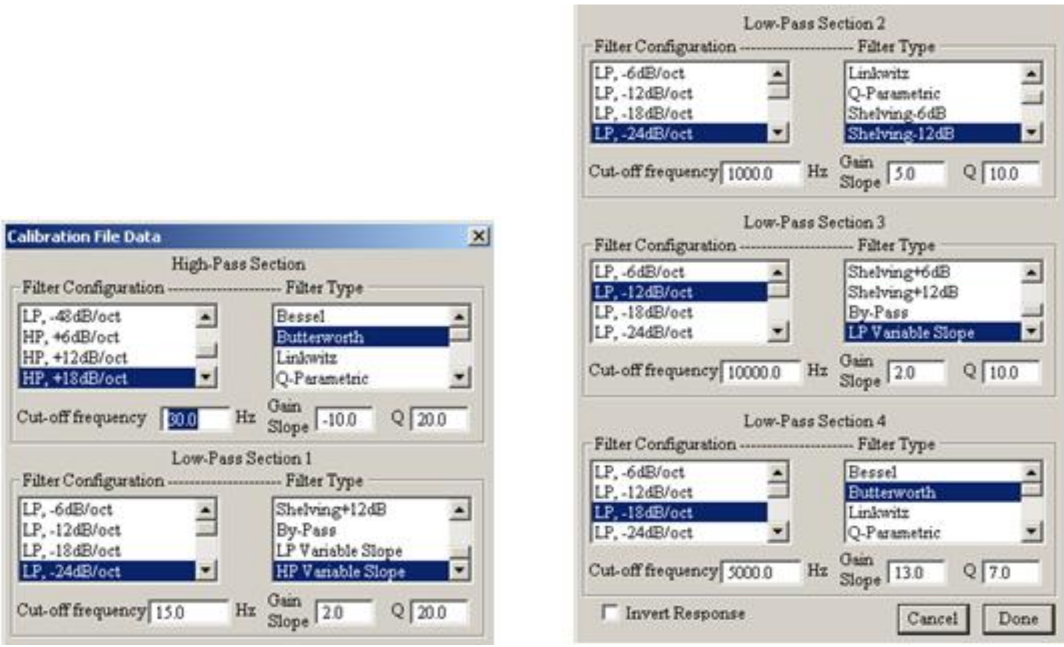


Figure 16.193. Initial parameters for band-pass filter template.

The first-cut band-pass filter will be used as the starting point for the optimizer. It is therefore sensible, to design such filter fairly close to the measured SPL curve, as this makes the optimization process easier to perform. The range of adjustments are shown below:

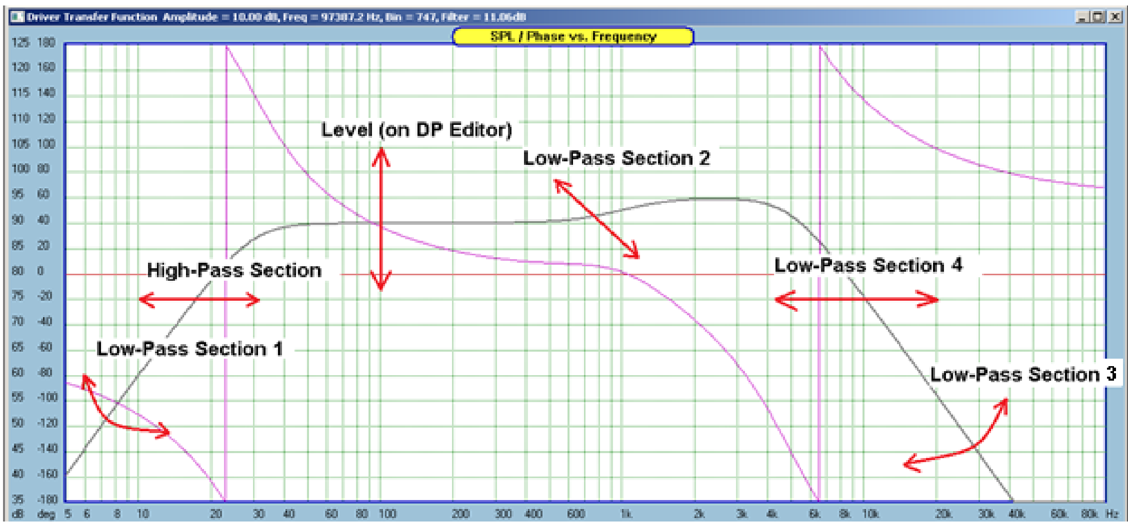


Figure 16.194. Resulting first-cut band-pass filter with range of optimizations.

1. High-Pass Section – select the initial slope and cut-off frequency of the high-pass filter approximation. Since the vented box is a QB3 alignment, then the high-pass slope was selected as +18dB/oct, with cut-off frequency of 30Hz.
2. Low-pass Section 1 – select HP Variable Slope filter located well below the cut-off frequency of the (1). This optimization will provide “fine trimming” of the HP slope – if necessary.
3. Low-Pass Section 2 – Examination of the measured SPL on Figure 6, indicates, that there is a step increase in SPL at around 1000Hz. This will be approximated by -12dB shelving filter at 1000Hz with 5dB gain and Q=10. This is only manual setting.
4. Low-pass Section 3 – Select LP variable Slope filter located well above the cut-off frequency of (4).). This optimization will provide “fine trimming” of the LP slope – if necessary.
5. Low-Pass Section 4 – Select the initial slope and cut-off frequency of the low-pass filter approximation. Visual examination of measured SPL reveals, that a -18dB/oct LP filter at 5000Hz could be a good starting slope.
6. Level – Enter 90dB as the starting value.

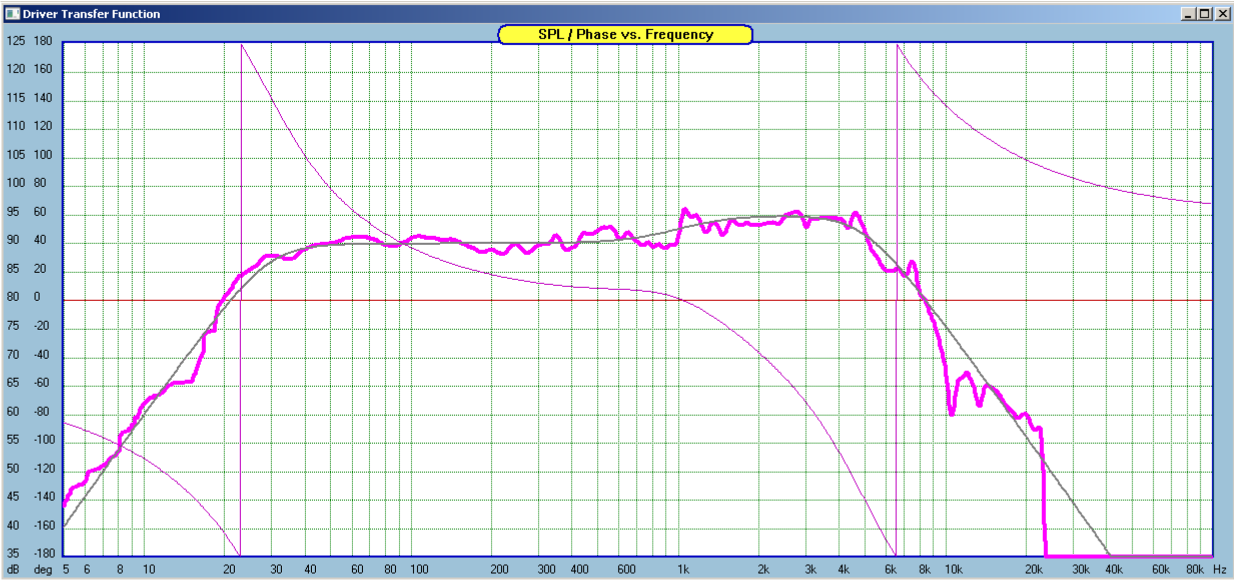


Figure 16.195. Initial matching of the filter with measured SPL.

In the next step, the frequency range of optimization is selected as: 6.0Hz – 20000Hz, and the Level = 90.0dB. Then, press the "**Start Filter**" button.

Optimizer takes two passes (Run=2). In the second pass it performed 66 optimizations. The resulting parameters are as follows:

1. High-Pass Section – cut-off frequency of 30Hz was optimized to 31.4Hz.
2. Low-pass Section 1 – select HP Variable Slope – not needed.
3. Low-Pass Section 3 – select LP variable Slope filter – only needed 0.43dB extra slope. Can be disregarded.
4. Low-Pass Section 4 – cut-off frequency was optimized to 4470Hz.
5. Level was optimized to 90.02dB.

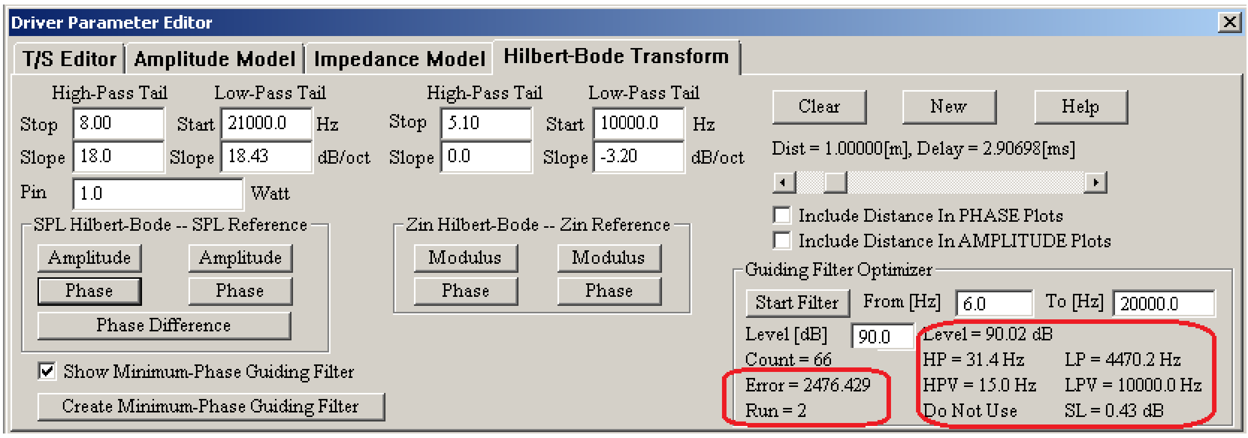


Figure 16.196. Optimization process completed.

Optimizer has shown, that the initial filter parameters were actually quite close to the final, optimized values. You'll get better at the initial estimations as you generate more minimum-phase curves.

We are now able to attach the high-pass and low-pass asymptotic slopes to the measured SPL.

The process is exemplified on Figure 16.197 below. The attachment points are:

For the high-pass slope: 18dB at 8Hz
For the low-pass slope: 18.43dB at 21000Hz.

We are now ready to calculate HBT, based on the available data.

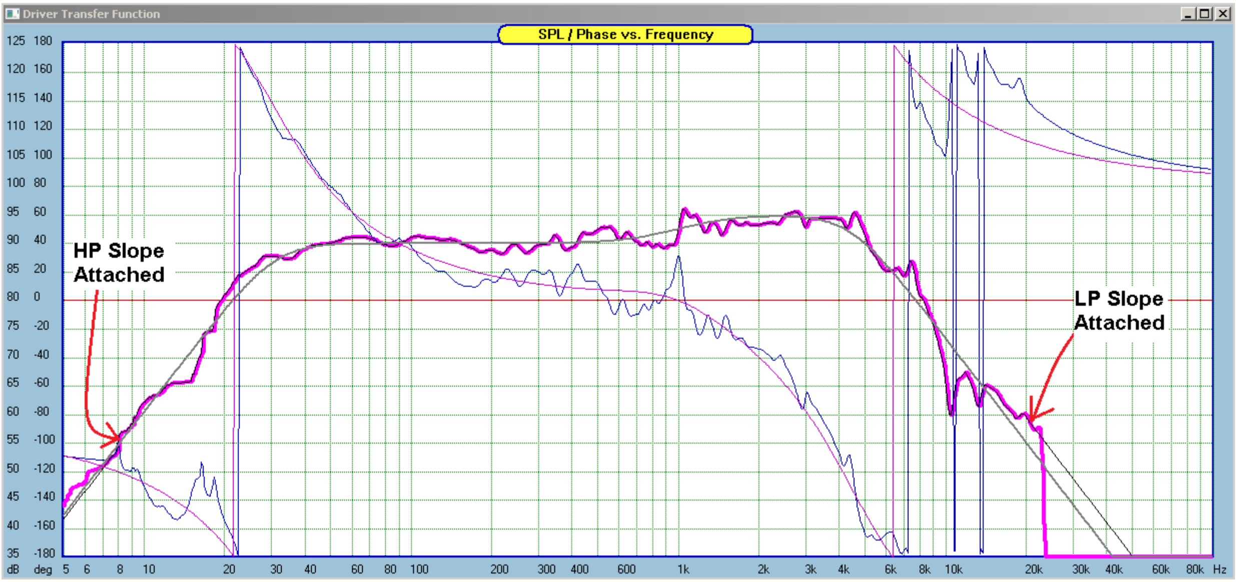


Figure 16.197. Calculated HBT – based on optimized band-pass filter.

Please note, that HBT phase (thin, pink curve) has early phase transition at 6kHz, due to neglecting of the small peak on the drivers measured SPL at 7.5kHz.

Fortunately, we can engage the “unused” Variable Slope filter to create a small, +5dB hump on the filter’s SPL at 7.5kHz. Replotting the filter, shows, that the first 180deg transition of the Guiding Filter and the HBT phase response overlap exactly.

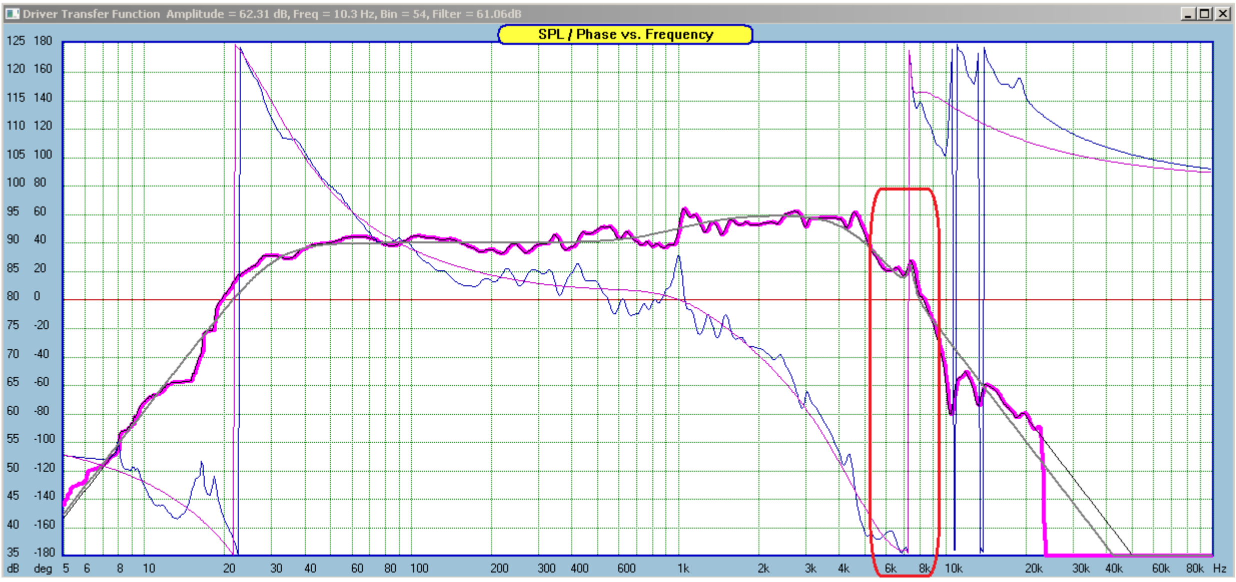


Figure 16.198. Calculated HBT – based on optimized band-pass filter with small correction at 7.5kHz .

If the initial slopes are estimated **incorrectly**, you may end up with the band-pass filter looking like the one on Figure 16.199 below. You can still run the optimization process, but the **Error value** will tell you, that initially selected filter parameters were way of mark.

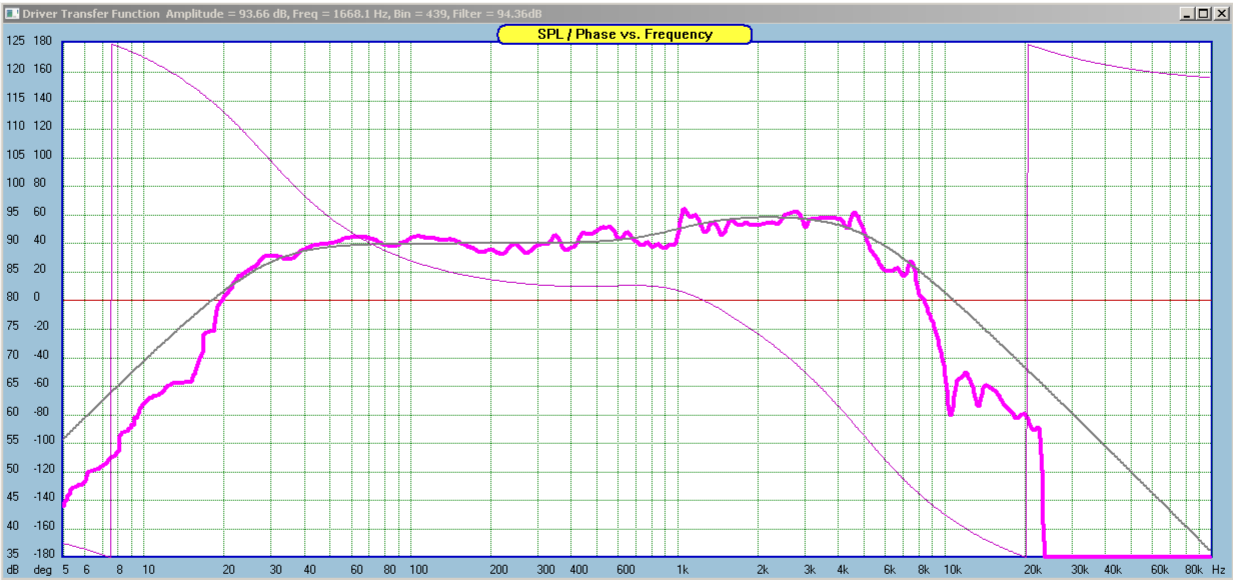


Figure 16.199. Incorrectly estimated slopes for the band-pass filter.

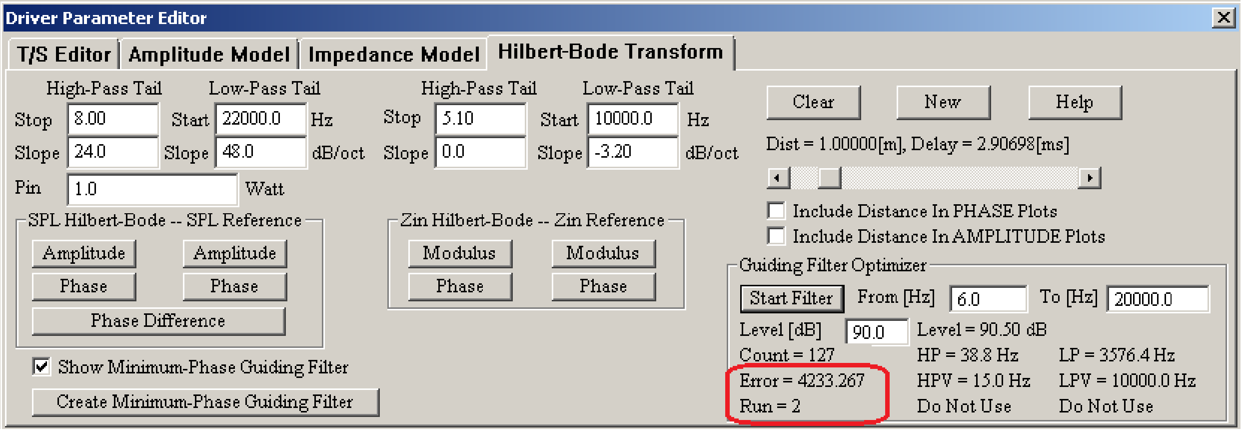


Figure 16.200. Optimization of incorrectly estimated slopes for the band-pass filter

The optimizer still attempts to minimize the error, but now the **Error is 4233** (previously 2476). And both Variable Slope Filters are not recommended. Smart move would be to revert to the +18dB/oct and -18dB/oct for the main slopes and 1dB/oct Variable Filters initial settings.

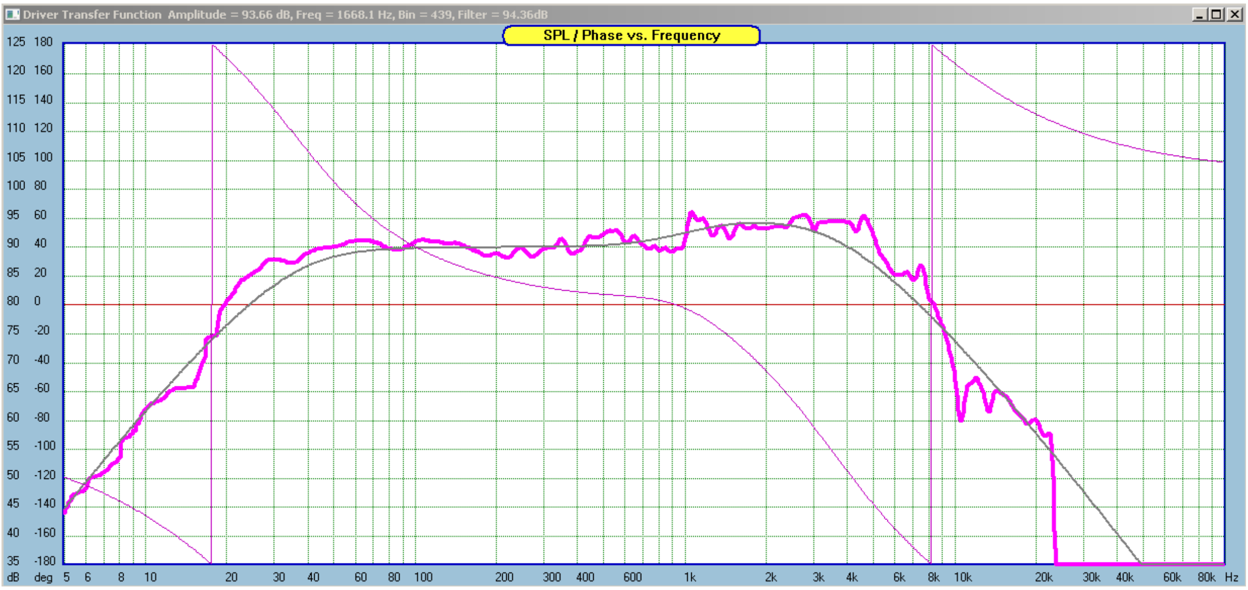


Figure 16.201. Optimized filter with incorrectly estimated slopes for the band-pass filter

Midrange Driver Example

Shown below is a midrange driver example. The process is exactly the same as for the woofer, so only the final results are show.

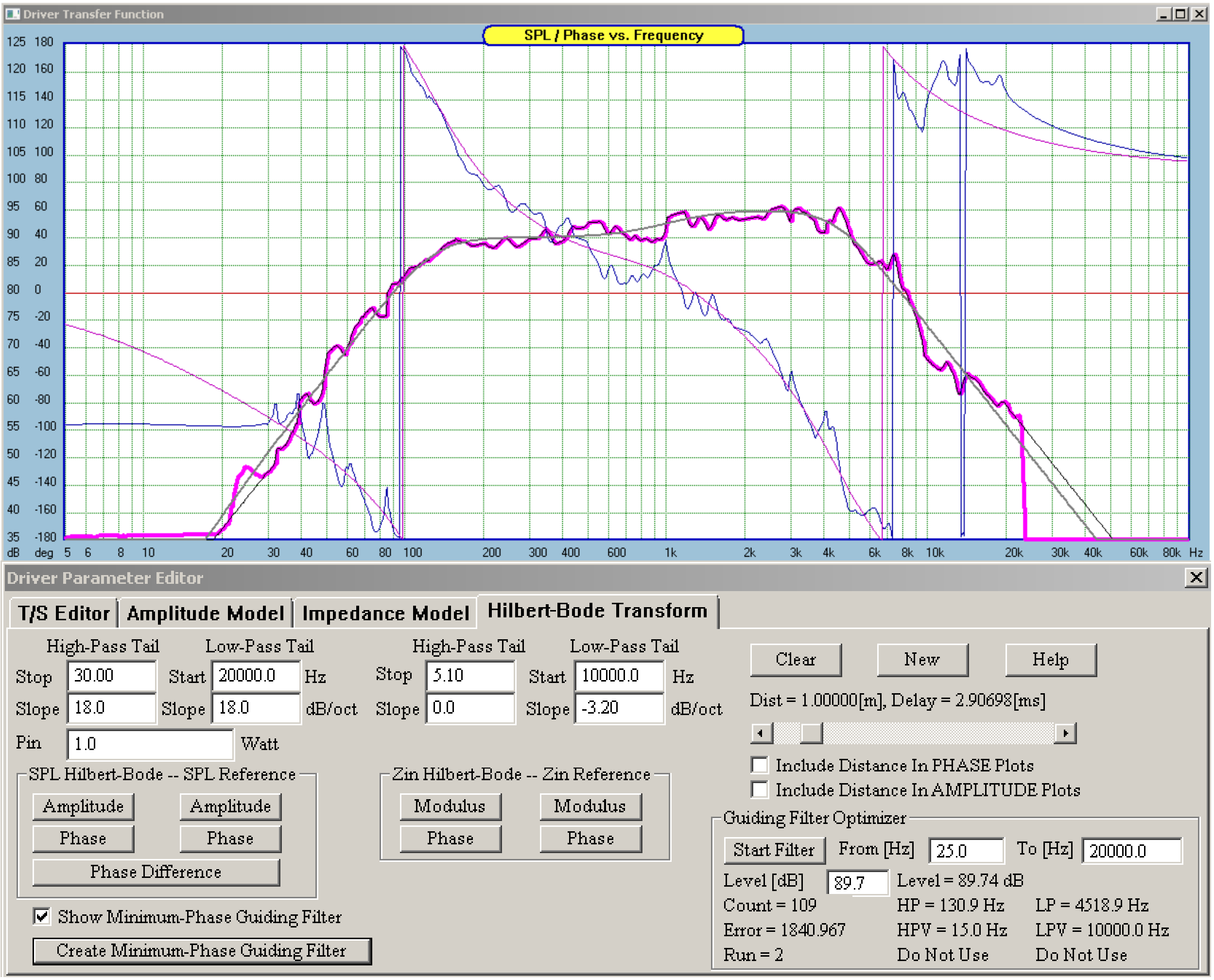


Figure 16.202. Optimized minimum-phase response of a midrange driver. Error = 1840.9.

Tweeter Example - Sampling 48kHz

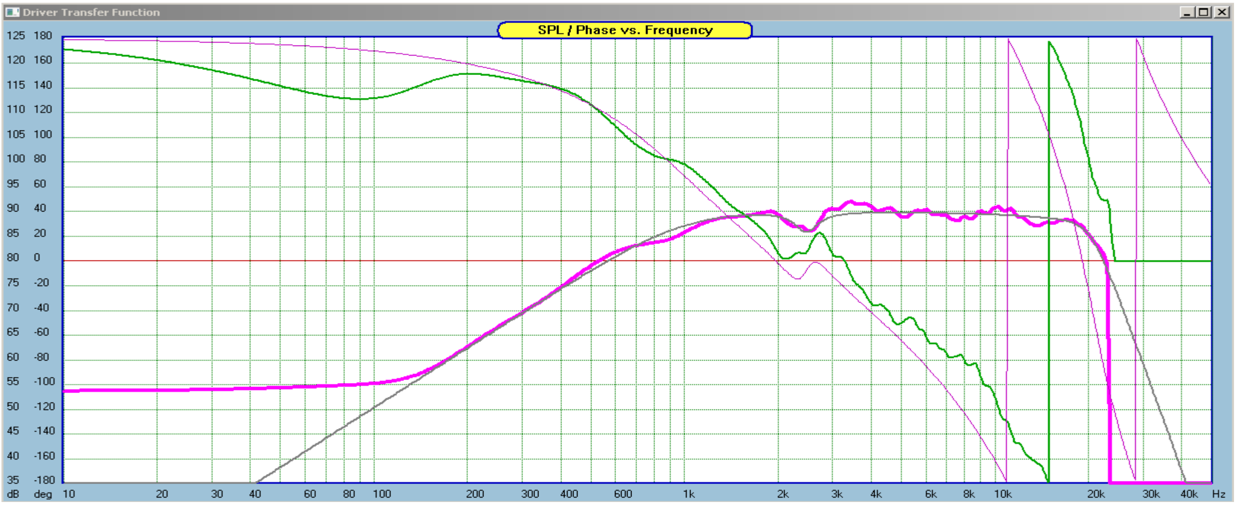


Figure 16.203. Measured SPL + band-pass filter of a tweeter driver. Green curve is the measured phase.

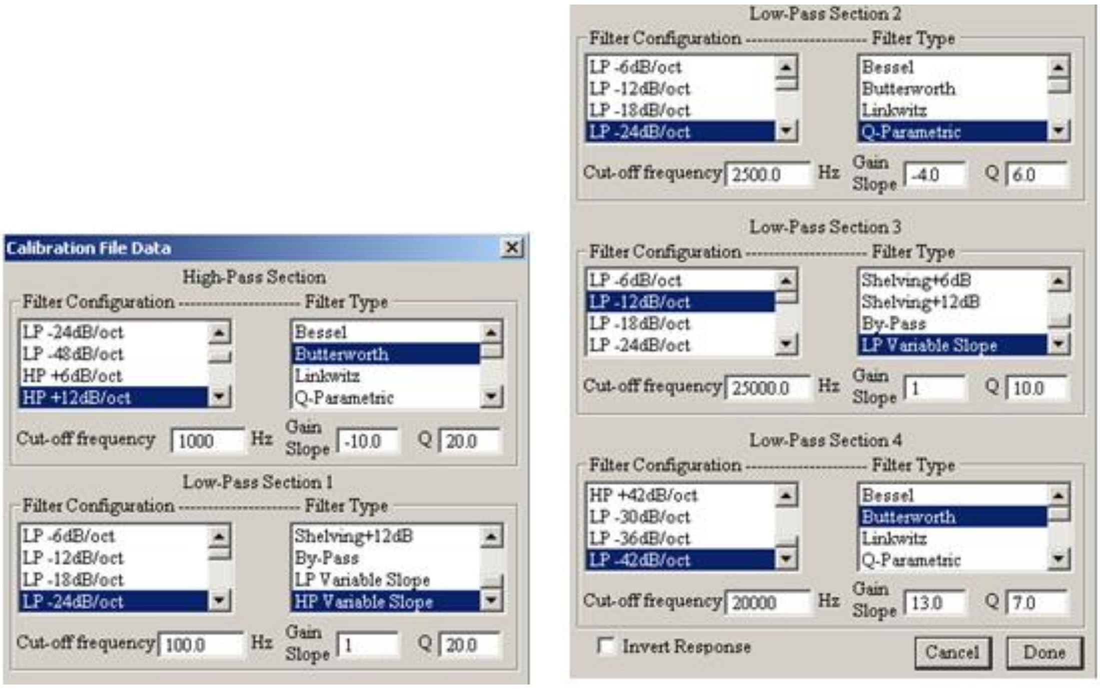


Figure 16.204. Selection of the band-pass filter initial parameters.

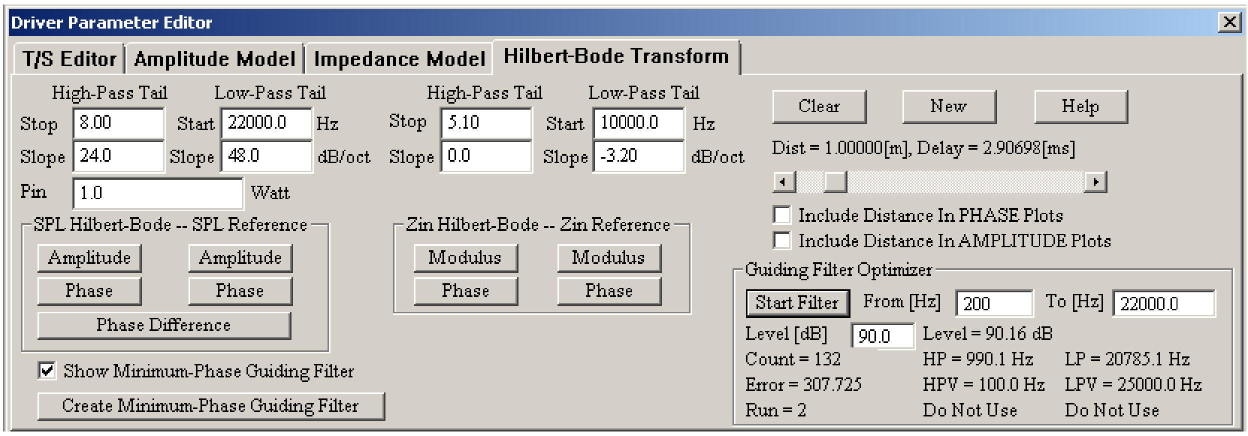


Figure 16.205. Optimization completed. Frequency range 200Hz – 22000Hz.

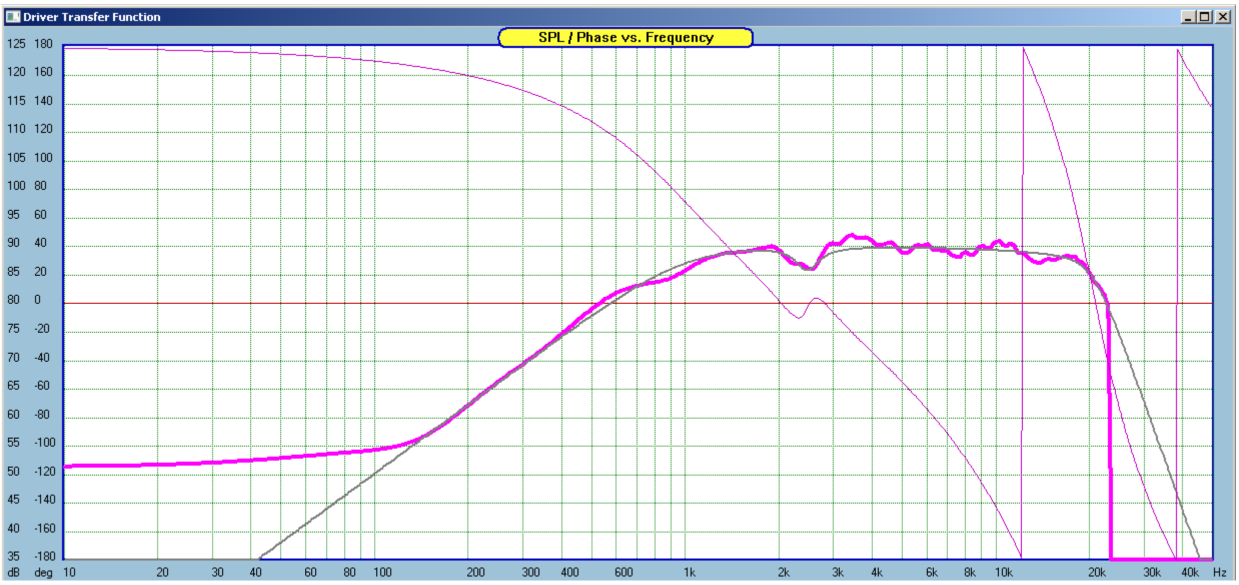


Figure 16.206. Optimized band-pass filter. Error = 307.

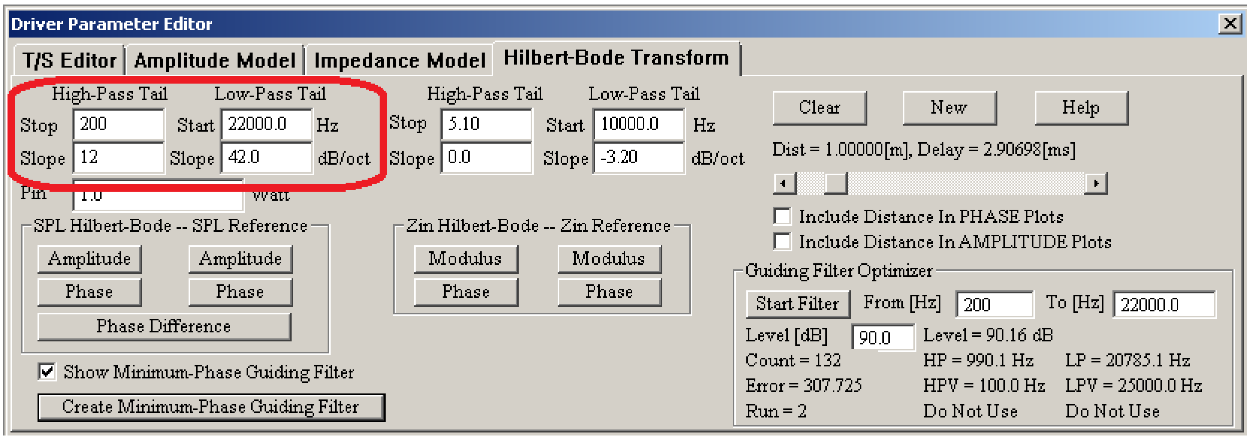


Figure 16.207. Attaching HBT tails to the measured SPL.

When the HBT parameters are entered into the HBT dialogue box, the matching between the Guiding Filter and HBT is remarkable.

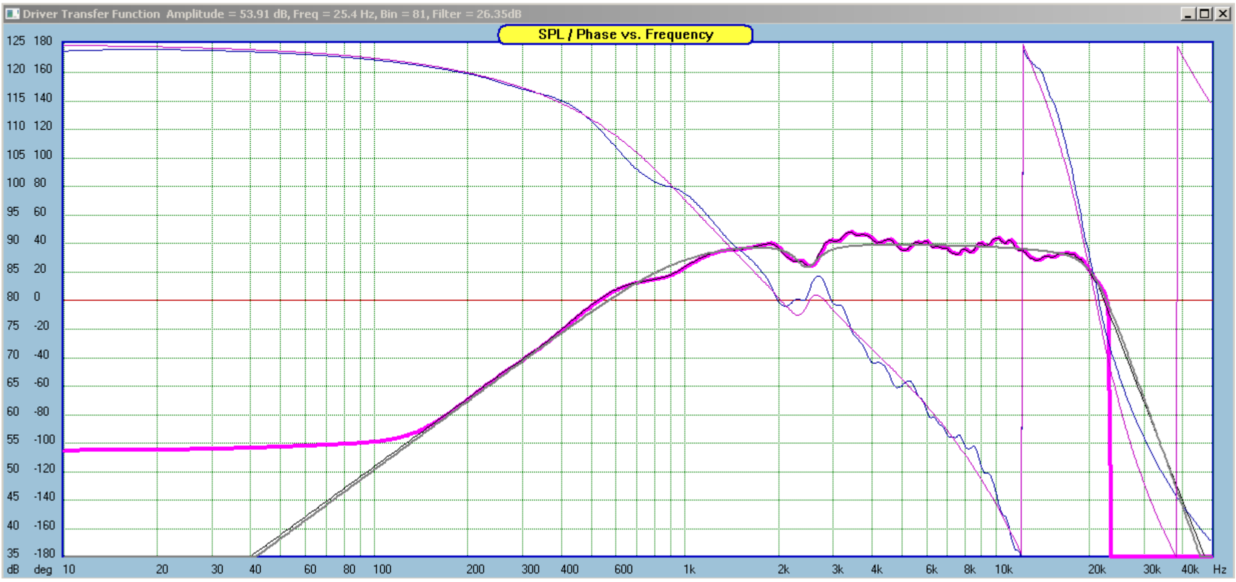


Figure 16.208. Pretty exact match between band-pass filter and HBT phase responses.

Phase response developed above can now be used as the "template" for setting the FFT window use in MLS and ESS measurement systems. Here is how it works.

Please go back the measurement system and compare phase of the measured SPL and HBT phase responses.

The green phase response shown on Figure 16.209 is the measured phase response. It makes the +180/-180 deg transition at about 12kHz – so it is too late. It appears, that the **FFT window has to be moved back by 1 bin, and a delay of 7.5 usec has to be added** to the measured phase to overlap the measured phase with the guiding filter model.

The 7.5usec delay is dues to the fact, that sampling frequency of 48kHz has coarse time steps of 20.833usec. Therefore, if the actual phase response results in FFT window placed between the time stamp points of the 48kHz sampling – this would not be possible. In a situation like this, you can still "manipulate the placement" of the FFT window by adding a delay to the Impulse Response thus affecting the phase of the FFT process.

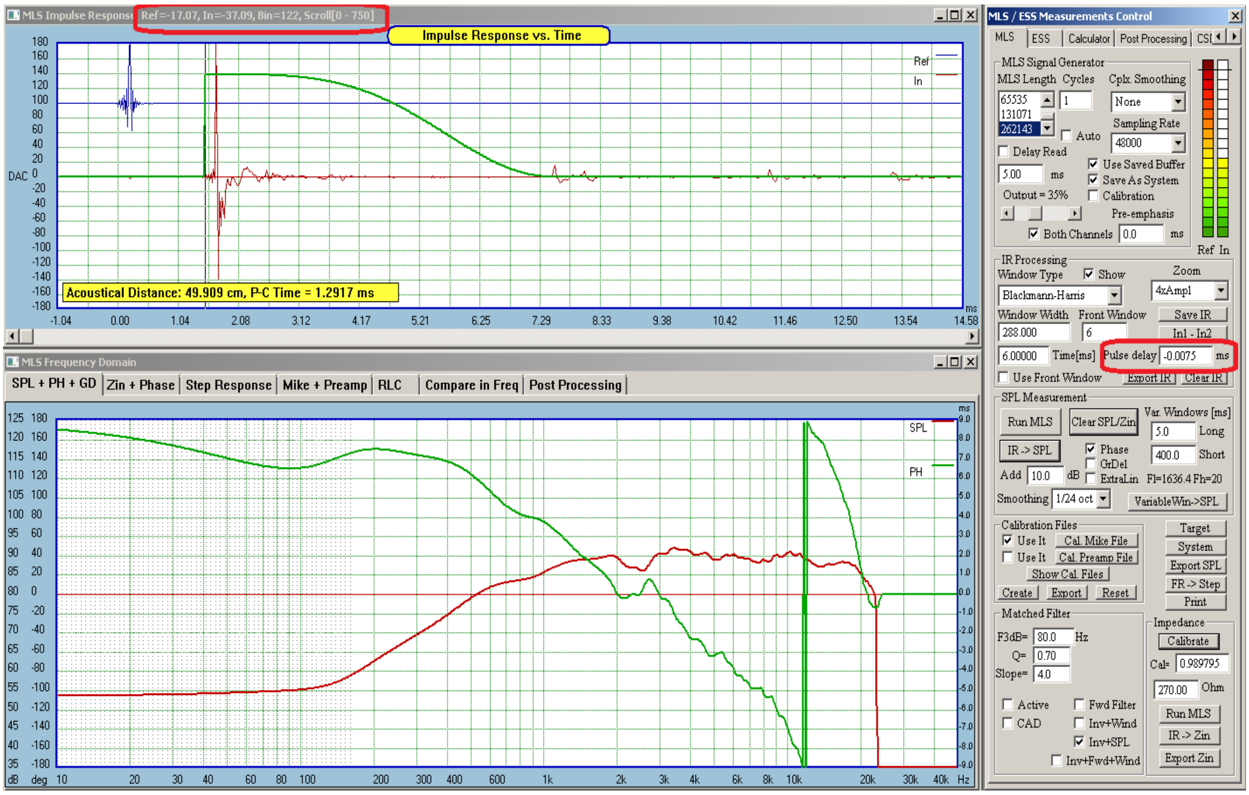


Figure 16.209. Initially measured phase response – green curve.

Once the FFT window was moved back and 7.5usc delay was added to the measured phase, we can go back to the Editor System and finally compare the phases.

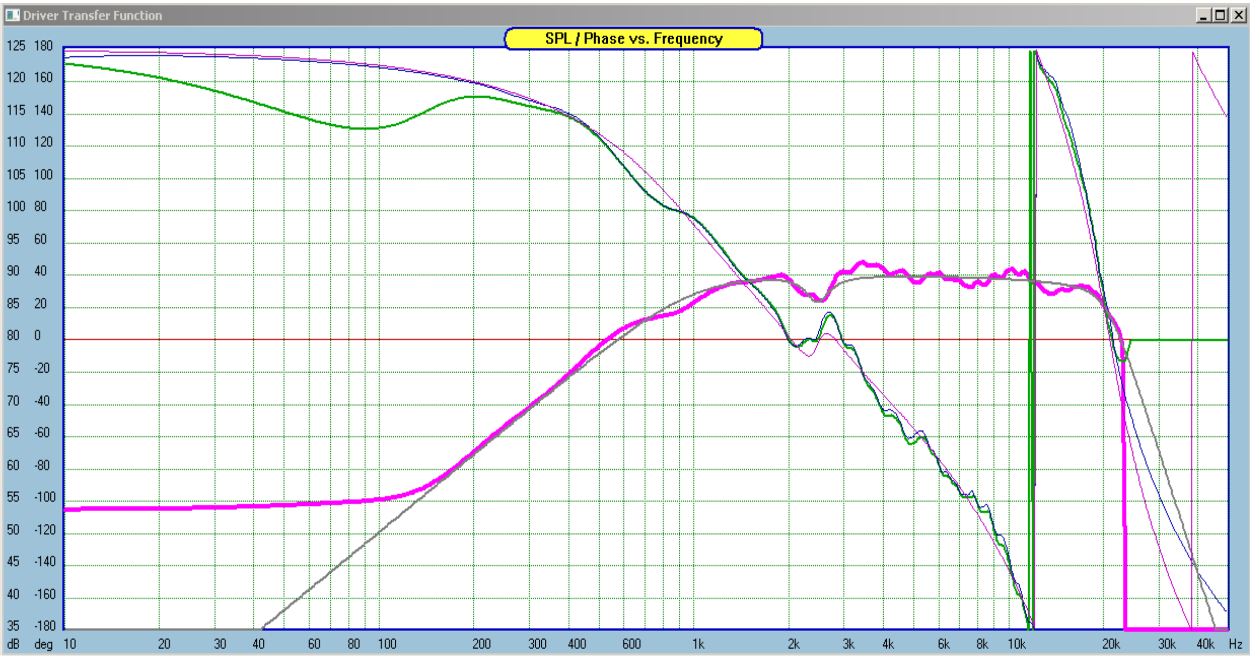


Figure 16.210. Perfect match of the band-pass filter, HBT and measured phase response.

You have now extracted minimum-phase phase response from the measurement. Please save the driver file.

Tweeter Example – Sampling 96kHz

In the next tweeter example, we examine the measurements made at 96kHz. Some development steps will not be repeated here, as they were explained in details before.

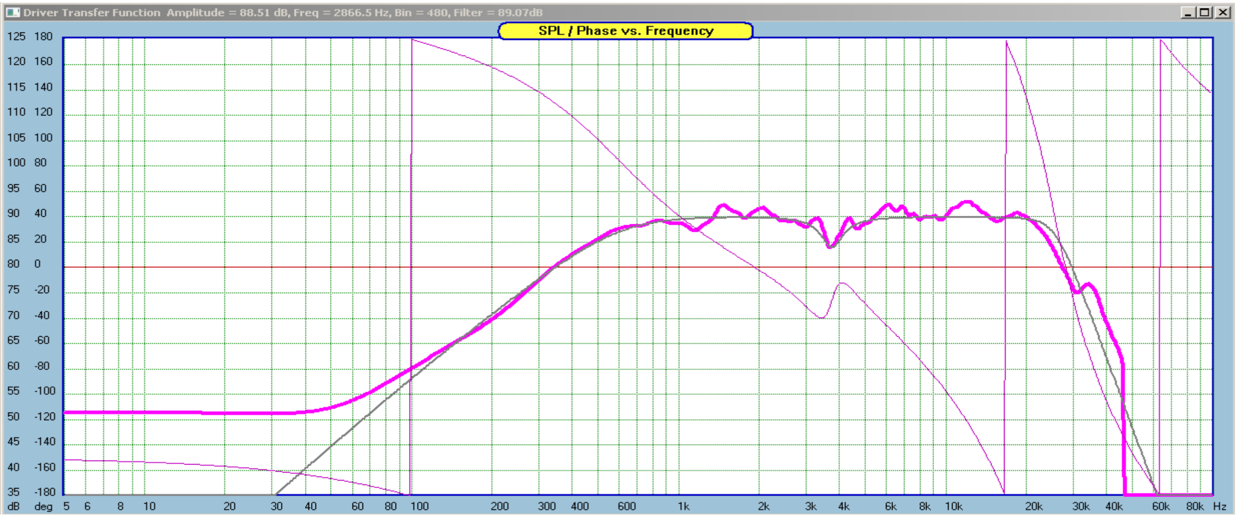


Figure 16.211. Measured SPL and band-pass filter with initial parameters shown below.

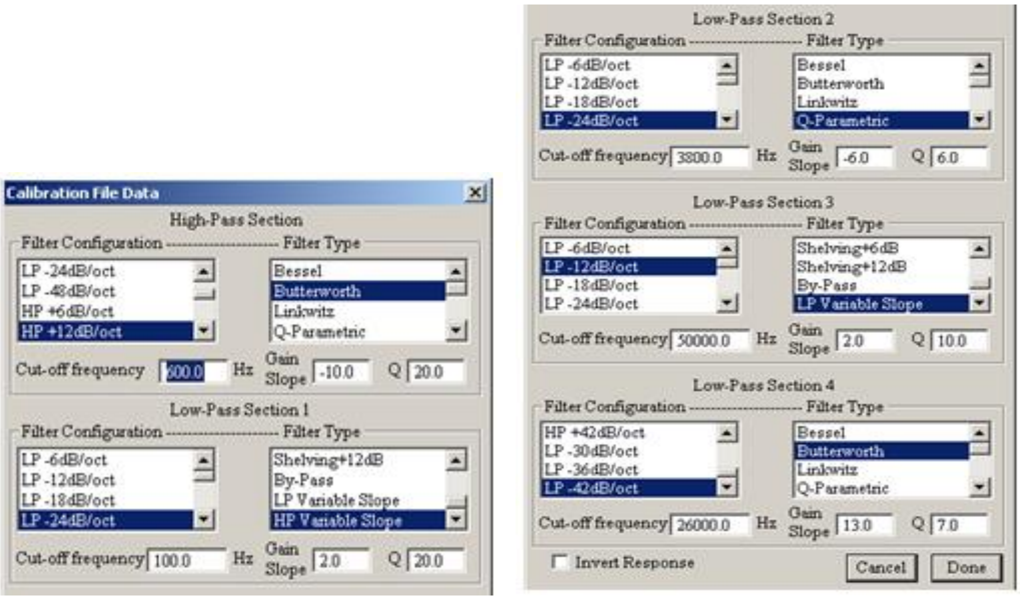


Figure 16.212. Initial settings of the Guiding Filter parameters.

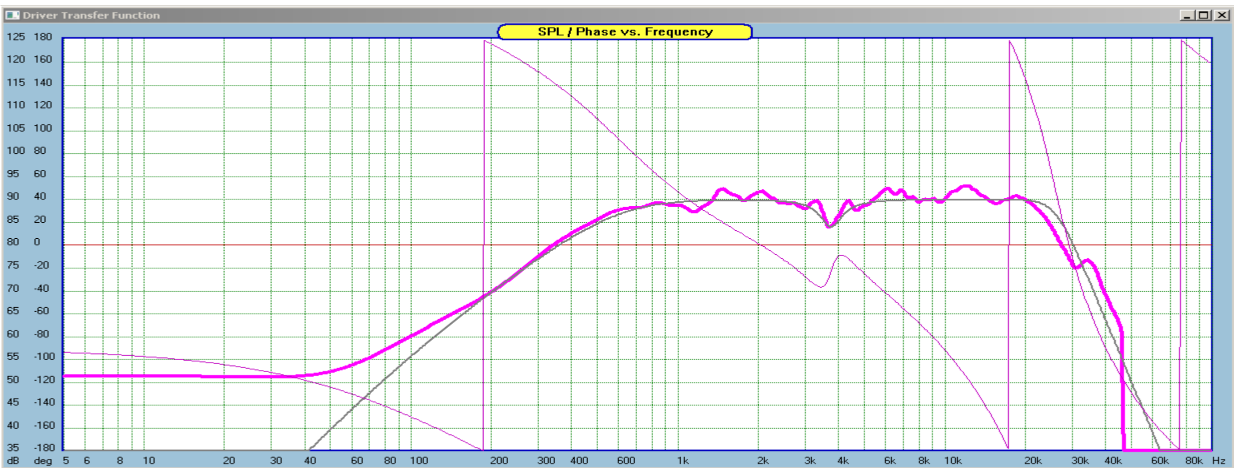


Figure 16.213. After optimization

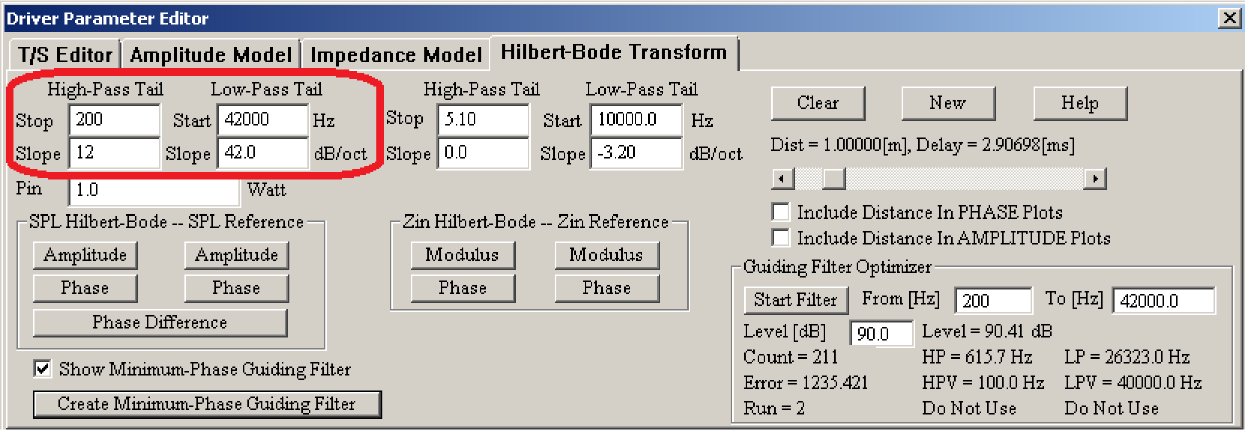


Figure 16.214. Optimization completed.

When the HBT parameters are entered into the HBT dialogue box, the matching between the Guiding Filter and HBT is remarkable.

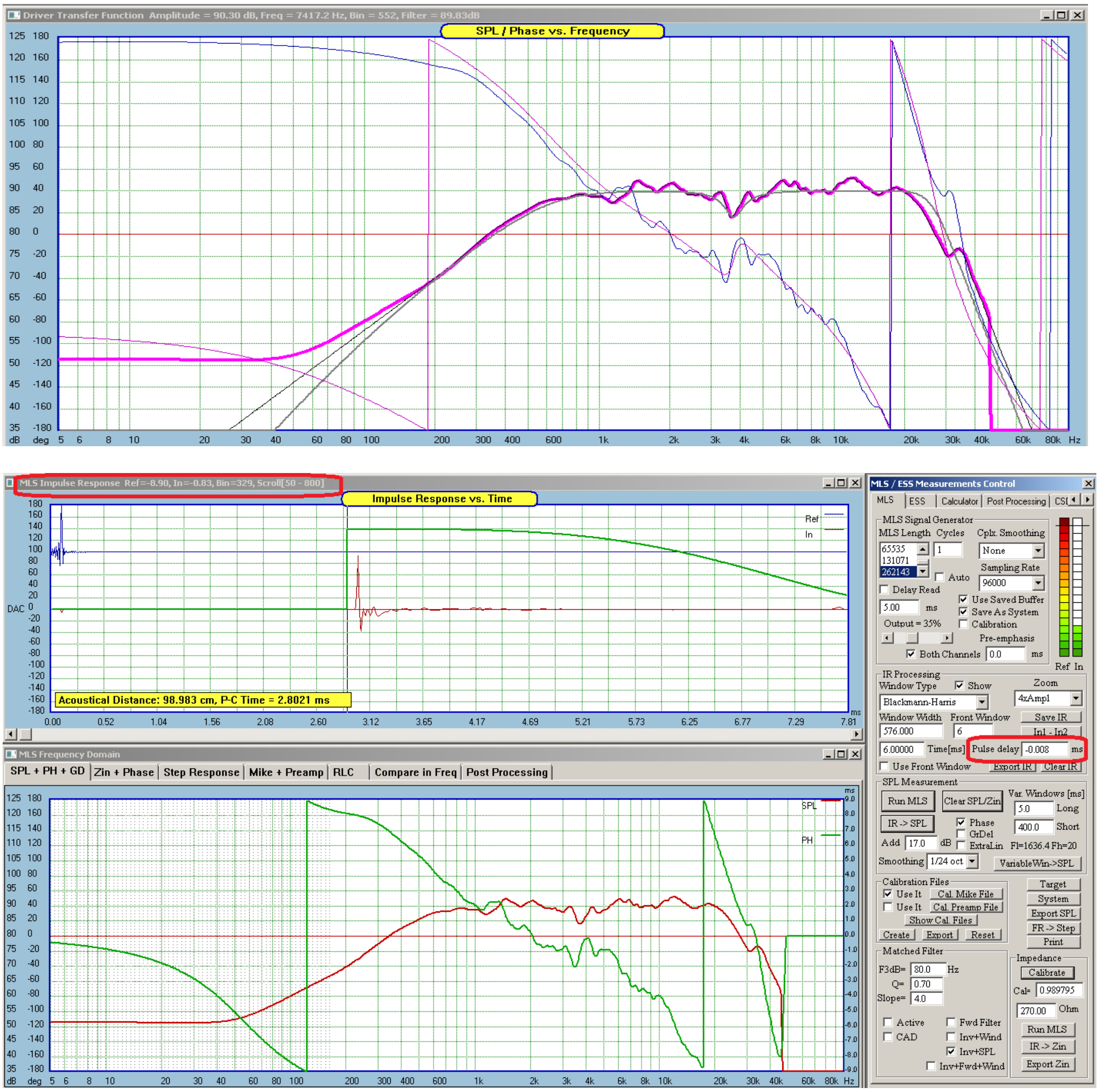


Figure 16.215. Measured SPL and Phase responses.

The FFT window has been moved and additional delay of 6 usec has been added to make the measured phase (green curve), overlap with the modelled phase. It is now clear what steps need to be taken for correct measurements of the phase response of this driver.

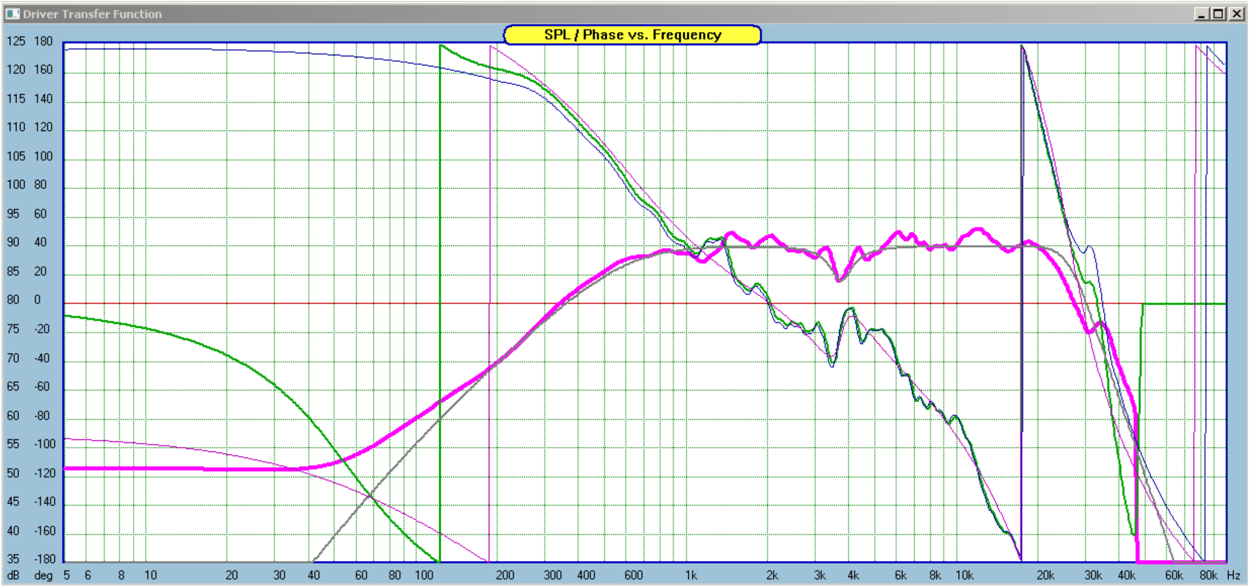


Figure 16.216. Perfect match of the band-pass filter, HBT and measured phase response.

In summary, the following steps are recommended.

1. Accurately measure the SPL of the driver. Needless to say, that environmental conditions need to be minimized as usual. Poor or insufficient data will affect the results.

2. Use the measured SPL as the template to develop simple band-pass model of the measured SPL.

3. Use curve-fitting optimizer to obtain the best fit between (1) and (2).

4. Use optimized asymptotic slopes as input parameters the HBT. The result is minimum-phase transfer function of the measured driver.

5. Go back to the measurement system. Use the phase response from band-pass model or HBT as the guiding template to place FFT window. Please note, that sampling rate chosen for the measurement may be too coarse to correctly place the FFT window. You may need to add/subtract some fine time delay to match the measured phase with the modelled one.

6. Now, all three components: (1) band-pass model, (2) HBT and (3) measured SPL/Phase must be in agreement over the operating frequency range.

Input Impedance At High Power

Let's, consider the following Zin measurement system.

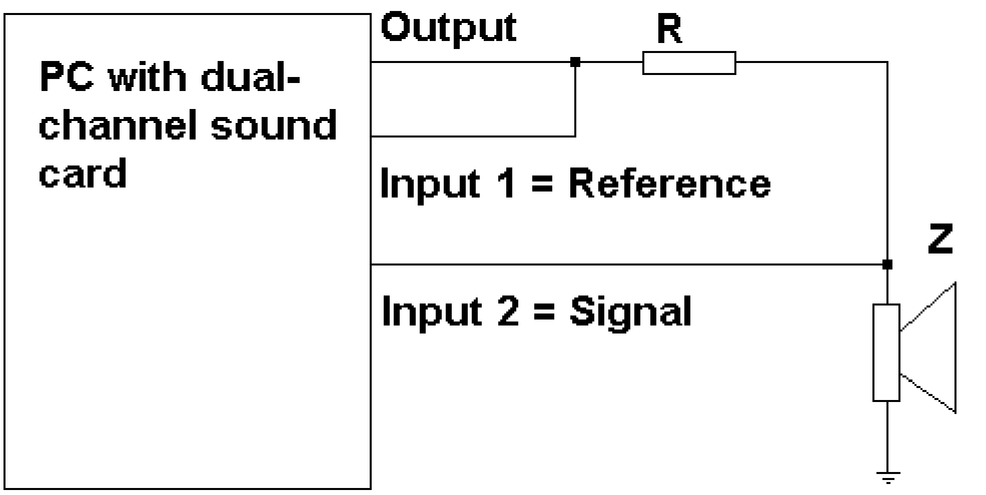


Figure 16.217. Simple distortion test circuit.

If the resistor, R is of relatively high value, say around 1kohm, then the current flowing through the loudspeaker, Z, is largely determined by the value of R. it can also be said, that the loudspeaker is driven by "constant current source". The driver level to the loudspeaker can not be large, because the current is limited by the large value of R.

An example of loudspeaker impedance, Z(ω) measured with 540ohm resistor is given below.

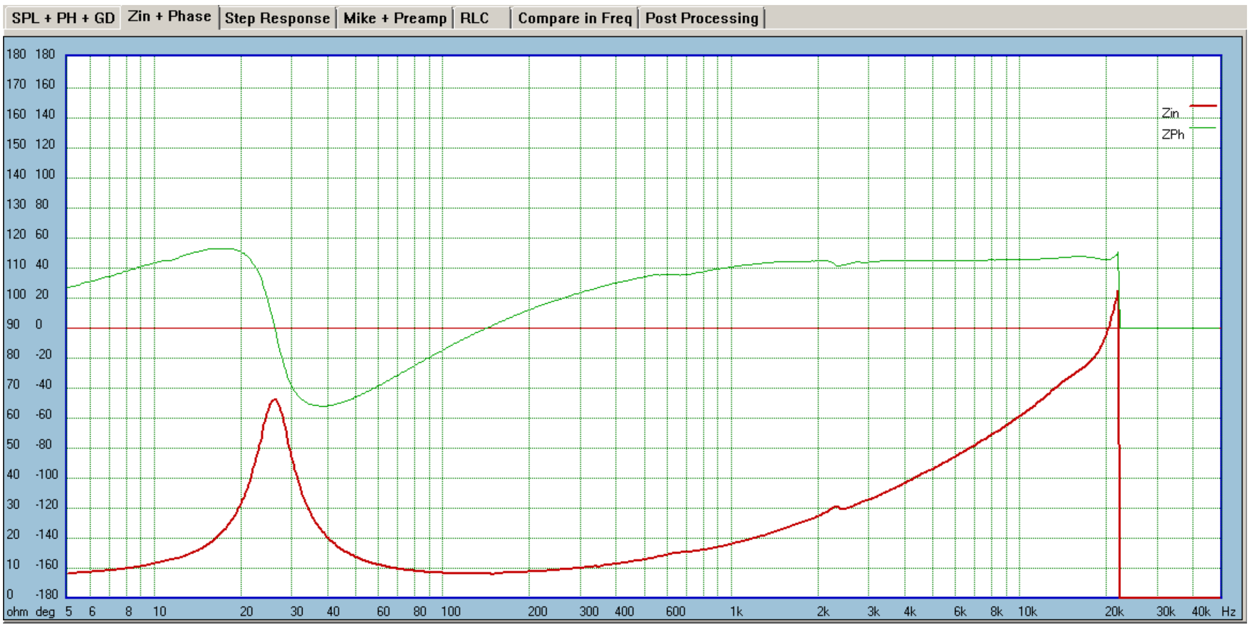


Figure 16.218. Input impedance curve measured with R=540ohm.

Loudspeakers operating in real-life conditions are rarely driven by such low power levels. Therefore, it may be beneficial to employ a measurement method, that would facilitate significant increase in power levels delivered to the loudspeaker under test. The test method relies on dual-channel measurement, employing the following formula:

$$Vsignal = Vref\frac{Z}{Z+R}$$

From which: $$Z = R\frac{Vsignal}{Vref - Vsignal}$$

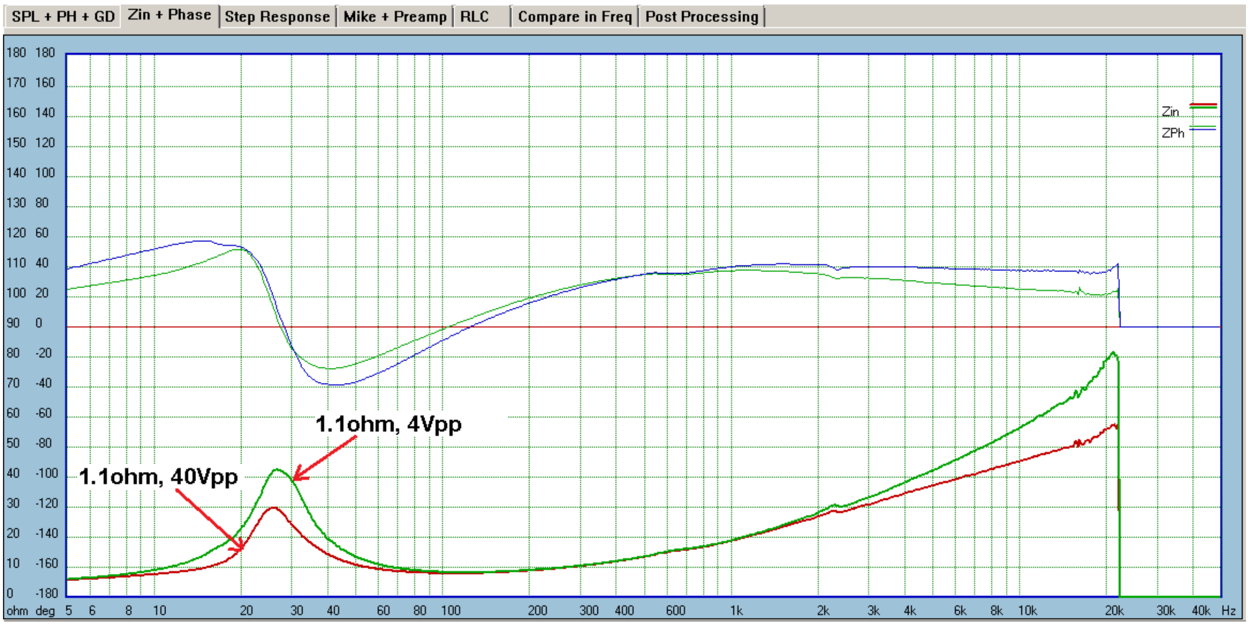


Figure 16.219. Input impedance curve measured with R=1.1ohm at different power levels.

The choice of 1.1ohm resistor is simply two 2.2ohm resistors connected in parallel. Such small resistor value allows for significant increase of power delivered to the loudspeaker during the test. In fact, an 8ohm loudspeaker can be driven to the full operating power in this test.

It is observable on Figure 16.219 above, that 40pp impedance/phase curve, exhibits somehow different shape than typical input impedance curve. Indeed, the loudspeaker was driven into non-linear BL region and the voice coil started to leave the magnetic gap flux.

Lastly, a word of caution when measuring Thiele-Small parameters. The T/S parameters are derived from linear loudspeaker model, where the parameters are measured using constant current method. Figure 16.218 shows such method in action, when the reference resistor was quite large: R=540ohm.

However, if your typical application is a high-power loudspeaker system, operating at elevated power levels for prolonged periods of time, you may still want to determine the T/S parameters at higher power levels. This is because enclosure design is based on the T/S parameters. Well, it’s a designer choice – as usual.

Semi-Inductance Impedance Model

The Semi-Inductance Model has been proposed by K. Thorborg, A.D. Unruh and C.J. Struck in 2007 in their AES paper ”An Improved Electrical Equivalent Circuit Model for Dynamic Moving Coil Transducers”.

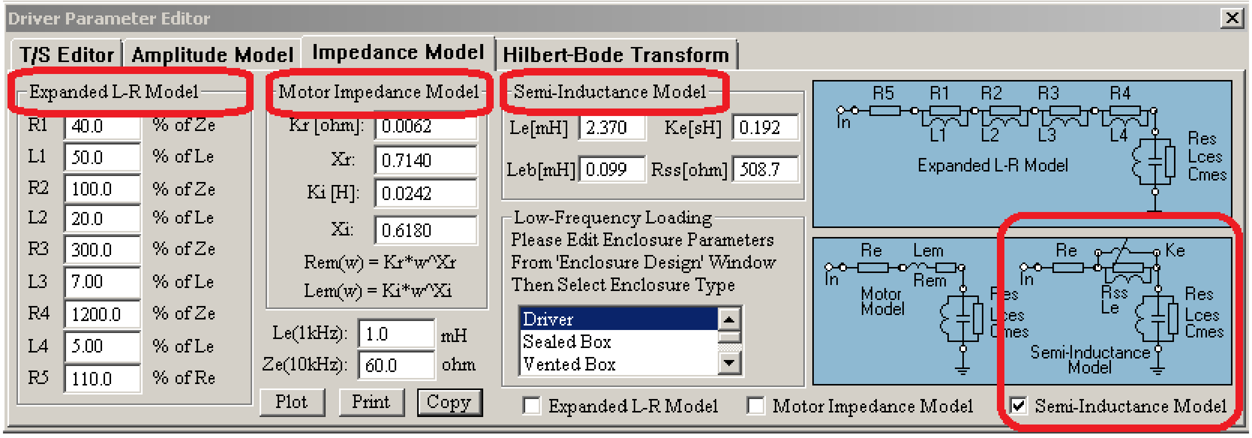
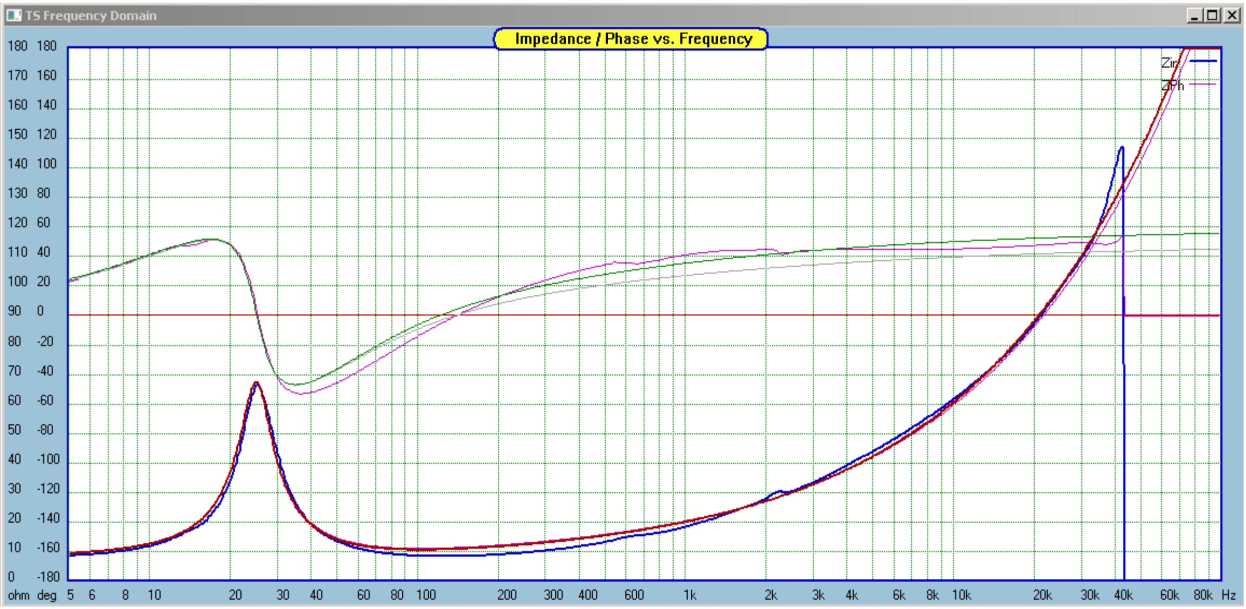


Figure 16.220. Three available Impedance Models.

There are three impedance models implemented in the program. The current addition – the Semi-Inductance Model is proving to be the most accurate. First, the Motor Impedance Kr, Xr, Ki and Xi parameters will be extracted and the degree of fit will be calculated – this is the **Error = 6793** figure.



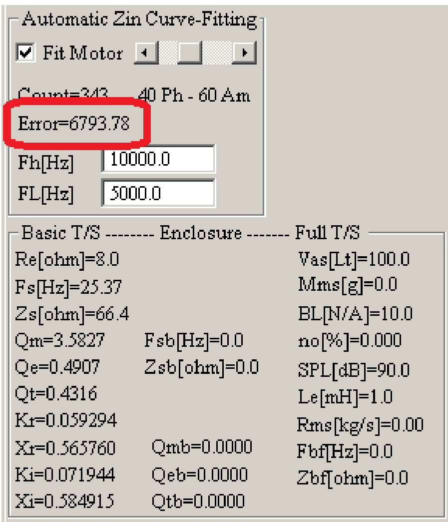


Figure 16.221. Extracting parameters for Motor Impedance Model.

Next, the Semi-Inductance method is employed and Rss, Leb, Ke and Le parameters are extracted, together with Lp, Cp,and Rp.

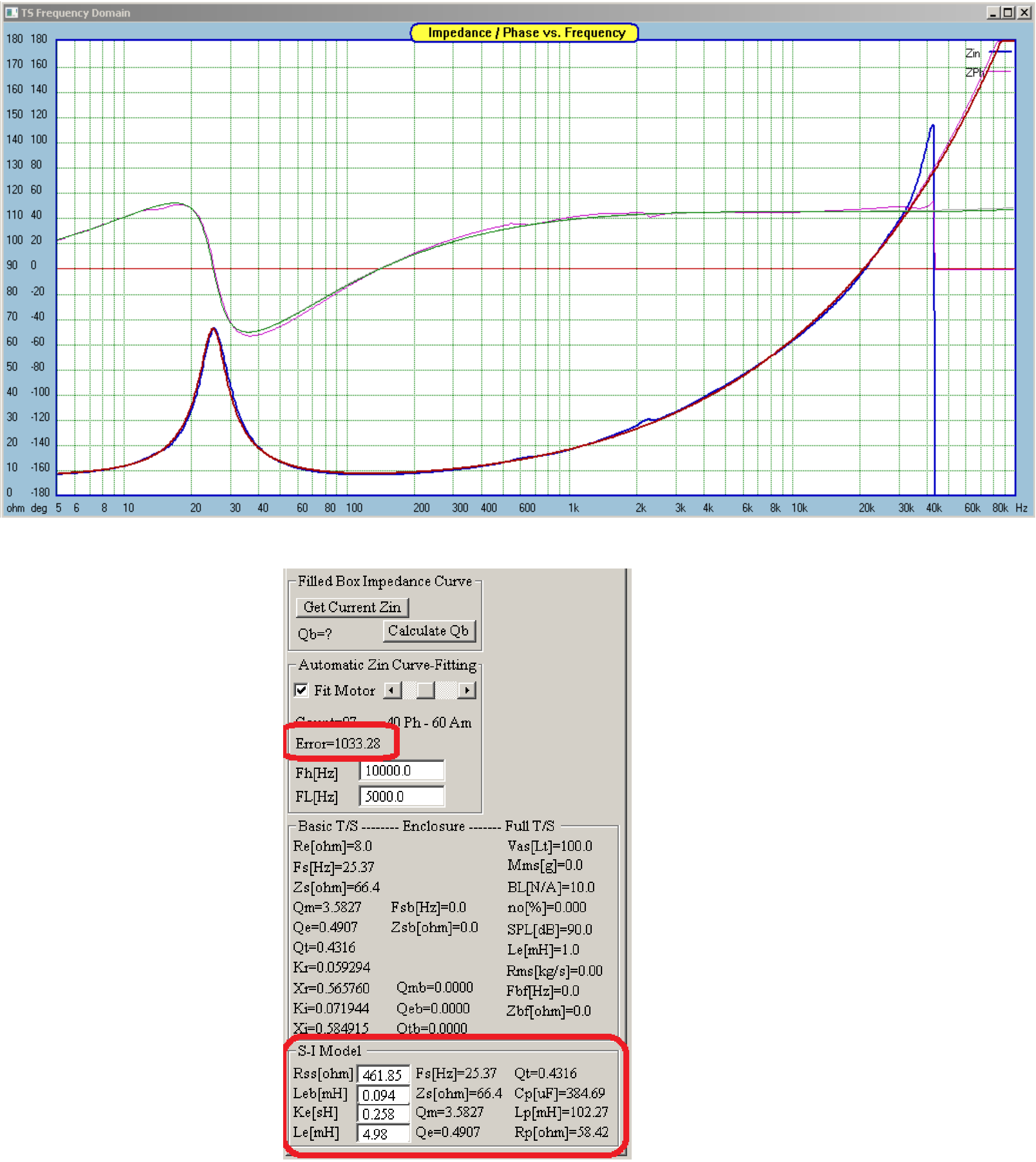


Figure 16.222. Extracting parameters for Semi-Inductance Model.

It is clearly observable, that **Error = 1033**, which is more than 6 times better than the degree of match provided by the Motor Impedance Method. All extracted parameters can be used for driver's modelling.

Shown below, are two examples of where the impedance model makes it's mark – these are sealed and vented enclosure SPL models, with Motor Impedance (Green Curve) and Semi-Inductance (Brown Curve) used as the basis for calculations.

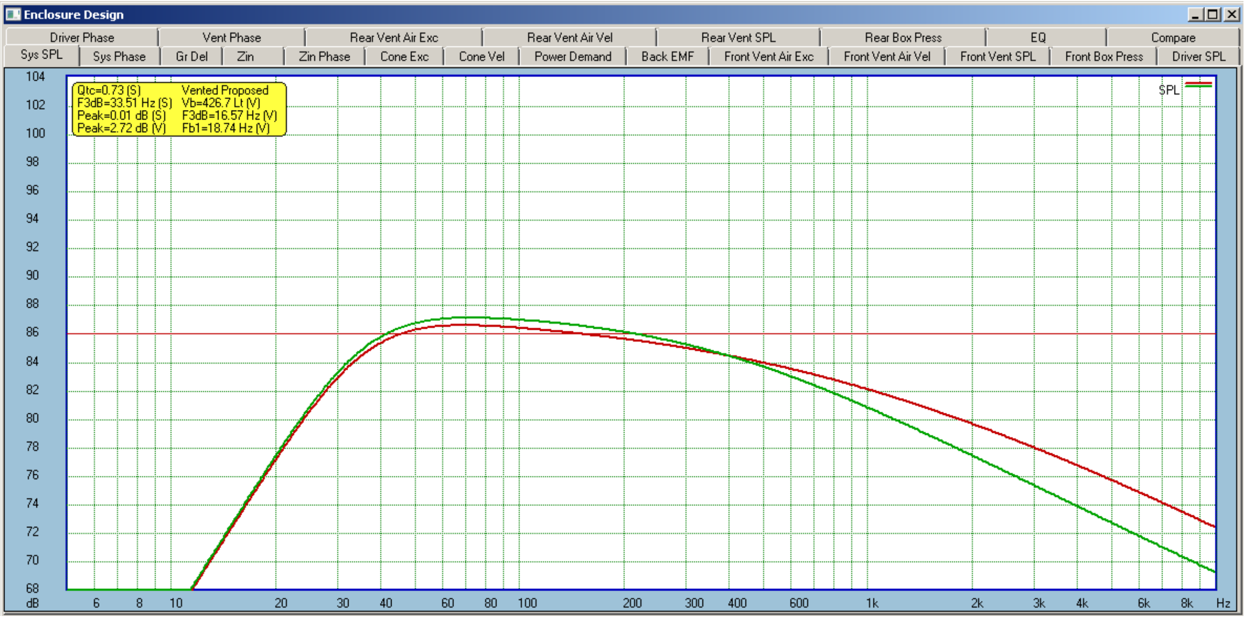


Figure 16.223. Green – Motor Impedance Model, Brown – Semi-Inductance Model.

CSD Comparisons

Over the years, the Cumulative Spectrum Decay (CSD) has proven itself to be a valuable development tool. It is easy to observe where any unwanted, and ringing resonances occur in the driver/enclosure combination. However, it is not easy to see fine changes in the CSD waterfalls when the designer attempts to fix the problems. A simple CSD Comparison feature would provide much needed tool to emphasise the most recent change in the design.

To illustrate this feature, two loudspeakers were measured and CSD waterfalls were developed.

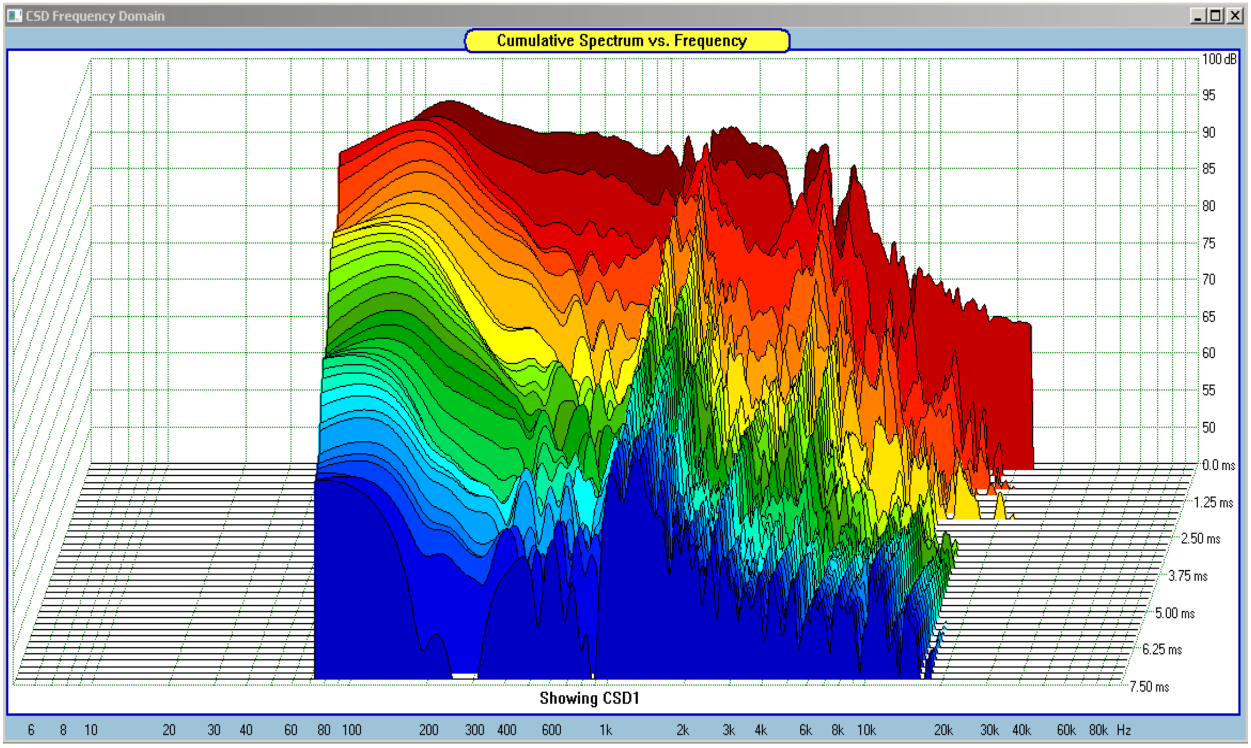


Figure 16.224. CSD1 – measured Driver 1.

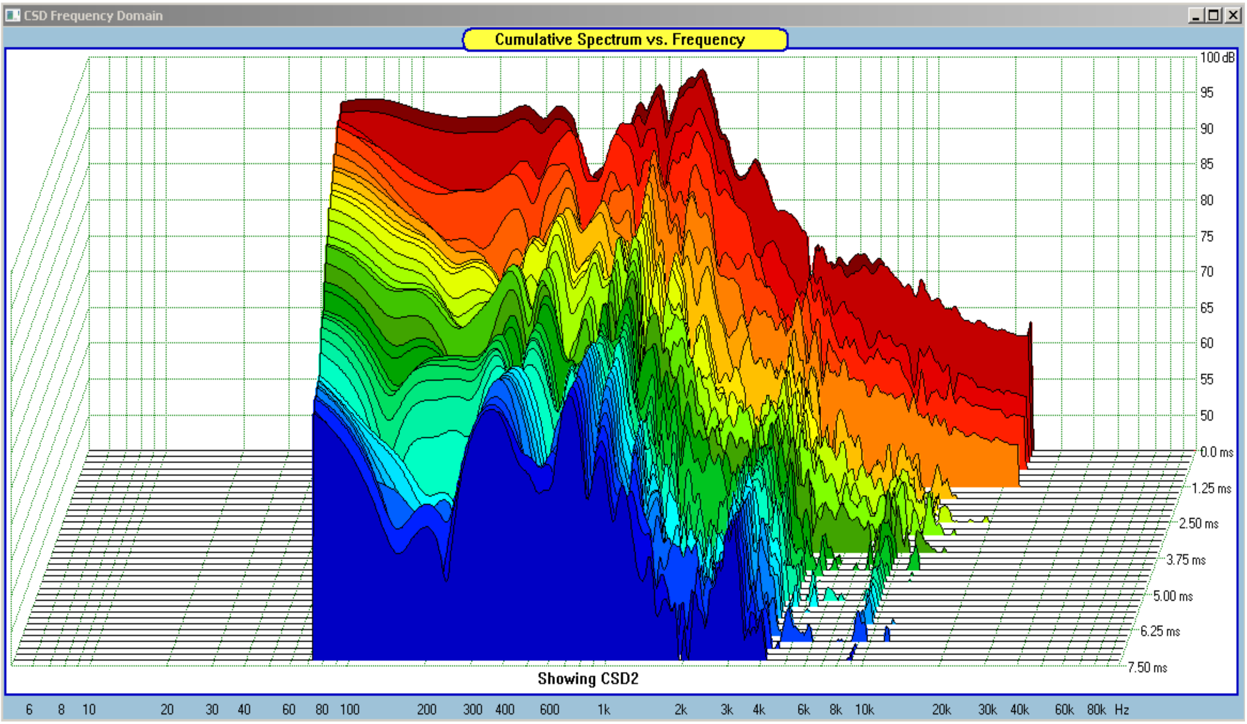


Figure 16.225. CSD1 – measured Driver 1.

Both CSDs were saved to HD using the controls provided – see Figure 16.226 below. Both measurements can be recalled as well.

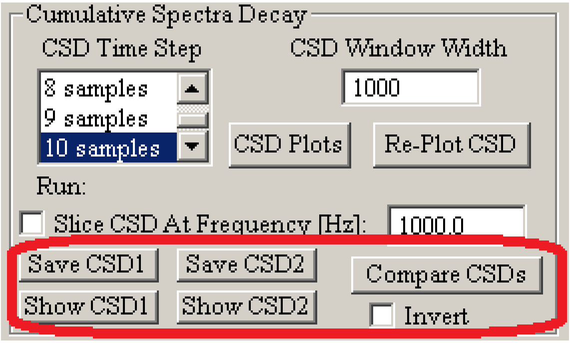


Figure 16.226. CSD controls.

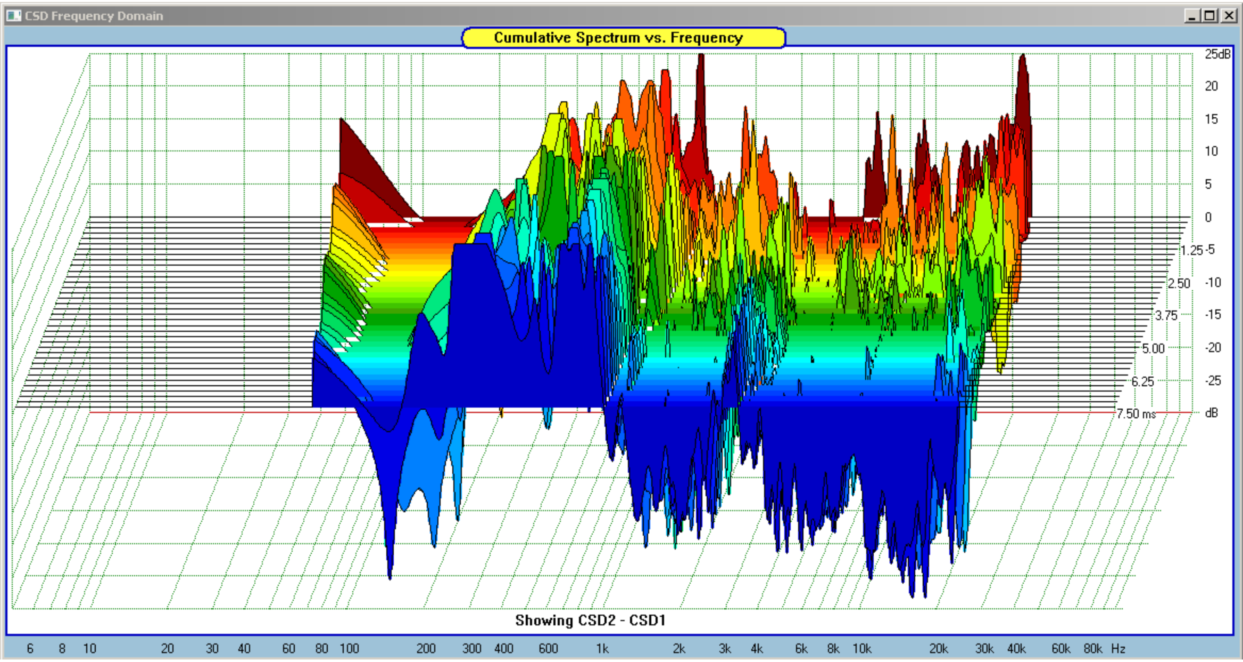


Figure 16.227. CSD1 – CSD2

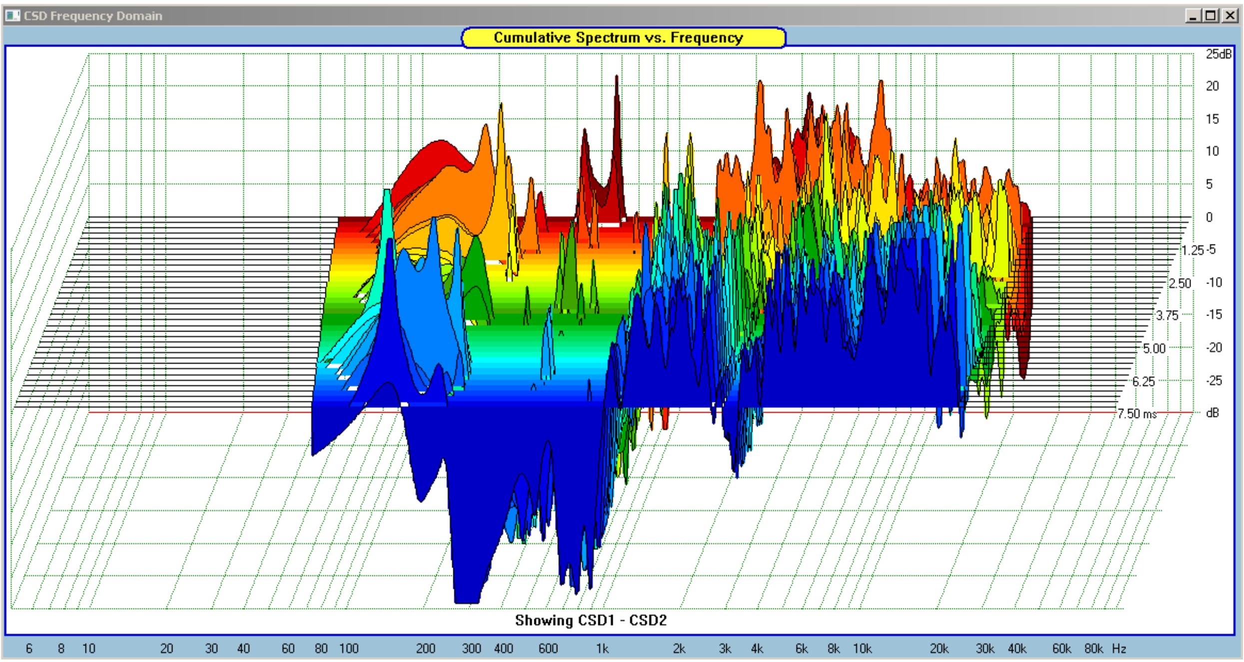


Figure 16.228. CSD2 – CSD1 (Inverted)

Inverse Hilbert-Bode Transform

Introduction and remainder on HBT

Before we introduce Inverse Hilbert-Bode Transform (IHBT), a short summary of the original Hilbert-Bode Transform (HBT) is presented.

The HBT was introduced to the DIY community about 20 years ago and it is a computational algorithm, that allows the user to extract phase response from the known magnitude response, or SPL curve. Since the HBT is based on an integral calculated from DC to infinity, one can immediately see the problem of supplying SPL data points over this impossible frequency range. However, it turns out, that calculating the integral over a narrower and much more manageable frequency range affects the total phase accuracy only in a minimal way. However, the other problem of extending the asymptotic slopes of the SPL curve towards zero frequency and infinite frequency still remains. Some substitute methods for determining the slopes have been proposed, and the method of "Optimized Guiding Filter" is possibly the most accurate approximation. The IHBT properties make it perhaps more suitable for solving the above dilemma.

Introduction to "Phase Slopes" Corresponding to SPL Slopes.

When dealing with traditional HBT the user needs to supply asymptotic SPL slopes, whose tangent is expressed in dB/oct. The SPL slopes, together with the measured SPL constitute input data into the HBT algorithm. We are all familiar with SPL slopes. Similarly, the IHBT requires **asymptotic phase slopes** to be attached to the phase response at selected frequency points. This combined phase response constitutes input data into the IHBT algorithm. Here is the problem: we are not really familiar with the concept of "phase slopes" expressed in deg/oct. In order to make the IHBT a bit more digestible, the phase slopes (or tails) are calculated from the equivalent SPL slopes. In the IHBT "language" this will be called "High-Pass PHASE Tail Equivalent To SPL" slope attached at say, 35Hz with 24dB/oct slope. From this data, the algorithm will calculate corresponding phase slope and will attach it to the measured phase at the specified frequency point. Same deal for "Low-Pass PHASE Tail Equivalent To SPL" slope.

SPL from Phase for Simple Filters

In order to start visualizing the operation of IHBT, a couple of simple filters were constructed and their phase responses were supplied to IHBT algorithm. Figure 1 below, shows a simple band-pass filter with +/-24dB/oct slopes and a shelving component. The thick black curve, which is the SPL calculated via IHBT, actually overlaps the original SPL curve, so it can not be seen. Figure 16.229 below, shows a simple band-pass filter with +/-48dB/oct slopes, a shelving component and a Q_parametric component. The thick black curve, which is the SPL calculated via IHBT, actually overlaps the original SPL curve, so it can not be seen again.

Figure 16.229 SPL extracted from phase response of a simple filter 1.

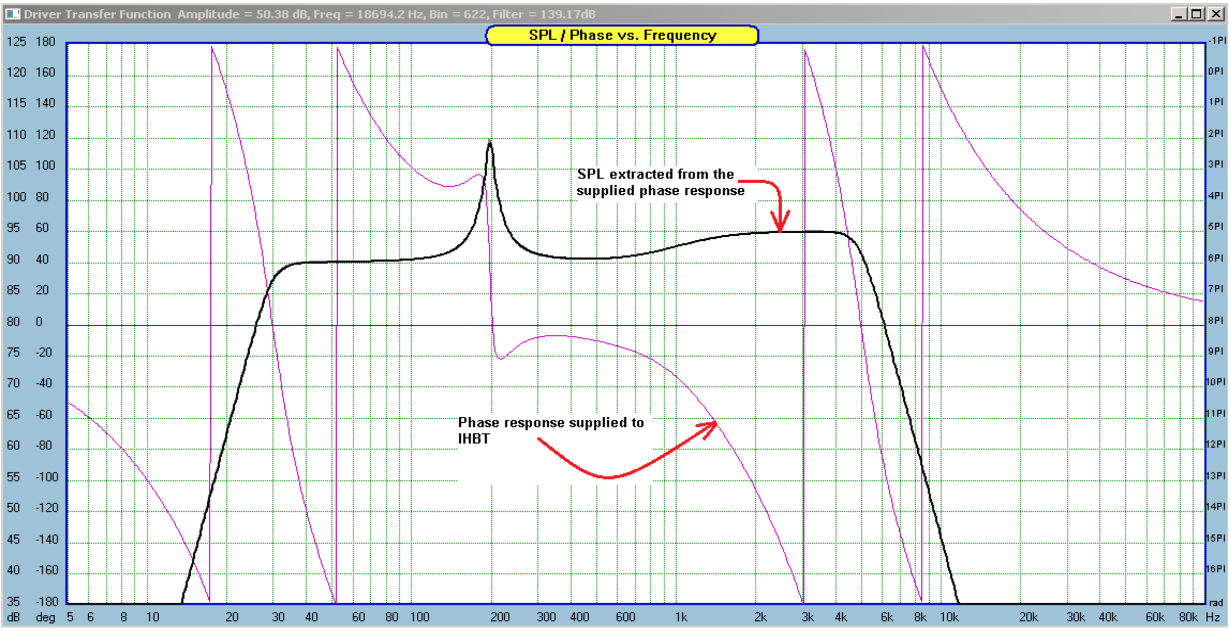
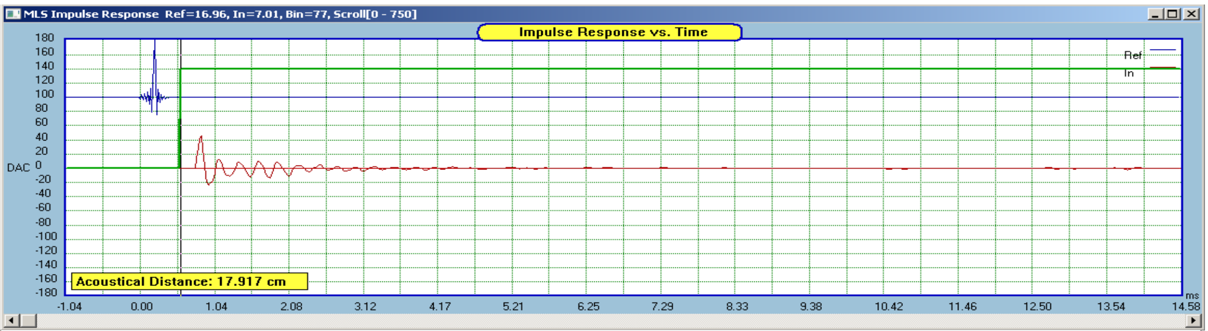


Figure 16.230. SPL extracted from phase response of a more complex filter 2.

Next, we will examine a real loudspeaker phase response.

SPL from Phase for Measured Example Loudspeaker

The MLS system was used to measure a 12" loudspeaker in vented enclosure, so the asymptotic slope on the low-frequency side would be 24dB/oct.



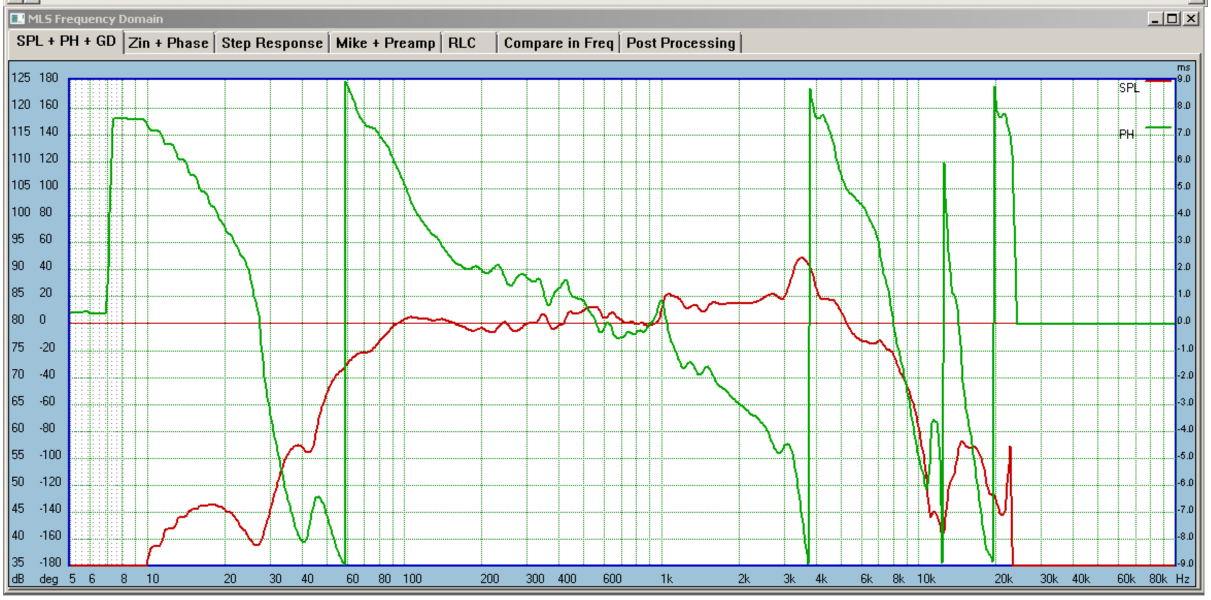
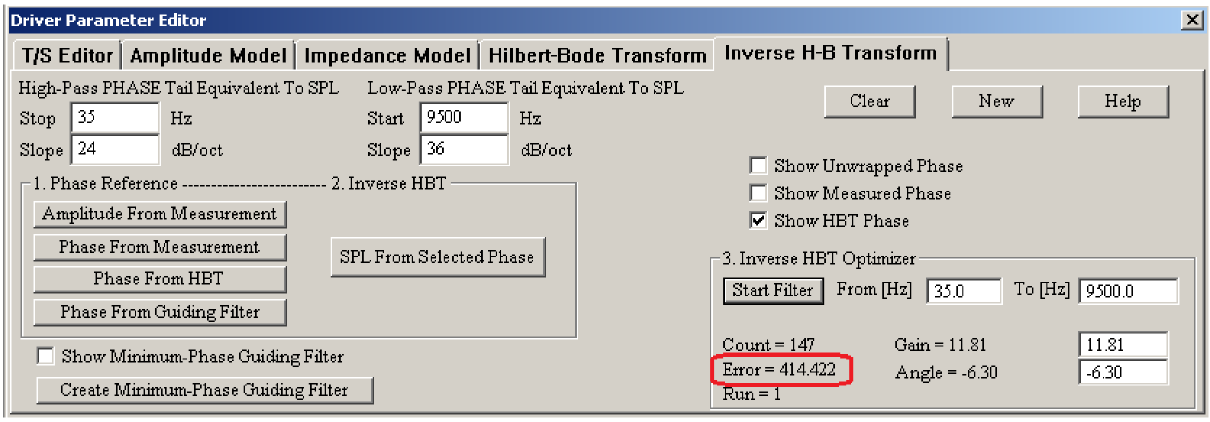


Figure 16.231. MLS measurement of SPL/Phase of a 12" loudspeaker with FFT window at Bin 77.

The above measured phase response (green), with FFT window set to start at Bin 77 was supplied to the IHBT algorithm. Settings for the IHBT algorithm were as follows: Stop = 35Hz, Slope = 24dB/oct and Start = 9500Hz, Slope = 36dB/oct, frequency range of IHBT optimization 35 – 9500, Start IHBT Optimizer with: Gain = 0.0, Angle = 0.0 After the run, **Error = 414**. It is observable, that SPL does not match the measured SPL from 35Hz - 3kHz, then the match is a lot better.



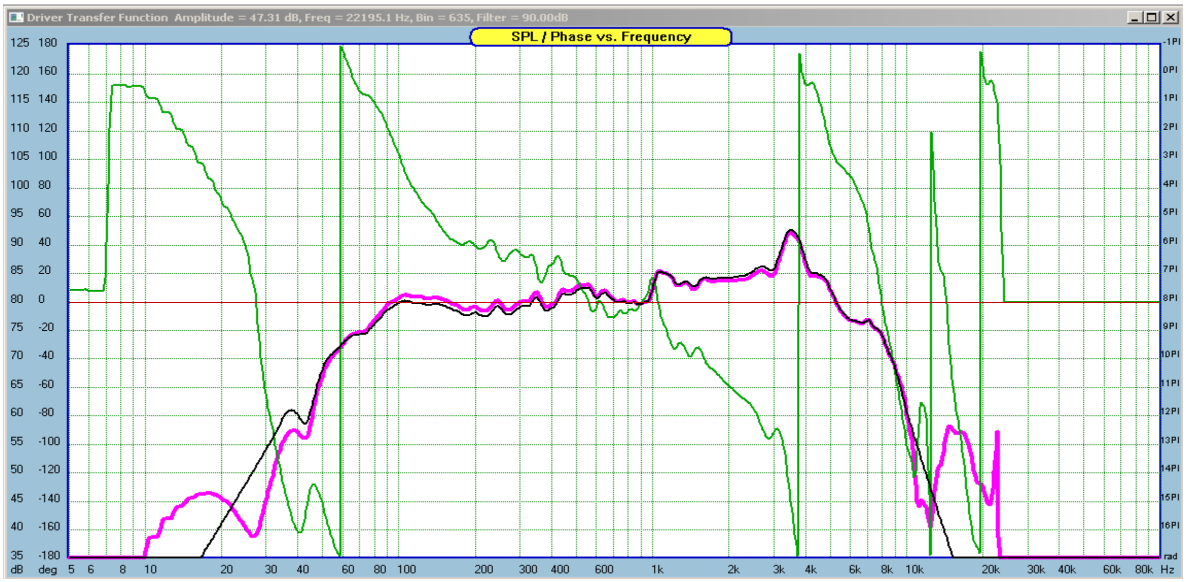
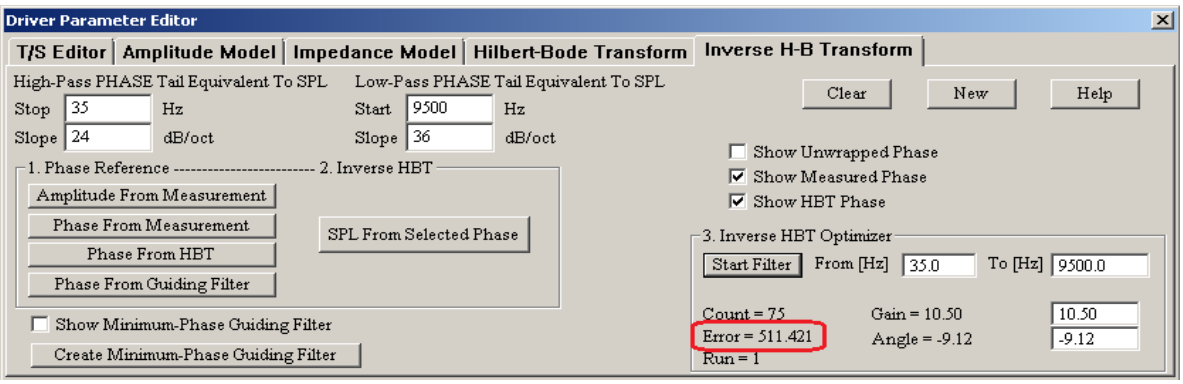


Figure 16.232 MLS measurement of SPL/Phase of a 12” loudspeaker with FFT window at Bin 77.

Then FFT window in MLS system was moved to Bin 76, and the process was repeated. The **Error = 511**, and was higher than for Bin 77.



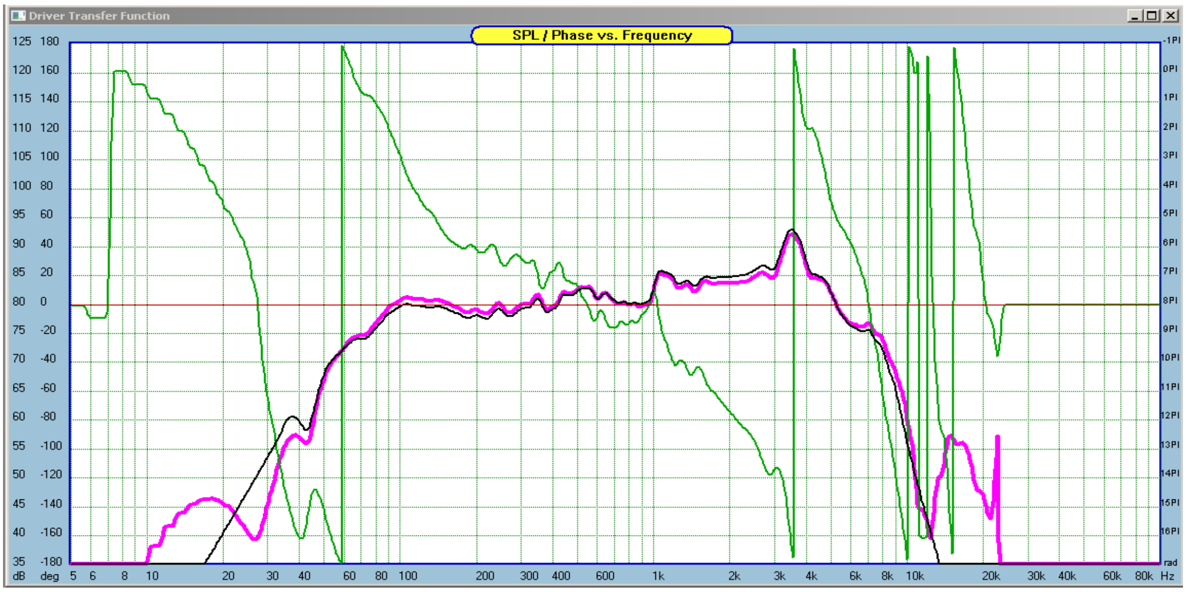


Figure 16.233 MLS measurement of SPL/Phase of a 12” loudspeaker with FFT window at Bin 76.

Finally, the FFT window was moved to Bin 78, and the process was repeated. The **Error = 567**, and was again higher than for Bin 77.

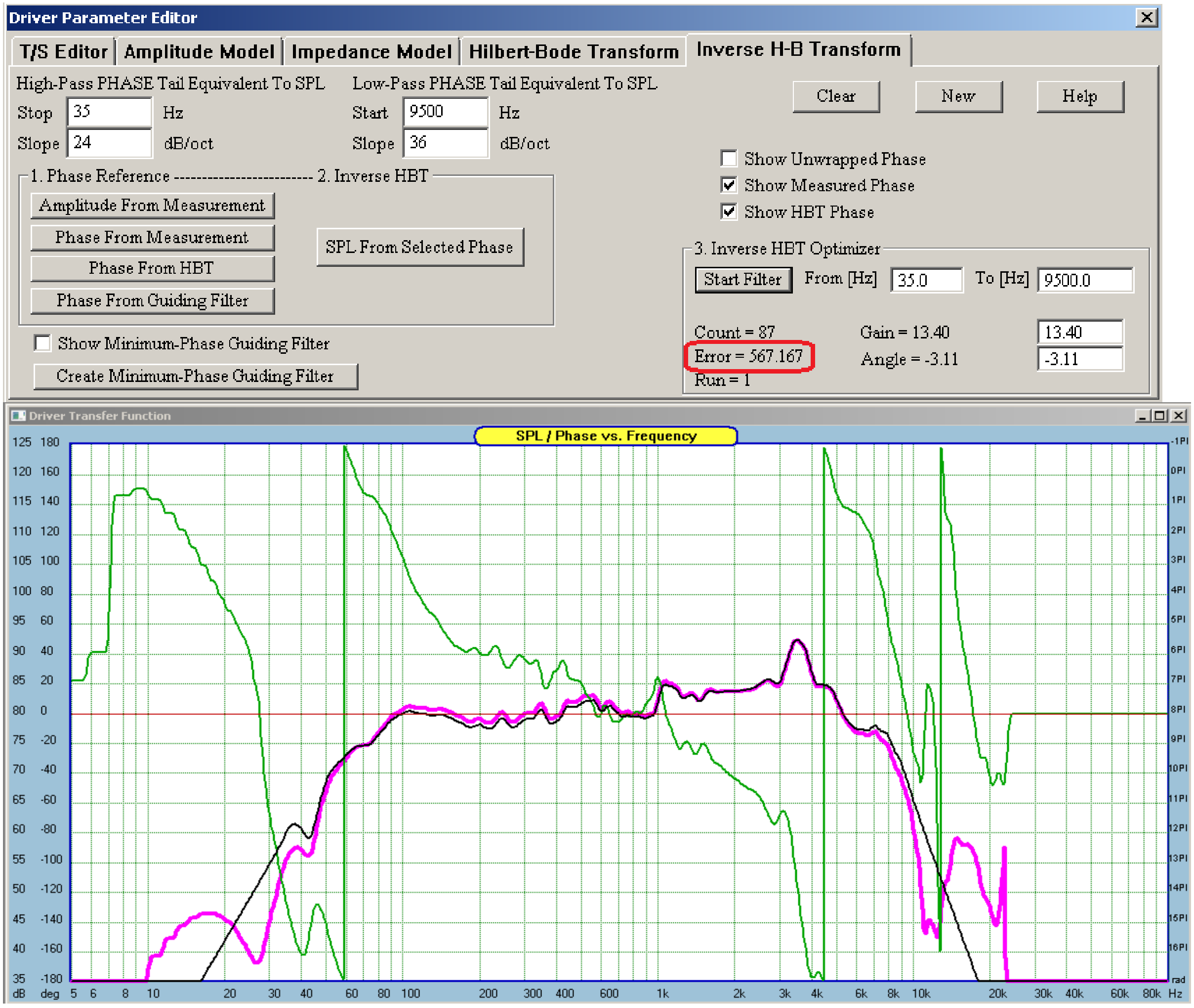
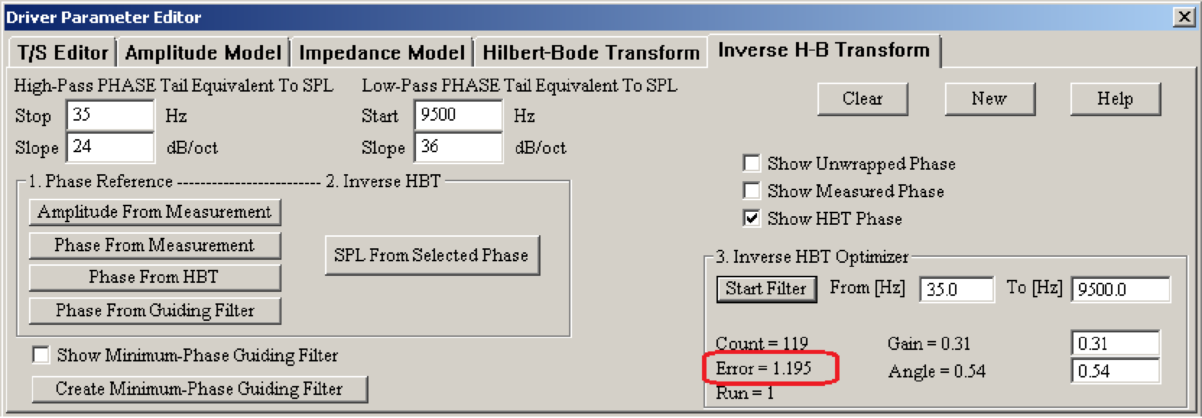


Figure 16.234 MLS measurement of SPL/Phase of a 12” loudspeaker with FFT window at Bin 78.

SPL from HBT Phase

However, when the measured SPL was used to calculate minimum-phase phase response, and this phase response was subsequently supplied to IHBT algorithm, the Error = 1.195 (very small) SPL does match the measured SPL from 35Hz – 9.5kHz



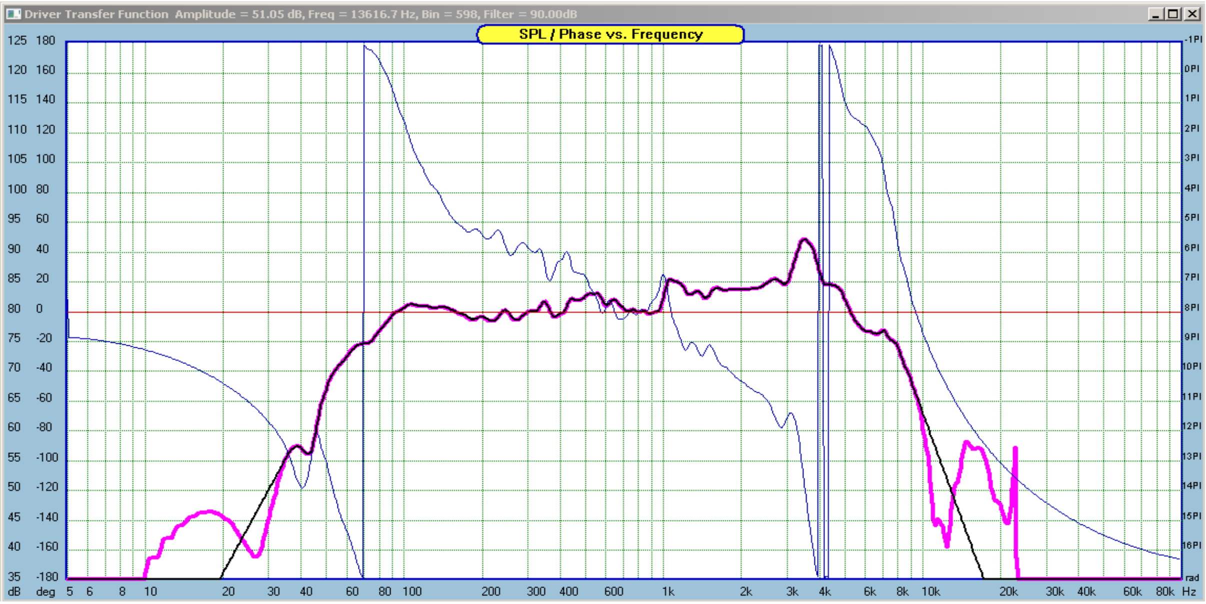


Figure 16.235 SPL of a 12" loudspeaker with Phase provided by HBT.

The Error is now 346 times smaller when the SPL is recovered from guaranteed minimum-phase response.

SPL sensitivity to attached phase slopes.

On minimum-phase systems, the accuracy of recovered SPL response can be tested by generating various phase responses and supplying such phase responses to the IHBT algorithm. In order to assure that all phase responses are indeed of the minimum-phase type, firstly, the phase response was generated by the HBT process. Then, resulting phase slopes corresponding to -18dB/oct and then -68dB/oct were attached at 9500Hz. Next, those two phase responses were supplied to the IHBT algorithm. It would be expected, that SPL would only change beyond the attachment point, as the two different phase slopes would guide the IHBT algorithm to produce corresponding SPL slopes.

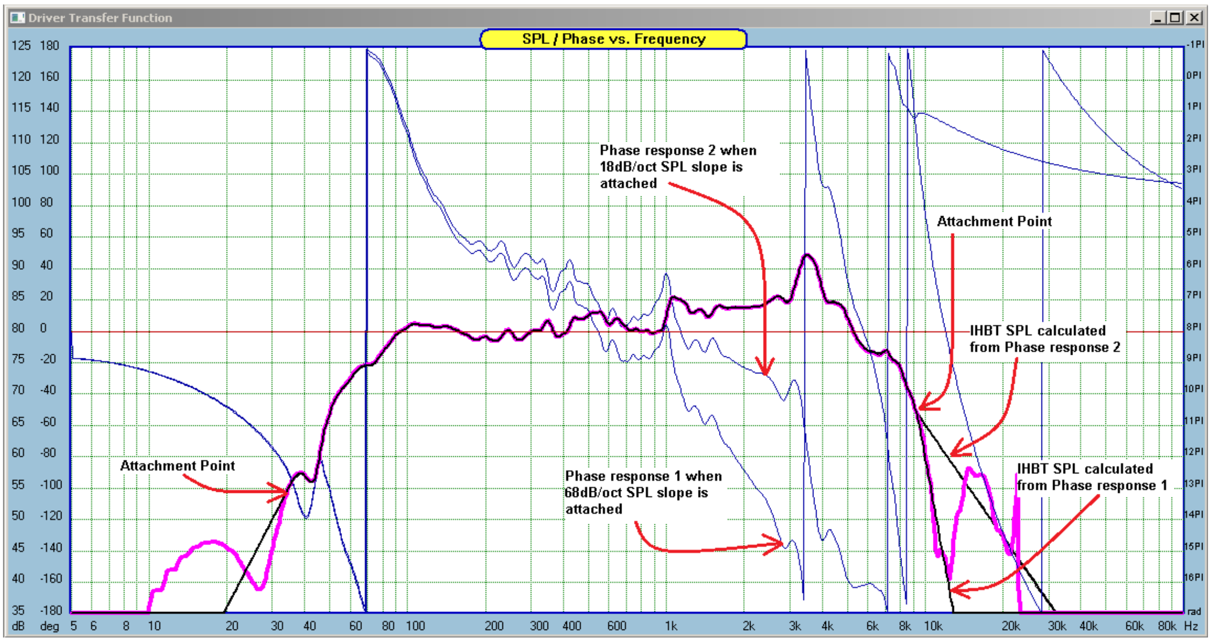


Figure 16.236 SPL extracted from HBT phase with attached slopes of -18dB/oct and 68dB/oct.

6

Indeed, for the example above, the LP slope attachment point is 9500Hz. For slopes of -18dB/oct to -68dB/oct – **NO change in extracted SPL between attachment points can be observed, even though the phase responses were dramatically different**. Similar SPL recovery accuracy can be shown for variety of SPL slopes attached at low-frequency attachment point of 35Hz. This is very important observation, as it shows, that we can use the degree of SPL match between the attachment points as the “detector” of minimum-phase characteristics, with complete disregard of the attached phase slopes. In other words, we can get the slopes wrong, but as long as the supplied phase is of the minimum-phase type, the SPL extracted between the attachment points will be perfect. Frequency range of the IHBT curve fitting mechanism should be selected to be the same as the phase slopes attachment points. As shown above, if the supplied phase response is of minimum-phase type, then the SPL will be always extracted accurately between the phase attachment points, and across the whole frequency spectrum, including the slopes. However, for the purpose of using the IHBT as a “minimum-phase detector”, we are interested in evaluating the degree of SPL matching (the Error value) between the attachment points. The curve fitting algorithm manipulates two additional parameters: Gain and Angle. It is not possible to uniquely define the amplitude response from the phase data, since an infinite number of amplitude characteristics, differing only be a fixed number of dB, have identical phase response. This is where the “Gain” parameter comes in. The “Angle” parameter is related to the way mathematical functions (like arctan(x)) are calculated on the computer.

SPL Sensitivity to Non-minimum-phase Phase Response

This a critical question for the IHBT algorithm. How well can the IHBT discriminate between minimum-phase and non-minimum-phase data.

Measurement procedure for the example woofer.

1. Measure SPL/Phase and set FFT window to 77 bin
2. On IHBT tab and select Stop = 35Hz, Slope = 24dB/oct and Start = 9500Hz, Slope = 36dB/oct
3. Select frequency range of IHBT optimization 200 – 9500
4. Press “Phase From Measurement”-> “SPL From Selected Phase”.
5. Start IHBT Optimizer with: Gain = 0.0, Angle = 0.0, **The Error = 32.579**
6. Go to MLS tab and replot SPL/Phase for 76bin
7. Go back to IHBT and press “Phase From Measurement”-> “SPL From Selected Phase”.
8. Start IHBT Optimizer. **The Error = 216.77**
9. Go to MLS and replot SPL/Phase for 78bin
10. Go back to IHBT and press “Phase From Measurement”-> “SPL From Selected Phase”.
11. Start IHBT Optimizer. **The Error = 35.225**

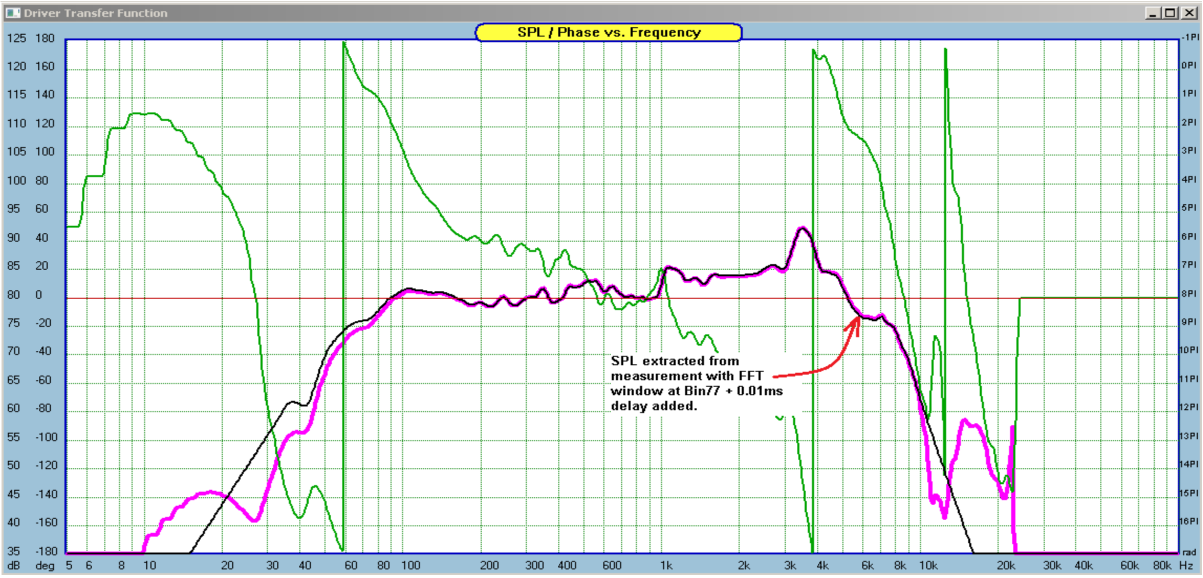


Figure 16.237 SPL extracted from measured phase with FFT window at Bin77+ 0.01ms delay.

7

Using FFT window at Bin 77, which already produced the lowest Error = 32.579, one can attempt to further reduce the error by going back to MLS system and adding small time delays. Subsequent trial errors were as follows:

Bin 78 -> Error = 35.225
Bin 77 -> Error = 32.579
Bin 77 + 0.008ms Error = 11.376
Bin 77 + 0.01ms Error = 10.547
Bin 77 + 0.012ms Error = 12.445
Bin 77 + 0.015ms Error = 15.969
Bin 76 -> Error = 216.777

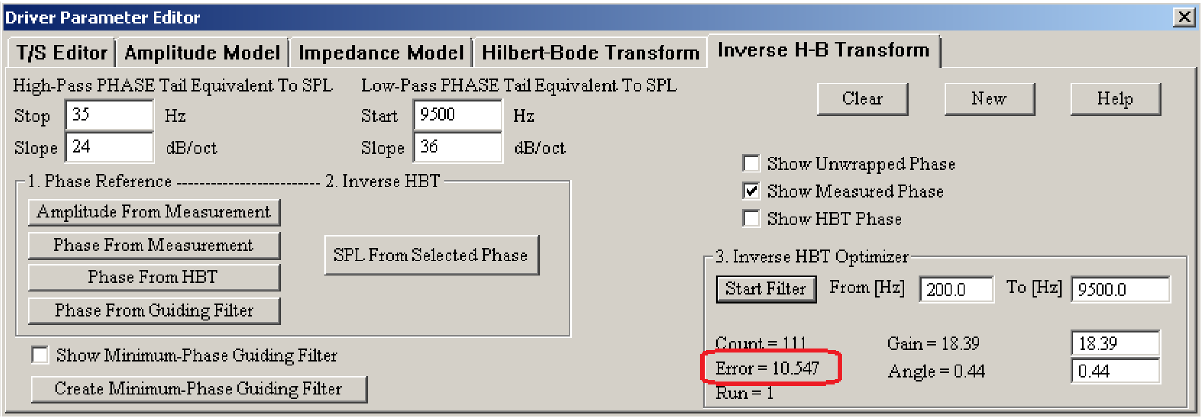


Figure 16.238 SPL extracted from measured phase with FFT window at Bin77+ 0.01ms delay.

In the example above, the effort was concentrated on the high-frequency range of the SPL/Phase curves. Therefore the optimization frequency range was set to 200-9500Hz. The IHBT parameters for phase slopes were set as for a vented box for the low-end, and it was demonstrated above, that attaching various phase slopes at high frequency end, does not affect the SPL response below the attachment point. Therefore an arbitrary slope of 36dB/oct was used for the high-end. With the above parameters, it was demonstrated, that IHBT is NOT sensitive to attached phase slopes, and will produce the same SPL curve between attachment points for a wide range of various phase slopes selected at the attachment points.

It was also demonstrated, that IHBT algorithm is quite sensitive to non minimum-phase distortions and will <u>easily detect phase errors introduced by single FFT bin shift</u>. Even more, in the example above, the IHBT combined with curve fitting algorithm easily detected of non-minimum-phase 1-2usc delay, which is 10 times better than single bin shift of 20.833usec at 48kHz sampling.

How does all of the above help with determining the minimum-phase phase response of the loudspeaker?.

It was demonstrated, that using IHBT with the final measured phase response and FFT window set to Bin 77+0.01ms delay generates the most accurate SPL above 1kHz. This would be indicative, that supplied phase response is of minimum-phase type above 1kHz, and we have removed the non-minimum-phase component (time-of-flight). All we need to do now, is to select HBT slope that matches the final measured phase up to the attachment point of 9500Hz. Turns out, that changing the Slope from 36dB/oct to 42dB/oct makes the measured and HBT-generated phase match quite well above 1kHz.

But what about the phase below 1kHz?.

Measured phase response at low-frequency end will be affected by the length and type of the FFT window used. Therefore, the phase will deviate from the minimum-phase trajectory. To avoid this issue, one could use really wide FFT windows (if the measurement environment allows it) of use close-mike techniques.

In summary, it is suggested, that the phase response on the figure below (blue line) is the accurate representation of minimum-phase response of the measured loudspeaker.

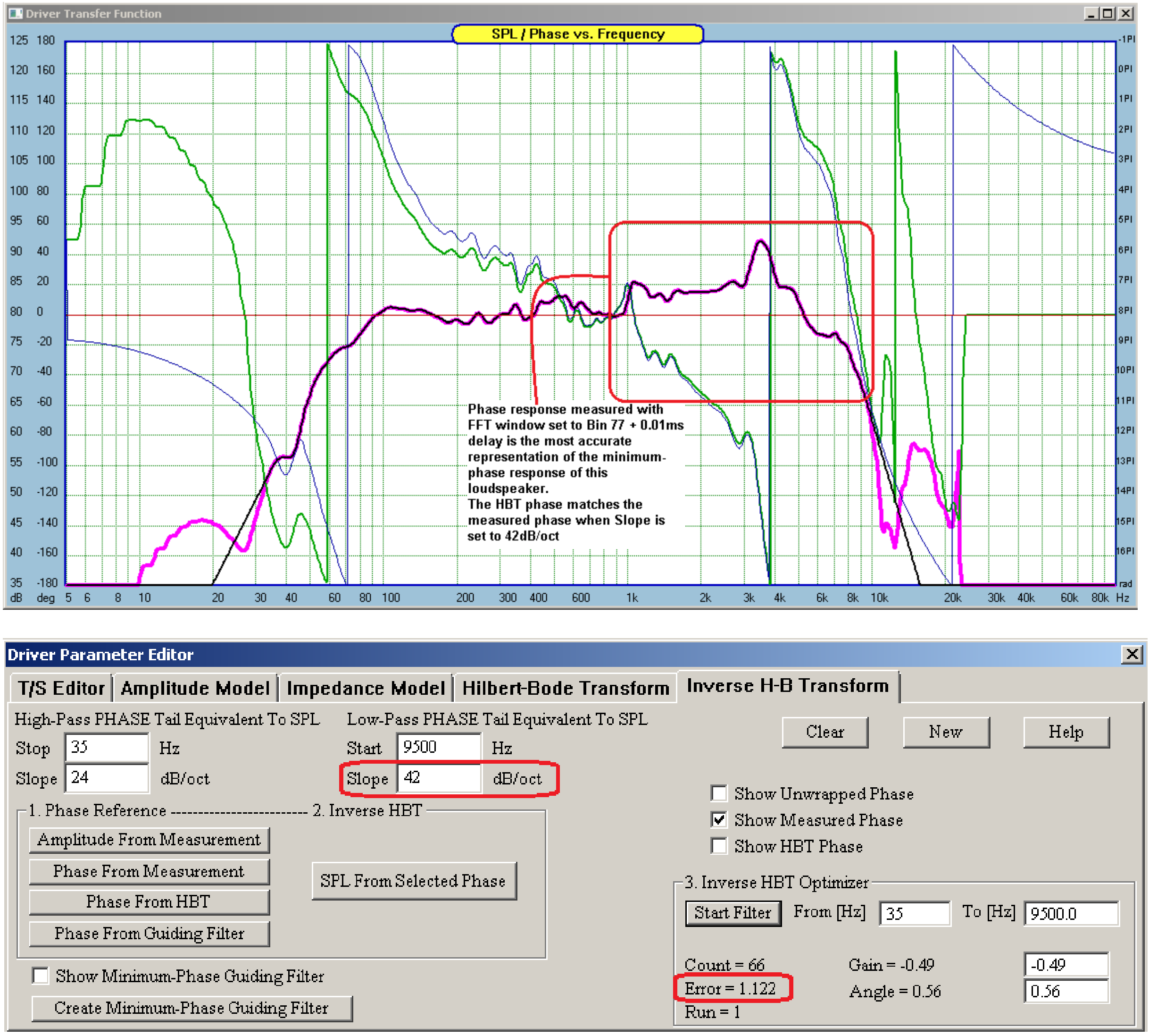


Figure 16.239 Final minimum-phase phase response – blue curve.

Conclusions

SPL can only be extracted accurately by IHBT, if the supplied phase is of minimum-phase type.

Between the phase attachment points, the IHBT is insensitive to attached phase (SPL) slopes, as the SPL does not change between attachment points.

Between the phase attachment points, the IHBT is quite sensitive to non minimum-phase phase response, behaviour, as the SPL does change quite significantly between attachment points.

The IHBT can be used as a “detector” to determine if the phase response is indeed the sought-after minimum-phase response.

Peak Power Dissipation Ratio

Peak power dissipation in amplifier output stages is significantly greater with loudspeaker loads than with “ideal” resistor loads. This is quite unfortunate and is due to real loudspeaker loads are typically quite reactive. Typical increase in power dissipation was found to be 150-200% and peaked at 270%. Functionality described in this chapter allows you to evaluate how much bigger power dissipation really is when compared to an ideal reference resistor.

The resistor is selected from ”CAD Frequency/Time Domain Scale” dialogue box – as shown below.

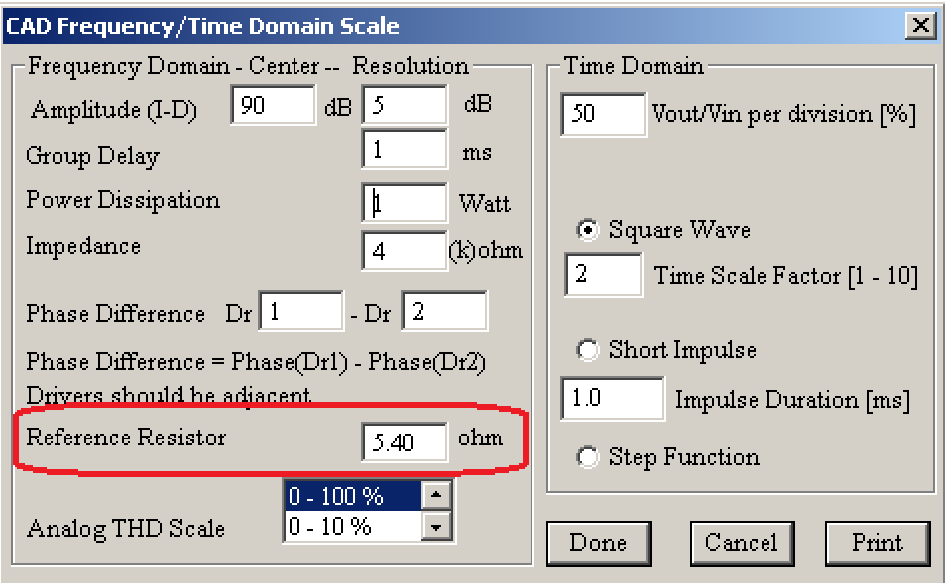


Figure 16.240 Selecting the Reference Resistor value for PPDR function.

A simple example of a tweeter driver input impedance (Zin) graphs explain the PPDR functionality. Here we have Zin (blue) and Zin phase (pink) plots versus frequency. It is observable, that phase response crosses zero in two points: 650Hz and 10kHz. At those two spots only, Zin is purely resistive. Everywhere else Zin is reactive, and one would expect that power dissipation ratio would be greater than that of a pure resistor R=5.4ohms. Indeed, the PPDR curve (green) exceeds unity in all other places, except in two spots, where the phase is zero. At those points, the PPDR value depends on the value of |Zin|. At 650Hz, the |Zin| is nearly 70 ohms, so the when referenced to 5.4ohm resistor, the PPDR value becomes quite small. At 10kHz, the |Zin| is around 5.4ohm, therefore the PPDR value referenced to 5.4ohm resistor is around 1.

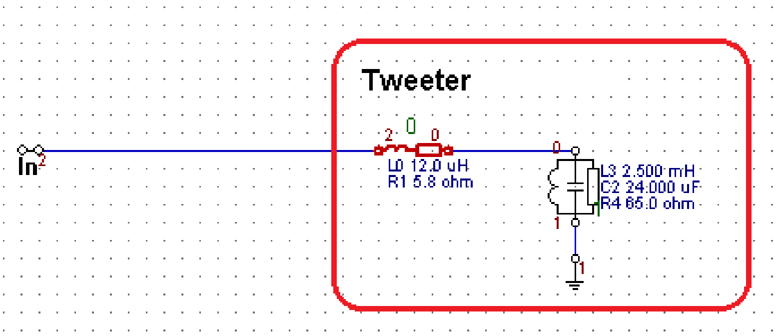


Figure 16.241 Tweeter Zin model.

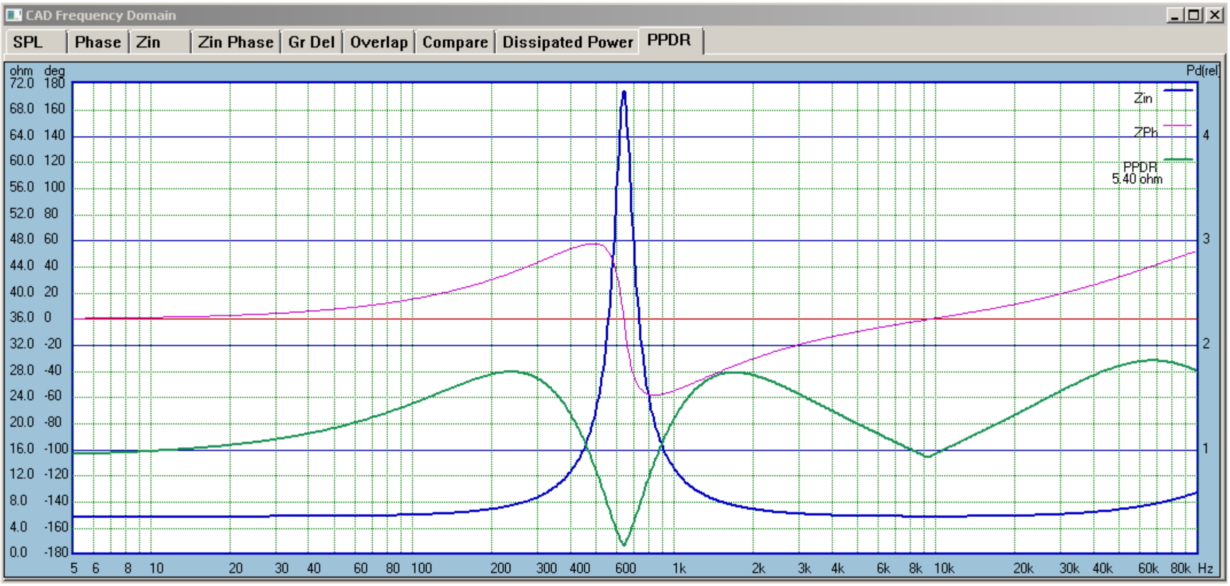
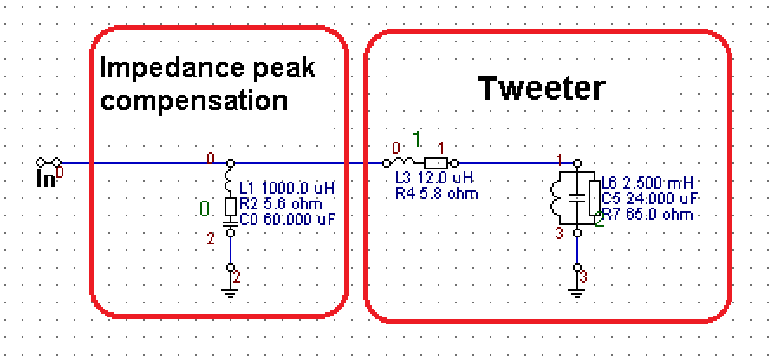


Figure 16.242 PPDR function referenced to 5.4ohm resistor.

The PPDR can be improved (reduced) by simply applying impedance compensation circuit to flatten the Zin curve. A crudely designed series resonant circuit is used to perform this function, and indeed, both Zin and Zin phase are flattened significantly, resulting in PPDR curve revolving around unit up until 5kHz – see figure below.



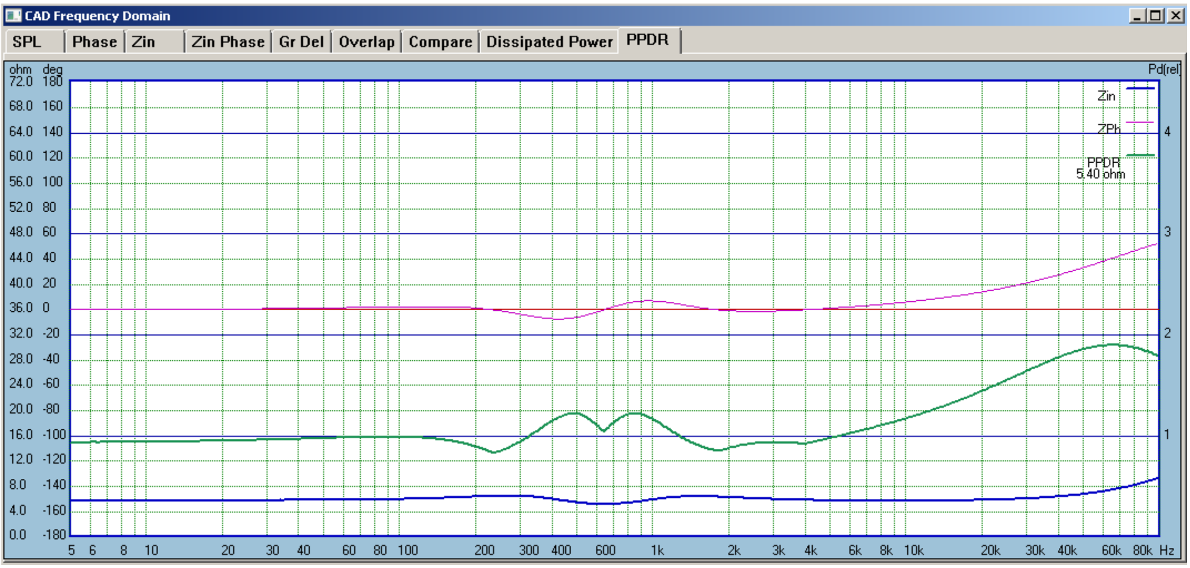


Figure 16.243 Impedance peak compensation for tweeter.

In the next example we examine an impedance model of a two-way speaker system with a crossover and impedance compensation circuits. This is Zobel network for woofer and an impedance compensation circuit for tweeter.

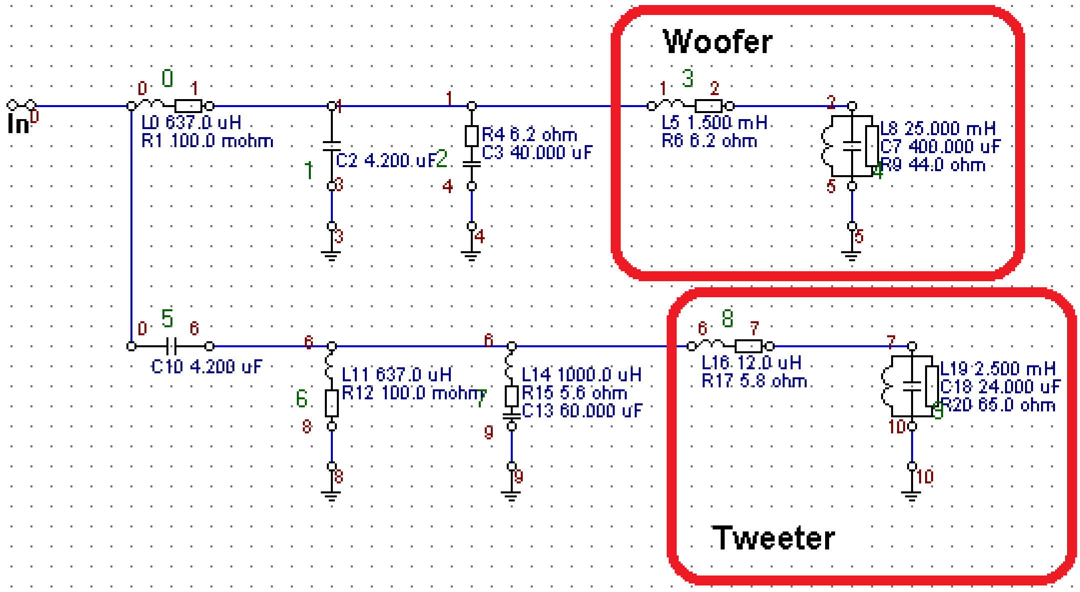


Figure 16.244. Impedance model of a two-way loudspeaker system.

Corresponding Zin, Zin phase and PPDR plots are shown on the Figure 16.245 below.

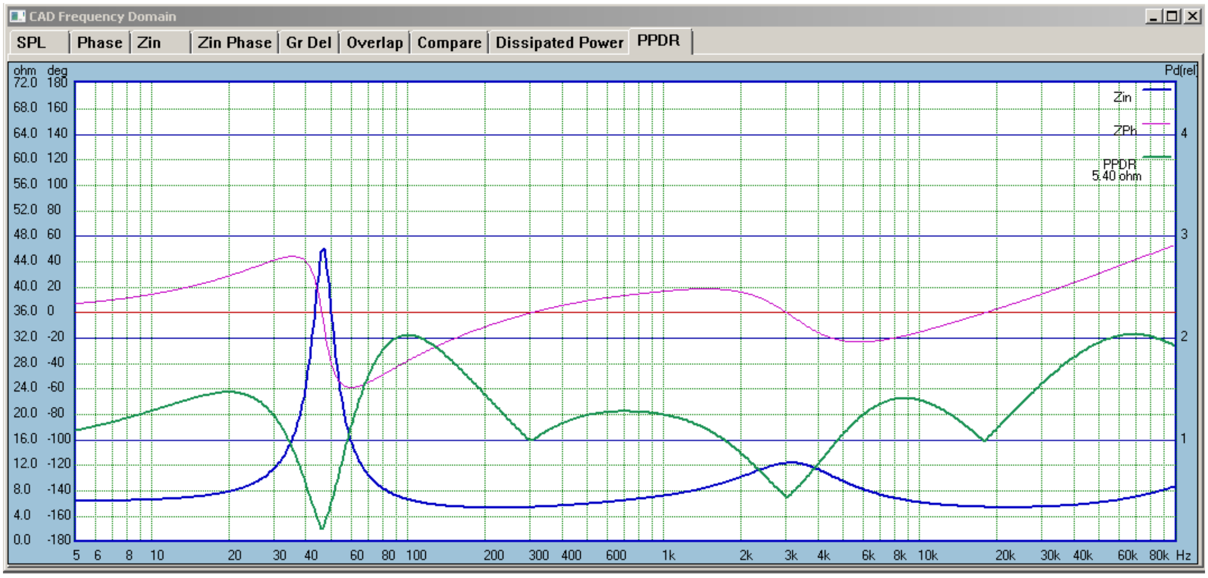


Figure 16.245 Zin, Zin phase and PPDR plots of a 2-way system.

The PPDR curve is valid for sine-wave excitation signals. It is observable from the above lots, that impedance compensation circuits are very effective in reducing power dissipation of the amplifiers.

Subtracting diffraction from SPL

At times, you may need to actually subtract diffraction curve from the measured SPL curve. A typical diffraction curve is shown on Figure 16.246 below

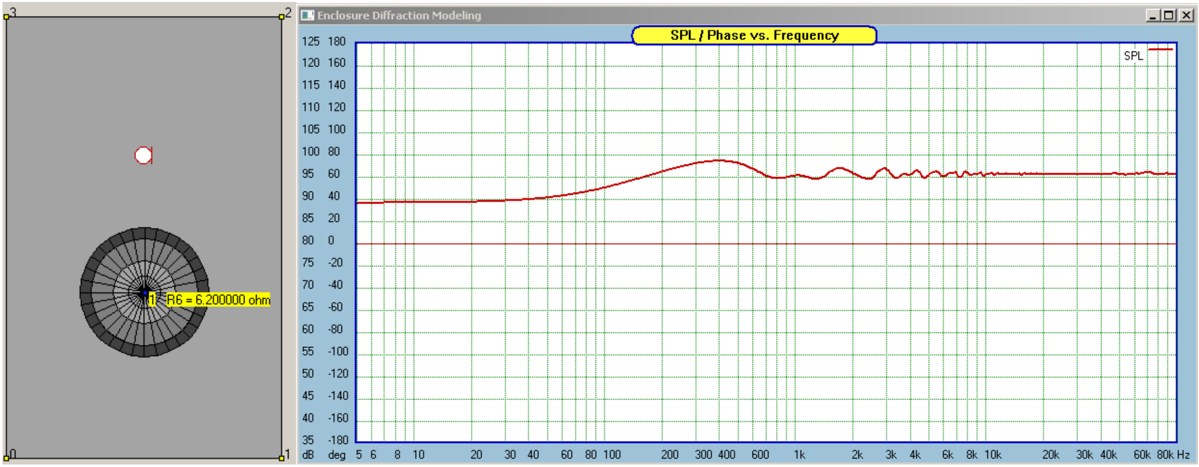


Figure 16.246 Example of a diffraction curve for the baffle shown on the left.

Diffraction subtraction is accomplished in the "Post-Processing" TAB of the MLS and ESS measurement systems. Figure below shows an example diffraction subtraction with measured SPL curve send to Buffer 5. You can test the subtraction process by executing the opposite function, "Add Diffraction" to have the now diffraction-less curve returned to its original shape.

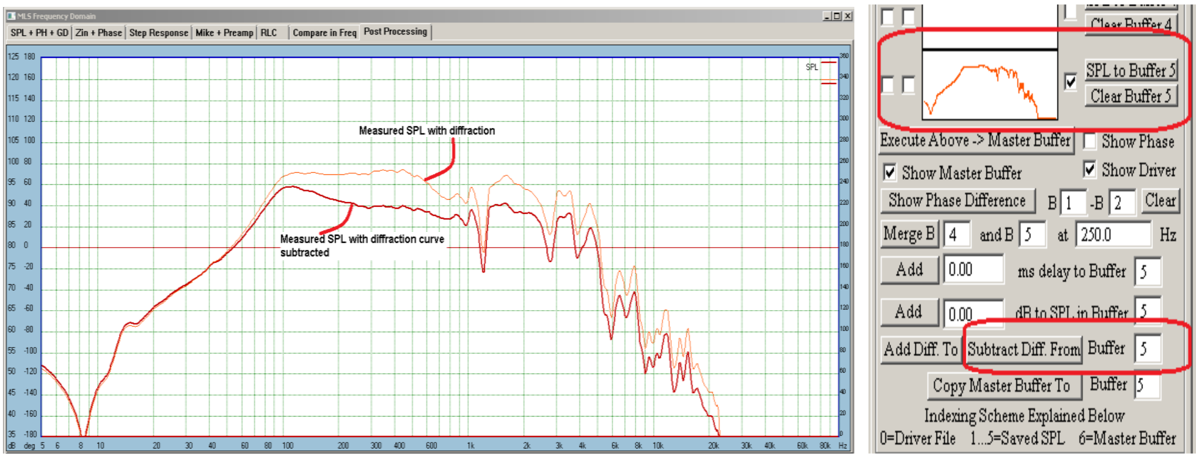


Figure 16.247 Example of subtracting diffraction from SPL in Buffer 5.

Saving MLS settings to Hard Disk

Saving/Loading the entire settings of MLS or ESS measurement systems to or from the HD may speed up your measurement system settings. In order to Save/Load the setting, please right-click the mouse button above the "MLS/ESS" TABs, and a small floating menu will pop up.

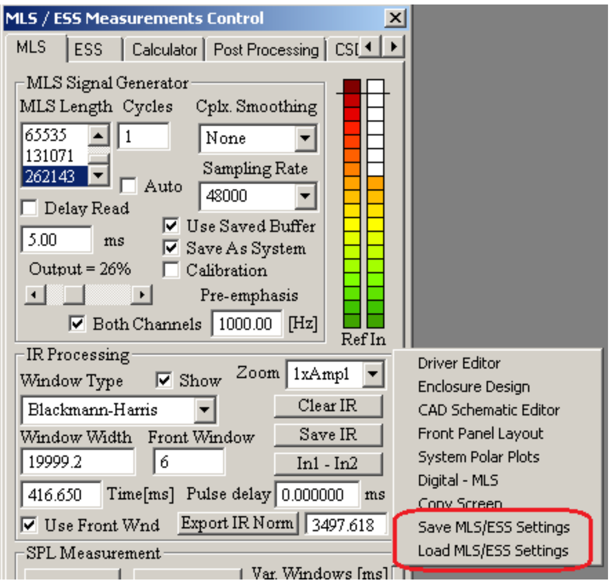


Figure 16.248. Right-click the mouse button above the “MLS/ESS” TABs

You can then select what function you wish to accomplish and then a standard file saving/loading dialog box will allow you to select the required file with MLS/ESS settings.

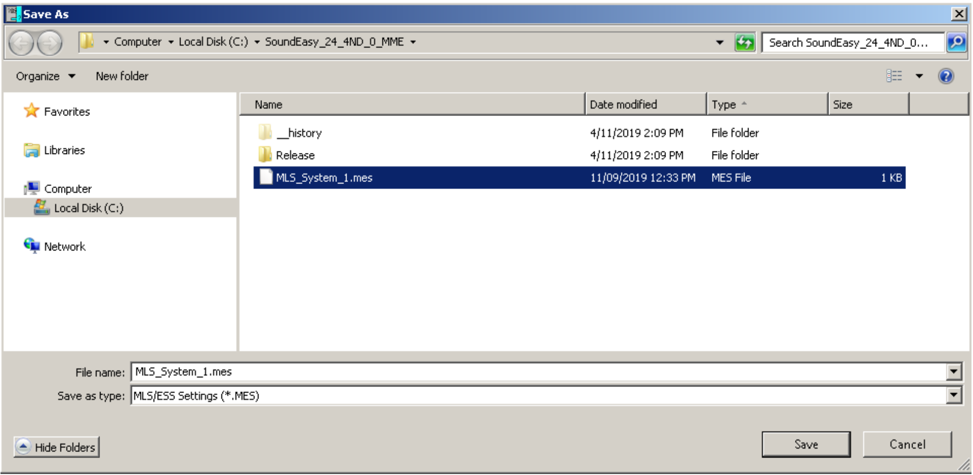


Figure 16.249. Saving/loading dialog box