

# **Measurement of Total Harmonic Distortion (THD) and Its Related Parameters using Multi-Instrument**

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Rev: 01  
June 8, 2020

This article presents how to measure Total Harmonic Distortion (THD) and its related parameters correctly using Multi-Instrument. Sophisticated mathematics is intentionally avoided in this article in order to make it easily understood by most software users.

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# 1. Introduction

## 1.1 Distortion Classification

Simply put, distortion is the alteration of the waveform of a signal when it passes through a system. Distortions can be classified as either linear or non-linear distortions.

### 1.1.1 Linear Distortion

A linear distortion is a distortion with no new frequencies added. It can be caused by the non-flat magnitude frequency response or the non-linear phase frequency response of the system.

### 1.1.2 Non-linear Distortion

A non-linear distortion is a distortion with new frequencies added. It can be classified as either harmonic or non-harmonic. Non-linear distortions are sometimes simply called “distortions”.

There are other ways to describe non-linear distortion, such as non-coherent distortion and GedLee Metric, etc.

#### 1.1.2.1 Harmonic Distortion

Harmonic distortions refer to those newly added frequencies which are integer multiples of the fundamental frequency.

#### 1.1.2.2 Non-Harmonic Distortion

Non-harmonic distortions refer to those newly added frequencies which are not integer multiples of the fundamental frequency, such as intermodulation distortion. Note that noises are not considered as non-harmonic distortions.

For more information on intermodulation distortion measurement, please refer to the article “Measurements of Various Intermodulation Distortions (IMD, TD+N, DIM) using Multi-Instrument” at <http://www.virtins.com/doc/Measurements-of-Various-Intermodulation-Distortions-IMD-TD+N-DIM-using-Multi-Instrument.pdf>.

## 1.2 Overview

This article describes the measurements of harmonic distortion related parameters, including THD, THD+N, SINAD, SNR, ENOB, NL, SFDR. The measurements of other types of distortion will be introduced elsewhere. THD (Total Harmonic Distortion) measurement is required in many fields such as audio, acoustics, power supply and vibration. Among them, the audio industry generally requires the highest measurement accuracy.

In Multi-Instrument, THD measurement mode can be selected by right clicking anywhere within the Spectrum Analyzer window and selecting [Spectrum Analyzer Processing]> “Parameter Measurement”> “THD, THD+N, SINAD, SNR, NL”. SFDR is measured under “Parameter Measurement”> “Peaks” option.

## 2 Definitions of THD、THD+N、SINAD、SNR、 ENOB、NL、SFDR

The definitions of THD、THD+N、SINAD、SNR、ENOB have some variants. Only the commonly used ones are provided below. When measuring these parameters, a signal generator is employed to generate a sine wave at a certain frequency with sufficiently low distortions. The generated signal is used as the stimulus to the Device Under Test (DUT). Meanwhile, the response from the DUT is sampled and then analyzed using Fast Fourier Transform (FFT). The signal power is decomposed into three parts: fundamental, harmonics and noise. The DC component is usually filtered out and not used in the calculation. Finally, these parameters are calculated based on their definition formulae, as shown below.

### 2.1 THD (Total Harmonic Distortion)

THD is one of the commonly used parameters to characterize amplifiers, ADC / DAC devices, sensors, transducers, and power supplies. It is usually defined as the square root of the ratio of the sum of the powers of all harmonic frequencies to the power of the fundamental, expressed in percentage:

$$\text{THD} = \frac{\sqrt{\sum_{i=2}^N V_i^2}}{V_1} \times 100\%$$

where  $V_i$  is the RMS amplitude of the  $i$ -order harmonic,  $V_1$  the RMS amplitude of the fundamental,  $N$  the order number of the highest harmonic used in the THD calculation.

It can also be expressed in dB:

$$\text{THD}_{\text{dB}} = 20\log_{10}(\text{THD})$$

For example, if  $\text{THD} = 0.0001\%$ , then  $\text{THD}_{\text{dB}} = -120$  dB. THD is an indication of the overall harmonic distortion. All harmonics have equal weights in THD. THD does not differentiate amplitude clipping and cross-over distortions. A THD rating must be accompanied by its testing conditions such as testing frequency (the fundamental), amplitude, and frequency range or the highest harmonic order used in the calculation.

### 2.2 THD+N (Total Harmonic Distortion Plus Noise)

THD+N is usually defined as the square root of the ratio of the sum of the powers of all harmonic frequencies plus noise to the total power. In other words, it equals to the square root of the ratio of the total power less the power of the fundamental frequency to the total power, expressed in percentage:

$$\text{THD} + \text{N} = \frac{\sqrt{V_{\text{total}}^2 - V_1^2}}{V_{\text{total}}} \times 100\%$$

where  $V_{\text{total}}$  is the RMS amplitude of the signal (including the fundamental, all harmonics, and noise),  $V_1$  the RMS amplitude of the fundamental.

It can also be expressed in dB:

$$(\text{THD+N})_{\text{dB}} = 20\log_{10}(\text{THD+N})$$

For example,  $\text{THD+N} = 0.0001\%$ , then  $(\text{THD+N})_{\text{dB}} = -120$  dB. A THD+N rating must be accompanied by its testing conditions such as testing frequency (the fundamental), amplitude, frequency range, as well as whether A, B, C, ITU-R 468 or ITU-R ARM frequency weighting is used.

## 2.3 SINAD (Signal to Noise and Distortion Ratio)

SINAD is usually defined as the ratio of the total power to the total power less the power of the fundamental frequency, expressed in dB:

$$\text{SINAD} = 20\log_{10}\left(\frac{V_{\text{total}}}{\sqrt{V_{\text{total}}^2 - V_1^2}}\right) \quad (\text{dB})$$

where  $V_{\text{total}}$  is the RMS amplitude of the signal (including the fundamental, all harmonics, and noise),  $V_1$  the RMS amplitude of the fundamental.

## 2.4 SNR (Signal to Noise Ratio)

SNR is usually defined as the ratio of the power of the fundamental frequency to the power of the noise, expressed in dB:

$$\text{SNR} = 20\log_{10}\left(\frac{V_1}{\sqrt{V_{\text{total}}^2 - \sum_{i=1}^N V_i^2}}\right) \quad (\text{dB})$$

where  $V_{\text{total}}$  is the RMS amplitude of the signal (including the fundamental, all harmonics, and noise),  $V_i$  is the RMS amplitude of the  $i$ -order harmonic,  $N$  the order number of the highest harmonic used in the SNR calculation.

## 2.5 ENOB (Effective Number of Bits)

ENOB can be derived from SINAD according to:  $\text{ENOB} = (\text{SINAD} - 1.76 \text{ dB}) / 6.02$ . This formula should be used only if the amplitude of the test signal reaches the full-scale of the ADC / DAC device under test. Otherwise, the SINAD and ENOB will be biased towards a smaller value, in which case the above ENOB should be corrected by adding  $1/6.02 \times 20\log_{10}\left(\frac{\text{Full-Scale Amplitude}}{\text{Signal Amplitude}}\right)$ .

## 2.6 NL (Noise Level)

NL is defined as the RMS amplitude of the noise.

## **2.7 SFDR (Spurious Free Dynamic Range)**

SFDR is defined as the ratio of the power of the highest peak to that of the second highest peak, expressed in dB. Please note that the second highest peak is not necessarily a harmonic of the highest one.

The above formulae are not so complicated. It may look like a simple task to obtain THD and its related parameters using FFT. However, things can easily go wrong if the sampling and analyzing parameters such as test frequency, sampling rate, sampling bit resolution, record length, FFT size, window function, frequency range are not set properly. How to configure these parameters correctly in order to get meaningful and accurate results as well as the reasons behind will be explained in the following sections.

## 3 How to Avoid or Suppress Spectral Leakage

### 3.1 What is Spectral Leakage?

To obtain a correct THD result, one common problem needs to be solved, that is, spectral leakage. The signal to be analyzed must be truncated and put into a FFT segment with a limited length in time. Spectral leakage is the result of the inherent assumption in the FFT algorithm that the time record of the signal in a FFT segment is exactly repeated throughout all time and that signal contained in a FFT segment is thus periodic at intervals that correspond to the length of the FFT segment. This is called periodic extension. If the time record in a FFT segment contains an integer number of cycles of the signal, then the periodically extended signal will be the same as the un-truncated original periodic signal, otherwise discontinuities will show up at the periodic extension boundaries, resulting in the spread of the energies of the periodic components to their adjacent frequencies. This phenomenon is called spectral leakage. Spectral leakage, if handled incorrectly, will cause underestimation of the fundamental and harmonics and overestimation of noises. For more information on FFT, please refer to:

[https://www.virtins.com/doc/D1002/FFT\\_Basics\\_and\\_Case\\_Study\\_using\\_Multi-Instrument\\_D1002.pdf](https://www.virtins.com/doc/D1002/FFT_Basics_and_Case_Study_using_Multi-Instrument_D1002.pdf)

The following figure shows the spectrum of an ideal 1 kHz sine wave with severe spectral leakage. The ideal signal is generated using the “iA=oA, iB=oB” software loopback mode of the Signal Generator of Multi-Instrument. The test parameters are: [Sampling Rate] = 48 kHz, [Signal Frequency] = 1 kHz, [FFT Size] = 32768, [Number of Cycles] =  $1000 / 48000 \times 32768 = 682.6667$ , [Sampling Bit Resolution] = 24, [Window Function] = Rectangle (i.e. no window), [Frequency Range] = 20 Hz ~ 20 kHz. No zero padding is applied as [Record Length] = 48000 which is greater than the FFT size. As the FFT segment contains a non-integer [Number of Cycles], severe spectrum leakage occurs. It can be seen from the figure that all the harmonics are submerged under the leaked energy from the fundamental. The measured results are: THD = 0.1664 % (-55.58 dB), THD+N = 14.2383% (-16.93 dB), SINAD = 16.93 dB, SNR = 16.84 dB, ENOB = 2.52 Bit. Obviously, these results are completely wrong for a 24-bit ideal sine wave.



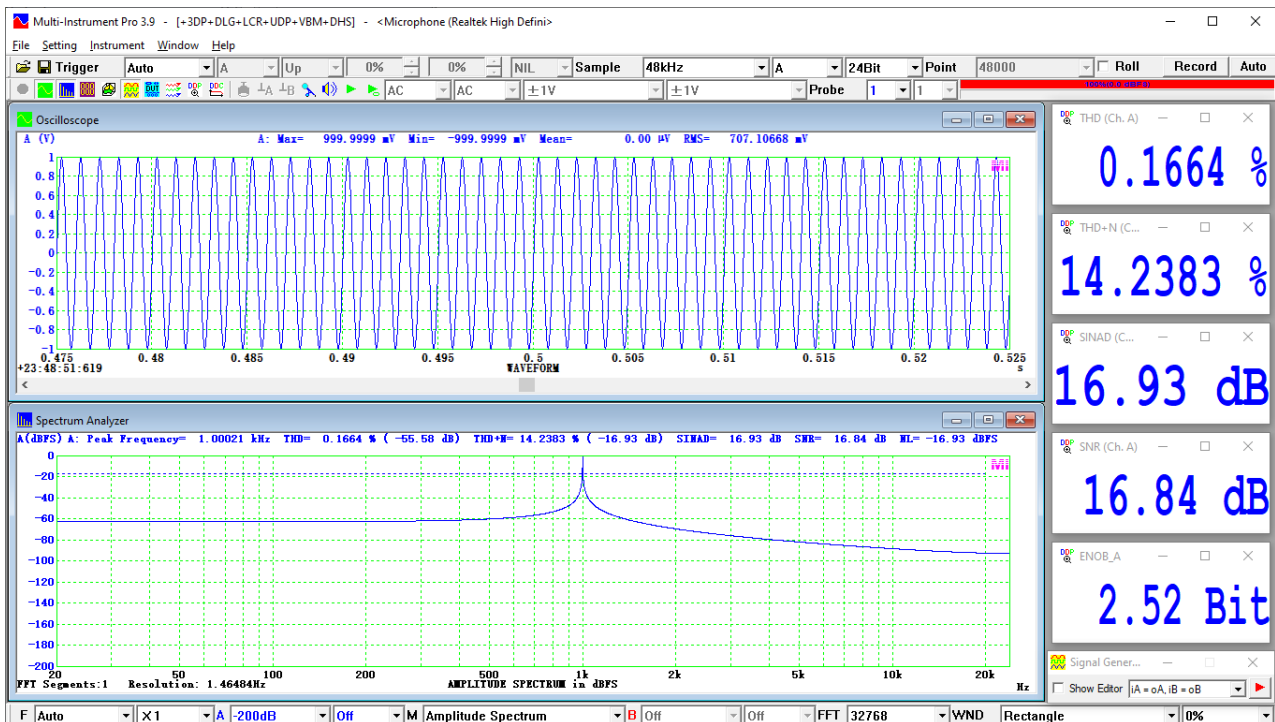


Fig. 1 Non Full-cycle Sampling with a Rectangle Window Leading to Severe Spectral Leakage

## 3.2 Spectral Leakage Solutions

### 3.2.1 Full-Cycle Sampling / Coherent Sampling

To avoid spectral leakage, a FFT segment must contain an integer number of signal cycles. This can be expressed in math as:

$$[\text{Sampling Rate}] / [\text{Signal Frequency}] = [\text{FFT Size}] / [\text{Number of Cycles}]$$

Sampling that satisfies the above equation is called full-cycle sampling or sometimes loosely called coherent sampling. The more stringent definition of coherent sampling requires  $[\text{FFT Size}] / [\text{Number of Cycles}]$  to be irreducible to ensure identical phases not to be sampled repeatedly in a FFT segment. Unnecessary repetition of samples at the same phase increases the test time and yields periodic quantization error (explained later). As the FFT size is a power of 2, the number of cycles must be odd in order to meet the stringent coherent sampling requirement.

The following figure shows the spectrum of an ideal 1 kHz sine wave with no spectral leakage using coherent sampling. The test parameters are:  $[\text{Sampling Rate}] = 48 \text{ kHz}$ ,  $[\text{Signal Frequency}] = 1000.48828125 \text{ Hz}$  (ideal signal generated using the Signal Generator of Multi-Instrument),  $[\text{FFT Size}] = 32768$ ,  $[\text{Number of Cycles}] = 1000.48828125 / 48000 \times 32768 = 683$ ,  $[\text{Sampling Bit Resolution}] = 24$ ,  $[\text{Window Function}] = \text{Rectangle}$  (i.e. no window),  $[\text{Frequency Range}] = 20 \text{ Hz} \sim 20 \text{ kHz}$ . No zero padding is applied as  $[\text{Record Length}] = 48000$  which is greater than the FFT size. As the FFT segment contains an integer  $[\text{Number of Cycles}]$ , no spectrum leakage occurs. The measured results are:  $\text{THD} = 0.0000032 \%$  ( $-149.80 \text{ dB}$ ),  $\text{THD+N} = 0.0000055 \%$  ( $-145.19 \text{ dB}$ ),  $\text{SINAD} = 145.19 \text{ dB}$ ,  $\text{SNR} = 147.04 \text{ dB}$ ,  $\text{ENOB} = 23.83 \text{ Bit}$ . These results represent more or less the best measurable values under 24-bit coherent sampling. They are far better than those of a HIFI device. This proves, from software point of view, that the above test parameters can be used in the measurement of a HIFI device. One question remains: why do distortions and noises still

show up in an ideal sine wave? It has something to do with the quantization and numerical calculation errors which will be discussed later.

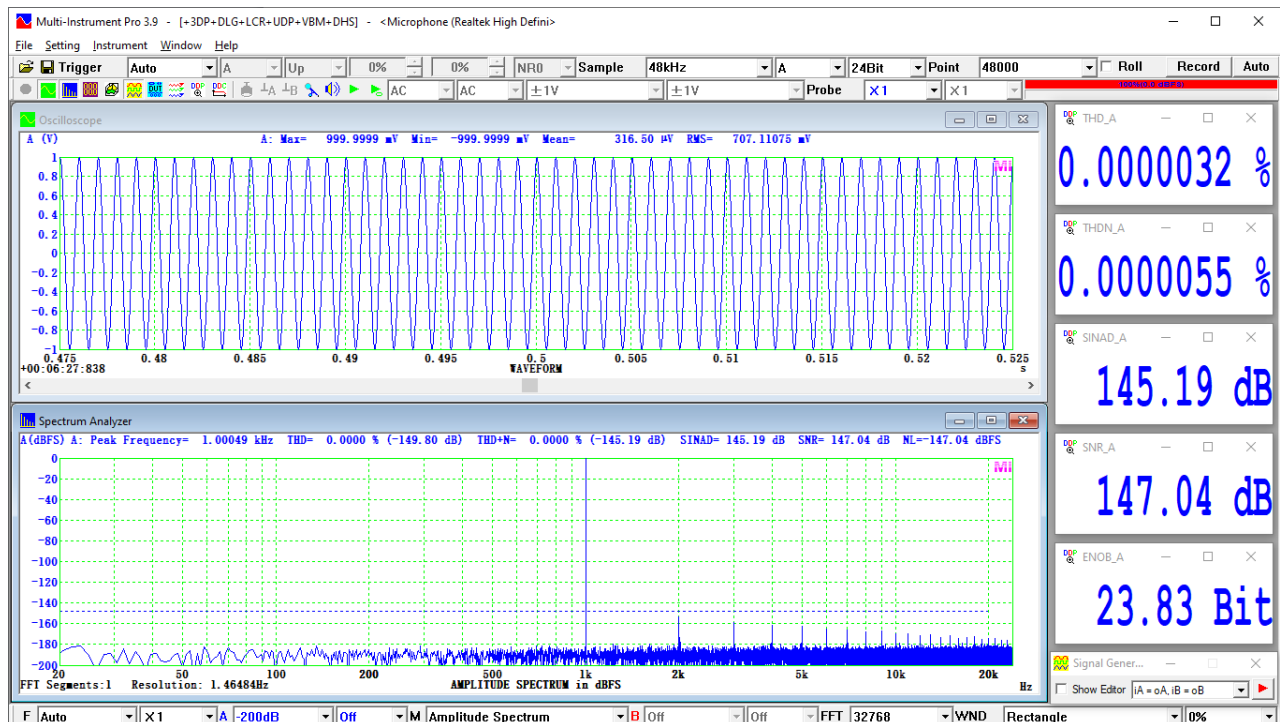


Fig. 2 Coherent sampling with a Rectangle window leading to no spectral leakage

It should be noted that zero padding should not be used in full-cycle and coherent sampling.

### 3.2.2 Window Sampling

If the conditions for full-cycle / coherent sampling cannot be met, to suppress spectral leakage, a window function must be applied to the sampled data in a FFT segment before FFT. The window function forces a smooth transition at the periodic extension boundaries. This method is called window sampling. In window sampling, the number of cycles in a FFT segment cannot be too small. The larger the number of cycles, the better the suppression effect. For THD measurement, window functions that are able to confine most of the energy of a periodic component in its neighboring FFT bins are preferred. These window functions usually have a big main lobe. Kaiser 6 ~ Kaiser 20, Blackman Harris 7, Cosine Sum 220, Cosine Sum 233, Cosine Sum 246, Cosine Sum 261 are recommended. More information on various window functions can be found at:

[https://www.virtins.com/doc/D1003/Evaluation\\_of\\_Various\\_Window\\_Functions\\_using\\_Multi-Instrument\\_D1003.pdf](https://www.virtins.com/doc/D1003/Evaluation_of_Various_Window_Functions_using_Multi-Instrument_D1003.pdf)

Fig. 3 below is derived from Fig. 1 by changing only the Rectangle window to Kaiser 8 window to suppress the spectral leakage. The test parameters are: [Sampling Rate] = 48 kHz, [Signal Frequency] = 1 kHz (ideal signal generated using the Signal Generator of Multi-Instrument), [FFT Size] = 32768, [Number of Cycles] =  $1000 / 48000 \times 32768 = 682.6667$ , [Sampling Bit Resolution] = 24, [Window Function] = Kaiser 8, [Frequency Range] = 20 Hz ~ 20 kHz. No zero padding is applied as [Record Length] = 48000 which is greater than the FFT size. As the FFT segment contains a non-integer [Number of Cycles], spectrum leakage is inevitable but suppressed greatly

by the Kaiser 8 window. The measured results are: THD = 0.0000055 % (-145.23 dB), THD+N = 0.0000055% (-145.23 dB), SINAD = 145.23 dB, SNR = 206.44 dB, ENOB = 23.83 Bit. These results represent more or less the best measurable values under 24-bit window sampling. They are far better than those of a HIFI device. This proves, from software point of view, that the above test parameters can be used in the measurement of a HIFI device. Again, the measured distortions and noises in an ideal sine wave are caused mainly by the quantization and numerical calculation errors which will be discussed later. Fig. 3 has a higher THD and a better SNR than Fig.2. This is because the quantization noise is concentrated at and thus measured as the harmonics of the signal, as a result of the integer ratio of the sampling rate to the signal frequency. This will be elaborated later.



Fig. 3 Non Full-cycle Sampling with a Kaiser 8 window suppressing greatly spectral leakage

It can be seen that the 1 kHz peak in Fig. 3 is lower but wider than that in Fig. 2. This is because a window function can only confine most of the energy of a periodic component in a few neighboring FFT bins but just cannot put all its energy into one single FFT bin, unlike the case of full-cycle or coherent sampling. Nevertheless, the software is able to calculate the correct heights of frequency peaks despite the spectral leakage.

Unlike full-cycle and coherent sampling, in window sampling, zero padding is allowed in THD measurement in Multi-Instrument.

### 3.2.3 Coherent Sampling or Window Sampling?

As shown in Figs. 2 and 3, both methods can be used for THD measurement as long as their respectively required conditions are met. The differences in the measured results will be negligible in any practical HIFI device measurements.

If there is a restriction on the FFT size and it is not possible to have a sufficient number of signal cycles in a FFT segment, then coherent sampling should be used as it work well even with one cycle.

If the ADC and DAC of the measuring device do not share the same sampling clock, then window sampling should be used, because the clock inaccuracy and unsynchronized jitter will violate the coherent sampling conditions and cause spectral leakage. This problem will not show up if the same sampling clock is used.

If a non-Rectangle window function is applied in coherent sampling, it will cause a very small suppressed spectral leakage. It should then be considered as window sampling. The measured THD and THD+N will still be correct as long as the window sampling requirements are met.

## 4. How to Avoid Quantization Noise Being Measured As Harmonic Distortion

### 4.1 Avoid Integer or Reducible Ratio of Sampling Rate to Signal Frequency

The quantization error of an ADC or a DAC device is often thought to be a white noise distributed uniformly from 0 Hz to half of the sampling rate. This is not true when sampling a periodic signal. In fact, the quantization noise can be heavily correlated to the signal frequency. When the ratio of sampling rate to signal frequency is an integer, the quantization noise will be periodized and concentrated at the harmonics of the signal, causing an overestimation of THD and underestimation of quantization noise. Fig. 3 in the previous section is such an example. The ratio of sampling rate to signal frequency in Fig. 3 is  $48000 / 1000 = 48$  (integer). If the signal frequency is changed to 997 Hz, then the ratio becomes  $48000 / 997 = 48.144433$  (non-integer and not reducible). This greatly reduces the correlation between the quantization noise and the signal frequency, and randomizes the quantization noise, as shown in Fig. 4. Compared with Fig. 3, the noise floor in Fig. 4 is higher, but its THD is lower, decreased from 0.0000055 % (-145.23 dB) in Fig.3 to 0.0000033 % (-149.70 dB) in Fig. 4. In fact, the window sampling results of Fig. 4 is almost the same as the coherent sampling results of Fig.2.



Fig. 4 A Non-Integer and Non-irreducible Ratio of Sampling Rate to Signal Frequency ( $48000 / 997 = 48.144433$ ) Randomizes Quantization Noise (24 Bits)

The THD measurement error introduced by the correlation between 24-bit quantization noise and signal frequency can be ignored in most practical THD measurements. This is because the error is very small compared with the THD value of a HIFI device. Also, the noise in the signal chain is usually sufficient to dither the 24-bit quantization process and randomize the quantization noise. However, the issue becomes worse in the 16-bit quantization and serious in the 8-bit one.

The following two figures illustrate two distinct spectral distributions of quantization noises in ideal 8-bit sine waves. The ratios of sampling rate to signal frequency are  $48000 / 1000 = 48$  in Fig. 5 and  $48000 / 997 = 48.144433$  in Fig. 6. The measured THD values are 0.2767 % (-51.16 dB) and 0.0910 % (-60.82 dB) respectively. The SNR in Fig. 5 is 204.66 dB which is significantly higher than 51.14 dB in Fig. 6 as most of the quantization noise becomes harmonics.



Fig. 5 An Integer Ratio of Sampling Rate to Signal Frequency ( $48000 / 1000 = 48$ ) Concentrates Quantization Noise at Signal Harmonics (8 Bits)

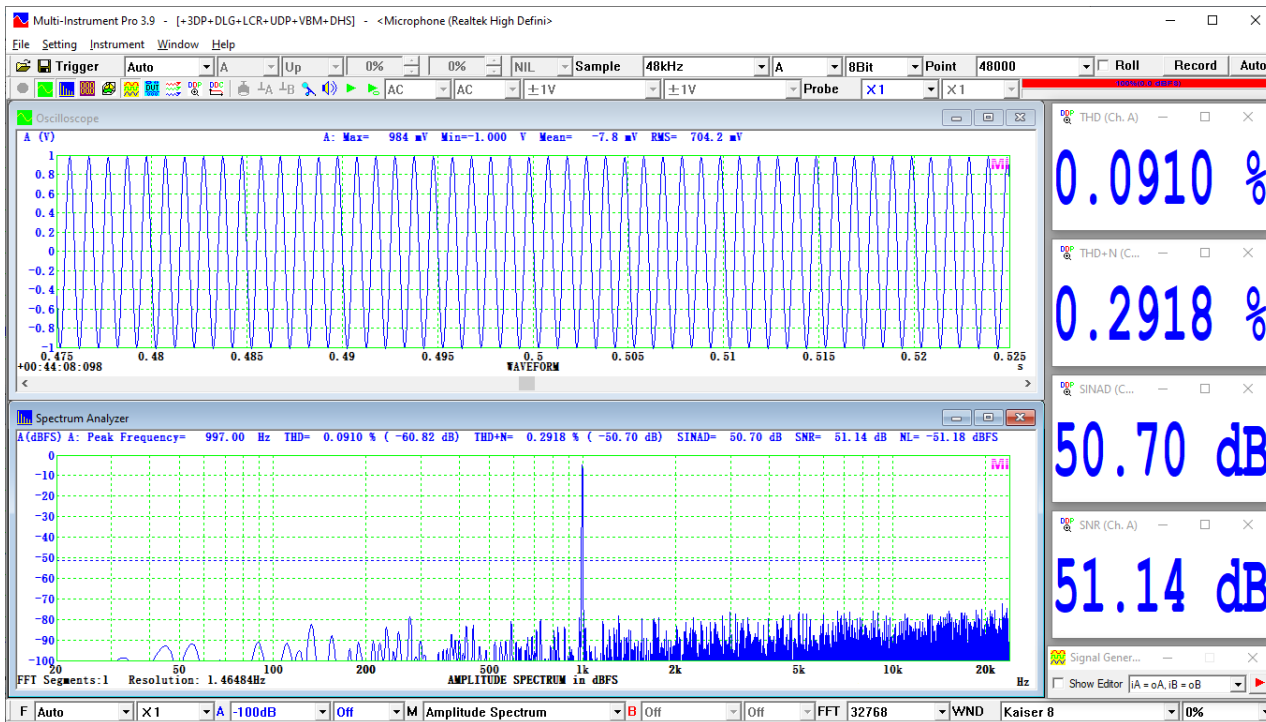


Fig. 6 A Non-Integer and Non-Reducible Ratio of Sampling Rate to Signal Frequency ( $48000 / 997 = 48.144433$ ) Randomizes Quantization Noise (8 Bits)

More generally, when sampling a periodic sine wave signal, the quantization noise is always concentrated at the fundamental and harmonics of a frequency equal to the greatest common factor of the sampling rate and signal frequency. If the greatest common factor is the signal frequency itself, then the ratio of sampling rate to signal frequency can be reduced to an integer and almost all quantization noise is concentrated at the harmonics of the signal itself. This can be observed in Figs. 3 and 5. In contrast, if the ratio of sampling rate to signal frequency is not reducible, then their greatest common factor is 1. The quantization noise is distributed quite uniformly at an interval of 1 Hz from 0 Hz to half of the sampling rate. This can be observed in Figs. 4 and 6.

The following figure shows an example where the ratio of sampling rate to signal frequency is not an integer but reducible. The test parameters are: [Sampling Rate] = 44100 Hz, [Signal Frequency] = 250 Hz (ideal signal generated using the Signal Generator of Multi-Instrument), [FFT Size] = 32768, [Number of Cycles] =  $250 / 44100 \times 32768 = 185.7596$ , [Sampling Bit Resolution] = 24, [Window Function] = Kaiser 8, [Frequency Range] = 20 Hz ~ 20 kHz. No zero padding is applied as [Record Length] = 48000 which is greater than the FFT size. The greatest common factor of 44100 and 250 is 50. It can be seen from the figure that the quantization noise is concentrated at 50 Hz and its harmonics. 1/5 of the harmonics of 50 Hz are in line with the harmonics of 250 Hz. In other words, a substantial portion of the quantization noise energy is included in THD.

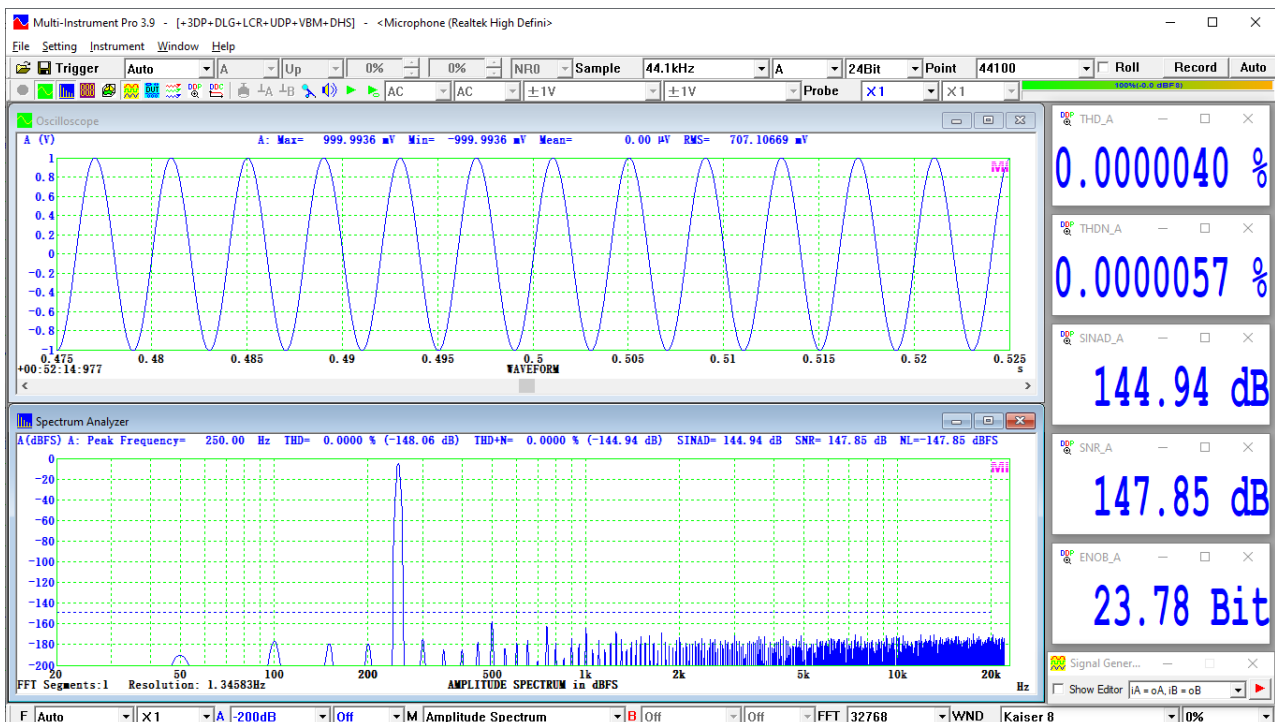


Fig. 7 Sampling Rate 44100 Hz, Signal Frequency 250 Hz, Greatest Common Factor 50 Hz, Quantization Noise is concentrated at 50 Hz and its harmonics.

In conclusion, to avoid the concentration of quantization noise at the harmonics of the test signal, which leads to an overestimation of THD, choose a test signal frequency that is not reducible with the sampling rate.

## 4.2 Add Dither to Signal before Quantization

When the ratio of sampling rate to signal frequency is an integer or reducible, especially when the amplitude of the signal is small, a small amount of white noise (dither) with an amplitude of 0.5 ~ 1 bit can be summed with the signal in order to randomize the quantization noise and void the overestimation of THD. Fig. 8 shows an ideal 1 kHz sine wave sampled at 48 kHz and quantized at 24 bits. It has the same test parameters as Fig. 3 except that a white noise with an amplitude of 0.5 bit is added into the signal. The multi-tone configuration of the test signal is shown in the Fig. 9 (Relative Amplitudes of the 1 kHz sine wave and the 0.5-bit white noise are 1 and  $1/2^{24} = 6 \times 10^{-8}$  respectively). It can be seen from Fig. 8 that, despite the integer ratio of sampling rate to signal frequency, dithering greatly reduces the correlation between the quantization noise and the signal frequency, and randomizes the quantization noise. Compared with Fig. 3, the noise floor in Fig. 8 is higher, but its THD is lower, decreased from 0.0000055 % (-145.23 dB) in Fig.3 to 0.0000033 % (-149.52 dB) in Fig. 8. Compared with Fig. 4, Fig. 8 has almost the same THD but a higher THD+N, suggesting that avoiding integer and reducible ratio of sampling rate to signal frequency is preferable to dithering in THD measurement.

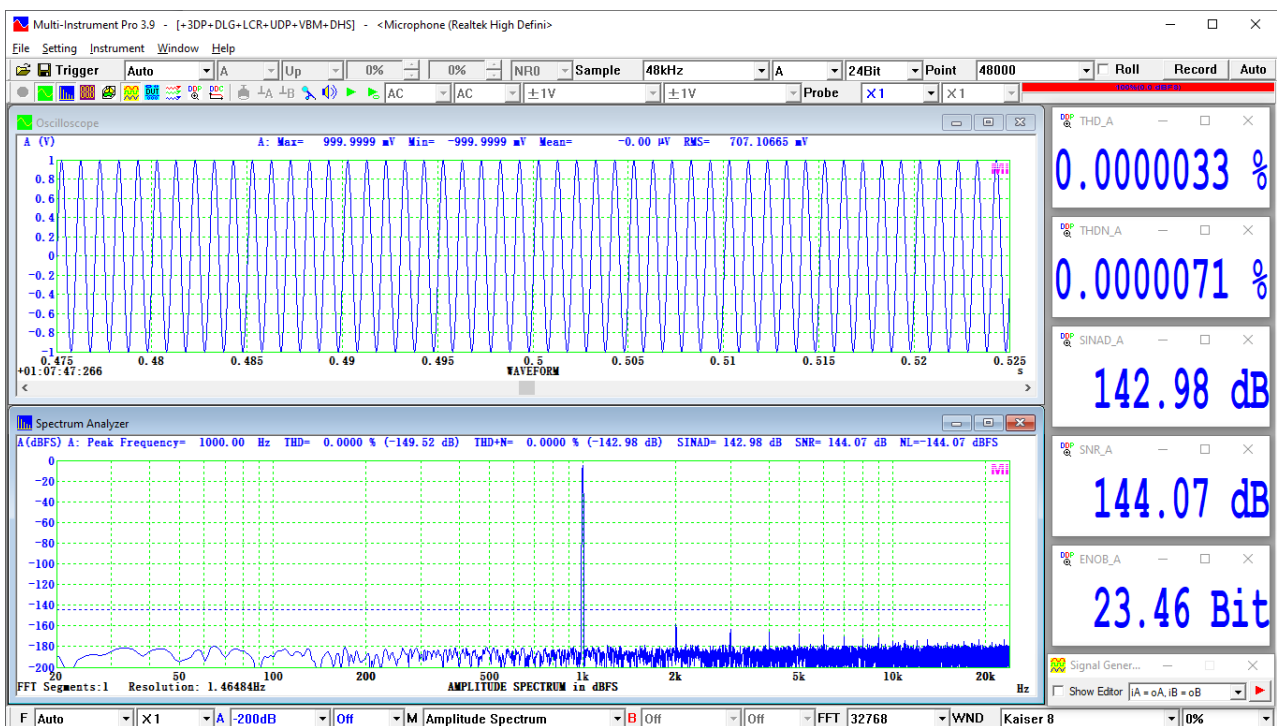


Fig. 8 Adding Dither Randomizes Quantization Noise When The Ratio of Sampling Rate to Signal Frequency is an Integer ( $48000 / 1000 = 48$ )



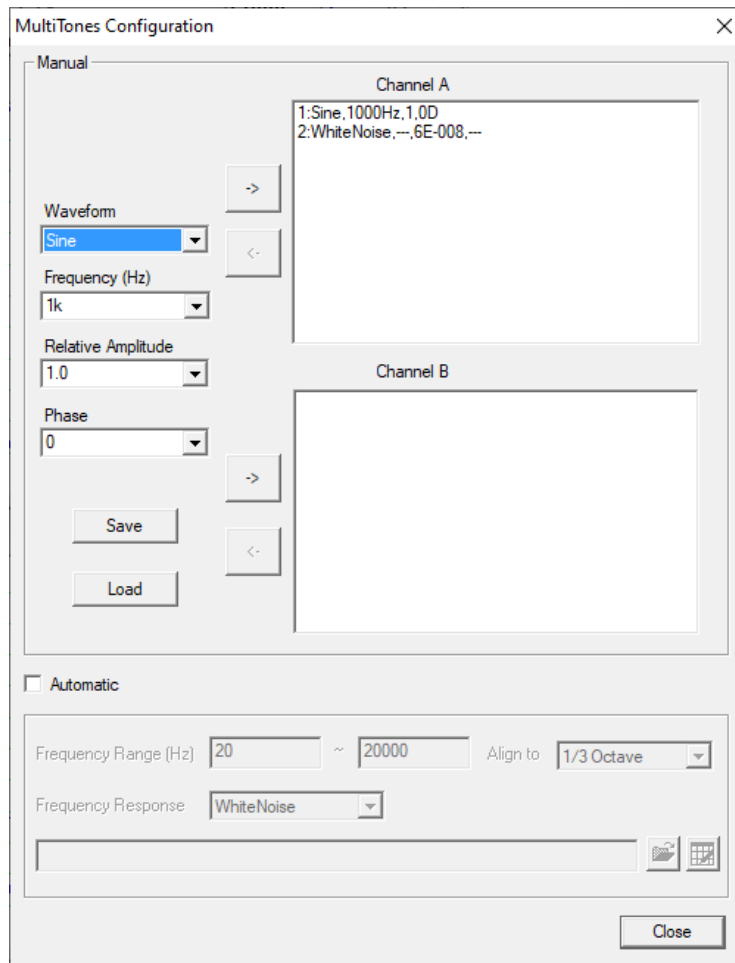


Fig. 9 Dithering Configuration using The Signal Generator Multitone Function of Multi-Instrument

## 5. Software and Hardware Loopback Verification

### 5.1 Why Software and Hardware Loopback Verification?

It is important to verify the correctness of the test parameters before conducting the actual measurements on the DUT (Device Under Test). Linear or non-linear distortion measurements normally involve spectrum analysis via FFT, which is notorious for producing artefacts if the test parameters are not set properly. In addition, as shown previously, measurement inaccuracy resulting from quantization noise can be pronounced if the test parameters are not optimized. Therefore there is a need to understand the performance of the test parameters to be used. This can be done by analyzing an ideal test signal, just as what has been done in the previous chapters. In Multi-Instrument, the ideal test signal (stimulus) can be generated and fed into the Oscilloscope and Spectrum Analyzer directly without going through hardware using the “iA=oA, iB=oB” mode (which means “input Ch.A = output Ch.A, input Ch.B = output Ch. B”) of the Signal Generator. This method is called software loopback. It is also possible to save an ideal test signal to a WAV file and then load it into the Oscilloscope and Spectrum Analyzer via [File]>[Open]. Through software loopback tests, the software measurement accuracy under the designated test parameters can be verified.

The next step is to verify the performance of the measuring device. This can be done by hardwiring directly the output of the DAC to the input of the ADC of the measuring device(s). The test signals are the same as the ones used in software loopback tests. In addition to the FFT artefacts, quantization noise and numerical computation error that have already been evaluated in the software loopback tests, distortions and noises of the measuring device itself are included in hardware loopback tests. The overall performance must be substantially better than that of the DUT to ensure the measurement accuracy.

The following figure illustrates the concepts of software and hardware loopback tests.

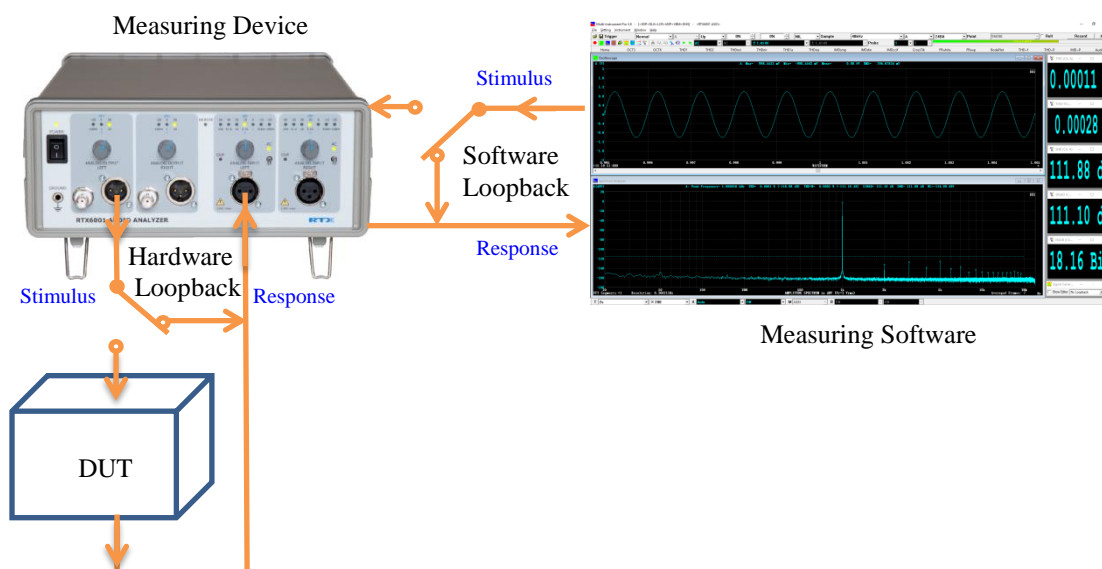


Fig. 10 Software and Hardware Loopback Tests

## 5.2 Examples: RTX6001 Audio Analyzer Hardware Loopback Tests

The following figures show the hardware loopback test results of RTX6001, a 24-bit high-performance audio analyzer. Figs. 11, 12 and 13 use: (1) Coherent sampling (Rectangle window) and a non-integer ratio of sampling rate to signal frequency, (2) window sampling (Kaiser 8 window) and an integer ratio of sampling rate to signal frequency, (3) window sampling (Kaiser 8 window) and a non-integer ratio of sampling rate to signal frequency, respectively. There is almost no difference between the measurement results of Figs. 12 and 13. Both measured a THD of 0.00013% (-117.9 dB) and a THD+N of 0.00028% (-111.2 dB). The reason is that the noise floor of the hardware exceeds 0.5 bit and it has a quite white spectrum. As a result, the quantization noise is sufficiently randomized. Fig. 11 measured a THD of 0.00012% (-118.1 dB) and a THD+N of 0.00032% (-109.8 dB). Its slightly increased noise level is mainly due to the wider skirt of the fundamental peak as compared to that in Figs. 12 and 13. It is a result of the marginal violation of the coherent sampling conditions due to the jitter of the sampling clock. It is software dependent as to where to draw the demarcation line between the fundamental and noise along the skirt. Nevertheless the differences in the three cases are still very small. The above hardware loopback test results are much worse than their corresponding software loopback ones in Figs. 2, 3 and 4, showing that the hardware is the bottleneck in improving the measurement accuracy.

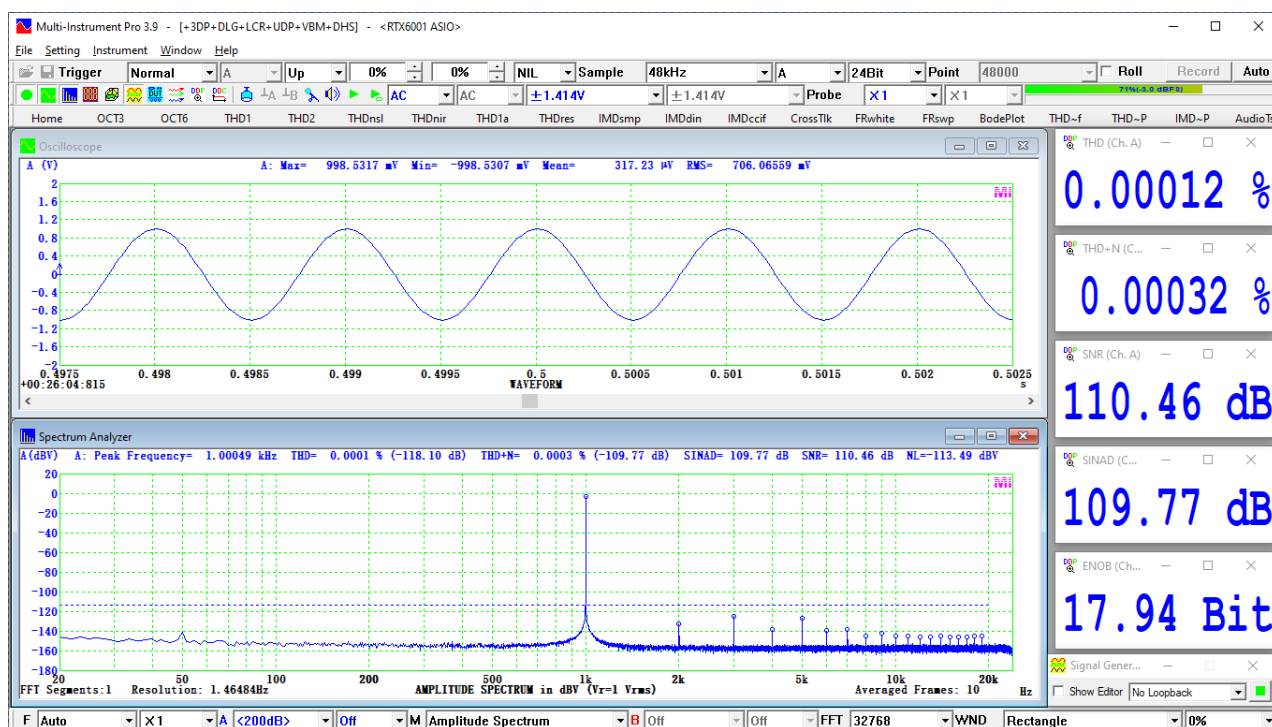


Fig. 11 RTX6001 hardware loopback test: [Sampling Rate] = 48 kHz, [Signal Frequency] = 1000.48828125 Hz, [FFT Size] = 32768, [Window] = Rectangle

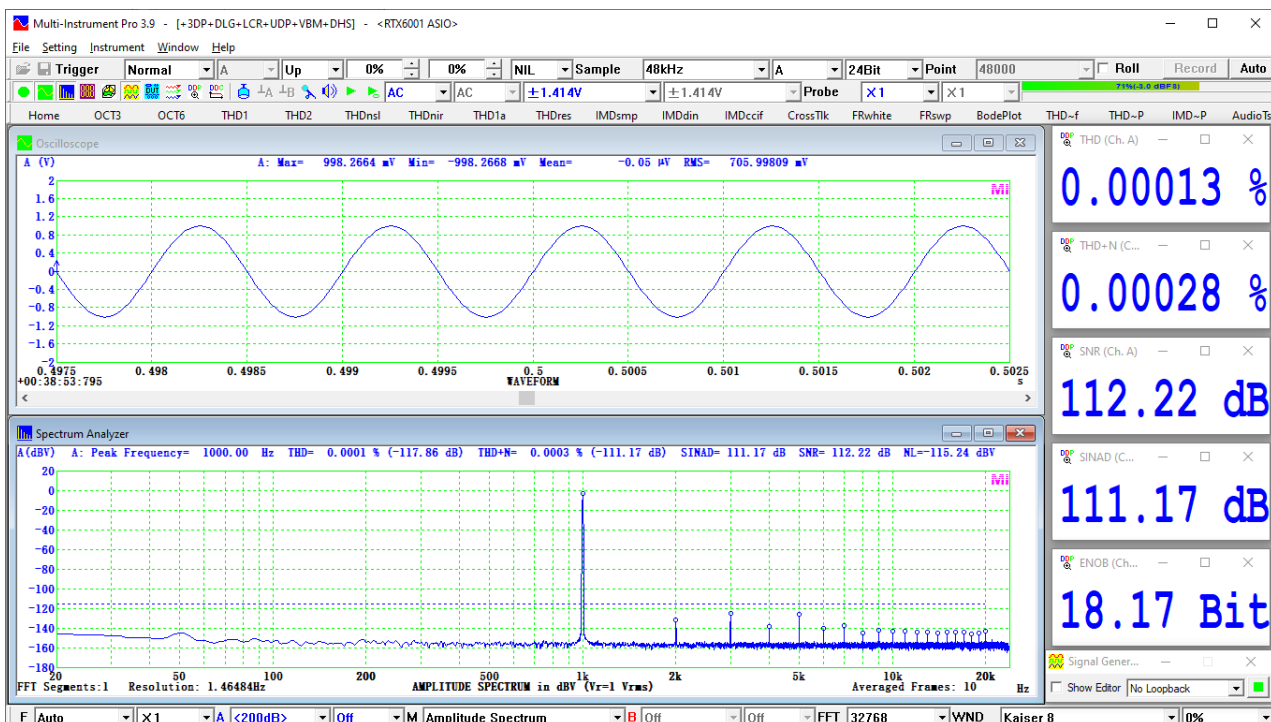


Fig. 12 RTX6001 hardware loopback test: [Sampling Rate] = 48 kHz, [Signal Frequency] = 1000 Hz, [FFT Size] = 32768, [Window] = Kaiser 8

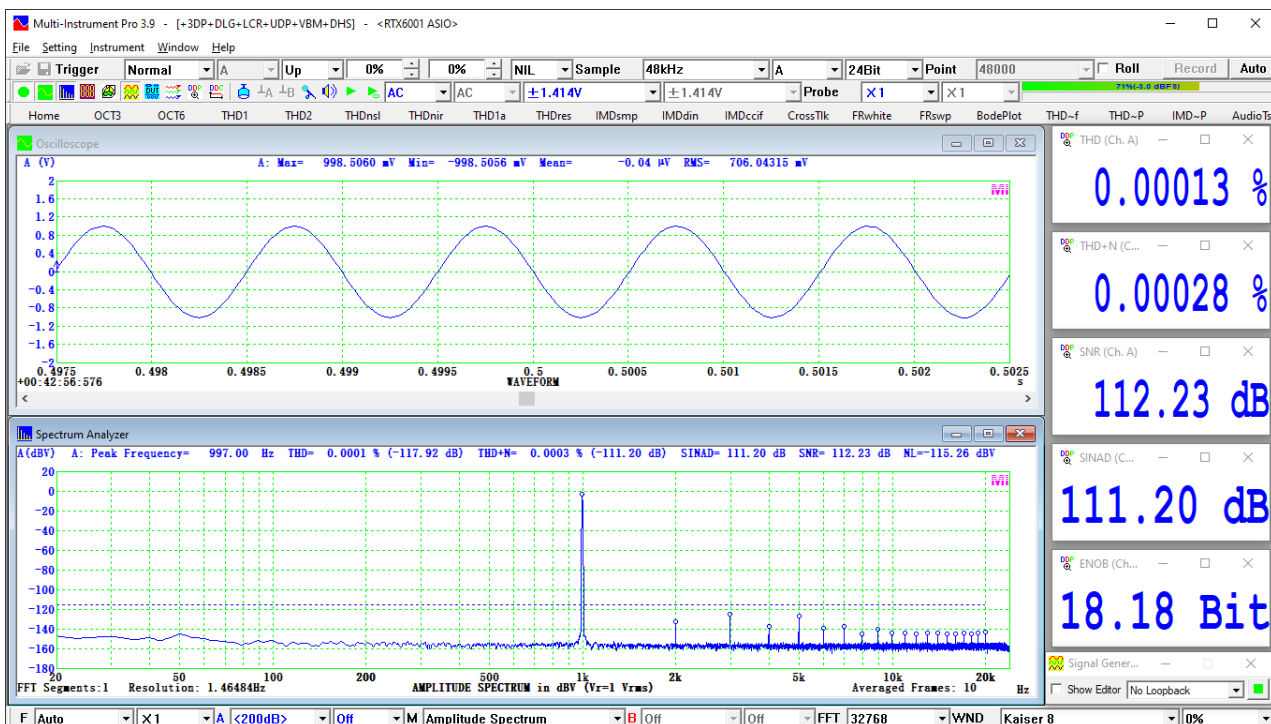


Fig. 13 RTX6001 hardware loopback test: [Sampling Rate] = 48 kHz, [Signal Frequency] = 997 Hz, [FFT Size] = 32768, [Window] = Kaiser 8

It is possible to apply frequency weighting to the THD measurement via [Spectrum Analyzer Processing]>“Weighting” in Multi-Instrument. This technique is often used to obtain an A-weighted THD+N. The following figure shows the A-weighted results of Fig. 12. It can be seen that THD+N is improved from 0.00028% (-111.2 dB) to 0.00024% (-112.6 dB) after A-weighting.

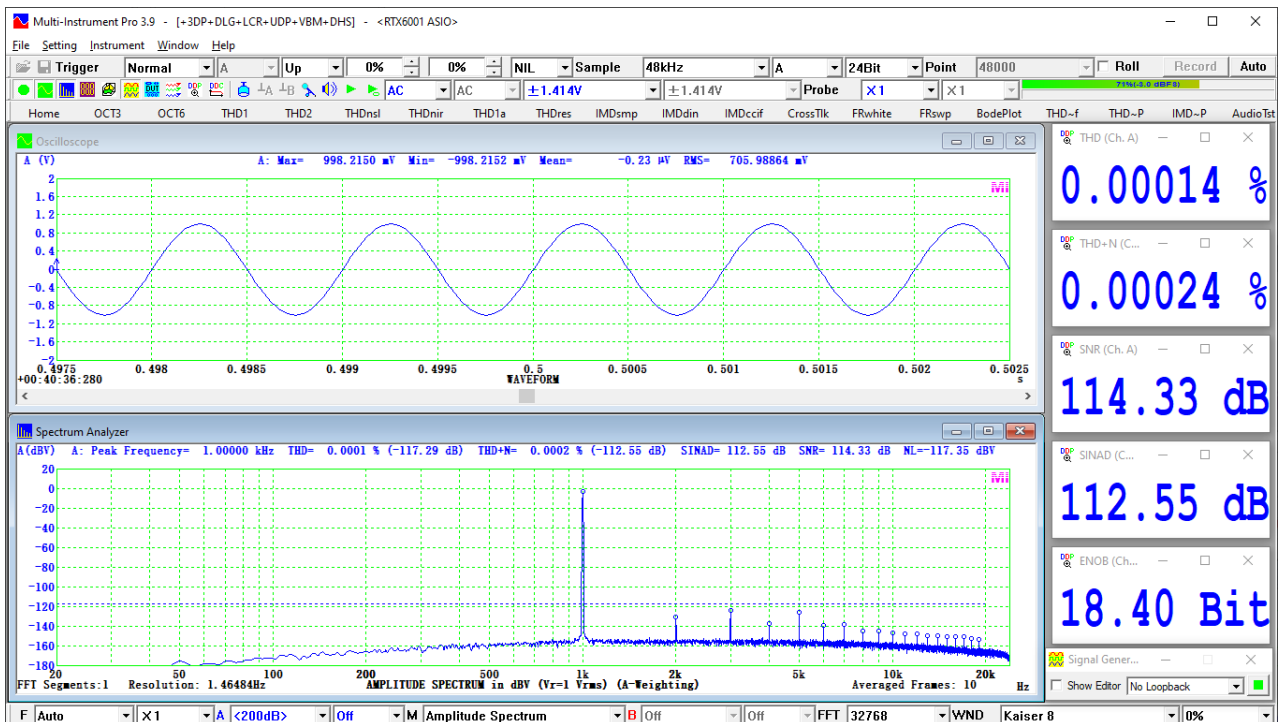


Fig. 14 RTX6001 hardware loopback test: [Sampling Rate] = 48 kHz, [Signal Frequency] = 1000 Hz, [FFT Size] = 32768, [Window] = Kaiser 8, A-weighted

The above results show that RTX6001 is truly a high-performance audio analyzer. Moreover, it is able to achieve similar results in a wide voltage range ( $\pm 141.4\text{mV}$ ,  $\pm 447.2\text{mV}$ ,  $\pm 1.414\text{V}$ ,  $\pm 4.472\text{V}$ ,  $\pm 14.14\text{V}$ ,  $\pm 44.72\text{V}$ ,  $\pm 141.4\text{V}$ ).

### 5.3 Examples: Additional Sound Card Hardware Loopback Tests

Professional audio analyzers are expensive. Some sound card compatible audio interfaces, such as RME ADI-2 Pro, RME ADI-2 Pro FS and RTX6001, coupled with Multi-Instrument, are able to achieve equally good performance. More sound cards can achieve comparable performance under controlled conditions. These sound cards include:

(1) Focusrite Scarlett Solo

THD: 0.0009% (-100.54 dB)  
 THD+N: 0.0029% (-90.86 dB)

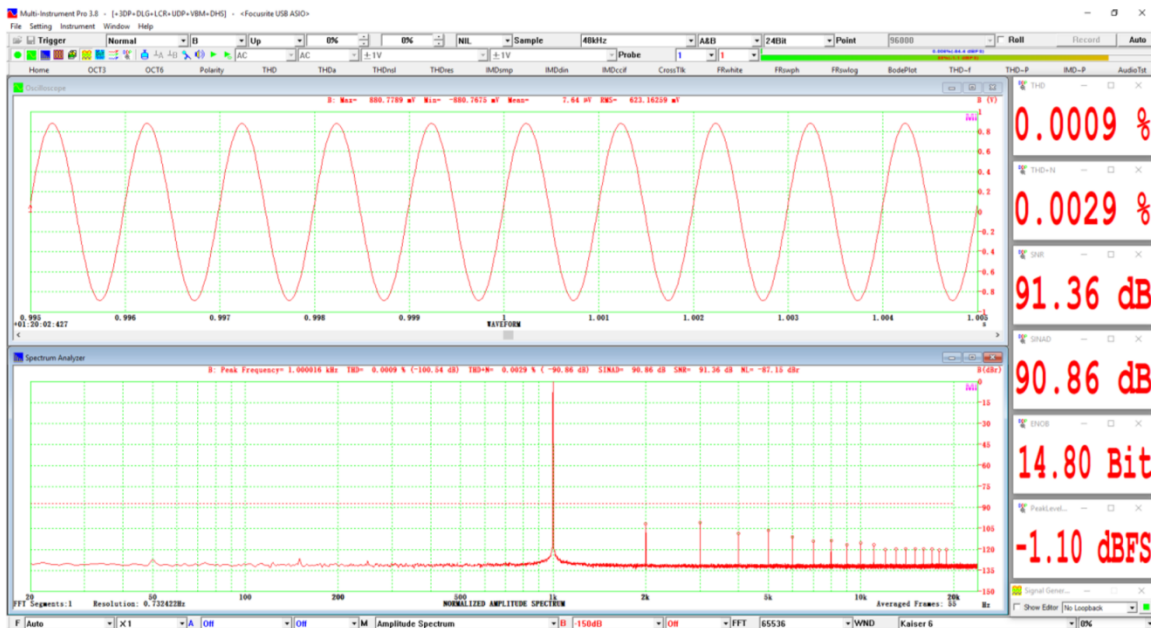


Fig. 15 Focusrite Scarlett Solo hardware loopback test

See the detailed test report at:

<https://www.virtins.com/doc/Focusrite-Scarlett-Solo-Test-Report-using-Multi-Instrument.pdf>

(2) EMU Tracker Pre

THD: 0.000351% (-109.1 dB)

THD+N: 0.001424% (-96.9 dB)

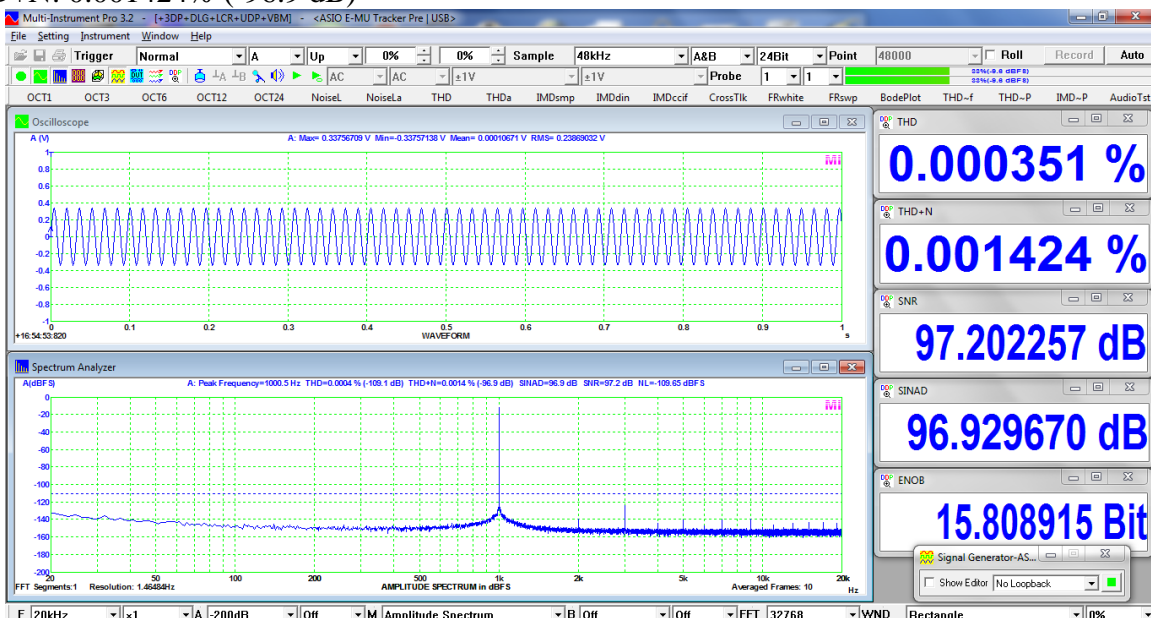


Fig. 16 EMU Tracker Pre hardware loopback test

See the detailed test report at:

[https://www.virtins.com/doc/D1004/EMU\\_Tracker\\_Pre\\_Report\\_D1004.pdf](https://www.virtins.com/doc/D1004/EMU_Tracker_Pre_Report_D1004.pdf)

(3) EMU 0204

THD: 0.000398% (-108.0 dB)

THD+N: 0.000921% (-100.7 dB)

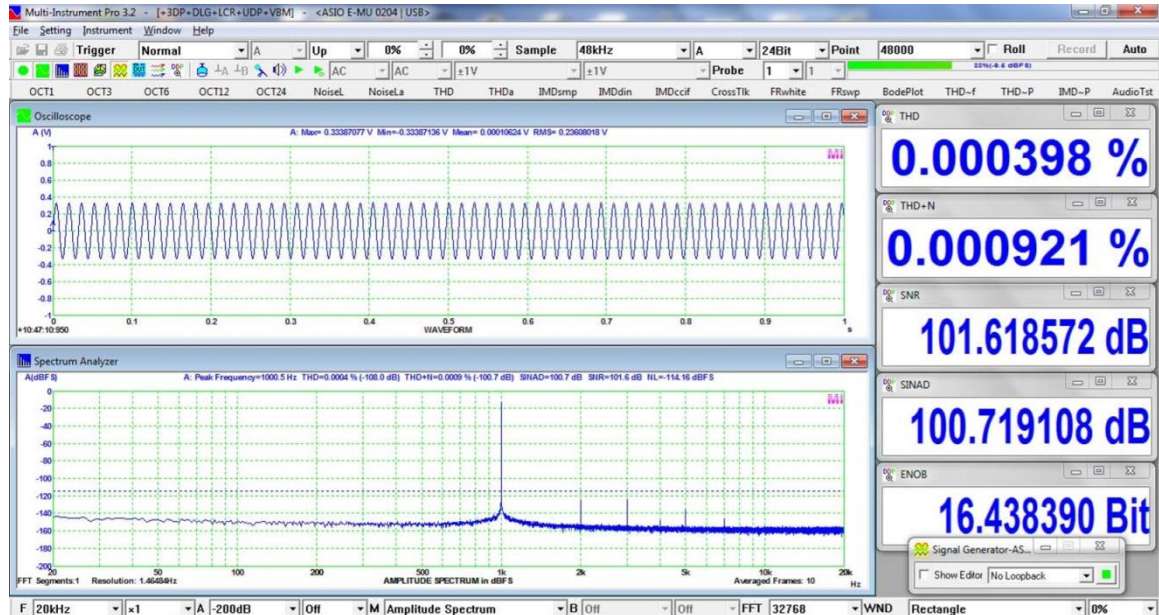


Fig. 17 EMU 0204 hardware loopback test

See the detailed test report at:

[https://www.virtins.com/doc/D1007/EMU\\_0204\\_Report.pdf](https://www.virtins.com/doc/D1007/EMU_0204_Report.pdf)

(4) ASUS Xonar Essence STX

THD: 0.00063% (-103.94 dB)

THD+N: 0.00073% (-102.71 dB)

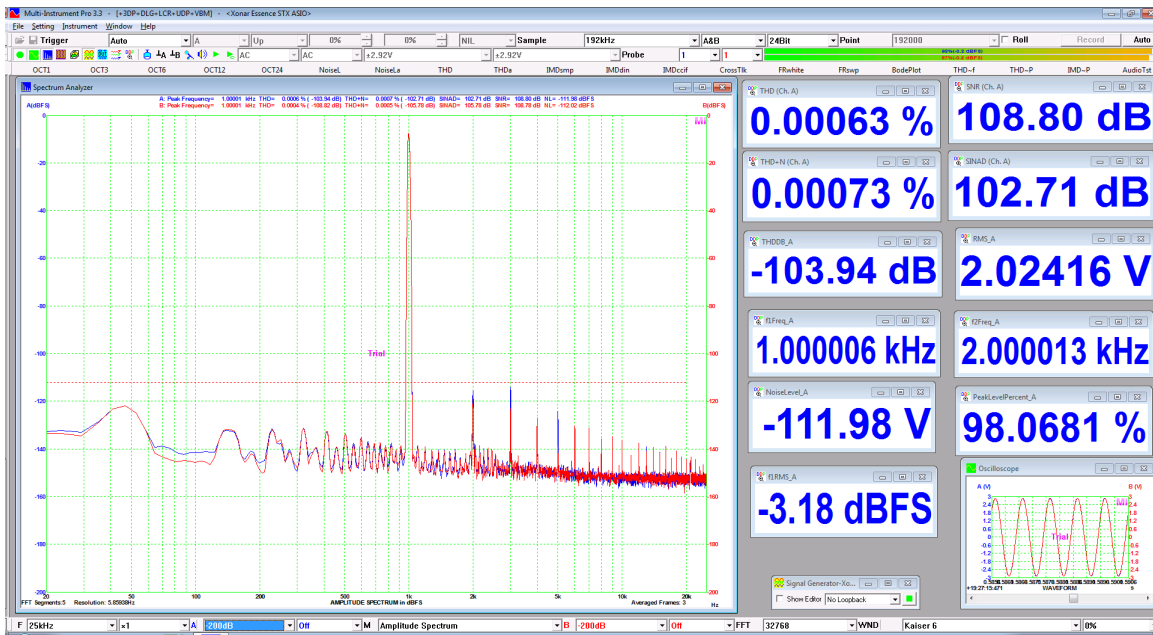


Fig. 18 ASUS Xonar Essence STX hardware loopback test (Obtained from internet)

(5) Prism Sound Lyra  
 THD: 0.00037% (-108.5 dB)  
 THD+N: 0.000456% (-106.7 dB)

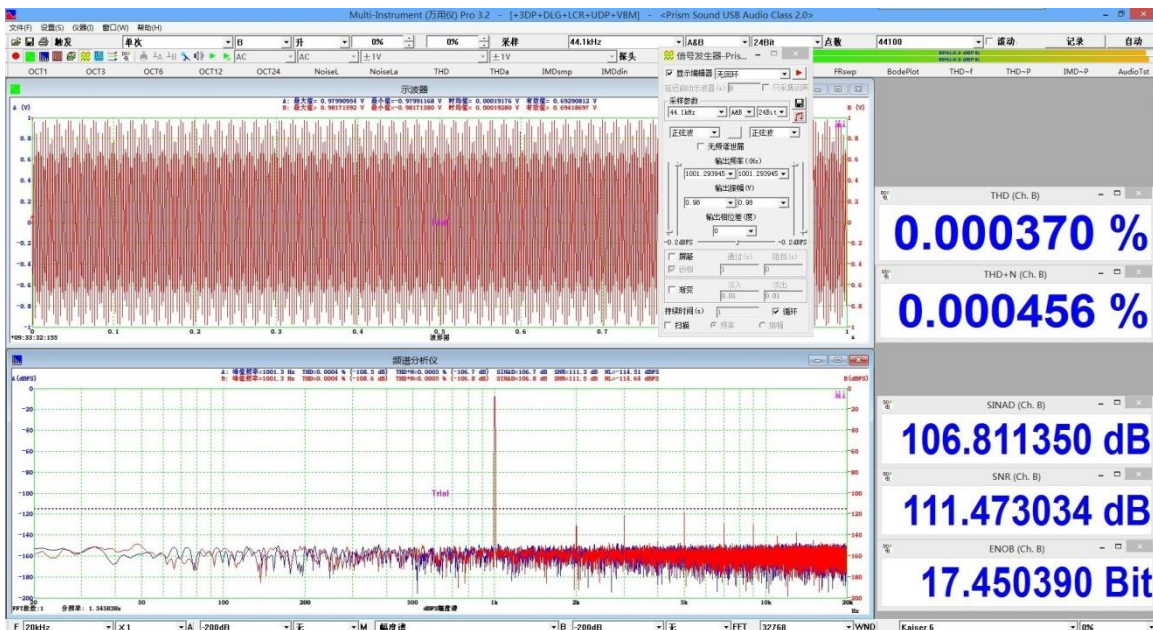


Fig. 19 Prism Sound Lyra hardware loopback test (Obtained from internet)

The price of a professional sound card is much lower than that of a professional audio analyzer, partially due to its high sales volume. The drawbacks of a professional sound card are that it usually has a limited input voltage range, is not voltage calibrated and might have a low input impedance. A voltage divider consisting of two resistors can be used to expand linearly its input



voltage range if required. THD measurement is based on relative values and thus voltage calibration is not absolutely required.

## 5.4 Estimation of Software Measurement Accuracy Using a Simulated Distortion Signal

As the performance of hardware continues to improve, it would be interesting to see how the 24-bit quantization noise and numerical computation error start to affect the measurement accuracy. This can be evaluated using a simulated test signal. Fig. 20 shows a multi-tone signal consisting of 997 Hz and 2991 Hz with an amplitude ratio of 1:0.000001. The measured THD and THD+N are 0.0000988% (-120.11 dB) and 0.0000989% (-120.10 dB) respectively. Their theoretical values are both 0.0001% (-120 dB). The software measurement errors are thus negligibly small at this distortion and noise level.

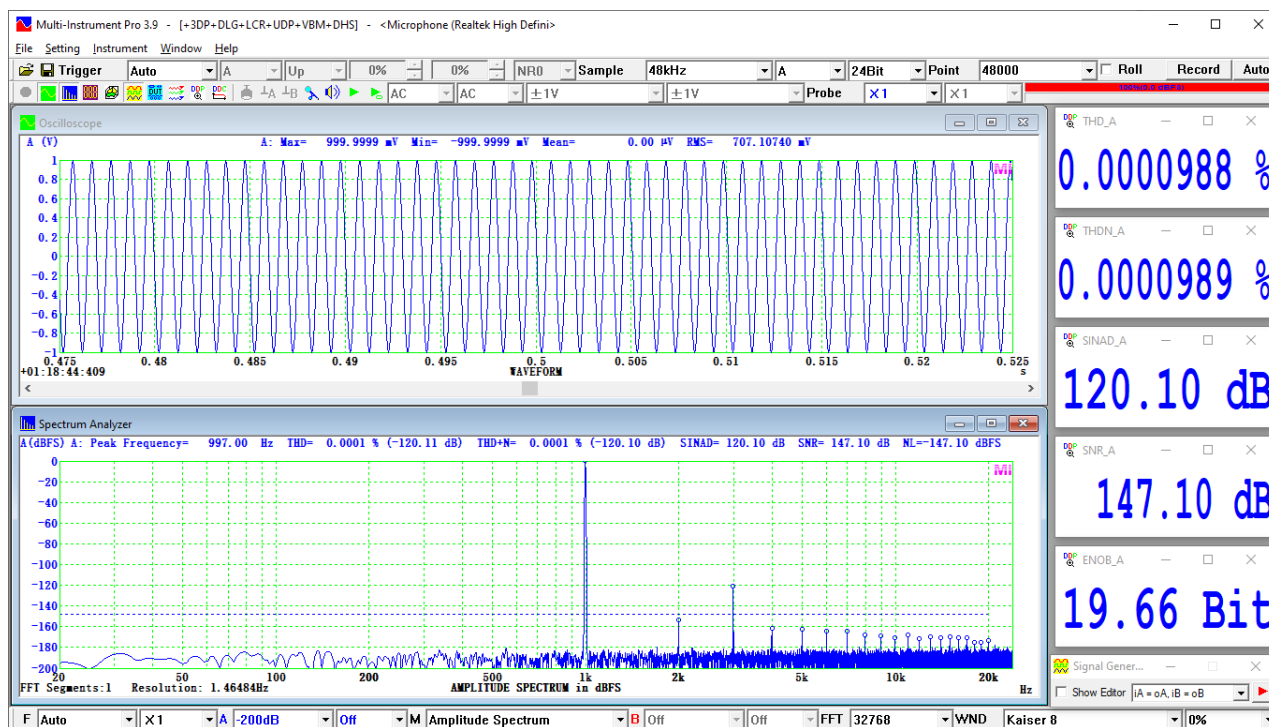


Fig. 20 [Sampling Rate] = 48 kHz, [Signal] = 997 Hz + 2991 Hz with Amplitude Ratio 1:0.000001, [FFT Size] = 32768, [Window] = Kaiser 8 (Simulation)

What if the harmonic distortion is 20 dB lower? Fig. 21 shows a multi-tone signal consisting of 997 Hz and 2991 Hz with an amplitude ratio of 1:0.0000001. The measured THD and THD+N are 0.0000093% (-140.7 dB) and 0.0000102% (-139.8 dB) respectively, which are still very close to their theoretical values 0.00001% (-140 dB). The software measurement errors are still very small when THD+N is as low as -140 dB. Note that the harmonic amplitude here is only  $0.0000001 \times 2^{24}/2 = 0.84$  bit. Till now, almost no hardware is commercially available to achieve this performance yet.



Fig. 21 [Sampling Rate] = 48 kHz, [Signal] = 997 Hz + 2991 Hz with Amplitude Ratio 1: 0.0000001, [FFT Size] = 32768, [Window] = Kaiser 8 (Simulation)

Fig. 22 shows how the multi-tone test signal in Fig. 21 is configured and generated in Multi-Instrument.

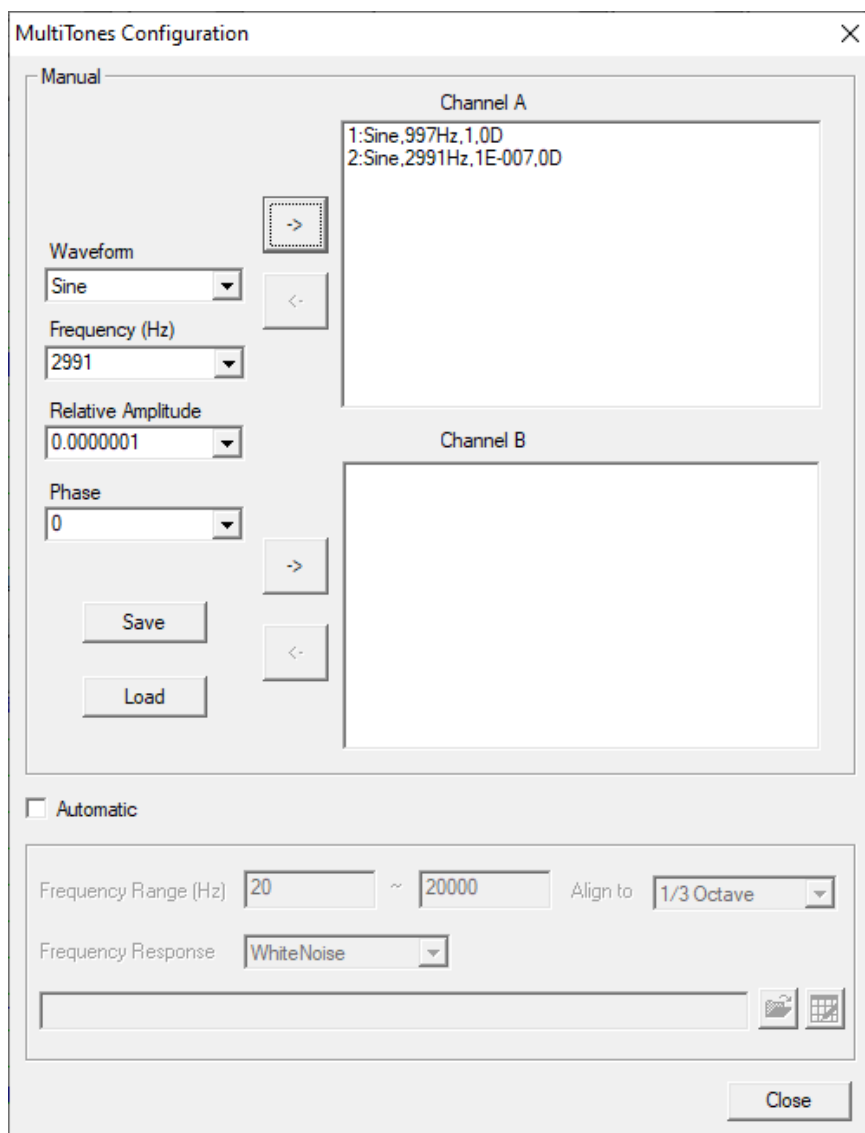


Fig. 22 Fundamental and Harmonic Configuration using The Signal Generator Multitone Function of Multi-Instrument

## 6. Distortion Residual Waveform

Different amplifier topologies and components have different harmonics creation mechanisms, resulting in different distorted waveform characteristics, such as clipping and crossover distortions. These time-domain characteristics will become unidentifiable if converted to frequency domain. When the distortion is large, they can be observed directly from the output waveform. However, in most of the cases, the distortion is very small compared with the fundamental. Thus there is a need to remove the fundamental in order to reveal the distortion's time-domain characteristics, which in turn can be used to tune the amplifiers or optimize their design. In the old days, an analogue notch filter was employed to do this job. However, it introduces additional distortion and noise and cannot be of true linear phase. With Multi-Instrument, the fundamental can be readily removed instead through a linear-phase FIR digital filter with almost ideal performance. It is also possible to synthesize the distortion residual waveform after harmonic decomposition.

### 6.1 Method 1: Removing the Fundamental using a FIR Digital Filter

To avoid phase distortion, a linear-phase FIR digital filter must be used. The commonly used linear-phase FIR digital filters are of even order and have a symmetric impulse response. A N-order FIR filter introduces a time delay of  $N/2/[\text{Sampling Rate}]$ . This delay must be corrected so that the distortion residual waveform can be time aligned with the original waveform for comparison.

In the following examples, a test signal with a fundamental frequency of 1 kHz was fed into the input channels A & B of Multi-Instrument. The signal in the input channel B was filtered by a 500Hz ~ 1500Hz band-stop, 1022-order, Kaiser 6 windowed, FIR filter with the FIR delay removed. This filter is able to attenuate the fundamental by 179 dB without affecting its harmonics, manifesting the advantages of digital filters over analogue ones. The signal in the input channel A remained unchanged. For easy viewing and explanation, the test signal in the following examples still contained a significant level of distortion. In Multi-Instrument, the above FIR parameters can be configured by right clicking anywhere within the Oscilloscope window and selecting [Oscilloscope Processing] (see Fig.23 below).

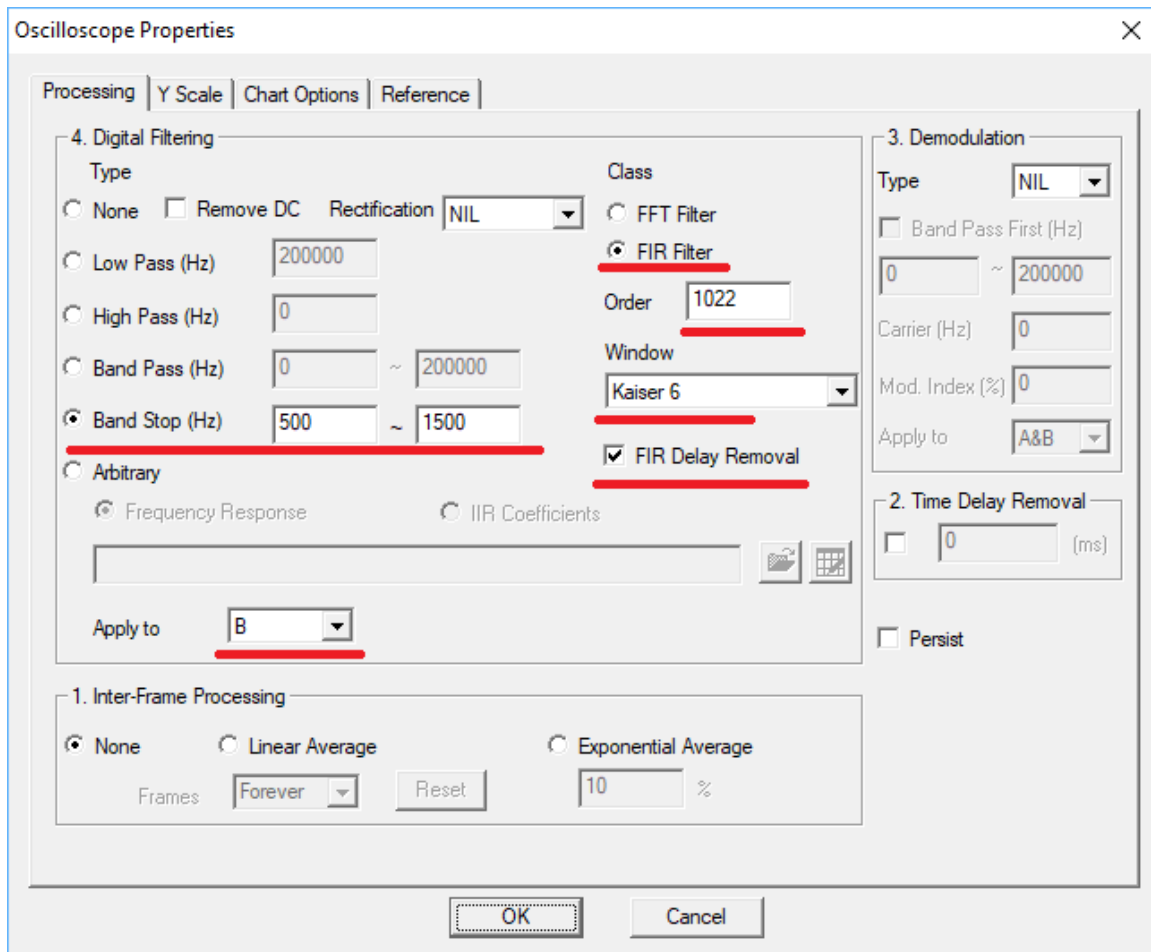


Fig. 23 Configuration of The Band-Stop FIR Filter for 1 kHz Fundamental Removal

For live signal measurement, Multi-Instrument provides a way to replicate digitally the signal from the input channel A to the input channel B. It can be done through [Setting]>[ADC Devices]>“Channel Operation”>“A=iA, B=iA”. This option is convenient for comparing the same live signal with different channel processing parameters simultaneously, without physically feeding the signal to both channels.

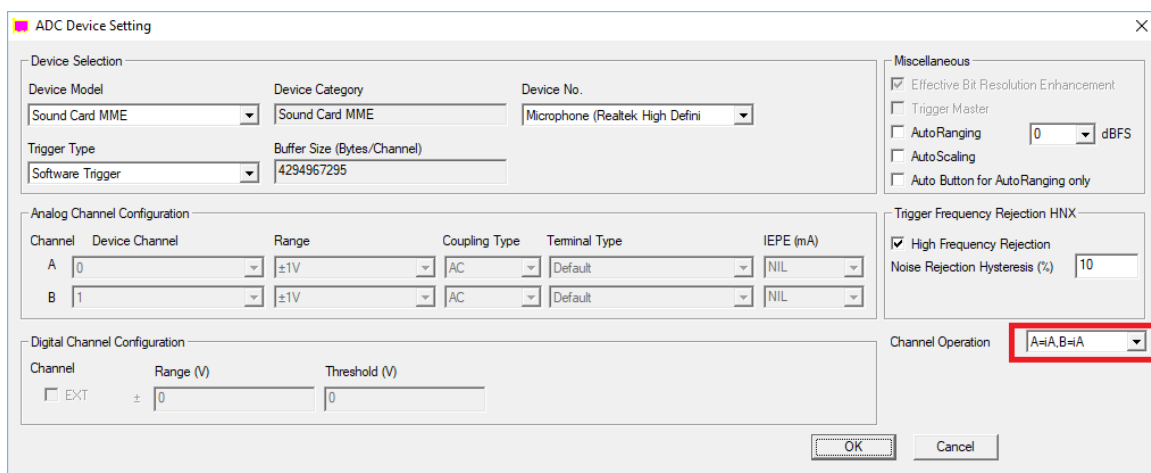


Fig. 24 Configuration for Replicating a Live Signal from Input Ch.A to Input Ch.B

For simulated signal measurement, the signal generated in the output channel A can be fed to the input channels A & B if the software loopback mode “iA=oA, iB=oA” is selected in the Signal Generator panel of Multi-Instrument (see Fig. 25).

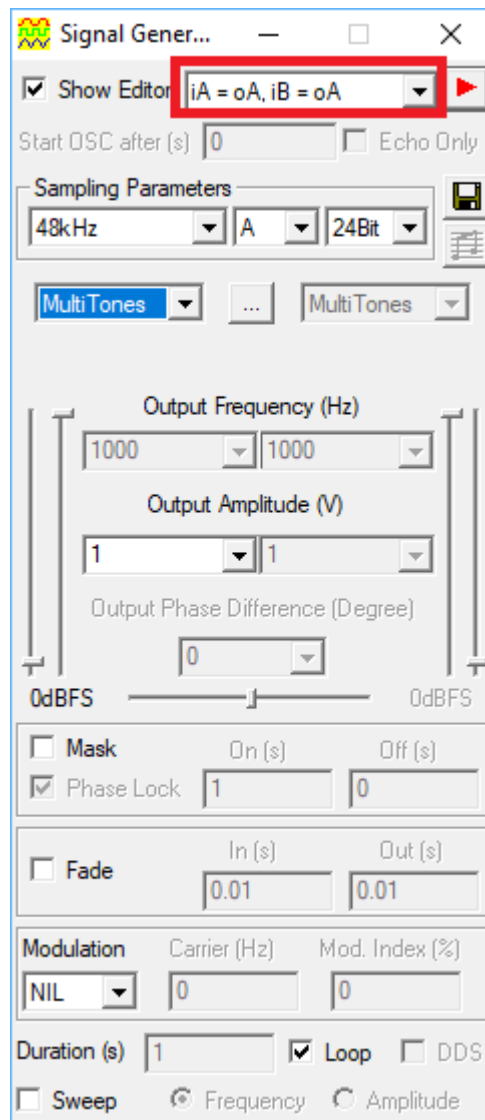


Fig. 25 Configuration for Feeding Simulated Signal in Output Ch.A to Input Ch.A and Input Ch.B

### 6.1.1 Harmonic Distortion with Only Even-Order Harmonics

An example of harmonic distortion with only even-order harmonics is shown in Fig. 26. The test signal is a simulated multitone signal consisting of 1 kHz and 2 kHz with an amplitude ratio of 1:0.1 and initial sine phases of  $0^\circ$  and  $-90^\circ$ . The blue waveform is the original test signal and the red one is the distortion residual. Their phase relationship can be clearly seen in the figure. The peaks of the second-order harmonic are located alternatively at the peaks and troughs of the fundamental, enhancing the peak height and reducing the trough depth in the overall waveform and thus making it asymmetric about the x axis. This type of distortion is often observed in single-ended Class A tube power amplifiers.

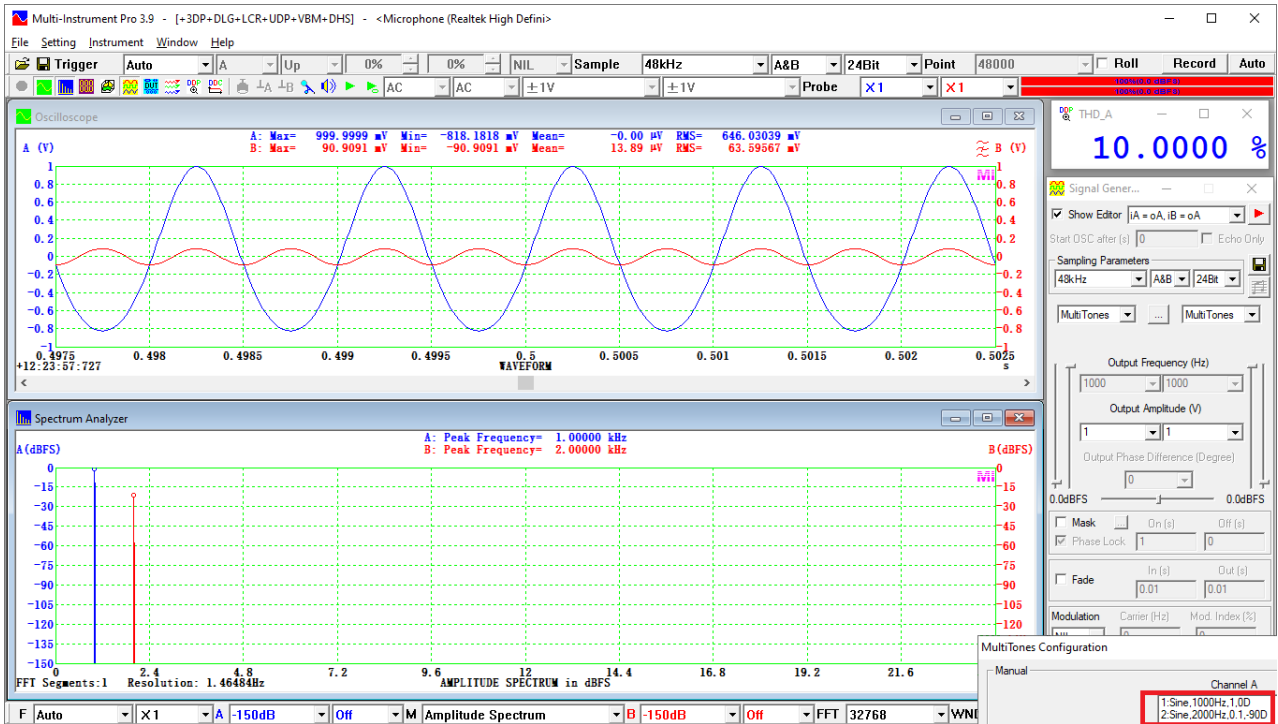


Fig. 26 Distortion Residual Containing only Second-Order Harmonics (Simulated, Fundamental Removed via FIR)

The transfer characteristic between the input and the output of a system can be shown intuitively using Lissajous Plot (X-Y Plot), as shown in Fig. 27 below. The input is an ideal 1 kHz sine wave. The output contains its second-order harmonic additionally. It can be seen from the figure that the curve is not symmetric about the zero-crossing point.

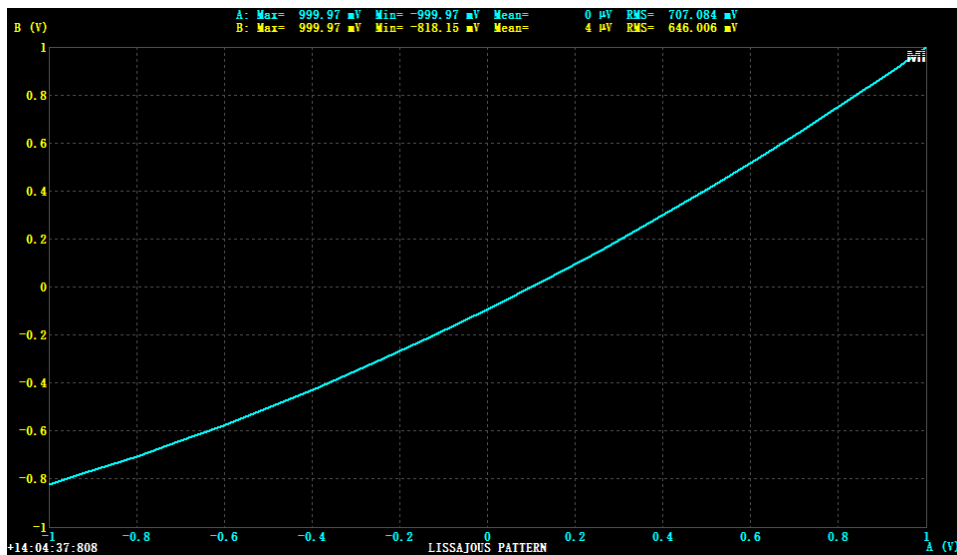


Fig. 27 Input-Output Plot of Harmonic Distortion Containing Only Second-Order Harmonics (Simulated)

### 6.1.2 Harmonic Distortion with Only Odd-Order Harmonics

Fig. 28 shows an example of harmonic distortion with only odd-order harmonics. The test signal is a simulated multitone signal consisting of 1 kHz and 3 kHz with an amplitude ratio of 1:0.1 and initial sine phases of  $0^\circ$  and  $0^\circ$ . The blue waveform is the original test signal and the red one is the

distortion residual. Their phase relationship can be clearly seen in the figure. The peaks and troughs of the fundamental are always moderated by the troughs and peaks of third-order harmonic and this alteration is symmetrical about the X axis. This type of distortion (called soft clipping) is sometimes observed in push-pull Class A tube power amplifiers.

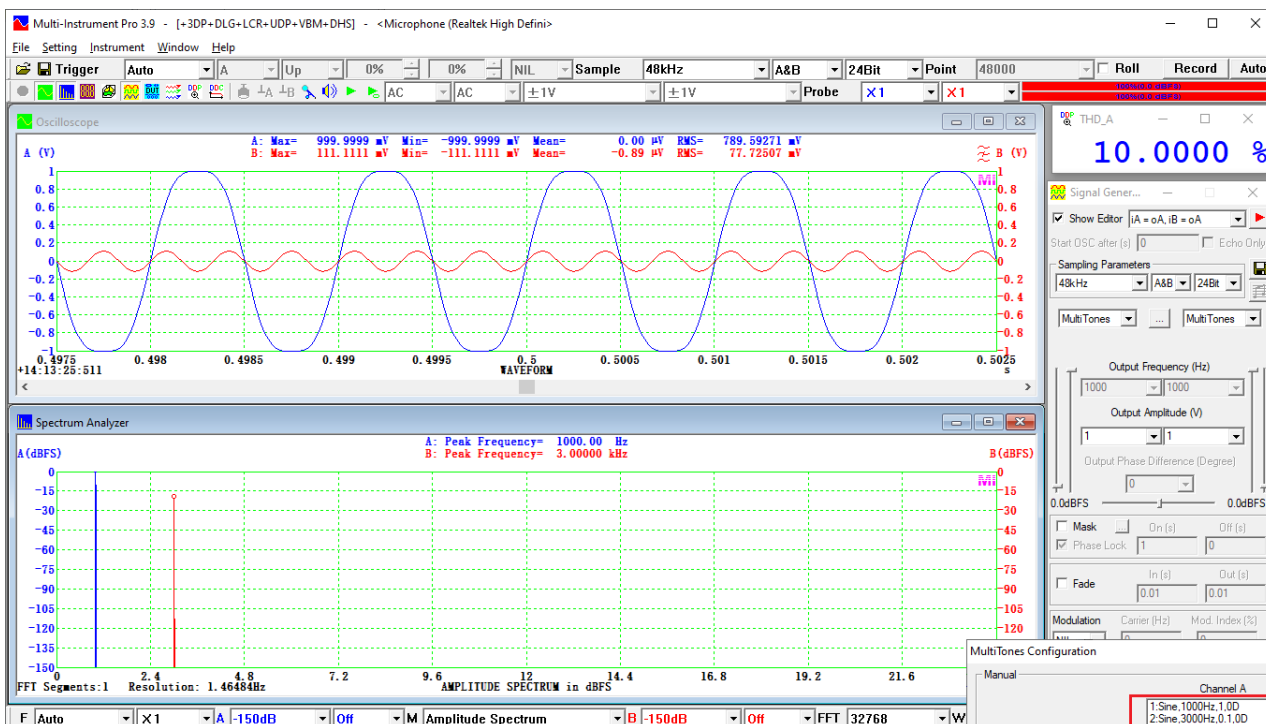


Fig. 28 Distortion Residual Containing only Third-Order Harmonics (Simulated, Fundamental Removed via FIR)

Fig. 29 shows the transfer characteristic between the input and the output based on the above signal. The input is an ideal 1 kHz sine wave. The output contains its third-order harmonic additionally. It can be seen from the figure that the curve is symmetric about the zero-crossing point.

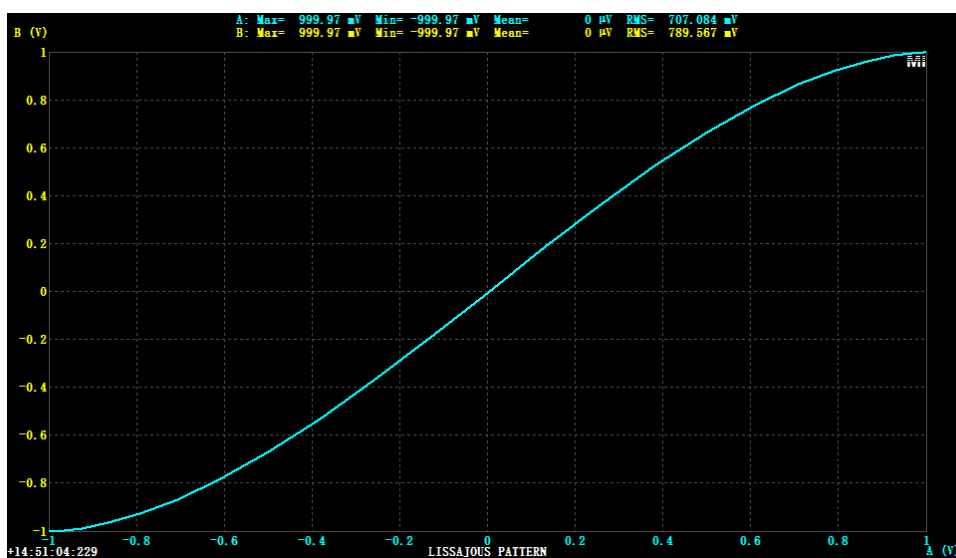


Fig. 29 Input-Output Plot of Harmonic Distortion Containing Only Third-Order Harmonics (Simulated)



### 6.1.3 Crossover Distortion

Crossover distortion is often seen in Class A and Class AB power amplifiers. Its magnitude does not decrease with the signal level. It contains many high-order harmonics and can degrade auditory perception of the generated sound significantly, especially when the signal level is low. Fig. 30 shows a simulated signal with crossover distortion. It contains both odd and even harmonics as follows (Original data are quoted from Keith Howard’s article “Weighting Up”).

Order N	Frequency (Hz)	Relative Amplitude	Initial Phase (°, sine summation)	Initial Phase (°, cosine summation)
1	1000	1	0	0
2	2000	0.000398	90	-180
3	3000	0.056234	180	0
4	4000	0.00075	-90	-180
5	5000	0.025119	180	-180
6	6000	0.001334	90	-180
7	7000	0.015849	180	0
8	8000	0.001334	-90	-180
9	9000	0.008913	180	-180
10	10000	0.001259	90	-180
11	11000	0.00631	180	0
12	12000	0.001	-90	-180
13	13000	0.004217	180	-180
14	14000	0.001	90	-180
15	15000	0.002371	180	0
16	16000	0.00075	-90	-180
17	17000	0.001	180	-180
18	18000	0.00075	90	-180
19	19000	0.000422	180	0
20	20000	0.000562	-90	-180

The multitone function of Multi-Instrument's Signal Generator uses sine summation. The above table also provides initial cosine phases if cosine summation is used. They can be derived from the initial sine phases by adding  $(N-1) \times 90^\circ$ . It can be seen from the initial cosine phases that the crossover distortion is characterized by alternating odd harmonic polarities while even harmonic polarities remain unchanged (Note:  $0^\circ$  Positive Polarity,  $-180^\circ$  Negative Polarity). In Fig. 31, the blue waveform is the original test signal and the red one is the distortion residual. Their phase relationship can be clearly seen in the figure. The distortion has its maximum value around the zero-crossing points.

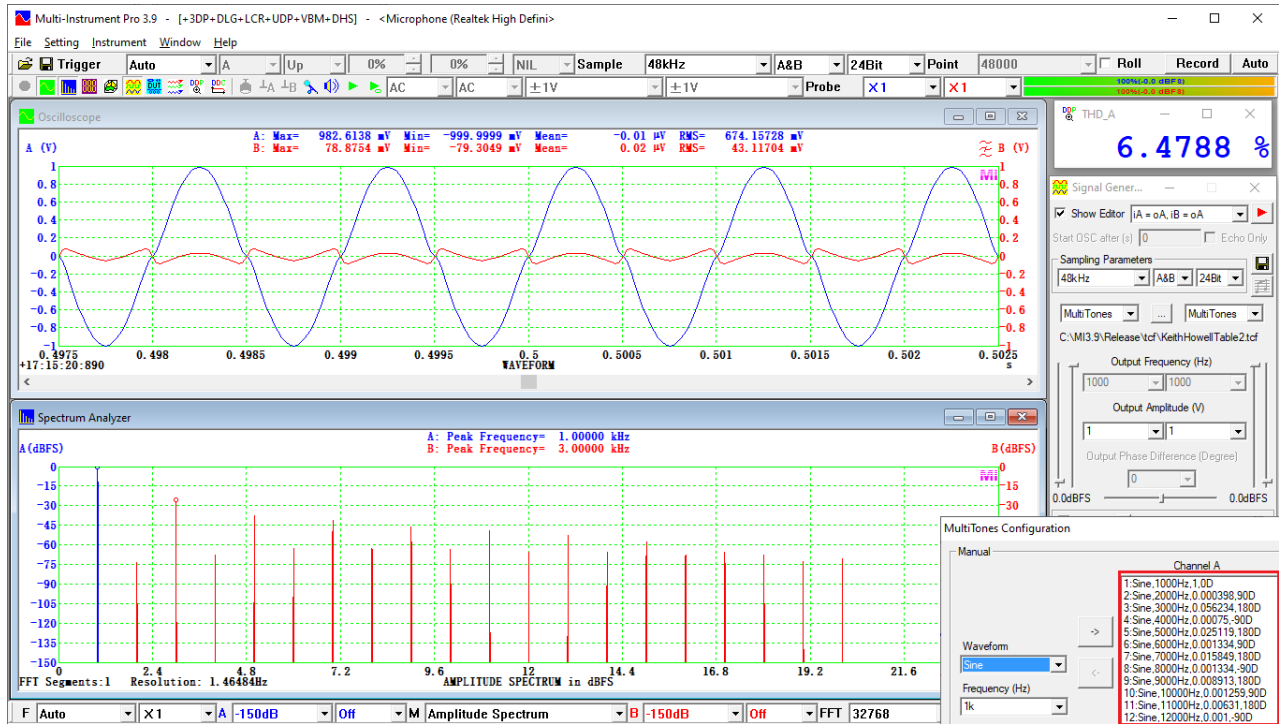


Fig. 30 Crossover Distortion Residual (Simulated, Fundamental Removed via FIR)

Fig. 31 shows the transfer characteristic between the input and the output based on the above signal. The input is an ideal 1 kHz sine wave. The output contains its harmonics additionally according to the aforementioned table. The crossover distortion around the zero-crossing point can be seen clearly in the figure.

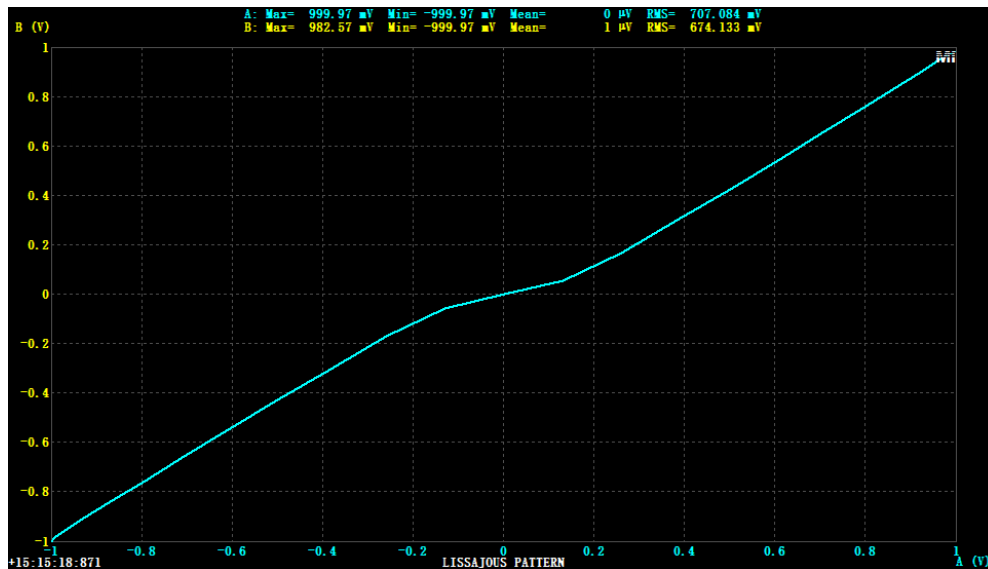


Fig. 31 Input-Output Plot of Crossover Distortion (Simulated)

### 6.1.4 Clipping Distortion

Clipping distortion occurs when the input level is too high for the output to reach what it is supposed to reach. Fig. 32 shows a real example of clipping distortion. The blue waveform is the

original test signal and the red one is the distortion residual. Their phase relationship can be clearly seen in the figure. The distortion is dominated by odd-order harmonics.

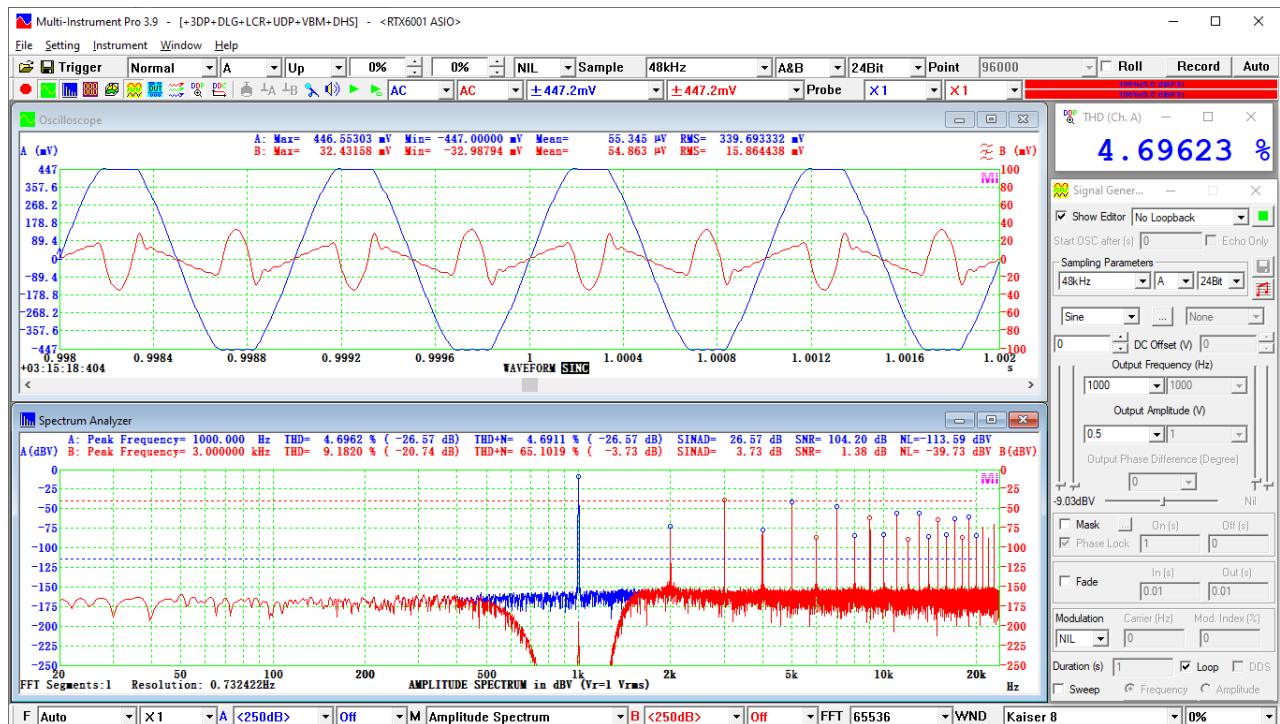


Fig. 32 Clipping Distortion Residual (Real Test, Fundamental Removed via FIR)

Fig. 33 shows the transfer characteristic between the input and the output based on the above signal. The input is an ideal 1 kHz sine wave. The output is its clipped signal.

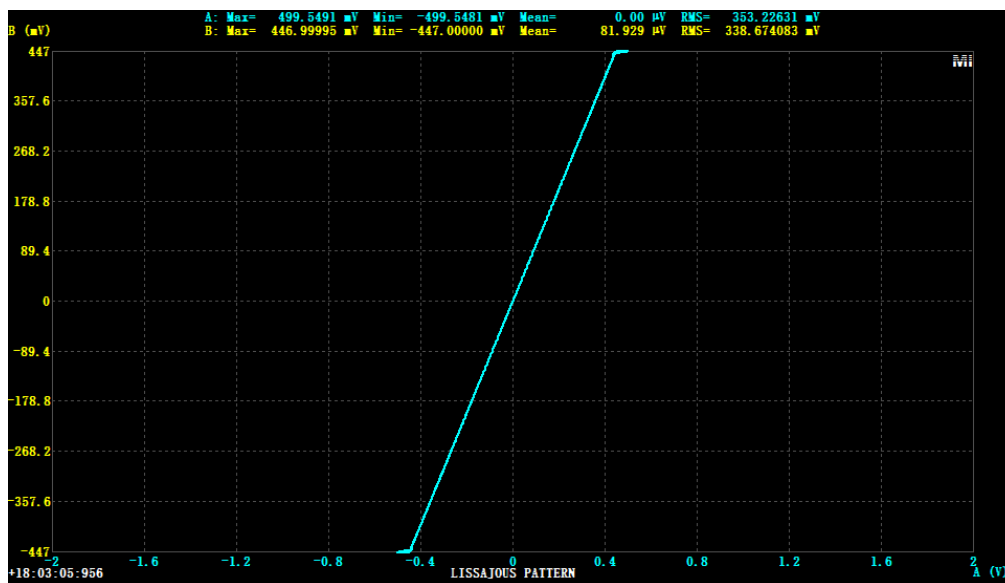


Fig. 33 Input-Output Plot of Clipping Distortion (Real Test)

It should be noted that only in an ideal memory-less system which has no frequency dependence or memory in time and thus has an infinite bandwidth, can the instantaneous output  $y$  be determined uniquely by the instantaneous input  $x$ . A real-world system is always bandlimited and the mapping between  $x$  and  $y$  may not be one-to-one, i.e. the X-Y curve may not be a single-valued function.

## 6.2 Method 2: Synthesizing Based on Harmonic Decomposition Results

The distortion residual obtained by removing the fundamental using a FIR digital filter contains both harmonics and noise. The distortion waveform characteristics can be submerged under the noise floor if the noise floor is high. This problem can be avoided if the distortion residual waveform is synthesized from the measured harmonics instead.

The THD test signal can be decomposed into the sum of the fundamental and harmonics with different amplitudes and initial phases using FFT. Distortion residual waveform can be synthesized by removing the fundamental from the sum.

### 6.2.1 Step 1: Harmonic Decomposition into a Multitone Configuration File

Fig. 34 shows the same test signal as Fig. 32. All settings in Fig.34 are the same as those in Fig. 32 except that the FIR digital filter is not used. Therefore both channels contain identical sampled data and analysis results. Click the DDP Viewer button in the second toolbar from the top, then click the DDP Array Viewer button in the DDP Viewer Configuration Page and select the report titled “A&B-Harmonic Frequencies, RMS, Phases”, the harmonic decomposition information of the test signal will be shown. Note that in Multi-Instrument, the condition  $[Record\ Length]/2 < [FFT\ Size] \leq [Record\ Length]$  must be satisfied in order for the software to calculate the phase of each frequency component. It is met here as  $[Record\ Length] = 96000$  and  $[FFT\ Size] = 65536$ . Right click anywhere within the above report and select “DDP Array Viewer Export As Multi-tone Configuration File” to save the report as a TCF (Tone Configuration File) file.

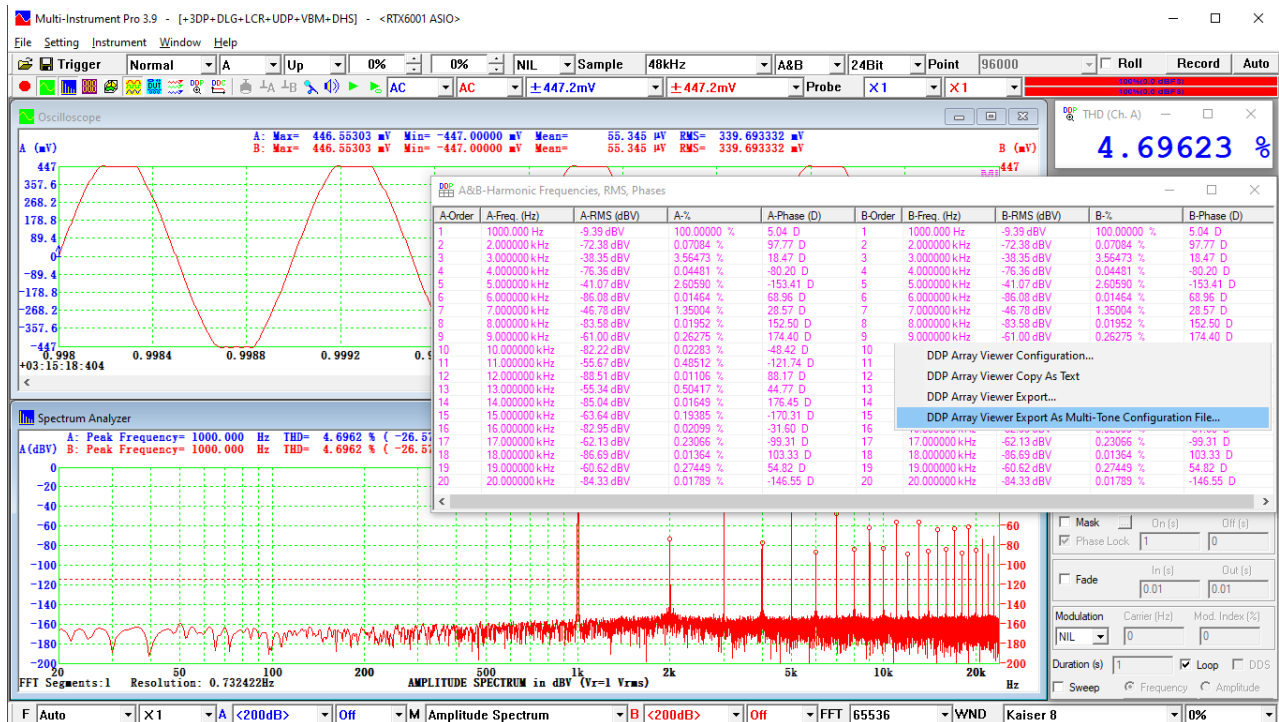


Fig. 34 Harmonic Decomposition of the Test Signal (Real Test)

## 6.2.2 Step 2: Distortion Residual Waveform Synthesis using the Multitone Configuration File

In the Multitone Configuration dialog of the Signal Generator, load the saved TCF file. Remove the fundamental in Ch.B, and then generate the signal and feed it into the Oscilloscope and Spectrum Analyzer via software loopback mode “iA=oA, iB=oB” (see Fig. 35). In Fig. 35, the blue waveform is the synthesized test signal and the red one is the synthesized distortion residual. Comparing Fig. 35 with Fig. 32, it can be seen that the distortion residual waveforms are almost the same except that the one in Fig. 35 does not contain any noise.

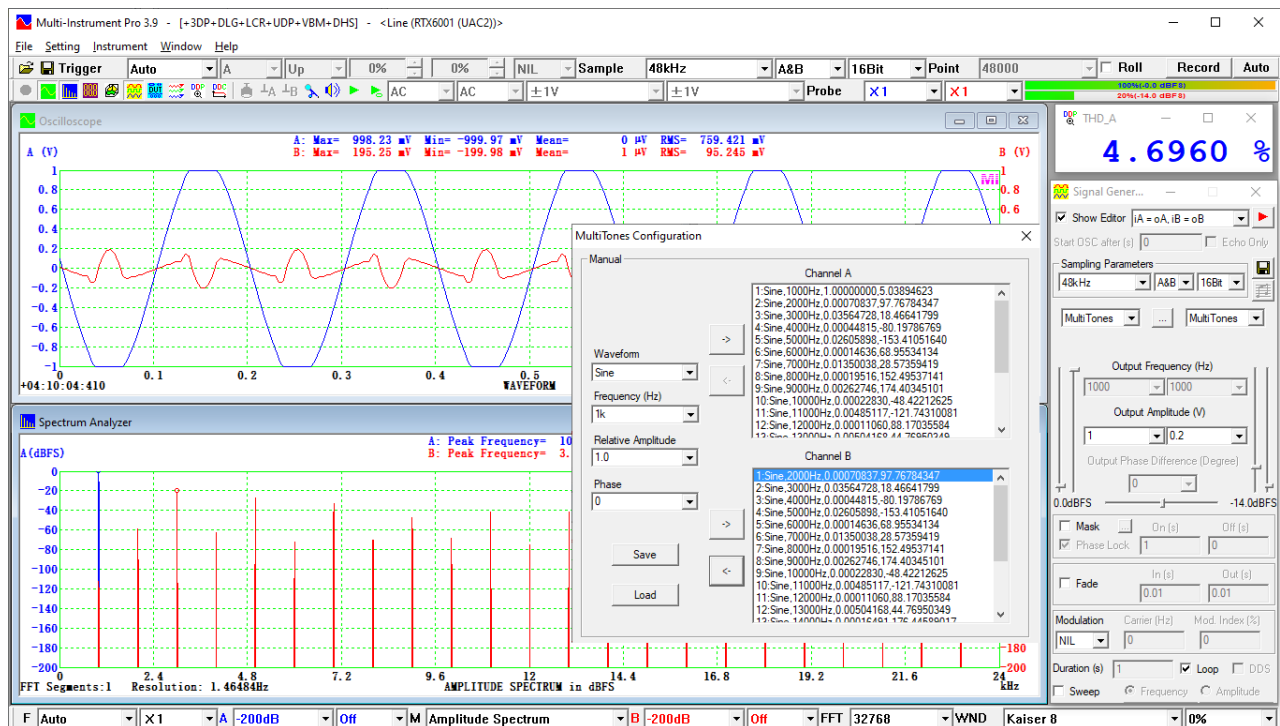


Fig. 35 Distortion Residual Synthesis

## 7. THD and THD+N vs Frequency or Amplitude (Power) Curve

THD and THD+N vs Frequency curve can be obtained using frequency stepped sine stimulus. Similarly, THD and THD+N vs Amplitude curve can be obtained using amplitude stepped sine stimulus. If the load is specified, then the curve can be converted to THD and THD+N vs Power curve. All these functions can be realized using Device Test Plan in Multi-Instrument. The following figure shows the THD+N vs Frequency Curve measured in a RTX6001 hardware loopback test. It can be seen that its THD+N is about 0.0003% from 20Hz to 20kHz.

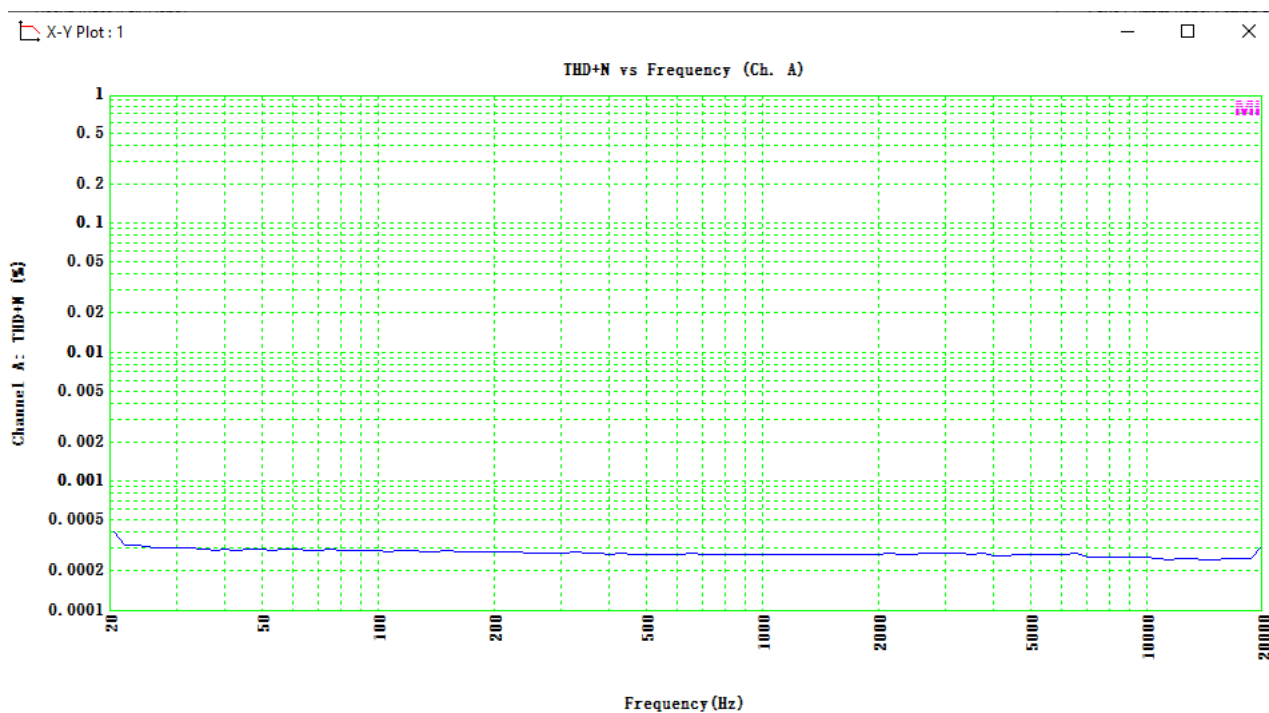


Fig. 36 RTX6001 Hardware Loopback Test: THD+N vs Frequency Curve

THD vs Output Amplitude (in dBFS, 0dBFS@1.414V) curve measured in a RTX6001 hardware loopback test is shown in Fig. 37. The THD value starts from about 50% at the output amplitude -130 dBFS, drops constantly to about 0.0001% at the output amplitude -13 dBFS, remains more or less unchanged until the output amplitude -3dBFS where it starts to rise again. At the maximum output amplitude 0 dBFS, the THD value becomes about 0.0004%.

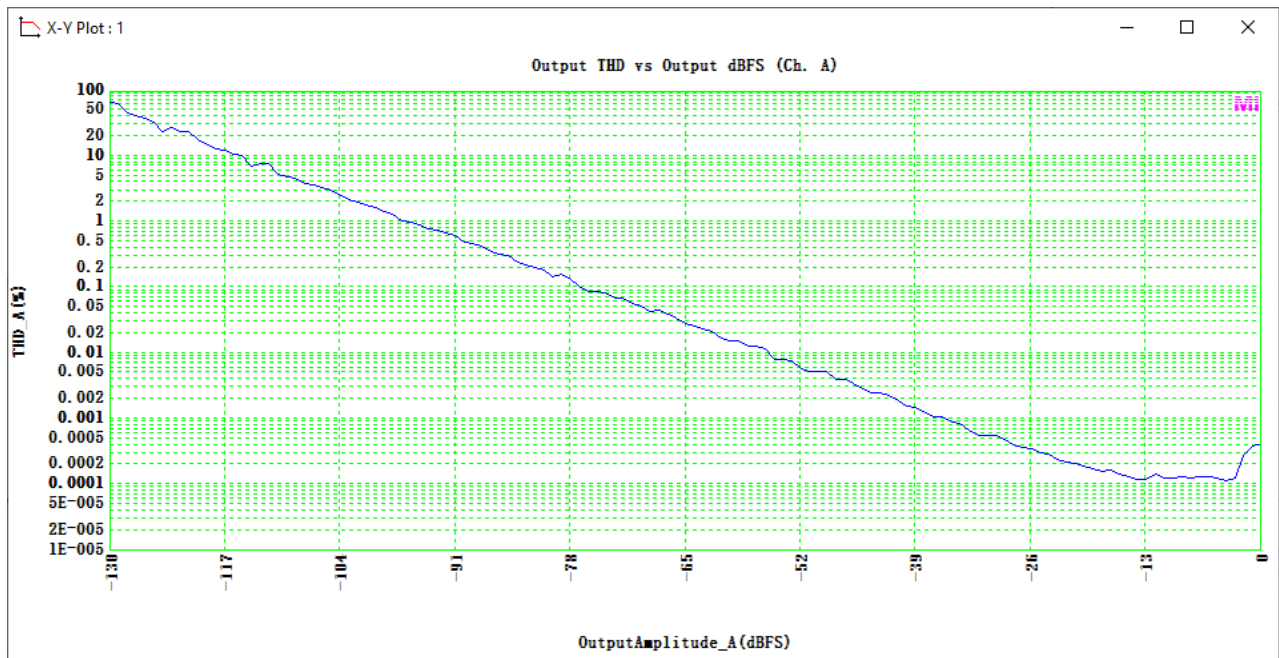


Fig. 37 RTX6001 Hardware Loopback Test : THD vs Output Amplitude (in dBFS, 0dBFS@1.414V) Curve

## 8. Common Misunderstandings of Harmonic Distortion Spectrum

Fig.38 shows a typical harmonic distortion spectrum in the Spectrum Analyzer window. The test signal is a 1 kHz sine wave with an amplitude of 1V. [Record Length] = 48000 and [FFT Size] = 32768, thus no zero padding is used and the extra data of (48000-32768) = 15232 samples is not used in FFT.

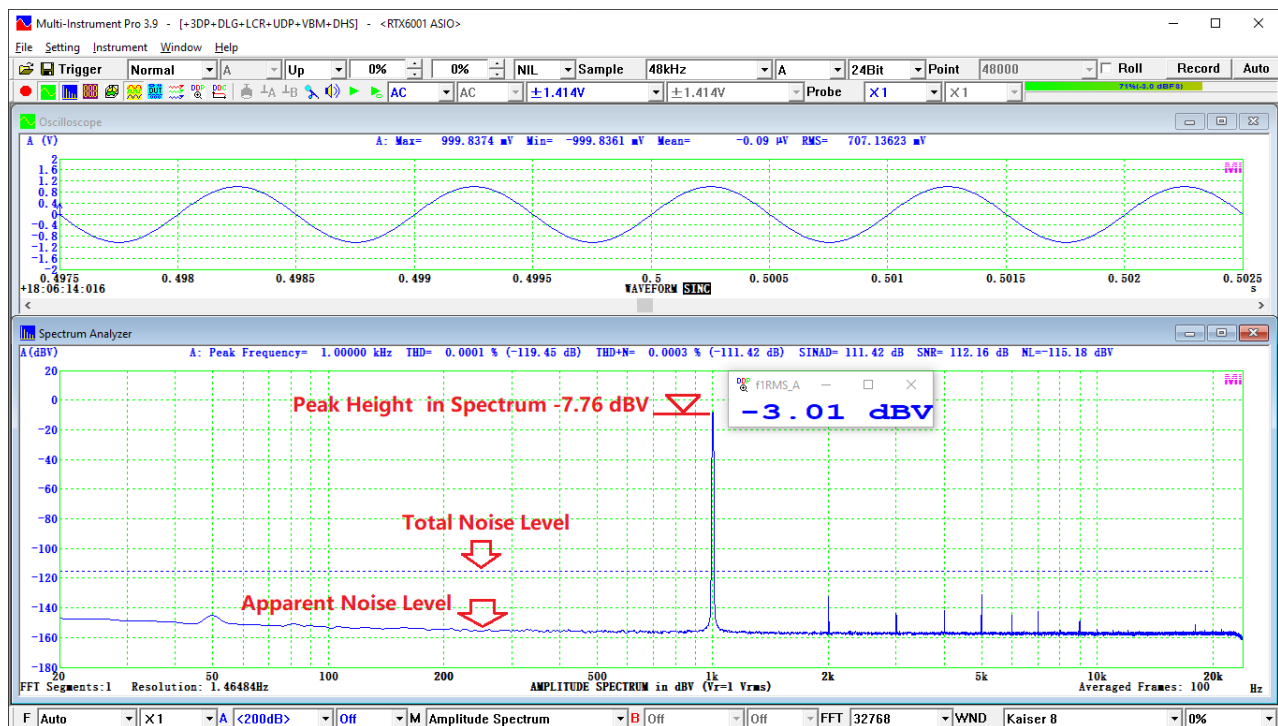


Fig. 38 Typical Harmonic Distortion Spectrum

### 8.1 Noise Level

The “apparent noise level” in Fig. 38 is often mistakenly thought to be the “actual noise level”. In fact, it represents the noise energy contained in each FFT bin, and thus varies with the FFT bin width (i.e. FFT frequency resolution), which is equal to [Sampling Rate]/[FFT Size]. The finer the FFT frequency resolution, the lower the “apparent noise level”. Note that, to keep things simple here, only FFT analysis without zero padding is discussed in this section. The effect of zero padding will be described separately later. For white noise which has a uniform energy distribution along the frequency axis, under the same sampling rate, if the FFT size is doubled, the FFT bin width will be halved, the energy contained inside a FFT bin will also be halved, the “apparent noise level” drops by 3dB as a result. Therefore, when comparing the apparent noise levels in different harmonic distortion spectra, the FFT frequency resolution must be specified. Similarly, it does not make sense to say a harmonic is x dB above the “apparent noise level” without mentioning the FFT frequency resolution. Due to this reason, in Multi-Instrument, the FFT frequency resolution is clearly indicated at the lower left corner of the spectrum graph. Fig. 39 superimposes three harmonic distortion spectra of the same device, with different FFT sizes: 4096, 65536, 1048576 and without zero padding.



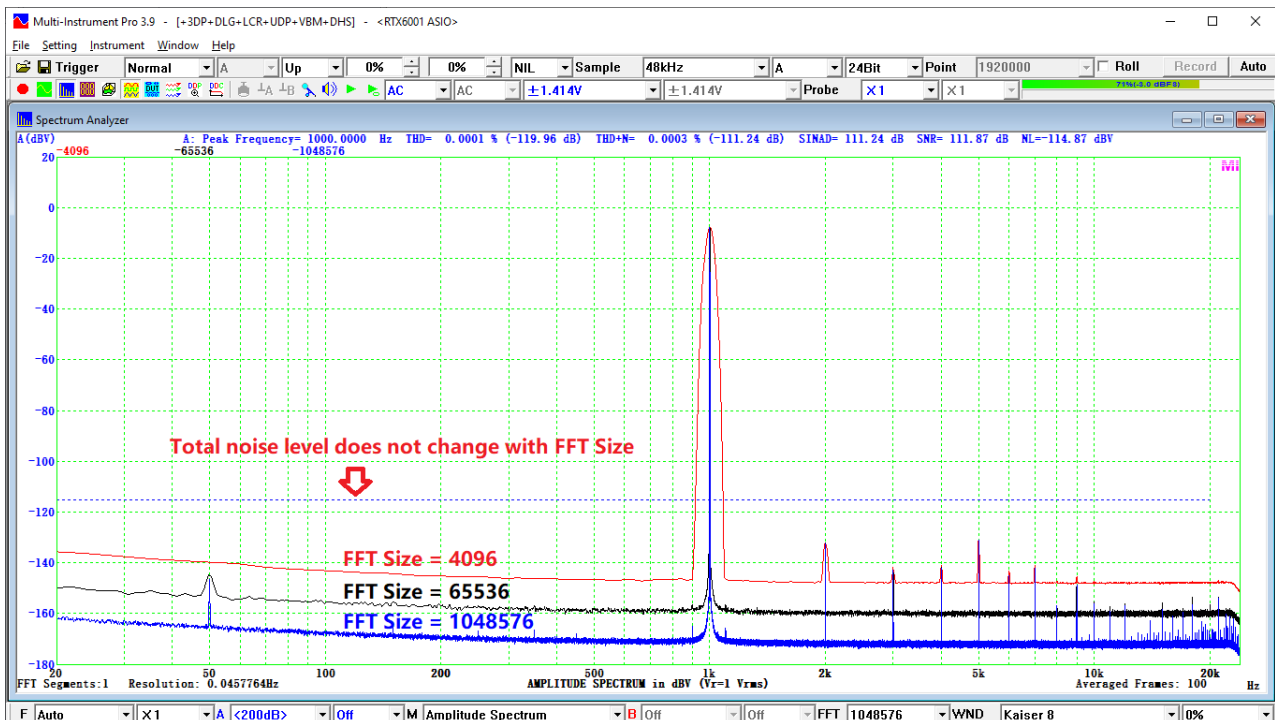


Fig. 39 Comparison of Harmonic Distortion Spectra with Different FFT Sizes (No Zero Padding)

Some important conclusions can be drawn from the above figure (Under the conditions of a fixed sampling rate and no zero padding):

- (1) “Apparent noise level” drops as FFT size increases. Every time FFT size is doubled, “apparent noise level” drops 3 dB in order to keep the noise energy per Hz constant. Thus when FFT Size is changed from 4096 to 65536 and then 1048576, the “apparent noise level” drops 12 dB each time and 24 dB in total. This can be seen in the figure.
- (2) Unlike the wide-band noise, the height of a periodic single-frequency peak does not change with FFT size. This is because its energy will always stay in the FFT bin (s) where the single-frequency is located, and will not be distributed evenly along the frequency axis. It can be observed in the figure that the heights of the fundamental and harmonics remain unchanged in the three cases. The width of a single-frequency peak is due to spectral leakage and is the result of convolution between the frequency peak and the window function spectrum. It decreases as FFT size increases in order to conserve the energy of the single-frequency component.
- (3) (1) and (2) shows that a larger FFT size increases the FFT frequency resolution, decreases the “apparent noise level” but does not reduce the heights of the fundamental and harmonics as well as other single-frequency components. This technique is often used to make the small-amplitude periodic components stand out above the noise level. It can be seen in the above figure that the 50Hz hum and the higher-order harmonics of 1 kHz become discernible and then prominent as FFT size increases. Hence, a large FFT size is often used in audio THD measurement in order to observe the smallest harmonic distortion.

- (4) The “total noise level” does not change with FFT size. Due to this reason, in Multi-Instrument, it is drawn as a dotted line in the spectrum graph, with its coverage of the frequency axis representing the frequency range used in the THD and noise measurement. To configure this range, right click anywhere within the Spectrum Analyzer window and select [Spectrum Analyzer Processing], as shown in Fig. 40. It is also possible to specify the highest harmonic order used in the THD calculation.

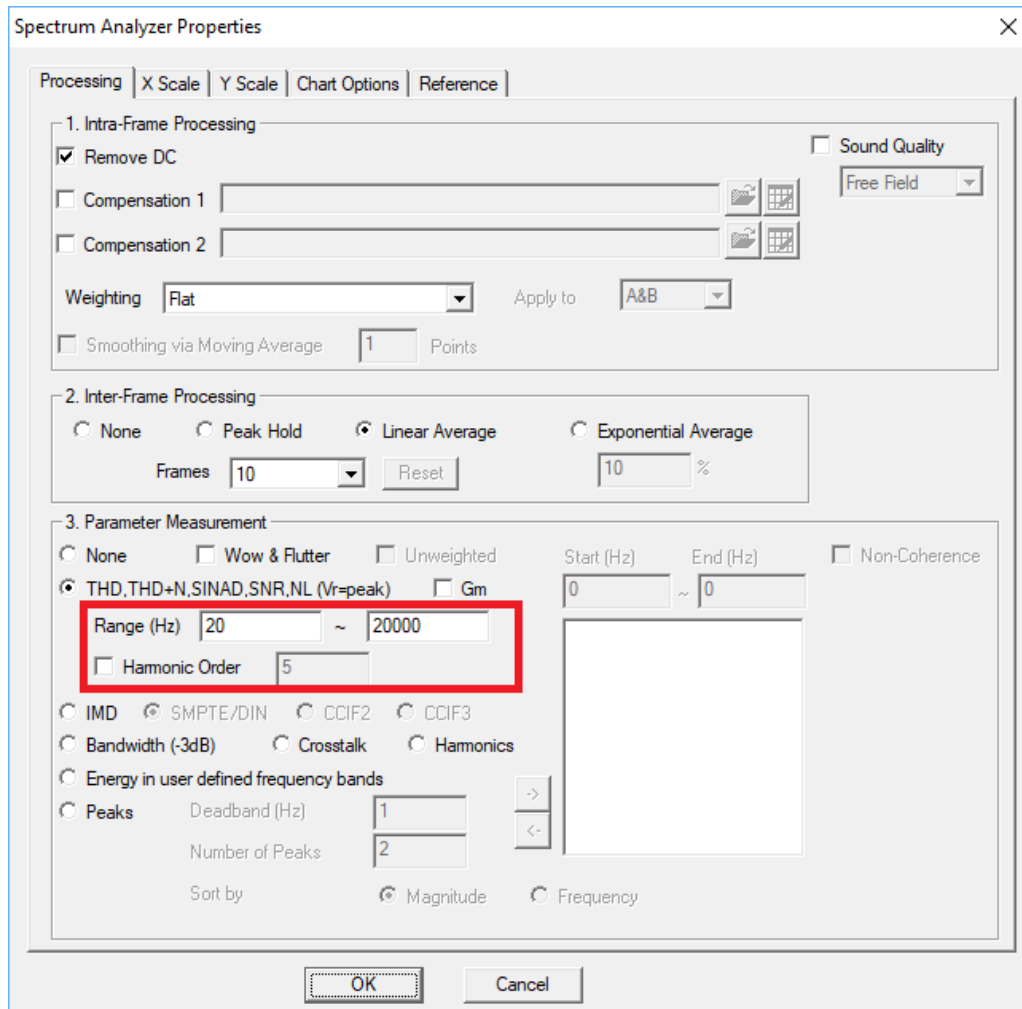


Fig. 40 Frequency Range and Highest Harmonic Order Configuration for THD and Noise Calculation

Multi-Instrument also supports power spectral density function which is independent of FFT frequency resolution. It can be used if only noise distribution along the frequency axis is of interest. The power spectral density is obtained by normalizing the energy in each FFT bin with the bin width. Its value represents the energy per Hz. Its unit can be dBV/Hz, dBu/Hz, dB/Hz, or dBFS/Hz. To convert the spectrum display to power spectral density display, right click anywhere within the Spectrum Analyzer window, select [Spectrum Analyzer Y Scale] and tick “Energy Per Hz” (see Fig. 41).

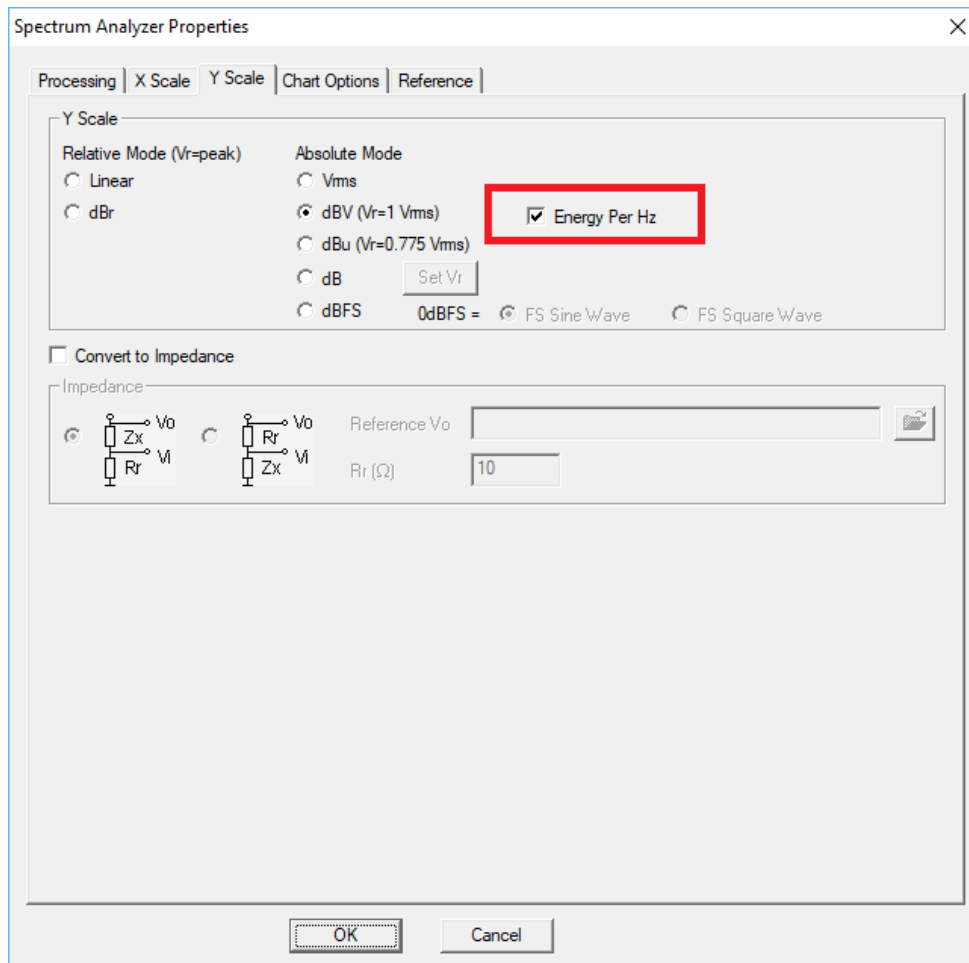


Fig. 41 Power Spectral Density Option

Note that the concept of power spectral density cannot be applied to a periodic signal. Otherwise, the energy of its single-frequency components will be normalized by the FFT bin width without a physical meaning. It will falsely make the height of a single-frequency component FFT-frequency-resolution dependent, as shown in Fig. 42. The figure is similar to Fig. 39 but with “Energy per Hz” option ticked. Although the “apparent noise level” does not change with the FFT size this time, the heights of the fundamental and harmonics as well as other single-frequency components do, which is misleading. Therefore, power spectral density graph should be used only for wide-band noises.

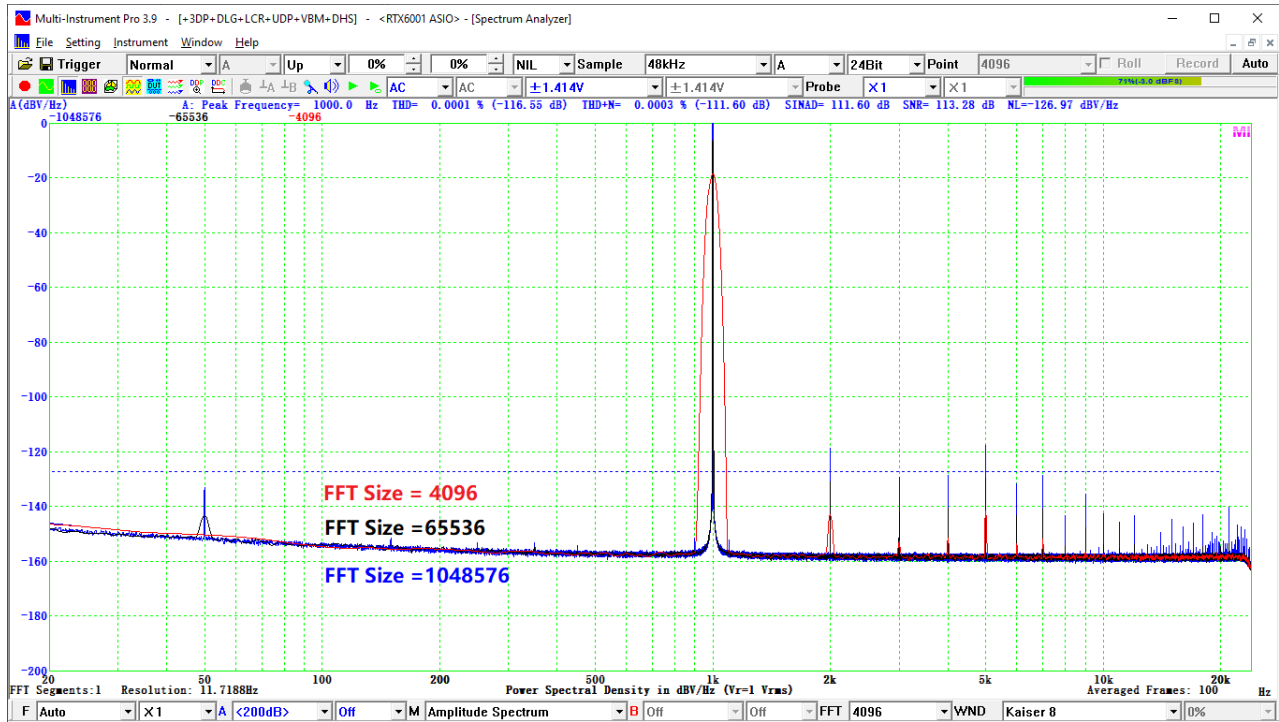


Fig. 41 Comparison of Harmonic Distortion Power Spectral Density with Different FFT Sizes (No Zero Padding)

## 8.2 Heights of Fundamental and Harmonics

Only when the aforementioned full-cycle sampling condition is met does the height of a single-frequency peak in a spectrum graph represent its true RMS value. Otherwise, spectral leakage will be unavoidable and the peak height will always be lower than its actual value. Using Fig. 38 as an example, the actual RMS value of the 1 kHz 0.707 V<sub>rms</sub> fundamental is -3.01 dBV. But the peak height in the spectrum graph is only -7.76 dBV. This may give you a wrong impression that something is wrong with the software. Actually, it simply reflects the fact that the energy of the single-frequency periodic component is not contained in one FFT bin due to spectral leakage. As long as a proper window function is used, the measurement of THD and its related parameters will not be affected, and the software is able to calculate the actual energy of a single-frequency component correctly. In Fig. 38, the DDP viewer displays the correct RMS value -3.01 dBV using DDP: f1RMS\_A(EU).

It is possible to display the correct heights of the fundamental and harmonics in the spectrum graph even under spectral leakage, by right clicking anywhere within the Spectrum Analyzer window and select [Spectrum Analyzer Chart Options]> “Mark Peaks”, as shown below. Note that this is only a cosmetic correction and will not alter the spectrum data which must conserve the energy of the time-domain data after FFT according to Parseval's theorem.

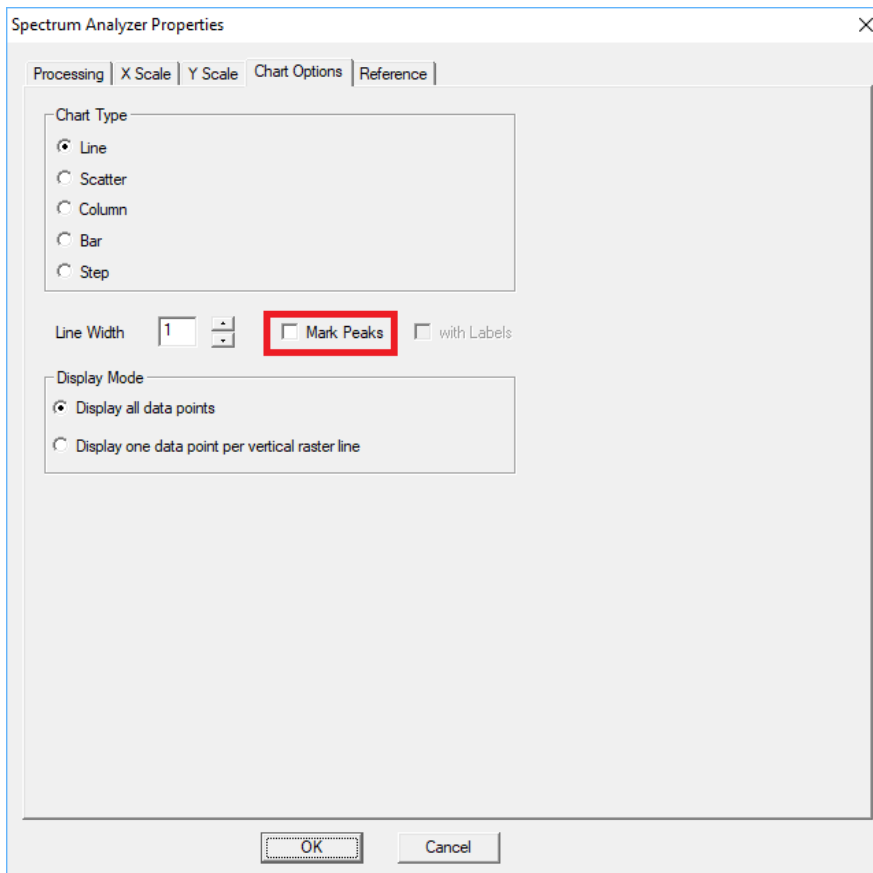


Fig. 42 “Mark Peaks” Option

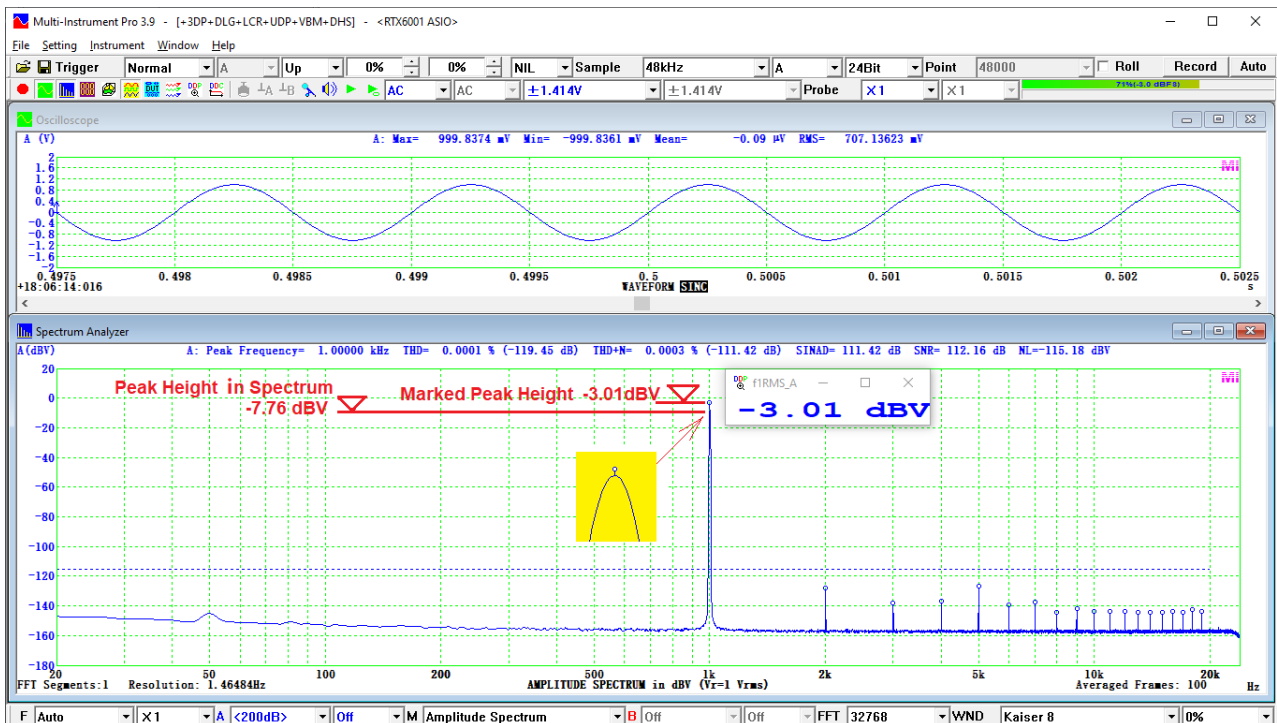


Fig. 43 Spectral Leakage Correction through Marked Peak Heights

### 8.3 Effect of Zero Padding

It is generally not recommended to use zero padding in THD measurement. Zero padding must not be used in full-cycle sampling. It can be used in windowed sampling only. In Multi-Instrument, when the FFT size is greater than the Record Length, [FFT Size]-[Record Length] zeros will be automatically padded at the end of the sampled data before FFT, and “FFT Segments <1” will be indicated at the lower left corner of the spectrum graph. Parseval's theorem is still followed when zero padding is used. That is, the energy sum of all frequency components in the Spectrum Analyzer is always equal to the energy sum of all the samples in the Oscilloscope. After this issue is settled automatically by the software, the remaining impact of zero padding on the harmonic distortion spectrum is illustrated in Fig. 44. Three measurements similar to those in Fig. 39 but with zero padding are superimposed in the figure. In the three cases, Record Length is fixed at 4096 while the FFT size is 4096, 65536 and 1048576, respectively. It can be seen that both the heights of single-frequency components, such as the fundamental and harmonics, and the “apparent noise level” decrease as more zeros are padded. The “apparent noise level” in each case is virtually the same as that in the respective case in Fig. 39, as dictated by energy conservation. But the single-frequency components are wider and lower than those in Fig. 39 in the two cases with zero padding, i.e. FFT size =65536 and 1048576. This is due to the fact that the window function was applied to the sampled data only. The energy of a single-frequency component is still conserved in this case.

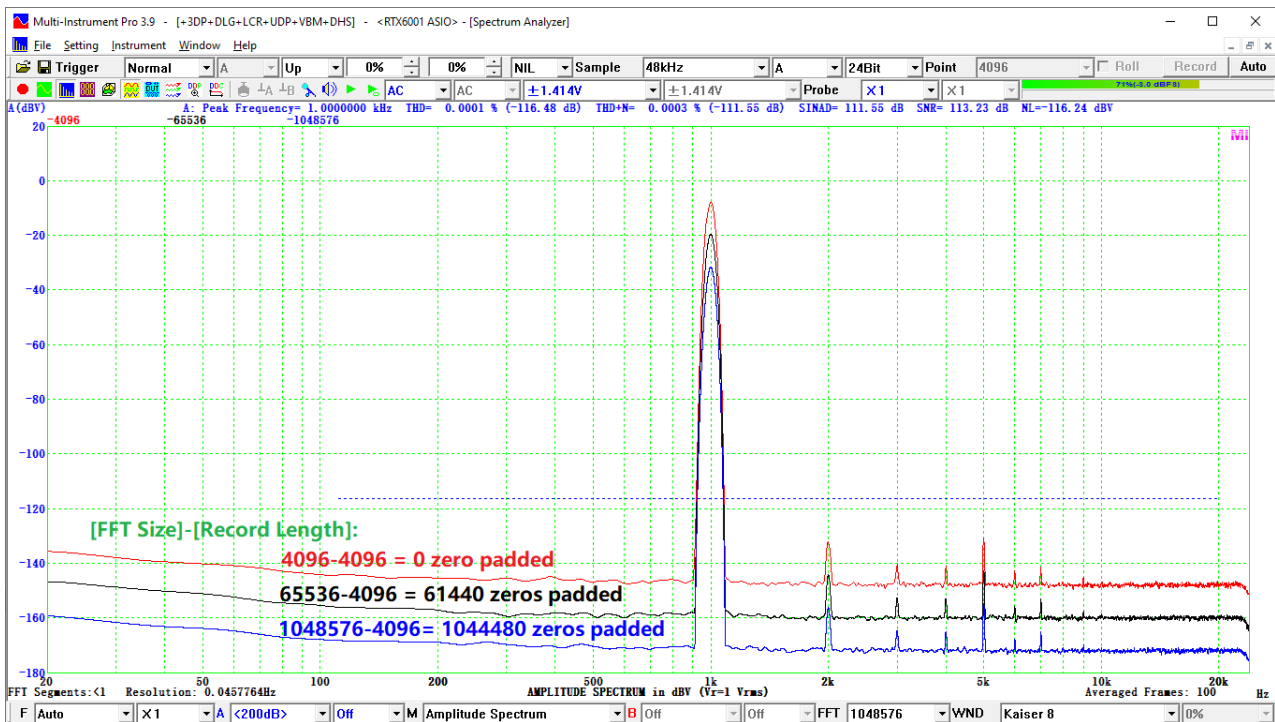


Fig. 44 Comparison of Harmonic Distortion Spectrum with a Fixed Record Length and Different FFT Sizes (i.e. with zero padding)

Despite the above alteration of spectrum by zero padding, Multi-Instrument is still able to calculate THD and its related parameters correctly, and show the heights of the fundamental and harmonics correctly if “Mark Peaks” option is ticked. The energy of a single-frequency component will still be calculated correctly in its respective DDP.

## 8.4 Effect of Removing DC

In Multi-Instrument, if [Spectrum Analyzer Processing]> “Remove DC” option is ticked, the mean value (i.e. DC component) of the data in the Oscilloscope will be subtracted from each sample before FFT in order to remove the DC component in the spectrum. This option is useful when there exists a true DC component in the sampled data and it is not of interest. However, as this mean value is calculated without applying any window function, if a window function other than Rectangle is applied in the Spectrum Analyzer, some DC residual may still exist after the mean subtraction, and under some conditions, the DC residual may be even greater than the case where the “Remove DC” option is not ticked. Fig. 45 shows such an example. The mean value in the oscilloscope is 2.72505 mV with a Record Length of 4096. In contrast, with the same 1 kHz test signal, the mean value in Fig. 43 is only -0.09  $\mu$ V with a Record Length of 48000. The reason for this difference is that the Oscilloscope in Fig. 45 contains a non-integer number of signal cycles and the incomplete cycle contributes most to the non-zero mean value. The non-zero mean value becomes prominent when the Record Length is small. Subtracting this “fake” DC component in the Spectrum Analyzer creates a hump in the vicinity of 0 Hz (see the leftmost part of the red curve in Fig. 45). The hump is the result of the convolution between the DC residual and the window function spectrum. If this DC hump shows up, untick “Remove DC” to see if it helps. In any event, Multi-Instrument is able to calculate the THD and its related parameters correctly with or without this DC hump.

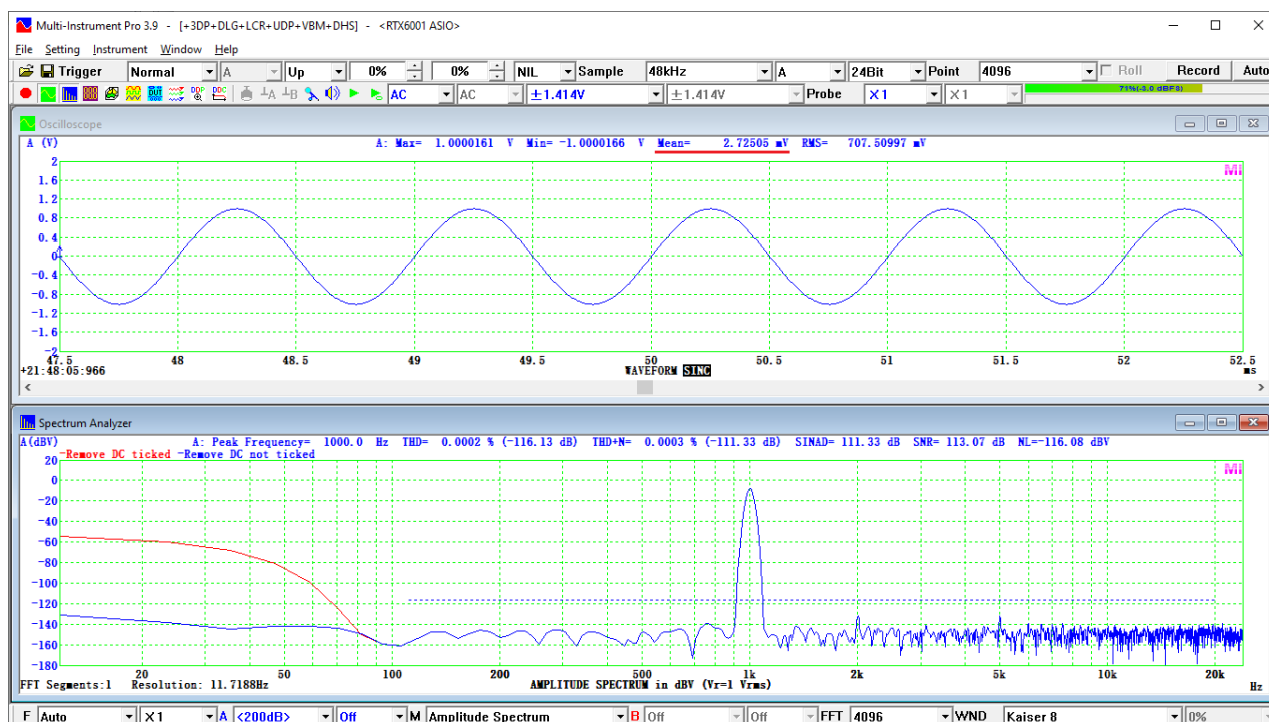


Fig. 45 DC Residual sometimes shows up even with [Spectrum Analyzer Processing]>“Remove DC” option ticked

## 9. Input-Output Linearity Graph

The output amplitude vs input amplitude graph can be used to show a system's linearity intuitively in time domain. It is especially useful in examining audio compressors and limiters for obvious reasons, and testing DA and AD converters in which nonlinearity usually becomes worse at low amplitudes. In Sections 6.1.1~6.1.4, Input-Output Lissajous Plots have been used to show the transfer characteristics between the input and the output. However, that method uses the instantaneous input and output values of a sine wave stimulus, and therefore may not be able to plot a single-valued curve owing to a real-world system's frequency response. The method here uses the input and output amplitudes of a sine wave stimulus instead, with the input amplitude stepped from the lowest, which is equivalent to the noise level, to the highest. Fig. 46 shows a RTX6001 loopback test using Device Test Plan of Multi-Instrument. Its X axis is the specified output amplitude, from -130 dBFS to 0 dBFS (0dBFS@1.414V). Its Y axis is the measured signal level in dBV, obtained from the time-domain RMS value (DDP: RMS\_A(EU)), which represents the total energy of the measured signal including noises. This explains the "Level-off" from -115 dBFS to -130 dBFS. It shows that at low amplitudes, the output signal contains a large portion of noises. From -115 dBFS to 0 dBFS, the relationship is basically linear.

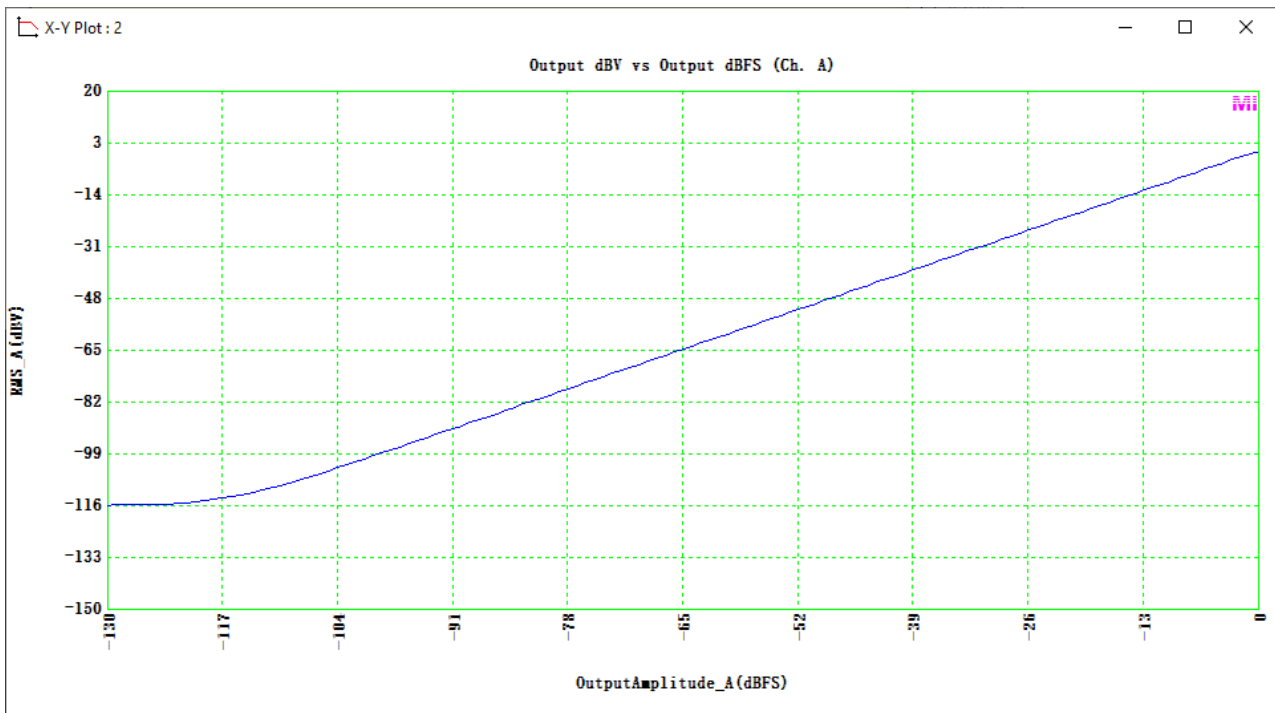


Fig. 46 RTX6001 Hardware Loopback Test: Input-Output Linearity Graph (Wide-band Method)

To eliminate the effect of noises on this linearity measurement, the output signal level can be fetched in frequency domain instead, using the fundamental's RMS amplitude (DDP: f1RMS\_A(EU)). This is equivalent to use a narrow-band filter to reject almost all the wide-band noises. As shown in Fig. 47, the "level-off" disappears and the curve becomes almost linear in the full range from -130 dBFS to 0 dBFS.



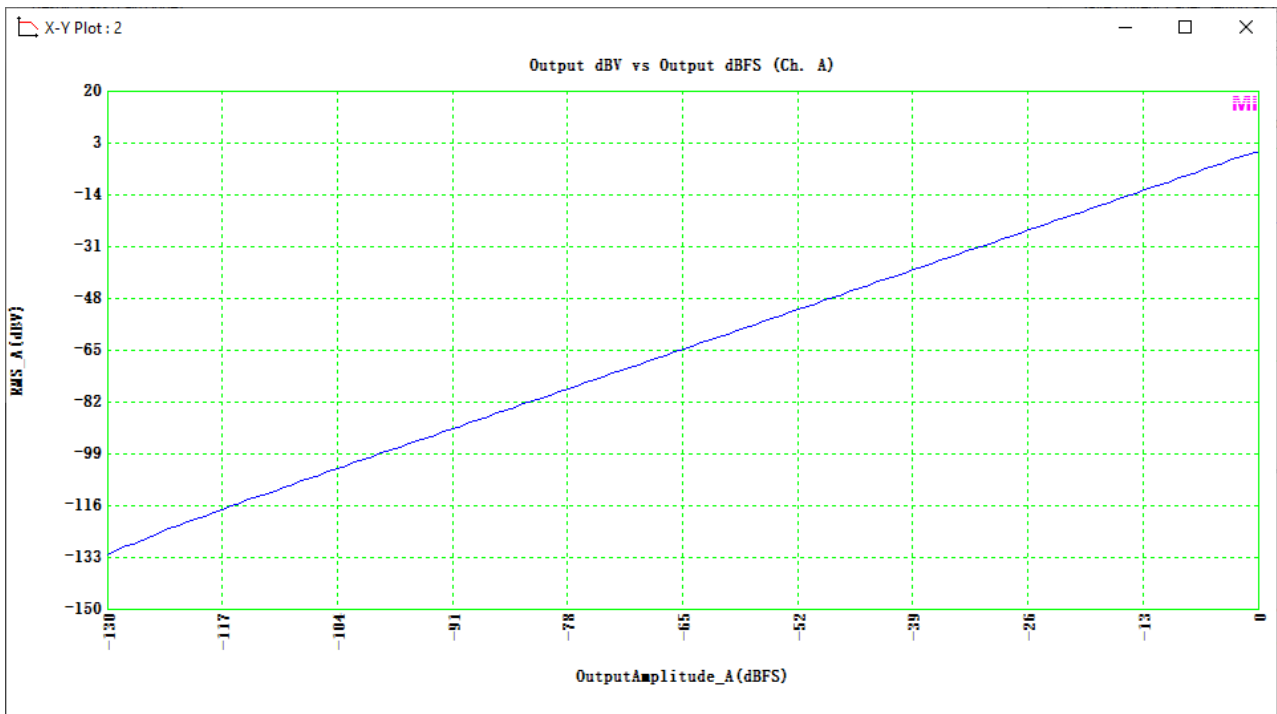


Fig. 47 RTX6001 Loopback: Input-Output Linearity Graph (Narrow-band Method)

# 10. Comparison between Measured THD with Analytical Result for Typical Waveforms

The THD values of some typical waveforms such as square wave, triangle wave, saw tooth, rectangle wave can be derived analytically. It would be interesting to compare the measured results using Multi-Instrument with the analytical ones. In the following sections, the test signals were generated and analyzed through the software loopback mode “iA=oA, iB=oB”. The sampling rate is 48 kHz. The test signal frequency is 100 Hz. The upper frequency limit for THD calculation is set to 1/2 of the sampling rate, i.e. 24 kHz, in order to take into account harmonics as many as possible, because these ideal signals have an infinite number of harmonics. The comparison shows that the two matches excellently well.

## 10.1 Square Wave



Fig. 48 Measured and Analytical THD of A Square Wave

## 10.2 Triangle Wave

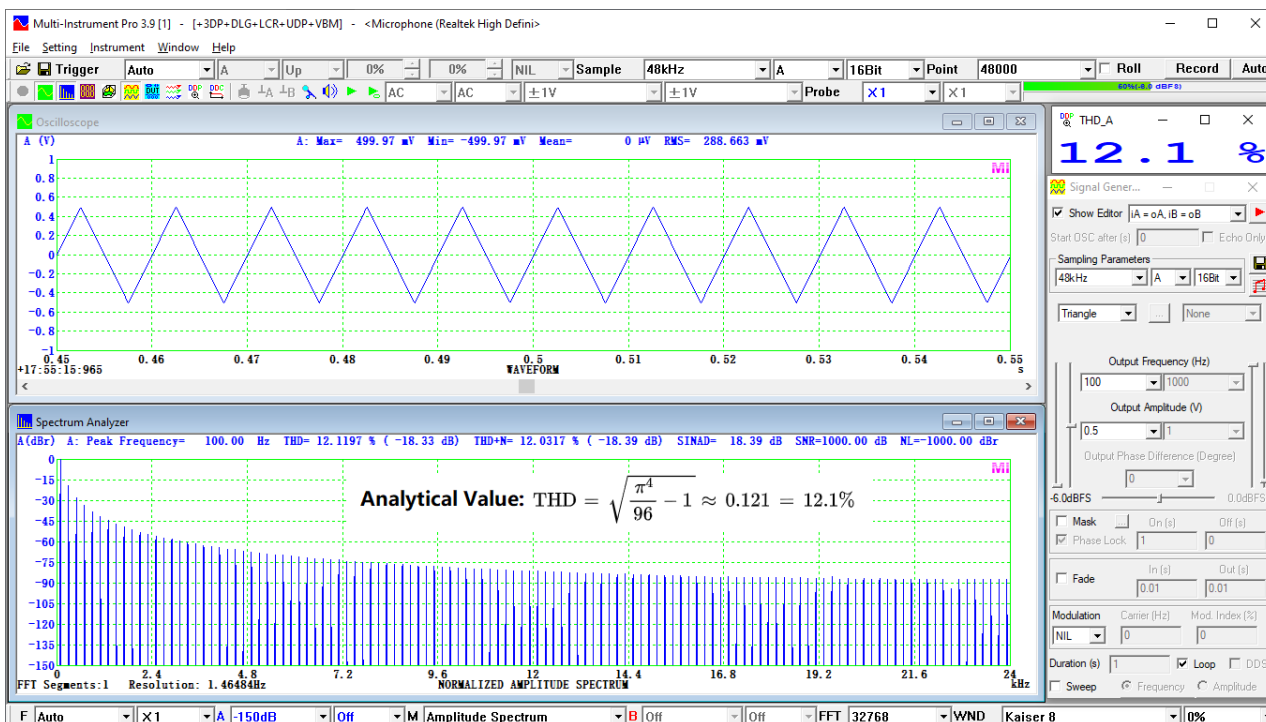


Fig. 49 Measured and Analytical THD of A Triangle Wave

## 10.3 Saw Tooth Wave

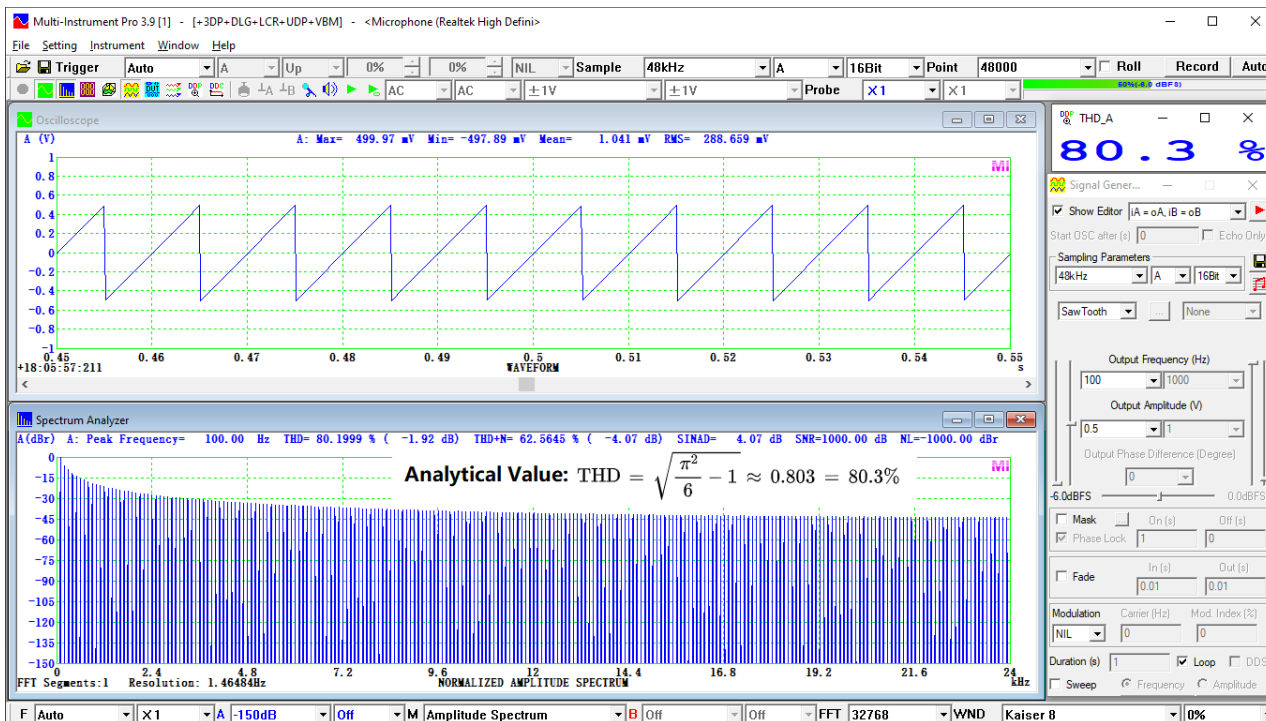


Fig. 50 Measured and Analytical THD of A Saw Tooth Wave

## 10.4 Rectangle Wave



Fig. 51 Measured and Analytical THD of A Rectangle Wave (Duty Cycle =10%)