

SECTION 8

DISTORTION MEASUREMENTS

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SECTION 8

DISTORTION MEASUREMENTS

HIGH SPEED OP AMP DISTORTION

Walt Kester

Dynamic range of an op amp may be defined in several ways. The most common ways are to specify Harmonic Distortion, Total Harmonic Distortion (THD), or Total Harmonic Distortion Plus Noise (THD + N).

Harmonic distortion is measured by applying a spectrally pure sinewave to an op amp in a defined circuit configuration and observing the output spectrum. The amount of distortion present in the output is usually a function of several parameters: the small- and large-signal nonlinearity of the amplifier being tested, the amplitude and frequency of the input signal, the load applied to the output of the amplifier, the amplifier's power supply voltage, printed circuit board layout, grounding, power supply decoupling, etc. Therefore, any distortion specification is relatively meaningless unless the exact test conditions are specified.

Harmonic distortion may be measured by looking at the output spectrum on a spectrum analyzer and observing the values of the second, third, fourth, etc., harmonics with respect to the amplitude of the fundamental signal. The value is usually expressed as a ratio in %, ppm, dB, or dBc. For instance, 0.0015% distortion corresponds to 15ppm, or -96.5dBc. The unit "dBc" simply means that the harmonic's level is so many dB below the value of the "carrier" frequency, i.e., the fundamental.

Harmonic distortion may be expressed individually for each component (usually only the second and third are specified), or they all may be combined in a root-sum-square (RSS) fashion to give the Total Harmonic Distortion (THD). The distortion component which makes up Total Harmonic Distortion is usually calculated by taking the root sum of the squares of the first five or six harmonics of the fundamental. In many practical situations, however, there is negligible error if only the second and third harmonics are included. This is because the RSS process causes the higher-order terms to have negligible effect on the THD, if they are 3 to 5 times smaller than the largest harmonic. For example,

$$\sqrt{0.10^2 + 0.03^2} = \sqrt{0.0109} = 0.104 \quad 0.1$$

DEFINITIONS OF THD AND THD + N

- V_s = Signal Amplitude (rms Volts)
- V_2 = Second Harmonic Amplitude (rms Volts)
- V_n = nth Harmonic Amplitude (rms Volts)
- V_{noise} = rms value of noise over measurement bandwidth
- $$\text{THD} + \text{N} = \frac{\sqrt{V_2^2 + V_3^2 + V_4^2 + \dots + V_n^2 + V_{\text{noise}}^2}}{V_s}$$
- $$\text{THD} = \frac{\sqrt{V_2^2 + V_3^2 + V_4^2 + \dots + V_n^2}}{V_s}$$

Figure 8.1

It is important to note that the THD measurement does not include noise terms, while THD + N does. The noise in the THD + N measurement must be integrated over the measurement bandwidth. In audio applications, the bandwidth is normally chosen to be around 100kHz. In narrow-band applications, the level of the noise may be reduced by filtering. On the other hand, harmonics and intermodulation products which fall within the measurement bandwidth cannot be filtered, and therefore may limit the system dynamic range. It should be evident that the THD+N approximately equals THD if the rms noise over the measurement bandwidth is several times less than the THD, or even the worst harmonic. It is worth noting that if you know only the THD, you can calculate the THD+N fairly accurately using the amplifier's voltage- and current-noise specifications. (Thermal noise associated with the source resistance and the feedback network may also need to be computed). But if the rms noise level is significantly higher than the level of the harmonics, and you are only given the THD+N specification, you cannot compute the THD.

Special equipment is often used in audio applications for a more sensitive measurement of the noise and distortion. This is done by first using a bandstop filter to remove the fundamental signal (this is to prevent overdrive distortion in the measuring instrument). The total rms value of all the other frequency components (harmonics and noise) is then measured over an appropriate bandwidth. The ratio to the fundamental is the THD+N specification.

Audio frequency amplifiers (such as the OP-275) are optimized for low noise and low distortion within the audio bandwidth (20Hz to 20kHz). In audio applications, total harmonic distortion plus noise (THD+N) is usually measured with specialized equipment, such as the Audio Precision System One. The output signal amplitude is measured at a given frequency (e.g., 1kHz); then the fundamental signal is removed with a bandstop filter, and the system measures the rms value of the remaining frequency components, which contain both harmonics and noise. The noise and harmonics are measured over a bandwidth that will include the highest harmonics,

usually about 100kHz. The measurement is swept over the frequency range for various conditions.

THD+N results for the OP-275 are plotted in Figure 8.2 as a function of frequency. The signal level is 3V rms, and the amplifier is connected as a unity-gain follower. The data is shown for three load conditions: 600ohm, 2kohm, and 10kohm. Notice that a THD+N value of 0.0008% corresponds to 8ppm, or -102dBc. The input voltage noise of the OP-275 is typically 6nV/rtHz @ 1kHz, and integrated over an 80kHz noise bandwidth, yields an rms noise level of 1.7μV rms. For a 3V rms signal level, the corresponding signal-to-noise ratio is 125dB. Because the THD is considerably greater than the noise level, the THD component is the primary contributor. Multiple plots with variable bandwidths can be used to help separate noise and distortion.

THD + N FOR THE OP -275 OVER 100kHz BANDWIDTH IS DOMINATED BY DISTORTION

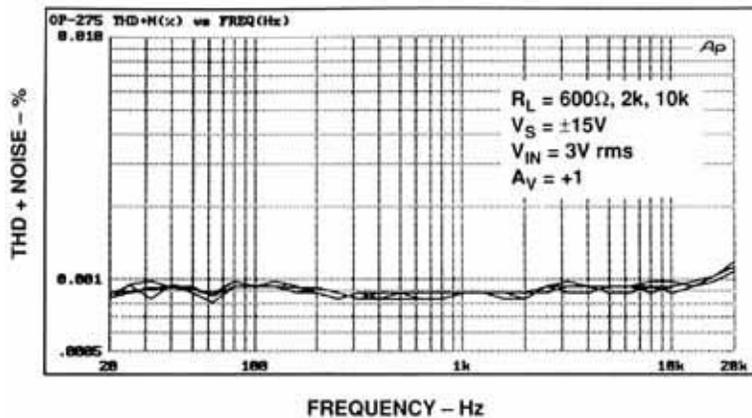


Figure 8.2

Now, consider the AD797, a low noise amplifier (1nV/rtHz) where measurement equipment distortion, and not the amplifier distortion, limits the measurement. The THD specification for the AD797 is 120dBc @ 20kHz, and a plot is shown in Figure 8.3. The distortion is at the limits of measurement of available equipment, and the actual amplifier noise is even lower by 20dB. The measurement was made with a spectrum analyzer by first filtering out the fundamental sinewave frequency ahead of the analyzer. This is to prevent overdrive distortion in the spectrum analyzer. The first five harmonics were then measured and combined in an RSS fashion to get the THD figure. The legend on the graph indicates that the measurement equipment "floor" is about 120dBc; hence at frequencies below 10kHz, the THD may be even less.

THD OF THE AD797 OP AMP SHOWS MEASUREMENT LIMIT AT -120dBc, WHILE AMPLIFIER NOISE FLOOR IS AT -140dBc (1nV/√Hz INTEGRATED OVER 100kHz BANDWIDTH)

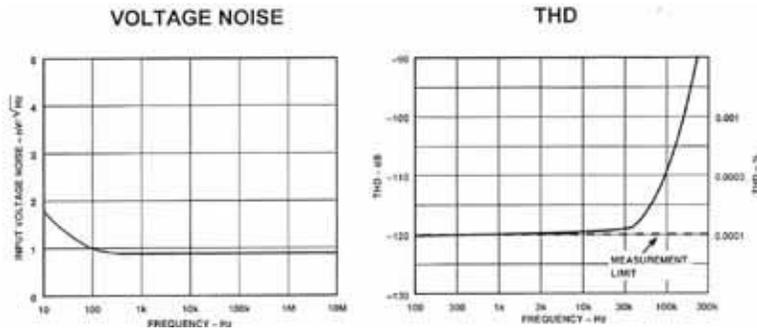


Figure 8.3

To calculate the AD797 noise, multiply the voltage noise spectral density (1nV/rtHz) by the square root of the measurement bandwidth to yield the device's rms noise floor. For a 100kHz bandwidth, the noise floor is 316nV rms, corresponding to a signal-to-noise ratio of about 140dB for a 3V rms output signal.

Rather than simply examining the THD produced by a single tone sinewave input, it is often useful to look at the distortion products produced by two tones. As shown in Figure 8.4, two tones will produce second and third order intermodulation products. The example shows the second and third order products produced by applying two frequencies, f_1 and f_2 , to a nonlinear device. The second order products located at $f_2 + f_1$ and $f_2 - f_1$ are located far away from the two tones, and may be removed by filtering. The third order products located at $2f_1 + f_2$ and $2f_2 + f_1$ may likewise be filtered. The third order products located at $2f_1 - f_2$ and $2f_2 - f_1$, however, are close to the original tones, and filtering them is difficult. Third order IMD products are especially troublesome in multi-channel communications systems where the channel separation is constant across the frequency band.

SECOND AND THIRD-ORDER INTERMODULATION PRODUCTS FOR $f_1 = 5\text{MHz}$ and $f_2 = 6\text{MHz}$

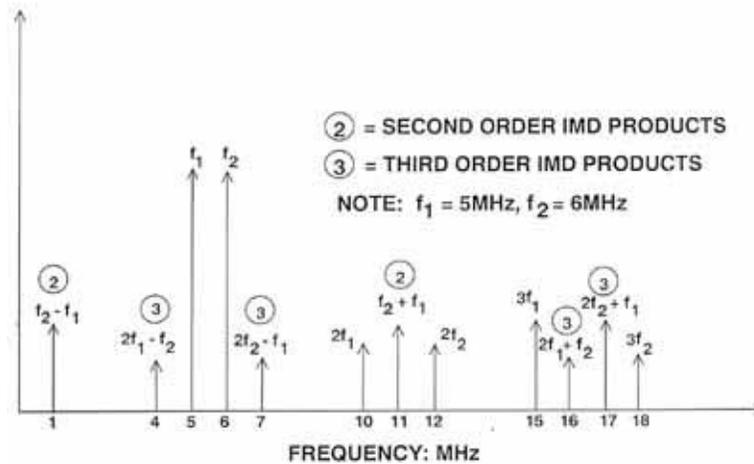


Figure 8.4

Intermodulation distortion products are of special interest in the RF area, and a major concern in the design of radio receivers. Third-order IMD products can mask out small signals in the presence of larger ones. Third order IMD is often specified in terms of the *third order intercept* point as shown in Figure 8.5. Two spectrally pure tones are applied to the system. The output signal power in a single tone (in dBm) as well as the relative amplitude of the third-order products (referenced to a single tone) is plotted as a function of input signal power. If the system non-linearity is approximated by a power series expansion, the second-order IMD amplitudes increase 2dB for every 1dB of signal increase. Similarly, the third-order IMD amplitudes increase 3dB for every 1dB of signal increase. With a low level two-tone input signal, and two data points, draw the second and third order IMD lines as are shown in Figure 8.5, because one point and a slope determine each straight line.

Once the input reaches a certain level, however, the output signal begins to soft-limit, or compress. But the second and third-order intercept lines may be extended to intersect the extension of the output signal line. These intersections are called the *second- and third order intercept points*, respectively. The values are usually referenced to the output power of the device expressed in dBm. Another parameter which may be of interest is the *1dB compression point*. This is the point at which the output signal is compressed by 1dB from the ideal input/output transfer function. This point is also shown in Figure 8.5.

INTERCEPT POINTS, GAIN COMPRESSION, AND IMD

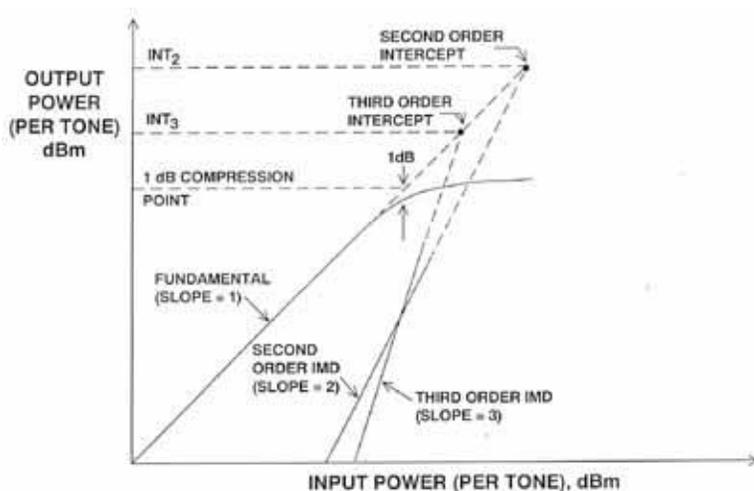


Figure 8.5

Knowing the third order intercept point allows calculation of the approximate level of the third-order IMD products as a function of output signal level. Figure 8.6 shows the third order intercept value as a function of frequency for the AD9622 voltage feedback amplifier.

**AD9622 THIRD ORDER IMD INTERCEPT
VERSUS FREQUENCY**

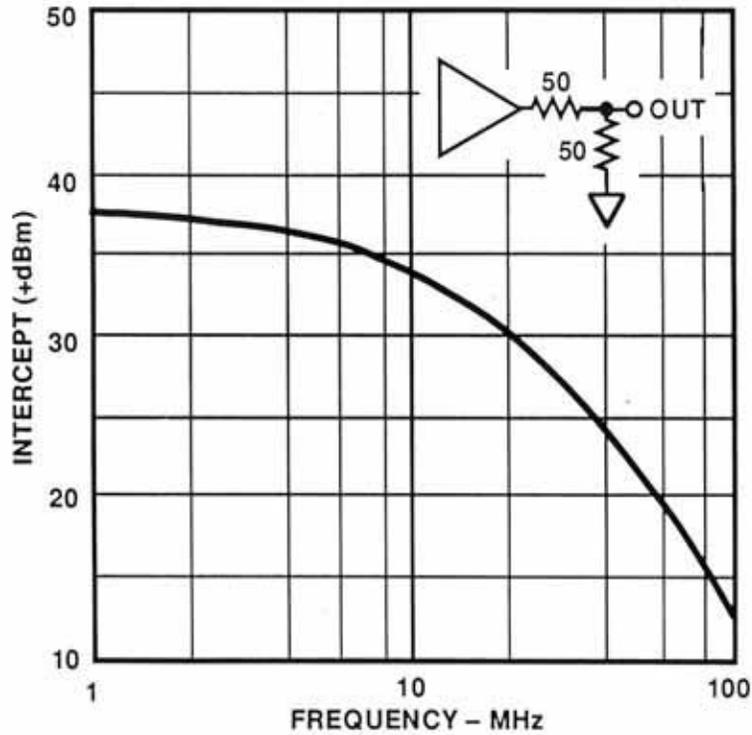


Figure 8.6

Assume the op amp output signal is 5MHz and 2V peak-to-peak into a 100ohm load (50ohm source and load termination). The voltage into the 50ohm load is therefore 1V peak-to-peak, corresponding to +4dBm. The value of the third order intercept at 5MHz is 36dBm. The difference between +36dBm and +4dBm is 32dB. This value is then multiplied by 2 to yield 64dB (the value of the third-order intermodulation products referenced to the power in a single tone). Therefore, the intermodulation products should be -64dBc (dB below carrier frequency), or at a level of -60dBm. Figure 8.7 shows the graphical analysis for this example.

USING THE THIRD ORDER INTERCEPT POINT TO CALCULATE IMD PRODUCT FOR THE AD9622 OP AMP

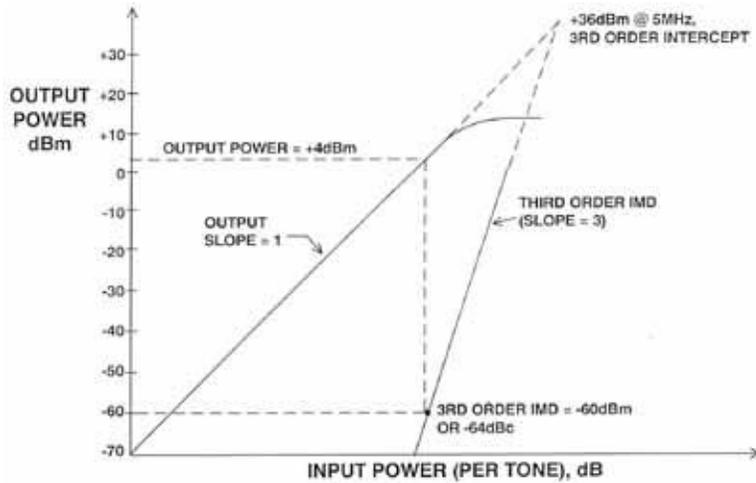


Figure 8.7

HIGH FREQUENCY TWO-TONE GENERATION

Generating test signals with the spectral purity required to make low distortion high frequency measurements is a challenging task. A test setup for generating a single tone is shown in Figure 8.8. The sinewave oscillator should have low phase noise (e.g., Marconi 2382), especially if the device under test is an ADC, where phase noise increases the ADC noise floor. The output of the oscillator is passed through a bandpass (or lowpass) filter which removes any harmonics present in the oscillator output. The distortion should be 6dB lower than the desired accuracy of the measurement. The 6dB attenuator isolates the DUT from the output of the filter. The impedance at each interface should be maintained at 50ohms for best performance (75ohm components can be used, but 50ohm attenuators and filters are generally more readily available). The termination resistor, R_T , is selected so that the parallel combination of R_T and the input impedance of the DUT is 50ohms.

LOW DISTORTION SINGLE-TONE GENERATOR

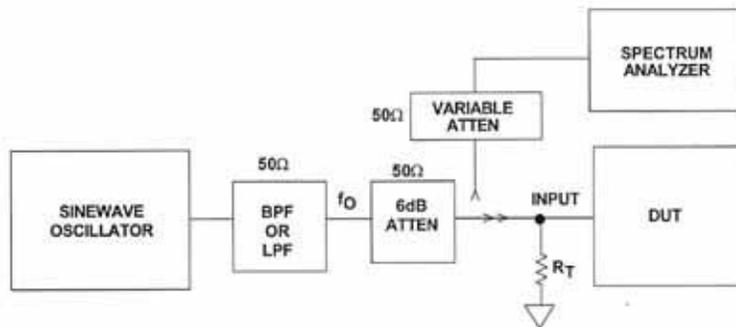


Figure 8.8

Before performing the actual distortion measurement, the oscillator output should be set to the correct frequency and amplitude. Measure the distortion at the output of the attenuator with the DUT replaced by a 50ohm termination resistor (generally the 50ohm input of a spectrum analyzer). Next, replace the 50ohm load with R_T and the DUT. Measure the distortion at the DUT input a second time. This allows non-linear DUT loads to be identified. Non-linear DUT loads (such as flash ADCs with signal-dependent input capacitance, or switched-capacitor CMOS ADCs) can introduce distortion at the DUT input.

Generating two tones suitable for IMD measurements is even more difficult. A low-distortion two-tone generator is shown in Figure 8.9. Two bandpass (or lowpass) filters are required as shown. Harmonic suppression of each filter must be better than the desired measurement accuracy by at least 6dB. A 6dB attenuator at the output of each filter serves to isolate the filter outputs from each other and prevent possible cross-modulation. The outputs of the attenuators are combined in a passive 50ohm combining network, and the combiner drives the DUT. The oscillator outputs are set to the required level, and the IMD of the final output of the combiner is measured. The measurement should be made with a single termination resistor, and again with the DUT connected to identify non-linear loads.

LOW DISTORTION TOW TONE GENERATOR

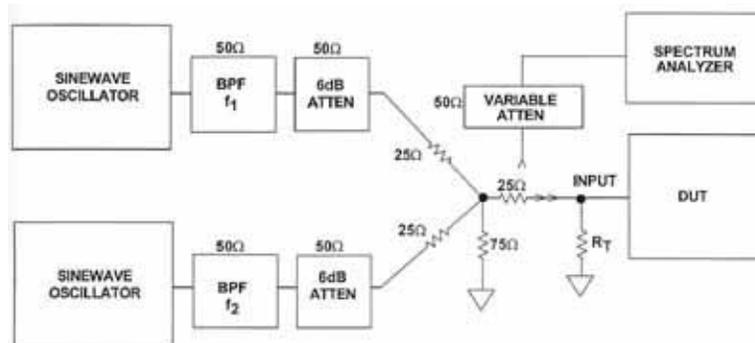


Figure 8.9

USING SPECTRUM ANALYZERS IN HIGH FREQUENCY LOW DISTORTION MEASUREMENTS

Analog spectrum analyzers are most often used to measure amplifier distortion. Most have 50ohm inputs, therefore an isolation resistor between the device under test (DUT) and the analyzer is required to simulate DUT loads greater than 50ohms. After adjusting the spectrum analyzer for bandwidth, sweep rate, and sensitivity, check it carefully for input overdrive. The simplest method is to use the variable attenuator to introduce 10dB of attenuation in the analyzer input path. Both the signal and any harmonics should be attenuated by a fixed amount (10dB, for instance) as observed on the screen of the spectrum analyzer. If the harmonics are attenuated by more than 10dB, then the input amplifier of the analyzer is introducing distortion, and the sensitivity should be reduced. Many analyzers have a

button on the front panel for introducing a known amount of attenuation when checking for overdrive.

MEASURING AMPLIFIER DISTORTION REQUIRES CARE TO PREVENT ANALYZER OVERDRIVE

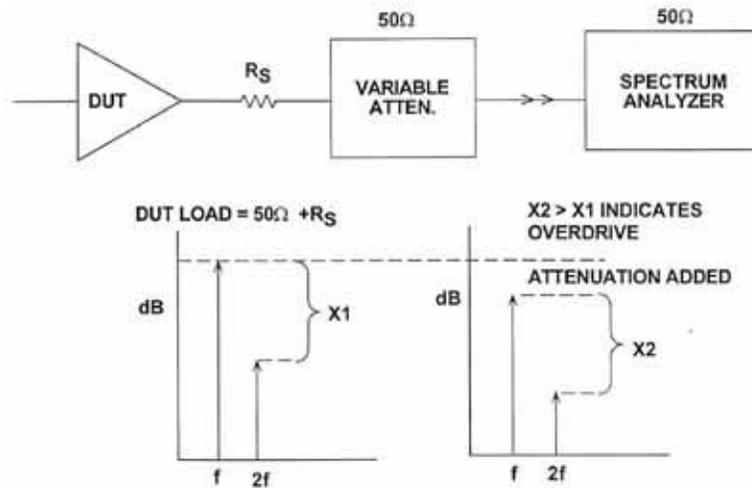


Figure 8.10

Another method to minimize sensitivity to overdrive is shown in Figure 8.11. The amplitude of the fundamental signal is first measured with the notch filter switched out. The harmonics are measured with the notch filter switched in. The insertion loss of the notch filter, X dB, must be added to the measured level of the harmonics.

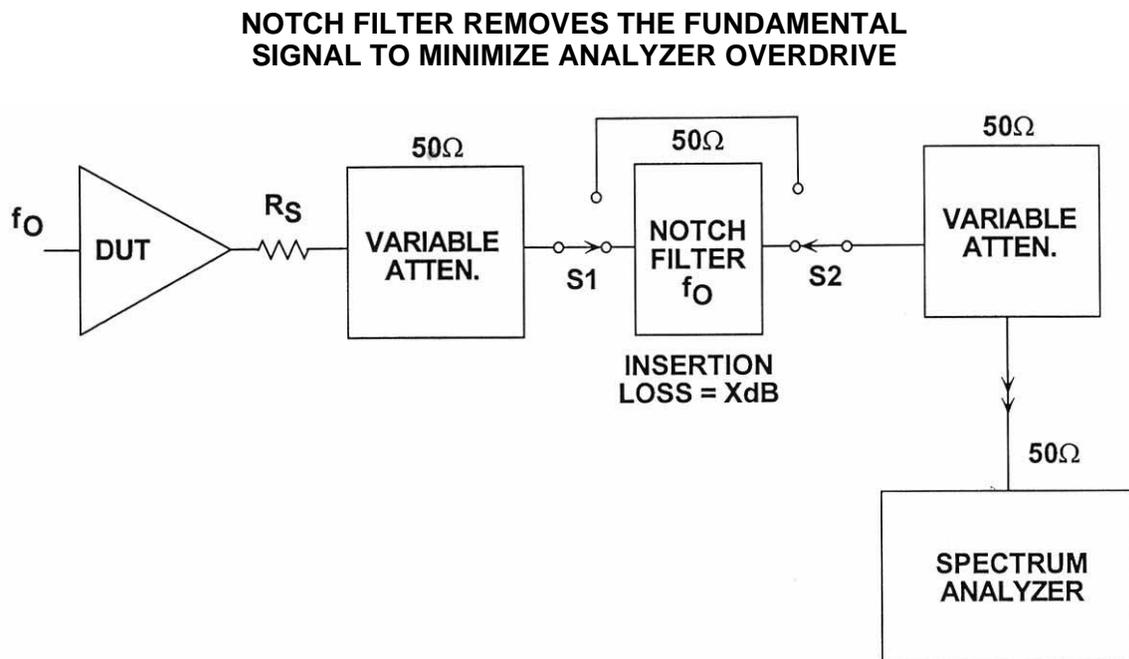


Figure 8.11

MEASURING ADC DISTORTION USING FFTS

The speed of personal computers and the availability of suitable software now makes DSP bench testing of high speed ADCs relatively easy. A block diagram of a typical DSP PC-based test system is shown in Figure 8.12. In order to perform any DSP testing, the first requirement is a high speed buffer memory of sufficient width and depth. High speed logic analyzers make a convenient memory and eliminate the need for designing special hardware. The HP1663A is a 100MHz logic analyzer which has a simple IEEE-488 output port for easy interfacing to a personal computer. The analyzer can be configured as either a 16-bit wide by 8k deep, or a 32-bit wide by 4k deep memory. This is more than sufficient to test a high speed ADC at sample rates up to 100MHz. For higher sample rates, faster logic analyzers are available, but are fairly costly. An alternative to using a high speed logic analyzer is to operate the ADC at the desired sample rate, but only clock the final output register at an even sub-multiple of the sample clock frequency. This is sometimes called *decimation* and is useful for relaxing memory requirements. If an FFT is performed on the decimated output data, the fundamental input signal and its associated harmonics will be present, but translated in frequency. Simple algorithms can be used to find the locations of the signal and its harmonics provided the original signal frequency is known.

A SIMPLE PC-BASED TEST SYSTEM

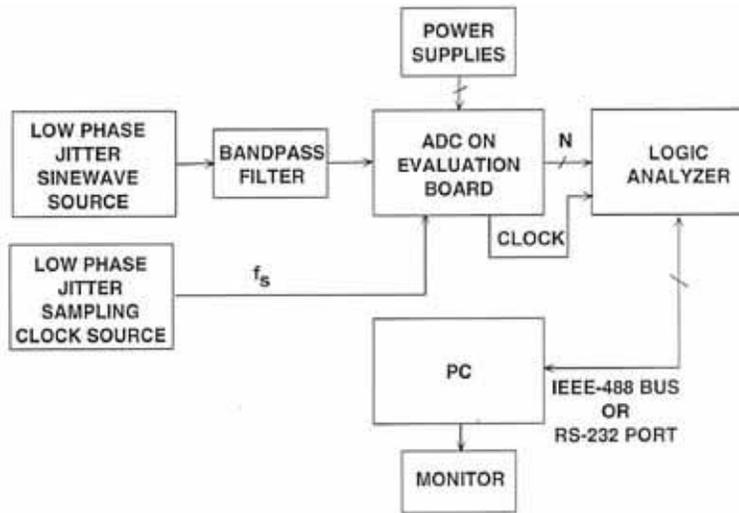


Figure 8.12

FFT TESTING

Easy to use mathematical software packages, such as Mathcad™ (available from MathSoft, Inc., 201 Broadway, Cambridge MA, 02139) are available to perform fast FFTs on most 486-based PCs. The use of a co-processor allows a 4096-point FFT to run in a few seconds on a 75MHz, 486 PC. The entire system will run under the Windows environment and provide graphical displays of the FFT output spectrum. It can be programmed to perform SNR, $S/(N+D)$, THD, IMD, and SFDR computations. A simple QuickBasic program transfers the data stored in the logic analyzer into a file in the PC via the IEEE-488 port (Reference 3, Section 16).

Properly understanding of FFT fundamentals is necessary in order to achieve meaningful results. The first step is to determine the number of samples, M , in the FFT record length. In order for the FFT to run properly, M must be a power of 2. The value of M determines the frequency *bin width*, $\Delta f = f_s/M$. The larger M , the more frequency resolution. Figure 8.13 shows the relationship between the average noise floor of the FFT with respect to the broadband quantization noise level. Each time M is doubled, the average noise in the Δf bandwidth decreases by 3dB. Larger values of M also tend to give more repeatable results from run to run (see Figure 8.13).

RELATIONSHIP BETWEEN AVERAGE NOISE IN FFT BINS AND BROADBAND RMS QUANTIZATION NOISE LEVEL

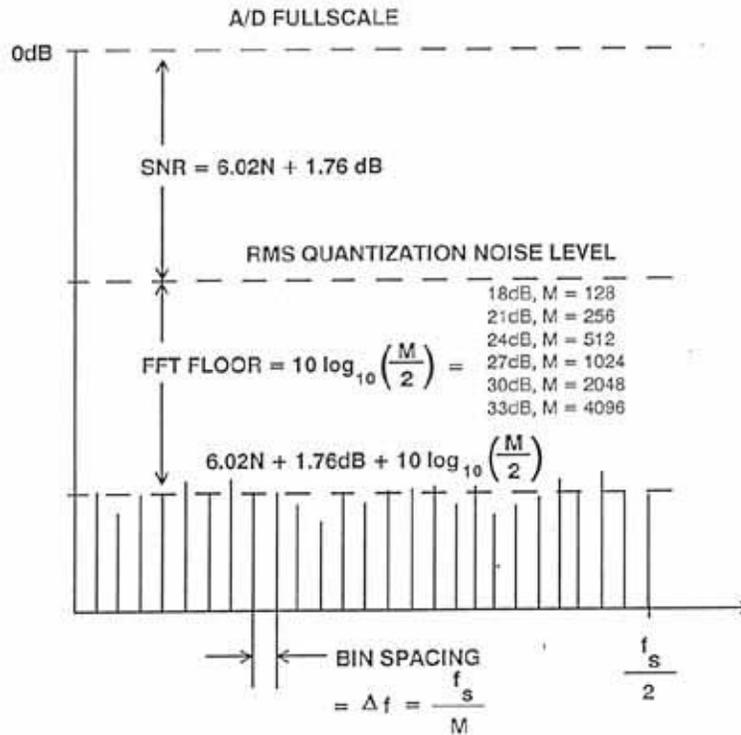


Figure 8.13

M values of 512 (for 8-bit ADCs), 2048 (for 10-bit ADCs), and 4096 (for 12-bit ADCs) have proven to give good accuracy and repeatability. For extremely wide dynamic range applications (such as spectral analysis) M=8192 may be desirable. It should be noted that averaging the results of several FFTs will tend to smooth out the noise floor, but will not change the average value of the floor.

In order to obtain spectrally pure results, the FFT data window must contain an exact integral number of sinewave cycles as shown in Figure 8.14. These frequency ratios must be precisely observed to prevent end-point discontinuity. In addition, it is desirable that the number of sinewave cycles contained within the data window be a prime number. This method of FFT testing is referred to as *coherent* testing because two locked frequency synthesizers are used to insure the proper ratio (coherence) between the sampling clock and the sinewave frequency. The requirements for coherent sampling are summarized in Figure 8.15.

FFT OF SINEWAVE HAVING INTEGRAL NUMBER OF CYCLES IN WINDOW

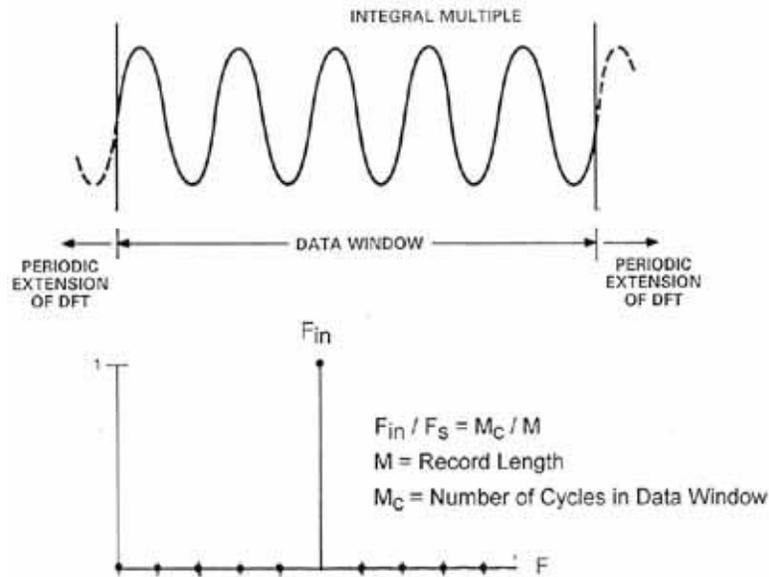


Figure 8.14

REQUIREMENTS FOR COHERENT SAMPLING

- $f_s =$ Sampling Rate
- $f_{in} =$ Input Sinewave Frequency
- $M =$ Number of Samples in Record (Integer Power of 2)
- $M_c =$ Prime Integer Number of Cycles of Sinewave During Record (Makes All Samples Unique)
- Make $f_{in} / f_s = M_c / M$

Figure 8.15

Making the number of cycles within the record a prime number ensures a unique set of sample points within the data window. An even number of cycles within the record length will cause the quantization noise energy to be concentrated in the harmonics of the fundamental (causing a decrease in SFDR) rather than being randomly distributed over the Nyquist bandwidth. Figure 8.16 shows a 4096-point FFT output for a theoretically perfect 12-bit sinewave. The spectrum on the left was made with exactly 128 samples within the record length, corresponding to a frequency which is $1/32$ times f_s . The SFDR is 78dB. The spectrum on the right was made with exactly 127 samples within the record, and the SFDR increases to 92dB.

**CHOOSING A PRIME NUMBER OF CYCLES WITHIN THE
FFT RECORD LENGTH ENSURES RANDOMIZATION OF
THE QUANTIZATION NOISE (IDEAL 12-BIT ADC)**

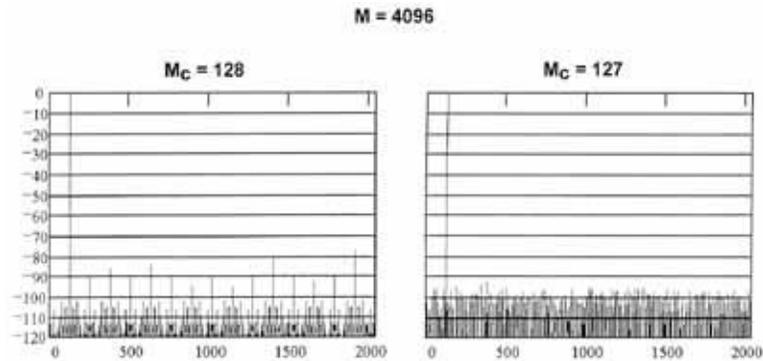


Figure 8.16

Coherent FFT testing ensures that the fundamental signal occupies one discrete line in the output spectrum. Any leakage or smearing into adjacent bins is the result of aperture jitter, phase jitter on the sampling clock, or other unwanted noise due to improper layout, grounding, or decoupling.

If the ratio between the sampling clock and the sinewave frequency is such that there is an endpoint discontinuity in the data (shown in Figure 8.17), then spectral leakage will occur. The discontinuities are equivalent to multiplying the sinewave by a rectangular windowing pulse which has a $\sin(x)/x$ frequency response. The discontinuities in the time domain result in leakage or smearing in the frequency domain, because many spectral terms are needed to fit the discontinuity. Because of the endpoint discontinuity, the FFT spectral response shows the main lobe of the sinewave being smeared, and a large number of associated sidelobes which have the basic characteristics of the rectangular time pulse. This leakage must be minimized using a technique called *windowing* (or *weighting*) in order to obtain usable results in non-coherent tests.

FFT OF SINEWAVE HAVING NON-INTEGRAL NUMBER OF CYCLES IN WINDOW

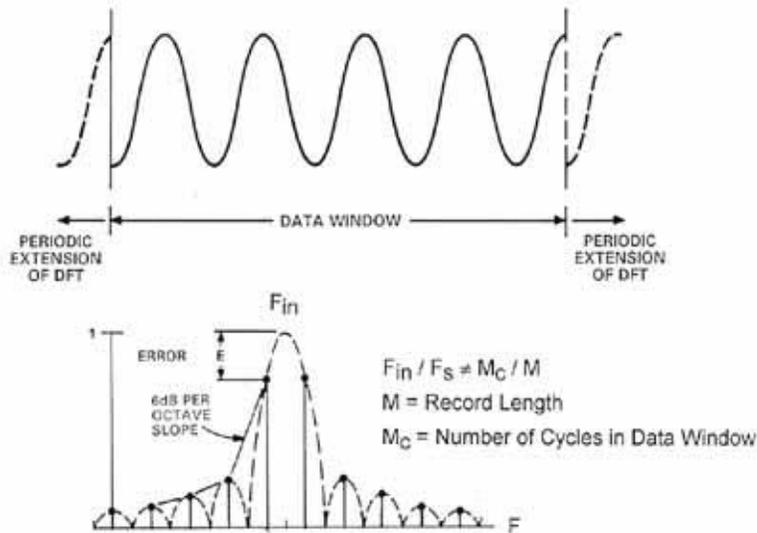


Figure 8.17

This situation is exactly what occurs in real-world spectral analysis applications where the exact frequencies being sampled are unknown and uncontrollable. Sidelobe leakage is reduced by choosing a *windowing* (or *weighting*) function other than the rectangular window. The input time samples are multiplied by an appropriate windowing function which brings the signal to zero at the edges of the window. The selection of an appropriate windowing function is primarily a tradeoff between main-lobe spreading and sidelobe rolloff.

The time-domain and frequency-domain characteristics of a simple windowing function (the Hanning Window) are shown in Figure 8.18. A comparison of the frequency response of the Hanning window and the more sophisticated Minimum 4-Term Blackman-Harris window is given in Figures 8.19 and 8.20. For general ADC testing with non-coherent input frequencies, the Hanning window will give satisfactory results. For critical spectral analysis or two-tone IMD testing, the Minimum 4-Term Blackman-Harris window is the better choice because of the increase in spectral resolution.

TIME AND FREQUENCY REPRESENTATION OF THE HANNING WINDOW

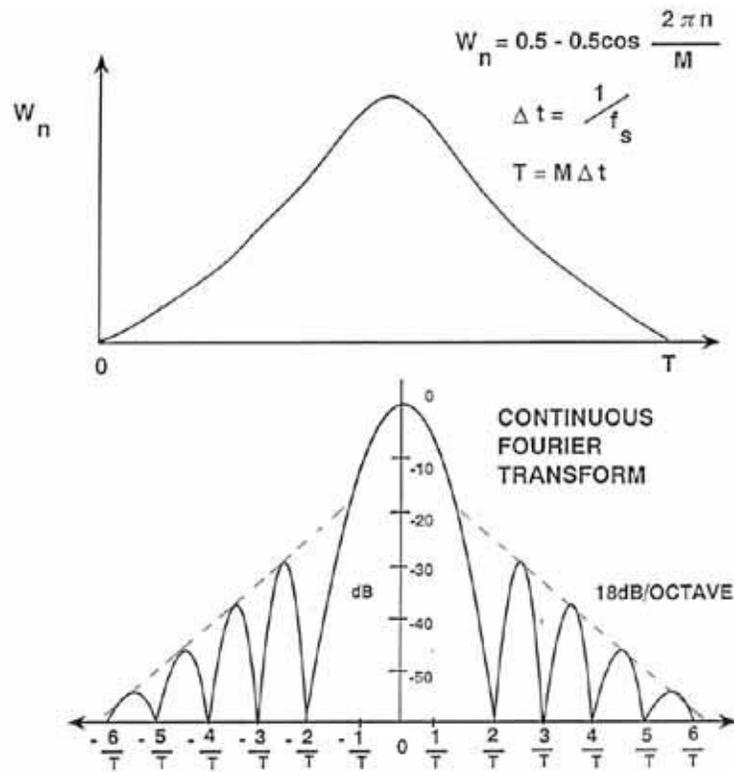


Figure 8.18

FREQUENCY RESPONSE OF THE HANNING WINDOW

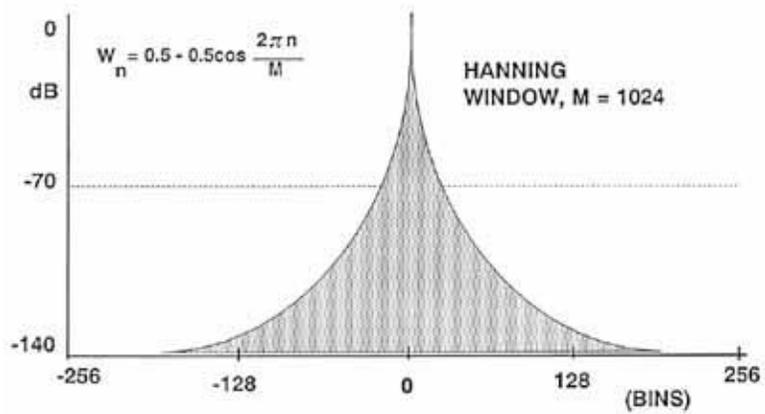


Figure 8.19

FREQUENCY RESPONSE OF THE MINIMUM 4-TERM BLACKMAN-HARRIS WINDOW

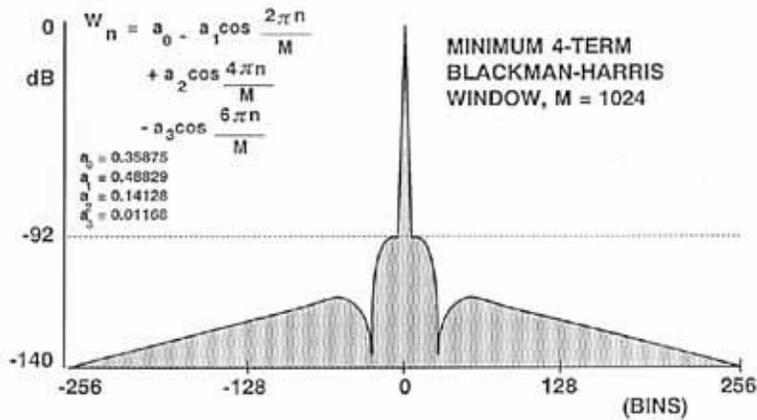


Figure 8.20

The addition of a windowing function to the FFT software involves first calculating the proper coefficient for each time sample within the record. These values are then stored in a memory file. Each time sample is multiplied by its appropriate weighting coefficient before performing the actual FFT. The software routine is easy to implement in QuickBasic.

When analyzing the FFT output resulting from windowing the input data samples, care must be exercised in determining the energy in the fundamental signal and the energy in the various spurious components. For example, sidelobe energy from the fundamental signal should not be included in the rms noise measurement. Consider the case of the Hanning Window function being used to test a 12-bit ADC with a theoretical SNR of 74dB. The sidelobe attenuation of the Hanning Window is as follows:

Bins From Fundamental	Sidelobe Attenuation
2.5	32dB
5.0	50dB
10.0	68dB
20.0	86dB

Therefore, in calculating the rms value of the fundamental signal, you should include at least 20 samples on either side of the fundamental as well as the fundamental itself.

If other weighting functions are used, their particular sidelobe characteristics must be known in order to accurately calculate signal and noise levels.

A typical Mathcad™ FFT output plot is shown in Figure 8.21 for the AD9022 12-bit, 20MSPS ADC using the Hanning Window and a record length of 4096.

MATHCAD™ 4096 POINT FFT OUTPUTS FOR AD9022 12-BIT, 20 MSPS ADC (HANNING WEIGHTING)

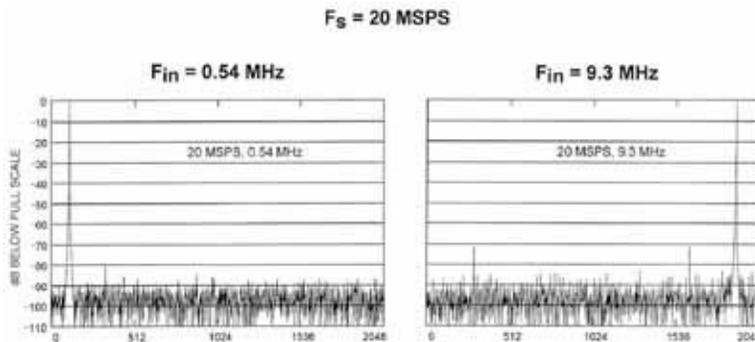


Figure 8.21

The actual QuickBasic routine for transferring the HP analyzer's data to a DOS file in the PC as well as the Mathcad™ routine are given in Reference 3, Section 16.

TROUBLESHOOTING THE FFT OUTPUT

Erroneous results are often obtained the first time an FFT test setup is put together. The most common error is improper timing of the latch strobe to the buffer memory. The HP1663A logic analyzer accepts parallel data and a clock signal. It has an internal DAC which may be used to examine a record of time samples. Large glitches on the stored waveform probably indicate that the timing of the latch strobe with respect to the data should be changed.

After ensuring correct timing, the FFT routine should produce a reasonable spectral output. If there are large values of harmonics, the input signal may be overdriving the ADC at one or both ends of the range. After bringing the signal within the ADC range (usually about 1dB below fullscale), excess harmonic content becomes more difficult to isolate.

Make sure that the sinewave input to the ADC is spectrally pure. Bandpass filters are usually required to clean up the output of most high frequency oscillators, especially if wide dynamic range is expected.

After ensuring the spectral purity of the ADC input, make sure the data output lines are not coupling to either the sampling clock or to the ADC analog input. Remember that the glitches produced on the digital lines are signal-dependent and will therefore contribute to harmonic distortion if they couple into either one of these two lines. As has been discussed previously, noise or digital modulation on the sampling clock can also produce harmonic distortion in the FFT output. The use of an evaluation board with separate sampling clock and analog input connectors will usually prevent this. The special ribbon cable used with the logic analyzer to capture

the ADC output data has a controlled impedance and should not cause performance degradation.

In addition to the above hardware checks, the FFT software should be verified by applying a theoretically perfect quantized sinewave to the FFT and comparing the results to theoretical SNR, etc. This is easy to do using the "roundoff" function available in most math packages. The effects of windowing non-coherent inputs should also be examined before running actual ADC tests.

In performing calculations with the FFT output, the term at dc and $f_s/2$ should be omitted from any calculations, as they can produce erroneous results. Input frequencies which are integer submultiples of the sampling clock can also produce artificially large harmonics.

TROUBLESHOOTING THE FFT OUTPUT

- **Excess harmonic distortion:**
 - ◆ Distortion on input signal
 - ◆ Signal outside ADC input range
 - ◆ Digital runs coupling into analog input or sampling clock
 - ◆ Poor layout, decoupling, and grounding
 - ◆ Buffer memory not clocked at correct time
 - ◆ Analog input frequency locked to integer submultiple of sampling clock

- **Excess noise floor:**
 - ◆ Noise or phase jitter on input signal
 - ◆ Noise or phase jitter on sampling clock
 - ◆ Poor layout, decoupling, and grounding
 - ◆ Eliminate dc and $f_s/2$ FFT components from calculations

Figure 8.22

ANALYZING THE FFT OUTPUT

Once the FFT output is obtained, it can be analyzed in a number of ways, similar to that of the display on an analog spectrum analyzer. The spurious free dynamic range (SFDR) is the ratio of the fundamental signal to the worst frequency spur. Total harmonic distortion (THD) is obtained by taking the ratio of the signal to the rms value of the first several harmonics (and then taking the logarithm of the ratio). Because of aliasing, however, locating the harmonics in the frequency spectrum can be difficult. For instance, if a 3MHz signal is sampled at 10MSPS, the second harmonic (6MHz) actually appears in the FFT output at 4MHz (10MHz - 6MHz). The third harmonic (9MHz) appears at 1MHz (10MHz - 9MHz). The fourth harmonic (12MHz) appears at 2MHz (12MHz - 10MHz). Software routines to perform these calculations are easily written.

Two tone intermodulation distortion can be measured by applying two spectrally pure tones to the ADC using the circuit previously shown in Figure 8.9. The concept of second- or third-order intercept point has little meaning when testing ADCs for two reasons. First, the ADC acts as a hard limiter for out-of-range signals, while an amplifier soft limits. Second, as the amplitude of the tones is reduced, the value of the signal-related frequency spurs tends to become somewhat constant because of the discontinuous nature of the ADC transfer function. A "hard distortion" floor is reached beyond which further reduction in signal amplitude has little effect on the spur levels.

Finally, signal-to-noise plus distortion (S/N+D) can be calculated by taking the ratio of the rms signal amplitude to the rms value of all other spectral components (excluding dc and $f_s/2$). From the S/N+D value, the effective number of bits (ENOB) can be calculated. In some applications, the value of S/N+D without the harmonics included is of interest.

Because of the statistical nature of the FFT analysis, there will be some variability in the output from run to run under identical test conditions. The data can be stabilized by averaging the results of several FFT runs. This will not lower the average noise floor of the FFT, but will reduce the variation in the results.

ANALYZING THE FFT OUTPUT

- **Single-to-Noise including Distortion: S/(N+D)**
- **Effective Number of Bits: (ENOB)**
- **Signal-to-Noise without distortion: SNR**
- **Spurious Free Dynamic Range: SFDR**
- **Harmonic Distortion**
- **Total Harmonic Distortion: THD**
- **THD + Noise (Same as S/N + D)**
- **Two-Tone Intermodulation Distortion**

Figure 8.23

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