Defining And Testing Dynamic ADC Parameters

Part 1 of this two-part series defines some of the key dynamic parameters of ADCs, including signal-to-noise ratio and total harmonic distortion.

Dynamic ADC Testing, Part 1

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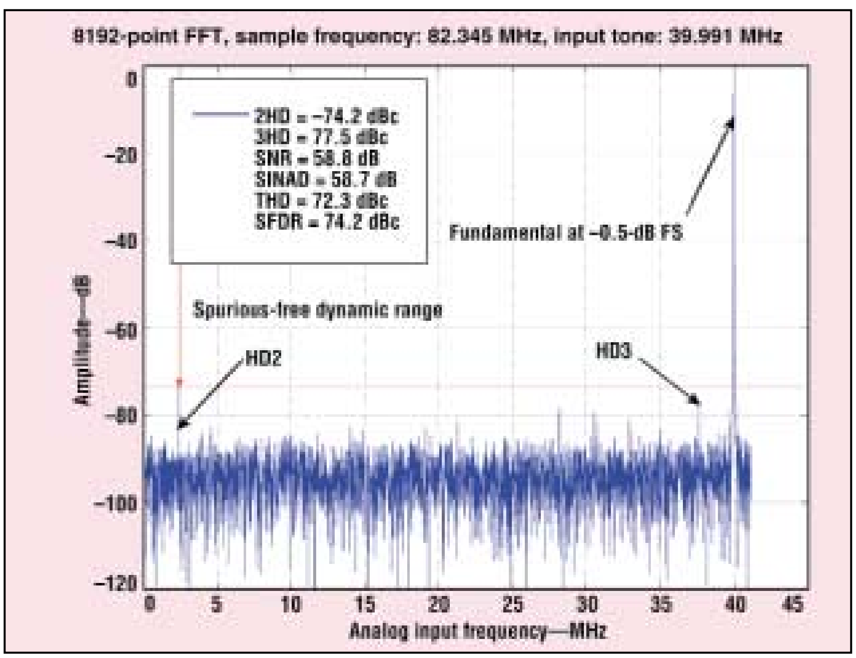
DYNAMIC specifications for high-speed analog-to-digital converters (ADCs) are very important in high-speed applications such as digital communications, ultrasound imaging, instrumentation, and intermediate-frequency (IF) digitization. This first of two articles provides definitions and mathematical foundations for each of the key dynamic ADC parameters, and explains how these dynamic parameters correlate with ADC performance. The key specifications include signal-to-noise ratio (SNR), signal to noise and distortion (SINAD), effective number of bits (ENOB), total harmonic distortion (THD), spurious-free dynamic range (SFDR), two-tone intermodulation distortion (TTIMD), multi-tone intermodulation distortion (MTIMD), and VSWR.

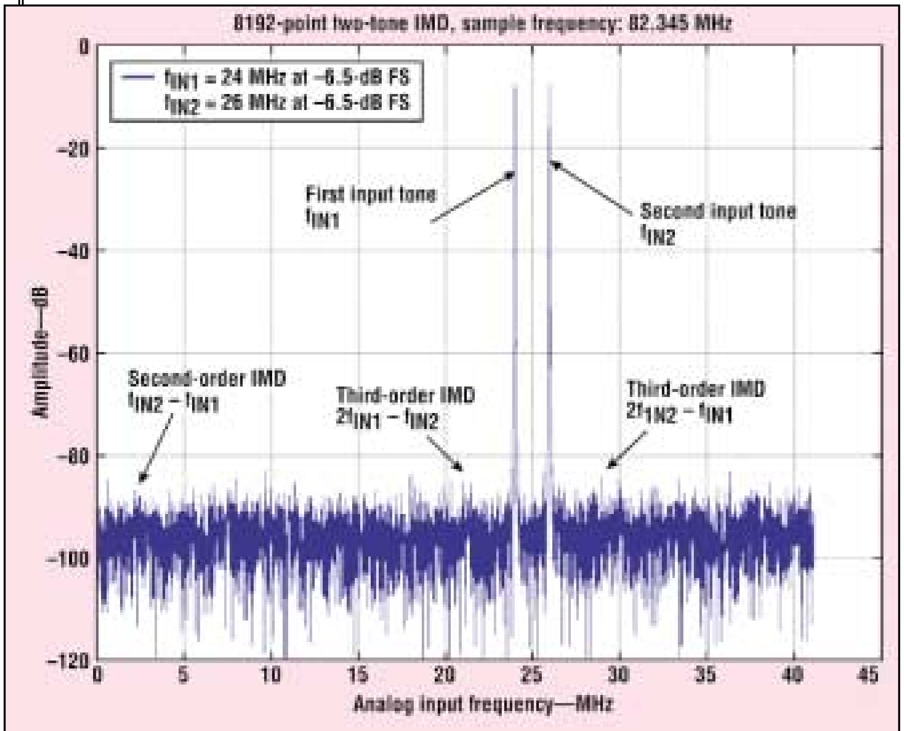
This short article series will conclude next month with insights into the practical aspects of dynamic performance testing. Note that some specifications allow more than one approach for measurement and even for definition. Thus, the test techniques of Part 2 represent one approach, but are not mandatory. Any of the methods described can be extended or altered as necessary to suit the application at hand.



1. This 8192-point FFT plot was generated for the MAX1448.

When testing high-speed ADCs, one emulates the operation of an instrument used to quantify linearity in analog circuits—the spectrum analyzer. For this instrument and for the test procedure, dynamic specifications are usually expressed in the frequency domain, using the Fast Fourier transform (FFT). In both cases, the data output represents the magnitude of this FFT. As an example (Fig. 1), consider the FFT plot for an 80-MSamples/s, 10-b ADC designed and optimized for ultrasound imaging and the digitization of baseband/IF signals. These FFT plots contain impressive amounts of information, and they can be quickly generated. But to make use of an FFT, one must understand how its parameters are defined.

For a waveform perfectly recon-



2. The two-tone intermodulation distortion for the MAX1448 was plotted with a sample frequency (f_{SAMPLE}) of 82.345 MHz.

structed from digital samples, the SNR is the ratio of a root-mean-square (RMS) full-scale analog input to its RMS quantization error, $A_{QUANTIZATION}$ [RMS] = $A_{LSB}/(12)^{0.5}$ = $A_{REF}/(2^N(12)^{0.5})$. The RMS value of a sine wave is one-half its peak-to-peak value divided by $\sqrt{2}$, and quantization error is the difference between an analog waveform and its digitally reconstructed replica, which is characterized by a staircase-shaped transfer curve. The difference function resembles a sawtooth wave that oscillates once per sample between the least-significant-bit (LSB) levels of +0.5 LSB and −0.5 LSB.

The difference function's RMS value is its peak value (0.5 LSB) divided by $\sqrt{3}$. For an ideal N-bit converter, the SNR is defined as:

$$SNR = 2^N \times (\sqrt{3}/\sqrt{2}) = 1.225 \times 2^N$$

Most of the dynamic specifications are expressed as a ratio of relative measurements rather than absolute units. Thus, the SNR for an ideal ADC, driven by a full-scale sinusoidal input with AC power equal to $A_{REF}/(2\sqrt{2})$ [in decibels], is:

$$SNR_{dB} = 20\log_{10}(A_{IN}[RMS] / A_{QUANTIZATION}[RMS])$$

$$SNR_{dB} = 20\log_{10}(A_{REF}[2\times\sqrt{2}] / A_{REF}/[2^N\times\sqrt{12}])$$

$$SNR_{dB} = 6.02N + 1.763. \qquad (1)$$

SNR is diminished by many noise sources in addition to quantization noise (see the sidebar "Decoding noise"). A data converter's resolution and quantization level help to establish its noise floor. The actual SNR for a sinusoidal input signal can therefore be described as:

$$SNR_{dB} = 20\log_{10}(A_{SIGNAL}[RMS] / A_{TOTAL\_NOISE}[RMS]), \qquad (2)$$

where:

A_{SIGNAL}[RMS] = the RMS amplitude for the analog input signal, and

$A_{TOTAL\ NOISE}$[RMS] = the RMS sum of all noise sources (thermal noise, quantization noise, etc.) that limit the converter's dynamic performance. Applying this definition to a 10-b ADC, such as the MAX1448 from Maxim Integrated Products (Sunnyvale, CA), yields a typical SNR value of 58.4 dB at the 40-MHz Nyquist frequency (f_{SAMPLE} = 80 MSamples/s). This SNR represents 94 percent of the ≈62-dB SNR exhibited by an ideal 10-b ADC.

For an ADC driven by a sinusoidal input with an amplitude equal to the ADC's full-scale input, the maximum theoretical SNR is:

$$SNR_{dB} = 6.02N + 1.763 + 10\log_{10} (f_{SAMPLE} / 2)(f_{MAX}), \quad (3)$$

where:

f_{MAX} = the maximum bandwidth of the input tone, and

f_{SAMPLE} = the converter's sampling frequency.

From this equation, note that SNR increases as the sampling frequency increases beyond the Nyquist rate of ($2f_{MAX}$). Known as processing gain, this effect is caused by spreading of the quantization noise power (which is fixed and independent of bandwidth) as the sampling frequency increases. This "oversampling" helps to minimize the effect of noise, which falls into the Nyquist bandwidth of DC to f_{MAX}.

For sinusoidal input signals, SINAD is defined as the ratio of RMS signal to RMS noise (including the first N harmonics of THD—usually the second- through fifth-order harmonics). For a particular sampling rate and input frequency, SINAD provides the ratio (in decibels) of the analog input signal to the noise plus distortion. SINAD describes the quality of an ADC's dynamic range, expressed as the ratio of the maximum amplitude output signal to the smallest increment of output signal that the converter can produce. Mathematically, SINAD is described as:

$$SINAD_{dB} = 20\log_{10} (A_{SIGNAL}[RMS] / A_{NOISE+HD}[RMS]), \quad (4)$$

where:

A_{SIGNAL}[RMS] = the RMS output signal level, and

$A_{NOISE+HD}$[RMS] = the RMS sum of all spectral components below the Nyquist frequency, excluding DC.

The quality of SINAD also depends on the amplitude and frequency of a sinusoidal input tone.

For actual (versus ideal) ADCs, a specification often used in place of the SNR or SINAD is ENOB, which is a global indication of ADC accuracy at a specific input frequency and sampling rate. It is calculated from the converter's digital data record as $N - \log_2$ of the ratio of measured and ideal RMS error:

$$ENOB = N - \log_2 (A_{MEASURED\_ERROR}[RMS] / (A_{IDEAL\_ERROR}[RMS]), \quad (5)$$

where:

N = the number of digitized bits,

$A_{MEASURED\ ERROR}$[RMS] = the averaged noise, and

$A_{IDEAL\ ERROR}$[RMS] = the quantization noise error, expressed as $q/(12)^{0.5} = A_{FS}/(2^N(12)^{0.5}$. Note that A_{FS} is the converter's full-scale input range as determined by the voltage reference, A_{REF}:

$$ENOB = \log_2 (A_{FS} / A_{MEASURED\_ERROR}[RMS]\sqrt{12} \quad (6)$$

or

$$ENOB = \log_2 (A_{REF} / A_{MEASURED\_ERROR}[RMS]\sqrt{12} \quad (7)$$

ENOB generally depends on the amplitude and frequency of the applied sinusoidal input tone, and both must be specified for this particular test. This method compares the RMS noise produced by the ADC under test to the RMS quantization noise of an ideal ADC with the same resolution in bits. If an actual 10-b ADC with a sine-wave input of a particular frequency and amplitude has an ENOB = 9 b, then it produces the same RMS noise level for that input as an ideal 9-b ADC would.

Directly related to SINAD, ENOB

is frequently expressed as ENOB = (SINAD − 1.763)/6.02. The error of an ideal ADC consists solely of noise. For actual converters, however, the measured error includes quantization noise along with aberrations such as missing output codes, AC-to-DC nonlinearity, and aperture uncertainty (jitter). Noise on the reference and power-supply lines also degrades the ENOB.

Dynamic errors and integral nonlinearities contribute to harmonic distortion whenever an ADC samples a periodic signal. For pure sine-wave inputs, the output harmonic-distortion components are found at spectral values whose non-aliased frequencies are integer multiples of the applied sinusoidal input tone. The amplitudes of the non-aliased frequencies, which depend on the amplitude and frequency of the applied input sine wave, are generally provided as a decibel ratio with respect to the amplitude of the applied sine-wave input. Their frequencies are usually expressed as a multiple of the frequency of the applied sinusoidal input signal.

THD is the RMS sum of all harmonics in the output signal's FFT spectrum. In communications, high-speed instrumentation, and RF/IF applications, THD is often a more important figure of merit for ADCs than are the DC nonlinearity specifications that describe the converter's static performance. THD is given by:

$$THD_{dBc} = 20log_{10} (\{A_{HD\_2[RMS]}{}^2 + A_{HD\_3[RMS]}{}^2 + ... + A_{HD\_N[RMS]}{}^2\}^{0.5} / A[f_{IN}]_{RMS}), \quad (8)$$

where:

$A[F_{IN}]_{RMS}$ = the RMS fundamental amplitude, and

$A_{HD\_2[RMS]}$ through $A_{HD\_N[RMS]}$ = the RMS amplitudes of the second- through Nth-order harmonics.

The choice of harmonic components included in a set is usually a trade-off between the desire to include all harmonics with a significant portion of the harmonic-distortion energy, and the exclusion of Discrete Fourier Transform (DFT) frequency bins, whose energy content is mainly dominated by random noise (see the sidebar "DFTs or FFTs?").

Unless otherwise specified (refer to the manufacturer's specification in the data sheet), THD normally consists of the lowest four to nine harmonics (the second through tenth harmonics, inclusive) of the sinusoidal analog input tone. Note that manufacturers may specify their THD values either in decibels (dB) or decibels with reference to the level of the carrier frequency or fundamental (dBc). Both units are commonly used, and THD is defined with respect to the analog input tone.

The term SFDR is usually applied to cases where the harmonic distortion and spurious signals are regarded as undesirable content in the output spectrum of a sampled pure sinusoidal input tone. SFDR indicates the usable dynamic range of an ADC, beyond which a spectral analysis poses special detection and threshold problems. Though similar to THD, SFDR addresses the converter's in-band harmonic characteristics.

SFDR is the ratio of RMS amplitude of the fundamental tone (the maximum signal component) to the RMS value of the largest distortion component in a specified frequency range. In well-designed systems, this spurious signal should be a harmonic of the fundamental. SFDR is important because noise and harmonics restrict a data converter's dynamic range. In an IF bandpass converter, for example, spurious signals may be interpreted as adjacent-channel information.

In other applications, signals of interest, such as low-level radar signals, cannot be distinguished from the harmonic content. To help determine the SFDR value, a spectrum analyzer with an integrated digital-to-analog converter (DAC) for reconstruction is recommended. The usual procedure is to apply a near full-scale input signal (the preferred input-tone amplitude is −0.5 dB to −1 dB FS), measure the response, and then acquire and measure the amplitude of the largest spurious component. SFDR is the ratio of the first to the second measurement. SFDR can also be determined by inspecting the FFT spectrum (plot) of an ADC under test.

For spectrally pure sine-wave inputs, SFDR is the ratio of the amplitude of the averaged DFT value at the fundamental frequency ($A[f_{IN}]$) to the amplitude of the averaged DFT value of the largest-amplitude harmonic ($A_{HD\_MAX}[RMS]$) or spurious signal component ($A_{SPUR\_MAX}[RMS]$) observed over the entire Nyquist band:

$$SFDR_{dBc} = 20log_{10}$$
$$(|A[f_{IN}]_{RMS}| / |A_{HD\_MAX}[RMS]|), \text{ or} \quad (9)$$

or

$$SFDR_{dBc} = 20log_{10}$$
$$(|A[f_{IN}]_{RMS}| / |A_{SPUR\_MAX}[RMS]|). \quad (10)$$

In general, SFDR is a function of the amplitude ($A[f_{IN}]$) and frequency (f_{IN}) of the analog input tone, and, in some cases, even the sampling frequency (f_{SAMPLE}) of the converter under test. When testing an ADC for its SFDR, therefore, the sampling frequency, as well as the input frequency and amplitude, should be specified.

TTIMD is generally caused by modulation, and it can occur when an ADC samples a signal composed of two (or multiple) sine-wave signals. IMD spectral components can occur at the sum (f_{IMF_SUM}) and difference ($f_{IMF\_DIFFERENCE}$) frequencies for all possible integer multiples of the fundamental (input frequency tone) or signal-group frequencies.

When performing TTIMD testing, the input test frequencies—f_{IN1} and f_{IN2}(f_2>f_{IN1})—are set to values that are odd numbers of the DFT bins, and away from the Nyquist frequencies ($f_{SAMPLE/2}$). These settings guarantee that the difference between the two input tones is always an even number of DFT bins. The resulting spectrum is the averaged amplitude spectrum, $A[f_{IMF}]_{RMS}$. The IMD amplitudes for a two-tone input signal are found at the specified sum and difference frequencies:

$$f_{IMF\_SUM} = |mf_{IN1} + nf_{IN2}| \text{ and}$$
$$f_{IMF\_DIFF} = |mf_{IN1} - nf_{IN2}|, \quad (11)$$

where:

m and n = positive integers.

The condition that m and n are greater than zero creates the second-order ($f_{IN1} + f_{IN2}$ and $f_{IN1} - f_{IN2}$) and third-order ($2f_{IN1} + f_{IN2}$, $2f_{IN1} - f_{IN2}$, $f_{IN1} + 2f_{IN2}$ and $f_{IN2} - 2f_{IN2}$, $3f_{IN1}$ and $3f_{IN2}$) IM products.

Since test parameters are generally application specific, no particular guidelines are necessary (or available) to specify the frequencies and signal amplitudes used for IM tests. The size of $| f_{IN2} - f_{IN1} |$ depends entirely on the application and the information desired. Note that small differences in the two input tones cause the IM frequencies to be clustered around the harmonic-distortion components of f_{IN1} and f_{IN2}.

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DFTs OR FFTs?

Testing high-speed analog-to-digital converters (ADCs) for their dynamic performance often requires a frequency transform of the captured data record, using Discrete Fourier transform (DFT) or Fast Fourier transform (FFT) analysis. An FFT produces the same results as the DFT, but minimizes the computation requirements by taking advantage of computational symmetries and redundancies within a DFT analysis. By speeding up the computation, this approach enables a spectral analysis in virtual real time.

Provided that a periodic input signal is sampled frequently enough (i.e., $\geqslant 2f_{MAX}$, where f_{MAX} is the maximum bandwidth of the sinusoidal-input test tone, not the bandwidth of the data converter to be tested), the DFT equation pair is defined as:

$$x[n] = 1/N \sum_{k=0}^{N-1} X[k] \times e^{-j(2\pi kn)} \quad (1)$$

$$X[k] = \sum_{n=0}^{N-1} x[n] \times e^{j(2\pi kn/N)} \quad (2)$$

The acquired data record usually contains sinusoidal input signals, harmonics, intermodulation (IM) products, and other spurious signals that must be analyzed to properly characterize an ADC. Assuming that all input signals are periodic, the DFT of a data record not containing an integral number of cycles of all sinusoidal input signals will contain spectral components at frequencies other than those corresponding to the chosen input tones. Also known as spectral leakage, these components should be avoided because they mask spurious performance of the ADC itself. For a precise characterization, spectral leakage must be kept at a minimum by choosing the proper input tones (with respect to f_{SAMPLE}), and by the use of low-noise, high-precision signal sources.

To avoid spectral leakage completely, the method of coherent sampling is recommended. Coherent sampling requires that the input- and clock-frequency generators are phase locked, and that the input frequency be chosen based on the following relationship:

$f_{IN}/f_{SAMPLE} = N_{WINDOW}/N_{RECORD}$,

where:

f_{IN} = the desired input frequency,

f_{SAMPLE} = the clock frequency of the data converter under test,

N_{WINDOW} = the number of cycles in the data window (to make all samples unique, choose odd or prime numbers), and

N_{RECORD} = the data record length (for an 8192-point FFT, the data record is 8192 points long).

Since the ratio of f_{IN} and f_{SAMPLE} is an integer value, the signal and clock sources must have adequate frequency tuning resolution to prevent spectral leakage.

(continued from p. 84)

TTIMD is generally a function of the amplitudes—$A[f_{IN1}]_{RMS}$ and $A[f_{IN2}]_{RMS}$ and frequencies (f_{IN1} and f_{IN2}) of the input components. Therefore, it is necessary to specify the input tones and amplitudes for which two-tone IMD measurements are performed. It is essential that the input test signal be virtually free of IM and harmonic distortion. For ADCs that have wide bandwidths and large dynamic ranges, this condition is increasingly difficult to achieve.

Two signal generators, containing output-leveling circuitry and linked through balanced or isolated outputs or any other coupling circuits, can easily generate IMD effects. In order to avoid IMD in the test signal, therefore, one should operate power splitters/combiners (used to combine or split two input tones) well within their linear range. Figure 2 depicts two-tone IMD with second- and third-order IMD products for a 10-b, 80-MSamples/s ADC. For best results, the two-tone envelope for this ADC was chosen to be −0.5-dB FS, and the amplitude for the two input tones was normalized to −6.5-dB FS.

Multi-tone IMD tests are often used in system design to determine limits for the signal dynamic range, useful frequency bands for different signal groups, and where to set the input signal's noise floor to mask small IM components for a particular ADC. The measurement of single-tone harmonic distortion is useful in obtaining general ideas about the linearity of a particular ADC, but these data do not lead directly to models for predicting useful measures of IM performance for independent input signal tones.

MEASURING NPR

Noise power ratio (NPR) defines the spectral power of contributed errors such as IMD and THD, in a small frequency band within the baseband of the composite input signal being analyzed.

For this test, one generates random noise whose spectrum is approximately uniform up to a predetermined cutoff frequency less than half the sampling frequency. Then, a notch filter removes a narrow band of frequencies from the noise. To improve the measurement, the notch depth is recommended to be at least 10 to 15 dB greater than the NPR value being measured. Compared to the overall noise bandwidth, the notch width should be narrow. With this notched noise applied to the ADC input, one computes the frequency spectrum of the resulting code sequence, and then calculates NPR as the ratio (in decibels) of the average power-spectral density inside the notched frequency band to that outside of the notched band.

TEST PROCEDURE

A typical test procedure features a computer-controlled DAC that generates a signal composed of a set of sine waves at DFT binary center frequencies. As the tone amplitudes are increased uniformly, beginning at the noise floor and continuing to the full-scale ADC level at which clipping begins, gaps between the tones serve as observation points to analyze any resulting IMD. These tests provide results similar to that of the noise-power-ratio (NPR) test (see the sidebar "Measuring NPR"). They support better simulation of the expected signal-group waveforms, however.

Seldom specified in the data sheets for high-speed data converters, VSWR is the ratio of mismatch between the actual impedance and the desired or expected impedance. It can be calculated by applying a test signal and measuring the reflection coefficient of the ADC input terminal. Calculated by the following, VSWR is directly related to the reflection coefficient, ρ, of a simple terminating impedance, Z_T:

$$VSWR = (1+|\rho|)/(1-|\rho|), \text{ where}$$
$$\rho = (Z_T - Z_o)/(Z_T + Z_o). \quad (12)$$

where:

Z_T = the ADC input termination impedance, and

Z_O = the transmission line impedance (nominally 50 Ω).

To compensate for circuit inaccuracies in the measurement, it is recommended to use calibration standards if available (typically short, open, and 50 Ω).

In addition to the test setup information, Part 2 will provide samples of source code based on the MATLAB (from the MathWorks, Natick, MA) and LabWindows/CVI (from National Instruments, Austin, TX) software tools. The software will enable designers to analyze the dynamic performance of an ADC by capturing data records quickly and processing them efficiently. • •

DECODING NOISE

The term "noise" is rather ambiguous if not qualified as to type. In general, it includes the effects of nonlinearities [such as integral nonlinearities (INLs) and differential nonlinearities (DNLs)], random and fixed-pattern effects, and sampling-time error. The total noise ($A_{TOTAL\ NOISE}$[RMS]) is any deviation of the output signal (converted to input units) from the input signal, excluding deviations caused by differential gain and phase errors, or DC-level shifts. Notable examples of these effects defined here as noise, include quantization error, harmonic and intermodulation distortion (IMD), and spurious distortion.

For Further Reading

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