Defining and Testing Dynamic Parameters in
High-Speed ADCs, Part 1

The first part of this article series discusses commonly known definitions most crucial for high-speed data converters (in this case analog-to-digital converter or short ADCs) used in communications, instrumentation and data acquisition applications. The purpose of this article is to help the reader gain a better understanding of common parameters such as signal-to-noise ratio (SNR), signal-to-noise-and-distortion (SINAD), total harmonic distortion (THD) and spurious-free dynamic range (SFDR). In the second part of this article series (see “Dynamic Testing of High-Speed ADCs” for further reading), these parameter definitions are put to the test by measuring them in real-world test scenarios.

Dynamic specifications for ADCs are very important in high-speed applications such as digital communications, ultrasound imaging, instrumentation, and IF digitization. The following discussion provides a definition and a mathematical foundation for each parameter, offers useful techniques for evaluating the dynamic performance of high-speed ADCs, and explains how the dynamic parameters correlate with ADC performance. Part 1 of this two-part discussion covers the definition of these specifications:

- Signal-to-noise ratio (SNR)
- Signal-to-noise and distortion ratio (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)
- Voltage standing-wave ratio (VSWR)

In explaining how to measure these parameters, Part 2 provides insight 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 method and are not mandatory. Any of the methods described can be extended or altered as necessary to suit the application at hand.

When testing high-speed A/D converters, one emulates the operation of a spectrum analyzer used to quantify linearity in analog circuits. 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 (Figure 1), consider the FFT plot for an 80Mps, 10-bit ADC designed and optimized for ultrasound imaging and the digitization of baseband/IF frequencies. Such FFT plots contain impressive amounts of information and can be generated quickly. However, to make use of an FFT, one must understand how its parameters are defined.
Signal-to-Noise Ratio (SNR)

For a waveform perfectly reconstructed from digital samples, SNR is the ratio of an rms (root mean square) full-scale analog input to its rms quantization error ($A_{\text{quantization}}[\text{rms}] = A_{\text{LSB}} / \sqrt{2} = A_{\text{REF}} / [2^N \sqrt{2}]$). 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 levels $+1/2\text{LSB}$ and $-1/2\text{LSB}$ (LSB being least-significant bit). The difference function's rms value is its peak value $(1/2\text{LSB})$ divided by $\sqrt{3}$. For an ideal N-bit converter, SNR is defined as

$$\text{SNR} = 2^N \cdot \sqrt{3} / \sqrt{2} = 1.225 \cdot 2^N$$

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

$$\text{SNR}_{20} = 20 \cdot \log_{10}(A_{\text{IN}}[\text{rms}] / A_{\text{quantization}}[\text{rms}])$$

$$\text{SNR}_{20} = 20 \cdot \log_{10}(A_{\text{REF}} / 2 \cdot \sqrt{2} / A_{\text{REF}} / 2^N \cdot \sqrt{2})$$

$$\text{SNR}_{20} = 6.02 \cdot N + 1.763$$
SNR is diminished by many noise sources in addition to quantization noise (See Appendix 1). A data converter's resolution and quantization level both help to establish its noise floor. The actual SNR for a sinusoidal input signal can therefore be described as

\[
\text{SNR}_{\text{dB}} = 20 \cdot \log_{10}(A_{\text{SIGNAL}}[\text{rms}]/A_{\text{TOTAL\_NOISE}}[\text{rms}])
\]

where \(A_{\text{SIGNAL}}[\text{rms}]\) represents the rms amplitude for the analog input signal, and \(A_{\text{TOTAL\_NOISE}}[\text{rms}]\) is 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-bit ADC such as the MAX1448 yields a typical SNR value of 58.4dB at the 40MHz Nyquist frequency (\(f_{\text{SAMPLE}} = 80\text{Msps}\)). This SNR represents 94% of the ~62dB SNR exhibited by an ideal 10-bit 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

\[
\text{SNR}_{\text{dB}} = 6.02 \cdot N + 1.763 + 10 \cdot \log_{10}(f_{\text{SAMPLE}}/2 \cdot f_{\text{MAX}})
\]

where \(f_{\text{MAX}}\) describes the maximum bandwidth of the input tone, and \(f_{\text{SAMPLE}}\) is the converter's sampling frequency. From this equation, note that SNR increases as the sampling frequency increases beyond the Nyquist rate (2 \(f_{\text{MAX}}\)). Called 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 \(f_{\text{MAX}}\).

Signal-to-Noise and Distortion Ratio (SINAD)

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 2nd through 5th-order harmonics). For a given sampling rate and input frequency, SINAD provides the ratio (in dB) 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

\[
\text{SINAD}_{\text{dB}} = 20 \cdot \log_{10}(A_{\text{SIGNAL}}[\text{rms}]/A_{\text{NOISE+HD}}[\text{rms}])
\]

where \(A_{\text{SIGNAL}}[\text{rms}]\) depicts the rms output signal level and \(A_{\text{NOISE+HD}}[\text{rms}]\) describes the rms sum of all spectral components below the Nyquist frequency, excluding DC. The quality of SINAD also depends on the amplitude and the frequency of a sinusoidal input tone.

Effective Number of Bits (ENOB)

For actual (versus ideal) ADCs, a specification often used in place of the SNR or the 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:
where \( N \) is the number of digitized bits, \( A_{\text{measured \_ error}}[\text{rms}] \) is the averaged noise, and \( A_{\text{ideal \_ error}}[\text{rms}] \) is the quantization noise error, expressed as \( q/\sqrt{12} = A_{\text{FS}}/2^N \cdot \sqrt{12} \). \( A_{\text{FS}} \) is the converter's full-scale input range as determined by the reference voltage \( A_{\text{REF}} \).

\[
\text{ENOB} = \log_2 \left( \frac{A_{\text{FS}}}{A_{\text{measured \_ error}}[\text{rms}] \cdot \sqrt{12}} \right)
\]

or

\[
\text{ENOB} = \log_2 \left( \frac{A_{\text{REF}}}{A_{\text{measured \_ error}}[\text{rms}] \cdot \sqrt{12}} \right)
\]

ENOB generally depends on the amplitude and the 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-bit ADC with a sine-wave input of a given frequency and amplitude has an ENOB = 9.0 bits, then it produces the same rms noise level for that input as would an ideal 9-bit ADC.

Directly related to SINAD, ENOB is frequently expressed as

\[
\text{ENOB} = \left( \frac{\text{SINAD} \cdot 1.763}{6.02} \right)
\]

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/DC nonlinearity, and aperture uncertainty (jitter). Noise on the reference and power-supply lines also degrades the ENOB.

**Total Harmonic Distortion (THD)**

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 nonaliased frequencies are integer multiples of the applied sinusoidal input tone. The amplitudes of the nonaliased frequencies, which depend on the amplitude and the frequency of the applied input sine wave, are generally given as a dB-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 Fast Fourier Transform (FFT) spectrum. All harmonics are included by definition, but the first three (in most cases) represent the major contribution to output distortion in a given converter. In communications 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} = 20 \cdot \log_{10} \left( \sqrt{A_{HD_2}[rms]^2 + A_{HD_3}[rms]^2 + \cdots + A_{HD_N}[rms]^2} / A[f_{IN}]_{rms} \right),

where A[f_{IN}]_{rms} is the rms fundamental amplitude, and A_{HD_2}[rms] through A_{HD_N}[rms] represent the rms amplitudes of the 2nd to 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 Appendix 2).

Unless otherwise specified (refer to the manufacturer's specification in the data sheet), THD normally consists of the lowest four to nine harmonics (2nd through 10th harmonics, inclusive) of the sinusoidal analog input tone. Note that manufacturers can specify their THD values either in decibels (dB) or decibels with reference to the carrier frequency or fundamental (dBc). Both units are in common use, and THD is defined with respect to the analog input tone.

**Spurious-Free Dynamic Range (SFDR)**

The term spurious-free dynamic range is usually applied for cases in which the harmonic distortion and spurious signals are regarded as undesirable spurs 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 thresholding problems. Though similar to THD, SFDR addresses the converter's in-band harmonic characteristics.

Spurious-free dynamic range is the ratio of rms amplitude of the fundamental (the maximum signal component) to the rms value of the largest distortion component in a specified frequency range. In well-designed systems, this spur 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, spurs can 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 -0.5dB to -1dB 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.

$$SFDR_{dBc} = 20 \cdot \log_{10}(A[f_{IN}]_{rms} / A_{HD_MAX}[rms])$$
In general, SFDR is a function of the amplitude and the frequency \(A[f_{IN}]\) of the analog input tone and, in some cases, even the sampling frequency \(f_{SAMPLE}\) of the converter under test. Therefore, when testing an ADC for its spurious-free dynamic range, you should specify the sampling frequency as well as the input frequency and amplitude.

**Two-Tone Intermodulation Distortion (Two-Tone IMD)**

IMD 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 both the sum \(f_{IMF\_SUM}\) and the difference \(f_{IMF\_DIFF}\) frequencies for all possible integer multiples of the fundamental (input frequency tone) or signal-group frequencies.

For the two-tone IMD test, the input test frequencies \(f_{IN1}\) and \(f_{IN2}\) \((f_{IN2} > f_{IN1})\) 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 (Figure 2) 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:

\[
\begin{align*}
|f_{IMF\_SUM}| &= |m \cdot f_{IN1} + n \cdot f_{IN2}| \\
|f_{IMF\_DIFF}| &= |m \cdot f_{IN1} - n \cdot f_{IN2}|
\end{align*}
\]

where \(m\) and \(n\) are positive integers. The condition that \(m\) and \(n\) are greater than zero creates the 2nd-order \((f_{IN1} + f_{IN2}\) and \(f_{IN1} - f_{IN2}\)\) and 3rd-order \((2f_{IN1} + f_{IN2}, 2f_{IN1} - f_{IN2}, f_{IN1} + 2f_{IN2}, f_{IN1} - 2f_{IN2}, 3f_{IN1}\) and \(3f_{IN2}\)\) intermodulation products.

Because test parameters are generally application-specific, no particular guidelines are necessary (or available) to specify the frequencies and signal amplitudes used for intermodulation 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 intermodulation frequencies to be clustered around the harmonic distortion components of \(f_{IN1}\) and \(f_{IN2}\).

Two-tone intermodulation distortion is generally a function of the amplitudes \(A[f_{IN1}]_{rms}\) and \(A[f_{IN2}]_{rms}\) and the frequencies \(f_{IN1}\) and \(f_{IN2}\) of the input components. You must therefore 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 intermodulation and harmonic distortion. For ADCs of larger dynamic range and wider bandwidth, this condition is increasingly difficult to achieve.
Figure 2. This plot illustrates a two-tone IMD spectrum with 2nd- and 3rd-order IMD products.

Two signal generators, containing output-leveling circuitry and linked via balanced or isolated outputs or any other coupling circuits, can easily generate IMD effects. Therefore, to avoid intermodulation distortion in the test signal, you should operate power splitters/combiners (used to combine or split two input tones) well within their linear range. Figure 3 depicts two-tone IMD with 2nd- and 3rd-order IMD products for a 10-bit, 80Msps ADC. For best results, the two-tone envelope for this ADC was chosen to be -0.5dB FS, and the amplitude for the two input tones was normalized to -6.5dB FS.
Multi-Tone Intermodulation Distortion (Multi-Tone IMD)

Multi-tone intermodulation distortion 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 intermodulation components for a given ADC. The measurement of single-tone harmonic distortion is useful in obtaining general ideas about the linearity of a given ADC, but such data does not lead directly to models for predicting useful measures of intermodulation performance for independent input-signal tones.

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. Such tests provide results similar to that of the noise-power ratio (NPR) test (see Appendix 3). They allow better simulation of the expected signal-group waveforms, however.

Voltage Standing-Wave Ratio (VSWR)

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 as follows, VSWR is directly related to the reflection coefficient $\rho$ of a simple terminating impedance $Z_T$: 
where $Z_T$ depicts the ADC input termination impedance, and $Z_0$ represents the transmission line impedance (nominally $50\Omega$). To compensate for circuit inaccuracies in the measurement, it is recommended to use calibration standards if available (typically short, open, and $50\Omega$).

**Conclusion**

The preceding discussion has been a review of the most important dynamic specifications for high-speed data converters. It will conclude in Part 2, which offers detailed insight into the tools most suitable for capturing data records and for using those records in testing the dynamic performance parameters defined above. In addition to the test setup information, Part 2 provides samples of MATLAB® and LabWindows/CVI® source code, enabling designers to analyze the dynamic performance of an ADC by capturing data records quickly and processing them efficiently.

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**Appendices**

**Appendix 1**

The term "noise" is rather ambiguous if not qualified as to type. In general, it includes the effects of nonlinearities (INL, DNL), random and fixed-pattern effects, and sampling-time error. Total noise ($A_{\text{TOTAL_NOISE}[\text{rms}])$ 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 such effects, defined here as noise, include quantization error, harmonic and intermodulation distortion, and spurious distortion.

**Appendix 2**

Testing high-speed 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., $\geq 2 \times f_{\text{MAX}}$, where $f_{\text{MAX}}$ 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

$$|\mathbf{H}| = \frac{(1 + |\rho|) + (1 - |\rho|)}{2}$$

where $\rho = \frac{(Z_T - Z_0)}{(Z_T + Z_0)}$
The acquired data record usually contains sinusoidal input signals, harmonics, intermodulation 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 $f_{\text{SAMPLE}}$) 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 you choose the input frequency based on this relationship:

$$f_{\text{IN}}/f_{\text{SAMPLE}} = N_{\text{WINDOW}}/N_{\text{RECORD}}$$

where $f_{\text{IN}}$ is the desired input frequency, $f_{\text{SAMPLE}}$ is the clock frequency of the data converter under test, $N_{\text{WINDOW}}$ is the number of cycles in the data window (to make all samples unique, choose odd or prime numbers), and $N_{\text{RECORD}}$ is the data record length (for an 8192-point FFT, the data record contains 8192 points).

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

**Appendix 3**

Noise-power ratio (NPR) is a figure of merit that 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 processed and 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 10dB to 15dB 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 dB) of the average power spectral density inside the notched frequency band to that outside the notched band.

Literature Sources

2. MAX1448 data sheet Rev. 0, 7/00, Maxim Integrated Products.
3. MAX1448EVKIT data sheet Rev. 0, 7/00, Maxim Integrated Products.

A similar version of this article appeared in the November 2000 issue of Microwaves and RF

Dynamic Testing of High-Speed ADCs, Part 2

Analog-to-digital converters (ADCs) represent the link between analog and digital worlds in receivers, test equipment and other electronic devices. As outlined in Part 1 of this article series, a number of key dynamic parameters provide an accurate correlation of the dynamic performance to be expected from a given ADC. Part 2 of this article series covers some of the setup configurations, equipment recommendations and measurement procedures for testing the dynamic specifications of high-speed ADCs.

The following is a discussion of the setups and procedures recommended for testing high-speed data converters. It includes the software tools, the hardware configurations, and the instruments for data capture and analysis needed to test a new family of 10-bit, +3V, high-speed data converters from Maxim. It also warns of traps you can encounter if equipment selection, setup configuration, layout, and FFT-based analysis are not performed with care. The following topics are covered:

- Dynamic specifications and definitions
- Board layout and hardware configuration
- Power spectrum, bins, spectral leakage, and window functions
- Software tools for testing SNR, SINAD, THD, SFDR, and TTIMD

Many approaches are available for acquiring output data from A/D converters (not just the high-speed ones) and for analyzing their dynamic performance. The methods presented here represent one proven approach, and readers are encouraged to modify them as necessary for the application at hand.
## Dynamic Specifications

For those who missed Part 1 of this discussion, the following is a brief overview of the definitions and the mathematical descriptions of important dynamic parameters for high-speed ADCs.

<table>
<thead>
<tr>
<th>Dynamic Parameter</th>
<th>Description/Definition</th>
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<tbody>
<tr>
<td>Signal-to-Noise Ratio (SNR)</td>
<td>$\text{SNR}_{\text{dB}} = 6.02 \cdot \text{N} + 1.763$.</td>
</tr>
<tr>
<td>Signal-to-Noise and Distortion Ratio (SINAD)</td>
<td>$\text{SINAD}<em>{\text{dB}} = 20 \cdot \log</em>{10} (A_{\text{SIGNAL[rms]}} / A_{\text{NOISE[rms]}})$.</td>
</tr>
<tr>
<td>Effective Number of Bits (ENOB)</td>
<td>$\text{ENOB} = (\text{SINAD} - 1.763) / 6.02$.</td>
</tr>
<tr>
<td>Total Harmonic Distortion (THD)</td>
<td>$\text{THD}<em>{\text{dB}} = 20 \cdot \log</em>{10} (\sqrt{\sum_{k=2}^{\infty}</td>
</tr>
<tr>
<td>Spurious-Free Dynamic Range (SFDR)</td>
<td>SFDR is the ratio expressed in decibels of the rms amplitude of the fundamental (maximum signal component) to the rms value of the next-largest spurious component, excluding DC offset.</td>
</tr>
<tr>
<td>Two-Tone Intermodulation Distortion (TTIMD)</td>
<td>$\text{TTIMD}<em>{\text{dB}} = 20 \cdot \log</em>{10} \left( \sum (A_{\text{IMF_SUM[rms]}} + A_{\text{IMF_DIFF[rms]}}) / A_{\text{FUNDAMENTAL[rms]}} \right)$. IMF_SUM and IMF_DIFF in a TTIMD setup contain two input tones only.</td>
</tr>
<tr>
<td>Multi-Tone Intermodulation Distortion (MTIMD)</td>
<td>$\text{MTIMD}<em>{\text{dB}} = 20 \cdot \log</em>{10} \left( \sum (A_{\text{IMF_SUM[rms]}} + A_{\text{IMF_DIFF[rms]}}) / A_{\text{FUNDAMENTAL[rms]}} \right)$. IMF_SUM and IMF_DIFF in an MTIMD setup contain more than two (usually up to four) input tones.</td>
</tr>
<tr>
<td>Voltage Standing-Wave Ratio (VSWR)</td>
<td>$\text{VSWR} = (1 +</td>
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### Board Layout and Hardware Requirements for the Test Setup

To perform adequate dynamic tests on high-speed data converters, you should use a test board pre-assembled by the manufacturer or follow the data sheet’s board-layout recommendations. This article considers the layout requirements for dynamic testing before delving into the details of hardware and software. An evaluation or characterization board for fast data converters (Figure 1a) must incorporate high-speed layout techniques (Figure 1b and 1c). You can usually replicate the dynamic performance specified in a data sheet by following these basic rules:
- Locate all bypass capacitors as close to the device as possible, preferably on the same side as the ADC, using surface-mount components to achieve minimum trace length, inductance, and capacitance.
- Bypass analog and digital supplies, references, and common-mode inputs with two 0.1µF ceramic capacitors in parallel and a 2.2µF bipolar capacitor to ground.
- Multilayer boards with separate ground and power planes produce the highest level of signal integrity.
- Consider the use of a split ground plane arranged to match the physical location of analog and digital grounds on the ADC’s package. The impedance of the two ground planes must be kept as low as possible, and, to avoid possible damage or latchup, their AC and DC voltage differences (or both) must be less than 0.3V. These grounds should be joined at a single point, so that noisy digital ground currents do not interfere with the analog ground plane. The ideal location of this connection can be determined experimentally, as the point along the gap between the two ground planes that produces optimum results. This connection can be achieved with a low-value surface-mount resistor of 1Ω to 5Ω, a ferrite bead, or a direct short.
- As an alternative (if the ground plane is sufficiently isolated from noisy digital systems such as the downstream output buffer and DSP), all ground pins can share the same ground plane.
- Route high-speed digital signal traces away from sensitive analog traces.
- Keep all signal lines short and free of 90° turns.
- Always consider the clock input as an analog input. Route it away from actual analog inputs and other digital signal lines.

![Figure 1a. MAX1448 EV kit circuit schematic](PDF, 210k)
A proper test setup and the right test equipment\(^1\) are necessary to realize the performance specified for a given converter (Figure 2a and 2b).
The following hardware has proven to be extremely efficient and is therefore recommended for the test setup:

**DC power supply:** Hewlett Packard E3620A dual-supply 0-25V/0-1A. Use separate supplies for the analog and the digital nodes. Each must provide 100mA of output drive current.

**Clock-signal function generator:** Hewlett-Packard HP8662A. The clock input for the device under test (DUT) accepts CMOS-compatible clock signals. This signal should have low jitter and fast rise and fall times, because the high-speed ADC has a 10-stage pipeline, and its interstage conversion depends on the repeatability of the rising and falling edges of the external clock. Sampling occurs on the falling edge
of the clock signal, so edge should have the lowest possible jitter. Significant aperture jitter limits the ADC’s SNR performance as follows:

\[
SNR_d = 20 \cdot \log\left(\frac{1}{2\pi \cdot f_N \cdot t_{AJ}}\right)
\]

where \(f_N\) represents the analog input frequency and \(t_{AJ}\) is the time of the aperture jitter. Clock jitter is especially critical for undersampling applications.

**Input-signal function generator:** Hewlett-Packard HP8662A. For proper operation, this function generator should be phase-locked to the clock-signal generator.

**Logic analyzer (LA):** Hewlett-Packard HP16500C. Depending on the number of points in the proposed FFT, you may be able to capture the data using an LA with less memory depth (such as the 4k data record available in the HP1663C).

**Analog bandpass filter:** TTE Elliptical Function Bandpass Filter, Q56 series. Cutoff frequencies are 7.5MHz, 20MHz, 40MHz, and 50MHz.

**Digital multimeters (DMMs):** Various Fluke, Keithley Instruments, and HP multimeters (including the hand-held HP2373A and the AC-powered HP34401A) were used in the setup to check for proper reference, supply, and common-mode voltages.

### Evaluating the DUT

To simplify evaluation of the DUT, it was tested with a performance-optimized, fully assembled and tested, surface-mount board. Follow the steps below to configure the setup and to operate this board. You should complete all the connections before turning on the power supplies or enabling the function generators.

1. Apply a +3.0V analog power supply to VAIN1 and VAIN2, and connect its ground terminal to AGND.
2. Apply a +3.0V digital power supply to VDIN1 and VDIN2, and connect its ground terminal to DGND.
3. Verify that no shunts are installed for jumpers JU1 (shutdown disabled) and JU2 (digital outputs enabled).
4. Connect the clock function generator to the CLOCK SMA connector.
5. Connect the output of the analog-signal function generator to the input of one of the bandpass filters.
6. To evaluate differential analog signals, verify that shunts are installed on pins 1 and 2 of jumpers JU3 and JU4. Connect the output of the bandpass filter to the DIFF IN SMA connector.
7. To evaluate single-ended analog signals, verify that shunts are installed on pins 2 and 3 of jumpers JU3 and JU4, and connect the output of the bandpass filter to the SINGLE IN SMA connector.

8. Connect one of the logic-analyzer interface cables (pods) to the square pin header J1.

9. Turn on both power supplies, and verify +1.20V across test points TP4 and TP5 with a voltmeter. If necessary, adjust potentiometer R34 to obtain +1.20V.

10. Enable the function generators. Set the clock function generator to its maximum output amplitude (999mV for the suggested HP8662A) and a clock frequency of $f_{\text{CLK}} = 80\text{MHz}$. Set the analog signal function generator to the desired input tone, with any amplitude between 10µV and 999mV. Note that input amplitude and frequency must be selected according to the bandpass filter's corner frequency. Bandpass filters used in evaluating high-speed data converters usually have a very narrow passband. To achieve optimum performance (depending on filter type and manufacturer, of course), you should set the input tone to within 5% of the corner frequency. Because the filter attenuates the generator's output signal, set the generator's amplitude slightly higher to achieve the desired full-scale input specification.

11. For proper operation, phase-lock the two (three, if testing for two-tone IMD) function generators.

12. Synchronize the LA with the external clock signal from the board, and set the LA to latch data on the clock's rising edge.

13. Enable the LA and begin collecting data. Data can be stored on a floppy disk, on the LA's hard disk, or on a data-acquisition (DAQ) board communicating through the LA's HPIB bus.

Now that the necessary steps for test setup and hardware configuration have been completed and the system is ready to capture data from the DUT, it is time to select the software tools for data capture and analysis:

- LabWindows/CVI™ serves as the required data capture and communications link between the LA and the DAQ controller board. (The C-based program routine used for this purpose will not be discussed in this article.)

- MATLAB™ is a powerful tool that performs the FFT and dynamic analysis of the captured data.

To help you understand how a MATLAB program routine analyzes and graphs the dynamic performance of a high-speed data converter, some of the FFT and power-spectrum basics are reviewed in the next section.

**Power Spectrum, Bins, Spectral Leakage, and Windowing**

The Fast Fourier Transform (FFT) and the power spectrum are powerful tools for measuring and analyzing signals from captured data records. They can capture time-domain signals, measure their frequency content, convert the results to convenient units, and display them. To perform FFT-based measurements, however, one must understand the issues and the calculations involved. Basic functions of an FFT-based signal analysis are the FFT itself and the power spectrum. Both are extremely useful for measuring the frequency content of stationary or transient signals. FFTs usually produce the average of a signal's frequency content over the time interval that the signal was acquired. Thus, FFTs are always recommended for stationary-signal analysis.
Two-Sided to Single-Sided Power-Spectrum Conversion

Among the most basic and important computations in signal analysis are the use of the FFT in converting from a two-sided to a single-sided power spectrum, adjusting the frequency resolution, and displaying the spectrum. A power spectrum usually returns a matrix containing the two-sided representation of the time-domain signal power in the frequency domain. The values in this matrix are proportional to the amplitude squared of each frequency component making up the time-domain signal.

A plot of the two-sided power spectrum usually contains both negative and positive frequency components. Actual frequency-analysis tools, however, focus on the positive half of the frequency spectrum only, noting that the spectrum of a real signal is symmetrical around DC. Negative frequency information is therefore irrelevant. In a two-sided spectrum, half the energy resides in the positive frequencies and half in the negative frequencies. Therefore, to convert from a two-sided spectrum to a single-sided spectrum, you discard the second half of the matrix and multiply every point (except DC) by two.

Bins and Frequency Resolution

The frequency range and the resolution on the x-axis of a spectrum plot (see the program-code extraction below) depend on the sampling rate and the size of the data record (the number of acquisition points). The number of frequency points or lines in the power spectrum is N/2, where N is the number of signal points captured in the time domain. The first frequency line in the power spectrum always represents DC. The last frequency line can be found at \( \frac{f_{\text{SAMPLE}}}{2} - \frac{f_{\text{SAMPLE}}}{N} \). Frequency lines are spaced at even intervals of \( \frac{f_{\text{SAMPLE}}}{N} \), commonly referred to as a frequency bin or a FFT bin (Figure 3).

```
Example:
Based on an 8192-point FFT and a sampling frequency of 80Mps, the MAX1448 provides a bin spacing of 9.768kHz.

N: ADC Resolution
N_{\text{RECORD}}: Number of points in the FFT
f_{\text{SAMPLE}}: ADC sampling frequency
```

**Figure 3. The representation of frequency/FFT bins in an FFT graph**
Bins can also be computed with reference to the ADC’s sampling period:

\[ \text{Bin} = \frac{f_{\text{SAMPLE}}}{N} = \frac{1}{(N \Delta t_{\text{SAMPLE}})} \]

For example, with a sampling frequency of \( f_{\text{SAMPLE}} = 82.345 \text{MHz} \) and a record length of 8,192 data points, the distance between each frequency line in the FFT plot is exactly 10.052kHz. (Refer to Figure 1 of Defining and Testing Dynamic Parameters in High-Speed ADCs, Part 1.)

The calculations for the frequency axis (\( x \)-axis) are proof that the sampling frequency determines the range or the bandwidth of the frequency spectrum. For a given sampling frequency, the number of points acquired in the time domain determines the resolution frequency. To increase the resolution for a given frequency range, the depth of the data record can be increased at the same sampling frequency (see the following program-code extraction).

```matlab
% Find the signal bin number, DC = bin 1
fin = find(Dout_dB(1:numpt/2)==maxdB);
% Span of the input frequency on each side
span = max(round(numpt/200),5);
% Approximate search span for harmonics on each side
spanh = 2;
% Determine power spectrum
spectP = (abs(Dout_spect)).^2*(abs(Dout_spect));
% Find DC offset power
Pdc = sum(spectP(1:span));
% Extract overall signal power
Ps = sum(spectP(fin-span:fin+span));
% Vector/matrix to store both frequency and power of signal and harmonics
Fh = [];
% The 1st element in the vector/matrix represents the signal, the next element represents the 2nd harmonic, etc.
Ph = [];
```

**Spectral Leakage and Window Functions**

Window functions are common in FFT analysis, and their proper use is critical in FFT-based measurements. The following discussion of spectral leakage stresses the need to select an appropriate window function and scale it properly for a given application. To accurately determine spectral leakage, however, it may not be enough to use adequate signal-acquisition techniques, convert a two-sided power spectrum into a single-sided one, and rescale the result. To gain a better understanding of this term, one should perform an N-point FFT on a spectrally pure sinusoidal input.

Spectral leakage is the result of an assumption in the FFT algorithm that the time record is precisely repeated throughout all time and that all signals contained in this time record are periodic at intervals corresponding to the length of the time record. However, a nonintegral number of cycles in the time record (\( f_{\text{IN}} f_{\text{SAMPLE}} \neq \frac{N_{\text{WINDOW}}}{N_{\text{RECORD}}} \)) violates this condition and causes spectral leakage (Figure 4).
Only two cases can guarantee the acquisition of an integral number of cycles:

- Synchronous sampling with respect to the input tone
- The capture of a transient signal that fits entirely into the time record

In most cases, however, the application deals with an unknown stationary input. This means there is no guarantee of sampling an integral number of cycles. Spectral leakage distorts the measurement by spreading the energy of a given frequency component over the adjacent frequency lines or bins. Selecting an appropriate window function can minimize the effects of this spectral leakage.

![Diagram showing the effects of windows on spectral leakage](image)

**Figure 4. The effects of windows on spectral leakage**

To fully understand how a given window function affects the frequency spectrum, one must take a closer look at the frequency characteristics of windows. Windowing of the input data is equivalent to convolving the spectrum of the original signal with the spectrum of the window. Even for coherent sampling, the signal is convolved with a rectangular-shaped window of uniform height. Such convolution shows a typical sine-function characteristic spectrum.

The real-frequency characteristic of a window is a continuous spectrum consisting of a main lobe and several side lobes. The main lobe is centered at each frequency component of the signal in the time domain. Side lobes approach zero at intervals on each side of the main lobe. An FFT, on the other hand, produces a discrete frequency spectrum. The continuous, periodic spectrum of a window is sampled by the FFT, just as an ADC would sample an input signal in the time domain. What appears in each frequency line of the FFT is the value of the continuous, convolved spectrum at each FFT frequency line.
If the frequency components of the original signal match a frequency line exactly, as is the case when you acquire an integral number of cycles, you see only the main lobe of the spectrum. Side lobes do not appear, because the window spectrum approaches zero at bin-frequency intervals on either side of the main lobe. If a time record does not contain an integral number of cycles, the continuous spectrum of the window is shifted from the main lobe center at a fraction of the frequency bin that corresponds to the difference between the frequency component and the FFT frequency lines. This shift causes side lobes to appear in the spectrum. Thus, the window's side-lobe characteristics directly affect the extent to which adjacent frequency components "leak into" the neighboring frequency bins.

Window Characteristics

Before choosing an appropriate window, it is necessary to define the parameters and the characteristics that enable users to compare windows. Such characteristics include the -3dB main-lobe width, the -6dB main-lobe width, the maximum side-lobe level, and the side-lobe rolloff rate (Table 1).

Side lobes of the window are characterized by the maximum side-lobe level (defined as the maximum side-lobe level in dB with respect to the main lobe's peak gain) and the side-lobe rolloff (defined as the asymptotic decay rate in dB/decade or dB/octave of frequency) of the side-lobe peaks.

Table 1. Characteristics of Frequently Used Window Functions
(Also refer to MATLAB program code)

<table>
<thead>
<tr>
<th>Window Type</th>
<th>-3dB Main-Lobe Width</th>
<th>-6dB Main-Lobe Width</th>
<th>Maximum Side-Lobe Level</th>
<th>Side-Lobe Rolloff Rate</th>
</tr>
</thead>
<tbody>
<tr>
<td>No Window</td>
<td>0.89 bins</td>
<td>1.21 bins</td>
<td>-13dB</td>
<td>20dB/decade, 6dB/octave</td>
</tr>
<tr>
<td>Hanning</td>
<td>1.44 bins</td>
<td>2.00 bins</td>
<td>-32dB</td>
<td>60dB/decade, 18dB/octave</td>
</tr>
<tr>
<td>Hamming</td>
<td>1.30 bins</td>
<td>1.81 bins</td>
<td>-43dB</td>
<td>20dB/decade, 6dB/octave</td>
</tr>
<tr>
<td>Flat Top</td>
<td>2.94 bins</td>
<td>3.56 bins</td>
<td>-44dB</td>
<td>20dB/decade, 6dB/octave</td>
</tr>
</tbody>
</table>

Selecting the Right Window

Different windows suit different applications. To choose the right spectral window, one has to guess the signal frequency content. If the signal contains strong interfering frequency components distant from the frequency of interest, you should choose a window whose side lobes have a high-rolloff rate. If strong interfering signals are close to the frequency of interest, a window with low maximum levels of side lobe is more suitable.

If the frequency band of interest contains two or more signals close to each other, spectral resolution becomes important. In that case, a window with a narrow main lobe is better. For a single frequency
component in which the focus is on amplitude accuracy rather than its precise location in the frequency bin, a window with a broad main lobe is recommended. Finally, coherent sampling (instead of a window) is recommended for a flat or broadband frequency spectrum (see the following program-code extraction).

%If no window function is used, the input tone must be chosen to be unique and with regard to the sampling frequency. To achieve this prime numbers are introduced and the input tone is determined by \( f_{IN} = \frac{f_{SAMPLE}}{\text{Prime Number}} \), with the input tone

%To relax this requirement, window functions such as HANNING and HAMMING (see below) can be introduced, however the fundamental in the resulting FFT spectrum appears ‘sharper’ without the use of window functions.

Doutw=Dout;
%Doutw=Dout.*hanning(numpt);
%Doutw=Dout.*hamming(numpt);

%Performing the Fast Fourier Transform
Dout_spect=fft(Doutw);

%Recalculate to dB
Dout_dB=20*log10(abs(Dout_spect));

%Display the results in the frequency domain with an FFT plot
figure;
maxdB=max(Dout_dB(1:numpt/2));

The Hanning window function, which provides good frequency resolution and reduced spectral leakage, yields satisfactory results in most applications. The Flat Top window has good amplitude accuracy, but its wide main lobe provides poor frequency resolution and more spectral leakage. The Flat Top window has a lower maximum side-lobe level than does the Hanning window, but the Hanning window has a faster rolloff rate.

An application consisting of only transient signals should have no spectral windows at all, because they tend to attenuate important information at the beginning of the sample block. In the case of a transient signal, you should choose a nonspectral window such as the Force or Exponential window.

Selecting an appropriate window is not easy, but if the signal content is unknown one can start with the Hanning characteristic. It is also an excellent idea to compare the performance of multiple window functions to find the one most suitable for a given application.

Table 2. Signal Content vs. Window Selection and Advantages

<table>
<thead>
<tr>
<th>Window Type</th>
<th>Signal Content</th>
<th>Window Characteristics</th>
</tr>
</thead>
<tbody>
<tr>
<td>No Window</td>
<td>Broad-band random, closely spaced sine-wave signals</td>
<td>Narrow main lobe, slow rolloff rate, poor frequency resolution</td>
</tr>
<tr>
<td>(Uniform)</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Table 2. Signal Content vs. Window Selection and Advantages
Hanning | Narrow-band random signals, nature of content is unknown, sine-wave or combination of sine-wave signals | High maximum side-lobe level, good frequency resolution, reduced leakage, faster rolloff rate
---|---|---
Hamming | Closely spaced sine-wave signals | Good spectral resolution, narrow main lobe
Flat Top | Sine wave with need for amplitude accuracy | Good amplitude accuracy, wide main lobe, poor frequency resolution, more spectral leakage

**Dynamic-Range Specifications SNR, SINAD, THD, and SFDR**

With the knowledge you’ve gained from the preceding sections of this article, the following program-code extraction should be easy to understand. Based on the FFT, the power spectrum, and attention to spectral leakage and window functions, the specifications SNR, SINAD, THD, and SFDR are calculated as follows, using MATLAB:

\[
\begin{align*}
\text{SNR} &= 10 \cdot \log_{10} \left( \frac{P_s}{P_n} \right) \\
\text{SINAD} &= 10 \cdot \log_{10} \left( \frac{P_s}{P_n + P_d} \right) \\
\text{THD} &= 10 \cdot \log_{10} \left( \frac{P_d}{P_h(1)} \right) \\
\text{SFDR} &= 10 \cdot \log_{10} \left( \frac{P_h(1)}{\max(P_h(2:10))} \right).
\end{align*}
\]

where \( P_s \) is signal power, \( P_n \) is noise power, \( P_d \) is distortion power caused by 2nd through 5th-order harmonics, \( P_h(1) \) is fundamental harmonic power, and \( P_h(2:10) \) is harmonic power of 2nd through 9th-order harmonics (see the following program-code extraction for the power-spectrum level).

```matlab
%Find harmonic frequencies and power components in the FFT spectrum
for har_num=1:10
%Input tones greater than f_{SAMPLE} are aliased back into the spectrum
tone=rem((har_num*(fin-1)+1)/numpt,1);
if tone>0.5
%Input tones greater than 0.5*f_{SAMPLE} (after aliasing) are reflected
tone=1-tone;
end
Fh=[Fh tone];
end
%For this procedure to work, ensure the folded back high order harmonics do not overlap
%with DC or signal or lower order harmonics
har_peak=max(spectP(round(tone*numpt)-spanh:round(tone*numpt)+spanh));
har_bin=find(spectP(round(tone*numpt)-spanh:round(tone*numpt)+spanh)==har_peak);
har_bin=har_bin+round(tone*numpt)-spanh-1;
Ph=[Ph sum(spectP(har_bin-1:har_bin+1))];
end
end

%Determine the total distortion power
Pd=sum(Ph(2:5));
```
Determine the noise power

\[ P_n = \text{sum(spectP(1:numpt/2))} - P_{dc} - P_s - P_d \]

Format;

\[ A = \frac{\text{max(code)} - \text{min(code)}}{2^n} \]

\[ A_{dB} = 20 \log_{10}(A) \]

\[ \text{SINAD} = 10 \log_{10} \left( \frac{P_s}{P_n + P_d} \right) \]

\[ \text{SNR} = 10 \log_{10} \left( \frac{P_s}{P_n} \right) \]

Disp('THD is calculated from 2nd through 5th order harmonics');

\[ \text{THD} = 10 \log_{10} \left( \frac{P_d}{P_h(1)} \right) \]

\[ \text{SFDR} = 10 \log_{10} \left( \frac{P_h(1)}{\max(P_h(2:10))} \right) \]

Disp('Signal & Harmonic Power Components:');

\[ \text{HD} = 10 \log_{10} \left( \frac{P_h(1:10)}{P_h(1)} \right) \]

Based on the MATLAB source code (see below), the MAX1448 was tested not only for its data sheet specifications but for many other over- and undersampling input frequencies as well. It achieved excellent dynamic performance under all conditions.

Example program routine to generate FFT plots and determine the dynamic performance of a high-speed data converter from the data records taken with a HP16500C Logic Analyzer. Data was extracted through the HPIB interface and read into the following MATLAB program routine. The same data can be extracted from the controller interface of the LA and simply be copied to a floppy disk — a rather time-consuming way, but possible.

Start MAX1448 Dynamic Performance Test Routine

Disp('HP16500C State Card');
filename=input('Type a:filename or Press RETURN for HPIB Data Transfer: ');
if isempty(filename)
    filename = 'listing';
end
fid=fopen(filename,'r');
numpt=input('Data Record Size (Number of Points)? ');
fclk=input('Sampling Frequency (MHz)? ');

MAX1448 - 10-bit data converter
numbit=10;

Discard first 13 lines from the data file, which do not contain data
for i=1:13,
fgetl(fid);
end
[v1,count]=fscanf(fid,'%f',[2,numpt]);
fclose(fid);
\[ \text{v1} = \text{v1}'; \]
\[ \text{code} = \text{v1}(:,2); \]

\% Display a warning, when the input generates a code greater than full-scale
\text{if} \ (\text{max(code)} == 2^\text{numbit} - 1) \ | \ (\text{min(code)} == 0) \\
\text{disp('Warning: ADC may be clipping!!!');} \\
\text{end} \\

\% Plot results in the time domain
\text{figure;} \\
\text{plot([1:numpt], code);} \\
\text{title('TIME DOMAIN')} \\
\text{xlabel('SAMPLES')} \\
\text{ylabel('DIGITAL OUTPUT CODE');} \\

\% Recenter the digital sine wave
\text{Dout = code - (2^\text{numbit} - 1)/2;} \\

\% If no window function is used, the input tone must be chosen to be unique and with \% regard to the sampling frequency. To achieve this prime numbers are introduced and the \% input tone is determined by \( f_{\text{IN}} = f_{\text{SAMPLE}} \times \text{Prime Number / Data Record Size}. \)
\% To relax this requirement, window functions such as HANNING and HAMING (see below) can \% be introduced, however the fundamental in the resulting FFT spectrum appears 'sharper' \% without the use of window functions.
\text{Doutw = Dout;} \\
\text{Doutw = Dout .* hanning(numpt);} \\
\text{Doutw = Dout .* hamming(numpt);} \\

\% Performing the Fast Fourier Transform
\text{Dout_spect = fft(Doutw);} \\

\% Recalculate to dB
\text{Dout_dB = 20*\text{log10(abs(Dout_spect));}} \\

\% Display the results in the frequency domain with an FFT plot
\text{figure;} \\
\text{maxdB = max(Dout_dB(1:numpt/2));} \\

\% For TTIMD, use the following short routine, normalized to —6.5dB full-scale.
\text{\% plot([0:numpt/2-1] \times fclk/numpt, \text{Dout_dB(1:numpt/2)}-maxdB-6.5);} \\
\text{plot([0:numpt/2-1] \times fclk/numpt, Dout_dB(1:numpt/2)-maxdB);} \\
\text{grid on;} \\
\text{title('FFT PLOT');}
xlabel('ANALOG INPUT FREQUENCY (MHz)');
ylabel('AMPLITUDE (dB)');
a1=axis; axis([a1(1) a1(2) -120 a1(4)]);

%Calculate SNR, SINAD, THD and SFDR values
%Find the signal bin number, DC = bin 1
fin=find(Dout_dB(1:numpt/2)==maxdB);
%Span of the input frequency on each side
span=max(round(numpt/200),5);
%Approximate search span for harmonics on each side
spanh=2;
%Determine power spectrum
spectP=(abs(Dout_spect)).*(abs(Dout_spect));
%Find DC offset power
Pdc=sum(spectP(1:span));
%Extract overall signal power
Ps=sum(spectP(fin-span:fin+span));
%Vector/matrix to store both frequency and power of signal and harmonics
Fh=[];
%The 1st element in the vector/matrix represents the signal, the next element represents
%the 2nd harmonic, etc.
Ph=[];

%Find harmonic frequencies and power components in the FFT spectrum
for har_num=1:10
%Input tones greater than fSAMPLE are aliased back into the spectrum
tone=rem((har_num*(fin-1)+1)/numpt,1);
if tone>0.5
%Input tones greater than 0.5*fSAMPLE (after aliasing) are reflected
tone=1-tone;
end
Fh=[Fh tone];
%For this procedure to work, ensure the folded back high order harmonics do not overlap
%with DC or signal or lower order harmonics
har_peak=max(spectP(round(tone*numpt)-spanh:round(tone*numpt)+spanh));
har_bin=find(spectP(round(tone*numpt)-spanh:round(tone*numpt)+spanh)==har_peak);
har_bin=har_bin+round(tone*numpt)-spanh-1;
Ph=[Ph sum(spectP(har_bin-1:har_bin+1))];
end

%Determine the total distortion power
Pd=sum(Ph(2:5));
%Determine the noise power
Pn=sum(spectP(1:numpt/2))-Pdc-Ps-Pd;
format;
A=(max(code)-min(code))/2^numbit
AdB=20*log10(A)
SINAD=10*log10(Ps/(Pn+Pd))
SNR=10*log10(Ps/Pn)
disp('THD is calculated from 2nd through 5th order harmonics');
THD=10*log10(Pd/Ph(1))
SFDR=10*log10(Ph(1)/max(Ph(2:10)))
disp('Signal & Harmonic Power Components:');
HD=10*log10(Ph(1:10)/Ph(1))

%Distinguish all harmonics locations within the FFT plot
hold on;
plot(Fh(2)*fclk,0,'mo',Fh(3)*fclk,0,'cx',Fh(4)*fclk,0,'r+',Fh(5)*fclk,0,'g*',...
Fh(6)*fclk,0,'bs',Fh(7)*fclk,0,'bd',Fh(8)*fclk,0,'kv',Fh(9)*fclk,0,'y^');
legend('1st','2nd','3rd','4th','5th','6th','7th','8th','9th');
hold off;

Dynamic-Range Specifications, TTIMD

Two-tone IMD can be a tricky measurement, because the additional equipment required (a power combiner to combine two input frequencies) can contribute unwanted intermodulation products that falsify the ADC's intermodulation distortion. You must observe the following conditions to optimize IMD performance, although they make the selection of proper input frequencies a tedious task.

First, the input tones must fall into the passband of the input filter. If these tones are close together (several tens or hundreds of kilohertz for a megahertz bandwidth), an appropriate window function must be chosen as well. Placing them too close together, however, may allow the power combiner to falsify the overall IMD readings by contributing unwanted 2nd- and 3rd-order IMD products (depending on the input tones' location within the passband). Spacing the input tones too far apart may call for a different window type that has less frequency resolution.

The setup also requires a minimum of three phase-locked signal generators. This requirement seldom poses a problem for test labs, but generators have different capabilities for matching frequency and amplitude. Compensating such mismatches to achieve (for example) a -0.5dB FS two-tone envelope and signal amplitudes of -6.5dB FS will increase your effort and test time (see the following program-code extraction).

%For TTIMD, use the following short routine, normalized to -6.5dB full-scale.
%plot([0:numpt/2-1].*fclk/numpt,Dout_DB(1:numpt/2)-maxdB-6.5);
plot([0:numpt/2-1].*fclk/numpt,Dout_DB(1:numpt/2)-maxdB);
grid on;
title('FFT PLOT');
xlabel('ANALOG INPUT FREQUENCY (MHz)');
ylabel('AMPLITUDE (dB));
a1=axis; axis([a1(1) a1(2) -120 a1(4)]);

Conclusion

Besides the points above, many other issues confront an engineer trying to determine the dynamic range of a high-speed ADC by capturing its signals and analyzing them. Unfortunately, mistakes are easily made in spectral-measurement procedures. But this task of data acquisition and analysis is greatly eased by an understanding of FFT-based measurement and related computations, the effect of spectral leakage and how to prevent it, and the necessary layout techniques and equipment.

The MAX1444/MAX1446/MAX1448 EV kit was selected to properly evaluate dynamic performance for the 10-bit, 80Msps MAX1448.

For any of the equipment suggested, you can substitute an item more suitable for your specific application.

A stationary signal is present before, during, and after the data capture.

Performing an FFT with no apparent window function selected is frequently referred to as performing the FFT with a "uniform" or "rectangular" window.

Literature Sources

1. MAX1448 data sheet, Rev. 0, 10/00, Maxim Integrated Products.
2. MAX1448 EV kit data sheet, Rev. 0, 0/00, Maxim Integrated Products.

A similar version of this article appeared in the December 2000 issue of Microwaves and RF.