ANALOG DEVICES

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APPLICATION NOTE

ONE TECHNOLOGY WAY • P.O. BOX 9106 • NORWOOD, MASSACHUSETTS 02062-9106 • 617/329-4700

Test Video A/D Converters Under Dynamic Conditions

by Walt Kester

To adequately characterize video-speed analog-to-digital converters, collect and evaluate the test results from comprehensive dynamic measurements.

To check out video-speed analog-to-digital converters—those operating at sampling rates exceeding 10 MHz—use dynamic performance testing. Such testing accumulates data that accurately describes these converters' ability to digitize fast-acting analog input signals.

An alternative test approach traditionally used with low- to medium-speed converters, static testing proves deficient for high-speed devices because video A/D converters demonstrate virtually ideal digitizing transfer functions under dc or low-frequency ac stimulation. However, under dynamic stress—ac test signals approaching the Nyquist sampling-rate limit $f_S/2$—video-speed units often exhibit otherwise hidden shortcomings, such as missing codes, nonmonotonic conditions or linearity errors.

On the debit side, though, dynamic testing lacks the standardization of industry-accepted static testing. Consequently, users tend to focus only on those ac specs that highlight a particular video-converter application. For example, differential gain and phase are paramount in digital-video uses, and noise-power-ratio checks dominate evaluation of parts aimed at data-communication tasks.

Moreover, in practice, no single specification or test can completely characterize a video A/D converter's performance under dynamic conditions. Therefore, you must implement a variety of probing analog and digital tests. Equipped with the test results, however, you can then generate a set of converter specs that accurately describe those parameters affecting your application.

DAC eases ADC testing

Connecting a high-performance D/A converter to the video A/D converter under test permits dynamic performance measurements with conventional analog test equipment. To obtain valid test results, however, the DAC's static and dynamic performance must exceed that of the A/D converter by two bits or more.

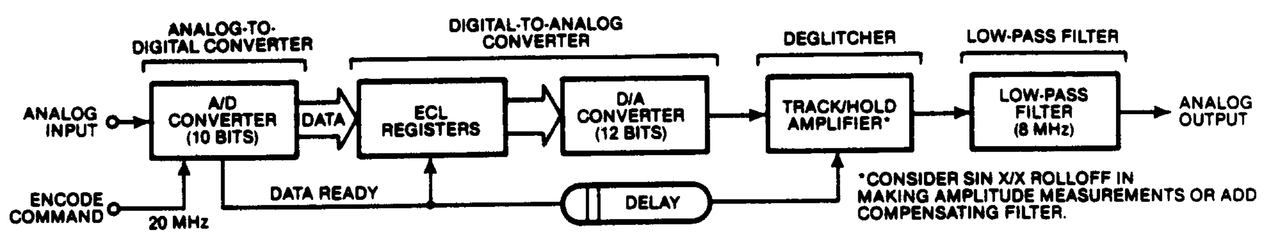

Fig 1—Connecting a high-performance digital-to-analog converter to the video A/D converter under test in a back-to-back setup allows you to perform dynamic measurements using common analog test equipment. The DAC, however, must function at least four times better than the A/D converter to ensure accurate test data. The track/hold amplifier/deglitcher removes undesirable harmonics or discontinuities from the output signal.

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Use a D/A converter to dynamically test an A/D converter

tunately, though, because many DACs achieve such higher performance, back-to-back A/D-to-D/A testing proves successful (**Fig 1**). In this method, the DAC must operate transparently over the input frequency range and with the A/D converter's sampling rate.

For accurate dynamic testing, the DAC's settling time must not surpass the period corresponding to the A/D converter's maximum sampling rate. In addition, a deglitcher connected to the DAC's output removes unwanted harmonics caused by DAC glitches or analog-output discontinuities. Essentially a track/hold amplifier, the deglitcher switches to its Hold mode immediately before the DAC gets updated. During DAC updating in the Hold mode, the glitch settles out. Then, the track/hold enters its Track mode and acquires the DAC's new output value, typically in less than 30 nsec.

Note that the deglitcher inserts a glitch in the DAC's output. However, the glitch occurs at the test system's sampling rate and is therefore filterable. For analysis purposes, a sinc function—a (sin x)/x curve—describes the deglitcher output's frequency response. This response stems from the deglitcher's signal-reconstruction process: Its output is a series of rectangular pulses whose widths equal the sampling frequency's reciprocal.

When this pulse stream passes through the low-pass filter, the attenuation at frequency f with a sampling rate of f_S equals

$$A_f = \frac{\sin(\pi f/f_s)}{\pi f/f_s}.$$

System measurements that depend on frequency response must take into account this theoretical rolloff or must be made using a compensating filter.

This sinc correction filter's output goes through the low-pass filter, which has a cutoff of approximately $f_S/2.2$. The filter's attenuation at frequencies equal to or greater than $f_S/2$ should be at least 10 dB higher than the A/D converter's dynamic range to prevent aliasing errors.

To ensure accuracy, how do you verify the DAC's and deglitcher's dynamic performance independently of the A/D converter? One method involves driving the DAC with a PROM loaded with an ideally quantized recirculating sine wave and examining the deglitcher's output for harmonics using a spectrum analyzer. If you store two sine waves of slightly different frequencies in PROM, you can measure dual-tone intermodulation products. A less desirable method centers on driving the DAC with an A/D converter having known characteristics.

SNR tests deserve priority

Probably the most powerful indicator of an A/D converter's dynamic performance is its signal-to-noise ratio (SNR) when stimulated by a spectrally pure sine wave. Ideally, an A/D converter generates $q/\sqrt{12}$ rms quantizing noise in an $f_S/2$ bandwidth, where q equals the weight of the least significant bit (LSB). This rms noise level is independent of the input sine wave's level and frequency so long as the level lies within the A/D converter's operating range.

Theoretically, the ratio of a full-scale sine wave's rms level to an ideal N-bit A/D converter's rms quantizing noise (measured over an $f_S/2$ bandwidth) equals

$$\text{SNR} = 6.02N + 1.8 \text{ dB}.$$

Practically, though, an A/D converter's noise floor increases with the full-scale sine wave's increasing input frequency; the corresponding SNR thus decreases. Conversely, holding the input frequency constant but reducing the sine-wave amplitude decreases the A/D converter's noise floor.

Fig 2 shows an analog test setup for measuring an A/D converter's SNR. During operation, a spectrally pure full-scale sine wave feeds the A/D converter, which is followed by several signal-processing stages. The output of these stages in turn drives a true-rms voltmeter, first bypassing the bandstop filter for

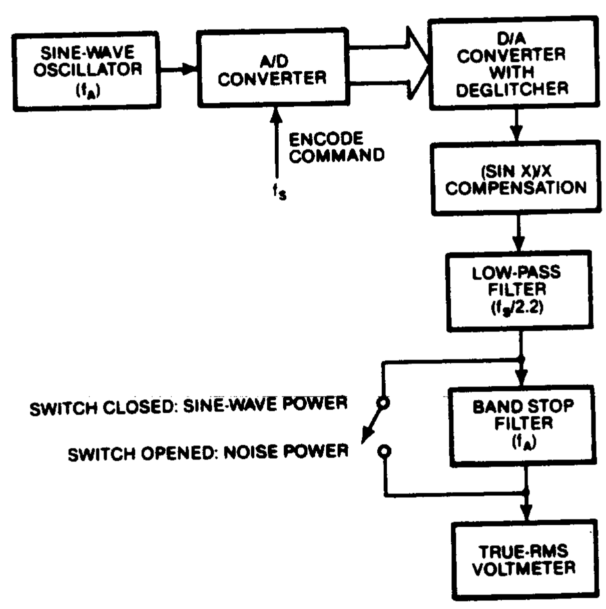

Fig 2—Use this test setup *to make analog signal-to-noise-ratio (SNR) measurements on video A/D converters. To operate, apply a spectrally pure full-scale sine wave to the converter. On the true-rms voltmeter, measure the signal-plus-noise level with the bandstop filter out of the circuit. Then, switch in the bandstop filter to remove the fundamental signal and to measure the rms noise level alone. The two readings, expressed in decibels, make up the signal-plus-noise-to-noise ratio.*

measurement of the signal's level plus noise. Switching in the bandstop filter then removes the fundamental signal to allow the rms voltmeter to measure the noise level alone. The ratio of these two readings expressed in decibels equals the ratio of the signal plus noise to the noise, which approximates the SNR for a signal at least 10 dB greater than the noise level.

Minimizing DAC effects

Fig 3 depicts a useful test method for reducing DAC effects in making back-to-back A/D-to-D/A measurements. The technique employs an analog input sine wave slightly lower in frequency than one-half the sampling frequency, and the registers driving the DAC get updated at an even submultiple of the sampling rate ($f_S/2N$). In turn, the DAC emits a sine wave having a frequency equal to the difference between one-half the sampling rate and the analog input frequency.

In test operation, the A/D converter receives a signal whose frequency approaches the device's Nyquist limit. The DAC, however, updates at a much lower rate ($f_S/2N$), thereby reducing the effects of glitches and other dynamic errors. You can therefore make signal-to-noise measurements over an $f_S/4N$ bandwidth. Use an oscilloscope to inspect the low-frequency beat for missing codes and other nonlinearities; a standard

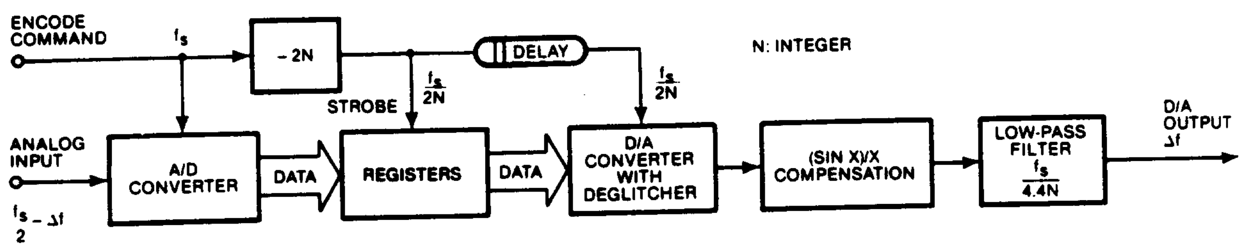

Fig 3—To reduce the DAC's effects on A/D-converter back-to-back measurements, consider this beat-frequency test setup. In this approach, the A/D converter, operating at its maximum sampling rate, receives a near-Nyquist-frequency signal ($f_S/2$). The DAC updates at a slower rate ($f_S/2N$), reducing glitches and dynamic errors. You can thus make signal-to-noise measurements over an $f_S/4N$ bandwidth. Examine the low-frequency beat on an oscilloscope for missing codes and nonlinearities and on a spectrum analyzer for harmonic content.

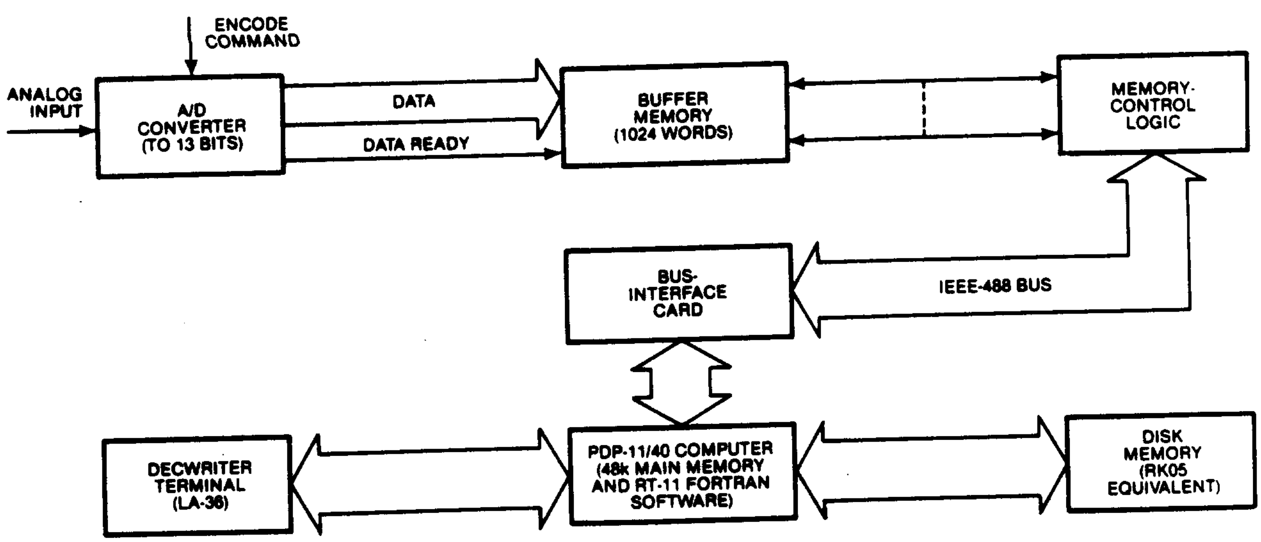

Fig 4—Digitally measuring a video A/D converter's signal-to-noise ratio (SNR) calls for a computerized test setup that eliminates the DAC and its potential errors. During operation, a spectrally pure sine wave feeds the A/D converter, and contiguous samples store in the high-speed buffer. The computer analyzes the stored data using a discrete Fourier transform to determine the A/D converter's true SNR and harmonic content.

·v-frequency spectrum analyzer lets you measure ·monic content.

Observe that the low-frequency-beat harmonics relate directly to the A/D converter's analog-input-frequency harmonics. In practice, a beat frequency of a few hundred kilohertz performs satisfactorily. Derive both the analog input sine wave and the sampling frequency from frequency synthesizers or crystals to prevent low-frequency-beat-signal smearing.

The beat-frequency test also proves effective in measuring the A/D converter's performance for input signals near the sampling frequency. For a typical test, make the A/D unit's input frequency slightly less than the sampling rate (f_S). Check that a low-frequency beat occurs even when the DAC updates at the sampling rate. Note, however, that updating the DAC at $f_S/2N$ reduces dynamic errors.

Measure SNR digitally

Measuring SNR using the methods previously described proves valid only when the DAC's static and dynamic performance exceeds that of the A/D converter under test by at least four times. To eliminate the DAC's errors, therefore, and to measure the A/D converter's SNR digitally, consider **Fig 4**'s test setup.

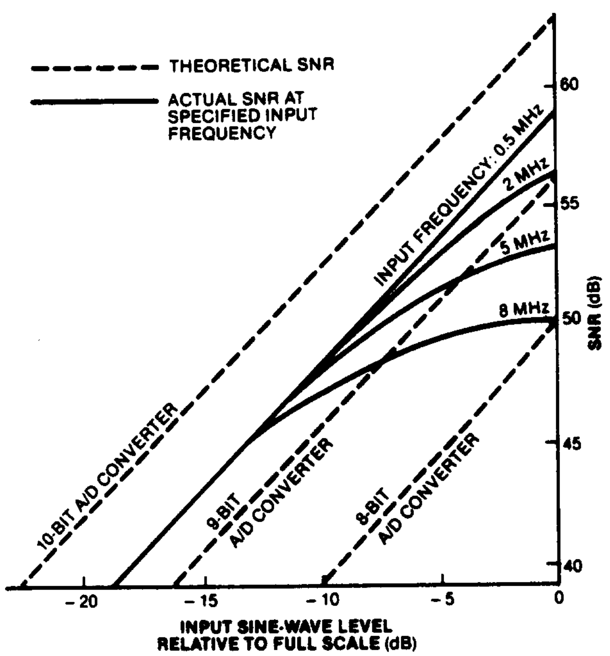

Fig 5—Plotted curves *help analyze a video A/D converter's theoretical and actual SNR test results. They track SNR values vs input sine-wave levels for 8-, 9- and 10-bit converters. With this information, you can determine the number of effective resolution bits for various input signal amplitudes and frequencies.*

To initiate testing, apply a spectrally pure sine wave to the A/D converter. Store contiguous samples in high-speed buffer memory for computer analysis using a discrete Fourier transform (DFT) or a fast Fourier transform (FFT). Note that although the FFT algorithm reduces processing time, the DFT mathematics involves less complexity and thus becomes the preferred method.

To use a DFT, first determine an appropriate time-weighting function for the A/D converter samples in order to reduce frequency side lobes. Exercise care in this determination, though: Selecting an improper weighting function spills main-lobe energy into other frequency bands and makes noise measurements impossible. Accordingly, a $\cos^2$ function without a pedestal (often termed Hanning weighting) is appropriate. The input data multiplied by the weighting function thus becomes

$$WD_n = D_n\left[0.5 - 0.5\cos\left(\frac{2\pi n}{N}\right)\right],$$

where:

WD_n=nth weighted data sample
D_n=nth input data sample
N=total number of samples.

This weighting function compresses the spillover energy into a few frequency bands centered on the fundamental sine-wave frequency. The result? It eliminates contamination over most of the spectrum.

Next, program a computer to establish the DFT of the sequence of weighted data samples for N/2 frequencies. The program must solve the following two equations for the Kth frequency:

$$A_K = \frac{1}{N}\sum_{n=1}^{N} WD_n \cos\left[\frac{2\pi k\,(n-1)}{N}\right]$$

$$B_K = \frac{1}{N}\sum_{n=1}^{N} WD_n \sin\left[\frac{2\pi k\,(n-1)}{N}\right],$$

where A_K and B_K represent the magnitudes of the Kth spectral line's cosine and sine parts. The line's magnitude therefore equals

$$MAG_K = \sqrt{A_K^2 + B_K^2}.$$

The computer uses this equation to determine SNR. It then computes the residual noise floor in all frequency bands except those occupied by the input signal. During execution, the program squares the magnitude of each frequency band that doesn't contain the input signal, adds all N/2 squared magnitudes and evaluates the following square root:

$$NOISE = \sqrt{\sum_{K=1}^{N/2} MAG_K^2}.$$

Set several frequency bands on either side of the fundamental sine-wave frequency band to zero to account for main-lobe widening caused by the $\cos^2$ weighting function.

For the final SNR test, measure harmonic distortion using the following expression:

$$\text{HARMONIC DISTORTION} = 20 \log \left[\frac{MAG_f}{MAG_{mf}} \right],$$

where:

MAG_f=magnitude of fundamental

MAG_{mf}=magnitude of mth harmonic.

In making digital SNR measurements, don't lock the fundamental input sine-wave frequency to the sampling frequency or its harmonics. If you do, the sample-to-sample variations in the sampled input signal and the A/D converter's encoded output voltage become predictable and repeatable. Obeying this rule achieves measurement results that indicate true SNR and harmonic content under normal A/D-converter operating conditions, where sample-to-sample variations are unpredictable and nonrepeatable.

Analyzing SNR results

You can display and analyze A/D-converter SNR test results in several ways. For example, you can plot SNR versus sine-wave frequency for a constant-amplitude full-scale sine-wave input. You can also plot SNR versus input-signal amplitude for a fixed sine-wave frequency. What's more, you can combine the two plots (**Fig 5**). Using these curves, you can determine the number of effective bits of resolution for various amplitudes and frequencies and compare them with the theoretical values.

Note that you can measure the SNR for input signals greater than the Nyquist limit ($f_S/2$). Be aware, however, that the fundamental sine wave then appears as an alias component within the $f_S/2$ bandwidth. For example, a 12-MHz sine wave's in-band component sampled at 20 MHz occurs at 8 MHz.

NPR tests demand attention

Another important indicator of an A/D converter's dynamic performance centers on noise-power-ratio (NPR) testing. This ratio testing finds extensive use in measuring the transmission characteristics of frequency-division-multiplexed (FDM) communications links. In a typical FDM system, 4-kHz-wide voice channels get stacked in frequency for transmission over coaxial, microwave or satellite equipment. At the receiving end, the FDM data becomes demultiplexed and returned to 4-kHz individual baseband channels. In an FDM system having more than about 100 channels, you can approximate the FDM signal by Gaussian noise with the appropriate bandwidth. **Fig 6** illustrates how to measure a 4-kHz channel for quietness using a narrow-band notch (bandstop) filter and a special tuned receiver. The receiver measures the noise power within the 4-kHz notch.

To implement testing, you first switch out the notch filter. Then using the receiver, measure the rms noise power of the signal within the notch. Next, switch in the notch filter and evaluate the residual noise within the slot. NPR thus equals the ratio of the two readings expressed in decibels. Check several slot frequencies across the noise bandwidth (eg, low, midband and high) to adequately characterize the FDM system.

For analysis purposes, plot NPR as a function of rms noise level referred to the FDM system's peak range. At low noise-loading levels, undesired noise exists primarily as thermal activity, independent of the input noise level. Over this region of the NPR curve, a 1-dB increase in noise level causes a 1-dB increase in NPR. At increased noise-loading levels, the FDM system amplifiers overload, creating intermodulation products that increase the noise floor. With further input-noise

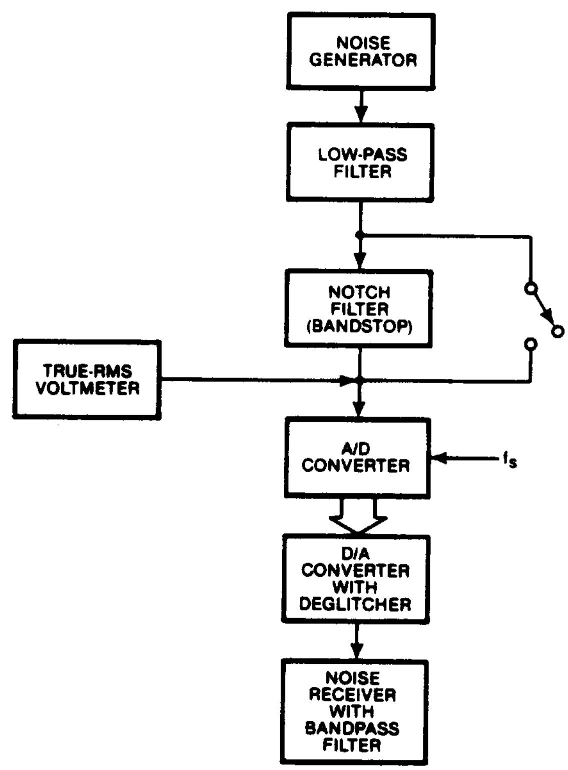

Fig 6—In this noise-power-ratio (NPR) test setup, *the special tuned receiver measures the input signal's rms noise power inside a narrow-band notch with the bandstop filter out of the circuit. Next, it measures the noise power with the bandstop filter included. NPR equals the ratio of the two readings expressed in decibels.*

increases, overload noise effects predominate and dramatically reduce NPR. In practice, FDM systems usually operate at a noise-loading level of a few decibels below the point of maximum NPR.

A digital system containing an A/D converter primarily generates quantizing noise within the notch filter's slot on receiving low values of noise input signals. The NPR curve tracks linearly in this region. As the noise input level increases, though, clipping noise caused by the A/D converter's hard-limiting action starts to prevail. **Fig 7** presents a set of theoretical NPR curves for various A/D converters.

In a practical A/D converter, however, dc or ac nonlinearities cause departures from the theoretical NPR values. Although the NPR's peak value occurs at a low input-noise level (rms noise=$\frac{1}{4}V_0$, where $\pm V_0$=A/D converter's input range), the broadband nature of the noise signal stresses the A/D converter. Consequently, NPR tests provide worthwhile dynamic-performance measurements.

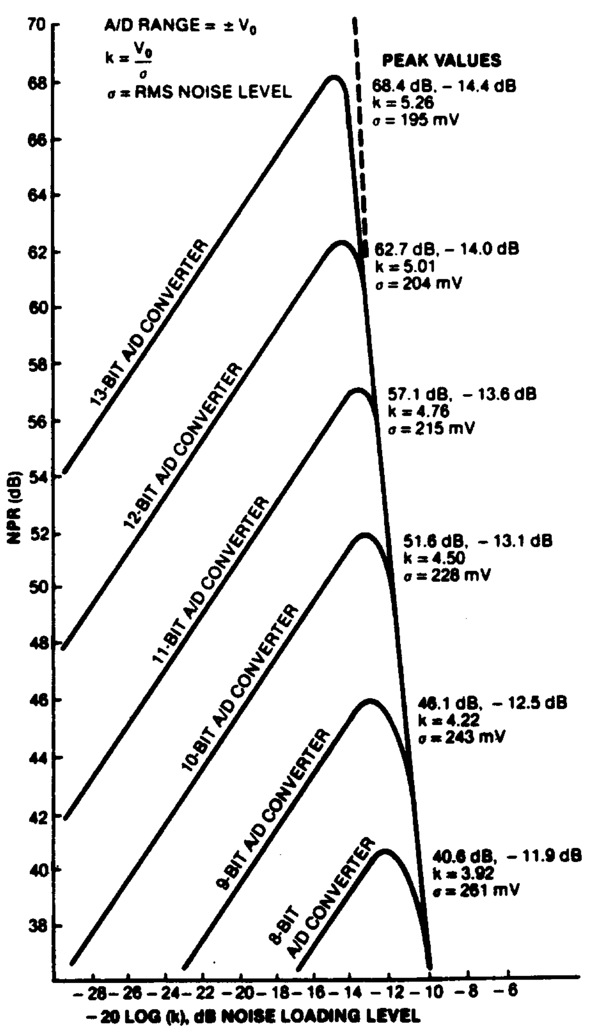

Fig 7—In A/D-converter-based digital systems, *theoretical NPR curves are linear at low noise-input-signal levels. As the noise input levels increase, however, the A/D converters produce clipping noise because of hard-limiting performance. Note that the NPR values peak at corresponding higher levels and at lower rms noise levels as converter bit resolutions increase.*

Histograms disclose nonlinearity effects

Another key parameter in video A/D-converter dynamic testing is differential nonlinearity, which often degrades as the input-signal frequency increases. You can employ histograms to help analyze this problem: To obtain a histogram for an A/D converter, make the sine-wave input frequency noncoherent with the sampling frequency. Then, take many A/D-converter samples and use a computer to calculate the number of times each output code occurs and to plot these results directly. Analyze the results for missing codes and differential nonlinearity.

Because the probability of obtaining digital codes around the zero-crossing point proves less than the probability of obtaining codes around the positive or negative peaks, the resulting curve traces a cusp shape rather than a flat one. Normalize this curve by using the sine wave's probability density function. Or apply a triangular waveform, rather than a sine wave, to the A/D converter.

Obtaining a histogram for a 10-bit A/D converter requires a large buffer memory. To get 100 samples per code, for example, buffer memory must contain 100×1024 or 102,400 locations.

The histogram approach furnishes accurate representations of an A/D converter's statistical differential nonlinearity. However, a video A/D converter's actual differential-nonlinearity characteristic might depend on the input signal's direction of change and rate of change. Thus, positive-slewing signals might produce different differential nonlinearity characteristics than negative-slewing signals. This subtle effect clouds the histogram's results for A/D converters used in transient- or pulse-analysis applications, where the encoded signals are not statistical in nature.

Aperture-time measurements differ

Heading the list of misunderstood and misused video-A/D-converter specifications is aperture time. Originally, aperture-time measurements centered on the classic sample/hold circuit (EDN, April 14, pg 41). Ideally, a sample/hold's switch possesses zero resistance when closed and opens instantly. In practice, though, the sampling switch transits from a low to high resistance over some finite time interval. Accordingly, the original definition holds—an error occurs because the input signal gets averaged over the finite time interval required for opening the switch. The sampled

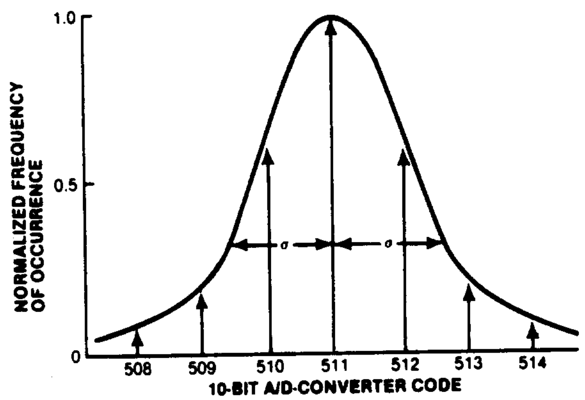

Fig 9—An ideal A/D converter *without aperture jitter shows only one code on a histogram. A practical converter exhibits a distribution of codes. The distribution's calculated sigma (σ) value corresponds to the rms error voltage produced by the rms aperture jitter.*

converters that operate in Continuous mode, in which the encode command is always present. Some applications, however, require that an A/D converter operate in Burst mode, in which the ADC operates on a finite number of encode-command pulses. This mode commonly occurs in radar return-pulse analysis because the A/D converter functions in an idle condition except for the brief time that encompasses each return pulse.

Burst-mode operation usually places more stringent requirements on the A/D converter than continuous operation. In a continuous mode, timing and track/hold circuits reach steady-state conditions. In Burst mode, however, samples taken before reaching steady-state conditions produce errors. A track/hold that uses a transformer to couple the switching pulses to the diode bridge proves particularly error susceptible.

Fig 10 outlines a test setup for evaluating Burst-mode converter operation. Use the variable delay to slide the burst of encode commands across the flat pulse. Analyze the results of each sample pulse with a computer or appropriate gating circuits on the A/D converter's output. Each sample's results should be identical except for the time delay that separates each individual pulse in the burst. Differences in results indicate the presence of analog memory, which might limit the A/D converter's use in certain applications.

A preferred method for handling Burst-mode applications calls for operating the A/D converter continuously and gating the digital outputs into a memory during the desired command intervals. This method relieves unnecessary stress on the A/D converter and accomplishes the same result as actual Burst operation.

Capturing bandwidth clipping

You can use **Fig 11**'s test setup to measure a video A/D converter's frequency response and bandwidth (small or large signal). During checkout, hold the input-signal amplitude constant while varying the input frequency over the desired range. Analyze the A/D converter's output digitally or with a DAC followed by an rms voltmeter. Using a voltmeter, however, limits the filtered DAC's output frequency to $f_S/2.2$.

A quick method for determining an A/D converter's large-signal bandwidth involves adjusting the input

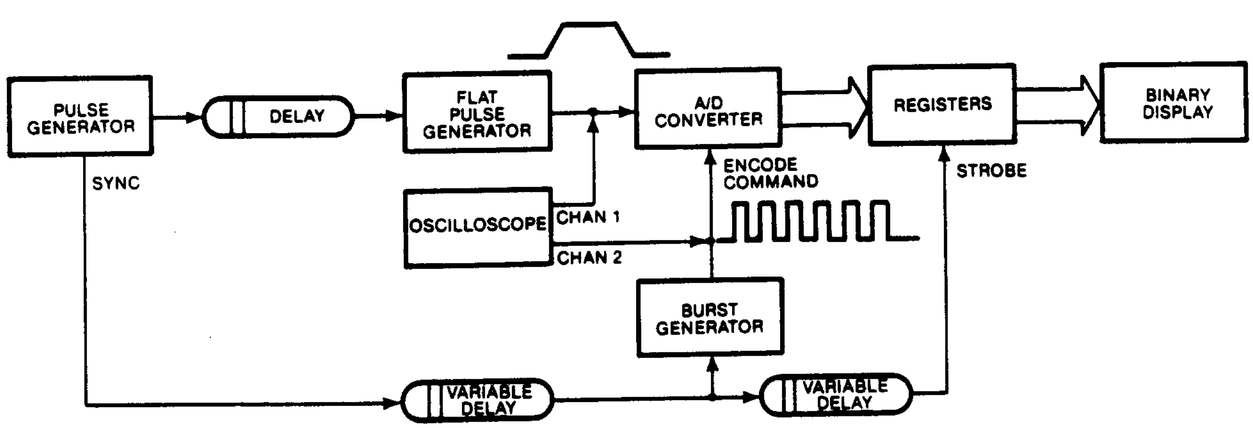

Fig 10—To check out an A/D converter's Burst mode, *in which it operates as directed by a finite number of encode-command pulses, you adjust the variable delays in this test setup to slide the burst of encode commands across the flat input pulse to the A/D converter. Analyze the sample pulses with a computer or appropriate gating circuits connected to the converter's output. Each sample should produce identical test results except for the time delay that separates each pulse in the burst.*

voltage, therefore, does not exactly correspond to the voltage at the instant the switch starts to open, and the time required to open the switch represents the aperture time. The aperture error, according to the original definition, therefore equals

$$E_a = t_a dV/dt, \qquad (1)$$

where:

E_a = aperture error

t_a = aperture time

dV/dt = rate at which the input signal changes.

A simple first-order analysis, which neglects nonlinear effects, reveals that no real *aperture* error exists for such a switch. Instead, so long as the switch opens repeatably, an effective fixed *sampling-time delay* exists—the time between encode-command arrival and the opening of the ideal zero-opening-time switch. (Note that you can include logic propagation delays in the encode-command path to modify this time interval.) Unfortunately, most manufacturers still refer to this delay as aperture time.

True aperture errors, on the other hand, result from variable time delays. These errors generally emanate from several sources: In a practical A/D converter, the encode-command signal often gets phase-modulated by an unwanted source—random noise, power-line frequency or digital noise—because of faulty grounding techniques. You can express the resulting error in terms of an rms time jitter, termed aperture jitter. The corresponding rms voltage error caused by aperture jitter qualifies as a valid aperture error.

An A/D converter's aperture-jitter spec sometimes gets interpreted as a measure of the device's ability to encode rapidly changing input signals accurately. However, although important, the aperture-jitter specification does not define the whole encoding process. Therefore, an A/D converter with an impressive aperture-jitter specification still might not accurately digitize a sine wave having a maximum slew-rate calculated from **Eq** 1.

Typical aperture-time test examples

For example, assume that a 20-MHz, 10-bit A/D converter has a $\pm V_0$ bipolar input range ($2V_0$ p-p) and a 10-psec rms aperture-jitter specification. To calculate the maximum aperture error, convert the rms aperture jitter into a maximum value rather than an rms one. If you assume that aperture jitter follows a Gaussian distribution similar to white noise, the rms aperture jitter (t_a) corresponds to the distribution's σ value. The distribution's 2σ point thus becomes the proper place to set the maximum value; the maximum aperture jitter therefore equals $2t_a$.

If you set the corresponding maximum voltage error (ΔV) at a full-scale sine wave's zero crossing to ½ LSB (where ½ LSB$=2V_0/2^{N+1}$ and N=number of resolution bits), you can then calculate the maximum full-scale sine-wave frequency (f_{MAX}) that produces the ½-LSB aperture error:

$$v(t) = V_0 \sin 2\pi ft \qquad (2)$$

$$\frac{dv}{dt} = 2\pi f V_0 \cos 2\pi ft \qquad (3)$$

$$\left.\frac{dv}{dt}\right|_{MAX} = \frac{\Delta v}{2t_a} = 2\pi V_0 f_{MAX},$$

thus yielding

$$f_{MAX} = \frac{\Delta v}{4\pi V_0 t_a} = \frac{2\pi V_0/2^{N+1}}{4\pi V_0 t_a} = \frac{1}{2\pi t_a 2^{N+1}}.$$

If t_a=10 psec rms and N=10, f_{MAX}=7.8 MHz. These calculations imply that the 20-MHz A/D converter under test can accurately digitize a 7.8-MHz FS sine wave. In practice, however, the A/D converter might suffer from skipped codes, decreased SNR and ac nonlinearities at much lower frequencies. These other limitations to good high-frequency performance, might, in fact, eclipse errors caused by the sample/hold's aperture jitter.

Similarly, you can calculate aperture-jitter effects on full-scale sine-wave SNR using **Eqs** 2 and 3 and the following expression:

$$\left.\frac{dv}{dt}\right|_{RMS} = \frac{2\pi f V_0}{\sqrt{2}}. \qquad (4)$$

Substituting rms error voltage (ΔV_{RMS}) and rms aperture jitter (t_a) into **Eq** 4 yields

$$\frac{\Delta v_{RMS}}{t_a} = \frac{2\pi f V_0}{\sqrt{2}}$$

$$\Delta v_{RMS} = \frac{2\pi f V_0 t_a}{\sqrt{2}}.$$

The rms-signal-to-rms-noise ratio expressed in decibels thus becomes

$$\text{SNR} = 20\log\left[\frac{V_0/\sqrt{2}}{\Delta v_{RMS}}\right] = 20\log\left[\frac{1}{2\pi f t_a}\right]\ \text{dB}. \qquad (5)$$

Consider again the 10-bit, 20-MHz A/D converter with a 10-psec rms aperture jitter. For an 8-MHz full-scale input, the SNR arising only from aperture jitter totals 66 dB, as calculated from **Eq** 5. The theoretical SNR due to quantizing noise in a 10-bit A/D converter equals 62 dB. Combining the two SNR decibel values results in a theoretical 60.5-dB SNR; this final value encompasses the ideal quantizing noise and the aperture-jitter noise. From **Fig** 5's curves, observe that a practical A/D converter having the same aperture-jitter spec achieves only a 50-dB SNR.

From this analysis, you can logically conclude that the SNR specification more closely defines video-A/D-converter performance than does the aperture-jitter specification. Relying solely on the aperture-jitter specification therefore creates an erroneous deduction about the converter's actual dynamic performance.

Measuring aperture jitter

Fig 8 depicts the test setup for measuring an A/D converter's aperture jitter. The encode-command signal and the analog input signal derive from the same low-jitter pulse generator to minimize the phase jitter between them.

To test, adjust the phase shifter until the A/D converter repetitively samples the sine wave at its midscale point of maximum slew rate. Set the bit switches to the A/D converter's midscale code: 0111...1. Adjust the phase shifter again for a maximum reading on the frequency counter. Examine the midscale code's frequency of occurrence as well as the frequency of occurrence of several codes above and below this value. Using this information, plot a histogram (**Fig 9**).

An ideal A/D converter with no aperture jitter would present only one code on the histogram. In contrast, a practical converter exhibits a distribution of codes. The distribution's calculated σ corresponds to the rms error voltage (ΔV_{RMS}) produced by the rms aperture jitter (t_a). Calculate the aperture jitter (t_a) as follows:

$$t_a = \frac{\Delta V_{RMS}}{dV/dt},$$

where dV/dt equals the sine wave's rate of change at a zero crossing.

If you attenuate the input sine wave sufficiently, the distribution spread around the nominal code results from intrinsic A/D-converter noise. As the input sine wave increases in amplitude, the slew rate (dv/dt) becomes proportionally greater, and the distribution begins to spread because of aperture jitter. Take care when interpreting a histogram for high-slew-rate inputs because high slew rates also affect the converter's ac differential linearity.

The offset trimming pot lets you position the sine wave at different A/D-converter range points. In this manner, you can plot histograms around several nominal codes and attribute variations to range-dependent differential-linearity characteristics. Don't exceed the A/D converter's range during this offset adjustment, however.

The tests presented so far aim at video A/D

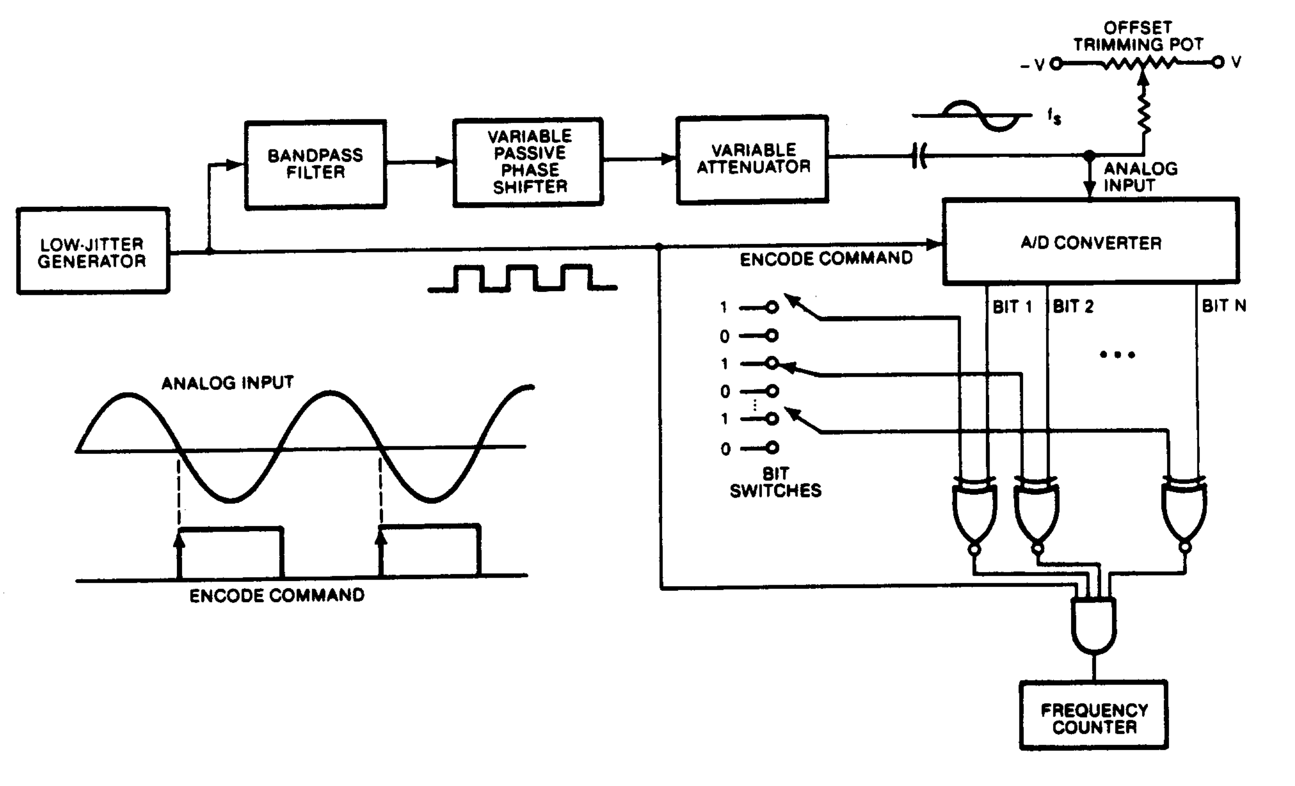

***Fig 8—This test setup measures an A/D converter's aperture jitter.** To operate, you adjust the phase shifter until the A/D converter repetitively samples the analog input sine wave at its midscale point of maximum slew rate. Next, you set the bit switches to the converter's midscale code. Then, you readjust the phase shifter for a maximum reading on the frequency counter. The frequency of occurrence of the midscale and adjacent codes indicate aperture jitter when plotted on a histogram.*

Be sure to thoroughly understand aperture-time-spec variations

sine wave until the DAC indicates A/D-converter output clipping; observe this condition on an oscilloscope connected across the DAC's output, before deglitching and filtering stages. Increase the input frequency and readjust the input level to cause clipping. An increase in input-signal level required to cause clipping corresponds to a decrease in A/D converter gain; a decrease in input level implies an increase in gain. This test method also yields frequency

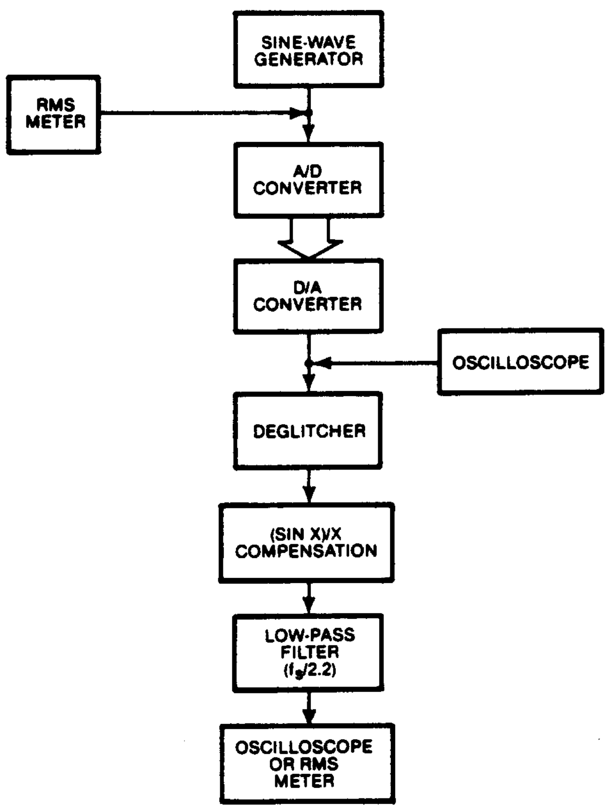

Fig 11—This test setup *measures an A/D converter's frequency response and small- or large-signal bandwidth. During testing, you hold the input signal amplitude constant while varying the input frequency over the desired range. You can analyze the converter's output digitally or with a DAC followed by an rms voltmeter.*

measurements above the Nyquist limit.

Note that a good bandwidth specification does not always ensure that the A/D converter's dynamic performance is satisfactory at higher input frequencies. Thus, an A/D converter having a wide bandwidth might still exhibit inadequate SNR, NPR or harmonic distortion at higher frequencies; carefully check it out.

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