SECTION 9

INTRODUCTION TO HIGH SPEED SIGNALS AND SYSTEMS

- HIGH SPEED SIGNAL LEVELS, IMPEDANCES, AND NOISE
- DYNAMIC RANGE OF HIGH SPEED SYSTEMS
- DEFINING HIGH SPEED SYSTEM DYNAMIC RANGE

SECTION 9

INTRODUCTION TO HIGH SPEED SIGNALS AND SYSTEMS

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In the first seven sections of this book, we examined systems for processing dc or low-frequency *precision* signals. As we saw, even though the frequency content of the signals is low, the signal processing *components* (amplifiers, multiplexers, ADCs) operate at much higher clock frequencies and have higher bandwidths. This requires close attention to fundamental high speed design techniques as well as the fundamental precision design techniques normally associated with precision measurement systems.

In the previous section (Section 8), we discussed various aspects of *audio* signal processing. Although audio signals generally occupy the bandwidth between 20 and 20kHz, we saw that the bandwidths of the amplifiers, ADCs, and other audio processing components may well extend into the hundreds of MHz. These high bandwidths are necessary to achieve the extremely low levels of distortion required in professional audio applications. For example, high speed amplifiers usually classified as *video* amplifiers, such as the AD811, are often used in audio applications to achieve low levels of distortion.

In this section and those to follow, we will primarily examine the processing of signals which lie above the voice and audio frequency range. These signals are somewhat arbitrarily classified as *high speed* signals, at least for the purposes of differentiation. With a few exceptions, most of the definitions and important characteristics also apply to *voiceband* and *audio frequency* signals, which are referred to in this section as *midband signals*.

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TYPICAL HIGH SPEED SYSTEM APPLICATIONS

- Instrumentation: Digital Oscilloscopes and Digital Spectrum Analyzers
- Video: Professional and Consumer
- Medical: Ultrasound and Digital X-Ray
- Radar Receivers
- Communications: Data Transmission, Broadband Receivers
- Direct Digital Synthesis
- Arbitrary Waveform Generation

Figure 9.1

The most important goal in high speed signal processing is to maintain the fidelity of the signal to the required accuracy (ac and dc) throughout the entire signal processing path. A good understanding of each element in the signal chain is therefore required in order to make intelligent design tradeoffs. Different applications (see Figure 9.1) require different levels of accuracy. In this section, we will first examine the key specifications which describe the fidelity of high speed signals. Some of these are applicable to low and mid-frequency applications, while others are commonly used only in high speed systems.

POPULAR MEASURES OF HIGH SPEED SIGNAL FIDELITY

- Bandwidth and Settling Time
- Harmonic Distortion and Total Harmonic Distortion
- Signal-to-Noise Ratio
- Two-Tone Intermodulation Products (IMD)
- Third Order IMD Intercept Point
- Spurious Free Dynamic Range (SFDR)

Figure 9.2

High speed signals may be specified in either the time or the frequency domain, or both. However, all elements in the signal processing path may not be specified in both domains. For instance, ADCs are often specified in terms of effective number of bits (ENOBs) at various input frequencies. However in digital oscilloscope and transient analyzer applications, one is also interested in the accuracy with which pulse edges are digitized. This type of data is usually not explicitly specified on ADC data sheets.

It is therefore important to be able to determine approximate time domain specs when only frequency domain specs are given, and vice versa. This may be done quite easily using the simple relationship between bandwidth and risetime for a first-order system: Bandwidth = 0.35/Risetime. Another useful formula relates the time constant and the risetime: Risetime = 2.2×Time Constant. It should be emphasized that these relationships should be used with caution when dealing with high speed, high resolution systems, nevertheless, they may be used for first-order approximations.

The pulse response of a single-pole system may be expressed using the well-known exponential relationship: $V_{out}(t) = V_0(1- e^{-t/\tau})$, where τ is the time constant. The settling time to within ε of the final value may be calculated by solving the expression for t using $t = -\tau \ln(\varepsilon)$. Applying this to systems having 10 or more bits of resolution, however, will probably give erroneous results due to second-order effects.

USEFUL RELATIONSHIPS BETWEEN PULSE AND FREQUENCY RESPONSE FOR FIRST-ORDER SYSTEM

- Bandwidth = 0.35 / Risetime
- Risetime = 2.2 × Time Constant
- Settling Time to Within ε of Final Value
  = – Time Constant × $\ln(\varepsilon)$
- Use These Approximations With Caution at High Speeds and / or High Resolutions Due to Second-Order Effects

Figure 9.3

HIGH SPEED SIGNAL LEVELS, IMPEDANCES, AND NOISE

High speed signal levels and impedances are typically lower than those encountered in mid- and low frequency applications. The low impedances (usually less than 1000Ω) are required in order to maintain sufficient bandwidth in the presence of stray capacitance. Transmission line effects require the use of terminated cables and microstrip printed circuit board techniques in order to minimize reflections. The lower impedances (typical cable impedances of 50, 75, and 93Ω) require proportionally larger drive currents in order to maintain reasonable signal amplitudes. While 10V signals at impedance levels of 600Ω or greater are common at low- and audio frequencies, high speed signal amplitudes tend to be limited to a few volts peak-to-peak. Filtering is critical in all signal processing systems. At high speeds, analog filters (usually passive) are generally designed to operate at standard impedance levels of 50, 75, or 93Ω.

There is an inherent tradeoff between signal amplitude and signal-to-noise ratio. The lower signal amplitudes associated with high speed signals, coupled with the wide bandwidths, increase the susceptibility of circuits to noise. Even though lower impedance levels imply lower resistor Johnson noise (a 100Ω resistor generates 1.29nV/√Hz voltage noise), this is somewhat offset by the wider bandwidths associated with high speed systems (thermal noise voltage is proportional to the square root of the bandwidth).

Another challenge associated with high speed systems is maintaining low levels of signal distortion at high frequencies. Many applications such as digital spectral analysis require that high speed signals be processed with minimal distortion. It is not uncommon for systems to require distortion levels compatible with 14 bits of resolution (84dB) at frequencies of 5 to 10MHz. This places severe constraints on any component in the signal processing chain.

A FEW OF THE MANY HIGH SPEED SYSTEM DESIGN CHALLENGES

- Impedance Levels are Low, Therefore Thermal Noise is Reduced

  BUT

- Signal Levels are Lower
- Bandwidths are High, and

  Thermal Noise = $\sqrt{4kTR \times \text{Bandwidth}}$

- Many Applications Require Low Levels of Distortion (>80dBc)

Figure 9.4

To illustrate the importance of low impedances for reduced thermal noise, consider the case of the AD9014 ADC which has 14 bits of resolution and a maximum sampling rate of 10MSPS. The input bandwidth is 60MHz, and the input signal range is 2V p-p, or 0.707V rms. The weight of the least significant bit (LSB) is 2V/16,384 = 125μV. This ADC has an input termination resistance of 75Ω. The thermal noise of a 75Ω resistor in the 60MHz bandwidth is approximately 9μV rms. The corresponding fullscale rms signal to rms noise ratio is 98dB (due to the resistor noise alone). If the input resistance were increased to 1000Ω, however, the rms noise over the same bandwidth would be 33μVrms, and the SNR (due to the resistor noise) would drop to 87dB. (Compare this to the theoretical 14 bit SNR of 86dB). For this reason, among others, optimum SNR performance with high speed devices is generally achieved at low impedance levels.

WHY IMPEDANCE LEVELS IN HIGH SPEED SYSTEMS ARE USUALLY LESS THAN 1000Ω

- Consider the AD9014 14bit, 10MSPS ADC:
  - Input Impedance = 75Ω
  - Input Signal = 2V peak-to-peak
  - Least Significant Bit = 125μV
  - Input Bandwidth = 60MHz
- A 75Ω Input Resistor Generates 9μV rms noise over 60MHz, Yielding SNR = 98dB
- A 1000Ω Input Resistor Generates 33μV rms over 60MHz, Yielding SNR = 87dB
- A Perfect 14 bit ADC has an SNR of 86dB

Figure 9.5

DYNAMIC RANGE OF HIGH SPEED SYSTEMS

Dynamic range requirements in high speed systems have increased dramatically over the last several years. Digital Signal Processing techniques are replacing traditional analog techniques. The need for large signal-to-noise ratios and low distortion have made the selection of appropriate amplifiers and ADCs a challenge.

HIGH SPEED SIGNAL PROCESSING DYNAMIC RANGE REQUIREMENTS (APPROXIMATE)

APPLICATION	SIGNAL BANDWIDTH	HARMONIC DISTORTION, SFDR	SIGNAL-TO-NOISE RATIO
Professional Video (HDTV)	6 to 30MHz	–50 to –60dBc	50 to 60dB
Medical Ultrasound Imaging	2 to 15MHz	–50 to –70dBc	45 to 60dB
Digital Ocsilloscopes	dc to 1GHz	–35 to –50dBc	35 to 50dB
Spectrum Analyzers	1 to 10MHz	–70 to –90dBc	60 to 70dB
Broadband Receivers	2 to 30MHz	–40 to –90dB	45 to 70dB

Figure 9.6

Several high speed applications are listed in Figure 9.6 along with approximate bandwidth, distortion, and noise requirements. Note that some applications such as spectral analysis can deal with noise using techniques such as averaging, or narrowing the scanning bandwidth. In these applications, harmonic distortion and spurious free dynamic range are the most important considerations. In other applications, however, both harmonic distortion and noise are important.

Defining High Speed System Dynamic Range

The dynamic range of a high speed system may be defined in several ways. We have already discussed the mid-frequency specifications Harmonic Distortion, Total Harmonic Distortion (THD), and Total Harmonic Distortion Plus Noise (THD + N). These definitions are summarized in Figure 9.7.

COMPARISON BETWEEN DYNAMIC RANGE TERMS FOR MID-FREQUENCY AND HIGH-FREQUENCY SYSTEMS

MID-FREQUENCY	HIGH-FREQUENCY
Harmonic Distortion	Harmonic Distortion
Total Harmonic Distortion THD	Total Harmonic Distortion THD
Total Harmonic Distortion Plus Noise THD + N	Signal-to-Noise Ratio Plus Distortion S/N + D, SNR
Signal-to-Noise Ratio SNR	Signal-to-Noise Ratio Without Harmonics, SNR
Spurious Free Dynamic Range SFDR	Spurious Free Dynamic Range SFDR
----	Intermodulation Distortion IMD
----	Third Order IMD Intercept

Figure 9.7

At high speeds, dynamic range may be defined using slightly different terminology. The spectral output of a signal (including noise components) is used to calculate these terms. The terms Harmonic Distortion and Total Harmonic Distortion have the same meaning at the higher speeds. The term *Signal-to-Noise Plus Distortion (S/N+D)* is often used instead of THD + N. The term *Signal-to-Noise (SNR)* is often used interchangeably with S/N + D. When the SNR measurement is made with the harmonics removed, it is usually referred to as *SNR Without Harmonics,* although in some cases *SNR* may be used. In all cases, careful study of the datasheet definitions will enable you to differentiate between these terms.

The distortion component which makes up Total Harmonic Distortion (THD) is usually calculated by taking the root sum of the squares of the first five or six harmonics of the fundamental. In many practical situations, however, there is negligible error if only the second and third harmonics are included.

DEFINITIONS OF THD AND THD + N

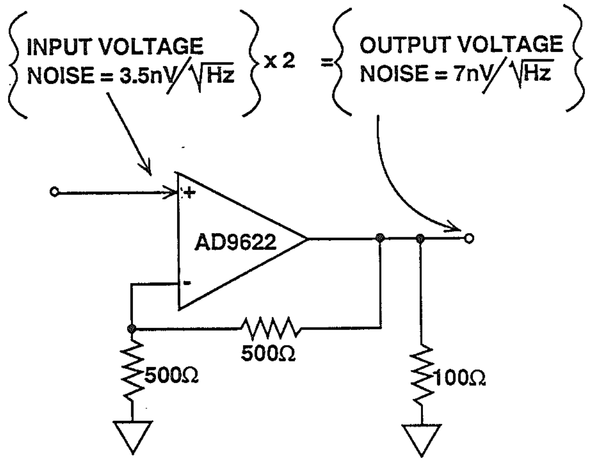
- V_S = Signal Amplitude (rms Volts)
- V_2 = Second Harmonic Amplitude (rms Volts)
- V_n = nth Harmonic Amplitude (rms Volts)
- V_{noise} = rms value of noise over measurement bandwidth
- $$\mathrm{THD+N} = \frac{\sqrt{V_2^2 + V_3^2 + V_4^2 + \ldots + V_n^2 + V_{noise}^2}}{V_S}$$
- $$\mathrm{THD} = \frac{\sqrt{V_2^2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

Figure 9.8

It is important to note that the THD measurement does not include noise terms, while THD + N does. The noise in the THD + N measurement must be integrated over the measurement bandwidth. In narrow-band applications, the level of the noise may be reduced by filtering. On the other hand, harmonics and intermodulation products which fall within the measurement bandwidth cannot be filtered, and therefore may limit the system dynamic range.

Consider the example of the simple wideband op amp circuit shown in Figure 9.9. The AD9622 is a high speed low distortion voltage feedback op amp optimized for use in a gain-of-two configuration. The input voltage noise of 3.5nV/√Hz is reflected to the output by multiplying by the noise gain, 2. The 7nV/√Hz is then integrated over the closed loop small signal bandwidth of the op amp, which is approximately 230MHz. This yields a total integrated output noise of 133μV rms. The output noise due to the op amp input current noise and the thermal noise of the resistors is negligible in this example. The corresponding signal-to-noise ratio (neglecting distortion) for a 2V peak-to-peak sinewave output is 74.5dB. Under these conditions, however, the harmonic distortion of the AD9622 is approximately –75dBc at 2MHz. With no filtering, the dynamic range is thus equally limited by the noise and the distortion. If, however, the output of the op amp is filtered, the dynamic range is limited by the distortion.

OUTPUT NOISE AND DISTORTION OF AD9622 WIDEBAND VOLTAGE FEEDBACK OP AMP



- CLOSED LOOP BANDWIDTH = 230 MHz
- RMS OUTPUT NOISE $= 7 \times 10^{-9}\sqrt{1.57 \times 230 \times 10^{6}}$

  $= 133\mu V$ rms
- FOR 2V p-p OUTPUT, SNR = 74.5dB
- HARMONIC DISTORTION = -75dBc @ 2MHz

Figure 9.9

Rather than simply examining the THD produced by a single tone sinewave input, it is often useful to look at the distortion products produced by two tones. As shown in Figure 9.10, two tones will produce second and third order intermodulation products. The example shows the second and third order products produced by applying two frequencies, f1 and f2 to a nonlinear device. The second order products located at f2 + f1 and f2 – f1 are located far away from the two tones, and may be removed by filtering. The third order products located at 2f1 + f2 and 2f2 + f1 may likewise be filtered. The third order products located at 2f1 – f2 and 2f2 – f1, however, are close to the original tones, and filtering them is difficult.

SECOND AND THIRD-ORDER INTERMODULATION PRODUCTS FOR f_1= 5MHz, f_2 = 6MHz

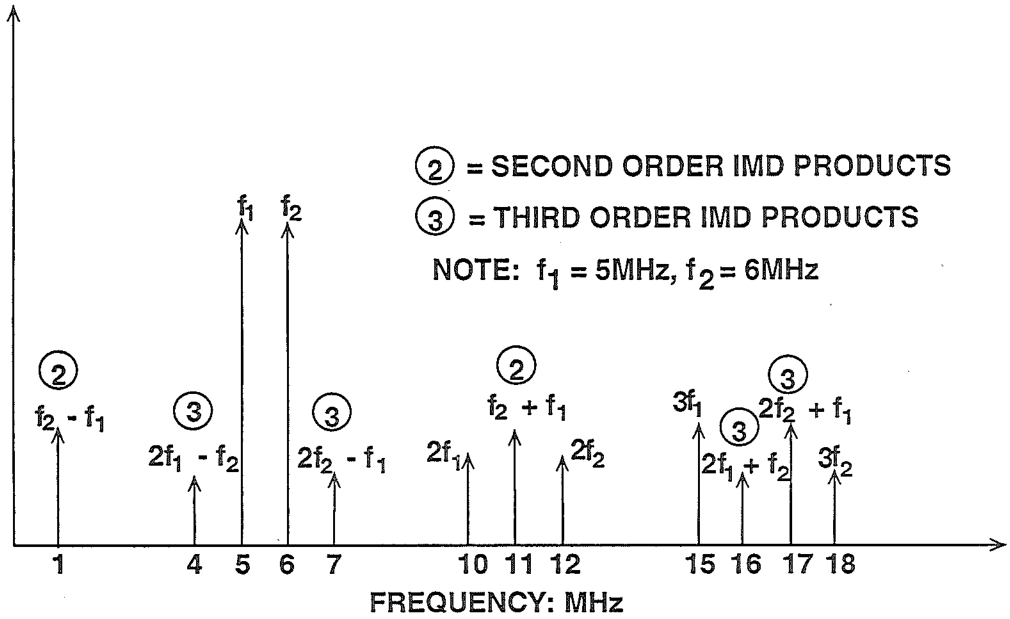


Figure 9.10

Intermodulation distortion products are of special interest in the RF area, and also a major concern in the design of radio receivers. Third-order IMD products can mask out small signals in the presence of larger ones. Third order IMD is often specified in terms of the *third order intercept* point as shown in Figure 9.11. Two spectrally pure tones are applied to the system. The output signal power in a single tone (in dBm) as well as the relative amplitude of the third-order products (referenced to a single tone) is plotted as a function of input signal power. If the system non-linearity is approximated by a power series expansion, you will find that the second-order IMD amplitudes increase 2dB for every 1dB of signal increase. Similarly, the third-order IMD amplitudes increase 3dB for every 1dB of signal increase. If you start with a low level two-tone input signal, and take a few data points, you can draw the second and third order IMD lines shown in Figure 9.11.

Once the input reaches a certain level, however, the output signal begins to soft-limit, or compress. If you extend the second and third-order intercept lines, they will intersect the extension of the output signal line. These intersections are called the *second-* and *third order intercept points*, respectively. The values are usually referenced to the output power of the device expressed in dBm. Another parameter which may be of interest is the *1dB compression point*. This is the point at which the output signal is compressed by 1dB from the ideal input/output transfer function. This point is also shown in Figure 9.11.

INTERCEPT POINTS, GAIN COMPRESSION, IMD

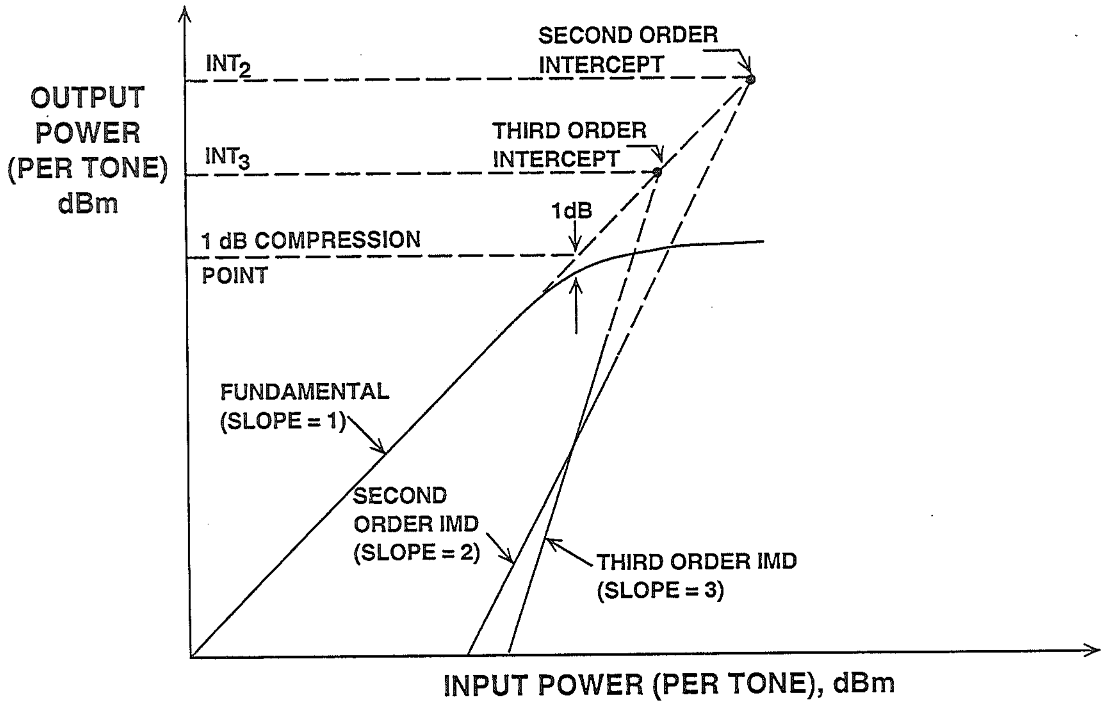


Figure 9.11

Knowing the third order intercept point allows you to calculate the approximate level of the third-order IMD products as a function of output signal level. Figure 9.12 shows the third order intercept value as a function of frequency for the AD9622 voltage feedback amplifier.

Assume the op amp output signal is 10MHz and 2V peak-to-peak into a 100Ω load (50Ω source and load termination). The voltage into the 50Ω load is therefore 1V peak-to-peak, corresponding to +4dBm. The value of the third order intercept at 5MHz is 36dBm. The difference between +36dBm and +4dBm is 32dB. This value is then multiplied by 2 to yield 64dB (the value of the third-order intermodulation products referenced to the power in a single tone). Therefore, the intermodulation products should be –64dBc (dB below carrier frequency), or at a level of –60dBm. Figure 9.13 shows the graphical analysis for this example.

AD9622 THIRD ORDER IMD INTERCEPT POINT VERSUS FREQUENCY FOR V_{out} = 1V p-p (+4dBm)

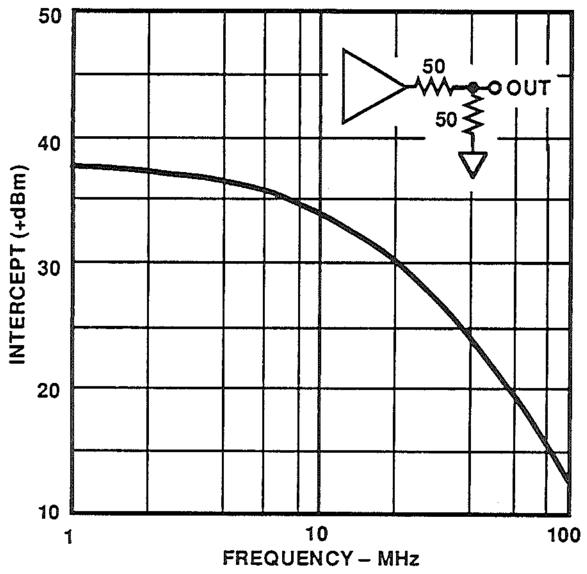


Figure 9.12

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

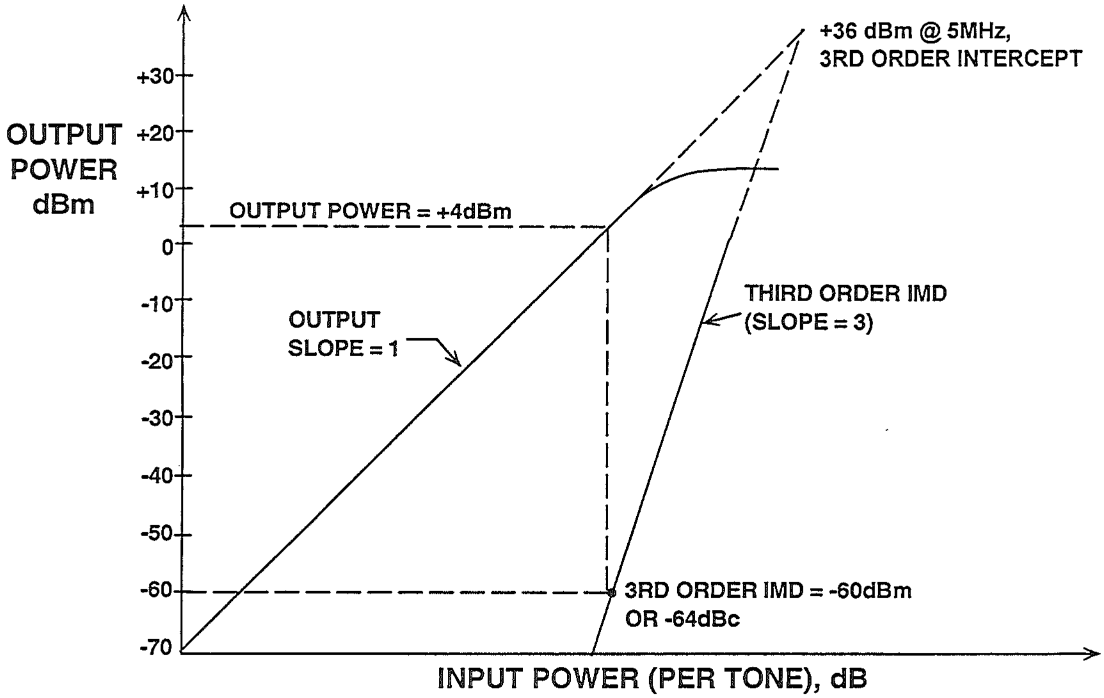


Figure 9.13

Another term often used to characterize the dynamic range of a high speed system is *spurious free dynamic range* (SFDR). The SFDR is defined simply as the difference, in dB, between the rms amplitude of a single-tone signal and the peak spurious signal within the bandwidth of interest. In an amplifier, the level of spurious signals is related to the amplitude of the original signal and is described in terms of harmonic distortion, THD, IMD, or intercept points.

DACs and ADCs, however, generally behave quite differently than amplifiers. In a sampled data system, it is therefore often difficult to examine the output spectrum and distinguish between spurs generated by the *soft* nonlinearity (low-order products) of amplifiers and the *hard* nonlinearity (low and high-order products) caused by quantization and quantization nonlinearity.

AD872 12-BIT, 10MSPS ADC EXHIBITS 73dB SPURIOUS FREE DYNAMIC RANGE FOR 1MHz, –0.5dBFS INPUT

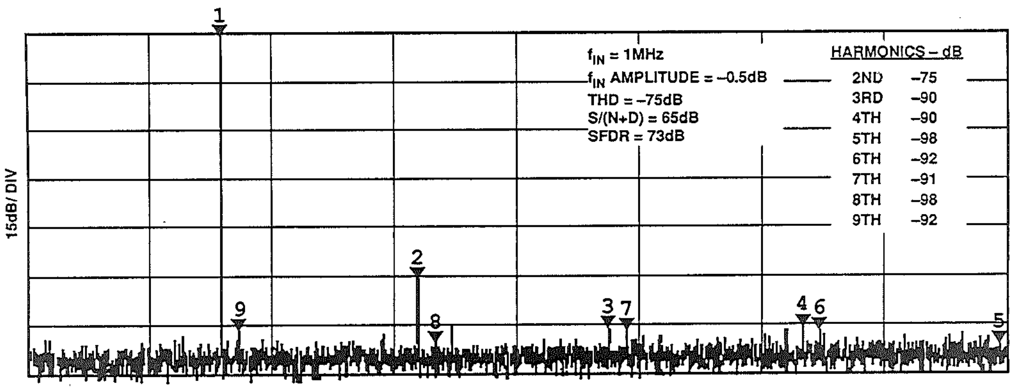


Figure 9.14

Because of this unpredictable nature of the spurs in a sampled data system, the second- and third-order intercepts generally cannot be used to accurately predict IMD performance at various signal levels. If the spurs due to quantization errors were entirely independent of signal amplitude, the point of maximum SFDR would occur for a fullscale input signal. Figure 9.15 shows a typical plot of the maximum spur level versus input signal. Notice that as the signal approaches fullscale, the level of the spurs begins to rise with the signal. This characteristic is typical of many high speed ADCs, especially for high frequency inputs. The point of maximum SFDR occurs at an input signal amplitude which is a few dB less than fullscale. In some systems, the gain is set such that the nominal signal amplitude into the ADC is equal to this value, thereby ensuring that the ADC provides maximum dynamic range possible.

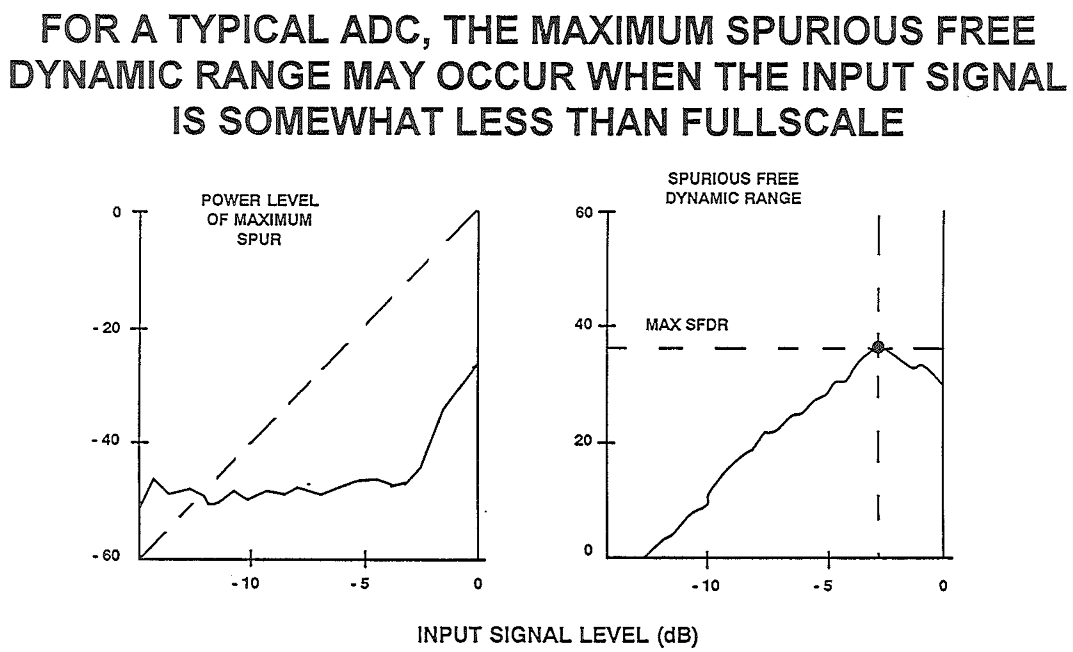


Figure 9.15

In summary, dynamic range in high speed systems is a function of all elements in the signal path and may be limited by distortion, noise, or both. Traditional methods of describing distortion and noise in amplifiers are not always applicable in sampled data systems containing ADCs and DACs. Achieving maximum dynamic range therefore requires a thorough understanding of how each system element contributes to the overall system performance. In the following sections, we will examine each in detail.

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2. Walt Kester, *Confused About Amplifier Distortion Specs?*, **Analog Dialogue**, 27-1, 1993, pp. 27-29.

3. **High Speed Design Seminar**, 1990, Norwood, MA: Analog Devices, Inc., Section 5, pp. 13-19.