Analog-to-Digital Converters and Their Applications in Radio Receivers

Rapid advances in hardware development of analog-to-digital converters (ADCs) have paved the way for development of radio receivers using digitization at the IF, and in some cases, at the RF. The constraints placed on these receivers due to hardware limitations of these devices are discussed and some examples of high-speed, state-of-the-art ADCs are given.

Jeffery A. Wepman

As advances in technology provide increasingly faster and less expensive digital hardware, more of the traditional analog functions of a radio receiver will be replaced with software or digital hardware. The ultimate goal in radio receiver design is to directly digitize the RF signal at the output of the receive antenna and hence implement all receiver functions in either digital hardware or software. Trends in receiver design have evolved toward this goal by incorporating digitization closer and closer to the receive antenna for systems at increasingly higher frequencies and wider bandwidths. Applications for these receivers are expected to increase rapidly in areas such as mobile cellular, satellite, and personal communications services (PCS) systems.

The analog-to-digital converter (ADC) is a key component in these radio receivers. ADCs most often used for wideband digitization at the RF or IF are organized as shown in Fig. 1. Key parameters of ADCs are effected by specific ADC circuit elements. For example, accuracy and linearity are primarily determined by the sample-and-hold circuitry, while jitter in the sampling clock can introduce noise in the desired output of the ADC. The quantizer establishes the resolution of the ADC. For a given burst rate, the buffers limit the sustainable throughput.

This article discusses some of the key parameters for ADCs used in radio receiver applications. The requirements, practical limitations, and potential problems for ADCs are also discussed. Conversion methods of practical ADCs such as flash, successive approximation, sigma-delta, bandpass sigma-delta, and subranging are not presented here. A discussion of these techniques can be found in [1].

Certain commercial equipment, components, instruments, or materials are identified in this article to provide some examples of current technology. In no case does such identification imply recommendation or endorsement by the National Telecommunications and Information Administration, nor does it imply that the material or equipment identified is necessarily the best available for the purpose. Furthermore, examples of technology identified in this article are not intended to be all inclusive; they represent only a sampling of what is available.

JEFFERY A. WEPMAN is an electronics engineer with the Institute for Telecommunication Sciences (ITS), National Telecommunications and Information Administration (NTIA), U.S. Department of Commerce.

Sampling Methods and Analog Filtering

The sampling process is of critical importance in radio receivers using digitization at the RF or IF. The content of the resulting sampled signal waveform is highly dependent on the relationship between the sampling rate employed and the minimum and maximum frequency components of the analog input signal. Some common sampling techniques that utilize a uniform spacing between the samples include Nyquist sampling, oversampling, quadrature sampling, and bandpass sampling (also called downsampling or direct downconversion). Sampling techniques with nonuniform spacing between the samples do exist, but they are not widely used and therefore are not considered in this article.

Any time a continuous-time analog signal is uniformly sampled, the spectrum of the original signal $F(f)$ is repeated at integer multiples of the sampling frequency (i.e., $F(f)$ becomes periodic). This is an inherent effect of sampling and cannot be avoided. This phenomenon is shown graphically in Fig. 2. Figure 2a shows the spectrum of the original analog signal $F(f)$. Figure 2b shows the spectrum of the sampled signal $F_s(f)$ using a sampling rate of $f_s = 2f_{max}$.

Nyquist Sampling

The general sampling theorem for sampling a bandlimited analog signal (a signal having no frequency components above a certain frequency f_{max}) requires that the sampling rate be at least two times the highest frequency component of the analog signal $2f_{max}$. This ensures that the original signal can be reconstructed exactly from the samples. A sampling rate of two times the highest frequency component of the analog signal is

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called the Nyquist sampling rate. Figure 2b shows an example of sampling a bandlimited signal with a maximum frequency of f_{max} at the Nyquist rate ($f_s = 2f_{max}$). Note that the copies of the spectrum of the analog signal $F(f)$ that are present in the spectrum of the sampled signal $F_s(f)$ do not overlap. As the sampling rate is increased beyond the Nyquist rate, the copies of the spectrum of the analog signal $F(f)$ that are present in the spectrum of the sampled signal $F_s(f)$ are spread even farther apart. This is shown in Fig. 2c. Sampling a bandlimited signal at rates equal to or greater than the Nyquist rate guarantees that spectrum overlap (often called aliasing) does not occur and that the original analog signal can be reconstructed exactly [2, 3].

Out-of-Band Energy

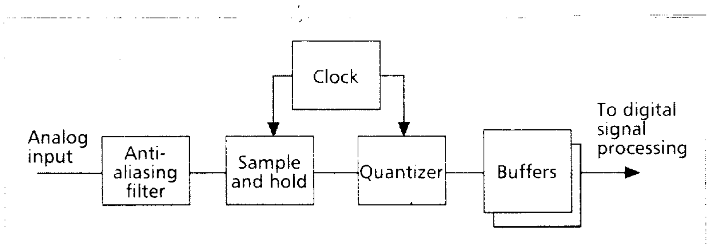
Two practical problems arise when sampling at the Nyquist rate: defining what a bandlimited signal truly is in a practical sense and analog filtering before the ADC stage. A theoretically defined bandlimited signal is a signal with no frequency components above a certain frequency. When considering real signals such as an RF signal at the input of a radio receiver, however, signals of all frequencies are always present. It is a matter of the amplitude of these frequencies that is important. In particular, the relative amplitude of the undesired signals to the desired signal is important. When digitizing an RF or IF signal at the Nyquist rate in a radio receiver, undesired signals (above one-half the sampling rate) of a sufficient amplitude can create spectrum overlap and distort the desired signal. This phenomenon is illustrated in Fig. 3. Figure 3a shows the spectrum of the analog input signal with its desired and undesired components. If this signal is sampled at two times the highest frequency in the desired signal f_d, the resulting spectrum of the sampled signal $F_s(f)$ is shown in Fig. 3b. Note that spectrum overlap has occurred here (i.e., the spectrum of the undesired signal occurs within the spectrum of the desired signal). This causes distortion in the reconstructed desired signal.

This effect raises an important question: What is the relative amplitude of the signals occurring above one-half of the sampling rate at which distortion of the desired signal due to spectrum overlap begins to predominate the distortion due to ADC nonlinearities? Nonlinearities in the ADC cause spurious responses in the ADC output spectrum. Distortion due to spectrum overlap can be said to predominate distortion due to ADC nonlinearities when the undesired signals appearing in the Nyquist band (DC to one-half the sampling rate) due to spectrum overlap exceed the largest spurious response of the ADC due to nonlinearities. Therefore, undesired signals appearing in the Nyquist band due to spectrum overlap must be lower in power than the largest spurious response of the ADC. In other words, distortion of the desired signal is predominated by ADC nonlinearities (and not spectrum overlap) if signals higher in frequency than $f_s/2$ are lower in power than the largest spurious response of the ADC. This can be quite a stringent requirement. Depending upon the specific type of radio system, it is possible that this requirement can be relaxed.

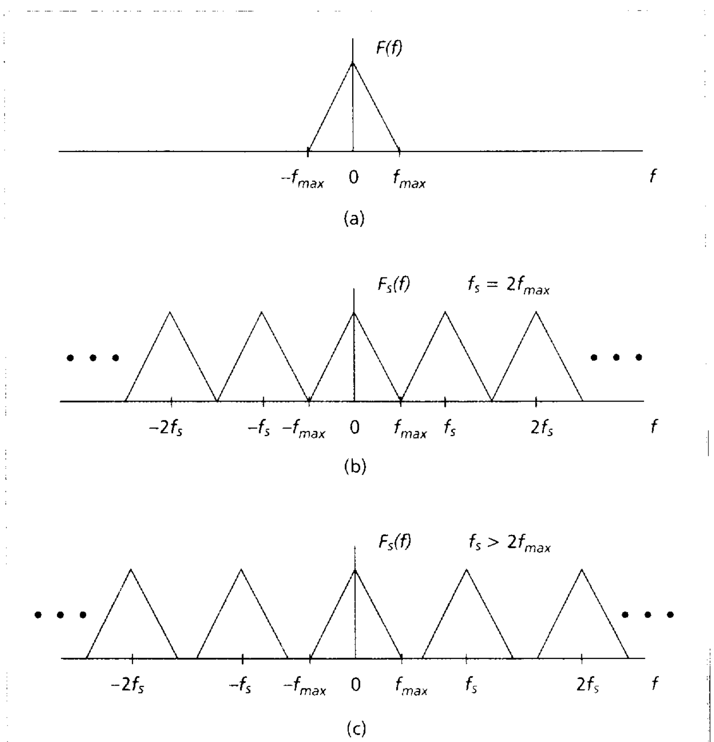
To determine ways to "relax" this requirement, the following questions should be asked: How much distortion of the desired signal is tolerable? Does the bandwidth and frequency content of both the desired signal in the Nyquist band and the undesired signals above the Nyquist band effect the distortion of the desired signal? These questions are best answered by considering the details of the specific radio communication system, such as the type of source information (voice, data, video, etc.), desired signal bandwidth, modulation and coding techniques, undesired signal characteristics (bandwidth, power, and type of signal), and finally, the performance criterion used to evaluate the quality of the reception of the desired signal.

Realizable Anti-Aliasing Filters

Analog filtering before the ADC stage is intimately related to the definition of bandlimiting. Where the definition of bandlimiting deals with the content of the signals that may be present, analog filtering before the ADC represents a signal processing stage



■ **Figure 1.** *Elements of ADCs used for wideband digitization at the RF or IF.*



■ **Figure 2.** *Spectrum of: a) a bandlimited continuous-time analog signal, b) the signal sampled at $f_s = 2f_{max}$, and c) the signal sampled at $f_s > 2f_{max}$.*

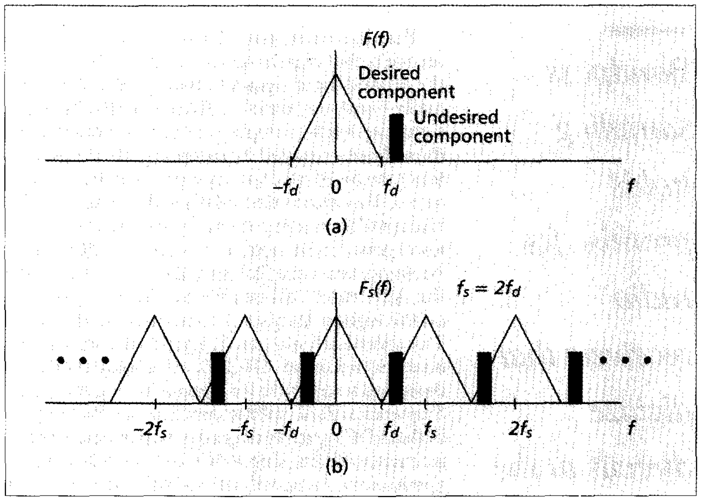
where certain frequencies can be attenuated. It is important to know both the signals that can be present before filtering and the amount of attenuation that the filter offers for different frequencies. With knowledge of both of these, the true spectrum of the signal to be digitized can be determined. Sampling at the Nyquist rate presents a large and often impractical demand on the filter used before digitization (anti-aliasing filter). Ideally, an anti-aliasing filter placed before an ADC would pass all of the desired frequencies up to some cutoff frequency and provide infinite attenuation for frequencies above the cutoff frequency. Then sampling at the Nyquist rate would be two times the cutoff frequency and no spectrum overlap would occur. Unfortunately, practically realizable filters cannot provide this type of "brickwall" response. The attenuation of real filters increases more gradually from the cutoff frequency to the stop band. Therefore, for a given cutoff frequency on a real filter, sampling at two times this cutoff frequency will produce some spectrum overlap. The steeper the transition from the passband to the stop band and the more attenuation in the stop band, the less the sampled signal will be distorted by spectrum overlap. In general, more complicated filters are required to achieve steeper transitions and higher attenuation in the stop band. Therefore, more complicated filters are required to reduce the distortion in the sampled signal due to spectrum overlap for a given sampling rate. Limitations on the practical implementation of analog filters make high-order, steep rolloff filters difficult to realize. Also, as the steepness of the rolloff is increased, the phase response tends to become more nonlinear. This can create distortion of the desired receive signal since different frequencies within a signal can be delayed in time by different amounts (i.e., phase shifted by amounts not proportional to frequency).

Oversampling Eases Requirements on the Anti-Aliasing Filter

Sampling at rates greater than the Nyquist sampling rate is called oversampling. One of the benefits of oversampling is that the copies of the spectrum of the analog signal $F(f)$ that are present in the spectrum of the sampled signal $F_s(f)$ become increasingly separated as the sampling rate is increased beyond the Nyquist rate. For an analog signal with a given frequency content and a given anti-aliasing filter with a cutoff frequency of f_c, sampling at the Nyquist rate (two times the cutoff frequency) produces a certain amount of distortion due to spectrum overlap. When sampling at a higher rate, a simpler anti-aliasing filter with a more gradual transition from passband to stop band and less stop band attenuation can be used without any increase in the distortion due to spectrum overlap. Therefore, oversampling can minimize the requirements of the anti-aliasing filter. The tradeoff, of course, is that increasingly faster ADCs are required to digitize relatively low frequency signals.

Quadrature Sampling Reduces Required Sampling Rate

In quadrature sampling the signal to be digitized is split into two signals. One of these signals is multiplied by a sinusoid to downconvert it to a zero center frequency and form the in-phase component of the original signal. The other signal is multiplied by a 90-degree phase-shifted sinusoid to downconvert it to a zero center frequency and form the quadrature-phase component of the original signal. Each of these components occupies only one-half of the bandwidth of the original signal and can be sampled at one-half the sampling rate required for the original signal. Therefore, quadrature sampling reduces the required sampling rate by a factor of two at the expense of using two phase-locked ADCs instead of one.



■ **Figure 3.** *Spectrum of: a) a continuous-time analog signal with a desired and undesired component and b) the signal sampled at $f_s = 2f_d$.*

Bandpass Sampling for Direct Downconversion

Sampling at rates lower than $2f_{max}$ can still allow for an exact reconstruction of the information content of the analog signal if the signal is a bandpass signal. An ideal bandpass signal has no frequency components below a certain frequency f_l and above a certain frequency f_h. For a bandpass signal, the minimum requirements on the sampling rate to allow for exact reconstruction are that the sampling rate be at least two times the *bandwidth* $f_h - f_l$ of the signal. To ensure that spectrum overlap does not occur, when sampling rates are between two times the bandwidth of the bandpass signal and two times the highest frequency in the bandpass signal, the sampling frequency f_s must satisfy

$$\frac{2f_h}{k} \le f_s \le \frac{2f_l}{(k-1)}$$

where k is restricted to integer values that satisfy

$$2 \le k \le \frac{f_h}{(f_h - f_l)}$$

and $(f_h - f_l) \le f_l$ [3].

Bandpass sampling can be used to downconvert a signal from a bandpass signal at an RF or IF to a bandpass signal at a lower IF. Since the bandpass signal is repeated at integer multiples of the sampling frequency, selecting the appropriate spectral replica of the original bandpass signal provides the downconversion function.

B*andpass sampling holds promise for radio receivers that digitize directly at the RF or IF, since the desired input signals to radio receivers are normally bandpass*

Bandpass sampling holds promise for radio receivers that digitize directly at the RF or IF, since the desired input signals to radio receivers are normally bandpass signals. Theoretically, bandpass sampling allows sampling rates to be much lower than those required by sampling at two or more times the highest frequency content of the bandpass signal. This means that ADCs with slower sampling rates (and therefore potentially higher performance, lower power consumption, or lower cost) may be used. An important practical limitation, however, is that the ADC must still be able to effectively operate on the highest frequency component in the signal. This specification is usually given as the analog input bandwidth for the ADC. Conventional ADCs are designed to operate on signals having maximum frequencies of one-half the sampling rate. Performance of the ADC typically degrades with increasing input frequency. When using ADCs for bandpass sampling applications, the specifications of the converter must be examined to determine the behavior at higher frequency inputs. In addition, when bandpass sampling, stringent requirements on analog bandpass filters (steep rolloffs) are needed to prevent distortion of the desired signal from strong adjacent channel signals.

Effects of Quantization Noise, Distortion, and Receiver Noise

This section addresses the relationships among quantization noise, harmonic distortion, and receiver noise. The ADCs best suited to RF and IF processing that have widespread availability use uniform quantization. In uniform quantization, the voltage difference between each quantization level is the same. Other methods of quantization include logarithmic (A-law and μ-law), adaptive, and differential quantization. These methods are currently used in source coding. A discussion of these quantization techniques may be found in [1].

In uniform quantization, the analog signal cannot be represented exactly with only a finite number of discrete amplitude levels. Therefore, some error is introduced into the quantized signal. The error signal is the difference between the analog signal and the quantized signal. Statistically, the error signal is assumed to be uniformly distributed within a quantization level. Using this assumption, the mean squared quantization noise power P_{qn} is

$$P_{qn} = \frac{q^2}{12R}$$

where q is the quantization step size and R is the input resistance of the ADC [4]. In an ideal ADC, this representation of the quantization noise power is accurate to within a dB for input signals that are not correlated with the sampling clock.

If, on the other hand, the analog input into an ADC is periodic, the error signal is also periodic. This periodic error signal includes harmonics of the analog input signal and results in harmonic distortion. Furthermore, harmonics that fall above the Nyquist frequency appear in the Nyquist band due to aliasing. Dithering is commonly used to reduce this harmonic distortion. In one implementation of this technique, thermal noise is added at the ADC input to provide a relatively flat noise power spectrum over the Nyquist bandwidth; the thermal noise plus the quantization noise combine so that the quantization error is uniform. A simple way to implement this technique is to provide amplifier gain to boost the receiver noise to, or several dB above, the level of the quantization noise. The implications of this are examined next.

Commercially available ADCs typically have a full scale range (FSR) of 1 to 20 V. The FSR of the ADC is the difference between the maximum and the minimum analog input voltages to the ADC. Dividing the FSR by the number of quantization levels 2^B, where B is the number of bits of the ADC, provides the quantization step size q. For an 8 b ADC with a FSR of 2.5 V, the quantization step size is 9.77 mV. To compute the quantization noise power, the effective input resistance of the ADC must be known.

Components in radio receivers typically have a 50-ohm input and output impedance. The input impedance of ADCs is usually higher than this and is not well specified. Therefore, when interfacing an RF component with an ADC, as is necessary for digitization at the RF or IF, this impedance mismatch must be considered. A simple method of impedance matching is to place a 50-ohm resistive load at the input of the ADC. This forces the effective input resistance of the ADC to be close to 50 ohms. The quantization noise power can then be computed. Assuming a 50-ohm effective input resistance R to the ADC in this example, the quantization noise power equals −38 dBm. For a noise-limited receiver, the receiver noise power P_{rn} can be computed as the thermal noise power in the given receiver bandwidth (BW) plus the receiver noise figure (NF). This is given as

$$P_{rn} = -174dBm + 10\log_{10}\mathrm{BW}(Hz) + \mathrm{NF}\,(dB).$$

For a receiver with a 10-MHz BW and a 6-dB NF, the receiver noise power is −98 dBm. Therefore, to boost the receiver noise to the quantization noise power level requires a gain of 60 dB. For an ADC of higher resolution, less gain would be needed, since the quantization noise power would be smaller. Also, wider receiver bandwidths and higher receiver noise figures would require less gain since the receiver noise power would be larger. Nevertheless, for most practical receiver and ADC combinations, automatic gain control is necessary to assure that the least significant bit (LSB) represents a uniform noise input while the peak power does not exceed the ADC's FSR.

Important Specifications

In this section, theoretical signal-to-noise ratio (SNR) due to quantization noise and aperture jitter is discussed. Practical specifications for real ADCs are then presented.

Theoretical Signal-to-Noise Ratio Specifications

For radio receiver applications where the amplitude of the desired signal falls within the ADC's FSR, and the bandwidth of the desired signal is equal to the Nyquist bandwidth, the SNR of an ADC is a useful specification. The theoretical SNR of ADCs is generally thought of as $6B$ (dB), where B

is the number of bits of resolution of the ADC. A more precise expression providing the maximum possible theoretical SNR can be derived based on some assumptions about the noise and the input signal. First, it is assumed that the noise present is due to quantization error only. The amplitude of this quantization noise is assumed to be a random variable uniformly distributed over one quantization step. Assuming a sinusoidal input with an amplitude equal to the FSR of the ADC, the maximum possible theoretical SNR is given as

$$\mathrm{SNR} = 6.02B + 1.76 + 10\log_{10}\left(\frac{f_s}{2f_{\max}}\right) \quad (dB)$$

where f_s is the sampling frequency and $f_{\max}$ is the maximum frequency of the input analog signal [5, 6]. The commonly stated theoretical SNR of $6B$ (dB) is an approximation to this equation when $f_s = 2f_{\max}$ and the 1.76 dB is neglected. From this equation, note that as the sampling frequency is increased beyond the Nyquist rate of $2f_{\max}$, the SNR increases. This occurs because the quantization noise power, which is fixed and independent of bandwidth, is spread out over an increasingly wider band as the sampling frequency is increased. This lessens the amount of the quantization noise that falls within the Nyquist band (DC to $f_{\max}$). Consequently, this means that oversampling increases the maximum possible SNR. Such oversampling is sometimes used to realize a greater maximum SNR than at first appears possible. An 8 b ADC, with a sampling rate of 20 Msamples/s, for example, can provide 68 dB rather than 48 dB of maximum SNR for 100 kHz signals in the passband if appropriate digital filtering is used to recover the 100 kHz signal.

Besides being limited by the quantization step size (resolution), the SNR of the ADC is also limited by aperture jitter. Aperture jitter is the variation in time of the exact sampling instant. Aperture jitter can be caused externally by jitter in the sampling clock, or internally since the sampling switch does not open at precise times. Aperture jitter causes a phase modulation of the sampled signal and thus results in an additional noise component in the sampled signal [7]. The maximum analog input frequency of the ADC is limited by this aperture jitter since the SNR due to aperture jitter (SNR_{aj}) degrades as the input frequency increases. The SNR_{aj} is given as

$$\mathrm{SNR}_{aj} = 20\log_{10}\left(\frac{1}{2\pi f_{\max}\, t_a}\right)$$

where t_a is the aperture jitter of the ADC [5]. For sampling at the Nyquist rate, where $f_s = 2f_{\max}$, both the SNR due to quantization noise and the SNR due to aperture jitter can be combined to give the overall SNR [8]. Distortion of the ADC output signal can also be introduced because each sample is taken over a finite time period in real sample and hold circuits. This distortion occurs as the input signal varies over this time period.

Practical Specifications for Real ADCs

The SNR in a real ADC can be determined by measuring the residual error. Residual error is the combination of quantization noise, random noise, and nonlinear distortion (i.e., all of the undesired components of the output signal from the ADC). The residual error for an ADC is found by using a sinusoidal input into the ADC. An estimate of the input signal is subtracted from the output of the ADC and the remaining signal is the residual error. The mean squared power of the residual error is then computed. The SNR is then found by dividing the mean squared (MS) power of the input signal by the mean squared power of the residual error.[1]

A specification sometimes used for real ADCs instead of the SNR is the effective number of bits. This specification is defined as *the number of bits required in an ideal ADC so that the mean squared noise power in the ideal ADC equals the mean squared power of the residual error in the real ADC.*

One definition of the spurious free dynamic range (SFDR) assumes a sinusoidal input to the ADC. In this case, the SFDR is the ratio of the sinusoidal signal power to the peak power of the largest spurious signal in the ADC output spectrum. SFDR allows one to assess how well an ADC can simultaneously detect a very small signal in the presence of a very large signal. Hence, it is an important specification for ADCs used in radio receiver applications. A common misconception is that the SFDR of the ADC is equivalent to the SNR of the ADC. In fact, there is typically a large difference between the SFDR and the SNR of an ADC. The SNR is the ratio between the signal power and the power of the residual error. The SFDR, however, is the ratio between the signal power and the peak power of only the largest spurious product. Since the power of the residual error includes quantization noise, random noise, and nonlinear distortion within the entire Nyquist band, the power of the residual error can be much higher than the peak power of the largest spurious product. Hence, the SFDR can be much larger than the SNR [4].

The SFDR specification is useful for applications when the desired signal bandwidth is smaller than the Nyquist bandwidth. In this case, a wide band of frequencies is digitized and results in a

Specification	Application	Definition
Signal-to-noise ratio (SNR)	Desired signal BW equal to Nyquist BW	MS signal power / MS power of residual error
Spurious free dynamic range (SFDR)	Desired signal BW less than Nyquist BW	MS signal Power / MS power of the largest spurious product
Noise power ratio (NPR)	Desired signal spectrum contains many narrowband channels	Power spectral density of noise outside frequency band of notch filter[1] / Power spectral density of noise inside frequency band of notch filter
Full-power analog input BW	Bandpass sampling	Range from frequency where output amplitude falls to 3 dB less than maximum[2]

[1] With an input signal having a bandlimited, flat noise spectrum and a narrow band of frequencies removed by a notch filter.

[2] For a full-scale input signal.

Table 1. *Summary of ADC specifications for radio receiver applications.*

[1] *The SNR is often (and more accurately) called the Signal-to-Noise Plus Distortion Ratio when distortion is included with the noise, as in this case.*

Resolution (number of bits)	Sampling rate (Msamples/s)	Manufacturer
6	4000	Rockwell International
8	750	Signal Processing Technology
8	2000	Hewlett-Packard
8	3000	*
10	70	Pentek
12	50	Hughes Aircraft
12	100	*
14	24	Hughes Aircraft
18	10	Hewlett-Packard

* Device in development. Work is being sponsored by the Advanced Research Projects Agency (ARPA) of the U.S. Department of Defense.

■ **Table 2.** *Examples of current high-speed ADC technology.*

given SNR. The desired signal is then obtained by using a narrowband digital bandpass filter on this entire band of frequencies. The SNR is improved by this digital filtering process since the power of the residual error is decreased by filtering. The SFDR specification for the ADC becomes very important since a spurious component may still fall within the bandwidth of the digital filter, and hence the SFDR, unlike the SNR, does not necessarily improve by the digital filtering process. However, several techniques are available to improve the SFDR. Dithering (discussed previously) improves the SFDR of ADCs. Additionally, post-digitization processing techniques such as phase-plane compensation [9], state variable compensation [10], and projection filtering [11] have been used to improve SFDR.

For an ideal ADC, the maximum SFDR occurs at a full-scale input level. In practical ADCs, the maximum SFDR occurs at input levels at least several dB below the full-scale input level. This occurs because as the input levels approach full-scale, the response of the ADC becomes more nonlinear and more distortion is exhibited. Additionally, due to random fluctuations in the amplitude of real input signals, as the input signal level approaches the FSR of the ADC, the probability of the signal amplitude exceeding the FSR increases. This causes additional distortion from clipping. Therefore, it is extremely important to avoid input signal levels that closely approach the full-scale level in ADCs. Prediction of the SFDR for practical ADCs is difficult, therefore measurements are usually required to characterize the SFDR of the ADC.

In the preceding discussion on SFDR, a sinusoidal ADC input signal was assumed. However, intermodulation distortion (IMD) due to multi-tone inputs is important in ADCs used for wideband radio receiver applications. To characterize this IMD due to multi-tone inputs, another definition of the SFDR could be used. In this case, the SFDR is the ratio of the combined signal power of all of the multi-tone inputs to the peak power of the largest spurious signal in the ADC output spectrum. A current example of test equipment to generate multi-tone inputs produces up to 48 tones.

The noise power ratio (NPR) specification is useful in applications such as mobile cellular radio, where the spectrum of a signal to be digitized consists of many narrowband channels and where adjacent channel interference can degrade system performance. Particularly, the NPR provides information on how well an ADC can limit crosstalk between channels [9].

The NPR is measured by using a noise input signal into the ADC. This noise signal has a flat spectrum that is bandlimited to a frequency that is less than one-half the sampling frequency. Additionally, a narrow band of frequencies is removed from the noise signal using a notch filter. This noise spectrum is used as the input signal to the ADC. The frequency spectrum of the output of the ADC is then determined. The NPR is then computed by dividing the power spectral density of the noise outside the frequency band of the notch filter by the power spectral density of the noise inside the frequency band of the notch filter [4].

When using an ADC in a bandpass sampling application where the maximum input frequency into the ADC is actually higher than one-half the sampling frequency, the full-power analog input bandwidth is an important specification. A common definition of full-power analog input bandwidth (although not universal) is the range from DC to the frequency where the amplitude of the output of the ADC falls to 3 dB below the maximum output level. This assumes a full-scale input signal to the ADC. Typically, the ADC is operated at input frequencies below this bandwidth. Aside from full-power analog input bandwidth, it is important to examine the behavior of the other specifications such as SNR, SFDR, and NPR at the desired operating frequencies since these specifications may vary with frequency. Table 1 provides a summary of the important ADC specifications for radio receiver applications.

ADC Applications Tradeoffs

The performance of ADCs continues to improve at a rapid rate. For radio receiver applications using digitization at the RF or IF, ADCs with both high sampling rates and high performance are desired. Unfortunately, there is a tradeoff between these two requirements. As a general trend, although not always true, the higher the performance of the ADC, the lower its maximum sampling rate will be. Table 2 shows some examples of current high-speed ADC technology for various ADC resolutions.

When selecting an ADC for a specific radio receiver application, in addition to the sampling rate, one must consider critical specifications that characterize the ADC performance such as the SNR, SFDR, and NPR. The ADC specifications most important for various applications are listed in Table 3. In certain applications such as channelized PCS, the Universal Mobile Telecommunication System (UMTS), the Future Public Land Mobile Telecommunication System (FPLMTS), and mobile cellular systems, instead of digitizing the entire band with a single high-speed ADC, parallel ADCs used to digitize narrower bandwidths are often practical ADC architectures. In this case, ADCs with better performance can be used since the demands of a high sampling rate are relieved.

Summary

This article provides an introduction to some of the important factors that must be considered when using ADCs in radio receiver applications. These factors include the choice of sampling method, the amount and effects of out-of-band energy, the analog filtering required, the effects of quantization noise, receiver noise, and distortion, and the critical ADC specifications for radio receiver applications. The differences between Nyquist sampling, oversampling, and bandpass sampling were discussed. It was shown that sampling at the Nyquist rate presents a large and often impractical demand on the anti-aliasing filter. Oversampling eases the requirements on the anti-aliasing filter. A very steep rolloff bandpass filter is required for bandpass sampling when there are strong signals present in adjacent channels. Quadrature sampling was shown to reduce the required sampling rate by a factor of two at the expense of using two phase-locked ADCs instead of one. It was also shown that, in general, some sort of gain control is required for proper digitization of signals at the RF or IF in a receiver. ADC specifications of particular importance to radio receiver applications such as SNR, SFDR, full-power bandwidth, and NPR were examined and some examples of high-speed, state-of-the-art ADCs were given. As ADC performance continues to improve, digitization at the RF and IF in radio receivers, at increasingly higher frequencies, will be used in an increasingly broad range of applications.

Acknowledgments

The author would like to thank Daniel Moulin of The MITRE Corporation for providing the table on critical ADC specifications and performance issues for typical applications.

Typical applications	Critical ADC specifications	Performance issues
Spread spectrum	SNR SFDR NPR	SNR for quantization of small signals in an environment with strong interference SFDR for spatial filtering NPR for interchannel crosstalk
Wideband digital receivers	SFDR	SFDR for accurate detection of low-level signals in an environment with strong interference
Radar	SNR SFDR Overvoltage recovery	SNR for clutter cancellation SFDR for Doppler processing
Cellular mobile and PCS	SNR SFDR NPR	SNR and SFDR for wide bandwidth channelized receivers NPR for interchannel crosstalk
Spectrum analysis	SNR SFDR	SNR and SFDR for high fidelity measurements
Digital sampling oscilloscopes	SNR DNL*	SNR for better amplitude resolution DNL for accurate representation of waveform

* Differential nonlinearity (DNL) is the maximum amount of deviation of any quantization step in the ADC from the theoretical quantization step size of $FSR/2^B$.

■ **Table 3.** *Critical ADC specifications and performance issues for typical applications.*

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Biography

JEFFERY A. WEPMAN received B.S. and M.S. degrees in electrical engineering from the University of Arizona in 1981 and 1985, respectively. He has completed all requirements except the dissertation for his Ph.D. in electrical engineering from the University of Colorado at Boulder. Since 1986, he has worked as an electronics engineer with the Institute for Telecommunication Sciences (ITS), National Telecommunications and Information Administration (NTIA), U.S. Department of Commerce, Boulder, Colorado. During this time, he worked on a wide variety of telecommunication projects, including propagation measurement system design, development, analysis, and implementation. From 1991 through 1994, he was a project leader for the Personal Communication Services (PCS) outdoor propagation measurements at ITS. He helped develop the recently patented *Digital Sampling Channel Probe*, ideal for making outdoor impulse response measurements to characterize wideband propagation in the radio channel. He was the recipient of two Department of Commerce Bronze Medal Awards for his work in advanced measurement system development and PCS measurements. His current work has included research on radio receivers using digitization at the RF or IF.