Wide-dynamic-range A/D converters pave the way for wideband digital-radio receivers

Brad Brannon, Analog Devices Inc

Digital radios are becoming more commonplace because of recent advances in data conversion, mixers, and DSPs. But what are digital radios, how do they work, and how do they differ from analog radios?

Digital radios offer many technical advantages over their analog predecessors, allowing flexible new designs and attractive features. Not only can digital radios detect digital modulation, they can also detect information applied to carriers by analog techniques, such as AM and FM. Furthermore, because the core of a digital radio is its DSP, one receiver can simultaneously detect analog and digital modulation if the hardware preceding the DSP is properly designed. The key to digital radios is the software that runs the DSP.

Software-programmable digital radios, in which software determines the receiver's characteristics, also offer many economic advantages over analog radios. A radio manufacturer can design a generic radio in hardware. As modulation-system standards change, for example, from FM to code- or time-division multiple access (CDMA or TDMA), manufacturers can quickly adapt their radios merely by reprogramming the DSP. By loading new software, users or service providers can upgrade software radios at little cost and without sacrificing investments in hardware. By changing only the software, manufacturers can also inexpensively tailor radios for specialized applications—something digital receivers can offer that analog receivers cannot.

What is a digital receiver?

Digital and analog receivers are basically the same except that in a digital receiver, some digital functions replace their analog equivalents. Figures 1 and 2
show examples of each type of radio. As these figures illustrate, the main difference is that a digital radio uses an A/D converter and a DSP, instead of a discriminator. Although this example is very simple, it shows the beginnings of what digital radios require. An added benefit of a digital radio is that it performs some of the filtering digitally, eliminating frequency-sensitive components, such as closely matched and tight-tolerance inductors and capacitors. In addition, because filtering takes place within the DSP, software (not costly, sensitive surface-acoustic-wave (SAW), ceramic, or crystal devices) determines the filter characteristics—bandwidth, passband ripple, and stopband rejection. In fact, digital techniques can implement filter characteristics that analog filters simply cannot match. The simple example in Figure 3 is only the beginning. Although existing technology allows digital implementation of much of the receiver, moving the digital portion of the radio as close to the antenna as possible offers so many advantages, that some people advocate direct RF sampling using an A/D converter located directly on the antenna. As attractive as this approach seems, it has some serious drawbacks, most notably receiver sensitivity. Because the A/D converter would be exposed to numerous signals, some larger than the desired signal and others smaller, the A/D converter would need a very wide dynamic range. Second, the A/D converter would have to provide this dynamic range at RF frequencies—a very difficult task. Finally, this approach would severely limit receiver selectivity and sensitivity.

**Eliminate analog filters**

Analog receivers achieve their selectivity in the IF stages through SAW or ceramic filters. Limiting the signal bandwidth allows large signal gains and achieves high sensitivity without amplifying undesired signals, which, if amplified, would overdrive a digital receiver's A/D converter. The notion of direct RF sampling does emphasize one key advantage of software digital radios, however: They are completely programmable and require little or no component selection or tweaking to achieve required performance.

A reasonable compromise in many receivers is to convert the radio to digital at the first or second IF stage. This architecture filters the out-of-band signals before the A/D converter and it also permits AGC in the analog stage prior to the A/D converter. AGC reduces the possibility of in-band overdrive and allows for maximum signal gain prior to data conversion, thus relaxing the dynamic-range requirements on the A/D converter. Also, IF sampling and digital-receiver designs work together to reduce cost by reducing component counts (for example, by eliminating IF stages) and to improve flexibility by replacing fixed analog-filter components with programmable digital circuits.

Analog receivers place much of the signal gain after the first IF stage. This approach prevents out-of-band or strong in-band signals from overdriving the front end. However, in an IF-sampling digital receiver, the front end provides all of the gain, necessitating great care to prevent in-band and out-of-band signals from saturating the A/D converter. The RF stage must not only provide gain, it must also include a way to attenuate large in-band signals. Although digital circuits following the A/D converter can provide additional gain, certain limitations apply (see the section entitled "Process gain"). Gain in the analog domain improves the S/N ratio and reduces performance only through the introduction of noise. Figure 3 shows a more detailed diagram of an IF-sampling digital receiver. This receiver includes RF gain, attenuation options, a high-performance A/D converter, a digital-filter chip, and the DSP.

Two important terms to consider are narrowband and wideband. A digital receiver can be either. Narrowband means that prefiltering reduces all undesired signals to the point of insignificance; the A/D converter sees only the signal of interest. Wideband means that adjacent channels are present in the A/D converter input and are digitally filtered in the receiver's later stages. Usually, a
wideband receiver receives an entire band, such as that of cellular telephony or other similar wireless services (personal communication services, for example). In fact, a single wideband receiver can receive all channels within the band simultaneously, allowing all channels to share most of the hardware (Figure 4). (Although this discussion focuses on wideband implementations, much of the discussion is equally relevant to IF-sampling narrowband receivers.)

**Wideband trade-offs**

Much debate has centered on the use of wideband digital receivers. Several analyses have shown that wideband receivers are not cost effective. A quick look at a Pareto chart of device cost in wideband and narrowband digital receivers (Table 1) identifies the reasons for this conclusion. Note that because both narrowband and wideband implementations use DSPs, DSP cost is irrelevant. Several devices make older wideband implementations particularly expensive. Because wideband architectures eliminate interstage filtering, the mixer outputs contain multiple signals, which the A/D converter digitizes. Therefore, the mixer and the A/D converter must produce exceptionally low third-order modulation products to prevent the creation of new spurious-frequency components that could interfere with receiver operation.

Traditionally, high-performance mixers and A/D converters have been discrete, hybrid, or modular devices with high selling prices. However, recent semiconductor advances and improvements in system architecture achieve higher levels of integration and performance at much lower cost. New hundred-dollar A/D converters perform better than did devices that cost $1000 in 1993 and they are raising interest in wideband radios.

Additional developments in digital design also enable continued improvements in DSPs. One such development is high-density on-chip memory, which eliminates significant amounts of pc-board real estate and memory-interface problems. DSPs with as much as 4 Mbits of on-chip memory are now available. The result is that a single DSP can handle multiple voice/RF channels with no external memory or interface requirements.

Other important digital advances include accelerators, such as digital tuner/filters (Figure 5) and forward-error-correction chips. These ICs are not only key architectural elements, they help lower system cost by reducing the demands on general-purpose DSPs, much as coprocessors do in personal computers.

As Figure 4 shows, in a wideband receiver, several channels share the bulk of the hardware. A recent study of cost models shows that the cost per channel quickly falls below that of a traditional narrowband receiver (Figure 6). Depending on the assumptions, the cost of a three-RF-channel wideband radio can be the same as that of three narrowband receivers. As you add channels, the wideband digital receiver's cost advantage grows.

The time needed to align traditional narrowband receivers can make such receivers quite expensive. Digital receivers have far fewer critical analog components and, thus, have lower assembly costs. The adjustments that remain are in the RF front end and are performed only once per receiver subsystem. Traditional narrowband receivers require alignment of each RF channel. Nevertheless, wideband receivers are not always the best solution, especially in cellular systems in which traffic volume is low.
Key component developments

For wideband digital radios to flourish, many technologies must come together. First, software radios require fast, powerful DSPs, and recent DSP developments are making a major impact on software radios. In addition, software radios' DSP chips must simply and effectively implement a range of algorithms. In addition, the chips should be highly integrated and have on-chip memory to simplify system design and minimize board space. DSPs such as the ADSP-2181 and ADSP-21062 have enough on-chip memory to implement complete receiver functions, including demodulation, equalization, and decoding.

Another key component is the digital tuner, or digital drop receiver. These tuners must have configurable data rates and bandwidths and must have outstanding filter characteristics that suit multiple modulation systems. These tuner chips are nothing less than special-purpose signal-processing chips with two main purposes: to filter out all but the signal of interest and to decimate the data rate to a value that falls within the capabilities of the DSPs that follow. Tuner chips also provide process gain that digitally improves the S/N ratio of the received signal.

Digital drop receivers also offer FIR filtering, which replaces analog receivers' LC, ceramic, and SAW filters. Analog filters use passive components that have initial tolerances, aging factors, and insertion loss. Receiver designs must account for these characteristics either through very rigid component selection or tweaking. Because digital filters are perfectly reproducible without tweaking and do not change with age, they hold down product life-cycle costs.

In the RF area, wide-dynamic-range mixers and low-noise amplifiers are a must. Key specifications are third-order input-intercept points (3OIPs), which determine spurious-frequency response, and noise figure, which determines in-band receiver noise.

Selecting an A/D converter

Perhaps one of the least understood components in an IF-sampling receiver is the data converter. For starters, the A/D converter specifications are derived from the characteristics of the modulation on the signal that the A/D converter must digitize. Because it digitizes all channels simultaneously, a wideband cellular base-station receiver must have excellent dynamic range. In such a receiver, no AGC can compensate for varying signal strengths; reducing the gain for strong signals reduces the sensitivity to weaker ones, possibly causing calls to drop out. Practical radios must use some form of AGC or step attenuation to prevent receiver overload from causing all calls to drop out. With a 10- or 20-dB gain drop, only the weaker signals are lost or handed off to another cell. This loss is much better than the complete signal loss that results from an overdriven receive channel.

When receiving European digital cellular Global System for Mobile Communications (GSM) signals, a receiver must digitize signals between –104 dBm and –13 dBm (–23 dBm for the DCS-1800 standard). In other words, the receiver must digitize accurately in the presence of other signals (Figure 7) over a dynamic range of 91 dB, not counting overhead for equalization. Thus, the SFDR of the converter and analog front end must exceed 100 dB relative to full scale (dBFS). When a base-station receiver receives signals from nearby cellular phones, SFDR is very important; it indicates how strong signals interfere with adjacent channels. Figure 7 shows that front-end nonlinearity causes strong signals to produce the largest spurious signals (spurs). When a receiver must simultaneously process strong signals and weak ones that originate near the cell fringes, the strong signals cause spurs that can mask the weaker signals. SFDR is also important because it shows how
the receiver performs as the signal approaches the noise floor. SFDR indicates the overall receiver S/N ratio—BER (bit-error rate) for digital modulation—where BER is proportional to the reciprocal of the remaining S/N and distortion (SINAD) at low signal levels. Although GSM is one of the more difficult standards to realize using wideband techniques, it provides an excellent example of the importance of certain converter specifications. Other standards, such as AMPS (Advanced Mobile-Phone Service) and N-AMPS (narrowband Advanced Mobile-Phone Service), are less demanding on receiver designs and permit simple implementations using wideband techniques. As a final note, third-order intercept is not a meaningful A/D converter specification, because third-order modulation products do not behave in the same way as do other linear products. A/D converters do exhibit intermodulation distortion (IMD), however (Figure 8).

When an A/D converter's input is a single signal, full-scale SINAD and S/N ratio are usually the specifications of choice. However, when an A/D converter digitizes broad bands of the spectrum, full-scale single-tone evaluations no longer provide a complete picture of the device's performance. Because myriad signals are present in wideband radios, multiple-tone testing and SFDR power sweeps describe converter performance better than do single-tone tests.

The most common form of multitone testing uses two tones, from which you can measure third-order IMD. IMD, which is related to a mixer and an amplifier's 3OIIP, is important in receivers that must process two large signals and many smaller ones. Nonlinearities cause the two larger signals to generate spurs. These spurs always fall within the band of interest and cannot be filtered. If significant, these spurs can override smaller desired signals at the same frequencies, just as harmonics can mask small signals. Therefore, IMD performance is important, not for how it affects the two larger signals, but for how it affects smaller adjacent signals or channels.

In Figure 8, the upper IMD product is clearly visible even though it has been aliased back within the band of interest. Although IMD is an important specification, Figure 8 also shows that other spurs often present receiver-design problems. In this case, two-tone SFDR measurements are just as important as two-tone IMD. Multitone testing is not limited to two tones, and many receiver manufacturers now use as many as 48 tones to test their products.

One other dynamic specification that is vital to good radio performance is jitter, the sample-to-sample variation in the periodicity of the sample-clock path. The effects of jitter, though important to baseband performance, are magnified when the A/D converter samples higher frequency or higher slew-rate signals. Such signals are found in undersampling applications. The overall effect of a poor jitter specification is a reduced S/N ratio at higher input frequencies. This effect is demonstrated in the equation

$$\text{SNR} = 20 \log_{10} \left( \frac{1}{2 \cdot \text{pftA}} \right),$$

where $t_A = \text{aperture time}$.

This equation shows that, if jitter is constant, the S/N ratio goes down as frequency increases. In fact, overall receiver sensitivity decreases as frequency increases. Therefore, as A/D converters sample earlier in the signal chain, sensitivity to jitter increases. You can achieve ideal 12-bit performance with a 70-MHz analog signal only when $t_A$ is less than or equal to 0.5 psec. Although minimizing jitter is a key goal, oversampling and decimation tend to reduce the overall effects of jitter.
A/D converters must have a high dynamic range and be capable of sampling rates as high as 65M samples/sec. The converters must also have low jitter so that the overall S/N ratio does not suffer and degrade receiver sensitivity. The A/D converters must also have dynamic ranges from 70 to 95 dB, depending on the modulation.

### Oversampling and process gain

The sampling theorem usually determines the sampling rate required for a given signal. Normally the lowest required sampling rate is twice the signal frequency. If you sample at this rate, you must sample a 70-MHz IF signal at 140 MHz. However, if the signal does not fully occupy the bandwidth from dc to 70 MHz, but occupies only 200 kHz at 70 MHz, a sampling rate of only 400k samples/sec, or twice the signal bandwidth, is adequate. Anything beyond this rate is called oversampling. Oversampling is very important because it allows for improving the S/N ratio in the digital domain.

In contrast, undersampling is sampling a signal at a rate lower than twice the signal frequency. A single sampling process can actually over- and undersample the same signal at the same time. For example, if the signal contains frequency components in a 200-kHz band centered at 70 MHz, and you sample at 3.2M samples/sec, you are oversampling the modulation by a factor of eight but undersampling the carrier by a factor of more than 40.

In any digitization, the faster the signal is sampled, the lower the noise floor, because noise spreads out over more frequencies. (The total integrated noise remains constant, however.) This noise floor follows the equation:

\[
\text{Noise floor} = 6.02 \cdot B + 1.8 + 10 \log(FS/2),
\]

where \(B\) = number of bits. This equation represents the level of the quantization noise within the converter and shows the relationship between noise and the sampling rate, \(FS\). Each time the sampling rate doubles, the effective noise floor improves by 3 dB.

Although increasing the sampling rate produces some improvements, you do not achieve the full effect of process gain until the signal is decimated and filtered. For example, a GSM oversampling application might use a sampling rate of 26M samples/sec. The band of interest is only 200 kHz, so, by decimation and filtering, only 0.2/13 of the noise passes through the filter. Because sampling folds all signals and noise around a frequency equal to half of the sampling rate, the noise bandwidth is 13 MHz. Expressed in dB, the process gain is

\[
-10 \log(200 \text{ kHz}/13 \text{ MHz}) = 18.13 \text{ dB}.
\]

For analog modulation standards with 30-kHz bandwidth, only a very small portion of the broadband noise passes through the digital-filter passband if you sample at 40.96M samples/sec. In this example, the noise in the passband is 0.03 MHz/20.48 MHz. Expressed in logarithmic form, the process gain is \(-10 \log(30 \text{ kHz}/20.48 \text{ MHz})\) or 28.3 dB (Figures 9 and 10).

When you account for process gain, the effective S/N ratio for a given signal is

\[
\text{SNR} = 6.02 \cdot B + 1.8 + 10 \log(FS/2\cdot BW).
\]

If you know it, you can use the A/D converter's actual S/N ratio specification instead of the number
of bits. If the converter has an S/N ratio of 67 dB, you can write the equation as

\[ \text{SNR} = 67 + 10 \log(F_S/2 \cdot BW). \]

### Undersampling and frequency translation

As previously stated, undersampling is sampling at a rate lower than twice the actual signal frequency. A 70-MHz IF signal sampled at 13M samples/sec is undersampled. Nevertheless, the sampling rate meets sampling-theorem criteria as long as the signal bandwidth is less than 6.5 MHz.

Undersampling is important because it can serve a purpose very similar to mixing. When a signal is undersampled, the frequencies are aliased into the baseband, or the first Nyquist zone, as if they were originally in the baseband. Table 2 shows signals aliased into the baseband and their spectral orientation. Although sampling (aliasing) is different from mixing (multiplication), the results are quite similar but are periodic about the sampling rate. Another phenomenon is that of spectral reversal. As in mixers, certain products become reversed in sampling. Table 2 also shows which cases cause spectral reversal.

One of the biggest challenges in designing a radio is selecting the IF. Drive amplifiers and A/D converters tend to generate unwanted harmonics that show up in the digital spectrum of the data and appear as false signals. These false signals compound the problem of IF selection. Whether or not the application is wideband, careful selection of sampling rates and IFs can place these spurs where they cause no problems to digital tuners and filters, such as the AD6620, which can select the signal of interest and reject all others. By carefully selecting the input frequency range and sampling rate, you can place the drive-amplifier and A/D converter harmonics out-of-band. Other components, such as filters and IF amplifiers, determine the dynamic range.

For example, if the second and third harmonics are unacceptably high, by carefully selecting the sampling rate and signal bandwidth, you can place these second and third harmonics out of band, even in a wideband application. For a sampling rate of 40.96M samples/sec and a signal bandwidth of 5.12 MHz, placing the IF between 5.12 and 10.24 MHz places the second and third harmonics out of band (Table 3). Although this example is very simple, you can tailor it to many applications.

Undersampling provides another example of frequency planning. If the analog input-signal range is from dc to FS/2, the amplifier and filter must perform to the system specification. However, if you place the signal in the third Nyquist zone (FS to 3FS/2), the amplifier no longer must meet the system's harmonic specifications because all harmonics fall outside of the filter passband. For example, in

![Figure 11](image.png)

the filter's passband ranges from FS to 3FS/2 and the second harmonic's span—from 2FS to 3FS—lies well outside the filter's passband. The burden, thus, passes to the filter design, provided that the A/D converter meets the basic specifications at the frequency of interest. In many applications, this approach is worthwhile because SAW or LCR techniques can easily realize many complex filters at these relatively high frequencies. Although this technique lets you relax drive-amplifier harmonic requirements, you cannot sacrifice intermodulation, because you must assume that amplifier and converter intermodulation products fall within the band of interest.

### Performance expectations

When you design a receiver, use the wideband techniques described here to determine what performance is possible. The following example shows how to proceed with the analysis. Bear in mind that, in a typical receiver, each RF channel requires its own combined decimator/DSP chip. For
The analysis in Figure 12 is based on several assumptions. The first is that the receiver is noise limited. Although this assumption is unrealistic in many ways, it provides a starting point for determining performance limits. The related assumption is that SFDR and IMD do not limit performance. As the conclusion shows, this assumption is not so bad. The final assumption is that the IF bandwidth is 25 MHz. Thus, a 25-MHz band is translated downward in frequency and applied to the A/D converter input.

To start the analysis, you must calculate the noise at the antenna. This noise is not in a 50Ω resistor but is the atmospheric thermal noise in a 25-MHz band. Calculate the noise from the following equation:

\[ PN = k \cdot T \cdot BW, \]

where \( k \) is Boltzmann's constant \((1.38 \cdot 10^{-23} \text{ J/K})\), \( T \) is the temperature in °C, and \( BW \) is the bandwidth in Hz.

When evaluated at 3008K, the available noise power is 103.5 fW or –99.85 dBm. This noise becomes the background noise at the receiver input. Remember that this noise is distributed across the entire 25-MHz band. The next portion of the receiver is the RF section, which consists of a preselect filter, an amplifier, and a mixer. From a study of several manufacturers' wideband-component data books, you can assume that this stage's combined noise figure (NF) is 10 dB. Good low-noise amplifiers and microstrip filters can achieve such performance. If you realize that most of the noise degradation occurs in the receiver's first few stages, you can use traditional computational methods to calculate the RF/IF processing stages' overall NF. Select a conversion gain of 30 dB—some standards may require more, others less. When you apply the noise figure and conversion gain to the noise, the total noise power at the A/D converter input becomes –99.85+10+30=–59.85 dBm.

For this application, use an A/D converter, such as the 12-bit AD9042. Although the sampling rate is only 41M samples/sec, units offering 65M samples/sec will soon be available. This rate is suitable for full-band AMPS digitization and is high enough to handle the GSM 53 reference clock. The sampling rate is more than adequate for AMPS, GSM, and CDMA. From the data sheet, the typical S/N ratio is 68 dB. Therefore, the next step is to calculate how the A/D converter degrades the receiver noise. Although the calculation seems difficult, the simplest method is to convert both the S/N ratio and the receiver noise into rms volts and obtain the total rms noise by determining the vector sum

\[ V_{\text{NOISE}} = 0.35 \cdot 10^{-\text{SNR/20}} \text{ or } 0.14 \text{ mV} \]

(the rms value of the input range is 0.35V). Similarly, converting –59.85 dBm to watts yields

\[ P_{\text{NOISE}} = 0.001 \cdot 10\text{dBm}/10 \text{ or } 1.035 \cdot 10^{-9} \text{W}. \]

Converting to volts, you get

\[ V_{\text{RMS}} = \pm R \cdot P_{\text{NOISE}} \]

or 0.2275 mV (50V load). This equation gives a total rms noise voltage of 0.2671 mV, which is the
total noise present in the A/D converter due to both receiver noise and A/D converter noise, including quantization noise.

**Process gain**

This noise voltage contributes to the overall A/D converter performance in the following way. Assume that only one RF signal is present in the receiver bandwidth. The S/N ratio is then 20 log(SIG/NOISE)=20 log(0.3535/0.0002671)=62.43 dB. The final noise calculation involves process gain. As discussed earlier, process gain results from eliminating all noise (and signals) outside the band of interest (Figures 9 and 10). In essence, decimation and FIR filtering digitally apply narrowband filtering to the wideband data. The process gain is 10 log(BWADC/BWFILTER).

The A/D converter bandwidth, which is the bandwidth that the A/D converter effectively digitizes, is always one half of the sampling rate. All of the sampled noise is distributed in this bandwidth, even if the A/D converter operates in an undersampling mode. For the AD9042 family of converters, the sampling rate is as high as 65M samples/sec. The modulation system determines the required bandwidth of the digital filter. US analog standards, which allow a deviation of ±12.5 kHz, require a filter bandwidth of 25 to 30 kHz. A worst-case bandwidth of 30 kHz gives a processing gain of 30.35 dB. Thus, the total S/N ratio for the full-scale input is 92.61 dB with process gain.

However, because this radio is wideband, the A/D converter must share its dynamic range with several RF carriers, and the signal must be less than full scale. If the receiver has eight active carriers (and many other smaller carriers from adjacent cell sites), and all active carriers have equal amplitude, each must be 18 dB below full scale (1/8 of full scale) to keep the converter from clipping (Figure 13). Lower power signals from adjacent cells, as well as larger signals that come inband unexpectedly (new calls, for example) require additional headroom, typically 3 to 15 dB. For this discussion, 3 dB suffice. Thus, headroom and multicarrier operation use a total of 21 dB of S/N ratio. The total S/N ratio in the single channel's multisignal environment is 71.61 dB. This value is equivalent to a carrier-to-noise (C/N) ratio of 71.61 dB in a multichannel environment. The minimum C/N ratio for adequate demodulation depends on the modulation scheme. If the scheme is digital, then you must consider the BER (Figure 14).

If the minimum C/N ratio is 10 dB, the input signal cannot be so small that the remaining S/N ratio is less than 10 dB. Thus, the signal level can fall 61.61 dB from its present level of 21 dB below the A/D converter's full scale, or 82.61 dB below the A/D converter's full scale. Because the A/D converter has a full-scale range of 4 dBm, the signal level at the A/D converter's input is then -78.61 dBm. Because of the 30-dB signal gain, this value is a noise-limited sensitivity of -108.61 dBm, referred to the antenna. Actual measurements of this receiver's performance support a sensitivity measurement of -104 dBm, very close to the theoretical limit. Possible sources of reduced sensitivity include lower A/D converter S/N ratio, higher analog-chain noise, and effects of aperture jitter, as shown in Equation 1.

This noise-limited example does not adequately demonstrate a wideband receiver's true limitations because other limitations, such as SFDR, are more restrictive than S/N ratio and noise. Assume that the A/D converter has an SFDR of -80 dBFS or -76 dBm (full scale=4 dBm). Also assume that a tolerable carrier-to-interferer (C/I—different from C/N) ratio is 18 dB. This ratio means that the minimum signal level is -62 dBFS (-80+18) or -58 dBm. At the antenna, this level is equivalent to -88 dBm. Therefore, SFDR (single or multitone) limits receiver performance long before the actual
noise limitation is reached.

### Using dither to improve SFDR

Adding an out-of-band noise signal called dither greatly improves SFDR, essentially eliminating spurs above the noise floor (Reference 1). Although the dither required depends on the converter, the technique applies to all A/D converters. In the AD9042, the added noise is only –32.5 dBm, or 21 codes rms. Plots with and without dither ([Figures 15](#) and [Figures 16](#)) provide insight into the potential for improvement. Dither works by randomizing the coherent spurs generated within the A/D converter. Because the energy of the spurs cannot change, dither transforms them into slight additional noise over the entire frequency band that the A/D converter digitizes. Dither thus removes all internally generated spurs at the cost of a slight decrease in the overall S/N ratio. The sensitivity loss amounts to approximately 1 dB compared with the noise-limited example. Sensitivity is much better than in the SFDR-limited example.

The RF section contributes to the receiver's spurious performance in several ways. Unless you take steps to limit them, spurs that originate in the RF section disrupt the receiver's performance. Dither does not affect these spurs. A review of the specification of GSM, perhaps the most demanding of receiver applications, indicates the performance that modern receivers demand of wideband RF components.

A GSM receiver must recover signals with a power level between –13 and –104 dBm ([Figure 7](#)). Assume also that the full scale of the A/D converter is 0 dBm and that losses through receiver filters and mixers are 12 dB. Also, because the receiver processes several signals simultaneously, it should not employ AGC. AGC reduces RF sensitivity and causes weaker signals to be dropped. Working with this information, RF/IF gain is calculated to be 25 dB (0=-13–6–6+x).

[Figure 17](#) shows the distribution of the 25-dB gain. Although a complete system would have additional components, the block diagram is adequate for this discussion, which provides a basis for determining whether the A/D converter introduces excessive noise or spurs. In this circuit, with a full-scale GSM signal of –13 dBm, the A/D-converter-input signal is 0 dBm; with a minimal GSM signal of -104 dBm, the A/D-converter-input signal is –91 dBm. With these signals and the required system gains, you can examine the performance of the amplifier and mixer with a full-scale, –13-dBm signal. Solving for the third-order products in terms of the full-scale signal yields:

\[
IIP = SIG + (3OP - SIG)/-3; \quad (2)
\]

where IIP=input-intercept point, SIG=full-scale input level, and 3OP is the required third-order-product level.

Assuming that overall SFDR must exceed 100 dB, solving this equation for the front-end amplifier shows a third-order input amplifier with a 3OIIP greater than 16 dBm. At the mixer, the signal level has been amplified by 10 dB, and the new signal level is –3 dBm. Therefore the mixer must have a 3OIIP greater than 35.3 dBm. At the final gain stage, the signal is attenuated to –9 dBm. For the IF amplifier, the 3OIIP is greater than 21.3 dBm. If these specifications are met, the SFDR should equal or exceed 100 dB. Solving Equation 2 for 3OP results in
3OP=SIG–3(IIP–SIG).

A quick check of available components shows 3OIIIs of 40, 38, and 25 dBm for the RF amplifier, mixer, and IF amplifier, respectively. These numbers and the signal levels for the appropriate stages result in 3OPs of -172, -126, and -111 dBm, respectively. The combined effect of these IMD-stage measurements is better than -100 dBm.

**Changing devices, changing architectures**

Semiconductor manufacturers are now integrating radio functions that once required multiple circuit boards, and integrated-receiver architectures now differ radically from older designs. The results are reduced size and cost and improved performance and flexibility. The pace of change in digital radio gives no indication of slowing any time soon, as higher levels of integration continue to emerge. Eventually, chip sets will meet even complex base-station requirements. Already, IC manufacturers are offering A/D converters, digital tuners, and DSPs that require little or no glue logic to form a base-station core.

The approaches embodied in the circuit of Figure 18 are but one alternative for a wideband receiver. Many options exist, but the circuit offers many advantages, including dynamic control of dither, which increases and decreases the dither to optimize performance as traffic volume changes. Lab tests of this design showed good sensitivity. A more rigorous design effort could produce further improvements. One idea worth trying is the use of lower loss filters, possibly implemented as microstrip elements, and lower noise active components.

Although the circuit lacks some key elements, this design provides the framework for a cost-effective wideband-radio architecture. With the inevitable performance improvements and price reductions in RF, converter, and digital devices, wideband software radios will become a reality in many commercial applications within the next year. Indeed, at the 1994 Cellular Telecommunications Industry Association show, no fewer than four system suppliers offered wideband radios. Even more of these products appeared at the 1995 show, and still more will follow, as new components continue to provide better performance at lower cost.

### Table 1—Costs of important narrowband and wideband receiver elements

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<thead>
<tr>
<th></th>
<th>A/D converter</th>
<th>Mixer</th>
<th>Digital down-converter</th>
</tr>
</thead>
<tbody>
<tr>
<td>Narrowband</td>
<td>$8</td>
<td>$10</td>
<td>not applicable</td>
</tr>
<tr>
<td>Wideband 1994</td>
<td>$250</td>
<td>$100</td>
<td>$100</td>
</tr>
<tr>
<td>Wideband 1995</td>
<td>$50</td>
<td>$50</td>
<td>$15</td>
</tr>
</tbody>
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### Table 2—How undersampling shifts signal frequencies

<table>
<thead>
<tr>
<th>Input signal</th>
<th>Frequency range</th>
<th>Frequency shift</th>
<th>Spectral sense</th>
</tr>
</thead>
<tbody>
<tr>
<td>1st Nyquist</td>
<td>dc to F_s/2</td>
<td>0</td>
<td>Normal</td>
</tr>
<tr>
<td>2nd Nyquist</td>
<td>F_s to F_s</td>
<td>F_s to F_Sig</td>
<td>Reversed</td>
</tr>
<tr>
<td>3rd Nyquist</td>
<td>F_s to 3F_s/2</td>
<td>F_Sig to F_s</td>
<td>Normal</td>
</tr>
<tr>
<td>4th Nyquist</td>
<td>3F_s/2 to 2F_s</td>
<td>2F_s to F_Sig</td>
<td>Reversed</td>
</tr>
<tr>
<td>5th Nyquist</td>
<td>2F_s to 5F_s/2</td>
<td>F_Sig to 2F_s</td>
<td>Normal</td>
</tr>
</tbody>
</table>

Note: F_s is the sampling rate, and F_Sig is the frequency of the input signal being sampled.

### Table 3—Choosing the sampling rate so that harmonics do not interfere with the fundamental frequency

<table>
<thead>
<tr>
<th>Sampling rate: (samples/sec)</th>
<th>Fundamental:</th>
<th>Second harmonic:</th>
<th>Third harmonic:</th>
</tr>
</thead>
</table>
Table 3—Choosing the sampling rate so that harmonics do not interfere with the fundamental frequency

<table>
<thead>
<tr>
<th>Sampling Rate</th>
<th>Frequency Bands</th>
</tr>
</thead>
<tbody>
<tr>
<td>40.96M</td>
<td>5.12 to 10.24 MHz</td>
</tr>
</tbody>
</table>

Note: With the values shown, undersampling can "fold" the third harmonic into the second-harmonic band but not into the fundamental-frequency band.

References

1. Brannon, Brad, Overcoming Converter Nonlinearities with Dither, Application note AN-410, Analog Devices Inc.
6. Analog Devices Inc, AD9042 Data Sheet.
7. Analog Devices Inc, AD6620 Preliminary Data Sheet.
8. Harris Semiconductor, HSP50016 Data Sheet.

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Author's biography

Brad Brannon, a development engineer, has worked at Analog Devices Inc in Greensboro, NC for the past 12 years. He holds a BSEE from North Carolina State University (Raleigh, NC) and has helped to develop the AD9042 and other high-speed, high-resolution A/D converters. His spare-time interests include his family, home projects, and computers.