Receiver Dynamic Range: Part 1

The task of the radio receiver has always been to "get the signal." However, with the proliferation of high-powered transmitters and the burgeoning growth of electronic noise pollution, often weak-signal reception is difficult, if not impossible. Receiver dynamic range is the measure of a receiver's ability to handle a range of signal strengths, from the weakest to the strongest. Because of the severe dynamic range requirements placed on modern receivers, it is imperative to define rational criteria for evaluating receiver performance. This two-part article provides a tutorial review of receiver dynamic-range specifications and measurements. It discusses the limits and applicability of the various measurements, highlighting potential errors and misleading specifications. Procedures for estimating and measuring true receiver performance are recommended.

PRIMARY MEASUREMENTS

Primary measurements that affect receiver dynamic range include: noise figure, second-order intercept, third-order intercept, 1-dB compression, phase noise, internal spurs and bandwidth. This group of receiver measurements is considered primary because most other receiver dynamic-range measurements can be predicted from them.

NOISE FIGURE

The most common expression of noise figure is the ratio (in dB) of the effective receiver input noise power with respect to -174 dBm/Hz. This single number dominates those receiver characteristics which are generally described as sensitivity. It also describes the "noise floor" of most dynamic-range measurements.

To determine noise figure accurately, it should be measured at a pre-detected output of the receiver: that is, at any output which is a version of the received input modified only by linear amplification, frequency translation, and bandwidth. Because noise figure degrades with each successive stage of the receiver, the most desirable measurement port is the audio output. Measurement at this port can be accomplished in either the cw or ssb mode because both of these are pre-detection modes. Note, however, that some receivers may not use a true product detector for cw detection, and the apparent noise figure will be degraded. In the case of receivers that do not have pre-detected audio outputs, the IF output may be used for noise-figure measurements. The most rigorous measurement will require the selection of the narrowest available IF bandwidth because it is under this condition that the largest number of receiver stages are in the signal path.

In general, the noise figure of most radios will vary both with receiver temperature and tuned frequency. Because of this, it is useful to note the frequency and temperatures over which the receiver data is valid.

SECOND- AND THIRD-ORDER INTERCEPT

Second- and third-order intercept, which are measures of receiver linearity, dominate the signal overload end of receiver dynamic-range specifications. It is tempting to define receiver dynamic range in terms of noise floor and overload level alone. However, measurement of second- and third-order intercept is somewhat more problematic than measurement of noise figure. Nonetheless, these measurements can be used to predict a wide range of receiver performance.

The second-order input intercept point (IIP2) is the receiver input level at which the curves of linear output and second-order distortion intersect. The third-order input intercept point (IIP3) is, similarly, the receiver input level at which the curves of linear output and third-order distortion intersect (see Figure 1).

In the most common measurement of these parameters, two equal-amplitude sinusoids are linearly combined and applied to the receiver input. The distortion products appear at the output as new frequency components whose relative magnitudes are measured and compared to the original inputs. A single set of data is then used to extrapolate the curves of receiver distortion. For conve-
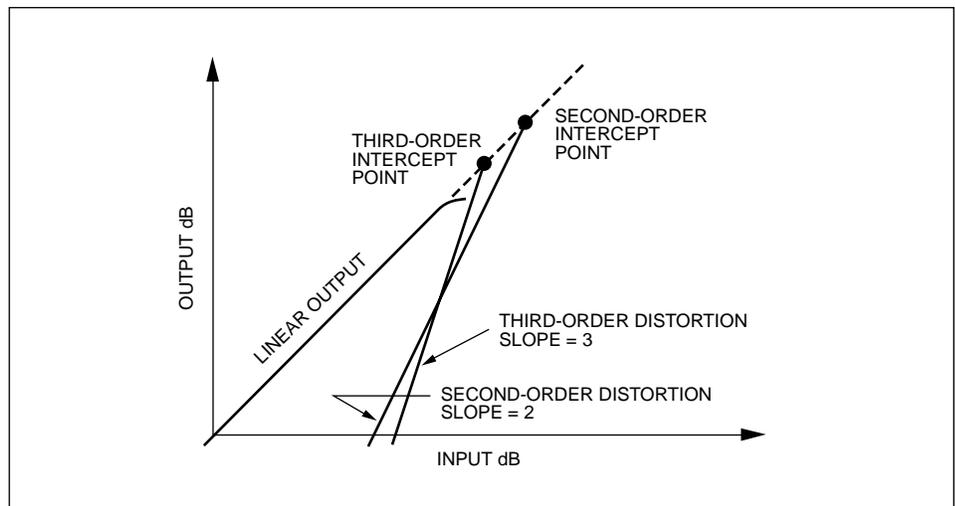

Figure 1. Receiver distortion vs. input power intercept point extrapolation (theoretical).

nience, these curves can be completely specified by the intercept points.

Several problems exist with intercept specifications. The first problem is that the intercept points are not directly measurable. Because the intercept points are mathematically extrapolated, their accuracy depends on the assumption that the curves of second- and third-order distortion are described by straight lines with slope values of two and three, respectively. To be useful, this assumption must be valid over the usable dynamic range of the receiver. Unfortunately, there are two potential errors associated with this assumption. First, as the receiver approaches overload compression, the actual distortion curves are no longer straight lines. This effect can be avoided by measuring the distortion products at relatively low input levels. Typically, the intercept measurements will be most accurate if measured at input levels where the distortion products are 60 dB less than the input signals. Second, certain nonlinear radio components do not seem to produce distortion curves of the appropriate slope. Examples of this are ferrite and GaAs FET components. This effect can be detected by measuring the intercept point at two different input levels and comparing the results for agreement. If they do

not agree, the validity of the intercept specification is in doubt.

Another problem of intercept measurements is their frequency dependence. Second-order distortion produces distortion components at twice the frequency of a single input signal (second harmonic distortion) and at the sum and difference frequencies of two input signals. It can be shown that a band-pass filter at the receiver input can suppress the undesired signal that would otherwise produce second-order distortion products at the receiver's tuned frequency. Because realizable filters are not ideal, rejection of second-order distortion will typically vary with the frequency of the undesired signals as well as with the receiver's tuned frequency.

Third-order distortion effects are even more frequency dependent than second-order distortion. This is because third-order distortion can produce distortion components from unwanted signals that a receiver input filter cannot remove. In this case, the distortion components from two input sinusoids occur at frequencies of twice the first frequency minus the second. While a receiver input filter can remove most of the undesired signals that can produce third-order distortion products at the tuned frequency, the signals which are within the passband of

the input filter can also produce distortion products (see Figure 2). The filtering in a typical receiver is produced by the cascade of several different parts of the receiver. For example, the filtering is provided by the input (rf) filter, followed by a narrower first IF filter, and finally, a still narrower final IF filter. Because as signals pass through the receiver they are increasingly distorted, it is necessary to specify at what frequencies the unwanted signals occur with respect to the tuned frequency. Typically, third-order distortion, due to signals at frequencies within the final IF passband, is worse than that due to signals in the first IF passband, but outside the final IF. Distortion from signals in the input filter passband, but outside the first IF, is even less, and distortion from signals with frequencies outside the input filter is the least. For this reason, measurements of receiver third-order intercept can vary radically depending on the frequencies of the test signals with respect to the receiver's tuned frequency (see Figure 3). It is typical for receiver manufacturers to specify the third-order intercept point for test tones at frequencies in the first IF. Another common specification is for test-tone frequencies outside the first IF, but inside the input filter. In this case, the manufacturer should specify the frequencies of the test signals with
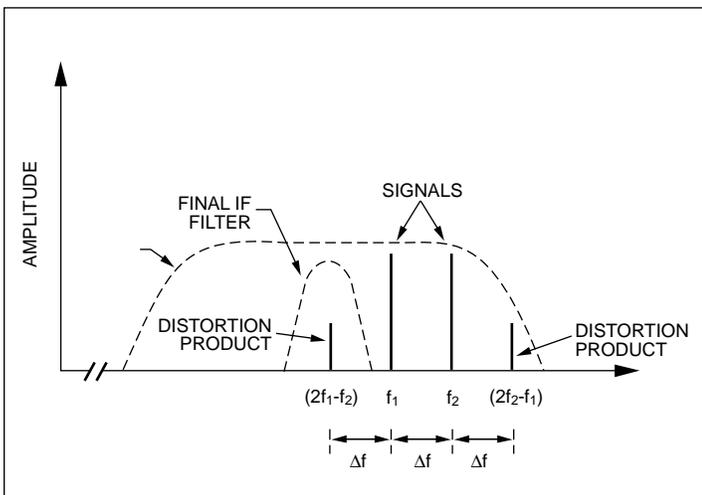

Figure 2. Third-order distortion products from two signals inside the receiver input filter.
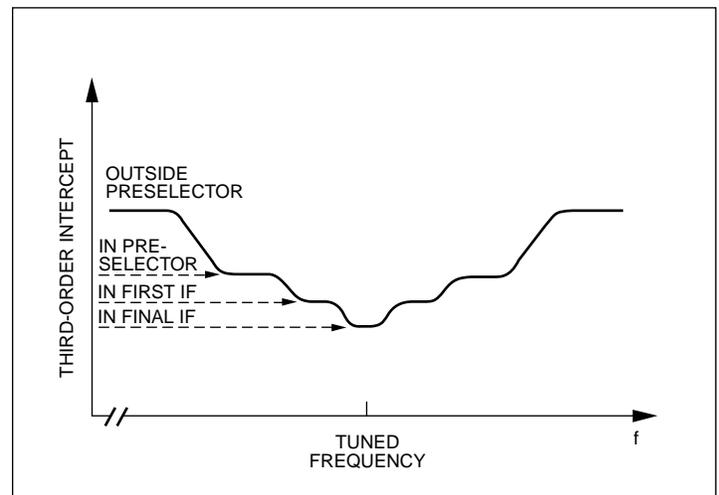

Figure 3. Third-order intercept as a function of test-tone frequency relative to receiver-tuned frequency.

respect to the tuned frequency. Like second-order distortion, third-order distortion can also vary with tuned frequency, so a proper specification should list a worst-case value or specify the frequency range of validity.

1-DB COMPRESSION

The 1-dB compression point is the measure of receiver performance that indicates the input level at which the receiver begins to deviate radically from linear amplitude response. In a linear device, for each dB of input-level increase, there is a corresponding dB increase in output level. In the case of input overload, the output does not continue to increase with each input increase, but instead, the output tends to limit. The input level at which the output deviates from linear response by 1 dB is known as the 1-dB compression point.

There are two general forms of 1-dB compression which are useful as receiver specifications. The first is the 1-dB compression of the desired signal due to its own signal power causing receiver overload. The second form is the 1-dB reduction of the output level of the desired signal due to a strong undesired signal causing receiver overload. This second form is usually called blocking or desensitization.

Measuring the 1-dB compression point of the receiver due to overload by the desired signal can be performed by noting the input level, in manual gain mode, at which an input-level decrease of 10 dB causes an output-level decrease of 9 dB (see Figure 4). A delta of 10 dB is a convenient value because smaller deltas may make the 1-dB compression point difficult to measure due to the gradual compression characteristics of some devices. Conversely, larger deltas may indicate 1-dB compression points at levels far from the onset of input overload.

This 1-dB compression point value is usually somewhat affected by the receiver tuned frequency, and is often strongly affected by the

receiver gain setting. The receiver-gain effect occurs because many receivers, as part of their gain-control scheme, attenuate signals early in the receiver signal path. If a receiver were to control its gain by rf attenuation alone, its 1-dB compression point could theoretically be unlimited. For this reason, the 1-dB compression point is best used to describe the upper limit of dynamic range for desired signals only.

Measuring the 1-dB compression point due to blocking can be accomplished by combining a small, desired sinusoid with a large, undesired sinusoid, and applying them to the receiver input. The desired sinusoid is at the receiver's tuned frequency and is adjusted for a 10-dB signal-to-noise ratio (SNR) in the narrowest receiver bandwidth at maximum receiver gain. The amplitude of the undesired, out-of-band sinusoid is then increased until the output of the desired sinusoid is reduced by 1 dB.

One-dB compression due to blocking measurements is strongly affected by the relative frequency of the interfering signal with respect to the receiver's tuned frequency. Like third-order intercept, the blocking performance of most receivers will improve as the undesired signal is moved away from the receiver's tuned frequency. Blocking is usually specified for undesired signals outside the first IF bandwidth. In part, this is because

undesired signals near the tuned frequency will interact with the receiver phase noise and degrade the SNR of the desired signal.

Receiver second- and third-order intercept points are typically much greater than the 1-dB compression points, but there is no reliable method for predicting one value from the others. Two receivers with identical 1-dB compression points may have third-order intercept points which differ from each other by as much as 20 dB, and vice versa. For this reason, the 1-dB compression point is a useful specification to supplement the other measurements as an indicator of receiver performance at high signal levels.

PHASE NOISE

Receiver phase noise is a measurement of phase and frequency perturbations added to the input signals by the receiver frequency-conversion oscillators. For signals inside the final IF bandwidth, the effect of this phase noise is to degrade angle-modulated signals. A second effect of phase noise is due to undesired out-of-band signals which mix with oscillator phase noise to produce in-band noise that degrades receiver sensitivity. This second effect is usually called reciprocal mixing. Receiver phase noise can be expressed as the amplitude of the phase noise sidebands added by the receiver to a spectrally pure input sinusoid. This is most
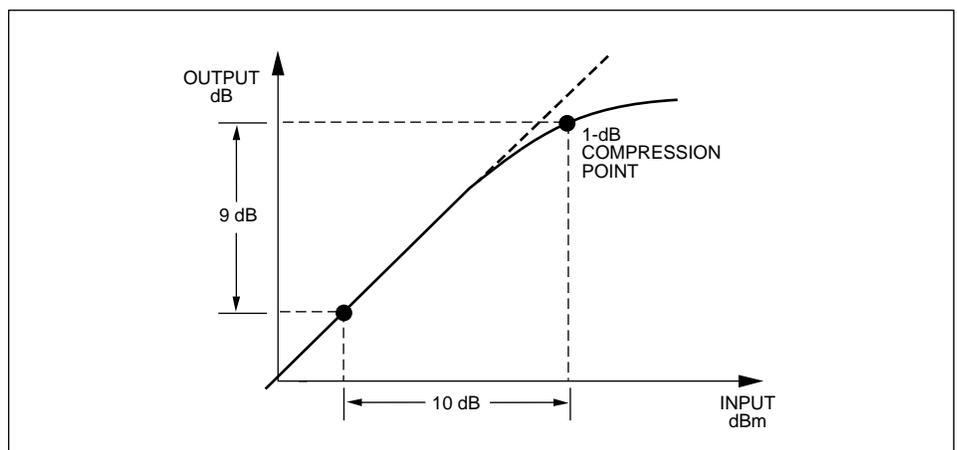
Figure 4. 1-dB compression point.

commonly specified as a single sideband noise spectral density expressed as dBc/Hz; that is, noise power in a 1-Hertz bandwidth compared to the signal ("carrier") power. Discrete spectral lines, usually called "spurs," are specified in dBc with no reference to bandwidth (see Figure 5). These spurs are troublesome because they can cause frequency translation of out-of-band signals into the receiver-tuned frequency band.

Receiver phase noise can be measured by connecting a spectrally pure sinusoid to the receiver input and measuring its phase-noise degradation. Phase noise close to the "carrier" can be measured by examining the IF output spectrum with a spectrum analyzer. However, the analyzer, like the test signal, must have better phase-noise specifications than the receiver under test. Phase noise out of the receiver passband; that is, farther removed from the "carrier," can be measured by observing reciprocal-mix effects. The receiver is set in manual-gain mode and the IF output is observed with a spectrum analyzer as the test sinusoid is moved in frequency with respect to the receiver's tuned frequency. The signal amplitude is adjusted until the reciprocal-mix phase noise can be measured at the IF output. The receiver phase noise (dBc) is equal to the output level minus the receiver gain, and then compared to the test-signal level. Care must be taken

to assure that the test signal does not exceed the receiver blocking 1-dB compression point and that the output noise is dominated by phase noise.

Receiver phase-noise performance is the product of both the oscillator phase noise and the receiver filtering. Consequently, the phase-noise performance is strongly affected by the frequency offset from the tuned frequency. A typical receiver's phase noise might be specified at offsets of 100 Hz, 1 kHz, 10 kHz, 100 kHz, 1 MHz, and 10 MHz.

INTERNAL SPURIOUS SIGNALS

Spurious signals internal to the receiver effectively degrade the receiver noise floor. The severity of this problem is a function of both the magnitude and number of the internal spurs. Unless otherwise indicated, it must be assumed that a spur specification (which lists only a spur level) indicates that the receiver has many spurs of that level. A more complete specification might, for example, indicate a maximum number of spurs per MHz with a certain power limit and a lower limit for all others.

Internal spur measurement requires that the receiver be set in maximum gain mode and scanned in its narrowest bandwidth over its entire range of tuned frequencies, while using the finest tuning resolution available.

For some receivers, this may be a daunting task. A broadband receiver may require measurements at over 10,000,000 discrete frequencies. Tuning manually at one frequency per second, a single test of all frequencies would take nearly four months of continuous testing. Unfortunately, it is generally necessary to test all possible frequencies because modern frequency-synthesized local oscillators generate a myriad of interacting spurs, each of which may appear only at a single tuned frequency. This type of spur is often colloquially called a "pop-up" spur because it pops up in a single frequency increment. Automation of testing can speed up the process, but practical testing must also be based on a deeper understanding of the spur mechanism so that fewer frequencies need to be tested.

BANDWIDTH

Bandwidth plays a dominant role in receiver dynamic range because it generally describes a receiver's ability to reject unwanted signals, which can reduce its ability to detect weak, desired signals. Undesired signal rejection includes: rejection of undesired signals by the final IF filter so that they do not reach the detector stages (adjacent channel rejection); rejection by the input and first IF filters to protect the receiver from overload and phase noise effects; and rejection by the input and first IF filters of undesired input signals at image and IF frequencies.

Measuring receiver filter bandwidths, unfortunately, can be a difficult or impossible task to accomplish without invading the "guts" of the receiver. Measuring the -3-dB bandwidth of the final IF filter is usually easy; however, measuring the preselector filter band width is much more difficult, and measuring the ultimate attenuation of the IF filters may be impossible.

The overall receiver bandwidth can be measured by setting the receiver to manual gain and frequency sweeping the input with a constant amplitude sinusoidal test signal.
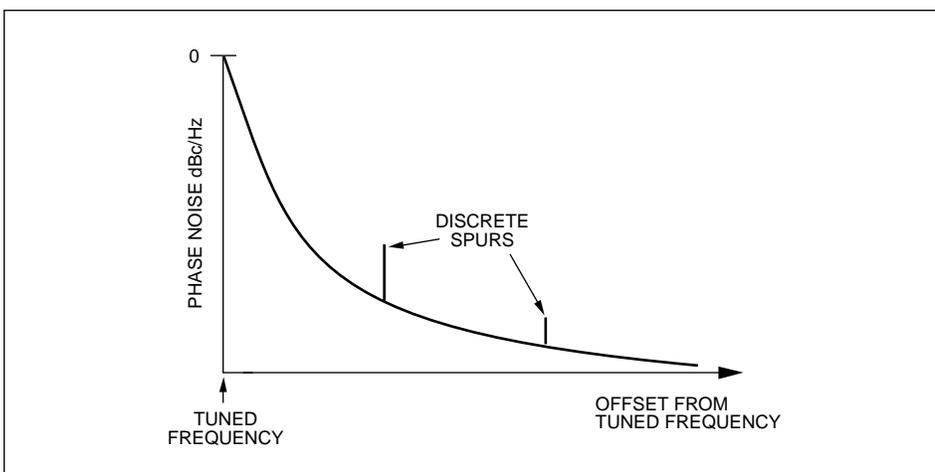

Figure 5. Receiver phase noise.

The IF output, measured relative to its output when the test tone is at the tuned frequency, determines the receiver's frequency response. A narrowband spectrum analyzer can be used to increase the sensitivity of the measurement because it can be used to look for the signal output "below" the IF noise. Because of possible receiver nonlinearity effects, this test should be performed with more than one input level, and the results compared. True filter response should be independent of input level.

Specification of receiver bandwidth should include the -3-dB and -60-dB bandwidths. These two numbers give a good indication of the signal bandwidth that the receiver will pass and the separation required between signals so that the receiver can reject them. The -3-dB bandwidth should be a guaranteed minimum bandwidth and the -60-dB bandwidth should be a guaranteed maximum. The choice of the -3-dB and -60-dB bandwidths is somewhat arbitrary, but not without reason. The -3-dB bandwidth is a more realistic estimator of usable signal bandwidth than the often-quoted -6-dB bandwidth. In addition, for most modern multipole filters, the -3-dB bandwidth is approximately equal to the filter noise equivalent bandwidth. The -60-dB bandwidth represents a level of undesired signal attenuation which is easily achieved with good bandpass filters; however, for receivers with good intercept and phase-noise specifications, a useful specification might include the -70-dB or -80-dB bandwidth.

SECONDARY MEASUREMENTS

Common secondary measurements of receiver dynamic range include: sensitivity, cross modulation, intermodulation distortion, and reciprocal mix. This group of measurements is considered secondary because, while they are useful, they can generally be predicted from the results of the primary measurements. In some cases, the wide variety of ways they are presented makes mean-ingful comparisons between receivers a difficult task at best.

SENSITIVITY

Sensitivity is probably one of the most confusing, misquoted and often most completely misunderstood of all receiver specifications. It attempts to indicate how well a receiver will capture weak signals, but unlike noise figure, it can be specified in many different ways.

Most sensitivity specifications list a required signal strength for a certain received signal quality with a specified bandwidth, modulation type and percentage of modulation. With so many variables, the possible number of different specifications is virtually unlimited. In order to minimize the confusion, it is useful to examine each of the variables independently.

Signal quality for sensitivity specifications is given most frequently in terms of signal-to-noise ratio (SNR); that is, the ratio of output signal power to the output noise power. A convenient - if somewhat arbitrary - commonly quoted value is an SNR of 10 dB. Because it is not always convenient to measure SNR directly, the related measurement, signal plus noise-to-noise ratio ([S+N]/N) is used. Mathematically, $(S+N)/N = S/N+1$. While the relationship is simple when expressed as simple ratios, expressed in dB, the difference between the two measurements depends on their value. For example, a ([S+N]/N) ratio of 3 dB equates to an SNR of 0 dB. At the more usual level of 10 dB ([S+N]/N), the SNR is 9.54 dB. Another related measure that is increasingly popular is signal-plus-noise-plus-distortion to noise-plus-distortion ratio (SINAD). This measurement is essentially the same as ([S+N]/N) with distortion included in the noise term. This is useful because, for most users, the distortion is no more usable than the noise.

In general, signal quality expressed as SNR can be predicted by knowing the signal type, signal strength, receiver bandwidth and noise figure. Because signal type and strength vary with the receiver application, the advantages of comparing receiver sensitivities on the basis of noise figure are apparent.

Signal strength is properly defined in terms of available signal power, which

is commonly given in dBm. Unfortunately, for historical reasons, signal strength is often specified in terms of signal voltage, commonly given in microvolts, millivolts, dB(V and dBmV. The first problem with voltage specifications is that it is not clear where the voltage should be measured. Some specifiers prefer to use source emf; that is, the unterminated (open circuit) output of the signal generator. Other specifiers measure the voltage across the receiver input terminals. In an impedance-matched system, this difference amounts to a 6-dB advantage to receivers whose sensitivities are specified with voltages at the input terminals. For this reason, most manufacturers who specify voltage sensitivity use the latter method.

A second problem with the voltage specifications is that the source and load impedances must be known. At one time in the United States, it was common practice to specify FM broadcast receiver sensitivities in terms of microvolts without direct reference to input impedance. Since all of these receivers used 300-ohm inputs, direct comparisons of sensitivity could be made. However, many manufacturers added 75-ohm inputs to match a common coaxial cable impedance. Soon after, some of the less scrupulous manufacturers began to specify their sensitivities in microvolts at the 75-ohm input instead of the 300-ohm input. To the unwary consumer, these receivers appeared to be twice as sensitive as their competition because the required voltage had been halved. In order to counter this sort of deception, the Federal Trade Commission required the use of sensitivity specifications in dBf (dB from a femtowatt). If, for typical communications

receivers, an input impedance of 50 ohms were specified, the problem would seem to be avoided. Again, this is not the case because even a relatively good VSWR specification of 2:1 allows input impedance variations which will affect the signal input power by up to (3 dB. Specifying signal strength in terms of available power (preferably in dBm) eliminates all of the ambiguities and is, therefore, the preferred method.

Bandwidth specification is mandatory for any sensitivity specification because the amount of noise power relative to the signal increases linearly with bandwidth. For example, cw sensitivity in a 1-kHz bandwidth is improved 10 dB in a 100-Hz bandwidth. An additional subtlety related to bandwidth is that post-detection "video" filtering can have a significant effect on output SNR. For example, the detected SNR of an AM signal with 1-kHz modulation, tuned in a 10-kHz IF bandwidth, can be improved 3 dB by reducing the post-detection bandwidth from the customary one-half IF bandwidth of 5kHz to an audio bandwidth of 2.5 kHz. This may explain the popularity of specifying receiver sensitivity with low-frequency modulations measured at the audio output.

Modulation type and percentage modulation are probably the most confusing components of the sensitivity specification, because the signal quality may not be linearly related to signal strength in the case of some modulations. For example, both AM and FM modulations exhibit "threshold" effects at low SNRs. This is especially true for FM signals with wide deviations and low modulation frequencies. Above threshold, the detected SNR is better than the predetected SNR, but below threshold, the reverse may be true. In general, for most signals, SNR is improved with increases in modulation level. In the case of AM and FM above threshold, the detected SNRs will increase as the square of the modulation percentage and modulation index increase, respectively. For high-percentage modulation AM signals, distor-

tion may be increased at low SNR, and this will be reflected in degraded SINAD values. FM SNR performance is often complicated further by the effect of de-emphasis filtering which, therefore, should be specified when used.

CROSS MODULATION

Cross modulation specifies the amount of AM modulation which is transferred from an undesired signal to a desired signal. This specification includes the percentage of modulation of the interfering signal, its signal power, and frequency offset from the tuned frequency. The conditions of this specification vary from receiver to receiver, but the level of cross modulation can be predicted by using the receiver third-order intercept point data. The percentage of modulation on the desired signal due to cross modulation is equal to the percentage of modulation of the undesired signal multiplied by four times its power and divided by the sum of the third-order intercept power and twice the undesired power. In algebraic notation, this can be expressed as:

$$\%d = \%u \ (4 \ P_u)/(P_{ip} + 2 \ P_u)$$

where:

$\%d$ is the percentage of modulation on the desired signal due to cross modulation

$\%u$ is the percentage of modulation on the undesired signal

P_u is the power of the undesired signal

P_{ip} is the receiver third-order input intercept point power

INTERMODULATION DISTORTION

Intermodulation distortion is used to describe the effects of receiver third-order distortion. It usually specifies the levels and frequency offsets of two test signals and the level of the resultant in-band distortion component. Like cross modulation, intermodulation distortion can be predicted from the receiver third-order intercept point. For two

equal-amplitude test tones, the power of the intermodulation distortion product is equal to three times the power of a single test tone minus two times the third-order intercept point, where all powers are in dBm. That is:

$$P_{im} = 3P_t - 2 \ P_{ip}$$

where:

P_{im} is the power of the intermodulation product in dBm

P_t is the power of a single test tone in dBm

P_{ip} is the power of third-order intercept point in dBm

RECIPROCAL MIX

The reciprocal-mix specification typically states that the magnitude of the noise power in a specific bandwidth is caused by an out-of-band undesired signal of a specified level and frequency offset mixing with receiver phase noise. This can be readily calculated from a knowledge of receiver phase noise. The reciprocal mix phase-noise power is equal to the amplitude of the undesired signal plus ten times the log of the measurement bandwidth plus the receiver phase noise at the frequency offset of the undesired signal. That is:

$$P_{pn} = P_u + 10 \log BW + P_{pnr}$$

Where:

P_{pn} is the equivalent input phase noise in dBm

P_u is the power of the undesired signal in dBm

BW is the receiver test bandwidth in Hz

P_{pnr} is the receiver phase noise at the undesired frequency offset in dBc/Hz

Part 2 of this article introduces comprehensive measurements, which attempt to characterize a receiver's dynamic range as a single number.

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