Watkins-Johnson Company
Tech-notes

Vol. 14 No. 1 January/February 1987

Receiver Dynamic Range: Part 1

Robert E. Watson

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).

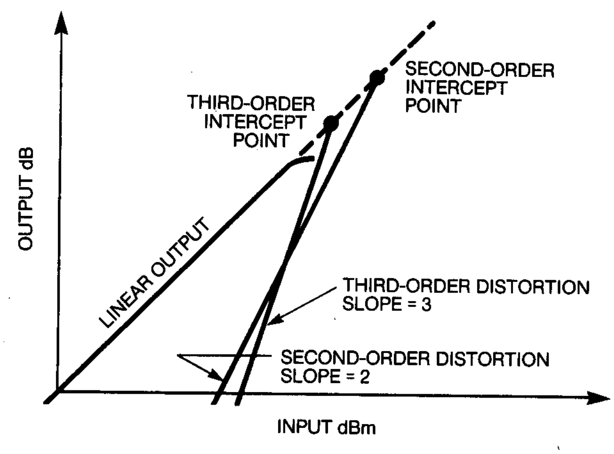


Figure 1. Receiver distortion vs. input power intercept point extrapolation (theoretical).

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 convenience, 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 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.

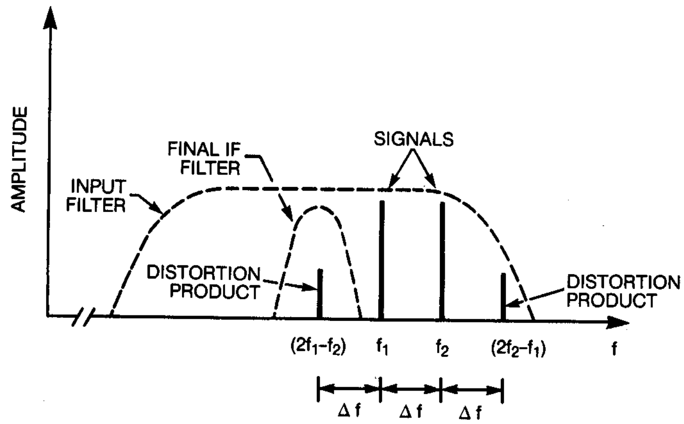


Figure 2. Third-order distortion products from two signals inside the receiver input filter.

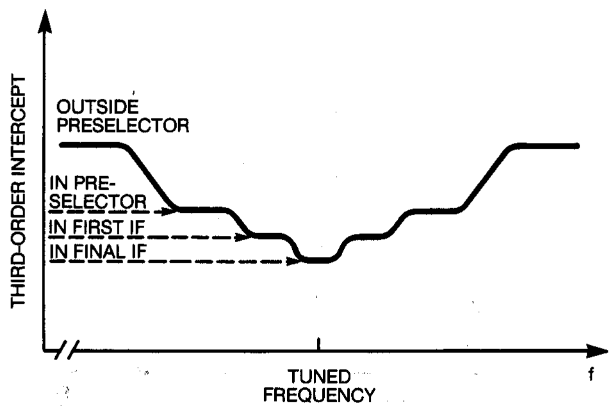


Figure 3. Third-order intercept as a function of test-tone frequency relative to receiver-tuned frequency.

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.

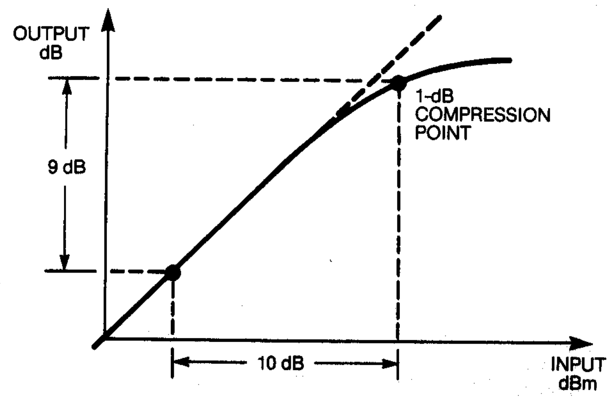


Figure 4. 1-dB compression point.

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 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.

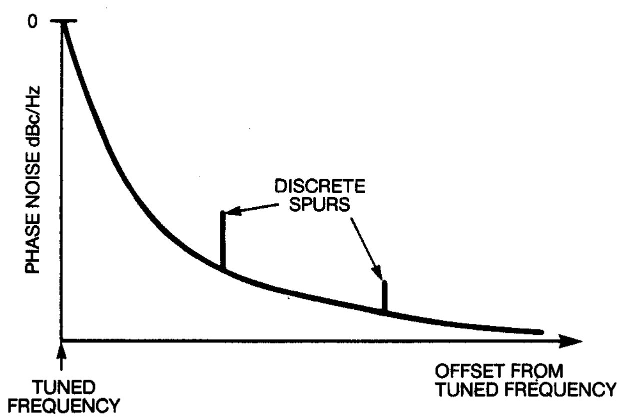


Figure 5. Receiver phase noise.

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. 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 meaningful 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, dBμV 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, distortion 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.

References

The following list of books and articles represents a sampling of the more readable literature relating to the subject of this paper.

1. Grebenkemper, C. John, "local Oscillator Phase Noise and Its Effect on Receiver Performance," *Tech-Notes,* Vol. 8 No. 6, November/December 1981.

2. Dexter, Charles E., "digitally Controlled VHF/UHF Receiver Design," *Tech-Notes,* Vol. 7 No. 3, June 1980.

3. McDowell, Rodney K., "High Dynamic Range Receiver Parameters," *Tech-Notes,*

Vol. 7 No. 2, March/April 1980.

4. Erst, Stephen J., *Receiving Systems Design,* Artech House, 1984.

5. Hsieh, Chi, “GaAs FET IMD Demands Better Standard,” *Microwaves,* Vol. 21 No. 6, June 1982.

6. Hirsh, Ronald B., “Knowing the Meaning of Signal-To-Noise Ratio,” *Microwaves and RE,* Vol. 23 No. 2, February 1984.

7. Fisk, James R., “Receiver Noise Figure Sensitivity and Dynamic Range - What the Numbers Mean,” *Ham Radio,* October 1975.

8. Schwartz, Mischa, *Information Transmission, Modulation and Noise,* McGraw-Hill, 1980.

Watkins-Johnson Company
Tech-notes

Vol. 14 No. 2 March/April 1987

Receiver Dynamic Range: Part 2

Robert E. Watson

Part 1 of this article reviews receiver measurements which, taken as a group, describe receiver dynamic range. Part 2 introduces comprehensive measurements that attempt to characterize a receiver's dynamic range as a single number.

COMPREHENSIVE MEASUREMENTS

The following receiver measurements and specifications attempt to define overall receiver dynamic range as a single number which can be used both to predict overall receiver performance and as a figure of merit to compare competing receivers. They include: 1-dB compression dynamic range, desensitization dynamic range, spur-free dynamic range, and NPR (noise-power ratio) figure of merit (NPRFOM). In general, they are based on the primary measurements of receiver performance, but the NPRFOM test attempts to simulate the actual signal environment in a way that combines all of the receiver dynamic range characteristics (see Table 1). This test is proposed as a practical and realistic measurement of receiver dynamic range.

100 kHz	10 MHz
200 kHz	20 MHz
500 kHz	50 MHz
1 MHz*	100 MHz*
2 MHz	200 MHz
5 MHz*	500 MHz*
	1 GHz
*Recommended minimum set.	

Table 1. Recommended standard filter frequencies for NPRFOM measurements.

1-dB Compression Dynamic Range

The receiver 1-dB compression dynamic range defines the range of signal levels that the receiver can process acceptably. In simplest terms, it is the difference in dB between the inband 1-dB compression point and the minimum-receivable signal level. The compression point is obvious enough; however, the minimum-receivable signal must be identified.

There are a number of candidates for minimum-receivable signal level, including: "minimum-discernable signal" (MDS), tangential sensitivity, 10-dB SNR, and receiver noise floor. Both MDS and tangential sensitivity are based on subjective judgments of signal strength, which differ significantly from author to author. They are mentioned here because of their historical significance, but the uncertainty limits their value as a part of receiver dynamic-range specifications. A more repeatable measurement is 10-dB SNR; but this, too, has disadvantages because of the variations of SNR due to type and percentage of modulation. The least ambiguous indicator of minimum receivable signal is probably receiver noise floor. This can be defined in two ways: noise floor in a 1-Hz bandwidth and total equivalent input noise power in the narrowest receiver bandwidth. The first is simply -174 dBm plus the receiver noise figure in dB; while the second has the additional factor of 10 times the log of the receiver bandwidth. For most purposes, the inclusion of the receiver bandwidth yields a better estimator of usable dynamic range. Using this definition, receiver dynamic range can be expressed as:

$$CDR = P_{ic} + 174 \text{ dBm} - 10 \log BW - NF$$

where:
CDR is the compression dynamic range in dB
P_{ic} is the 1-dB input compression power in dBm
BW is the narrowest receiver bandwidth in Hz
NF is the receiver noise figure in dB

This dynamic range definition has the advantage of being relatively easy to measure without ambiguity but, unfortunately, it assumes that the receiver has only a single signal at its input and that the signal is desired. For deep-space receivers, this may be a reasonable assumption, but the terrestrial sphere is not usually so benign. For specification of general-purpose receivers, some interfering signals must be assumed, and this is what the other definitions of receiver dynamic range do.

Desensitization Dynamic Range

Desensitization dynamic range (DDR) measures the receiver degradation effects due to a single, dominant, out-of-band interferer. In many “real world” signal environments, a single, strong signal may be the major source of interference due to the effects of receiver phase noise and out-of-band signal compression. In this test, a signal that produces an output SNR of 10 dB is injected at the receiver input. An interfering sinusoid is added to the input at a particular frequency offset from the tuned frequency and its magnitude is increased until the output SNR degrades 1 dB. The DDR is then the power ratio (in dB) of the undesired signal power (in dBm) to the receiver noise floor in dBm per Hertz. The DDR can be calculated using the equation:

$$DDR = Pi - NF + 174$$

where:
DDR is the desensitization dynamic range in dB
Pi is the interfering signal power in dBm
NF is the receiver noise figure

DDR is a true measure of dynamic range because it includes both noise figure and measurement of overload/ interfering signal power. The use of receiver input attenuation will improve large signal-handling capability, but noise figure will be degraded commensurately. The DDR, however, is not affected by input attenuators. When it is desirable to determine the absolute signal power in dBm required to cause desensitization for a particular receiver configuration, the following equation can be used:

$$Pi = DDR + NF - 174$$

Note that the noise figure must include the effects of input attenuation as the receiver is intended to be used.

DDR is strongly affected by the frequency offset of the interfering signal. At small frequency offsets, the DDR is dominated by the effects of receiver phase noise reciprocal mixing. In this region, the DDR is approximately 6 dB less than the magnitude of the single-sided phase noise spectrum in dB per Hertz below the “carrier” (dBc). For example, if a receiver’s phase noise at 100 kHz from the tuned (carrier) frequency is -130 dBc, the DDR at 100 kHz offset will be about 124 dB. In general, receiver phase noise improves with frequency offset so that in some receivers, interfering signals well removed from the tuned frequency, will begin to cause signal compression before the effects of phase noise reciprocal mixing are observed. In this case, the DDR will be worse than 6 dB less than the magnitude of the phase-noise suppression at these frequencies. Because of these frequency effects, it is necessary to specify the DDR at several different offset frequencies. The best presentation of this data would be in the form of a graph, as shown in Figure 1.

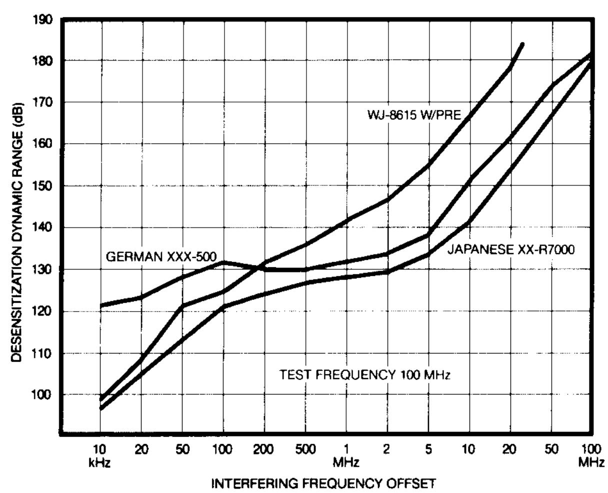


Figure 1. Desensitization dynamic range (DDR) as a function of frequency offset of the interfering signal.

This figure compares the DDR of three popular vhf/uhf receivers. At small frequency offsets, the DDR is typically dominated by receiver phase noise. At larger frequency offsets, in receivers with modest signal input filtering (rf preselection) 1 dB compression due to signal overload may occur. This is the effect seen in both the XX-R7000 and

XXX-500 receivers.

The significance of DDR is somewhat dependent on signal environment. If the interfering signals have significant phase noise of their own, it is only necessary for the receiver's phase noise to be better than the interferer's. Most radio transmitters have significant amounts of phase-noise sideband energy at modest offsets from the carrier frequency. This is especially true for variable frequency oscillator (vfo) and most frequency-synthesized frequency sources. A notable exception, which may have very low levels of small offset phase noise, is crystal oscillator signal sources. At large frequency offsets, many transmitters will have low phase noise because of the filtering properties of tuned power output stages and narrow antenna bandwidths. For this reason, more attention should be given to obtaining a good DDR at large frequency offsets.

A test setup for DDR measurement is shown in Figure 2. The receiver is tuned to the test frequency and set for maximum gain in the narrowest available bandwidth with a bfo detection mode. In some receivers, it will be necessary to use the ssb mode to activate the bfo and to achieve narrow bandwidth. The receiver is tuned to center the test signal in the IF passband and to produce an audio "beat note" of 1 kHz.

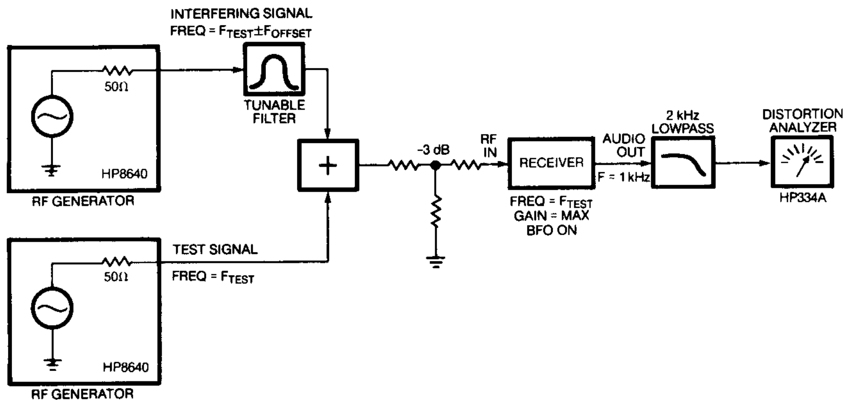


Figure 2. Desensitization dynamic-range test configuration.

If the receiver does not have a "predetection" demodulation mode like "cw" or ssb which uses a bfo frequency conversion to audio, the narrowest available IF output may be used with a spectrum analyzer. In this case, the signal is monitored for a 1-dB amplitude decrease due to compression, and the noise floor is monitored for a 1-dB increase due to phase-noise reciprocal mixing.

The interfering signal generator must have phase noise much better than that of the receiver under test. The tunable bandpass filter will help eliminate any residual generator phase noise at large frequency offsets. The audio lowpass filter is not required, but serves to minimize the effects of variations in audio response from receiver-to-receiver.

Spur-free Dynamic Range

Spur-free dynamic range (SFDR), as generally used, attempts to define receiver dynamic range in terms of two undesired interferers and the receiver noise floor. As with the 1-dB compression dynamic range, it is based on a mathematical manipulation of the primary measurements of receiver range. In this case, the spur-free dynamic range is the difference in dB between the receiver noise floor and the level of each of two equal-amplitude, out-of-band interfering tones that produce an in-band spurious product equal in power to the noise floor. Generally, the receiver third-order intercept point is used to predict the spurious product, but often the second-order distortion dominates. In any case, the SFDR can easily be expressed as:

$$SFDR3 = 2/3\ (IIP3 + 174 - NF - 10 \log BW)$$

or

$$SFDR2 = 1/2\ (IIP2 + 174 - NF - 10 \log BW)$$

where:

SFDR3	is the third-order spur-free dynamic range in dB
IIP3	is the receiver third-order input intercept point in dBm
NF	is the receiver noise figure in dB
BW	is the narrowest receiver bandwidth in Hz
SFDR2	is the second-order spur-free dynamic range in dB
IIP2	is the receiver second-order input intercept point in dBm

Spur-free dynamic range has become a very popular specification because it seems to give a single number which can be used to compare the overall dynamic-range performance of competing receivers. Unfortunately, the SFDR specification overlooks several important factors which influence dynamic range. First, it attempts to model interference by using just two interfering signals. This overcomes some of the objections to single-tone testing, but the real signal environment is usually populated by a multitude of signals. Second, it does not reveal the effects of reciprocal mixing or compression like the desensitization dynamic-range test. Third, it does not effectively test the effects of receiver input filtering (preselection). Finally, SFDR, as it is ordinarily specified, considers only the third-order distortion. In fact, for many receivers, especially those with modest input filters, the second-order products may dominate. For example, for a receiver with a bandwidth of 100 Hz, a noise figure of 10 dB, a third-order intercept of +20 dBm, and a second-order intercept of +50 dBm, the second- and third-order SFDRs will be 97 dB and 109.3 dB, respectively. Surely, the lesser of the two values is more valid; however, it is often not specified.

Because spur-free dynamic range is derived from the primary measurements, it would seem to provide no new information. Instead, it merely adds to the confusion of receiver dynamic range specifications. However, there is a continuing need — or at least a desire — for a truly comprehensive measurement of receiver dynamic range. A possible candidate is the NPR figure-of-merit (NPRFOM).

NPR Figure of Merit

The NPR figure-of-merit dynamic range measurement attempts to overcome some of the shortcomings of earlier measurements by better simulating the signal environment with high-power white noise.

NPR is an abbreviation for Noise Power Ratio, a term familiar to those involved in FDM telephone work. NPR testing simulates the signal environment by a broad band of "white noise with the total noise power adjusted to equal the total signal power that can be expected at the receiver input. The noise is removed at the receiver's tuned frequency by a notch filter. Due to receiver distortion, this notch tends to fill with intermodulation products. The apparent notch depth, as seen by the receiver, is the noise power ratio; that is, the ratio of out-of-notch noise to in-notch noise measured in dB (see Figure 3). This test has been used for many years in FDM telephone measurements and is specified by the CCIR and CCITT.

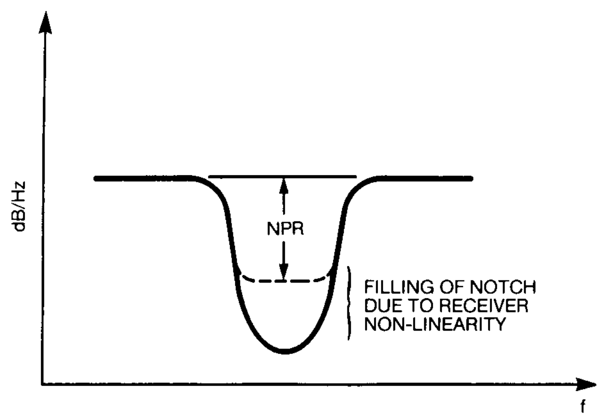


Figure 3. Noise power ratio (NPR).

NPRFOM is defined as the sum of the input noise spectral density which produces a receiver NPR of 40 dB plus 174 dB minus the receiver noise figure. For this measurement, noise figure is measured at the manual gain setting, which produces nominal receiver output when tuned to the notch frequency. NPRFOM can be expressed as:

$$NPRFOM = P_{npr} + 174 - NF$$

where:
NPRFOM is the Noise Power Ratio Figure of Merit in dB
P_{npr} is the input power spectral density in dBm/Hz for the white noise that produces a 40 dB NPR
NF is the receiver noise figure in dB measured at the receiver-gain setting which produces the nominal receiver output in the narrowest available predetection output

The 40-dB value for NPR was chosen because it is somewhat inconvenient to obtain noise sources with notch depths significantly greater than -50 dB.

Noise figure is included in the NPRFOM to give a relative indication of receiver dynamic range. It is measured with receiver gain set for a nominal output when tuned to the noise notch because this eliminates the effects of receiver input attenuation. Adding attenuation to the input of a receiver linearity increases its power-handling capability, but adding attenuation also linearly increases the receiver noise figure. Consequently, there is no net change in receiver dynamic range due to input attenuation. The factor of 174 dB normalizes the measurement to the theoretical noise floor of -174 dBm/Hz. The NPRFOM measurement produces a single measurement of dynamic range which can be used to directly compare the effective performance of competing receivers. This comparison is not affected by the nonideality of third-order intercept extrapolations, or by the vagaries of preselector specifications, but instead highlights the differences between receivers that appear in operation, but not on a manufacturer’s data sheet.

A basic NPRFOM test setup is shown in Figure 4. The test is performed using the following step-by-step procedure (Steps 1,2 and 3 preset the receiver and the noise level):

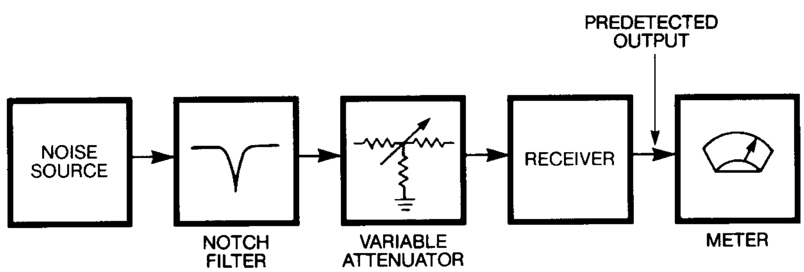


Figure 4. NPRFOM test configuration.

1. Set the receiver for the minimum available bandwidth and a predetection mode (CW or SSB). Monitor the audio output level with the meter. If these predetection modes are not available, monitor the narrowband IF output. Set the audio gain for mid-range and tune the receiver to a frequency outside of the notch.

2. Set the attenuator for maximum and temporarily disconnect the noise source. Increase the receiver RF/IF gain until the meter indicates -3 dB with respect to the nominal output, or until the gain is maximum. *(Nominal output* is the output produced by a strong signal in agc mode, or the output specified by the manufacturer.)

3. Reconnect the noise source and decrease the attenuation until the nominal output level is indicated on the meter.

Now that the receiver and test setup have been preset, Steps 4, 5 and 6 are repeated until the noise level at the receiver input produces a 40-dB NPR.

4. Tune to the notch frequency and decrease the attenuation by 40 dB. If the output is less than the nominal level, go to Step 5. If the output is equal to the nominal value, go to Step 7. If the output is greater than the nominal value, the receiver has insufficient dynamic range to achieve an NPR of 40 dB.

5. Decrease the attenuation until the output is at the nominal level.

6. Tune the receiver to a frequency outside of the notch and increase the attenuation by 40 dB. Adjust the receiver manual gain for nominal output and return to Step 4.

Steps 7 and 8 measure the noise level and receiver noise figure which are then used to calculate NPRFOM.

7. Without changing the receiver settings, disconnect the signal from the receiver input and measure the out-of-notch noise spectral density at the attenuator output with a spectrum analyzer. Enter this value into the NPRFOM equation.

8. Without changing the receiver settings, measure the receiver noise figure. Enter this value into the NPRFOM equation and calculate NPRFOM.

The measured values for NPRFOM are somewhat dependent on the test configuration. While the relative ranking of several receivers is generally unaffected by the test configuration, direct comparisons of NPRFOM values are possible only when the test conditions are standardized. In particular, the characteristics of the notch filter and noise source must be specified.

The noise source must be spectrally “flat” or “white.” A desirable limit of noise amplitude spectrum flatness is 1 dB. The band-of-noise frequencies, in the most desirable case, would cover the range from less than one-third the receiver’s lowest tunable frequency to greater than twice the highest tunable frequency. For wide-tuning receivers, this range may be difficult to obtain. A more reasonable noise frequency range would be from one-third the test frequency to two times the test frequency. For some receiver frequencies, noise generator limitations make an even more restricted noise bandwidth necessary. In this case, a noise band covering the range of ±25% of the tuned frequency is recommended.

The notch filter is the most critical component of the test configuration. The ideal notch filter would have a -3 dB bandwidth only slightly wider than the receiver final IF bandwidth, and a -50 dB bandwidth equal to the final IF bandwidth. Such a narrow notch would allow the noise to simulate a uniformly dense signal environment. This would also maximize the test’s ability to discriminate between the performances of differing receivers. This is because a narrow notch maximizes the stress on the receiver’s input filter and first IF stages. Unfortunately, the component Q’s for such a filter would make it difficult or impossible to realize. It is more practical to use a set of standardized filters similar to those specified by the CCITIT for FDM NPR measurements. These filters typically have -3 dB bandwidths, which are a constant percentage of their center frequencies. The constant percent bandwidth characteristic tends to produce NPRFOM values which do not vary greatly with test frequency. Also, this type of filter can be realized with limited values of component Q. Figure 5 gives recommended bandwidth values for both LC filters and crystal filters. Table 1 lists recommended center frequencies. Where feasible, the narrower crystal filters are preferred for the reasons stated above. These filters are relatively simple, but a full set is somewhat expensive. Lower-cost notch filters are shown in Figures 6 and 7.

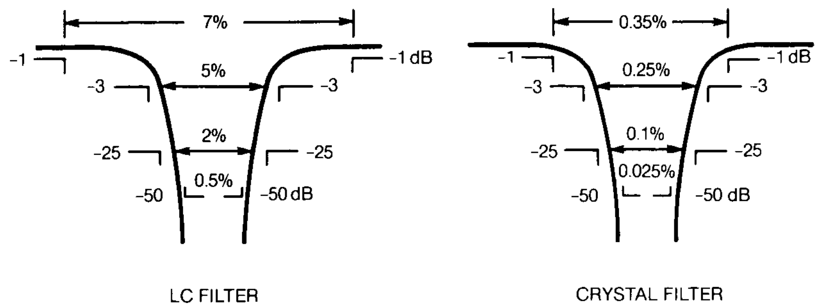
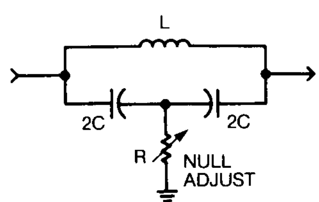


Figure 5. Recommended notch-filter bandwidths for NPRFOM measurements.

The LC filter of Figure 6 is most useful for frequencies below 50 MHz. It is relatively easy to construct and requires components of only modest Q. The capacitor values are specified to produce constant percentage bandwidth of approximately 6%. The inductor should have a Q greater than 50 and the capacitors should be matched within 5%.



$$C = \left(\frac{16.6\,Q_L}{Q_L - 16.6}\right) \frac{1}{200\,\pi f}$$

$$L = \frac{1}{(2\,\pi f)^2\,C}$$

$$R = \frac{3}{2\,Q_L}$$

Figure 6. Notch filter with 6%, -3 dB bandwidth for 50-ohm source and load.

The delay-line filter of Figure 7 is most useful for frequencies from 50 to 1000 MHz. Its major advantage is repeatability, but a major disadvantage is the production of multiple notches (Figure 8). However, the power loss due to the extra notches is only about 2.5 dB, and this small amount can be subtracted from the resulting NPRFOM measurement without causing a major error. The ninth notch in the series (counting from zero frequency), was selected for this test because it is a reasonable compromise considering notch -3 dB bandwidth, -50 dB bandwidth, and cable loss. Higher notch numbers have narrower -3 dB bandwidths, but the -50 dB bandwidth becomes very narrow and the increased cable length causes excessive frequency loss "tilt." For the ninth notch, the -3 dB bandwidth is approximately 6.06% of the center frequency, and the -50 dB bandwidth is approximately 0.024% (see Figure 9). The out-of-notch frequency for measuring noise spectral density and for tuning during the NPRFOM test is at plus or minus 6% from the notch frequency. These frequencies correspond to the nearest filter transfer maxima.

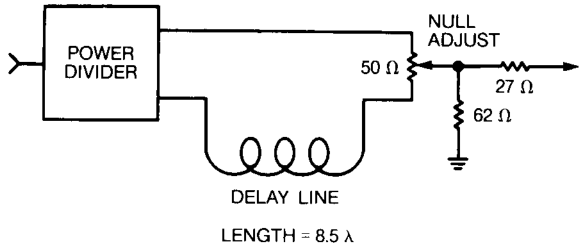


Figure 7.Delay-line notch filter.

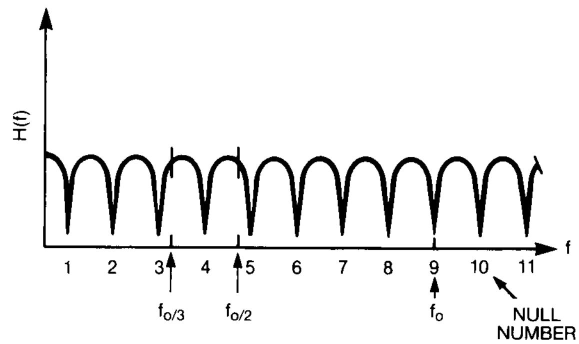


Figure 8. Frequency response of delay-line filter.

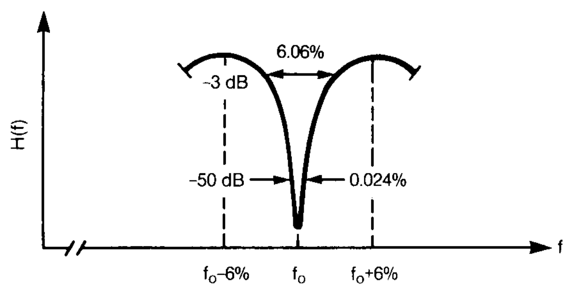


Figure 9. Detail of delay-line filter frequency response for ninth null.

The effectiveness of NPRFOM in differentiating the dynamic range performance can be seen in Table 2. Several communications quality vhf/uhf receivers were tested for NPRFOM using filters similar to those in Figure 6. The differences in performance can be readily explained by the differences in receiver design. Receiver A is a proven, older design which was optimized for low noise figure. The first stage of preselection is low loss and rather broad. Also the tube-technology input amplifier does not have the signal-handling capability of more modern high dynamic-range designs. Receiver B is a compact vhf/uhf receiver, without the preselector option, is overloaded by the broad-band input noise because of both second- and third-order intermodulation effects. Receiver C has a modest tracking preselector, but its poor third-order performance allows only a slight improvement in NPRFOM. Receiver D performs much better due to its internal switched suboctave preselector and very good third-order performance. The Receiver E performs slightly better than D receiver because it has a tracking preselector which is slightly narrower than a switched suboctave preselector. Receiver F, with an external suboctave filter, has a NPRFOM similar to E because it has a significantly better noise figure. Receiver G has significantly improved dynamic range because of the relatively narrow bandwidth of the tracking preselector. Finally, receiver H, with the tracking preselector option, performs best of all. This is because the preselector eliminates most of the broadband noise of the test signal. In addition, because the preselector is integral to the receiver, the entire signal path has been optimized for maximum dynamic range.

Receiver	Configuration	Noise PWA dim/Hz For 40 dB NPR		Noise Figure at Test Gain		NPRFOM (dB)	
		@39 MHz	@111 MHz	@39 MHz	@111 MHz	@39 MHz	@111 MHz
A	Normal	-118	-114	5	5.5	51	54.5
B	No Preselector	-99	-104	10	10	65	60
C	Normal	-95	-98.5	13	13	66	62.5
D	Normal	-88	-82	8.5	9	77.5	77
E	Normal	-83	-86	10	10	81	78
F	External Suboctave Preselector	-86	–	7	–	81	–
G	Tracking Preselector	-82	-82	10	10	82	81
H	Internal Tracking Preselector	-77	-79	10	10	87	85

Table 2. NPRFOM measurements for several communications receivers.

Conclusion

Satisfactory assessment of receiver dynamic range requires careful measurement of several parameters. The primary receiver measurements of noise figure, second-order intercept, third-order intercept, 1-dB compression, phase noise and internal spurious signals can provide information for determining receiver dynamic range. Secondary receiver measurements, such as sensitivity, cross modulation, intermodulation distortion, and reciprocal mix, can provide supplemental information, but they should not be substituted for the primary measurements. Comprehensive measurements of receiver dynamic range include noise-power ratio figure-of-merit and desensitization dynamic range. Together, these two measurements and the noise figure give an excellent indication of receiver dynamic range.

References

The following list of books and articles represents a sampling of the more readable literature relating to the subject of this paper.

1. Grebenkemper, C. John, "Local Oscillator Phase Noise and Its Effect on Receiver Performance," *Tech-Notes,* Vol. 8 No. 6, November/December 1981.

2. Dexter, Charles E., “Digitally Controlled VHF/UHF Receiver Design,” *Tech-Notes,* Vol. 7 No. 3, June 1980.

3. McDowell, Rodney K., “High Dynamic Range Receiver Parameters,” *Tech-Notes,* Vol. 7 No. 2, March/April 1980.

4. Erst, Stephen J., *Receiving Systems Design,* Artech House, 1984.

5. Hsieh, Chi, “GaAs FET IMD Demands Better Standard,” *Microwaves,* Vol. 21 No. 6, June 1982.

6. Hirsh, Ronald B., “Knowing the Meaning of Signal-To-Noise Ratio” *Microwaves and RF,* Vol. 23 No. 2, February 1984

7. Fisk, James R., ‘Receiver Noise Figure Sensitivity and Dynamic Range — What the Numbers Mean,” *Ham Radio,* October 1975.

8. Schwartz, Mischa, *Information Transmission, Modulation and Noise,* McGraw-Hill, 1980.