Receiver noise figure sensitivity and dynamic range - what the numbers mean

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A complete discussion of receiver sensitivity, intermoduation distortion, and gain compression, and what they mean in terms of performance

When it came to receivers, the earliest amateur operators were concerned primarily with sensitivity and experimented almost endlessly with different types of crystals, trying to find the one that was the most sensitive. Then came DeForest's Audion, and Armstrong's regenerative detector, and amateurs who could afford the tubes found they had all the sensitivity they could use. However, as the hobby grew, and more and more amateurs started populating the band below 200 kHz, the simple regenerative detector simply wasn't up to the task. Selectivity, with simple tuned input circuits, was practically nonexistent, and the regenerative detector hopelessly overloaded in the presence of strong signals.

In the early 1920s amateurs worked to improve their tuners, but even the so-called "Low-Loss" tuners were only marginally acceptable. Although several superheterodyne designs were described in the amateur magazines, it wasn't until low-cost, commercial i-f transformers became available in the late twenties that the superhet saw widespread amateur use. Selectivity against interfering signals was still a problem, however, and James Lamb revolutionized receiver design in 1932[1] with his "single-signal" CW circuit which used an i-f stage with extremely high selectivity - provided by regeneration or a simple crystal filter.

The single-signal, single-conversion superhet of the late 1930s suffered from poor RF image response at the higher frequencies, but it wasn't too severe on 14 MHz and few amateur receivers of the day, in fact, tuned much above 18 or 20 MHz (15 meters was not yet assigned to amateur use and most 10-meter operators used specialized receivers or converters). When the 10- and 15-meter bands opened up after the war, however, the poor RF image response of the single - conversion superhet with a 455-kHz i-f had to be faced - it was solved by going to a double-conversion layout with a first conversion to 2 or 3 MHz to minimize RF image response, and a second conversion to 455 kHz or lower for adjacent channel selectivity.

Although amateur radiotelephone operation in the 1930s was relatively limited, the huge growth of a-m activity after the war demanded improved adjacent-channel phone selectivity. While the crystal filter provided excellent selectivity for CW operation, it was of little or no use on a-m or ssb and some phone operators started using a Q5er an outboard 80-kHz i-f strip - for improved phone selectivity. This led to the triple-conversion superhets which were the rage of the 1950s.

As pointed out by Goodman[2], however, the multiple-conversion design had many shortcomings, including high-selectivity i-f which made it practically impossible the large number of stages between the antenna and the to attenuate strong, adjacent signals.

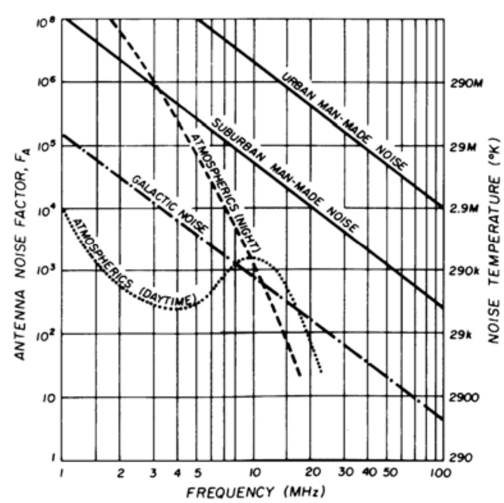

fig. 1. Receiver sensitivity is limited by the external available noise power which varies with frequency. For quiet, rural locations galactic noise is the limiting factor down to about 18 MHz, and atmospheric noise dominates below 18 MHz.

And, with at least three oscillators running at the same time, it was difficult to avoid the many spurious signals which were generated within the system. He advocated a return to the single-conversion superhet using highly-selective, high-frequency, crystal lattice filters which were just then becoming commercially available.

The 1960s saw a return to single conversion designs, the use of high frequency, crystal-lattice filters and the widespread use of semiconductors. With modern devices receiver sensitivity was no longer limited by the RF amplifier (or mixer) stage, but by the external galactic and man-made noise. Cross modulation and overload, on the other hand, were becoming a serious problem as more and more amateurs started using high-power Ii nears and large, directive antennas.

Modern communications receivers, therefore, in addition to meeting stringent frequency accuracy, stability, sensitivity, and selectivity requirements, must provide freedom from cross modulation, intermodulation distortion and blocking. Some modern, solid-state solutions to these design goals were discussed recently by Rohde[3].

The specifications for a typical modern communications receiver might list sensitivity of 0.5 µV for 10 dB signal plus-noise-to-noise (S+N/N) ratio, intermodulation distortion of -65 dB, "wide" dynamic range, and "virtual elimination" of overload from adjacent signals.

However, is 0.5 µV sensitivity for 10 dB S+N/N adequate for operation on the high-frequency bands? For satellite communications on 10 meters? What is - 65 dB intermodulation distortion in terms of signal strength? "Wide dynamic range" and "virtual elimination of overload" are obviously advertising superlatives without definition but what, exactly, can you expect from a high quality, modern receiver design? Perhaps, if these performance data were defined, and amateurs understood what they meant, manufacturers would be encouraged to use no-nonsense numerical data. Only then can amateurs compare the dynamic range and cross-modulation performance of one receiver against that of another.

sensitivity

The minimum usable signal or sensitivity of a receiver is determined by the noise in the receiver output. This can be noise generated within the receiver, thermal noise generated by losses in the transmission line, or atmospheric, manmade or galactic noise picked up by the antenna. As shown in fig. 1, external noise sources are likely to be the limiting factor up to 100 MHz or so.[4] In urban areas man-made noise predominates and measurements indicate the average level of man-made noise in suburban areas is about 16 dB lower. In a quiet, rural location which has been chosen with care the man-made noise may be near the galactic noise level, but few amateurs are so fortunate.

Atmospheric noise usually predominates in quiet locations at frequencies below about 20 MHz and is produced by lightning discharges so the level depends upon a number of variables including frequency, weather, time of day, season and geographical location. This type of noise is particularly severe during rainy seasons near the equator and generally decreases at the higher latitudes. More complete data on high-frequency atmospheric noise is given in reference 5.

Galactic or cosmic noise is defined as RF noise caused by disturbances which originate outside the earth or its atmosphere. The primary causes of this noise, which extends from 15 MHz well into the microwave region, are the sun and a large number of noise sources distributed chiefly along the Milky Way. Solar noise can vary as much as 40 dB from "quiet" sun levels (low sunspot activity) to periods of "disturbed" sun (high sunspot activity).

Galactic noise from the center of the Milky Way is about 10 dB below the noise from a "disturbed" sun, whereas noise level s from other parts of the galaxy can be as much as 20 dB lower. This is important in satellite communications and will be discussed later.

thermal noise

The free electrons in any conductor are in continuous motion - motion that is completely random and is the result of thermal agitation. The effect of this electron motion is to cause minute voltages which vary in a random manner to be developed across the terminals of the conductor.

$$e^2 = 4kTBR \qquad (1)$$

where e2 = mean square noise voltage
k = Boltzmann's constant = 1.38 X 10-[23] joules/o K
T = absolute temperature, ° K
B = bandwidth, Hz
R = resistance, ohms

Since this phenomenon was first demonstrated by J. B. Johnson in 1928[6], thermal noise is sometimes known as Johnson noise. At the same time, H. Nyquist showed, on the basis of the statistical theory of thermodynamics, that the mean square noise voltage generated in any resistance can be expressed as [7]

Note that the noise voltage is dependent upon the bandwidth across which it is measured. This implies that noise is evenly distributed across all frequencies which, for all practical purposes, it is. *

*** At extremely high frequencies statistical mechanics is no longer valid, and eq. 1 must be revised on the basis of quantum theory. This equation is valid, however, to at least 6000 GHz.[8]**

Although noise bandwidth is not precisely the same as the 3-dB bandwidth of a receiver, in modern receivers with high skirt selectivity the 3-d B bandwidth can be used in **eq. 1** with little error.

The equivalent circuit of any impedance as a source of noise voltage is shown in **fig. 2A**. Note that the thermal noise voltage is dependent only on the resistive component and is independent of any reactance in the circuit. As might be expected, maximum noise power is transferred from a thermal noise source when the load impedance presents a conjugate match to the source impedance . This is represented in fig. 28 where the load resistance, R_L, is equal to the source resistance Since R = R_L, the noise voltage developed across the load is e/2, and from Ohm's law:

$$P = \frac{E^2}{R} = \frac{(e/2)^2}{R} = \frac{e^2}{4R}\ watts \qquad (2)$$

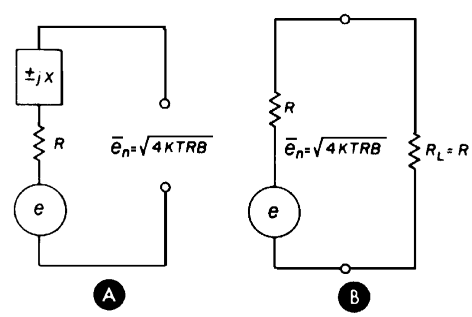

Fig. 2. Mean noise voltage depends on temperature, resistance and bandwidth, and is completely independent of reactance as shown in (A). Maximum noise power is transferred to the load when the load resistance is matched to the source resistance (B).

Substituting the value of e2 from eq. 1 into eq. 2, the power which can theoretically be transferred under such conditions is called the available noise power and is given by

$$P_n = kTB \qquad (3)$$

The factor of ***4R*** has cancelled out so the available noise power does not depend upon the value of the resistance. This is significant because it means that the available noise power of any resistor (or any noise source), if measured over the same bandwidth, can be represented by a resistor at temperature T. Thus, every noise source has an equivalent noise temperature.

The actual noise power dissipated in the load resistance may be affected by loss in the connecting leads, noise power generated in the load resistor itself, or a less than perfect match to the original resistance. This property is sometimes used in low-noise uhf amplifiers by creating a deliberate (but carefully determined) mismatch between the input termination and the detection device so that something less than the available termination noise power is coupled into the detector.

signal-to-noise ratio and receiver noise figure.

The relation of signal amplitude to noise is commonly referred to as the signal-to-noise (S/N) ratio. Unfortunately, this ratio has not been well standardized and is often used interchangeably to mean the ratio of rms signal voltage to rms noise voltage, the ratio of peak signal voltage to peak noise voltage and, in pulse systems, the ratio of peak signal power to average noise power. Therefore, when discussing SIN ratio, it's important to determine exactly which ratio is being referred to.

Although the minimum discernible signal (MDS) that can be heard above the receiver noise level is sometimes used as an indication of receiver sensitivity, it is extremely subjective because it differs many dB from measurement to measurement, and from one operator to another (some experienced weak-signal operators can detect signals as much as 20 dB below the noise level while other operators may have difficulty discerning signals which are equal to the noise level).[9]

Receiver sensitivity has also been defined in terms of a signal-to-noise ratio of unity (signal equals noise)* or equivalent noise floor, but this is difficult to measure unless you have a calibrated signal generator and a spectrum analyzer.

***This is sometimes erroneously referred to as tangential sensitivity. Tangential sensitivity, however, corresponds to a signal-to-noise ratio of 6.25 and is about 8 dB higher.[10]**

Noise figure or noise factor, on the other hand, is less susceptible to measurement errors than sensitivity and, since its introduction in 1944 by Friis,[11] it has become the accepted figure of merit for receiver sensitivity. Noise figure, NF, is simply noise factor, F, expressed in dB.

$$NF = 10 \log F \text{ (dB)} \qquad (4)$$

The concept of noise factor allows the sensitivity of any amplifier to be compared to an ideal (lossless and noiseless) amplifier which has the same bandwidth and input termination. As far as noise is concerned, that part of a receiver between the antenna and the output of the i-f amplifier can be regarded as an amplifier. The fact that the mixer stage shifts the frequency of the noise does not change the situation – it merely causes the noise to lie in a different place in the spectrum from the input noise. The only exception is when the receiver has poor RF image rejection. I n this case the noise figure of the receiver is 3 dB worse than it would be if the same receiver had good RF image rejection because the image noise appears at the output along with noise associated with the desired received frequency. This effectively doubles the noise at the output of the i-f amplifier.*

The noise factor, F, of a receiver is defined as:

$$F = \frac{S/N\ (\text{ideal receiver})}{S/N\ (\text{practical receiver})} = \frac{S_i/N_i}{S_o/N_o} \qquad (5)$$

S_i = available signal input power
N_i = available noise input power
S_o = available signal output power
N_o = available noise output power

Using this definition, it can be seen that an ideal receiver adds no noise to a signal so its output signal-to-noise ratio is the same as that at the input and the noise factor, $F = 1$.

***The noise figure is always defined at the input of the final detector(i-f output) because the noise output of a detector (but not of a mixer) is affected by the presence of a signal. An fm signal, for example, will suppress weak noise but will be suppressed itself by strong noise.**

Since the available noise input power, N_i is defined as $kToB$ in **eq. 3**, and the power gain of the system $G = S_o/S_i$, eq. 5 can be rewritten as

$$F = \frac{N_o}{GkT_oB} \quad \textbf{(6)}$$

Where T_o is 290^0 kelvin (IEEE definition). With the receiver noise factor defined in terms of noise output power, N_o, power gain, G, and noise input power, kT_oB, noise factor can be easily correlated to receiver sensitivity. Consider the case where the output signal to-noise ratio, $\boldsymbol{S_o / N_o}$**,** is unity.

$$\boldsymbol{S_o = N_o}$$

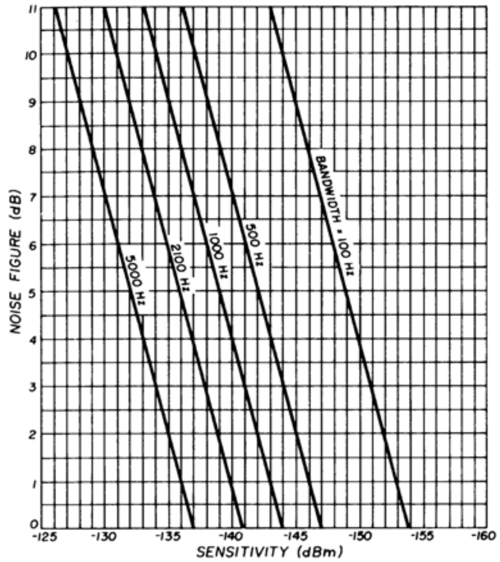

Fig. 3. Receiver sensitivity (-dBm) vs receiver noise figure and bandwidth for unity signal. to-noise ratio (signal equals noise). Add 10 dBm for a SIN ratio of 10 dB. Table 1 lists microvolt sensitivity for 50- and 75-ohm systems in terms of dBm sensitivity.

For this specialized case, eq. 6 can be rewritten

$$F = \frac{S_i}{kT_oB} \quad \textbf{(7)}$$

When the temperature is 290^0 K and the bandwidth is in kHz, $kT_oB = 4 \times 10^{-15}$ mW per kHz. Rewriting **eq. 7** in terms of dBm (dB referenced to 1 mW)

$$\boldsymbol{S_i = 10 \log kT_0 + 10 \log B + NF = -144 + 10 \log B_{kHz} + NF \text{ (dBm)}} \quad \textbf{(8)}$$

This function is plotted graphically in **fig. 3** for bandwidths commonly found in amateur communications receivers . For example, assume a high-frequency receiver has an 8 dB noise figure at 14.2 MHz and a bandwidth of 2.1 kHz. From **eq. 8** or **fig. 3**, the noise floor of the receiver at 14.2 MHz is at about -133 dBm. An input signal of -123 dBm (10 dB greater) would be required for a 10 dB SIN ratio.***** To convert dBm to microvolts, recall that:

E **=** the square Root of ***RP*** **(9)**

where E is in volts, R is resistance in ohms and P is power in watts. Since -123 dBm is 5.01×10^{-16} watts, -123 dBm is equivalent to 0.16 µV *across* a 50-ohm input termination. However, for a matched signal source, as shown in **fig. 4**, where the source resistance, R_S, is equal to the load resistance, R_L the source voltage must be twice the voltage across R_L because of the voltage dividing effect of the two series resistors in the network. For a matched 50-ohm source, therefore, an input signal of -123 dBm requires a source voltage of 0.32 µV. A chart of dBm vs microvolts for matched 50- and 75-ohm systems is presented in **table 1**.

***This is the signal-to-noise ratio in ssb and CW reception. The SIN ratio of a-m and nbfm signals is somewhat less because a-m (and nbfm) detection use only the envelope as a useful output and the SIN ratio must be reduced by a factor which is related to percentage of modulation (or modulation index).**

Because of this two-to-one voltage dividing effect, you must be very careful when comparing the sensitivity of one receiver against that of another. An input of +119 dBm, for example, implies an input directly at the receiver terminals and is 0.25 µV rms across 50 ohms. Sensitivity, on the other hand, implies the use of a matched signal generator so sensitivity of 0.25 µV corresponds to an input of -125 dBm.

This is a 6 dB difference. Since most amateur receiver manufacturers tend to use sensitivity specifications, there is no advantage, but the difference must be considered when you calculate the receiver noise figure and dynamic range. I n this article inputs will be stated in dBm as this eliminates the 6 dB conversion factor - **table 1** can be used to convert to *sensitivity*.*

cascaded stages

A relatively simple equation for the noise factor of a receiving system, in terms of the individual stage gains and noise factors is

$$F_T = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} \qquad \textbf{(10)}$$

F_T = overall system noise factor
F_1 = noise factor of the first stage
F_2 = noise factor of the second stage
F_3 = noise factor of the third stage
G_1 = power gain of the first stage
G_2 = power gain of the second stage
G_3 = power gain of the third stage

µV dBm 50 ohms		µV 75 ohms		P.V µV dBm 50 ohms	75 ohms
-76	70.8	86.7	-111	1.26	1.54
-77	63.2	77.4	-112	1.12	1.38
-78	56.2	69.1	-113	1.00	1.23
- 79	50.2	61.5	-114	0.90	1.09
- 80	44.8	54.9	-115	0.80	0.97
- 81	39.8	48.7	-116	0.71	0.87
- 82	35.6	43.6	-117	0.63	0.77
- 83	31.6	38.7	-118	0.56	0.69
- 84	28.2	34.5	-119	0.50	0.62
- 85	25.2	30.9	-120	0.45	0.55
- 86	22.4	27.4	-121	0.40	0.49
- 87	20.0	24.5	-122	0.36	0.44
- 88	17.8	21.8	-123	0.32	0.39
- 89	15.8	19.4	-124	0.28	0.35
- 90	14.2	17.3	-125	0.25	0.31
- 91	12.6	15.4	-126	0.22	0.27
- 92	11.2	13.8	-127	0.20	0.25
- 93	10.0	12.3	-128	0.18	0.22
- 94	9.0	10.9	-129	0.16	0.19
- 95	8.0	9.7	-130	0.14	0.17
- 96	7.1	8.7	-131	0.13	0.15
- 97	6.3	7.7	-132	0.11	0.14
- 98	5.6	6.9	-133	0.10	0.12
- 99	5.0	6.2	-134	0.09	0.11
-100	4.5	5.5	-135	0.08	0.10
-101	4.0	4.9	-136	0.071	0.087
-102	3.6	4.4	-137	0.063	0.071
-103	3.2	3.9	-138	0.056	0.069
-104	2.8	3.5	-139	0.050	0.062
-105	2.5	3.1	-140	0.045	0.055
-106	2.2	2.7	-141	0.040	0.049
-107	2.0	2.5	-142	0.036	0.044
-108	1.8	2.2	-143	0.032	0.039
-109	1.6	1.9	-144	0.028	0.035
-110	1.4	1.7	-145	0.025	0.031

Table 1. Microvolt sensitivity vs dBm for *matched* 50- and 75-ohm receiving systems.

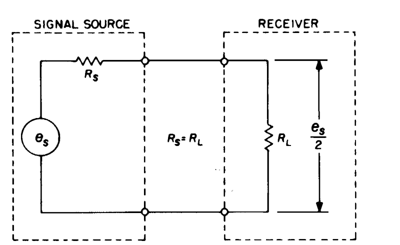

fig. 4. When a signal source is matched to a load, the voltage across the load is one-half the source voltage because of the voltage. dividing effect of the source and load resistors. When making receiver sensitivity measurements in the laboratory. a G-dB attenuator is placed in the line so sensitivity can be read directly from the signal generator's internal calibrated attenuator.

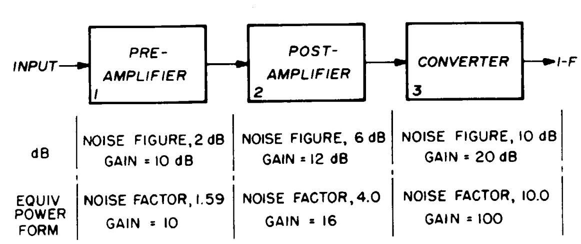

fig. 5. Noise factor of cascaded stages can be calculated by using equation 10 where all quantities are in power ratios. Noise factor at the input of the three cascaded stages shown here is 1.946 (noise figure = 2.9 dB).

Note that all terms are in power ratios. If the gain of the first stage is high, and the noise factor of the second stage is low, the overall system noise factor is determined primarily by the first stage and the third term of **eq. 10** may be dropped.

For example, consider the block diagram of the first three stages of a typical vhf receiving system shown in **fig. 5**. The overall system noise factor is:

$$F_T = 1.59 + \frac{4-1}{10} + \frac{10-1}{10 \cdot 16}$$

$$= 1.59 + 0.3 + 0.056 = 1.946$$

$$NF_T = 10 \log 1.946 = 2.9 \text{ dB}$$

For this receiver the overall system noise figure is 0.9 dB higher than that of the preamplifier alone. Raising the preamplifier gain to 13 dB or dropping the noise figure of the second stage to 4 dB (F = 2.5) would reduce the system noise figure to approximately 2.45 dB (*F* = 1.75). Depending on the frequency and the application, this may be a worthwhile improvement.

noise temperature

It is often convenient when working with very low noise uhf and microwave receivers to represent the noise figure of the receiver as an equivalent noise temperature. This is because the noise temperature of a receiving system varies over the range from 0^0 to 174^0 K as the noise factor varies from 1.0 to 1.6 (0 to 2 dB noise figure) and noise calculations using equivalent noise temperatures will provide better accuracy.

As mentioned above, the noise figure of an ideal receiver is 1, so the component of receiver noise figure which is due to internally generated noise, N_r, is $F - 1$. Therefore $N_r = (F - 1)\, GkT_oB$ **(11)**

The internally generated noise, N_r, can be represented by a noise power, GkT_rB, where T_r is the equivalent noise temperature. Substituting into **eq. 11**, this equivalent noise temperature can be expressed in terms of the reference noise temperature, T_o = 2900K. $T_r = (F - 1)\, T_o$ **(12)**

Eq. 12 can be easily rearranged to express noise factor as a function of the receiver noise temperature, T_r, and the reference noise temperature, T_o

$$F = 1 + \frac{T_r}{T_o} \qquad \textbf{(13)}$$

Although noise temperature is seldom used by amateurs, it is a more basic unit than noise factor and is actually easier to deal with, both in understanding concepts and making practical noise calculations. For a more complete discussion of noise temperature, see reference 12.

***The usual procedure for measuring receiver sensitivity is to place a 6 dB attenuator between the signal generator and the receiver. Receiver sensitivity can then be read directly from the signal generator's calibrated output attenuator.**

transmission lines

The transmission line’s contributions to receiver noise come from a common source: line losses. The first of these is the more obvious. When a signal travels down a lossy transmission line, the signal is attenuated. This reduces the signal-to-noise ratio and is equivalent to increasing the noise factor of the receiver. This increase can be calculated by introducing a loss factor, L, which is the loss of the cable expressed as a power ratio.

The second effect is due to the noise factor of the transmission line. The fact that the line has losses implies that there is a loss resistance associated with it (which is distinct from characteristic impedance). Since the line is warm it generates noise due to thermal agitation.

The noise factor of the line, F_t, is related to the loss factor, L, and the physical temperature of the line, T_t, by the following equation:

$$F_t = \frac{\left(\frac{1}{L} - 1\right) T_t}{290} + 1 \qquad \textbf{(14)}$$

The degradation of the receiver noise figure due to transmission line contributions may be calculated by considering the transmission line and the receiver as cascaded stages and using a form of **eq. 10**

$$F_{tr} = F_t + \frac{F_r - 1}{L} \qquad \textbf{(15)}$$

where F_{tr} = noise factor of the receiver and transmission line
F_r = noise factor of the receiver
F_t = noise factor of the transmission line
L = loss factor of the transmission line

For example, a receiver with a noise factor of 4 (NF = 6 dB) is used with a transmission line which has a loss factor of 0.63 (2 dB loss). The physical temperature of the line is 300^0 K (Ft = 1.61). What is the combined noise figure?

When line losses are low but receiver noise figures are 3 dB or greater, line loss is the predominate contributor to increased noise figure. When receiver noise figure is very low, the thermal effect predominates.

$$F_{tr} = 1.61 + \frac{4 - 1}{0.63} = 6.37$$

$$NF_{tr} = 8.04\ dB$$

antenna noise

Of all the contributions to system noise, antenna noise is probably the least understood. Assuming the antenna is built of good conducting materials, it contributes virtually no thermal noise of its own to the receiving system. The noise power the antenna does deliver to the receiver depends almost entirely on the temperature and other physical characteristics of the material lying in the antenna's field of view.

Table 2

frequency	noise factor at antenna	noise figure	external available noise power	receiver input signal for 10 µV dB S+N/N	acceptable noise figure
1.8 MHz	15.8	12.0	- 93 dBm	15.3µV	45 dB
3.5 MHz	16.2	12.1	-101 dBm	12.6µV	37 dB
7.0 MHz	16.7	12.2	-111 dBm	4.0µV	27 dB
14.0 MHz	17.6	12.5	-113dBm	3.1µV	24 dB
21.0 MHz	18.3	12.6	-118 dBm	1.8µV	20 dB
28.0 MHz	18.9	12.8	-123 dBm	1.0µV	15 dB
50.0 MHz	20.9	13.2	-129 dBm	0.5µV	9 dB
144.0 MHz	26.9	14.2	-139 dBm	0.2µV	2 dB

Table 2 Performance of a receiver with 0.5 µV sensitivity for 10 dB S+N/N with 100 feet (30.5m) of RG 8 A/U transmission line is shown in first two columns. Third column lists external available noise power for quiet receiving locations on each of the amateur bands. Fourth column shows receiver signal (50-ohms) required for 10 dB S+N/N on each of the bands (based on external noise). Last column lists acceptable noise figure for each of the bands (see text). Bandwidth = 2.1 kHz.

A 432-MHz moonbounce receiving antenna looking out into space, for example, may deliver only as much noise power as a resistor at 10° K (noise factor = ".03). If this same antenna is rotated so that the warm earth comes into its field of view, the antenna noise temperature would rise to about 300° K (noise factor = 2).

A ten-meter Oscar receiving antenna which is pointed into "cold" space, on the other hand, will see a noise temperature of about $10{,}000^0$K minimum (noise factor = 35). Pointed at the horizon, however, the antenna noise temperature may be ten or fifteen times higher, depending upon the amount of man-made noise.

There is little that can be done to improve the situation of a terrestrial radio circuit, but an antenna that looks at the sky, such as a satellite antenna, deserves careful design. This is because the effect of the earth is still present, and any sidelobe that sees the earth will pick up thermal noise. This is sometimes quite serious and sidelobes are of major concern in many deep-space communications and radio astronomy systems. Careful attention to antenna design with respect to sidelobes can provide antenna temperatures significantly under 50° K, while poor design can result in much higher values.

minimum usable sensitivity

With an understanding of receiver noise factor and its relationship to signal-to-noise ratio, it's now possible to determine the minimum usable sensitivity (MUS) of a receiving system, and how the performance of your own equipment affects your ability to receive weak signals. Let's first consider that modern communications receiver mentioned earlier which had a specified sensitivity of 0.5 /.LV for 10 dB S+N/N ratio.*

From table 1 a 0.5 µV sensitivity for 10 dB S+N/N is equivalent to a sensitivity of -119 dBm in a 50-ohm system (-129 dBm noise floor). Assuming a bandwidth of 2.1 kHz, the noise figure of the receiver is about 11.8 dB (noise factor = 15.1). Assuming 100 feet (30.5m) of RG-8A/U transmission line at 300^0K, the noise factor at the antenna terminals may be calculated from **eq. 15**, and is shown in **table 2** for the six high-frequency amateur bands, (calculated on the basis of 0.5 µV sensitivity for 10 dB S+N/N on all bands, which may be optimistic). Even at 28 MHz, where the line loss has increased the noise factor by 25 per cent, the system noise factor is still well below the available noise power seen by the antenna (see **fig. 1**).

At 50 MHz, however, the system is limited by receiver noise and a lower noise figure would be required for weak signal work (system noise at 50 MHz in this example is about 2.4 dB higher greater than external noise for a quiet location). If the receiver was connected directly to the antenna terminals to eliminate transmission line losses the system noise figure would be essentially that of the receiver alone and a 0.5 µV signal would provide the desired 10 dB S+N/N ratio. Although 100 feet (30.5m) of RG-8A/U coaxial cable has only about 1.35 dB loss at 50 MHz, it degrades the noise figure sufficiently that the system is no longer limited by external noise sources. This points up the importance of using low-loss transmission lines (or mounting a receiving preamp at the antenna).

Assuming a quiet, rural location that is limited primarily by galactic noise down to about 18 MHz, and atmospheric noise below 18 MHz, what is the minimum usable receiver sensitivity for terrestial communications?

As can be seen from **table 2**, rather poor receiver sensitivity is acceptable on 40, 80 and 160 meters because the external noise at these frequencies is very high. This also explains why the simple receivers of the 1920s were relatively successful. The high external noise levels also make it possible to use rather inefficient receiving antennas on the lower frequencies.[13]

It's important to note that a 0.5 µV signal is of little practical use on 160, 80 or 40 meters because it would be buried in the noise level.

The required sensitivity on 20, 15 and 10 is not difficult to obtain with modern devices, but receivers which are optimized for the lower frequencies may not offer top performance on 10 meters. It should be pointed out that the "acceptable" noise figure in the last column of table 2, is somewhat arbitrary and is based on setting the receiver noise floor about 3 dB below the external noise floor. This is probably adequate 90 per cent of the time, but since noise varies randomly, a statistical analysis indicates there may be times when a lower noise figure may be desirable. However, it is generally agreed that a 10 dB noise figure is more than adequate up to 22 MHz and an 8-dB noise figure may occasionally prove useful on 10 meters. Why design a high frequency receiver for extraordinary sensitivity when its performance is limited by external noise over which you have no control? A very sensitive receiver is more prone to intermodulation and cross-modulation effects, and these may be more important.

At vhf the external noise levels are much lower and low-noise receivers are required for good weak-signal performance. Since it's relatively easy to build low-noise receivers for 50 MHz with modern semiconductors, there's no excuse for being limited by system noise figure on this band.
A receiver with a 5 dB noise figure at 50 MHz, for example, when used with 100 feet (30.5m) of RG-8A/U transmission line, will provide a system noise factor of 4.38 (NF = 6.4 dB) at the antenna terminals. This is well below the external noise.

The 144-MHz example in **table 2** is hopelessly inadequate and represents at least 12 dB degradation over what can be obtained in practice. A receiver with a 1.5 dB noise figure on this band, when used with the 100 feet (30.5m) of RG-8A/U, will provide a system noise factor of 2.54 (NF = 4.1 dB) at the antenna terminals which is still inadequate. A transmission line with 0.7 dB loss would bring receiver noise figure within acceptable limits, but it might be easier and less expensive to install a low noise preamp at the antenna.

As pointed out earlier, the noise temperatures of antennas that are pointed into space for satellite communications (or EME) are much lower than for terrestial communications where the antenna is pointed at the horizon. This means that the receiver noise figures must be lower for maximum performance. Some parts of the sky are noisier than others due to the presence of noise sources, but the noise figure of the receiver should ideally be low enough that the system is galactic-noise limited. Following are the receiver noise figures to shoot for when designing receivers or converters for satellite communications on vhf.

frequency	**galactic noise floor**	**noise figure**
28 MHz	-125 dBm	8 dB
50 MHz	-130 dBm	5 dB
144 MHz	-139 dBm	1 dB
220 MHz	-140 dBm	0.7 dB
432 MHz	-141 dBm	0.2 dB

These figures are based on a 2.1-kHz bandwidth and assume a lossless transmission line. For more accurate calculations at low noise figures, the use of noise temperatures is recommended.

Although the topic of noise figure measurement is beyond the scope of this article, the simplest method of making the measurement is to compare receiver noise to the noise generated by a temperature-limited vacuum diode. This technique is easily applied in the home workshop and has been discussed many times in the amateur radio magazines.14,15,16 Guentzler also described a noise measuring system which used a pilot lamp as a noise source.[17]

intermodulation distortion

Amplitude distortion occurs in an amplifier when the magnitude of the output signal is not exactly proportional to the input signal. Although amplifiers can be designed to be nearly perfectly linear over a portion of their operating range, every amplifier has nonlinearity which can cause distortion products or harmonics of the driving waveform. Intermodulation distortion or IMD is a type of amplitude distortion which occurs when a nonlinear amplifier is driven by more than one discrete frequency. Although the discussion here is limited to IMD in receivers, this is the same distortion which is used to define the linearity of ssb linear power amplifiers.

When an rf signal with varying amplitude is passed through a nonlinear device, many new products are generated. The frequency and amplitude of each component can be calculated mathematically since the nonlinear device can be represented by a power series expanded about the zero-signal operating point.18 Although many products are generated, the ones of primary concern are the second and third. This can be demonstrated with a two-tone signal with outputs at f_1 and f_2 at 14001 and 14003 kHz.

$f_1 = 14001$ $2f_1 = 2802$ $3f_1 = 42003$ $f_2 = 14002$ $2f_2 = 28004$ $3f_2 = 42006$

Although each of the harmonics fall well outside the passband of a receiver which is tuned to pass 14001 and 14002 kHz, the harmonics mix together to produce intermodulation products which do fall within the passband.

The third order products consist of:

$2f_1 - f_2 =$ 14000 kHz $2f_2 - f_1 =$ 14003 kHz

The fifth-order products consist of:

$3f_1 - 2f_2 =$ 13999 $3f_2 - 2f_1 =$ 14004 kHz

The output spectrum is shown in **fig. 6**. Unless the nonlinearity of the amplifier is particularly severe, fifth-order IMD is not usually a problem and can be ignored in receiver applications.

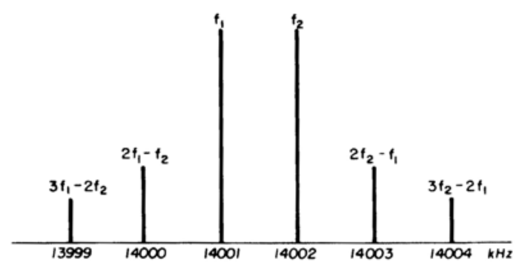

fig. 6. Third- and fifth-order intermodulation products generated by input signals at 14001 and 14002 kHz. In receiver stages fifth-order IMD is usually small enough to be neglected.

Although the IMD distortion products which are generated by two discrete frequencies are used here because they're easy to visualize, exactly the same sort of thing occurs with complex speech waveforms.

In a receiver rf amplifier or mixer stage IMD may be caused by two adjacent CW signals or by a ssb signal. Furthermore, in a mixer where the input must be wideband (such as the double-balanced mixer which is currently finding wide use). Two input signals, f_{R1} and f_{R2} may mix with the local oscillator (f_L) to produce in-band, two-tone, third-order intermodulation products $(2f_{R1} - f_{R2}) \pm f_L$ and $(2f_{R2} - f_{R1}) \pm f_L$. Third-order inter modulation products also occur at $(2f_{R1} - f_{R2})$ and $(2f_{R2} - f_{R1})$. Two input frequencies at 14210 and 14230 kHz, for example, with a 5.2-MHz local oscillator (9 MHz i-f). will produce two-tone, third-order intermodulation products at 8990 and 9050 kHz (i-f passband) and 14190 and 14250 kHz (rf passband).

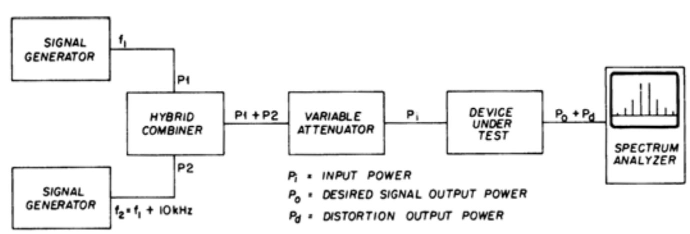

fig. 7. Block diagram of the test setup for evaluating the IMD performance of an amplifier or mixer with a spectrum analyzer.

Third-order IMD is measured in the laboratory with a spectrum analyzer using the test setup in fig. 7. However, the concept of the third-order intercept is finding increased use to describe the IMD response of mixers, and can also be used to describe the linearity of amplifiers.* The third-order intercept is the theoretical poi nt where the two-tone, third-order response is exactly equal to the two-tone input. Amplifiers and mixers are not operated at this level in practice, but the intercept point offers an internationally recognized figure of merit for comparison of devices, both active and passive.

In addition, the intercept point permits comparison of amplifiers and mixers where the intermodulation specifications are given at different two-tone levels. Once the intercept point is known, you can calculate the two-tone, third-order response at any input level by simply remembering that every 1-dB change in the two-tone input produces a 3-dB change in the third-order output. With this information it is possible to predict the maximum rf input level which is allowable.

***Class A amplifiers. Linear class AB or B amplifiers often exhibit two-tone, third-order intermodulation products which follow an S-shaped curve that both increases and decreases with additional input signal level so they cannot be compared by the intercept point method.**

With each 1 dB decrease in the IR input level, for example, the third-order product is decreased an additional 2 dB. As shown in **fig. 8**, a high-level double balanced mixer will suppress third-order products about 65 dB when both signals are at zero dBm (224 mV across 50 ohms) and 85 dB when both input signals are at -10 dBm (71 mV across 50 ohms). The third-order intercept point for these mixers is +27.5 dBm, relative to the output. Relative to the input, the intercept point is at +32.5 dBm. This is 17 dB higher than the intercept point for a low-level double-balanced mixer such as the Minilabs SR1A or Anzac MD108. The 3-dB compression point shown on the graph is a combination of both conversion compression and desensitization.

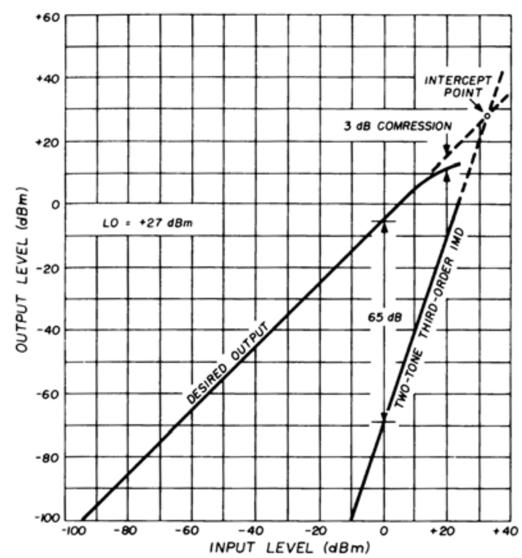

fig 8 Third-order intercept point of high level double balanced mixer (SRA 1H) at 50 MHz is +32.5 dBm relative to the input, 3 dB compression occurs at +20 dBm and third-order IMD is suppressed 65 dB when both input signals are 0 dBm (224 mV across 50 ohms).

intercept point

The third-order intercept point, IP, can be calculated from the relationship

$$IP = 1/2(P_o - P_d) + P_i \qquad \textbf{(16)}$$

IP = third-order intercept, dBm
P_o = desired output, dBm
P_d = third-order distortion products, dBm
P_i = input power, dBm

Since the third-order IMD is defined as (P_o - P_d). eq. 16 can be rewritten as

$$IP = 1/2\ IMD + P_i \qquad \textbf{(17)}$$

For the spectrum display shown in **fig. 9**, for example, the third-order intermodulation distortion products with two input signals of 4 mV (-35 dBm) are 80 dB down, and the third-order intercept is $IP = 0.5 + 80 - 35 = +5\ dBm.$

Most amateurs don't have spectrum analyzers, but if good intermodulation distortion information is provided on the receiver data sheet, the intercept point can be easily calculated with eq. 17 (in all too many cases, however, amateur receiver manufacturers ignore IMD completely, and when they do provide IMD data, it is incomplete).

How ever, assume the specifications for an amateur-band receiver list -75 dB IMD for an input of 1 mV (- 47 dBm). The third-order intercept is $IP=0.5 + 75 - 47 = -9.5dBm$

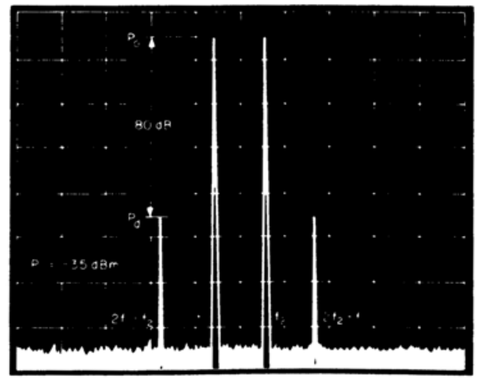

fig 9. Spectrum analyzer display of communications receiver shows IMD is 80 dB down with a two- tone input 0f –35 dBm (4 mV across 50 ohms). Intercept point is at +5 dBm.

Once the intercept point is known, the IMD performance at any input level can be found by rearranging **eq. 17.**

$$IMD = 2(IP - P_i) \qquad \textbf{(18)}$$

For an intercept point of -9.5 dBm, for example, the IMD at various input levels is shown to the right.

input signal	**IMD**
100 µV (-67 dBm)	115 dB
500 µV (-53 dBm)	87 dB
1000 µV (-47 dBm)	75 dB
5 mV (-33 dBm)	47 dB
10 mV (-27 dBm)	35 dB

Compare this with the state-of-the-art receiver front end described on page 17 of this which has -74 dB IMD at an input of 100 mV (-7 dBm). The intercept point is at +30 dBm.

input signal	**IMD**
100 µV (-67 dBm)	194 dB
500 µV (-53 dBm)	166 dB
1000 µV (-47 dBm)	154 dB
5 mV (-33 dBm)	126 dB
10 mV (-27 dBm)	114 dB

The superiority of this receiver is obvious - for all but the strongest signals the IMD products are at or below the receiver noise level. Assuming a 10 dB noise figure and 2.1-kHz bandwidth, an input signal of -23.7 dBm (14.7 mV across 50 ohms) will produce IMD products just equal to the noise level. In the receiver with an intercept point at -9.5 dBm, however, the IMD products are already 3 dB greater than the noise with an input of -49 dBm (800 µV). This will be discussed further under the subject of dynamic range.

double-balanced mixers

Often a mixer data sheet does not specify the third-order intercept point, but a rule-of-thumb estimate can be easily made by examining the 1-dB compression point. As the rf input is increased, the i-f output should follow in a linear manner. However, after a certain point, the i-f output increases at a lower rate until the mixer output becomes fairly constant. The point at which the i-f output deviates from the linear curve by 1 dB is called the 1.dB compression point. At this point the conversion loss is 1 dB greater than it was when the rf input was smaller.

The importance of the 1-dB compression point is its utility in comparing the dynamic range, maximum output and two-tone performance of various double-balanced mixers. As a rule of thumb, the third-order intercept point is approximately 10 to 15 dB higher than the 1-dB compression point19 (about 15 dB at the low frequencies and 10 dB at higher frequencies). This is shown in **fig. 10**.

To properly use a double-balanced mixer, it is necessary to relate the two tone input and third-order output levels to avoid generating excessive distortion which would compromise the final design. This is equally valid for amplifiers. Also important, but not as obvious, is the effect higher operating frequencies have on the double-balanced mixer's two-tone, third-order distortion characteristics. Performance is usually better at the lower frequencies and drops off as frequency is increased. For typical high-frequency double-balanced mixers with a maximum frequency specification of 500 MHz, performance starts to fall off somewhere between 50 and 100 MHz.

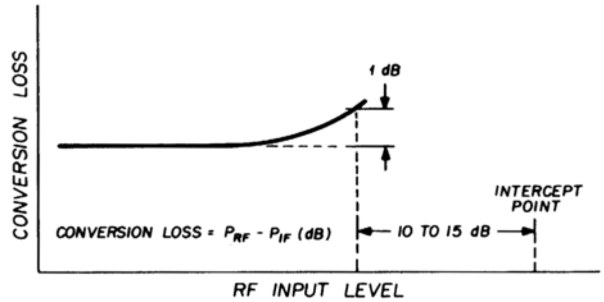

fig. 10. When working with double-balanced mixers, the third-order intercept point may be estimated by using the rule of thumb that the intercept point is 10 to 15 dB above the I-dB compression point (see text).

It is now possible, using a double balanced mixer in the front end, to build high-frequency communications receivers with a third-order intercept at +30 dBm. Using the rule of thumb that the 1-dB compression point is 15 dB below the intercept point, 1-dB compression occurs at an input of approximately + 15 d Bm or 1.25 volts across 50 ohms.

By comparison, the 1 dB compression point of many commercial amateur receivers is in the vicinity of - 20 dBm (22 mV across 50 ohms) and some solid state receivers with bipolar rf amplifiers go into compression at -40 dBm (2 mV across 50 ohms).

cross modulation

Another type of amplitude distortion which can occur in tuned amplifiers is *cross modulation*. This is related to IMD and is produced when the modulation from an undesired signal is partially transferred to a desired signal in the passband of the receiver. The 3.dB compression point in **fig. 8** describes the start of cross-modulation effects.

The cross-modulation effect is independent of the desired signal level and is proportional to the square of the undesired signal amplitude. Because of this relationship, an RF attenuator which lowers the signal level at the input to the receiver may provide a great improvement in cross-modulation performance. A 6 dB attenuator at the receiver input terminals, for example, will reduce cross modulation by 12 dB. If the desired signal is at least 6 dB above the level at which the receiver provides a satisfactory S/N ratio this results in a marked improvement in received signal quality.

Cross modulation is measured in the laboratory by setting one signal generator to deliver a CW output and another generator is set up for 30% amplitude modulation. The output of the a-m generator is increased until 1% modulation appears on the CW signal as measured with a spectrum analyzer. This represents a cross-modulation ratio of about 30 dB (cross-modulation level 30 dB below the reference level).

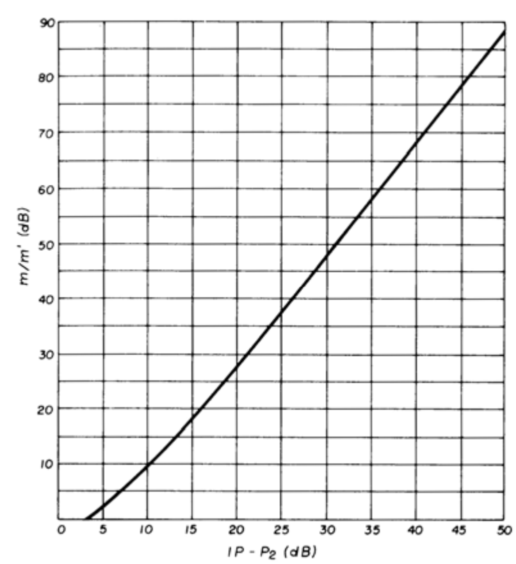

fig 11. Graph of cross modulation vs the ratio of the intercept point to the interfering signal level in dB.

Cross modulation is related to the intercept point by the relationship $m/m^1 = (P_{ip}/4P_c) - 1/2$ **(19)**

m/m^1 = ratio of cross modulation transferred from a large signal to a smaller one
P_{ip} = intercept point power
P_c = interfering signal power

Cross modulation in dB is simply: $m/m^1 (dB) = 20 \log (m/m^1)$ **(20)**

In **fig. 11** cross modulation is plotted against the difference between the intercept point and the cross-modulating signal in dBm. Cross modulation of 30 dB, for example, corresponds to 21 dB difference between the intercept point and the signal producing the cross modulation. For a receiver with an intercept point at +30 dBm, a modulated input signal at a level of +9 dBm (630 mV across 50 ohms) will produce 30 dB cross modulation. For a receiver with an intercept point at -9.5 dBm (the more usual case), an interfering signal level of -30.5 dBm (6.7 mV) will produce 30 dB cross modulation.

gain compression

When a receiver is tuned to a weak signal, a strong, adjacent signal may cause an apparent decrease in receiver gain. This is called compression or desensitization and occurs when the input voltage from the undesired signal is large enough to exceed the bias on an RF amplifier or mixer and drives the base (or grid) into conduction. This reduces gain, as shown by the compressed curve of fig. 8, and increases distortion. The rectified base (or grid) current can also be coupled back to the receiver's agc system which results in a further reduction in overall receiver gain.

Compression is measured by setting one signal generator to produce a CW signal and another generator, at a given frequency spacing, is adjusted to depress the desired signal a certain amount, usually 3 dB. Like cross modulation, however, a compression specification has little meaning if the frequency separation between the two signals is not specified.

Since both cross modulation and compression are caused by strong, undesired signals which are adjacent to the receiver passband, they can be controlled to a certain extent by the selectivity at the front end of the receiver. High receiver sensitivity, of course, is the antithesis of good cross-modulation and compression performance - this reinforces the argument for receiver noise figures on the order of 10 dB for the high frequency range.

dynamic range

The front end of a receiver is subjected to a multiplicity of input signals which tend to intermodulate to produce a level of distortion products which is dependent on the magnitude of the incoming signals. Therefore, the upper end of a receiver's dynamic range is defined by the input signal level which produces third-order 1M 0 products just equal to the receiver's noise level.

At the lower end dynamic range is limited ultimately by the noise figure of the receiver. Also important, however, is the way the receiver handles weak signals. Some linear RF amplifiers and mixers give good performance in the middle of their operating range but exhibit transfer curves that introduce considerable distortion at low signal levels. With proper design, however, this is not a problem, and the dynamic range of a receiver is usually defined as the spurious-free dynamic range where the maximum input signal is as defined above and the minimum input signal is at the noise floor of the receiver (**eq. 8**).

$$DR = 2/3\ (IP - N_o) \quad \textbf{(21)}$$

where DR = spurious-free dynamic range, dB
IP = intercept point, dBm
N_o = receiver noise floor, dBm

The dynamic range of a receiver is important because it allows you to directly compare the strong-signal performance of one receiver against that of another. On today's crowded bands, and the high incidence of kilowatt transmitters and directive antennas, strong signal performance is usually much more important than sensitivity. Although dynamic range can be used as a figure of merit, it's also useful to know the maximum input signal level, $P_{i(max)}$, which will produce third-order IMD products just equal to the noise level. This can be calculated from

$$P_{i(max)} = 1/3\,(2IP + N_o) \qquad \textbf{(22)}$$

$P_{i(max)}$ = maximum input signal, dBm
IP = intercept point, dBm
N_o = receiver noise floor, dBm

For example, assuming a noise figure of 10 dB and 2.1-kHz bandwidth, the spurious-free dynamic range and maximum input signal of a receiver with an intercept point at +30 dBm are

$$DR = 2/3[30 - (-131)] = 107.3\ dB$$

$$P_{i(max)} = 1/3[\ 2 \times 30) + (-131)]$$

$$= -23.7dBm$$

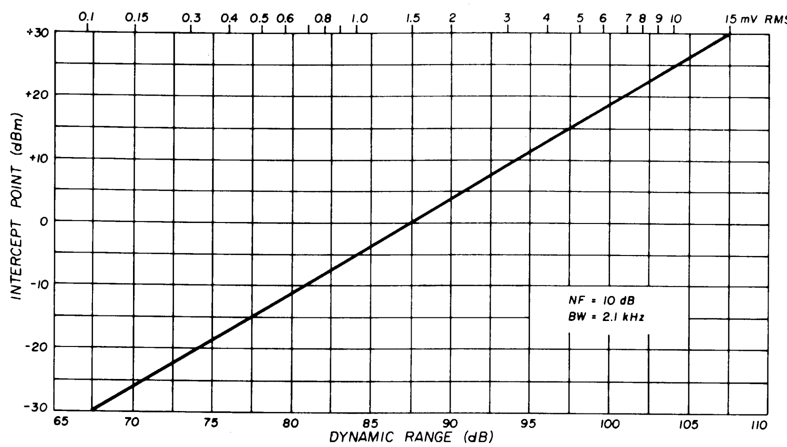

fig. 12. Spurious-free dynamic range vs intercept point (based on 2.1.kHz bandwidth, 10 dB noise figure). Maximum input signal (millivolts rms across 50 ohms) is shown at top.

Thus, the IMD products will be well below the noise level for all input signals below -23.7 dBm (14,000 µV or about S9+43 dB)! As a comparison, consider a receiver with a third-order intercept at -9.5 dBm (10 dB noise figure, 2.1-kHz bandwidth):

$DR = 2/3[-9.5 - (-131)] = 81\ dB$ $\qquad$ $P_{i(max)} = 1/3[2(-9.5) + (-131)] = -50$ dBm

With this receiver the distortion products are equal to the noise level with an input signal of about 710 µV or S9+17 dB. This may represent adequate strong-signal performance if you live out in the country, but it's doubtful. If you live in an urban area, you'll have a lot of trouble digging weak signals out of the morass of IMD products which effectively raise the noise floor of the receiver.
A graph of intercept point vs dynamic range and maximum input signal is presented in fig. 12 for a receiver with a 10 dB noise figure and 2.1-kHz bandwidth (typical for modern amateur communications receivers).

summary

Although modern amateur receivers are no longer performance limited by noisy vacuum tubes or poor selectivity, the published performance specifications have changed little since the 1940s and are still limited essentially to data on sensitivity and selectivity. Specifications on intermodulation distortion, cross modulation, desensitization and dynamic range, if they're mentioned at all, provide insufficient information for direct buyer comparison. Purchasing a new receiver under these conditions is a bit like buying a new car without knowing gas mileage or how many passengers it will carry.

The performance specifications for the high-frequency receiver shown in table 3 leave no question as to receiver performance and are recommended as a guide for receiver manufacturers to follow in the future. Well informed amateurs should demand nothing less.

table 3. Specifications for a high-performance, high-frequency communications receiver provide a good format for amateur equipment manufacturers to follow. These specifications give complete reference and qualifying data and leave little question as to actual receiver performance.

Frequency range	500 kHz to 30 MHz
Tuning accuracy :	± 500 Hz relative to the frequency of the desired signal
Sensitivity	CW and ssb: 0.5 µV for 10 dB SIN ratio in a 2.4-kHz bandwidth (11 dB noise figure)
I-f selectivity	2.1 kHz at -6 dB, 4.2 kHz at -60 dB (2.0 shape factor)
Intermodulation products	**Out of band**: With two 20 mV signals separated and removed from the desired signal by not less than 25 kHz the third-order IMD products are not less than 90 dB below either of the interfering signals. Intercept point = + 24 dBm. **In band:** Two In-band signals of 20 mV will produce third order IMD products not greater than -50 dBm
Dynamic range	102 dB
Cross modulation	With a desired signal greater than 100 µV In a 2.4-kHz bandwidth, an unwanted signal, 30% modulated, removed not less than 25 kHz, must be greater than 175 mV to produce an output 30 dB below the output produced by the desired signal.
Compression	With a desired signal of 500 µv, an unwanted signal more than 25 kHz removed must be greater than 300 mV to reduce the output by 3 dB
Spurious response	**External signals** 25 kHz or more removed from the desired signal must be at least 85 dB above the level of the desired signal to produce an equivalent output **Internal** spurious signals are not more than 3 dB above the noise level measured in a 2.1-kHz bandwidth
AGC range	An increase in input of 110 dB above 1 µV will produce an output change of less than 6 dB

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