A Precision Two-Tone RF Generator for IMD Measurements

Crystal oscillators plus crystal filters yield extremely low phase noise in a high-IP_3 IMD test generator for measuring the dynamic range of HF receivers.

By Stuart Rumley, KI6QP

There have been numerous articles in amateur literature on the significance of high-dynamic range performance in HF receivers.[1] Without sufficient dynamic range, a receiver's other important virtues, mainly selectivity and sensitivity, soon become ineffective. Dynamic range can be specified and measured as either blocking dynamic range or intermodulation distortion (IMD) dynamic range. Blocking dynamic range refers to a receiver's ability to not be desensitized when strong signals are present. IMD dynamic range, on the other hand, refers to the receiver's ability to not generate false signals whenever there are two or more strong signals present. The IMD measurement leads directly to a convenient figure of merit for dynamic range known as the third-order intercept or IP_3. The purpose of this article is to show how to construct an inexpensive source for making the IMD measurements and how to make the measurements and calculate IP_3, as well as to compare some typical receivers.

Making the IMD measurements requires two high-stability, low-phase-noise RF signal sources at the frequencies of interest. The typical setup for measuring IMD usually looks something like Fig 1.[2] Unfortunately, most enthusiastic radio amateurs do not have one, let alone two, low-phase-noise synthesized signal generators such as the HP8640A shown. Two synthesizers are required because the two-tone IMD measurement requires two frequency stable sources set only a few kilohertz apart.

Attempting to make this measurement with unstable sources proves quite frustrating. If you are prepared to trade frequency agility for cost, two crystal oscillators operating at the desired test frequencies can be just as useful and, to some extent, more convenient than a pair of synthesizers. Fig 2 shows a block diagram of the essential functions of such a test system. There are two crystal oscillators, one for generating each tone.

Each crystal oscillator is followed by a crystal filter network tuned to the same frequency as the oscillator. The output of each crystal filter network is attenuated, then summed together in a hybrid combiner. The combined output from the hybrid is further filtered to remove harmonics. I selected 14.20 and 14.22 MHz as the optimum frequencies for the two crystal oscillators. This is approximately mid-range for most continuous-coverage HF

[1]Notes appear on page 12.

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receivers (0.5 to 30 MHz) and is also in the middle of the 20-meter band so it may be used with ham-band-only receivers.

The two-tone IMD generator concept is rather simple, but a number of subtleties must be taken into consideration. In order to accurately measure receiver performance, the two-tone source must possess the following attributes:

1. Low equivalent IMD to ensure that the distortion products measured are only from the receiver or system under test, not from the test generator.
2. Low phase noise. The phase noise of both tones, at the frequency offset of the expected IMD products, must be much less than these IMD products.
3. Low harmonic energy. Because third-order IMD products do contain energy at the second and third harmonics of F1 and F2, it is essential that the harmonic energy from the two-tone generator be extremely low.
4. Careful filtering and shielding to ensure that external signals and noise do not interfere with the measurements.

Circuit Description

The detailed circuit schematic is shown in Fig 3. Q1 and Q2 are fundamental-mode Colpitts crystal oscillators operating on 14.200 and 14.220 MHz, respectively. The selection of the actual frequencies is determined by two bounding requirements. One, the frequencies cannot be so close together that the test receivers cannot resolve the IMD products because of either selectivity limitations or phase noise from the receiver's local oscillator. Two, the frequencies cannot be so far apart that the input band-pass filters will cause unequal attenuation of either tone. A good choice is 20 to 25 kHz. The output from each oscillator is taken from a tuned collector circuit in order to minimize harmonic energy and provide a low-impedance source to the crystal filter network. The oscillators are operating at a rather high power level, with over 10 mA of collector current in each transistor. The high power is required in order to provide –10 dBm at each tone to the output. This power level was selected so that at least 10 dB of attenuation could always be left in the step attenuators and still allow adequate signal energy to create measurable IMD products in high-dynamic-range receivers. The reason it is desirable to leave some attenuation between the IMD generator and receiver under test is because the receiver cannot be trusted to provide a good 50-Ω match to the IMD generator's output filter. The 10 dB of attenuation guarantees at least 20 dB of return loss to both the IMD output filter and the receiver's input band-pass filters.

Crystal oscillators generally have quite low phase noise, but operation above 1 mA can degrade their performance to some degree. In order to maintain extremely low phase noise, I took a couple of additional measures. The emitter degeneration resistors, R3 and R7, provide negative feedback, which reduces the oscillator phase noise. Following each oscillator is a narrow-band crystal filter network C1, C5, Y2 and C10, C15, Y3. The crystal filters provide an additional 30 dB of phase-noise attenuation at a carrier offset of ±10 kHz. The addition of the crystal filters might seem a bit excessive but is necessary in order to be absolutely sure that the phase-noise performance of the two-tone generator does not preclude the ability to measure low-level IMD products in high-performance receivers. An additional benefit of having two very clean sources is the ability to evaluate the phase noise and reciprocal mixing of the receiver under test.

The outputs from each of the crystal filter networks are passed through 6-dB attenuators comprised of R12, R13, R14 and R15, R16, R17 and summed in the hybrid combiner network made up of T1, R9, R10 and R11. The attenuators are required in order to isolate the combiner from the crystal filters and to provide a constant impedance for both. The combiner sums the two frequencies with an insertion loss of 6 dB and approximately 40 dB of isolation. This combination of filters, attenuators and combiner provides a great deal of isolation between the collector circuits of the two oscillators (more than 90 dB). Together with good physical isolation, this topology prevents the generation of any significant internal IMD products.

The output harmonic filter is a seven-section, all-pole design using the same inductors used for the oscillator. This filter adds approximately 1 dB of insertion loss at the operating frequency, more than 50 dB of loss at the second harmonic and more than 70 dB at the third harmonic. From the output of the crystal filter, the second harmonic is down more than 40 dB and the third harmonic is down more than 55 dB. The resulting output has a second harmonic of less than –100 dBm and a third harmonic less than –135 dBm. These levels are so low that they will not contribute in any measurable way to the IMD values of the receiver under test. Because the second harmonic is less than –100 dBm, it's practical to search for second-order products at 28.400 MHz, 28.420 MHz and 28.440 MHz.

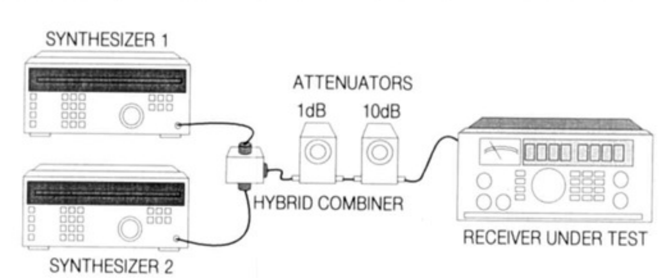


Fig 1—IMD measurement setup.

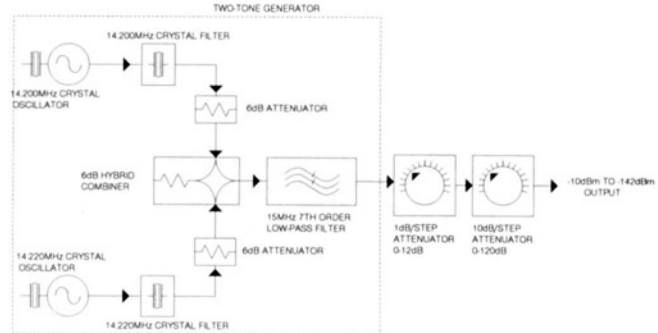


Fig 2—Block diagram of the two-tone source.

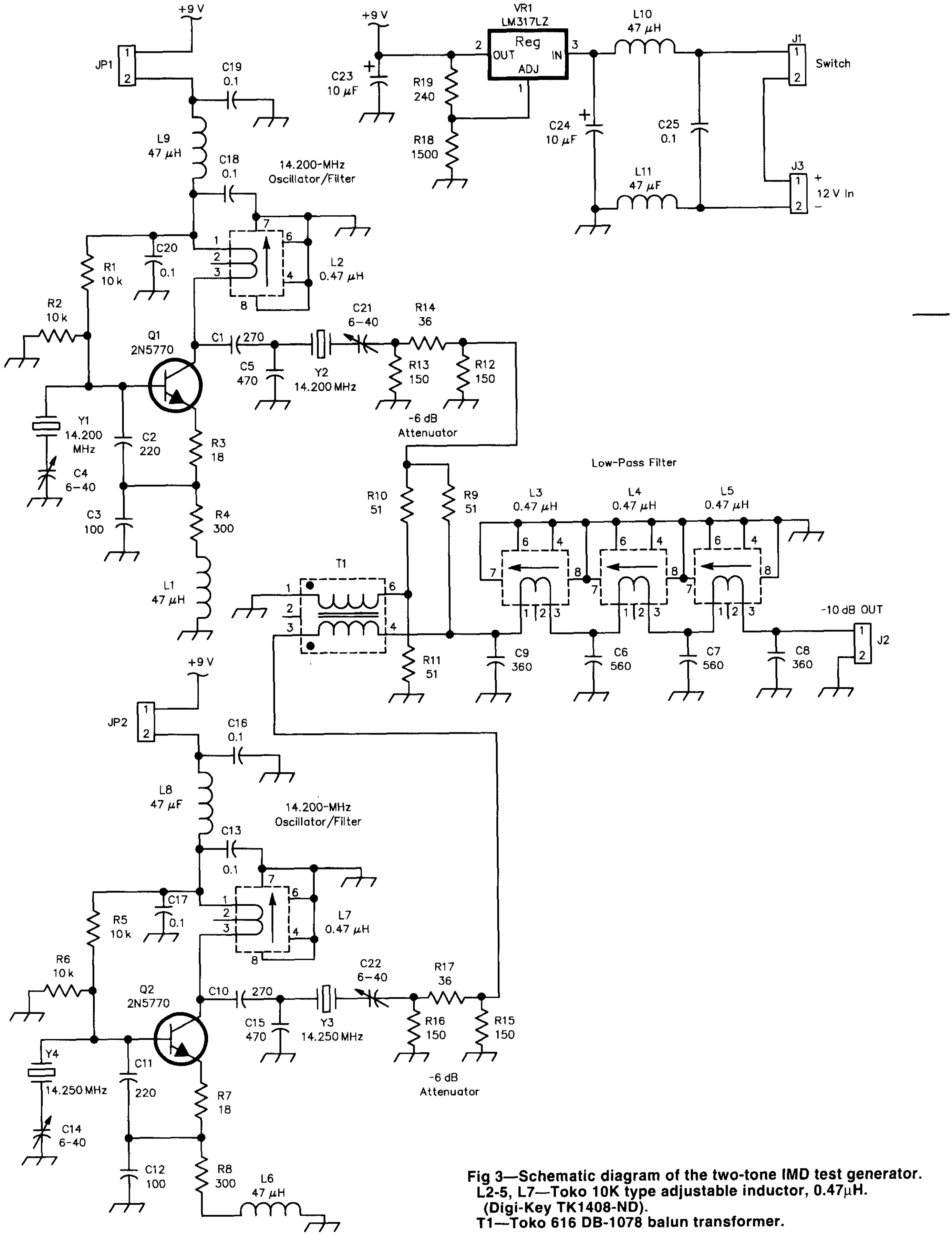
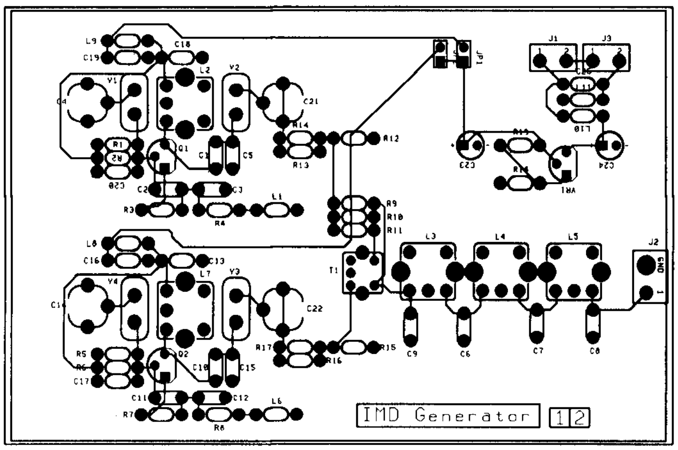


Fig 3—Schematic diagram of the two-tone IMD test generator.
L2-5, L7—Toko 10K type adjustable inductor, 0.47µH.
(Digi-Key TK1408-ND).
T1—Toko 616 DB-1078 balun transformer.



Use good RF prototyping techniques in constructing your unit: short lead lengths, good grounding and shielding. Be particularly careful to provide some physical separation between the oscillators, crystal filters, combiner and harmonic filter. If inadequate shielding or isolation is provided, IMD products will form in the oscillators from cross coupling. Similarly, isolation is required around the crystal and low-pass filters in order to maintain their good cut-off characteristics.

Because there are a number of coils used in this project, I choose not to use hand-wound toroid inductors. I think they would be too tedious to wind and lack adjustability. The inductors used, however, are a quality product, inexpensive and readily available.

Alignment

Alignment of the generator is straightforward: only an oscilloscope, frequency counter and a 50-Ω feedthrough termination is required. The feedthrough termination is used with the oscilloscope to provide the proper load impedance to the generator.

First, each oscillator and filter combination is tuned up independently. Begin by disabling one of the oscillators by removing its supply jumper (JP1 or JP2). Optimize the output of the other oscillator by adjusting its collector inductor (L2 or L7) for maximum output on the oscilloscope. The oscillator is then set on frequency with the frequency counter by adjusting the appropriate trimmer, either C4 or C14. After the oscillator frequency is set, remove the counter and connect the oscilloscope again. Now carefully adjust the corresponding filter-tuning trimmer (C21 or C22) for maximum output. Repeat this process for the other oscillator and filter combination.

Adjustment of the output filter is not critical. Alternately adjust each inductor, L3, L4 and L5, for maximum output from either oscillator. Adjustment by this method should give second-harmonic attenuation values within a dB or so of what you might achieve using a network analyzer to set this filter.

Finally, with the oscilloscope and termination still connected, (by the way, the termination should be at the scope end of the cable), disable one of the oscillators at a time and again trim L2 or L7 for −10 dBm of output at each frequency. Minus 10 dBm corresponds to 200 mV pk-pk into 50 Ω.

Intermodulation Distortion (IMD) Measurements

By definition, any completely linear circuit element would produce no IMD products (see **Appendix A**). But any real receiver circuit exhibits some degree of nonlinearity. It is precisely this degree of nonlinearity we wish to measure by making IMD measurements. The most troublesome of the intermodulation products are the so-called third-order products. These are the $2F1-F2$ and $2F2-F1$ signals shown on the hypothetical spectrum analyzer of Fig 4. If you were using a receiver with similar IMD performance to copy a weak signal at or near one of these intermod frequencies, you would suffer some interference. The higher your receiver's IP_3 value (in dBm), the lower these third-order products—and the consequent interference—will be. Notice that the third-order IMD product ($2F1$-$F2$) in Fig 4 is shown as 80 dB below the two signals $F1$ and $F2$. If the power levels of $F1$ and $F2$ were to decrease by 10 dB, the power levels of the third-order IMD products would decrease by 30 dB. Because the levels of the third-order IMD products are dependent on the input signal level as well as the nonlinearities of the system, the third-order intercept or, IP_3, is a more useful figure of merit for system performance; it is independent of the signal amplitude.

Fig 5 is a graphical representation of the IP_3 concept. If the intercept point is known, the level of the third-order intermodulation product can be determined from the graph. In the example in Fig 5, if two tones at −38 dBm are applied to this hypothetical receiver system with an IP_3 of +3 dBm, their fundamental signal amplitudes would measure +40 dB over S9 and the third-order intermodulation products would measure S1.

Making the Measurements

To make third-order IMD measurements, connect the

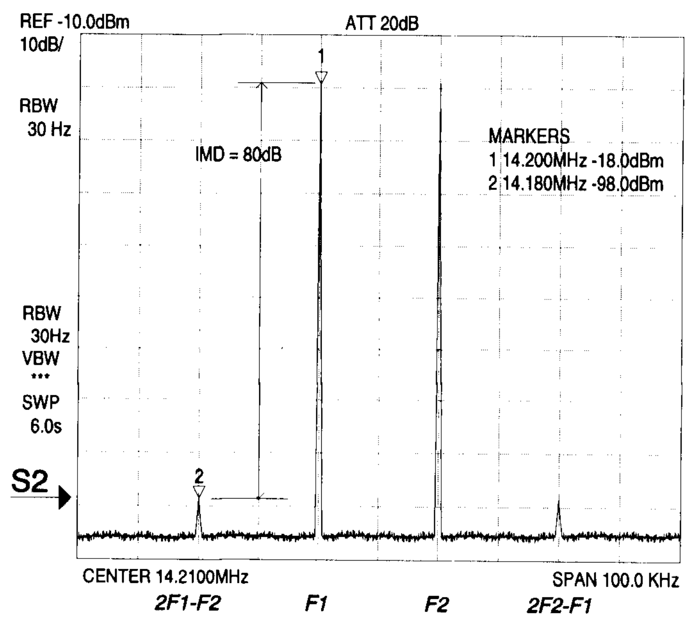


Fig 4—Typical third-order IMD products as seen on a spectrum analyzer.

Appendix A: Why is It Called Third Order?

The reason the particular type of distortion I have been referring to is called third order has to do with the derivation of the mathematical model for an imperfect amplifier. You will not need to be a math professor to follow this explanation, nor is it necessary that you be completely comfortable with a bunch of trigonometric relations. Just keep in mind that ω_1 and ω_2 are the two input signal frequencies in terms of radians per second and are equivalent to $2\pi f_1$ and $2\pi f_2$. Watch what happens to these terms as the power series is expanded for the cubic (third-order) term.

All amplifiers have some amount of distortion. In contrast to most general-purpose wide-band or video amplifiers, the outputs of RF or IF amplifiers are generally filtered. As shown in the amplifier configuration below, the filters effectively remove harmonically related signals caused by the nonlinear behavior of the amplifier. However, the third-order products are typically within the band-pass of these filters and therefore of particular concern.

If an amplifier had no distortion, its transfer function would be:

$$V_{out} = A_0 + A_1 V_{in}$$

where A_0 is just the dc offset and A_1represents the coefficient of the desired linear gain. Because most real amplifiers do have some distortion, their transfer functions can better be represented by a power series polynomial:

$$V_{out} = A_0 + A_1 V_{in} + A_2 V_{in}^2 + A_3 V_{in}^3 + A_4 V_{in}^4 \cdots$$

For

$$V_{in} = V_1 \cos(\omega_1 t) + V_2 \cos(\omega_2 t)$$

the desired first order term, $A_0 + A_1 V_{in}$, gives the fundamental products

$$V_{out} = A_0 + A_1 V_1 \cos(\omega_1 t) + A_1 V_2 \cos(\omega_2 t)$$

The second order term, $A_2 V_{in}^2$, determines the second order products

$$A_2 V_{in}^2 = \frac{A_2 V_1^2}{2} + \frac{A_2 V_2^2}{2} +$$
$$\frac{A_2 V_1^2}{2}\cos(2\omega_1 t) + \frac{A_2 V_2^2}{2}\cos(2\omega_2 t) +$$
$$\frac{A_2 V_1 V_2}{2}[\cos(\omega_1 t + \omega_2 t) + \cos(\omega_1 t - \omega_2 t)]$$

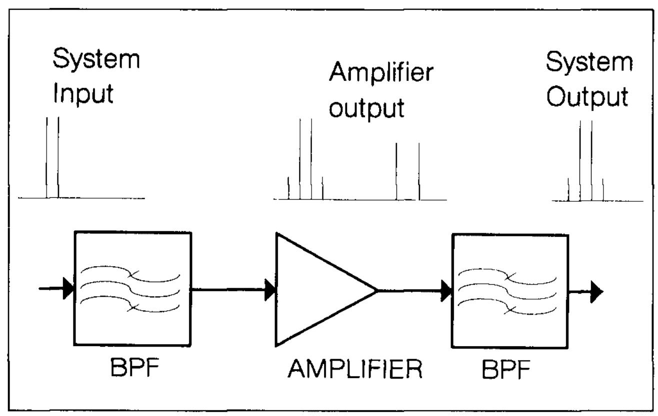
The first line shows the dc terms, the second line shows the second-harmonic terms, and the last line has the second-order IMD terms.

Here is where it gets interesting, the third-order term, $A_3 V_{in}^3$, gives us

$$A_3 V_{in}^3 = \frac{3A_3}{2}\left[V_1 V_2^2 + \frac{V_1^3}{2}\right]\cos(\omega_1 t) + \frac{3A_3}{2}\left[V_1^2 V_2 + \frac{V_2^3}{2}\right]\cos(\omega_2 t) +$$
$$\frac{A_3 V_1^3}{4}\cos(3\omega_1 t) + \frac{A_3 V_2^3}{4}\cos(3\omega_2 t) +$$
$$\frac{3A_3 V_1^2 V_2}{4}[\cos(2\omega_1 t + \omega_2 t) + \cos(2\omega_1 t - \omega_2 t)] +$$
$$\frac{3A_3 V_1 V_2^2}{4}[\cos(2\omega_2 t + \omega_1 t) + \cos(2\omega_2 t - \omega_1 t)]$$

The first line consists of terms at the fundamental (input signal) frequency. The second line again shows harmonic terms—third harmonics this time. The last two lines give the two third-order IMD products.

The difference terms in the third-order products are the troublemakers, and they usually turn up right next to a weak signal you are trying to copy. Notice that as the amplifier approaches the ideal, the coefficients (A_2, A_3,...A_n) would approach zero.



test receiver and attenuator set to the two-tone IMD source as shown in Fig 6. If you don't already have a set of step attenuators, you can build your own from designs in *The ARRL Handbook.* To get a feel for what to expect, begin by setting the attenuators for a combined value of 20 dB. Now tune the test receiver from 14.175 to 14.245 MHz; most receivers will have a noticeable third-order intermod at 14.180 and 14.240 MHz, plus the two very strong fundamental signals at 14.200 and 14.220 MHz. Your receiver should be set to either USB or LSB mode with the AGC on, RF attenuator to 0 dB and any preamplifier off. You may argue that the AGC should be off when making the measurements so that the RF and IF amplifiers will be at maximum gain. In principle I would agree, but the problem is that the S-meter will probably not be working with the AGC off, and it is required in order to make the necessary measurements.

Now that you have found the intermod signals, begin the measurement process by calibrating the receiver's S-meter. It is only necessary to calibrate the S-meter at one low-level value, say S1 or S2. This is done by setting the attenuators to a combined value of approximately 100-dB (–110 dBm) and tuning the receiver to either 14.200 or 14.220 MHz. Adjust the 1-dB step attenuator until the S-meter reads exactly S1 or S2. The choice of which to use depends on how responsive the meter is to a 1-dB attenuation change; some receiver's meters will not respond well at S1, so try S2 instead. Try to avoid using anything higher than S2 because the AGC will affect the linearity of the receiver's input circuits. The idea is that you should be able to resolve a

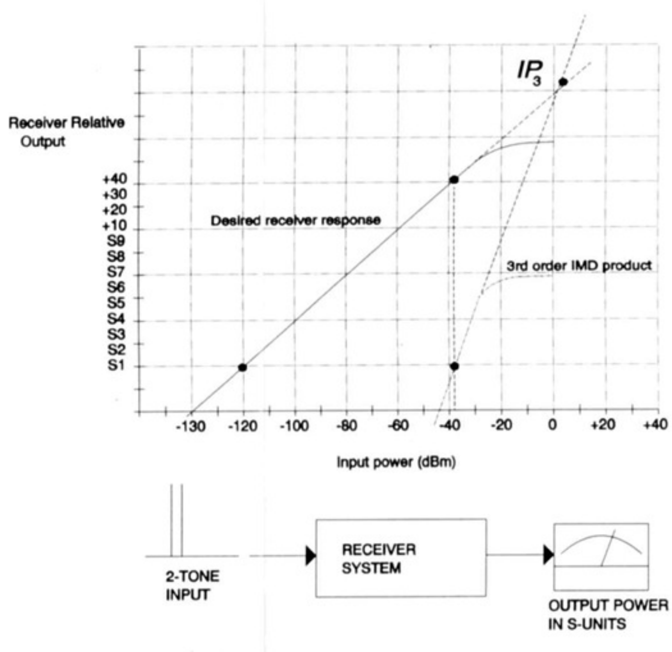


Fig 5—Third order intercept.

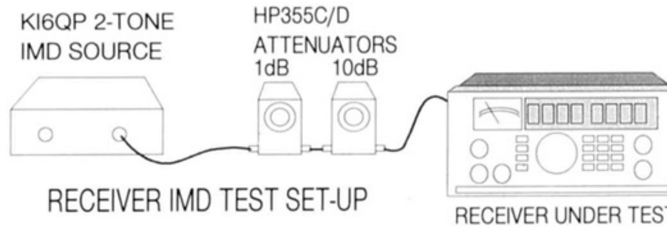


Fig 6—IMD measurement setup.

Table 1—Receivers Tested.

Manufacturer	*Model*	P_A *(dBm)*	P_{IM} *(dBm)*	IP_3 *(dBm)*
Drake	R-4C	–29	–92 at S3	+2.5
ICOM	IC-R70	–30	–96 at S1	+3.0
ICOM	IC-765	–29	–93 at S2	+3.0
ICOM	IC-781	–19	–95 at S1	+19.0
Kenwood	R599D	–50	–107 at S1	–21.5
Kenwood	TS-830S	–42	–107 at S1	–9.5

reference level within ±1dB of variation in input signal strength at some very low level near the receiver's noise floor. Now you should have an absolute signal level corresponding to a particular S-meter value. Record the signal level (P_{IM}) as –10 dBm minus the combined attenuator values.

Now, tune the receiver to either 14.180 or 14.240 MHz and increase the input signal level by decreasing the value of attenuation until the IMD signal is equal to the previously established S-meter calibration point. Record this signal (P_A) as –10 dBm minus the combined attenuator values. The IMD at this particular signal level is the difference in attenuator values $P_A - P_{IM}$. The IP_3 is equal to P_A plus half of the IMD value. Sounds confusing, so let's do an example:

With the receiver tuned to 14.200 MHz, let's say an S2 reading requires an attenuator setting of 88 dB. Therefore, P_{IM} = –10 dBm –88 dB = –98 dBm. And with the receiver tuned to IMD frequency of 14.180 MHz, the attenuator setting is found to be 18 dB. Therefore P_A = –10 dBm –18 dB = –28 dBm. So, IP_3 =[–28 dBm – (–98 dBm)] / 2 + (–28 dBm) = +7 dBm. This is derived from the more generalized form of intercept point:

$$IP_n = \frac{nP_A - P_{IM_n}}{n-1}$$

which in this particular example would look like:

$$IP_3 = \frac{3(-10\text{dBm} - 18\text{dB}) - (-10\text{dBm} - 88\text{dB})}{3-1} = 7\text{dBm}$$

You can use either method you prefer. I like the first example because I can do it in my head. As another example, consider the hypothetical spectrum analyzer shown in Fig 4 and see if you can determine its IP_3 in your head: –18 dBm +80 dB/2 = +22 dBm.

Table 1 shows how a few familiar receivers compare in terms of IP_3 performance. Use this as a benchmark when testing your receivers.

For more insight on IMD characteristics, switch the test receiver's preamplifier on, if it has one, and run the test again. You should notice IP_3 decrease by a value approximately equal to the gain the preamp added. This is a good reason for leaving the preamplifier off! Try it again, but this time switch in the receiver's attenuator. The IP_3 should increase—at a detriment to the receiver's sensitivity. You may notice that either of the two IMD products ($2F1$–$F2$ or $2F2$–$F1$) is significantly lower than the other. This happens in most receivers as a result of the way the IMD products add in cascaded front-end stages. You should assume, however, that a receiver system can be no better than worst case (ie, don't average the two IP_3 results).

Conclusion

I hope you have found this article interesting and useful.

Sources

International Crystal Manufacturing Co, Inc (ICM), PO Box 26330, 10N Lee, Oklahoma City, OK 73126-0330. Tel: 800-725-1426. Specify a holder capacitance of 23 pF and order all four crystals with a matched series resistance of 15 Ω. Four crystals at the time of this writing were $40, including shipping.

Digi-Key Corporation, 701 Brooks Ave South, PO Box 677, Thief River Falls, MN 56701-0677. Tel: 800-344-4539. All parts except the crystal should be available from Digi-Key; no minimum purchase required; service charge on orders under $25.

The techniques described here should give you further insight into receiver design details as well as helping you select a commercial unit. Bare PC boards as well as assembled and tested units will be available soon. For more information contact the author.

Notes

[1]Rohde, Ulrich L., DJ2LR, "High-Dynamic Range Active Double-Balanced Mixer," *Ham Radio*, November 1977.

[2]Rohde, Dr. Ulrich L., KA2WEU, "Testing and Calculation Intermodulation Distortion in Receivers," *QEX*, July 1994.

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Hawker, Pat, G3VA, "G3SBI's H-Mode Receiver Design," *Communications Quarterly*, Fall 1994.

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