Spectrum Analyzer Updates

Most recent update, 17 Oct 08, comments 26april11

Updates deal with the Phono Plugs for VHF connectors (K3NHI) [29Aug06]
Tracking Generator modifications [31Aug06] [Update 5Jan07].

The 70 MHz low pass filter in this mod is recommended for the front end of the analyzer.

Minor Log Amp/IF Amp changes in red on page 14 of this note.
A 17Nov06 entry deals with diode ring mixers that are "backward."
A detailed Calibration Procedure [October, 2008]. Also included is an optional gain stage for the IF and a low current calibrator circuit.
Please scroll down to the section of interest for discussion.
(Follow the colors.)

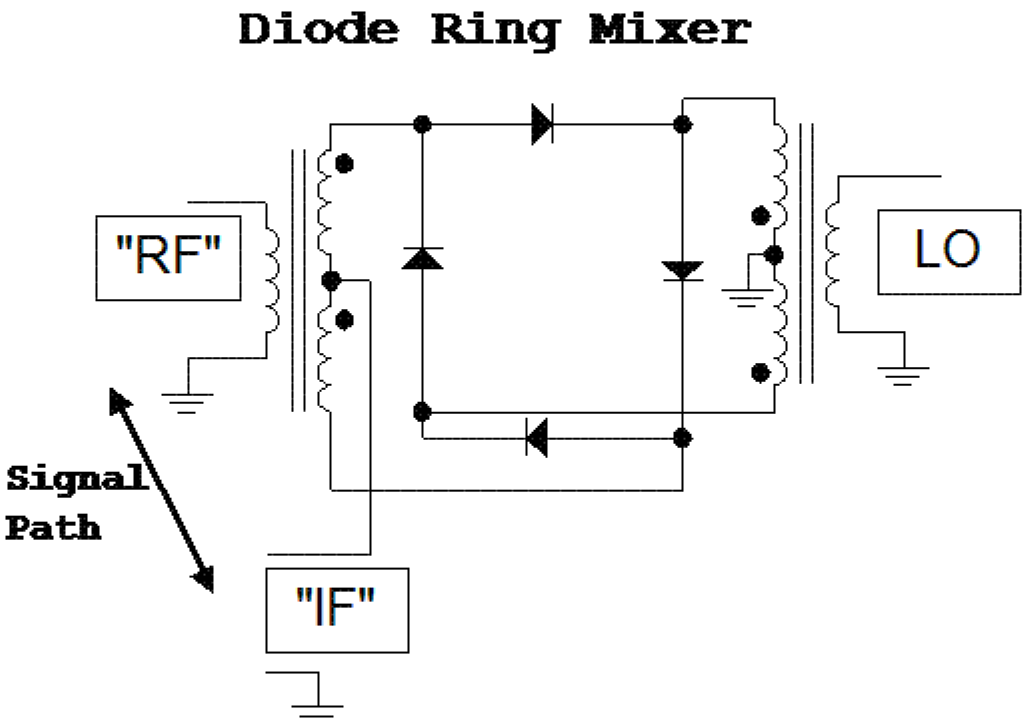
A companion document, "Early SA Updates," shows the state of this file in June, 2000 and is included in this list. This file contains information on the crystal filters used with this analyzer as well as some applications information. Much of the data from that file is now included in **Experimental Methods in RF Design** (ARRL, 2003).

This part of the web page deals with the spectrum analyzer that Terry White (K7TAU) and I described in QST for August and September of 1998, and the related Tracking Generator appearing in November 1999.

The complete article is available to be viewed and downloaded in PDF format from ARRL. Go to the ARRL web page and do a search on *spectrum analyzer*. This will then get you to a point where the two part paper can be downloaded. Also see **kanga** for information regarding kits of parts.

Backward Mixers? --- 17Nov06

A common question that I see regarding the spectrum analyzer is something like "Is there a typo in the way the mixer is drawn in the schematic? You show the signal from the input going into the terminal that we usually use for the IF."   No, there is no typo, but it is a good question.



Signals can travel in either direction in a passive, switching mode mixer.   Almost all of the properties are retained with changes in direction including port-to-port isolation, gain, and intermodulation distortion.   The one property that changes is frequency response.   Each port (rather than an entire path) dictates the frequencies that the mixer can process.   A transformer coupled terminal pair, which is the one we usually call the "RF Port," is only functional at frequencies where the transformer is an operating component.   The usual transformers we use for RF mixers consist of a small ferrite core (toroid, bead, or balun) with a trifilar winding. A typical minimum frequency is around 1 MHz. In contrast, the usual "IF" port will function all the way to DC.   So it is common to use the DC coupled port as the input for a spectrum analyzer.   The lowest frequency is then dictated by the blocking capacitor that we usually find at the input and by local oscillator feedthrough.

update, 4Dec02, 29Aug06

The methods that K3NHI has developed for using Phono plugs at VHF are now listed in a companion file, "RCA Plugs-Sockets."

Some Tracking Generator modifications are in the attached file

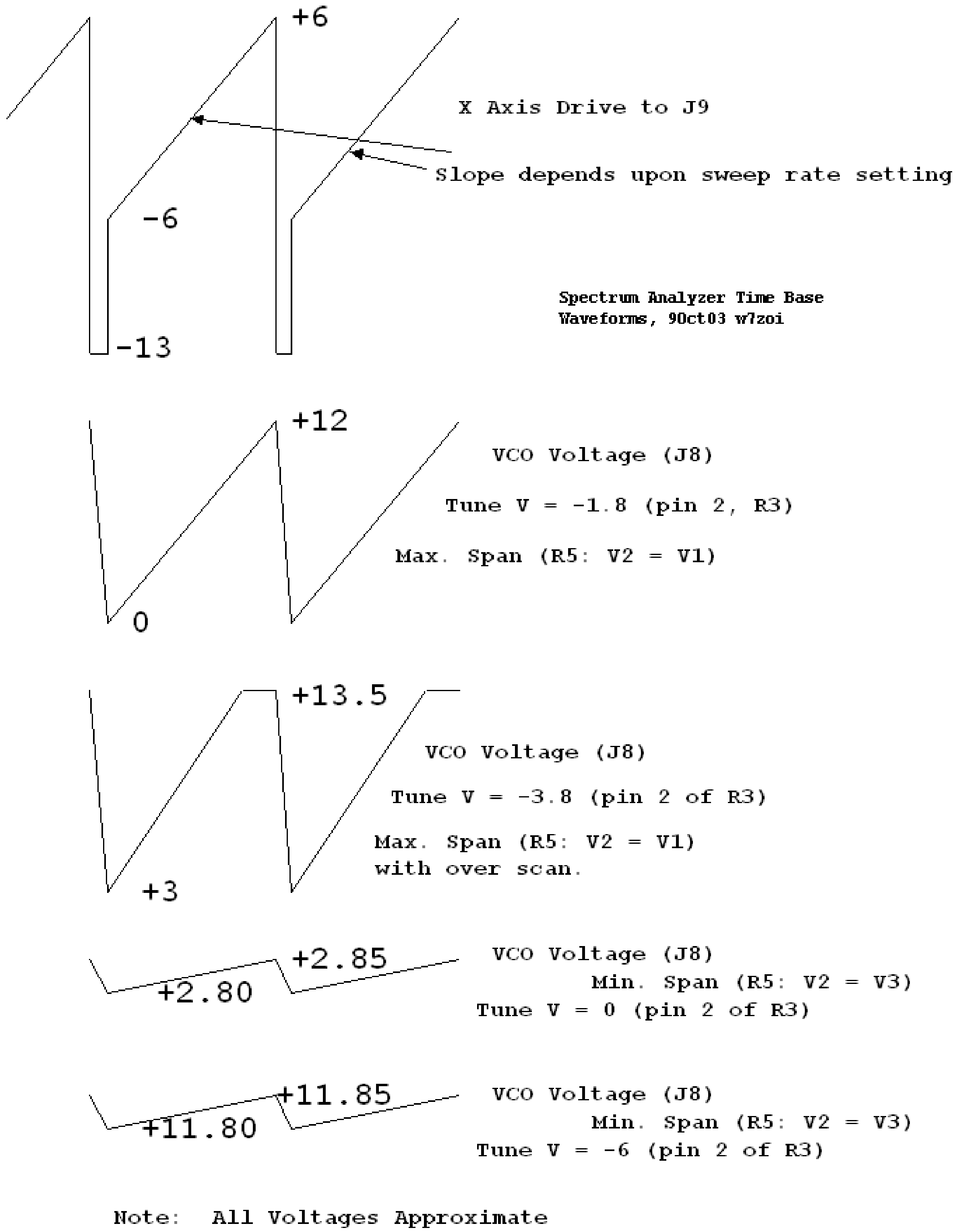
"Tracking Generator Modifications."

(TGMOD.pdf) **(31August06)** **[Update -- 5Jan07]**


See the file **"VHFBPF"** for information on an experimental bandpass filter.

Timebase Waveforms. **27Dec03**

Some builders have written with questions about the proper wave forms for the time base. Here are some that we measured.



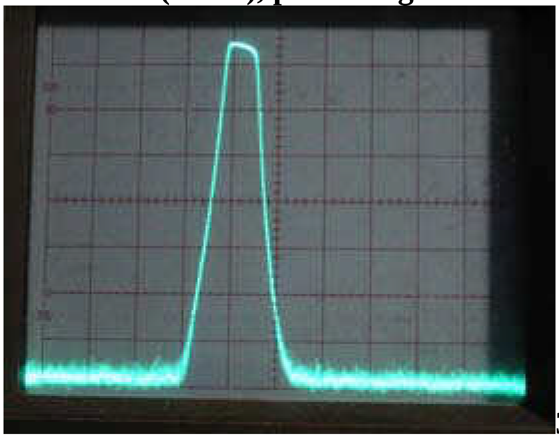
Note that the details will depend on the frequency tuning.

Regarding the monolithic crystal filters: Mouser Electronics lists the 520 -107-15B as a four pole filter. This is actually two 2-pole filters sold as a unit. See the **Kanga** page for information on using these filters in the spectrum analyzer.

"Ringing Distortion" **(April 6, 2003)**

Several builders have written to ask about an apparent distortion that they see when using the narrow resolution bandwidth in the spectrum analyzer. This is not really a distortion. That is, the display seen is a natural consequence of interaction between filter bandwidth and sweep speed. If the sweep is too fast, the full amplitude of the signal does not appear on screen and the shape of the peak is distorted from the actual filter shape.   The interaction results from both narrow resolution filters and narrow video filtering.   Some of this behavior is illustrated in the photos below with my analyzer.

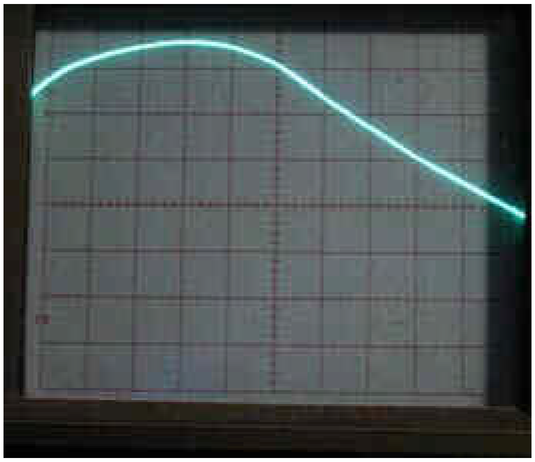
A signal was injected from a signal generator and the amplitude was adjusted for a response near the top of the screen. The analyzer was set for a 30 kHz resolution bandwidth (RBW), producing this result:



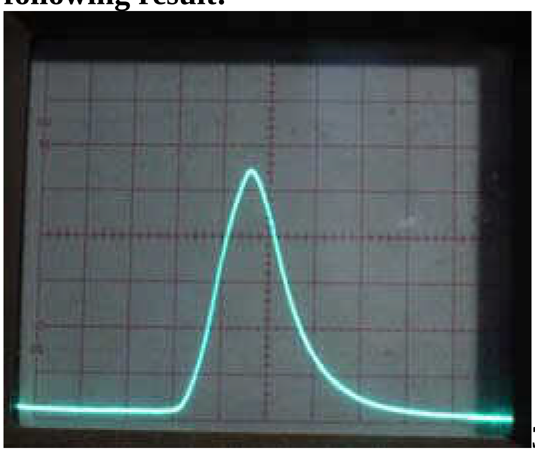
30 kHz RBW

The span was expanded so we were tuning over a range of less than 500 kHz.

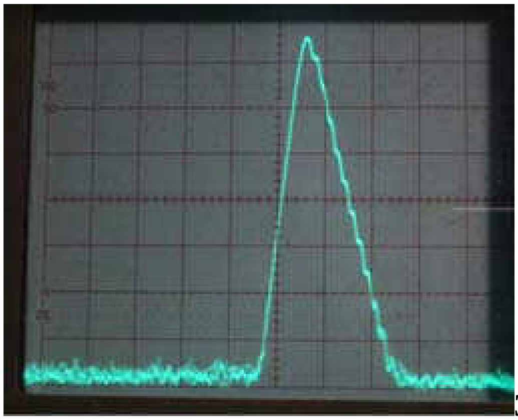
Switching RBW to 300 kHz produced the following response.

 300 kHz RBW.
With the wider RBW, we don't even see the stopband of the filter.

Returning to RBW of 30 kHz, we switched in the video filter to produce the following result:

30 kHz with narrow video filtering. Two distortions are seen. First, the amplitude is about 20 dB low. Second, there is a tail on the right side. Both problems disappear as the sweep rate is decreased. In the extreme, my analyzer will produce 2 or 3 sweeps per second. But this makes it very difficult to photograph the result.

The next experiment removes the video filtering and expands the span with a faster sweep rate. The result is a narrow line on the screen that seems to be distorted. It's was difficult to see in detail until the horizontal sensitivity of the monitor oscilloscope was changed from 1 volt/div to 0.1 volt/div. The result is then:


30 kHz with fast sweep and no video filtering.

This display shows shape distortion, although the peak amplitude is nearly correct. The behavior is classic "ringing," a phenomena we often hear in our receivers. In this case, the analyzer sweeps until a signal appears to dump energy into the filter. But the high sweep rate causes the trace to move to the right before the energy is dissipated. This is the classic effect we hear when a bell is rung.

These are all natural consequences of filter behavior and do not indicate a problem with the analyzer. Commercial spectrum analyzers get past the problems with different displays. Some use a CRT with a long persistence phosphor. Others use special CRTs that will store an image for a while. One popular format uses a so called "variable persistence" presentation. Modern analyzers use a digital storage display. The user will usually never see the problems outlined here, for the sweep rate will be automatically adjusted to be commensurate with the span and bandwidth.

If you have built a homebrew analyzer based on ours, you will have to reduce the sweep rate when the distortion appears. Be careful not to use excessive video filtering. Some builders have used a capacitor with an excessive C at C317 at the low amp output. (This is the original log amp shown in the QST paper.)

The formal details relating span, sweep rate, and bandwidth are presented in the text by Morris Engelson, "Modern Spectrum Analyzer Theory and Applications," Artech House, 1984.

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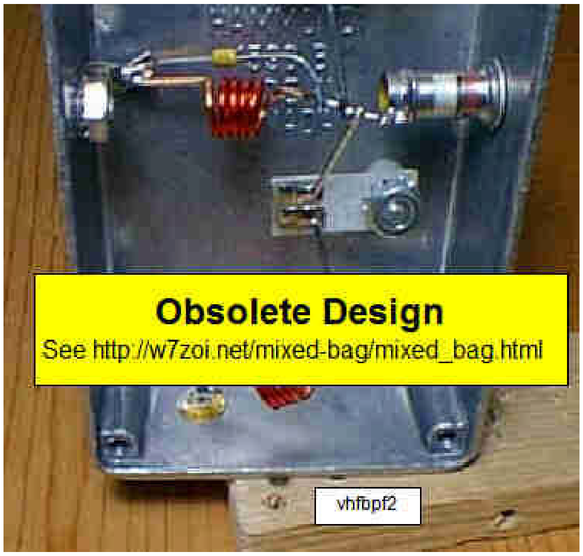
110 MHz Bandpass Filter.
(July 21, 2002)

The 110 MHz bandpass filter has been one of the more difficult construction tasks related to the spectrum analyzer project. The most common mistake seems to be in literal interpretation of the schematic diagram. That is, the schematic (Figures 1, 2, and 4 of the article in QEX, March/April 1998 and Figure 9 in the QST paper of August, 1998) showed coupling capacitors of 0.23pF. This would work just fine if (a) we could easily obtain these low values and (b) there was no additional coupling. But both problems are present, so we must resort to experimental methods. These methods are outlined in the QEX paper.

The photo below shows a view of the bandpass filter. The middle tuned circuit and a resonator at one end are shown. The end resonators are close to their respective walls while the middle one is physically close to the box middle. There is INDUCTIVE coupling between coils, but of an unknown amount that is not well modeled. This means that the 0.23 pF values for capacitive coupling are not valid. But this makes no difference. It DOES NOT MATTER how we get the coupling. All that is required is to adjust the net coupling to produce the desired bandwidth. Filter bandwidth is also controlled by the end section loading.



An end section and middle resonator for the 110 MHz triple tuned filter.

In this realization of the filter the coupling capacitor is built from wires that attach to the resonators. These wires are soldered to the tuning capacitors and to the small scrap of circuit board. The circuit board is nothing more than a handy place to physically hold the wires in a fixed position.

We used some multiple turn glass piston trimmer capacitors. The bandspread afforded by these parts is very useful, allowing easy tuning. Other trimmer caps may be much more difficult to adjust, especially if they are all internal to the box. Through-the-wall mounting is preferred.

Remember that this filter is only 2 or 3 MHz wide. A bandwidth this narrow is needed, for this filter must provide a stopband attenuation of 90 dB or more at the 90 MHz image related to the second mixer in the SA. This makes adjustment challenging.

The filter used in our SA was measured with both a network analyzer and then with a signal generator and spectrum analyzer. The network analyzer measurements, shown in

Fig. 5 of the QEX paper, confirmed the shape of the peak. The signal generator measurements with another spectrum analyzer confirmed the needed stopband attenuation. This measurement can also be done with a sensitive power meter such as the one that W7PUA and I described in June 2001 QST. The filter shape was quite clean with no other peaks in the VHF region. Some peaks occurred at UHF resulting from the coils acting as helical resonators with a length of several odd multiples of a quarter wavelength. These responses are of little consequence, for the front end low pass filter keeps such signals from reaching the bandpass filter.

This filter design is something of a "trick." That is, we were able to realize a working design while using but one enclosure. The coils had to be small to realize this, but not too small, for that degraded Q. A more refined triple tuned bandpass filter design is presented on this web site and can be adapted to 110 MHz if desired. No matter what you build for the spectrum analyzer, we hope that you will study the more general design methods.

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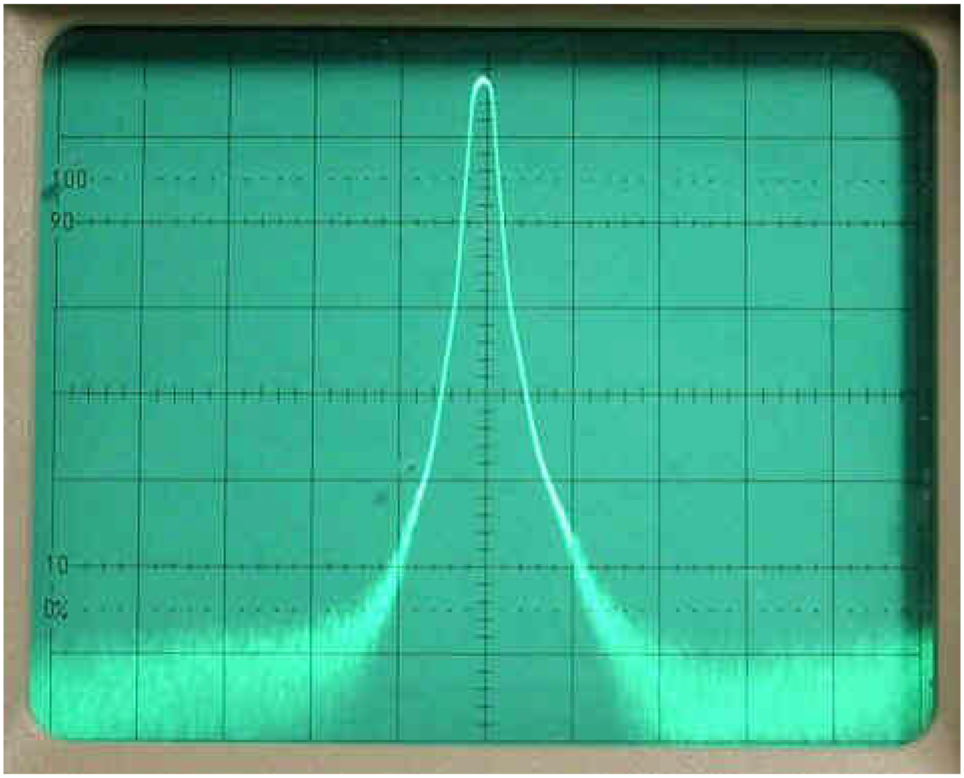
110 MHz Bandpass Filter Update, January 7, 2004:

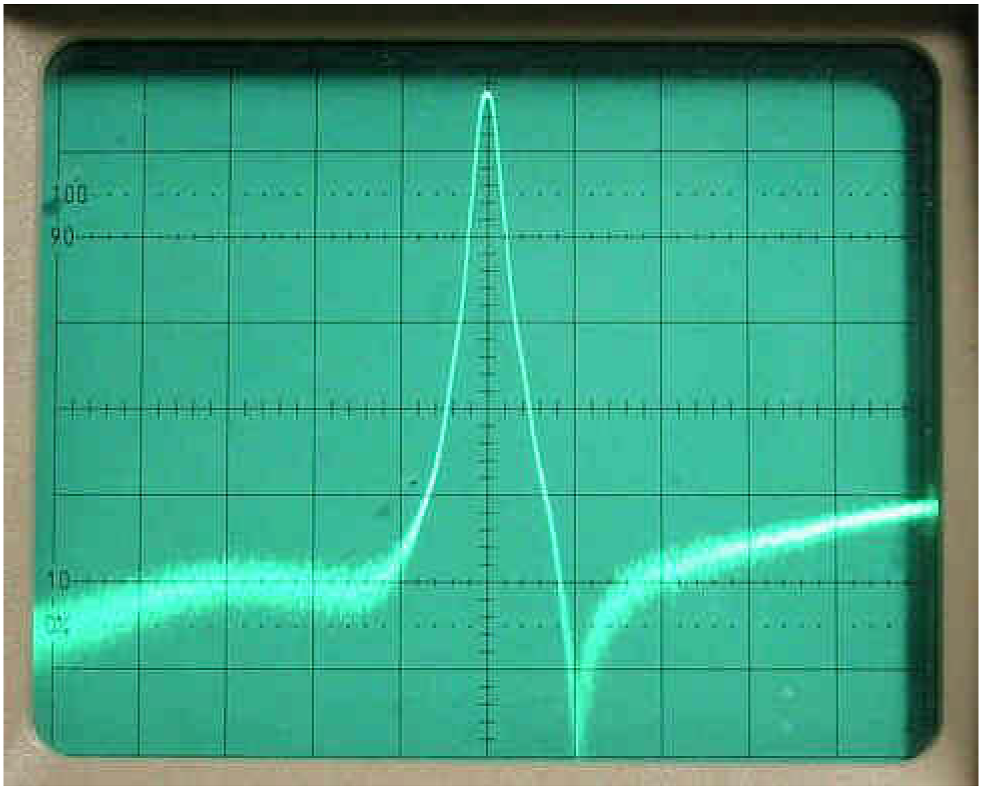
Some builders have been encountering problems with this filter. In a sense that is not surprising, for this filter is probably the most difficult circuit used in the spectrum analyzer. In at least one case, the difficulties arose from changes in the grounding related to the two end resonators. We built a second version of the filter to investigate these problems.

This is an inside view of the original filter. Note that the end resonators are crowded toward the box ends. The angle is of no significance other than allowing a fit. The goal was to maintain as much isolation between resonators as possible, providing coupling primarily through the small interstage capacitors. The loading at the ends can be adjusted by altering the length of the one turn inductive loops while interstage coupling could be altered by removing some of the tinned circuit board area, or by adding wire to the floating pads.



This is the measured response for the filter. This was obtained with a VHF spectrum analyzer with a tracking generator. The system is set for a center frequency of 110 MHz and a span of 10 MHz per division. While there were compromises in the stopband attenuation, they did not generally occur until well past the VHF part of the spectrum.

This is an inside view of our modified filter. The inductive end loading wires are now straight with the ground point moved toward the center of the box. This will increase stray inductive coupling between resonators. When we first built the filter we found that we could not align it at 110 MHz. The center resonator would hit 110 (barely) but the ends were confined to 105 or lower. The filter was easily aligned at 100 MHz where the shape looked reasonable. We removed 1 turn from each of the three coils in order to move resonance higher. We also spread the turns slightly on the left most resonator. The overall result is now a filter that we could align at 110 MHz.



This is the filter shape we observed after tuning at 110 MHz. The sharp dip on the edge of the passband indicates coupling completely across the filter. The stopband attenuation is severely compromised, especially in the VHF area above the region show, which still uses a span of 10 MHz per division.

One builder, upon discovering problems, added shields to his filter, quickly regaining the poor stopband performance. Either kind of construction will work, although measurements will be required. A VHF spectrum analyzer and tracking generator are ideal instruments, but are certainly not required. We did all of the early work with nothing more than a signal generator and homebrew power meter. A power meter like those based on the AD-8307 (QST, June, 2001) is ideal.

Be sure to read the section preceding this one as well as the later one dealing with **Some Spurious Response Problems**.

The Spectrum Analyzer Zero Spur.

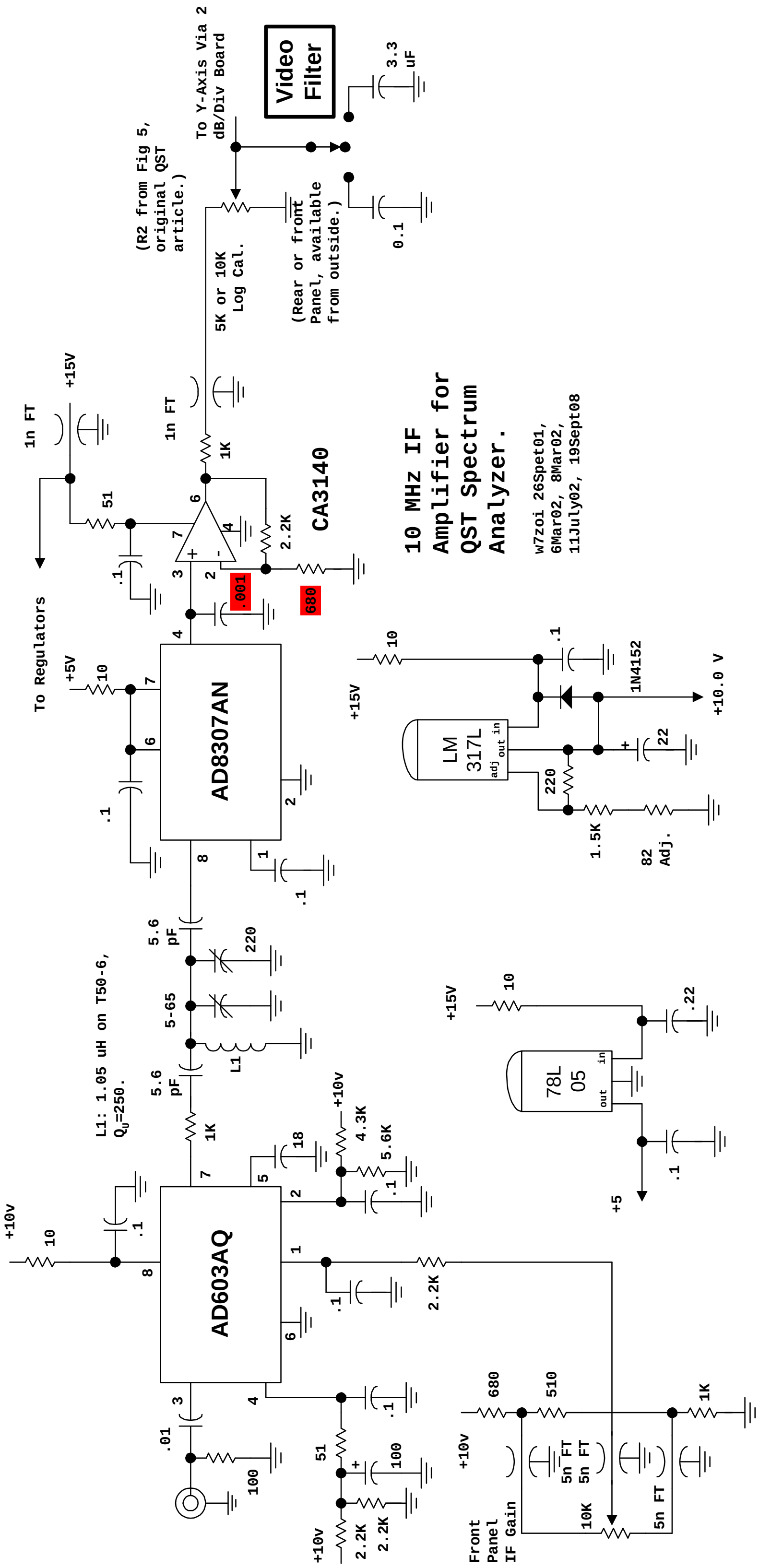
July 21, 2002.

All spectrum analyzers of this type have a large response at zero frequency. Think of the analyzer as consisting (in our case) of a fixed receiver at 110 MHz preceded by a tunable (swept) converter. The converter uses a mixer and an oscillator tuning from 110 MHz up to a higher value. The spur is merely the result of the LO sweeping through 110 MHz.

It may overload the analyzer. That is, the peak of the response may well be above the top of the screen. THIS IS OK. There is not enough power available to do any damage, and there is no signal in the downstream circuits when the first LO is above 110 MHz. Pay no attention to this spur when setting up the analyzer for a reference level response at the TOP of the screen. The magnitude of the zero spur is determined by the LO to IF port suppression of the balanced mixer used in the 2nd mixer position.

New IF and Log Amplifier (26Sept01, 6March02, 8March02, 12July02, 7April04, Mod 10Oct08.)

The original article presented data on a variety of integrated circuits that would function in the log amplifier. The Motorola MC3356 was the featured part, for it was available at the time we wrote the paper. The AD8307, which was still in a pre-production phase, was yet to become generally available. It is now available and is a far superior part. Hence, we dropped it into the analyzer. While doing the design work, we decided to try a new variable gain IF amplifier as well, the Analog Devices AD603. The overall circuit is shown below.



Note: This figure is included in Chapter 7 of <u>Experimental Methods in RF Design</u> (ARRL, 2003).

Changes of October 2008 are highlighted in Red.

Note that the feedback resistor around the CA3140 really is 2.2K and not 22K as was used in the original circuit. The new log amp has higher output than the old one.

The IF gain will be from 10 to 50 dB, depending on the setting of the control pot. This pot is on the front panel of our analyzer. This circuit replaces the circuitry in Fig. 5 in the original article.   The circuitry is built on a scrap of circuit board material mounted inside a Hammond 1590B box. The box output drives R2, the "log amp cal" pot, which then is routed to the added 2 dB per division board and then to the oscilloscope Y axis.

L1 is indeed 1.05 uH, as measured with an AADE LC Meter. Other powder iron toroids can be used, but keep Qu high. The bandwidth of this circuit, when using the two 5.6 pF coupling caps, is about 200 kHz and this filter is important in keeping the noise low in the log amplifier.

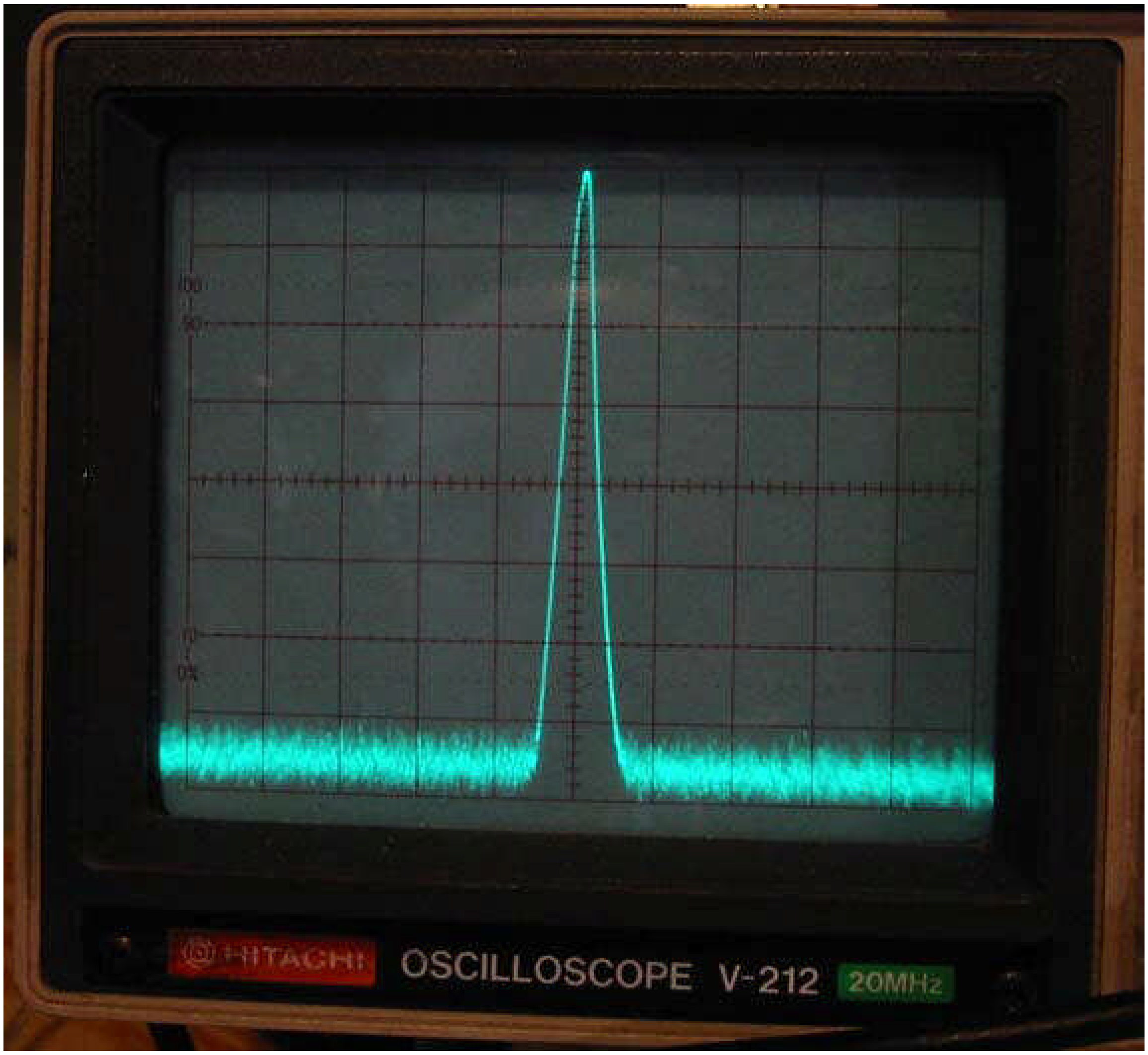
The AD603 was intended to be a video AGC amplifier for use in medical instrumentation. In this application, gain stability and accuracy are of great importance. Here we use the part in a narrow bandwidth IF system. As such, we can live with the slight gain slope at the operating frequency. See the AD603 data sheet on the Analog Devices web site. We do this all the time with simpler IF amplifiers. We have found the gain stability to be excellent in the analyzer at W7ZOI.   The tuned circuit at the '603 output has a loaded Q of about 50. It is tuned to 10.0 MHz in the circuit. Modifications will be required for slight IF changes, such as 10.7 MHz.

The front panel gain control pot is attached back to the IF through feedthrough capacitors in the box containing the IF. This is a place where you can probably get by without the expensive (and terribly difficult to find) caps. Be sure to decouple carefully and run all three lines from the IF amplifier board to the front panel without attaching to panel ground.

The board can be checked out before installation in the analyzer. A DC voltmeter is attached to the output and a 10 MHz signal is injected at the input and the tuned circuit is peaked.   The maximum output should be around 5 volts DC with an input signal of about -28 dBm. The alignment should happen at 10.7 MHz if this is used in your analyzer.   It may be necessary to adjust the resistors related to the LM317L regulator to obtain an output close to 10.0 volts.

The new log amp and IF is then dropped into the spectrum analyzer.   Alignment is again straight forward. A calibration signal of -30 dBm is applied to the analyzer input at 14 MHz and the analyzer is tuned to see that signal. The Log gain should be near maximum. The IF gain pot and the vertical position on the 'scope are tweaked to put the response at the top of the screen at the "Reference Level." Then, the attenuator is used to reduce the signal in 10 dB steps. Ideally, the on screen response should move down at 1 major division for each 10 dB step. It will probably move more than that. If so, reduce the log gain a little and increase the IF gain to again calibrate the instrument at the reference level. Keep repeating the process until you get a good match to the response on screen. It will be amazingly accurate with this AD8307 Log Amp IC.

Once the basic calibration is done, you will see that the IF gain and the vertical position can both be adjusted to move traces around. By moving toward highest IF gain with a downward oscilloscope offset, you will get the maximum on screen range. Leave enough gain adjustment left to take care of the narrower filter bandwidth when it is switched in . The noise floor is an indicated -105 dBm when in the 300 kHz resolution bandwidth position with my analyzer, set for a -30 dBm reference level.   (Note that this is 7 dB lower than measured when the noise filter bandwidth of L1 was wider.) It is possible to set the instrument up for a -20 dBm reference level, which may be more convenient for some measurements.



The above photo shows our spectrum analyzer with a -30 dBm 14 MHz input signal and no front end attenuation. The Resolution Bandwidth is in the widest position of 300 kHz with no video filtering selected. Although obscured by shadows, the peak is on the top line in the display. The noise fills about two thirds of one of the "boxes," a condition that is maintained as IF gain is added to bring the trace further up on the screen. There is about 10 dB more gain available in my IF system in this condition.   The noise is easily reduced to become a smooth, narrow line by adding video filtering. This also means that the sweep speed must be reduced. (As an aside, digital photographs are much easier with fast sweeps as shown.) Some experimenters have reported a changed character of the displayed noise when going to the newer AD8307 Log Amp system. This results from quite different terminations at the output of the respective log amps. It is often useful to have the ability to do faster sweeps, even with the higher noise levels.

The AD603 and the AD8307 integrated circuits were purchased from the Analog Devices web site. The '8307 arrived right away, but the '603 took several months. Hopefully the parts are now available "off the shelf."

If you don't want to use the AD603, the amplifier portion of the original circuit can be adapted. A bit more gain may be needed, which could be added with a simple feedback amplifier.

6April 2004: Several builders have fabricated modified versions of the analyzer that cover wider frequency spans. This results in faster sweeps in MHz per second, producing filter responses that are compressed. There are a few things that we can do to get past this problem. First, replace the capacitor on the output of the log amp. It's now a .01 uF. Drop to 1000 pF or even something as low as 100 pF. We probably want a little bit of video filtering here, but not as much as afforded by .01 uF. Video filtering can then be added later as needed during narrow sweeps. Another thing that I have done with my own analyzer is to change the main sweep capacitor, C401 of Figure 2 in the original QST paper, August 1998, p36. This capacitor was 1 uF, non-polar. I've gone up to 10 uF. This slows the sweep accordingly, but allows a cleaner on screen response. Finally, you may want to decrease R417 which would allow a narrower minimum span. See what works for you and your application.

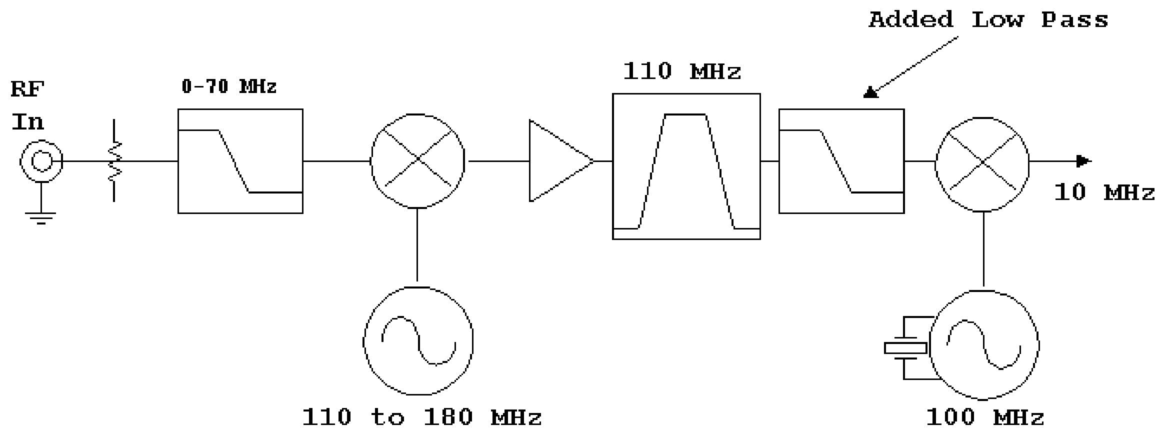
The AD603 IF amplifier used in our analyzer uses a 1K output resistor to set a termination impedance. Other values are certainly possible. If the value is dropped to much lower values, the available power gain increases considerably. This merely means that the user will then back off on the IF gain during normal operation. Some redesign in the filter will be required, an option useful to those building other related, but significantly modified spectrum analyzers.

Some Spurious Response Problems, 21 August 02.

We have recently heard from two different builders who have encountered an interesting spur problem with the recent QST Spectrum Analyzer (QST, August and September, 1998.) The most recent correspondence was a very detailed e-mail from Bill Evans, W3FB. He had built the analyzer using the Kanga boards and Hammond boxes. He had used quality feedthrough capacitors and SMA connectors throughout and had the system running. However, two spurs occurred high in the band at 76.25 and 78.75 MHz. Although the VCO in my own analyzer just allows tuning this high, the low pass filter response begins to role the response off at 70 MHz.

Bill reported that the two spurs he saw were roughly equivalent to -80 dBm signals. Terry (K7TAU) and I saw no spurs in our analyzers.

The source of the spurious responses becomes more clear after a careful examination of the design. The front end is shown below: (Ignore the "added low pass filter" for now.)



The incoming signals are up-converted to 110 MHz where they are bandpass filtered in a third order bandpass filter with a bandwidth of a couple of MHz. The result is then down converted to 10 MHz. The first LO is swept while the second is fixed at 100 MHz.

The spurious responses are reported to appear narrow. This indicates that they result from harmonics of the swept local oscillator. A harmonic will sweep through a fixed frequency N times as fast as the fundamental where N is the harmonic number. Experiments with the analyzer suggested that the frequencies were fairly stable. They did not change in frequency as one probed the circuitry with a finger, although amplitude changes could be produced.

It was speculated that harmonics of the 100 MHz 2nd LO were leaking into the front end. They would then be converted down to appear in the response band. The figure below is a spread sheet showing the harmonics of the VCO when it is tuned to the spur frequencies of 76.25 or 78.75 MHz.

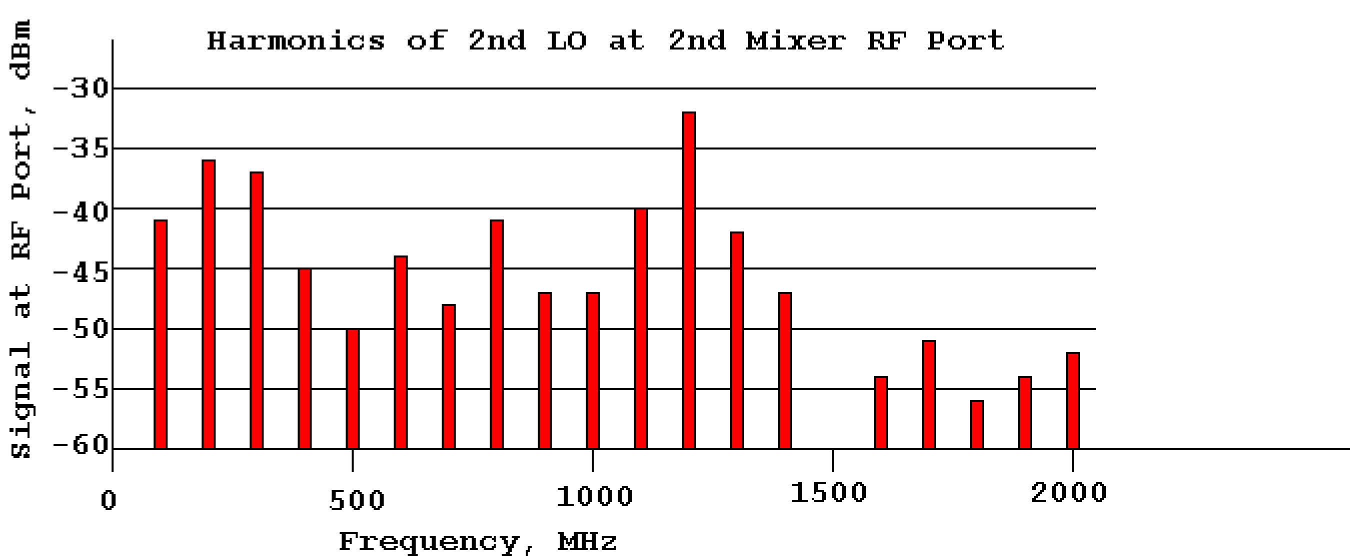
SA_SPUR.WB1

	A	B	C	D	E	F	G	H	I	J	K	L	M
1	SA Spur Analysis, 18 Aug 02 w7zoi												
2	IF1=	110	Spur F =	78.75	LO F =	188.75		IF1=	110	Spur F =	76.25	LO F =	186.25
3													
4	N	NxLO	RF+IF	RF-IF				N	NxLO	RF+IF	RF-IF		
5	1	188.75	298.75	78.75				1	186.25	296.25	76.25		
6	2	377.5	487.5	267.5				2	372.5	482.5	262.5		
7	3	566.25	676.25	456.25				3	558.75	668.75	448.75		
8	4	755	865	645				4	745	855	635		
9	5	943.75	1053.75	833.75				5	931.25	1041.25	821.25		
10	6	1132.5	1242.5	1022.5				6	1117.5	1227.5	1007.5		
11	7	1321.3	1431.25	1211.25				7	1303.8	1413.75	1193.8		
12	8	1510	1620	1400				8	1490	1600	1380		
13	9	1698.8	1808.75	1588.75				9	1676.3	1786.25	1566.3		
14	10	1887.5	1997.5	1777.5				10	1862.5	1972.5	1752.5		
15	11	2076.3	2186.25	1966.25				11	2048.8	2158.75	1938.8		
16													
17				1400						1600			

Note the left case showing a spur at 78.75. The LO is 188.75 for this example, with the 8th harmonic occurring at 1510 MHz. This will mix with a 1400 MHz signal to produce a 110 MHz output, which is the spur.

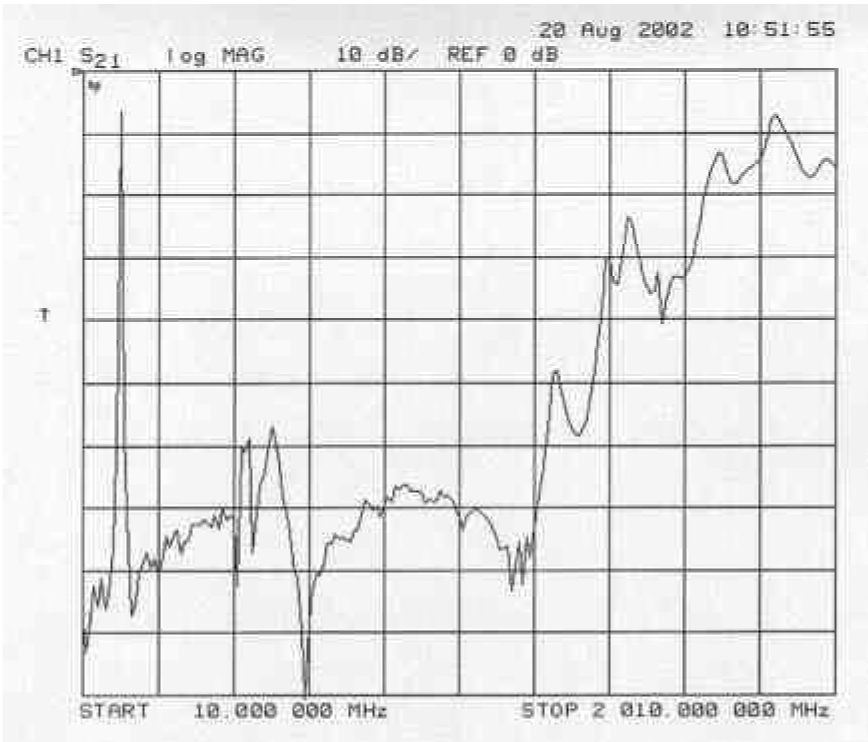
But why should these responses occur? We would hope that the bandpass filter would prevent this from happening, blocking 1400 MHz signals from reaching the first mixer module. But what if the stopband attenuation of this filter is less than ideal?

A drive out to TriQuint Semiconductor and Terry's measurement tools produced the needed results. First we measured the available harmonics from the RF port in the 2nd mixer. These are shown below:



Signal levels from –32 to –67 dBm (and less) are available at the various harmonics. Although the 100 MHz component is well suppressed from the +17 dBm level applied to the mixer (here by 58 dB,) the harmonics are still numerous and fairly strong at the RF port.

The more interesting measurement is shown below, a swept response of the bandpass filter from 10 to 2010 MHz, obtained with an HP network analyzer:

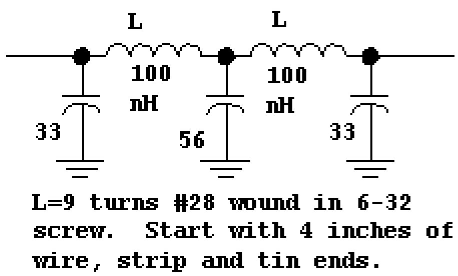


The desired response at 110 MHz is shown and is selective enough to provide the needed image attenuation. However, the stopband attenuation begins to degrade above 400 MHz and is really very poor above 1200 MHz. This is the result of the open box construction that was used. (See the photo near the top of this section of the web page.) Open construction worked for this particular filter because the coils were small and well spaced from each other. The 110 MHz energy is attenuated in the box when no resonators are present, for the box behaves like a waveguide that is too small for the frequency, propagating only in an evanescent "waveguide beyond cutoff" mode. However, at higher frequency the box is no longer too small. Indeed, as we get close to 2 or 3 GHz, the box will become resonant in a non-TEM waveguide mode.

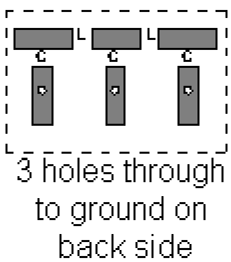
How do we fix the problem? One solution is to use a different bandpass filter. The early analyzers that Terry and I built both used three section helical resonator filters. These were early developmental "bench sweepings" components eventually used in the Tektronix 492 series of Spectrum Analyzers. Those parts used true helical resonators with aperture coupling between tuned circuits. Although not perfect, the stopband attenuation is much better than that of the filter shown above. A three section filter for

260 MHz is shown elsewhere on this web page and could be adapted to 110 MHz, offering an ideal solution.

Another solution should be the addition of a simple low pass filter in cascade with the 110 MHz bandpass. This would best be built using SMT components. If you can't find 100 nH inductors, they can be wound with #28 wire. Extremely short lead lengths should be used, especially for the capacitors. Remember that a typical 39 pF chip capacitor will have a series resonant frequency around 900 MHz. A suitable 5th order low pass filter is shown below:

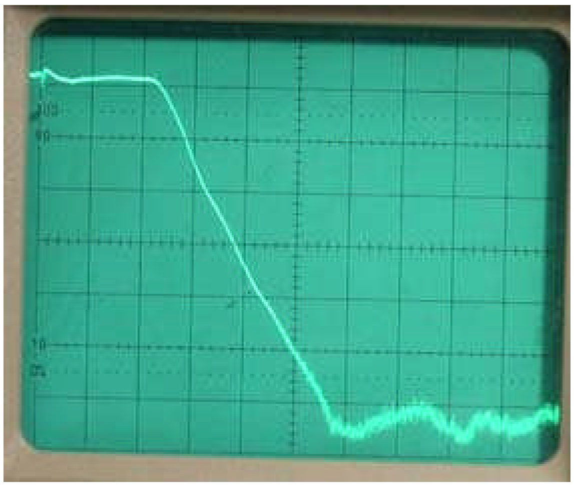


We built a version of this filter using surface mount technology, SMT, with the layout shown.



It is important to drill holes in the ground runs from the capacitors that are close to the capacitor body. If the runs are too long, the excess inductance in series with the capacitors will compromise filter shape. The layout shown uses a 1 inch square of double sided circuit board material. After the filter is built, this piece is one side of a box that houses the filter. I used some 0603 size Coilcraft inductors from an available kit. A good part to order might be a slightly larger Panasonic 5% 0805 part, DigiKey part number PCD1172CT-ND.

 This design used no connectors. The photo shows the ground side of the board. The measured response from DC to 500 MHz is shown below.

 The dip near 25 MHz is a result of the spectrum analyzer and tracking generator used for the measurement.

Understanding and isolating spurious responses is one of the greatest headaches of the spectrum analyzer (and receiver) designer.

Many thanks to VE7CA and W3FB for their observations and interactions!

Calibration of the 1998 QST Spectrum Analyzer

September 19, 2008, updated October 16, 2008. Wes Hayward, w7zoi

Some recent correspondence generated questions about the calibration of the August 1998 QST spectrum analyzer.   We were not as complete as we should have been in specifying the way the instrument was calibrated.   Hopefully, presenting this information will help to fix this problem.   This note describes the CAL procedure I used with my own version of the instrument.

The Goal

What we mean by "calibration" in this case is to set up the spectrum analyzer so we can meaningfully measure a signal ***amplitude***. Frequency calibration would be nice, but this is a feature that we bypassed in this instrument.

The instrument is designed to have a reference level of -30 dBm when there is no attenuation in front of the first mixer other than the input low pass filter. Recall that ***reference level*** is the analyzer input power corresponding to the ***top*** of the CRT screen. I usually use the instrument with at least 10 dB attenuation during most measurements, but that is not required for calibration using 50 Ohm sources. Not only do we want to have the top of the display correspond to -30 dBm, but we want all on-screen signals to be measurable by inspection of the screen. This means that a signal on screen should drop by 10 dB per major division as it is attenuated. So if we see a signal that is two major divisions below the top line, we want to be able to call that -50 dBm with confidence. This *log compliance* is easily tested with a step attenuator with 10 dB steps.

Some folks merely use their analyzer as a relative measurement tool. That is, they adjust the gain and attenuation to place an arbitrary reference at the display top and then examine the relative responses, having set up the log gain to achieve the desired 10 dB per division. This is fine, although I like to take it further with regard to amplitudes. I usually run my analyzer set up for -30 dBm at screen top as a reference level. This is then offset with attenuation for the measurements of higher power. Virtually ALL commercial spectrum analyzers use a calibrated reference level. Even when the amplitude sensitivity is changed from 10 to 5, 2, or even 1 dB per division, it is done in a way that a known power is retained for the reference level.

The goal we seek is to generate an instrument that is a frequency selective power meter. The calibration procedure outlined provides that behavior.

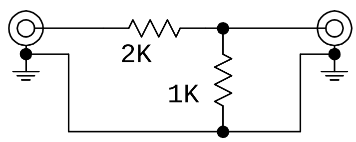
The reference level is what we see at the top of the screen. But what about the bottom? That's 80 dB lower, but those last few dB are obscured by the noise floor of the log system and noise in the IF. Optimization would allow one to extract more performance here. What we looked for here was not so much precise RF levels, but well defined DC levels and oscilloscope offsets during setup. The values suggested are those I found with my analyzer and are by no means optimum or unique. On the other hand, these levels should let the builder get started with greater ease.

Two Calibration Sources

The August 1998 spectrum analyzer uses an oscilloscope as the display. The 'scope is set up to operate in an XY mode with the time-base built into the analyzer. That time-base and VCO drive system establishes the tuning for the analyzer center frequency and the span. The scope Y sensitivity is set to 0.5 volt per division for the spectrum analyzer measurements.

There are two measurements that we will use in the calibration process. One applies a DC signal to the Y axis of the oscilloscope without a Y-channel connection to the analyzer. The other applies RF to the analyzer input with a well defined level.

The DC voltages needed are in the range from near zero up to 5. I have a simple variable voltage power supply that uses a LM-317 regulator. See EMRFD Chapter 1, p 1.15. That circuit has a minimum output of 1.25 volt. To get to lower voltage levels, I used a 3:1 divider consisting of 2K and 1K resistors. The ‘scope input impedance is 1 Meg, so the divider is more than stiff enough to drive the ‘scope Y axis.



The RF source I used is an HP-8640B signal generator with low harmonic distortion. This was set up to provide an output of -30 dBm. If you don’t have a good signal generator (yet), you can approximate it with a home brew instrument. Build an oscillator, probably one that is crystal controlled. Set it up for an output power of around 1 mW (0 dBm) into 50 Ohms. Then add a low pass filter so that the output is relatively pure. A 5th order LC circuit with a cutoff about 20% higher than the crystal will be just fine. Next build an attenuator into the generator to produce an output of -30 dBm.

The RF source must have a known power, so it must be pre-calibrated. One of many reasonable procedures to do this measures the output of the oscillator and low pass filter into 50 Ohms with your oscilloscope. Affixed the source to the ‘scope with 50 Ohm coax, using a 50 Ohm terminator right at the ‘scope input. Once the output power is established, calculate the resistor values needed for a 50 Ohm pi-pad that will produce an eventual value of -30 dBm. (EMRFD Fig 7.19, p 7.9) If you can’t hit the attenuation exactly, don’t worry; you can compensate for the differences later. For example, assume you ended up with an output of -2.8 dBm from your low pass filtered crystal oscillator. You would need a pi pad with an attenuation of 27.2 dB to get to -30 dBm. This pad would use a series resistor of 572 Ohms and two shunt resistors of 54.6 Ohms each. But all you can find in the junk box is a 620 and a couple of 56 Ohm resistors. This will yield a pad of 27.8 dB attenuation, producing an output of -30.6 dBm. Close enough.

For the present, assume we have exactly -30 dBm. We will make adjustments later, if needed.

Calibration

The amplitude response of any swept instrument, especially a spectrum analyzer, can be distorted by narrow bandwidth and/or excessively fast sweep speeds. So when doing calibrations, it is important to decrease the sweep speed to the point that the displayed signal is not changing in amplitude with sweep speed. If there is a display flicker, you will just have to live with it. In the same vein, set the resolution bandwidth to the wider 300 kHz and switch video filters out of the system.

The first thing done is to set up the level for the **bottom** of the screen. For this determination, we assume that the analyzer is approximately working. That is, we assume that you can see signals and determine that they are clean (or not) with regard to harmonics. Signals near the top of the screen, including the ubiquitous zero-spur, can be seen. Don't worry if the zero spur is over the top of the screen, for it usually is.

There are three pots in our overall system that must be adjusted: One is the position control on the oscilloscope associated with the Y axis. The other two are the gain controls in the spectrum analyzer. The first of these is R2, which is the so called Log Gain, or "Log Amp Cal" found in Fig 5 of the original QST article. Even if you have changed to the new AD-8307 based log amp, this gain control will still be present. The third pot is the IF gain control. This is R1 of Fig 5 of the original article, or the pot in the lower left corner of EMRFD Fig 7.53, p 7.28. All of these controls should be accessible from the outside of the spectrum analyzer and should best not be internal, screwdriver sets.

Disconnect the cable from the SA that goes to the Y-Input and attach the DC source to the oscilloscope. The spectrum analyzer should be running and the X-Input should still be attached to the analyzer. Then, adjustment of the DC supply voltage will produce horizontal lines with a vertical position that moves with the DC source. Set the DC source so that the voltage at the 'scope Y-Input is 0.6. Measure it with a voltmeter to confirm the reading. Then adjust the 'scope vertical position to place the resulting line at the bottom of the screen. (I'm assuming that your 'scope has 8 divisions and is set for a Y sensitivity of 0.5 volt /division.) Note that the calibration procedure called for in the original paper (page 38, 3rd column, 3.6 inches down from the text top) was to calibrate the bottom to be 0 volts. This procedure negates that older one.

While the DC supply is still connected, turn the DC power supply up to generate a line that is at the top of the screen. The DC voltage should now be 4.6 volts at the 'scope. If it is off a bit, it won't matter, for the slop will eventually be taken up with R2. If it is way off, you may want to investigate 'scope calibration.

We now set the DC power supply aside, attach the SA vertical output to the 'scope Y-Input, and attach the -30 dBm RF source to the analyzer RF input. My calibrator is at 7 MHz, but frequency is not important. I saw a major line appearing on screen at the 7 MHz position. I could also see a band of noise at the bottom of the display and the ever present zero spur. Adjusted the span so that the 7 MHz signal takes up about 10% of the screen and check the sweep rate to be sure that the amplitude is not changed by sweep rate. The resolution bandwidth should be at 300 kHz and there should be little video filtering. Twist both of the gain controls a little and note that they both will alter the amplitude of the -30 dBm signal on screen.

We now begin an iterative process that should, we hope, produce a calibrated spectrum analyzer. I'm going to number the steps that I took during the calibration of my analyzer with the understanding that you may take more steps, or fewer. Begin by turning the Log Gain (R2) to maximum. This will probability drive the 7 MHz, -30 dBm signal over the top of the screen. That's OK.

1. Reduce the IF Gain (R1) to put the 7 MHz signal exactly at the top of the screen. Now insert attenuation with your step attenuator into the analyzer RF input. Do this in 10 dB steps and watch the signal drop on screen. With 40 dB attenuation, I saw an **indicated** response that was 48 dB below the top. That is, the peak was now 4.8 major 'scope divisions down from the screen top.
2. Remove the 40 dB attenuation in the RF input. Decrease the Log Gain a little, which will drop the response a few indicated dB down from the screen top. Increase the IF Gain until the signal is again at the top of the screen. Now insert 40 dB of front end attenuation. This produced a signal that was at an indicated 46 dB below the top of the screen.
3. Remove the 40 dB attenuation and decrease the Log Gain a little more. Increase the IF Gain to bring the signal to the top of the screen. Insert 40 dB RF attenuation. I now saw a signal that was at -44 dB with respect to the reference level.
4. One more round produced a -42 dB response.
5. One additional iteration yielded -41 dB.
6. And finally, one final round produced a response at -40 dB with respect to the reference level. Moreover, the signal came close to hitting all of the major grid lines on the display as I ran the attenuation down in 10 dB steps to 70 dB below the top. The final step to 80 dB down was into the noise.

Refinements

The analyzer is now pretty well calibrated and should be a useful instrument. There are a few additional details and experiments that we did.

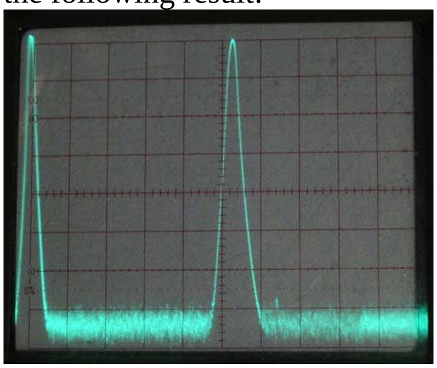
What if we didn't have exactly a -30 dBm source. The example that we used earlier was -30.6 dBm. All that is really required now is to attach this signal to the analyzer input and adjust the IF gain to put the signal 0.6 dB down from the screen top.

The system gain will probably change when the analyzer resolution bandwidth is changed. In my case, switching the 30 kHz crystal filter into the system drops the response by about 4 dB. This can be compensated merely by increasing the IF gain.

What if we use a different starting DC level for the beginning of the process? Essentially, this means that we offset the noise end of the display in one direction or the other. Having calibrated with a starting DC level of 0.6 volt, I proceeded to investigate a number of other starting levels. The first experiment used 1.0 volt as the DC level for the bottom of the screen. I then found that the noise was below the bottom most line. Moreover, there was some obvious gain compression at the top end, easily detected as a distorted shape of the top of he filter response.

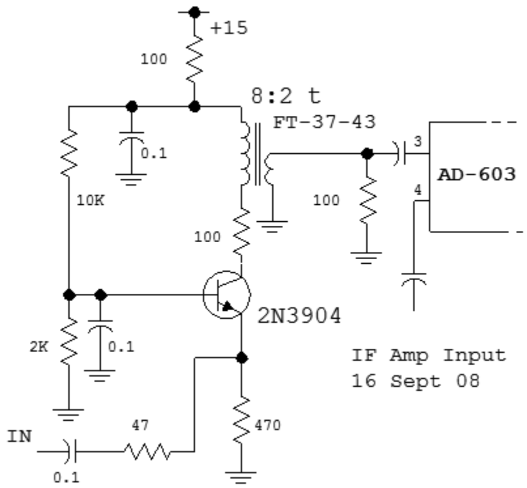
The next investigation used an especially low starting DC level of 0.42 volt. I found that I could easily obtain a clean calibration by jockeying the IF Gain and Log Gain pots. However, the noise floor was abnormally high.

I finally ended up with a starting DC value of 0.7 volt with my design. This produced the following result:



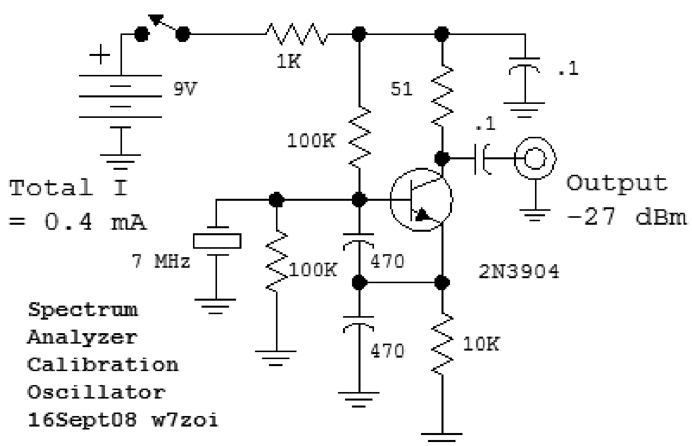
Spectrum Analyzer calibrated to a -30 dBm 7 MHz signal. RBW = 300 kHz. The left response is the zero spur, slightly over the top of the screen.

I had observed in the past that slightly more gain in the IF part of the analyzer would be useful. This was not vital, but it would have been handy for some experiments. I inserted an amplifier just prior to the input of the AD-603 variable gain IF stage. A common base stage was used, which offers good reverse isolation and very well defined 50 Ohm input impedance. This amplifier is:



Added amplifier at the IF Input. Net gain is about 9 dB. This precedes the circuitry of EMRFD Fig 7.53, also page 13 of this note.

I often find need for a general purpose oscillator that can be used to adjust receivers as well as a source to be used to calibrate the spectrum analyzer. The following circuit serves that purpose. The output is -27 dBm at 7 MHz. The harmonics at 14 and 21 MHz are also quite robust. Total current is about 0.4 mA for this block. Although away from the topic of this discussion, this circuit illustrates the interesting characteristic of the bipolar transistor oscillator that the DC emitter-base voltage is much less than the usual 0.65 volt that we might expect. The reason is that much of the biasing signal required for transistor operation is supplied by the RF signal, so much less is required of the DC bias circuitry. In this case, we obtain a good match between simple bias calculations (BIASnpn.exe, EMRFD CD/software/ladpac2002) and measured total current only if we set V(e-b) to 0.25 volt.



Spectrum analyzer calibrator and general purpose signal source. The output is -27 dBm at 7 MHz with low order harmonics that are only a few dB down from this level. The output impedance is well defined by the 51 Ohm collector resistor.

Many thanks to all the builders who have offered feedback.