Linrad: *New Possibilities for the Communications Experimenter, Part 2*

From the Analog World into the Digital: How do we get the desired signal from RF to the sound card?

By Leif Åsbrink, SM5BSZ

Linrad is a very flexible computer program that can work with more or less any digital data stream that can be moved into the computer. This is not unique to *Linrad*. Once the signal is present in a digital format, the flexibility of digital systems should make it reasonably straightforward to interface it to any software-defined radio. Bringing the radio signal into the digital world can be done in many different ways. The hardware descriptions below and the discussion of their advantages and disadvantages are an introduction to *Linrad*, but everything is fully applicable to any other software-defined receiver.

Jaders Prastgard 3265
63505 Eskilstuna, Sweden
leif.asbrink@mbox300.swipnet.se

The Filter Method

Part 1 of this series pointed out that frequency conversion and A/D conversion are closely related. One way of understanding that is to think about what would happen if the output of an A/D converter were routed to a D/A converter without any digital processing at all in between.

Half the frequency of the sampling clock is called the Nyquist frequency. It limits the highest frequency that can be represented by the digital data. The A/D converter is like a frequency mixer in which the LO is a signal at the Nyquist frequency and at its overtones. As in any other mixer, an output at a frequency f may be caused by an input at the frequency $(LO - f)$, $(LO + f)$, $(2LO - f)$, $(2LO + f)$, $(3LO - f)$ and so forth. In the analog world, one of the pair $(LO - f)$ and $(LO + f)$ is usually the desired signal while the other of this pair is the mirror-image frequency. The remaining false responses are denoted *spurs*, and they are weak if the analog mixer is good. In the digital world, all the unwanted responses are called *aliases* and many of them may be equally strong.

It is well-known that RF filters are needed in an analog radio to suppress the mirror image and spurs caused by the first LO. For an A/D converter, the corresponding filters are known as *anti-aliasing* filters.

If the RF filter were tuned to a response caused by an LO overtone, as is sometimes done in microwave receivers, the analog mixer would be called a harmonic mixer. If the filters in front of an A/D converter selected one of the responses above the Nyquist frequency, the A/D would be *under-*

sampling. Under-sampling is done to convert signals above half the maximum sampling speed for a particular A/D converter. It only works if the analog bandwidth of the A/D is high enough, typically several times higher than the maximum frequency for the sampling clock. Such A/D converters are known as radio A/Ds.

It is up to the system designer to decide what frequency range is desired and how much undesired signals must be suppressed. The problem is the same for an A/D converter as for an analog frequency mixer.

In the case where the lowest response, dc to the Nyquist frequency, is selected for an A/D, one needs a steep IF filter to suppress the image frequency that occurs in the mixing process from RF to audio. No filter is perfectly steep, so one will typically lose 15% of the possible digital bandwidth or more. Using an IF filter so that low frequencies reaching the A/D only come from IF signals at one side of the last LO (the BFO) is called the *filter method* because a qualified filter is necessary. The anti-aliasing filter needs a similar steepness, but since it is at a low frequency, it is comparatively simple. When using the filter method, one typically uses the IF-filter to improve alias suppression also.

The Phasing Method

The phasing method or direct-conversion method does not need any qualified filters. It is similar to transmitters wherein the phasing method can be used to generate SSB without crystal filters. By generating the complete (complex) baseband signal from the RF (or IF) signal, one feeds the computer with enough information to decide—for each signal—from which side of the BFO it comes. To do this, one needs two A/D channels. Each of them needs an anti-aliasing filter, but that is a simple filter—often included in the A/D converter board.

In the phasing method, two mixers in quadrature (the relative phases differ by 90°) generate an I/Q pair. A signal 1 kHz above the LO will produce a signal of 1 kHz with equal amplitude in both I and Q and their relative phase will be 90°. The computer program will combine them and correctly place the signal 1 kHz above zero. The analog circuitry suffers from tolerance problems that create a mirror-image spur at –1 kHz with an amplitude that is typically 30 to 40 dB below the main signal. It is possible to balance out this mirror image with analog phase and amplitude controls, but it can be dealt with equally well in the digital processing.

Desired Bandwidth Selects Technology

When selecting hardware to convert from RF to digital data, the most important aspect is the bandwidth. For many purposes, it is enough to move a bandwidth of 2 kHz into the computer. That is extremely easy: A conventional SSB radio and a simple audio-interface board will do the job. The dynamic range within the 2-kHz passband will be very poor, but the tolerance to strong signals outside the passband is at least as good as the analog radio specifications.

If more bandwidth is desired, say 10 kHz or so, one needs a solution with much better linearity than the product-detector and audio-amplifier circuitry of a normal receiver. The easiest way of doing this is to use the filter method with over-sampling. One uses an IF filter to select a passband that comes out as frequencies between something like 0.45 and 0.85 times the Nyquist frequency. The main problem of audio nonlinearities and overtones then won't be present and one can use simple circuitry with good results. With a 96-kHz audio board, one can easily get a bandwidth of about 20 kHz this way. The digital signal will be over-sampled but the extra load on a reasonably modern computer is not a problem. It is possible to get 20-kHz bandwidth using the filter method while sampling at 44.1 kHz, if a special crystal filter is used.

With PC sound cards, it is possible to get bandwidths up to about 95 kHz. If a radio A/D were used, the bandwidth limitation would come from the processing capabilities of the digital system and from the analog filter in front of the A/D converter. An AD6644 sampling at 65 MHz, preceded by a crystal filter with 200-kHz bandwidth at 10.7 MHz, 70 MHz or any other frequency, would make an excellent IF-to-digital converter. As far as I know, there is no standard for interfacing a radio A/D to a PC. The rest of this article shows what we can do with the sound cards that are around now.

The crucial factor is the desired bandwidth. For normal usage as a radio receiver, Table 1 gives an idea how dynamic range requirements increase with bandwidth.

It is extremely easy to make an RF-to-audio converter with a dynamic range of at least 40 dB, independent of the bandwidth. If *Linrad* were used to monitor microwave bands, low-cost solutions for the hardware would be appropriate. Occasional occurrences of strong local signals causing a few spurs on the screen then would not be a problem.

Below is a discussion of some circuits I have used to move radio signals into a PC via sound cards. Any of these solutions will work fine with *Linrad* if your local radio environment is compatible with the dynamic-range limitations of the particular solution.

The cost of increasing the bandwidth while maintaining adequate dynamic range is substantial. It is great fun to monitor nearly 100 kHz on the screen but for practical operation, chasing DX stations in a contest or just ordinary chatting, a bandwidth of about 2 kHz might be fully adequate for CW.

The main reason for bringing large bandwidths into the computer is that it allows the computer to do efficient noise blanking in a way that analog systems can never do. In a location where a better noise blanker than the one included in your current SSB receiver is not needed, *Linrad* will work happily with the output from your SSB receiver and allow you to dig out weak CW signals from the noise.

If power-line noise is not a problem, a 10-kHz bandwidth may be enough. With very difficult power-line noise problems, even 100 kHz is marginal because the interference pulse rate can become very high.

Table 1

Bandwidth (kHz)	*Dynamic Range (dB)*	*Comment*
2	40	Analog hardware stops strong signals.
10	70	Offending signal quality often makes better performance useless.
20	80	Having to keep really strong signals outside the passband is not a serious limitation at modest bandwidths. Is often necessary anyway because of offending signal quality.
100	100	With a large bandwidth one has to allow the local strong stations within the passband.

2-kHz Bandwidth: A Conventional SSB Receiver

It may be a good idea to insert a high-pass filter between the radio and the sound card to reduce 50/60-Hz hum. The audio signal that is routed to the sound card is actually an IF signal and it is a good idea to shift it upwards by a few kilohertz as one usually does when working high-speed meteor scatter. Placing the passband in the range 3-5 kHz rather than 0.3-2.3 kHz by a shift of the BFO greatly improves dynamic range. The dominating nonlinearity is usually IM_2 in the product detector and audio amplifiers, and it will be completely eliminated this way. The lowest possible IM_2 product would be 6 kHz, and that is above the passband. By setting the sound card to sample at 15 kHz or above, all overtones of 3 to 5 kHz are stopped by the anti-aliasing filter or by the digital processing. A shifted audio range also helps to get rid of overtones of the power-line frequency, which often go as high as the tenth harmonic and above when the volume is high and the RF gain is low. Those are the settings for a good dynamic range into the computer. A resistive voltage divider at the computer input is necessary to set the level low enough for the sound card. The volume usually cannot be set low enough without loss of dynamic range because of the noise in the AF amplifier after the volume control.

10-kHz Bandwidth: The Filter Method with Oversampling

Bob Larkin, W7PUA, has designed the DSP-10 to use the Analog Devices EZ-Kit for all the signal processing while the PC is used for control only. The analog hardware of that radio would work well for use with a sound card and PC replacing the EZ-Kit. The schematics (described in *QST,* Sept 1999, pp 33-41) include a crystal filter with the –6-dB points at 19.6591 and 19.6708 MHz. With the last LO at 19.680 MHz, the audio passband is placed from 9.2-20.9 kHz and that fits well to a standard sound card sampling at 48 kHz. Alias signals originate in audio frequencies above 27.1 kHz, and such signals are well suppressed by the anti-aliasing filter of the sound card. Such high audio frequencies originate in IF signals below 19.6529 MHz, so the IF filter gives an additional suppression of about 25 dB. Very strong signals at 19.670 MHz produce audio at 10 kHz with an overtone spur at 20 kHz that will be falsely interpreted as a signal at 19.660 MHz, just at the opposite end of the passband.

This is not a problem: One can simply avoid using the highest frequencies within the passband to receive weak signals. It is also possible to reduce the level of this spur by shifting the last LO downwards from 19.680 kHz until the alias spur starts to be visible at a similar IF level as the IM_2 spur. I have not tried this solution myself, but I see no reason why it should not be perfectly adequate as long as a 10-kHz bandwidth is enough.

20-kHz Bandwidth: The Filter Method without Oversampling.

My first DSP radio in a PC used this method. At that time, standard audio boards were limited to a maximum sampling speed of 44.1 kHz. I needed as much bandwidth as possible for two reasons. First, I suffer from terrible power-line noise now and then. The analog noise blankers do not work when the moon is low, and my local EME friends become very strong, so I needed something clever in the computer. Second, seeing a 20-kHz bandwidth rather than only 10 kHz is an advantage in EME contesting.

To get 20-kHz bandwidth with a sampling rate of 44.1 kHz, requires extremely good crystal filters to avoid spurious responses. How I did that is described in detail at **ham.te.hik.se/~sm5bsz/pcdsp/pcif.htm**. The filter contains 27 surplus crystals for each RF channel, and it is designed for easy tuning rather than low component count. It is flat over 20 kHz and attenuates by more than 100 dB at 2 kHz or more outside the passband. With a more-conventional crystal filter, one must accept somewhat stronger spurious responses at the passband ends, and one must reduce the passband slightly. When using much more than one octave for the audio signal, the mixer becomes a very critical part of the system. Most frequency mixers suffer from second-order distortions. The second harmonic of the IF signal mixes with the second harmonic of the LO to produce an audio signal at twice the frequency of the desired signal. The strengths of such products are proportional to the square of the IF signal level. If the desired signal were reduced by 10 dB, IM_2 would go down by 20 dB only. To get low IM_2 levels from a mixer, one must operate very far below the point of saturation or use some kind of feedback or other technique to reduce IM_2 at high signal levels.

For my first PC-based system, I found Schottky-diode mixers to have far too much IM_2 to be useful. A very simple CMOS mixer (as shown in Fig 1) is good enough despite the rather high noise level of the TL074 op amp. For IFs below 15 MHz or so, this solution offers lower cost and better performance than a Schottky-diode mixer.

40-kHz Bandwidth: A Low-Cost Direct-Conversion Receiver

Using two standard mixers and a crystal oscillator, one can get a

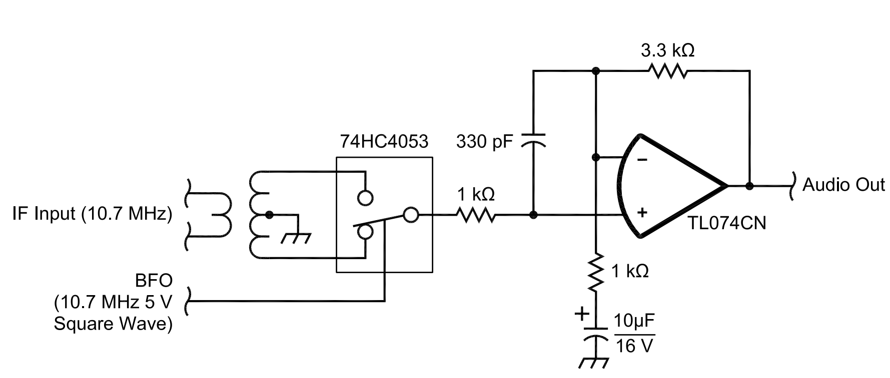


Fig 1—A simple CMOS mixer. The input transformer is not critical and it is typically loaded by a 50 Ω resistor to define the impedance seen by earlier stages.

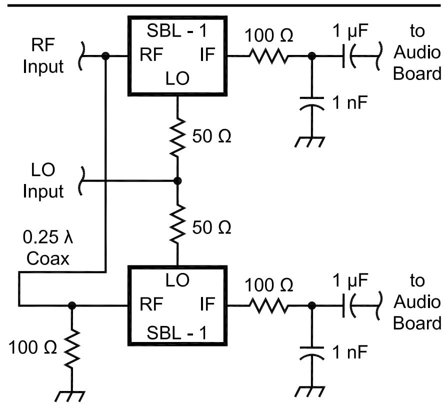


Fig 2—A very easy way to get 40 kHz of the radio spectrum into a PC.

surprisingly good radio receiver, as shown in Fig 2. No anti-aliasing filters are needed; the ones built into the sound card are sufficient if dynamic-range requirements are modest. Without any amplifier at all, the noise figure is far from acceptable; gain on the order of 40 dB needs to be added. Part of the gain is preferably added at baseband. With a very-low-noise amplifier following each mixer, the sensitivity can be made good enough for use without any RF amplifier. Since the LO power level is +10 dBm and one might expect leakage to the RF port about 30 dB below this level, one should never use this circuitry without an RF amplifier. The RF amplifier should also provide some selectivity because a Schottky-diode mixer is sensitive to signals at overtones of the LO frequency. On the HF bands, quite a lot of overtone attenuation is needed. The response at three times the LO frequency is typically only 12 dB below the response at the fundamental. For further details about this simple receiver and performance data, check **ham.te.hik.se/~sm5bsz/linuxdsp/hware/sbl1.htm**.

90-kHz Bandwidth: Another Low-Cost Direct-Conversion Receiver.

Fig 3 shows a low-noise, low-cost audio amplifier that is suitable to insert between a Schottky-diode mixer and the input of an audio board. The input transistor operates at a relatively high current, so it produces a very low noise figure at a source impedance around 50 Ω. The transistor should be rated at 1 A with an F_t of at least 100 MHz. BC489A, BF452 and many others are fine. The op amp that follows the transistor provides high gain, which is fed back to make the voltage swing small at the collector of the input transistor. This way, the second-order distortion is made very low. The second op amp provides 10 dB more gain and stabilizes the operating point of the input transistor through a low-pass filter.

With a low-noise amplifier following a Schottky-diode mixer, one easily runs into problems with the stability and purity of the local oscillator. At 144 MHz, even an IC-202 is not good enough, and that's the purest commercial transmitter I know. It is possible to lower the noise floor of an IC-202 by about 10 dB: First, replace the 470-Ω emitter resistor of the LO with a 150-Ω resistor in series with an RF choke. Then, decouple the base of the first frequency-multiplier stage for low frequencies by installing a 1-µF capacitor in series with an RF choke.

Modified like this, it is just about good enough to be used as the LO in a direct-conversion 144-MHz receiver, but only if a low-level Schottky-diode mixer were used. A high-level mixer needs about 15 dB more of LO power and would therefore be much more susceptible to LO sideband noise. Besides the desired mixing of the signal at the LO port with the signal at the RF port, a mixer will mix the signal at the LO port with itself and its own overtones. As a result, a Schottky-diode mixer will detect both AM and FM undesired modulation that may be present on the LO signal. With more LO power, the undesired AM/FM detection will produce more audio noise at the mixer port. One way of understanding it is like this: If a high-level mixer were used, about 15 dB more LO power would be needed. Both the LO carrier and the sidebands that surround it would then be 15 dB stronger so the mixing product, audio noise, would be 30 dB stronger. It is difficult, if not impossible, to make a local oscillator that is good enough for a high level Schottky mixer in a direct-conversion radio on 144 MHz.

The amplifier of Fig 3 works fine for signals up to about 75 kHz. If higher frequencies were allowed to enter the amplifier, the input transistor would go into nonlinear operation because the feedback is not fast enough. A low-pass filter is required to protect the amplifier from strong signals at high frequencies. For further details about this reasonably good 144-MHz receiver, including performance data, check **ham.te.hik.se/~sm5bsz/linuxdsp/hware/optiq.htm**.

95-kHz Bandwidth: A No-Compromise Design

An analysis of the previous design made it clear that conversion to baseband should not be made directly from VHF frequencies. It is much better to use an intermediate frequency at which a very good local oscillator can be made easily. I have started to design dedicated hardware for *Linrad*, or for any other SDR software that would interface to 96-kHz sound cards. I'm building this hardware primarily for my own use, so the units are rather big and have substantial power consumption, none of which matters at all to me personally. The first unit is a

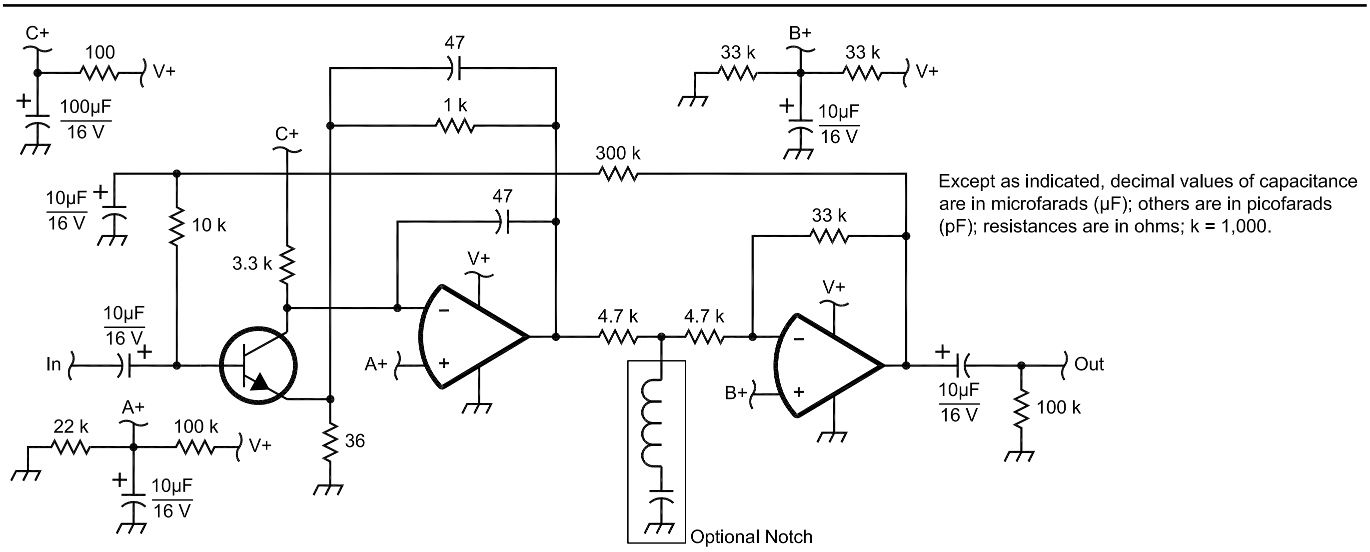


Fig 3—A low-cost audio amplifier with low noise and good linearity. An AD797 with two resistors can replace the entire circuit, but it is expensive and may be hard to find. This circuit gives very good performance with junk-box-grade components. A, B and C are common to I and Q and to both channels in a two-channel system.

2.5-MHz-to-baseband converter with a usable bandwidth of 95 kHz. This unit will be available from Svenska Antennspecialisten **www.antennspecialisten.com**. There will also be some other units to convert to 2.5 MHz from the different amateur bands via higher IF frequencies at 70 MHz and at 10.7 MHz. All the units will be described in detail with links from **ham.te.hik.se/~sm5bsz/linuxdsp/optrx.htm**.

To avoid any problems with sideband noise on the local oscillator, the last IF is placed as low as 2.5 MHz. The LO is obtained by dividing a 10-MHz oscillator by 4. Refer to Fig 4 for the circuit diagram.

A crystal oscillator can be viewed as an amplifier that has positive feedback through a filter. A 10-MHz crystal has a resistance of about 4 Ω at series resonance. When a crystal is used to connect the output of an amplifier to its input, both the input impedance and the output impedance of the amplifier must be well below 4 Ω, not to degrade (increase) the bandwidth. The voltage at the collector of the oscillator transistor is divided by about 50 times by the essentially capacitive voltage divider. This means that the source impedance feeding the crystal is about 2000 times smaller than the impedance at the collector, or about 1 Ω. The input impedance of a grounded base BF240 is very low, so this oscillator has the potential of being very good in terms of sideband noise.

The frequency divider, a 74AC74, must be fed from a low-impedance source. The clock input of this chip is not an amplifier with extremely low noise figure, and it helps to make the source impedance low to avoid phase jitter caused by the equivalent noise current at the input. It is also a good idea to feed the clock input with a high signal level. In this circuit, the pk-pk voltage at the clock input is 4.5 V, and the impedance is about 150 Ω, which is low enough at 10 MHz to do the frequency mixing from 2.5 MHz to baseband. An extremely linear mixer is needed to preserve the very good linearity and high dynamic range of 24-bit, 96-kHz sound cards. The four-phase AF feedback mixer shown in Fig 5 has an extremely low level of IM_2. Fig 6 shows a two-tone test with this unit. Two equally strong signals, both at –22 dBm, are fed into the 2.5-MHz-to-baseband converter connected to a modified Delta44 board. The two signals combine for a peak amplitude of –16 dBm, which is 1.8 dB below saturation for the A/D converters.

The two signals are at 2.4993 and at 2.5092 MHz and show up at 47.3 and 57.2 kHz, respectively, in the *Linrad* spectrum at a level of about 120 dB. The strongest spurs are at the mirror frequencies. In normal operation when *Linrad* is properly calibrated, these spurs are pushed down from about 85 to 45 dB in software.

IM_2 occurs at 46.6, 49.5, 66.4 and 38.8 kHz. The amplitude and phase relation of the IM products does not fit what is expected for a baseband signal, so each IM_2 spur is split into two frequencies. The level of the IM_2 products is about 37 dB, 83 dB below the main signals. The IM_2 products originate in the Delta44.

IM_3 occurs at 37.4 and 67.1 kHz at a level of 32 dB, 88 dB below the main signals. Since the input signal is –22 dBm, we may use the numbers to extract IP_3, which turns out to be +22 dBm. Unlike a conventional radio that typically does not misbehave too badly until the level is somewhere around 15 dB below IP_3, the 2.5-MHz unit makes the Delta44 saturate at –14 dBm; saturation is fatal to sensitivity. A DSP radio is not well characterized by IP_3 numbers. The level of the intermodulation products at the point of saturation, given in decibels below the main signals, is the best way to specify system linearity as determined by a two-tone test. The 2.5-MHz receiver has IM_2 at –81 dBc and IM_3 at –84 dBc. To fully characterize the in-band properties of this system, one must add that the noise floor is at –146 dBc/Hz at a frequency separation of 5 kHz. It is –147 dBc/Hz when no strong signal is present. Knowing that saturation is at –14 dBm, one finds that 147 dBc/Hz corresponds to –161 dBm/Hz, which is 13 dB above the noise from a room-temperature resistor (the theoretical limit). The noise figure is thus measured to be 13 dB by use of the *Linrad* S-meter. A conventional noise-figure meter gives a similar result.

At frequency separations below about 25 kHz, the 2.5-MHz-to-baseband converter competes well with any modern transceiver because of the reciprocal mixing problems associated with variable-frequency oscillators in analog receivers. But at frequency separations above 25 kHz with the offending signal within the *Linrad* passband, analog receivers may be better because they do not saturate like A/D converters. If the offending signal is more than 47 kHz from the center frequency, the 2.5-MHz-to-baseband converter does not saturate until at about +5 dBm because of the very sharp anti-alias filter following the mixer. The saturation level grows with the frequency separation to about +15 dBm at 100 kHz. This is an impressive 176 dBc/Hz above the noise floor. This kind of level for saturation is required to keep IM_3 low for signals outside the visible passband. The local oscillator is good enough to not

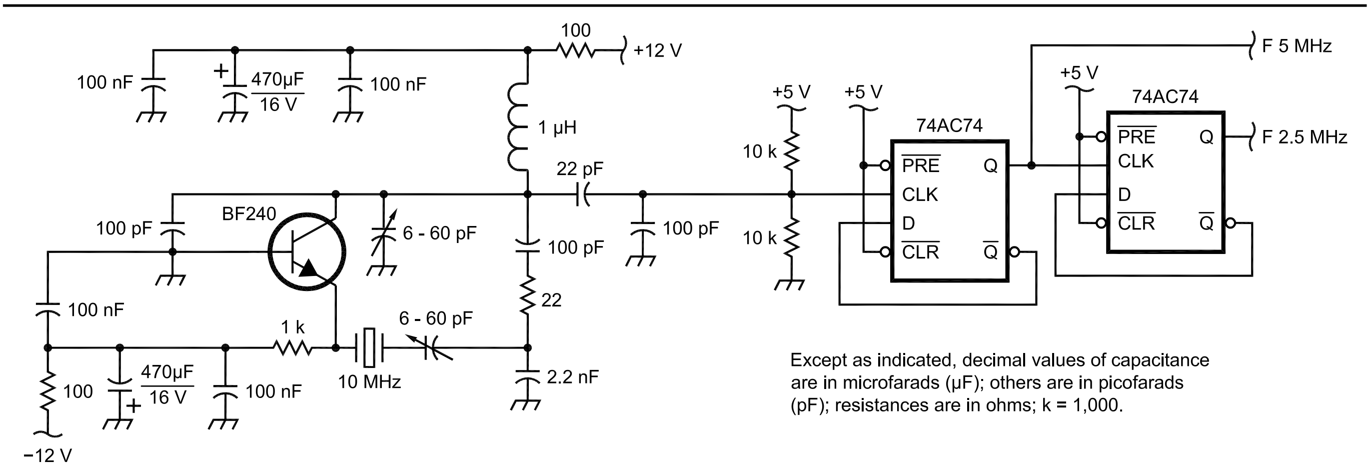


Fig 4—A low-noise crystal oscillator to drive a four-phase CMOS mixer.

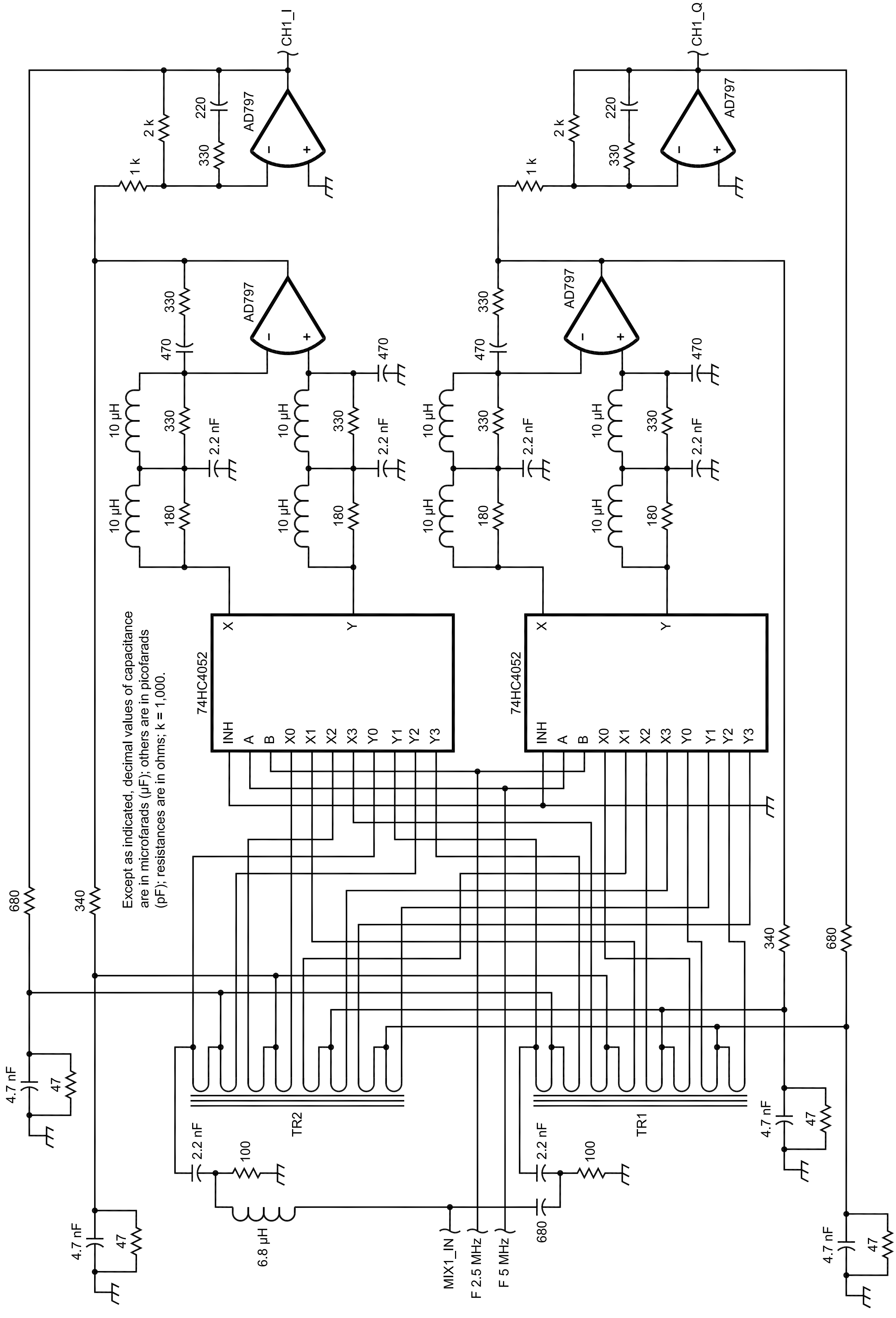

CH1_I
CH1_Q
AD797
2 k
220
330
1 k
470
10 µH
180
2.2 nF
74HC4052
INH
A
B
X0
X1
X2
X3
Y0
Y1
Y2
Y3
X
Y
Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k = 1,000.
680
340
4.7 nF
47
TR2
TR1
100
6.8 µH
MIX1_IN
F 2.5 MHz
F 5 MHz

Fig 5—(left) The four-phase AF feedback mixer.

degrade the noise floor by reciprocal mixing even for a +15dBm signal at 100 kHz separation.

The dynamic-range numbers given here are for a modified Delta44. The Delta44 has a design error, in that the analog ground of the board is grounded to the computer backplane through the screws that hold the D-sub connector. Some voltages are insufficiently decoupled. With larger capacitors and with the ground loop removed, the noise floor is improved by about 3 dB. In the case where Delta44 is intended for software-radio use only, one can improve by one more decibel. This is done by routing the signal directly to the A/D chip through a voltage divider, thereby excluding the noise from the input amplifier of the Delta44. For details about modifying the Delta44 board, look at **ham.te.hik.se/~sm5bsz/linuxdsp/hware/delta44.htm**.

Problems Associated with Large Bandwidths

The need to allow strong undesired signals within the passband creates several different problems that must be solved when a wide bandwidth system is designed. One must make the system linear up to signal levels as high as possible. At the same time, one wants the noise floor as low as possible. These two requirements are sometimes in conflict. If, for example, a Schottky-diode mixer were used, one can improve linearity by selecting a more expensive mixer that operates with a higher level of LO power. With more LO power, such a mixer becomes a more sensitive AM and FM detector that produces audio noise from the noise sidebands associated with the LO. The sideband noise of the LO may then become the limiting factor for the noise floor, in which case the performance actually is better with a low-cost, low-level mixer. Here is a list of problems associated with wide-bandwidth conversion to audio:

Ground Loops: The mains 50/60Hz and its overtones may flow in the analog ground to the computer sound card. At modest bandwidths, this problem is easily cured by shifting the spectrum upwards as mentioned above. Having a single grounding point (star configuration) for the computer and the rig will help to make the ground-loop currents small. For large bandwidths, when full audio response is required well below 1 kHz, it is often necessary to break up the ground reference. One way is to use a differential amplifier as shown in Fig 7. Any voltage difference between mixer ground and sound-card ground will be compensated for automatically. Another possibility is to break up the ground at the mixer, having separate HF and AF grounds as shown in Fig 8. If the mixer does not have separate ground returns for IF and HF signals, as does the SBL-1, one can isolate RF and LO from the common ground of the mixer by use of small coupling capacitors, typically 10 nF or less.

If the op amp is operated from a single-rail supply, R1 and R2 are made equal. Otherwise R2 is made very large, to make sure the voltage across the electrolytic capacitors does not go the wrong way. R1 should be at least 1 kΩ to keep a high impedance at 50/60Hz between sound-card ground and RF ground.

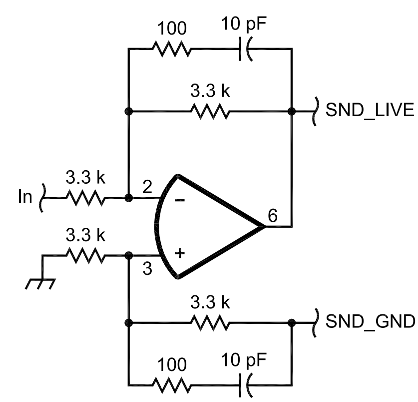
Mixer Linearity: At large signal levels, a mixer saturates. The sine wave that is the ideal response to a pure carrier at some offset from the LO becomes distorted; its overtones constitute spurious responses. A Schottky-diode mixer should not be allowed to give more than about 1% of its saturated output level to make these spurs reasonable in amplitude. FET-switch mixers may produce second harmonics because the current through the switch may have a large component of overtones to the RF frequency. This is particularly true if the switch has capacitors at both sides, an LC filter at RF and a capacitor to ground at the audio side.

RF Amplifier Linearity: An RF amplifier will produce overtones when forced to give large output signals. These overtones mix with the overtones of the LO that circulate in a saturated mixer and produce audio responses that behave exactly as mixer nonlinearities. A good filter that suppresses the harmonics of the desired RF is needed to reduce this problem.

AF Amplifier Linearity: It is obvious that the audio amplifier between the mixer and the sound card must be extremely linear. Distortion creates spurs. It is important to make sure very little RF signal reaches the audio-amplifier input. Less than a 100 mV of RF may be enough to severely degrade linearity.

AF Amplifier Noise: The audio amplifier should not add much noise. Feedback is necessary in order to get good linearity, but it is not necessarily a good idea to arrange the feedback to present a matched load to the mixer output impedance. The AF feedback mixer in Fig 5, as one example, does not load the mixer output at all. In this circuit, two mixers are used in antiphase to make use of both of the op-amp inputs. This way, the signal is increased by 6 dB while the noise is unaffected. The audio frequency impedance at the op-amp inputs is very low. This way, the thermal noise associated with resistive components at room temperature is minimized. The feedback resistor is only 47 Ω to ground. It does contribute some noise, but it

Fig 7—(below) A differential amplifier converts the voltage difference between input and ground to a voltage difference between SND_LIVE and SND_GND. Any voltage difference between ground and SND_GND will not appear at the output.



Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k = 1,000.

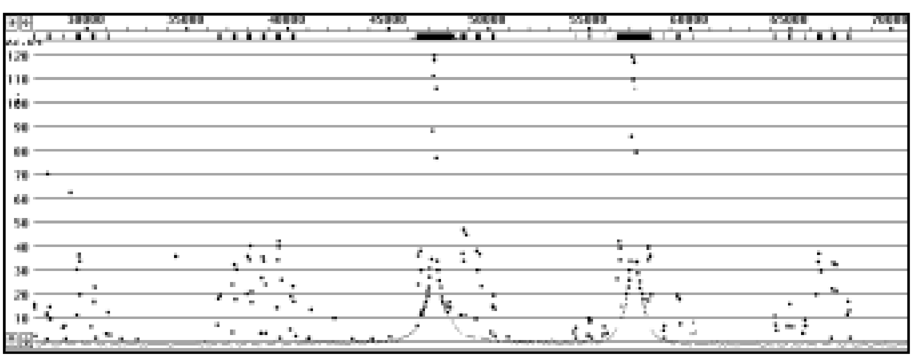


Fig 6—The main spectrum display of *Linrad* when two signals, both of level –22dBm are fed into the 2.5-MHz-to-baseband converter. The A/D is a modified Delta44 board.

cannot be made smaller because the AD797 cannot drive lower impedances than those used here, and more gain would lead to amplifier saturation.

LO Amplitude Noise Modulation: The noise sidebands of a local oscillator are usually referred to as LO phase noise. Amplitude noise is also a problem, and it can easily cause much bigger problems than the phase modulation. When the supply voltage is not extremely well filtered, there will be some AM present on the LO. A Schottky-diode mixer is sensitive to AM on the LO. AM sidebands at –80 dBc typically generate 1 µV of audio at the IF port. A 7812 regulator decoupled by the standard 1-µF capacitor does not produce a sufficiently stable dc voltage. It is necessary to use an additional RC link with something like 10 Ω and 470 µF. Since mixers to audio frequencies are usually very sensitive to AM on the LO, one must make sure that the LO is not modulated by mechanical vibrations. AM sidebands on the LO will also cause noise sidebands on strong signals because of reciprocal mixing, but this is not a major problem since the mixer suppresses AM to some extent in the reciprocal mixing process.

LO Phase Noise: Phase noise on the local oscillator causes noise sidebands on strong signals by reciprocal mixing. If the RF bandwidth is large enough to allow strong undesired signals, reciprocal mixing is a problem regardless of receiver technology. Phase noise on the LO also causes some audio noise on the IF port even when there is no signal present at the RF port. A Schottky mixer detects FM to some extent.

Amplifier AM: All RF amplifiers may introduce AM if the supply voltage is not carefully decoupled. The resulting noise sidebands on strong signals may be difficult to distinguish from noise sidebands caused by reciprocal mixing of LO noise sidebands.

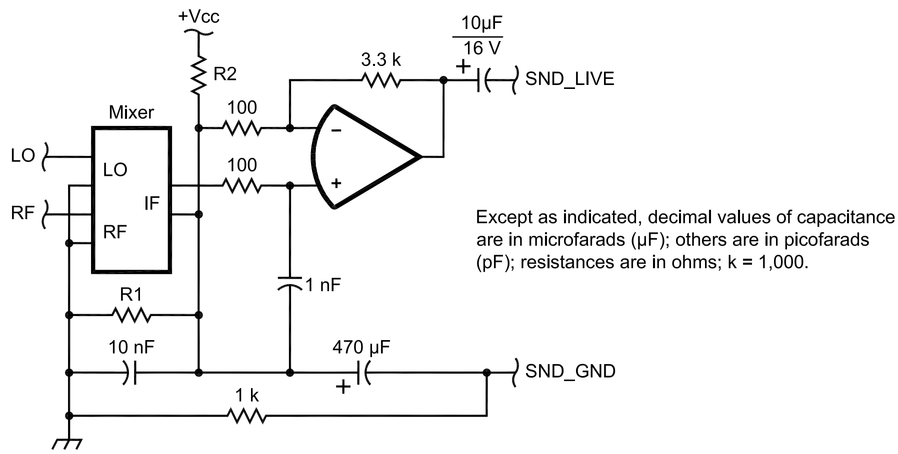


Fig 8—By making the impedance between SND_GND and ground high for audio frequencies, the current through the SND_GND wire will be small. Essentially, all the voltage induced by stray magnetic fields will develop across the 1 kΩ resistor.

Most of the problems on the list above disappear if a radio A/D is used to sample RF signals directly. I hope there will soon be hardware and drivers available to amateurs at affordable costs. Such systems will require good roofing filters to avoid very large voltages at the A/D input. Maybe a fixed-frequency filter (like 70 MHz) will be the popular solution rather than separate filters for each HF band. High-performance converters from HF to 70 MHz are not difficult. For highest performance, crystal oscillators may be used.

Summary

This article has concerned itself with details in design of the hardware required to move radio signals into the computer by use of sound cards. In the next segment, I shall discuss *Linrad* specifically: how to install it and get it running, what the basic architecture is and how it can be modified by setting various parameters.

Leif was born in 1944 and first licensed in 1961. He holds a PhD in physics and worked with research on molecular physics for 15 years at the Royal Institute of Technology in Stockholm. Since 1981, he has been running his own company. He is the inventor of the intermodulation EAS system now owned by Checkpoint. Leif is a board member of Svenska Antennspecialisten, but is essentially retired and working full time with Linrad *development.* □□