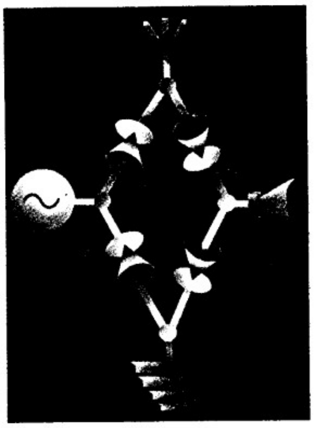
SSB:

third method, fourth explanation



In spite of the occasional burst of enthusiasm in recent years, the direct conversion SSB receiver has been overlooked. And, in a neglected corner of a neglected subject lies the Weaver receiver. It deserves better. For the professional, it is a natural candidate for use with digital signal processing techniques. For the radio amateur, it is a receiver design that will be novel to most, and is fascinating to play with. By Nic Hamilton G4TXG.

There are three ways of explaining how the Weaver receiver works in literature. Donald Weaver, in his paper *A Third Method of Generation & Detection of Single-Sideband Signals*[1] uses plenty of sin and cos maths. This is fine for mathematicians, but not so good for engineers. The *Radio Communication Handbook*[2] uses diagrams that resemble articles of evening dress. The method of explanation is reminiscent of a Victorian manual of etiquette on how to tie a bow tie. It starts off with a tie and neck. Then there a few illustrations showing fingers, neck and tie in impossible positions, and finally a perfectly formed knot surmounted by a huge grin.

Single-Sideband Systems and Circuits[3] uses the concept of negative frequency with a complex number topping.

Although these concepts are rather tricky to master, this last method is the best; the Weaver receiver becomes simple to understand, and there are many other uses for these concepts[4]. An alternative explanation follows.

The receiver system works by converting the RF input down to audio frequencies in one direct step, without using intermediate frequency stages. For reasons to be discussed later, this is done twice, using two RF mixers, one of which is supplied with an LO signal phase shifted in comparison to the other. The resulting audio signals are then passed through low-pass filters and finally combined in a second 'rotary' AF mixer stage, which is driven by a second LO.

Because the output of the RF mixers is at audio frequency, the Weaver receiver is classed as a direct conversion receiver. However, the receiver is best viewed as a type of superhet, with two parallel audio IF stages between the first (RF) and second (AF) mixers. This article uses superhet terminology. **Fig. 1** shows the block diagram of a Weaver receiver.

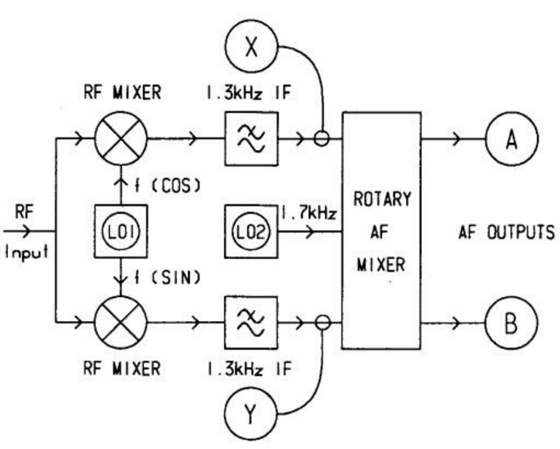
Consider one of the IF low pass filters. It has a cutoff frequency of 1.3kHz, half the width of the final audio output. This may seem rather surprising, but look at **Fig. 2**. This shows the signals to be found at various points on the block diagram. The left hand column shows that there are two possible RF input frequencies that will generate an IF output of 1kHz. One is 1kHz above the 1st LO frequency, and one 1kHz below the 1st LO frequency. So, although the IF audio filter is only 1.3kHz wide, the information passing through the filter is due to an RF bandwidth of 2.6kHz, half of which is below the 1st LO frequency, and half above. This process of getting a quart into a pint pot is achieved by the mixer folding the audio spectrum over.

Unfortunately, this folded signal is unintelligible as ordinary speech, so the job that the Weaver receiver performs is to unfold the audio spectrum into something intelligible.

To do this, a second RF mixer is used, but it is provided with an LO that is phase shifted by 90° with respect to the LO of the first mixer (sin and cos are 90° apart). This extra mixer also provides a folded audio output, however the output is folded differently. The second column of Fig. 2 shows the waveforms on the outputs of the two mixers. Note that, while the waveform at *X* is the same for both RF input frequencies, the waveform at *Y* is phase inverted. This is not to say that the waveform at *X* is useless, on the contrary, it serves as the reference that enables the phase inversion on point *Y* to be seen.

At this point in the discussion, it is simpler to consider these *X* and *Y* IF signals as being connected to the X and Y plates of an oscilloscope to form a Lissajous figure. The result for both RF input frequencies will be a circle (This is why sin and cos are called circular functions). However, the RF input that is 1kHz above the LO frequency will give an anti-clockwise rotating spot, and the RF input that is 1kHz below the LO frequency will give a clockwise rotating spot.

Fig. 1. The Weaver receiver principle. Although it looks like a direct conversion to baseband system, it is in fact a heterodyne arrangement with an intermediate frequency of 1.7kHz. Individual sidebands are resolved by quadrature product detection at 1.7kHz. Although the IF audio filter is only 1.3kHz wide, the information passing through the filter is due to an RF bandwidth of 2.6kHz, half of which is below the 1st LO frequency, and half above. This process of getting a quart into a pint pot is achieved by the mixer folding the audio spectrum over. The second mixer unfolds it to its full width.



Explaining AF mixing

To recover the original audio signal, the *X* and *Y* IF signals are connected to the rotary AF mixer. This imparts an extra clockwise or anti-clockwise twist to the spot's motion. The speed of this extra twist is 1.7kHz, which is the second LO frequency. The resulting outputs are shown in the right-hand two columns of Fig. 2. The outputs are still circles, but with differing rates of rotation, depending on whether the original RF signal was greater or less than the first LO frequency.

Assume that the receiver is to be used to demodulate USB. The upper row of Fig. 2 shows that an RF input of 1.001MHz gives an IF of 1kHz, and the output from the rotary mixer is 2.7kHz. The lower row shows that an RF input frequency 1MHz–1kHz also gives an IF of 1kHz, but this time gives a rotary mixer output of 700Hz.

If the Weaver receiver is used to demodulate LSB, the direction of the extra twist from the rotary AF mixer is reversed. The RF input of 1.001MHz now gives a 700Hz AF output, and the RF input frequency 1MHz–1kHz an output of 2.7kHz.

The circuit of the rotary mixer that gives this extra twist is discussed later. Note that, for the purposes of this illustration, the receiver has two audio outputs with a 90° relative phase shift. This allows the rotary AF mixer outputs to be discussed as circular Lissajous figures. However, for an SSB receiver, only one of the two outputs is needed.

Mixers

For the direct conversion receiver, the RF mixers' performance is vital: apart from the LO, there is very little other RF circuitry. If a direct conversion receiver does not work satisfactorily, the RF mixer is usually to blame.

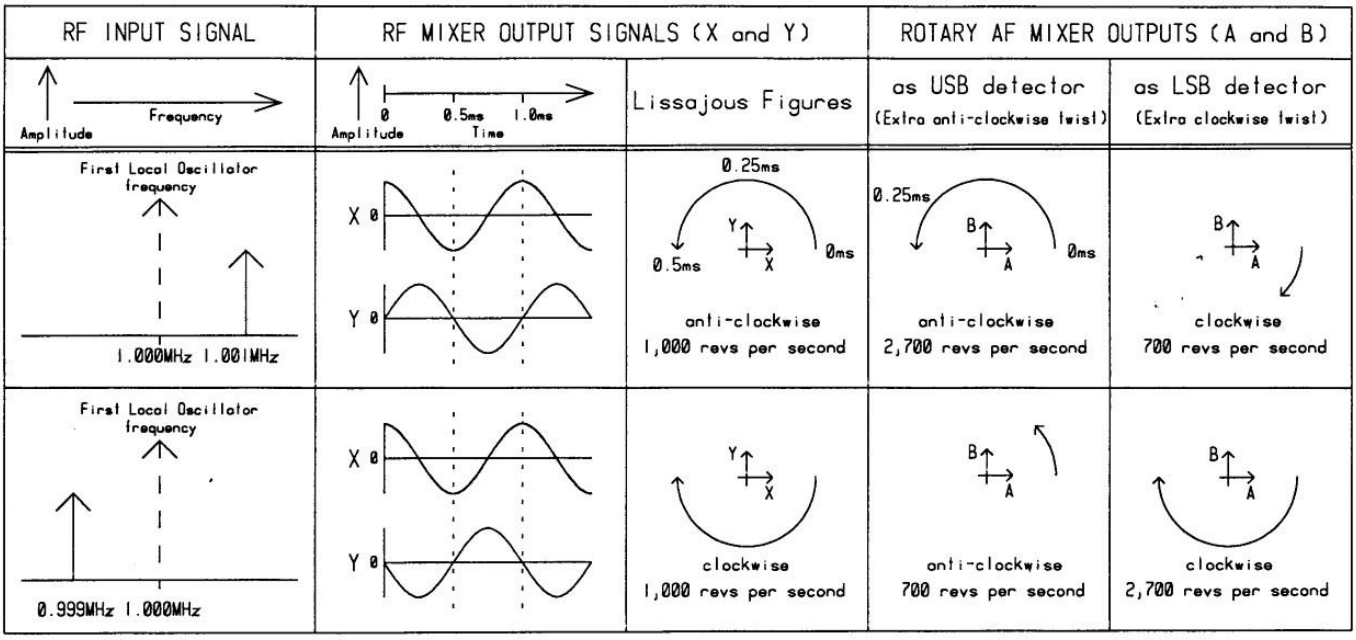
The wanted output frequency from an RF mixer can be either LO+RF or LO–RF. In the direct conversion receiver, the mixer must translate the RF input down to audio frequencies. So it will be assumed that LO–RF, the difference frequency, is the wanted output.

The ideal RF mixer would have two inputs, RF and LO, and the output would consist of just one frequency, the difference between the input frequencies. It would have no harmonic responses. How might this be achieved?

To generate a lower frequency output, the cycles of the input waveform must be lengthened thus the output must be phase-retarded each cycle with respect to the input. To do this, a voltage variable phase shifter is required. It must be able to shift the phase of the input signal by a full 360°, and be continuously variable. A block diagram of a circuit which does this is shown in **Fig. 3**.

The circuit works by splitting the incoming RF signal into a 0° and a 90° component. The two signals are then passed to two balanced mixers. These act both as phase inverters and as voltage controlled attenuators. The phase inverter action means that the phase of the upper mixer's output can be either 0° or 180°, and the phase of the lower mixer's output can be either 90° or 270°. A judicious mixture of these four phases results in a continuously variable phase from the output of the sum-

Fig. 2 shows the signals to be found at various points on the block diagram. The left hand column indicates that there are two possible RF input frequencies that will generate an IF output of 1kHz. One is 1kHz above the 1st LO frequency, and one 1kHz below the 1st LO frequency.



Design considerations

The Weaver receiver has been ignored as a design for analogue receivers. For receivers using digital signal processing it has been considered and rejected[11]. The reasons for this rejection are quoted as these.

"The problem of DC offsets would necessitate AC coupling. However, this would place a notch in the effective receiver passband which could be troublesome for certain modulation modes."

This design proves that the central notch can be made as narrow as 10Hz. The resultant degradation of SSB is negligible. Even when tuning around a strong carrier wave, the notch is quite hard to find by ear. However, the notch does result in a short burst of 1.7kHz from the receiver output each time there is a large change in LO frequency.

For the HF band, the synthesiser would have to cover over two decades of frequency range and provide quadrature outputs. The VCO must have a one octave frequency range. To receive the full HF frequency range, extra divide by two circuits may be used. It is simple to make the frequency dividers provide the necessary 90° outputs.

Gain and phase matching of the (cos and sin) channels have to be accurately maintained over a very wide bandwidth to avoid sideband image problems. This receiver achieved an AF distortion suppression of 30dB without difficulty, and this is almost inaudible. DSP techniques would reduce the distortion.

1/f noise in the mixers and audio amplifiers should be minimised. The receiver's noise figure is about 20dB, which is about 10dB higher than a standard HF receiver. This is mainly due to the choice of amplifier circuit at the input of the LP IF filter. The noise floor could be lowered, but this would reveal 1/f noise and greater hum sensitivity. The advantage would be that, given a clean LO, lowering the receiver's noise floor would further increase its spurious free dynamic range.

RF sub-octave filtering would be essential to prevent unwanted signals at harmonics of the input signal from mixing with harmonics of the local oscillator. This design makes the point that, by running the mixers with an accurate square wave drive, and by using 3rd harmonic response cancellation, the complexity of the preselector is considerably reduced.

Remember that it is standard practice to quote the suppressed carrier frequency of an SSB signal. So, for LSB, the receiver should display LO+1.7kHz, and for USB, the receiver should display LO–1.7kHz.

ming junction. This judicious mixture is arrived at by yet another 0° and 90° splitting of the LO signal.

Squarewave LO drive

Using a double balanced diode ring mixer as a voltage variable RF attenuator is likely to cause intermodulation products. Imagine that the LO input is a sine wave. As the instantaneous LO voltage nears 0V, the RF voltage has a greater effect on the diode current, so the mixer attenuation varies depending on the RF input waveform. This is another way of saying that it is generating intermodulation products. To avoid this, the mixers must be provided with a square-wave LO drive[5] in order to achieve the required strong signal handling performance. This works because the diode current is at its saturation value most of the time, and passes through the 0V danger zone much more quickly.

In this receiver, the square wave input to the mixer is generated by a frequency divider with a 1:1 duty cycle output waveform. The LO mixer drive of the lower mixer in Fig. 3 can be expressed as a Fourier series:

$$S(t)=4/\pi[\sin\omega t+{}^1/_3\sin 3\omega t+{}^1/_5\sin 5\omega t+{}^1/_7\sin 7\omega t+...]$$

This shows that the second harmonic component of the mixer drive waveform is theoretically zero. In fact it will be present, but will be very small. The result is that the mixer's 2nd harmonic response will be similarly small.

This is not the case for the 3rd harmonic response, which will have a conversion loss only $20\log(^1/_3)$ or 9.5dB larger than the fundamental frequency conversion loss. The upper mixer drive in Fig. 3 may also be expressed as a Fourier series:

$$C(t)=4/\pi[\cos\omega t-{}^1/_3\cos 3\omega t+{}^1/_5\cos 5\omega t-{}^1/_7\cos 7\omega t+...]$$

Compare the two series, and notice that the terms for the 3rd and 7th harmonics change signs. This results in the two mixers giving the same AF amplitude output in response to a 3rd or 7th harmonic RF input frequency, but the outputs are phase inverted with respect to each other. When they are added together in the summing junction, these harmonic responses cancel. The result of this is a mixer in which all the harmonic responses cancel, with the exception of the 5th, 9th 13th etc.

It is difficult in practice to build the RF input phase shifter so that the accuracy is better than 4°. This gives a theoretical 3rd harmonic signal cancellation of 29dB (see box). Add to this the 3rd harmonic conversion loss (see above) of 9.5dB. Thus the minimum 3rd and 7th harmonic rejection should be about 40dB. Inaccuracies of the 0/180° phase shift in the mixer limit the even order harmonic rejection to about 50dB.

The effects of error

For a pure sine wave input to the receiver, the outputs of the IF amplifier/filters should be equal and have 90° phase shift. However, the phase and amplitude are always slightly in error, so the Lissajous figure, which should be a perfect circle, is always slightly elliptical. This elipticity can be viewed as the result of a small circular component rotating in the opposite direction to the main component. The demodulator interprets this as a small signal of the opposite sideband to the main signal, so, when demodulating USB, the distortion takes the form of an LSB image. This is the worst case in-band AF distortion referred to in **Table 1**.

Note that the harmonic responses of the mixers give a residual output from the two IF filters. Although the resultant frequency seen in the filters will be the same, the relative phase and amplitude between the two signals will be random for any given frequency. Because the two IFs have signals that are not of identical amplitude and 90° phase shift, the rotary mixer will not be certain whether to

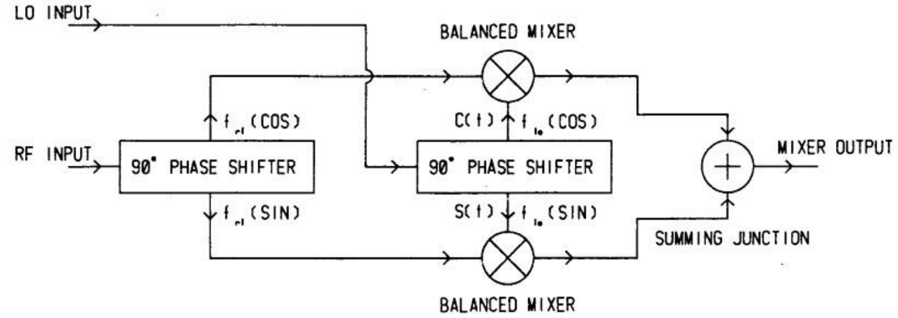


Fig. 3. The ideal RF mixer for a Weaver principle front end. The circuit works by splitting the incoming RF signal into a 0° and a 90° component. The two signals are then passed to two balanced mixers. These act both as phase inverters and as voltage controlled attenuators. The phase inverter action means that the phase of the upper mixer's output can be either 0° or 180°, and the phase of the lower mixer's output can be either 90° or 270°. A judicious mixture of these four phases results in a continuously variable phase from the output of the summing junction.

give a USB or an LSB output, and so gives a mixture of both.

Spurious demodulation is possibly the worst fault of direct conversion receivers: strong signals (usually 7MHz AM broadcast) are directly demodulated to audio frequencies, irrespective of the LO frequency. Spurious demodulation occurs because of imbalance in the mixer. The *SBL1* is inherently well balanced: it has all its diodes fabricated on one substrate. The *SRA1H* uses a ring of eight matched diodes in separate packages, and will not be so well balanced. This may explain why the spurious demodulation performance of the two types is similar, even though the *SRA1H* has a higher 3rd order input intercept point.

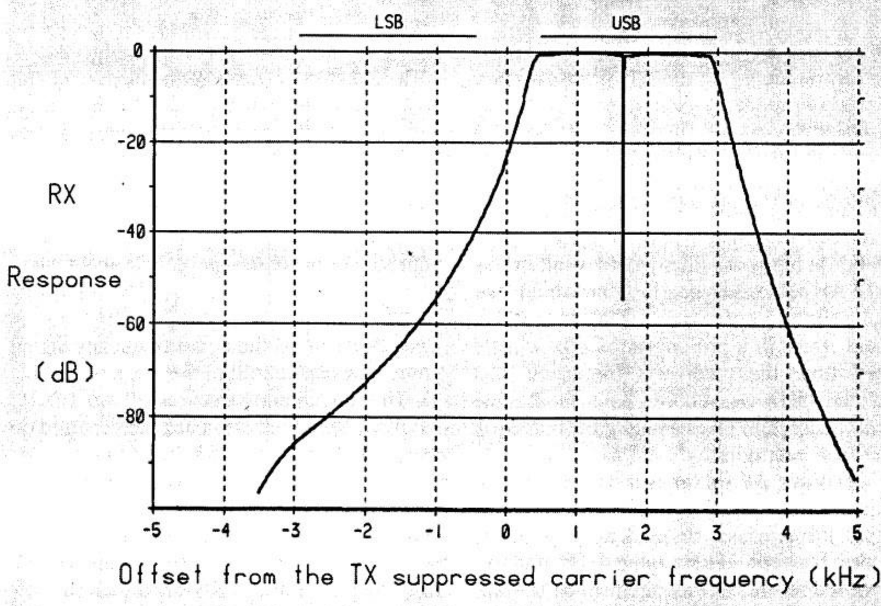
However, the third order intermodulation products start appearing above the receiver noise floor at the same signal level as the onset of spurious demodulation. In fact, for a superhet, an AF detector on the mixer output would be a good indication of mixer RF overload.

In the Weaver receiver, the spuriously demodulated signals are frequency shifted by the second mixer, making them unintelligible. This is an advantage: off tune SSB interference is subjectively far less annoying than, for example, being able to hear the BBC World Service in the background on every frequency.

All direct conversion receivers are prone to pick up power line hum. This is due to a combination of direct pickup and LO radiation[6]. To reduce hum, homodyne and phasing receivers have a 400Hz high pass filter on the mixer outputs. This removes 50Hz and its first seven harmonics without any loss of SSB signal information. The Weaver receiver cannot do this because the IF filters must pass frequencies down to 5Hz. In this receiver, any hum is converted by the AF mixer into two tones close to 1.7kHz, which are impossible to filter out. As a result, the Weaver receiver's mixers must have a hum output much smaller than other direct conversion receiver types.

Table 1. Performance figures for the prototype receiver. The measurements were made with the RF attenuator and preselector disconnected.

Frequency coverage	1.5 to 30MHz
Detector type	SSB (USB or LSB)
RX 3rd order input intercept point	+16dBm
Selectivity	see graph
Centre frequency notch width	10Hz at 3dB points
Sensitivity	–108dBm for 12dB SINAD
Harmonic responses (receive frequency 2MHz w.r.t. –108dBm):	
	34 dB at 6MHz (3rd)
	12 dB at 10MHz (5th)
	18 dB at 18MHz (9th)
	21 dB at 26MHz (13th)
	50±3dB the rest
Sub-harmonic response (2MHz in) w.r.t. –108dBm:	
	83dB (RX at 4MHz)
AF distortion	30dB below wanted tone
Spurious demodulation (100% AM)	–30dBm for 12dB SINAD



Receiver selectivity graph.

Implementation

Looking at block diagram **Fig. 4**, the main receiver components are these:

The **attenuator** reduces the signal levels arriving at the mixer when necessary. A 7.2MHz half-wave dipole at night can produce signal levels as high as 0dBm (223mV across 50Ω or 1mW) – causing severe direct demodulation. It is high time automatic RF attenuators were standard in HF receiver design.

The **preselector** performs three functions. The first is simply to minimise the out-of-band RF energy arriving at the mixer. This reduces the susceptibility of the receiver to interference from intermodulation products or direct demodulation. The second function is to attenuate sub-harmonic frequencies. These will suffer harmonic distortion in the isolator and mixer, and give spurious products at the receiver frequency. The third function is to attenuate frequencies at harmonics of the receiver frequency. These may produce harmonic responses from the mixer.

The **isolator** passes the signals from the preselector to the mixer at unity gain, but attenuates LO leakage from the mixer attempting to travel in the opposite direction by up to 80dB. Reference 6 explains the use of an isolator to reduce local oscillator radiation and RF generated hum and microphony.

The **90° phase shifter** is a broadband device, but, because its only function is to cancel the 3rd harmonic response of the mixers, the frequency range of the cancellation need only be from the 3rd harmonic frequency of the lowest input frequency to the receiver's top frequency. This phase shifter gives 90°±4° over a 4 to 30MHz frequency range.

The **complex mixer** (**Fig. 5**) is an implementation of Fig. 3, one to provide the sine output and one the cosine. The mixers are driven by a square wave derived from a divide by four flip-flop stage. The two square waves have a 90° relative phase shift. This phase shift is used to give the correct phase relationship between the two IF filter signals in order that the correct USB or LSB detection can be applied. It must therefore work over the full frequency range of the receiver.

The **1.3kHz LP IF filter and amplifier stages** have a 6th order 0.1dB Chebyshev frequency response. A more complex filter would give better selectivity, but it would make the phase and amplitude matching of the two stages harder. It is instructive (and pretty) to connect the outputs of these filters to the X and Y plates of an oscilloscope. The true phasor nature of the RF input signals can then be displayed. It is, for instance, easy to spot the broadcast stations that have AM to PM conversion, or dynamic carrier control.

The **5Hz high pass filter** removes the DC component of the IF signal that is to be connected to the rotary mixer. Most of this DC component arises from mixer imbalance, and the residue from amplifier offsets. The DC level controls the amount of 1.7kHz in the output, so the spurious content must be removed.

The mixer DC offset depends on the LO frequency, so, for each large frequency change, a step function is applied to the HP filter. The

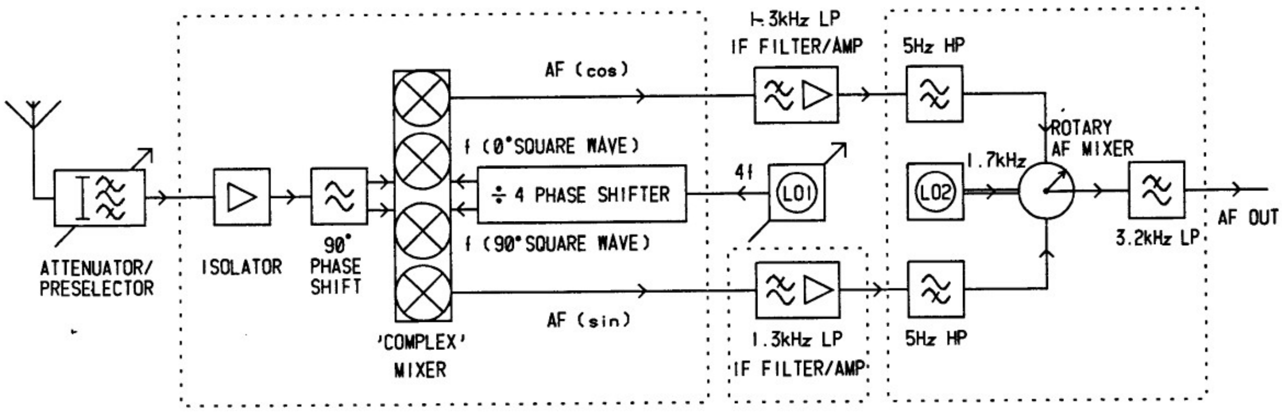
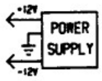


Fig. 4. Receiver block diagram.



transient response of the HP filter causes a 1.7kHz ping at the receiver output. So the cut-off frequency of the filter is a compromise between the desire to use the narrowest possible notch width in the centre of the AF passband, and the desire for the ping caused by each frequency change to be shortest.

A 0.5Hz high pass filter would result in less of the AF passband being lost, but would take a long time to arrive at its final value. This would result in a prolonged 1.7kHz whistle every time the receiver's frequency was changed. 5Hz was chosen as a reasonable compromise, and gives a negligible reduction in speech intelligibility[7].

The **rotary AF mixer** is a rotating switch which selects one of eight phases for the AF output. If four phases are used, an audio image appears between 3.8kHz and 6.4kHz, and this imposes a severe filter requirement on the subsequent LP filter. With eight phases, this image is moved up to 10.6kHz to 13.2kHz, well out of harm's way.

The **3.2kHz LP filter** removes the 3rd and higher harmonics of the 2nd LO frequency, and some high frequency mixing products which are generated by the Rotary AF mixer because it operates in discrete phase steps.

Circuits are not given for the attenuator, preselector, LO, or AGC stages since there is nothing unusual or design specific about them.

Receiver frequency response

The effects of all the audio frequency signal processing can be seen in the graph (see Table 1). The fundamental shape is of two 1.3kHz LP filters glued back-to-back, and centred on the 2nd LO frequency of 1.7kHz. In the centre of the passband is a 10Hz wide notch formed by the 5Hz HP filters. The 3.2kHz LP filter's cut-off can be seen at ±3.2kHz; it causes the receiver response to roll-off faster on the high frequency side of the passband than the low frequency side.

At the AF output shown on Fig. 4, all frequencies less than 400Hz will contain interference and no signal. The subsequent loudspeaker amplifier will therefore be AC coupled. The frequency response of the receiver from the RF input to the loudspeaker output will thus resemble the response graph, but will have an extra notch 800Hz wide centred on 0Hz.

RF circuit description

The isolator stages use two dual grounded gate JFETs. The circuit[6] has been adapted to a push-pull type in order to reduce the second harmonic distortion. Both live and ground of the RF input are connected to the receiver using 100nF capacitors. The LO input is similarly decoupled. This stops 50Hz hum current from flowing from the ground of the antenna connector or external LO, past the low noise AF input, and into the mains supply earth. A small potential due to this current would be developed at the low noise input of the 1.3kHz LP IF amplifier.

The 90° phase shifter, if made with perfect

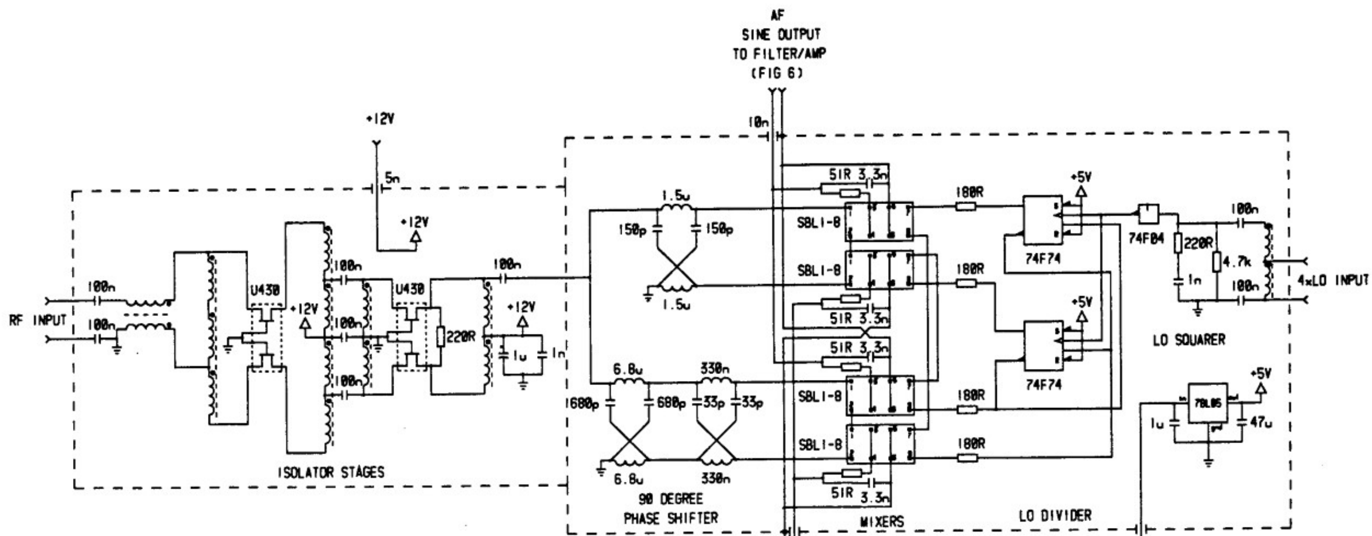


Fig. 5. The complex 1st mixer and front end. The full circuit would include an attenuator and preselector. It has the property of cancelling 3rd and 7th harmonics.

components, would give the response shown in the graph, right – a computer simulation. Unfortunately, the most accurate inductors to hand had a tolerance of 10%. As previously explained, a 3rd harmonic rejection of about 40dB could be expected depending on the accuracy of the 90° passive phase shifter. This has not been achieved, see Table 1. Note that this harmonic rejection was measured at 6MHz, which is where the inaccuracy of the 90° phase shift is greatest, so the phase error due to component tolerance was positive, and added to the predicted error.

The phase shifter design is simplified by using a branch impedance of 100Ω, which means that the inductors must be exactly 10,000 times the capacitor values, so the E6 series of component values can be used.

The LO divider is a *74F74*. This provides a square wave drive current of 14mA for the mixers. The *74F* logic series is not guaranteed to operate at a clock rate higher than 100MHz; this would limit the maximum receiver frequency to 25MHz. In practice, all the devices so far tried have operated at greater than 120MHz. For higher guaranteed speeds, use ECL.

The mixer inputs are connected in series, giving an input impedance of 100Ω to match the phase shifter impedance. The IF outputs of the mixers are matched at RF by a 51Ω resistor. At AF, the mixers are mismatched. This reduces the audio band 3rd order input intercept point of the receiver, but improves its noise figure.

LP IF filter/amp circuit description

The input to the 1.3kHz low pass IF filter/amplifier stages could have used a diplexer[6]. In these, the filtering provided by the diplexer is calculated to be part of the overall filter response. For more complex designs of diplexer using Chebyshev/Inverse Chebyshev or Cauer/Cauer pairs, see reference 8. Be warned that Cauer diplexers have a nasty habit of requiring negative values of output capacitance on the high pass side. The actual circuit is shown in **Fig. 6**. These have to be realised using transformers and some mathematical sleight of hand, and transformers are microphonic.

However, we live in an age where the aim is to avoid using inductors at all costs. So this design uses a second order filter stage[9]. The input resistance is formed by the 50Ω source impedance of the complex mixer circuit. This low input resistance results in a high input capacitance to ground, which provides an excellent opportunity to prevent LO signal escaping from the mixer compartment.

The second audio stage is similar to the first, and has a gain of 20dB. The third stage has the poles with the highest Q, so this was designed using a low sensitivity section. The design procedure for this section produces a fixed R/C ratio, where R and C may be chosen at will.

The 20dB amplifier output stage raises the signal level to a value that is just sufficient to

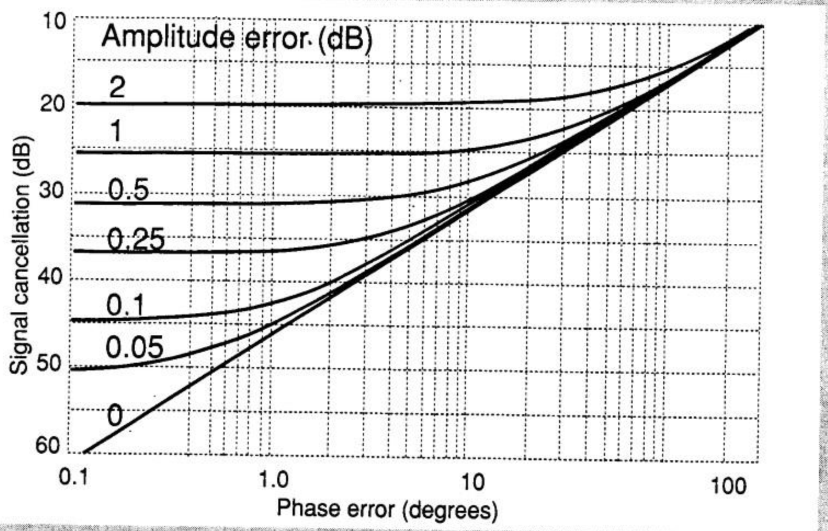
Signal Cancellation

There are two places in the receiver architecture where the signal path splits into two, which then rejoin. At each of these joining places, the aim is to cancel out unwanted signals. This happens in the RF mixer where 3rd and 7th harmonic responses of the mixers are cancelled out, and again in the rotary mixer where the two low frequency IF signals are combined in order to cancel out the unwanted sideband.

In the first instance cancellation depends on the accuracy of the 90° phase shifting network. This is composed of all pass filters, so the amplitude error will be very small, and the phase error dominant. However, the amplitude error can be dominant; for example, there is a 90° hybrid junction design that has constant phase difference, but has amplitude ripple.

This is the exception. The phase error is dominant in most circuits. Imagine the errors to be due to an RC LP filter. At one tenth of the cutoff frequency, the amplitude error is 0.044dB, and the phase error is 5.7°. Look at the graph; it shows the amount of signal cancellation that may be expected for various phase and amplitude errors. An amplitude error of 0.044dB with no phase error gives a signal cancellation greater than 50dB, whereas a 5.7° phase error with no amplitude error gives a signal cancellation of 27dB. The lower value of signal cancellation will prevail.

There is also a less obvious splitting and combining point in each of the balanced mixers. These work by selecting a phase of either 0° or 180°, depending on the direction of the instantaneous LO current flow through the mixer. The phase shift between these two states must be exactly 180° and the amplitude of the two states must be exactly equal for the even order harmonic responses to cancel completely. In this case, the two paths through the mixer may be thought of as splitting and rejoining in time.



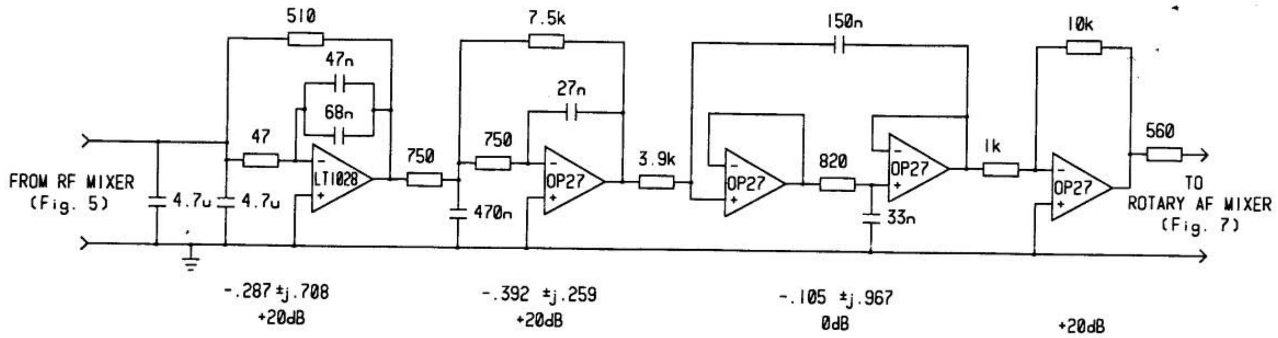
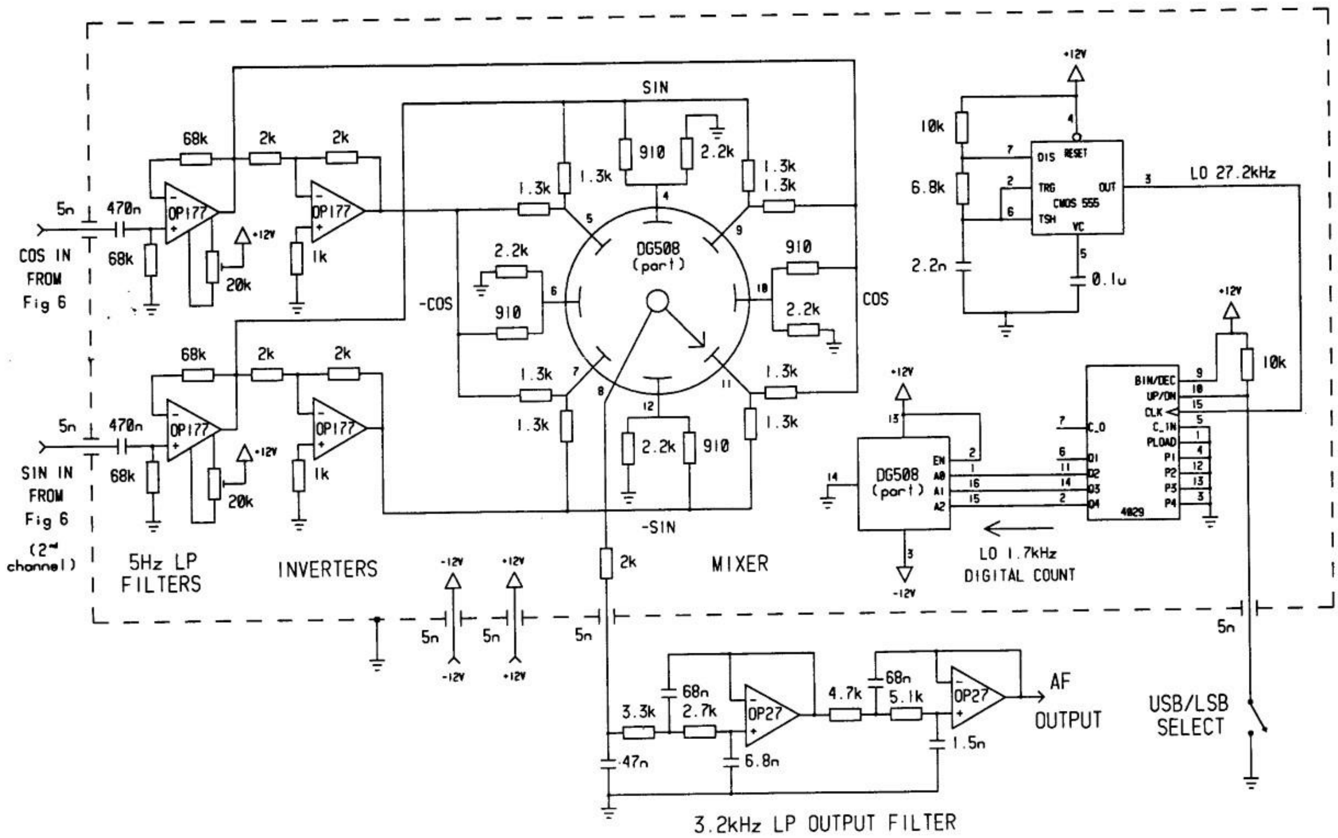


Fig. 6. 1.3kHz low pass IF filter/amplifier stages



make the leakage of the 2nd harmonic of the 1.7kHz LO inaudible in the front-end noise.

Rotary AF mixer description

The 5Hz high pass filters must be first order to keep the transient response free of overshoot[10]. For this a simple resistor/capacitor time-constant is used. An alternative but more complex candidate for this circuit is the critically damped filter, which has a transient response that arrives at the final value rather more promptly than a resistor/capacitor filter.

The *OP117* op-amps have very low DC offset, and would normally require no adjustment. However, a small amount of 1.7kHz leaks from the subsequent mixer circuit into the AF output. The offset potentiometers on the op-amps allow a small amount of signal to pass in anti-phase to the mixer leakage thus cancelling it out.

The **rotary mixer** contains two inverters: one on the sine channel and one on the cosine channel, their outputs are –sine and –cosine creating four phases to choose from, **Fig. 7**.

Four more phases are interpolated by the resistor networks around the analogue inputs of the *DG508* analogue switch. The resistor values are chosen such that, at the *DG508*, the source impedance of all eight phases are equal. Also, the sine, cosine, –sine and –cosine phases are attenuated by a voltage ratio of 0.5, so that all the phases have an equal RMS voltage. There are now eight phases for the *DG508* rotary mixer to choose from.

The second LO is generated by a *555* oscillating at 27.2kHz. This signal is connected to the *4029* up/down counter. As the *DG508* has only three digital inputs, one of the four counter outputs is unused. This gives a choice of using a clock rate of 27.2kHz, and not using the Q_1 output of the *4029*, or of using a clock rate of 13.6kHz, and not using the Q_4 output. While this last option provides identical control of the *DG508*, some of the 850Hz generated by the *4029* inevitably leaks into the audio output, so the 27.2kHz option is to be preferred. Note that the majority of low frequency circuitry is enclosed by a screened box, and that all the supply and signal connections are made using feed-through filters. This must be done to prevent RF harmonics of the 2nd LO from leaking into the RF stages of the receiver.

The 3.2kHz low pass output filter is a 5th order 0.1dB Chebyshev filter[9]. The filter input resistor has been reduced in value to allow for the mixer source impedance.

In the output of the conventional mixer (balanced modulator), the sum and difference frequencies are present at the same time. This is because the conventional mixer approximates to the rotary mixer by using two phases: 0 and 180 degrees. The mixer cannot distinguish between clockwise and anti-clockwise rotation. The output contains the two frequencies as if the mixer were rotating both clockwise and anti-clockwise simultaneously.

Fig. 7. Four phase product detector described by the author as a rotary mixer. It contains two inverters: one on the sine channel and one on the cosine channel, their outputs are –sine and –cosine creating four phases to choose from. Four more phases are interpolated by the resistor networks around the analogue inputs of the DG508 analogue switch.

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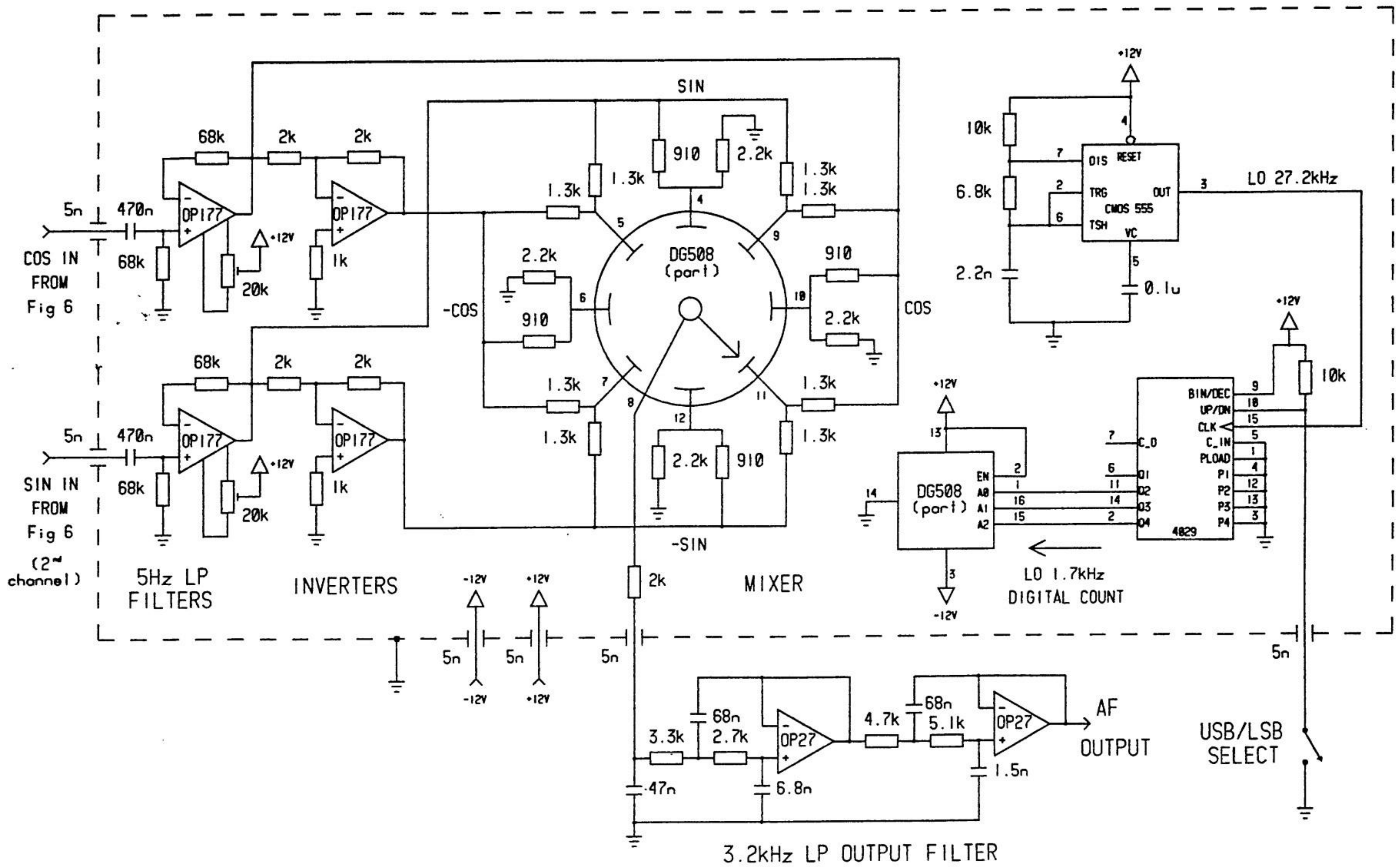
COS IN
FROM
Fig 6
SIN IN
FROM
Fig 6
(2nd channel)
5n
470n
68k
OP177
20k
+12V
2k
1k
5Hz LP
FILTERS
INVERTERS
SIN
-COS
COS
-SIN
DG508
(part)
910
2.2k
1.3k
MIXER
2k
-12V
+12V
5n
10k
6.8k
2.2n
0.1u
DIS
RESET
TRG
TSH
VC
OUT
CMOS 555
LO 27.2kHz
EN
A0
A1
A2
C_O
Q1
Q2
Q3
Q4
4029
BIN/DEC
UP/DN
CLK
C_IN
PLOAD
P1
P2
P3
P4
LO 1.7kHz
DIGITAL COUNT
3.3k
2.7k
68n
6.8n
.47n
OP27
4.7k
5.1k
1.5n
AF
OUTPUT
USB/LSB
SELECT
3.2kHz LP OUTPUT FILTER