

Summer 2006

Construction project done as additional “learning curve” towards
RSGB Advanced Licence Exam by Martin Juhe 2E0MJD

P.W. Portland VFO & Buffer

The circuit and information came from **Practical Wireless March and May 2006** editions and should be attributed to **Tony Nailer G4CFV**.

My Advanced exam tutor **Dave McQue G4NJJU** suggested, when we were covering the topic of VFO's in the Advanced syllabus, that I build one to further understand the basic principles, which I thought was a good idea.

I started the project in March and took the exam in May.

Please note that unlike the Intermediate exam, you do not have to build a project as part of the Advanced course, but I think it was a good idea.

I further thought at the time, that this could be a good project for the Intermediate exam. I wouldn't say that the project is particularly complicated from the construction point of view!

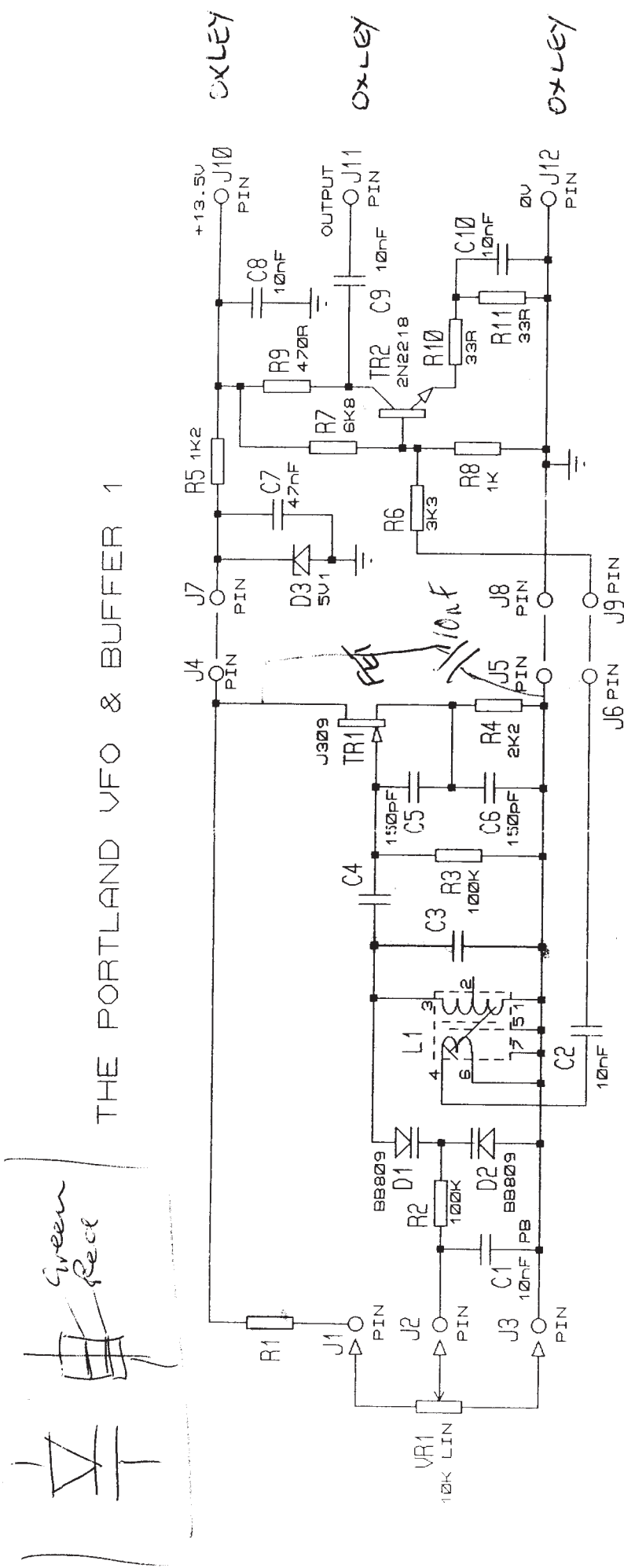
I built the the VFO and buffer for 7.88 - 8.42MHz as part of a mixerVFO system for a 4 metre project that I still have to do when I get time!

This VFO project, as the article states can easily be built for 6.99 - 7.21MHz for a 40 metre direct conversion local oscillator. All that changes are the value of R1 and the deletion of C3.

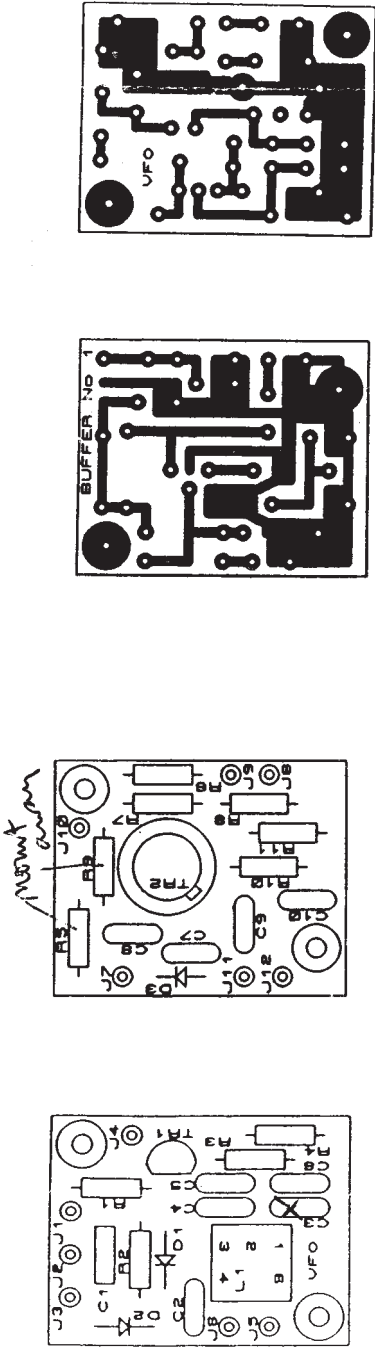
All in all it was a very interesting and worthwhile project and I am please that I did it.

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M0XJP / 2E0MJD
13/1/07

Article reprint, courtesy of Practical Wireless

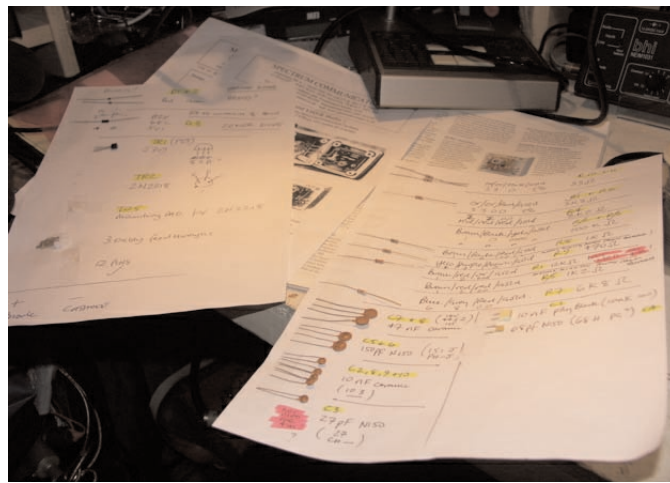


FREQUENCY MHZ	R1	L1	C3	C4	USE
6.985 - 7.215	12K	K2027	27pF	68pF	40M
7.880 - 8.420	3K3	K2027	-	68pF	4M

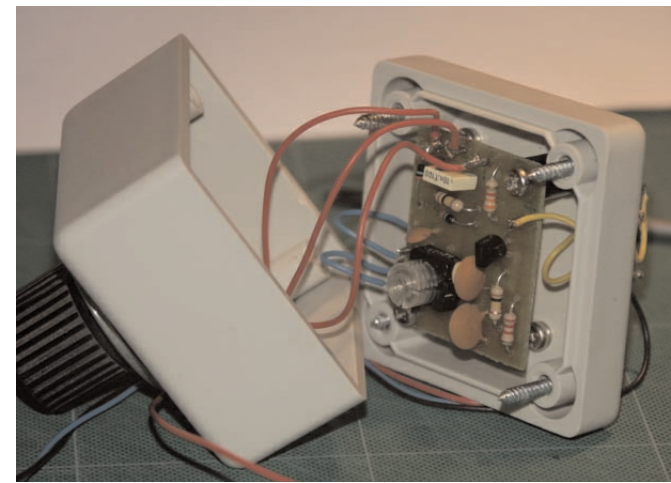




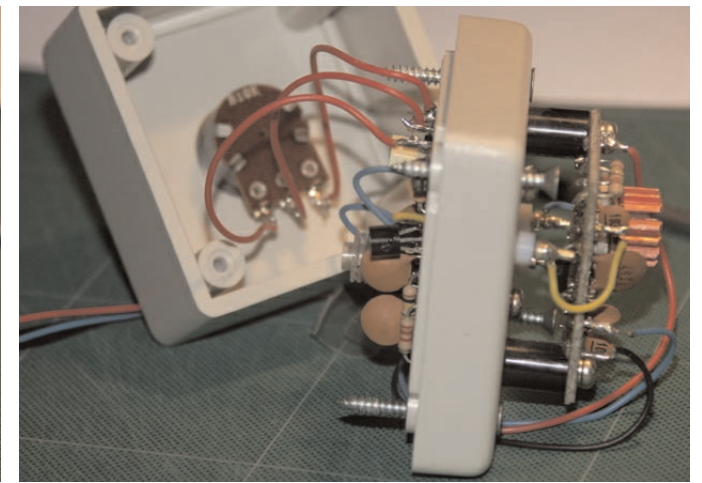
At the early stages



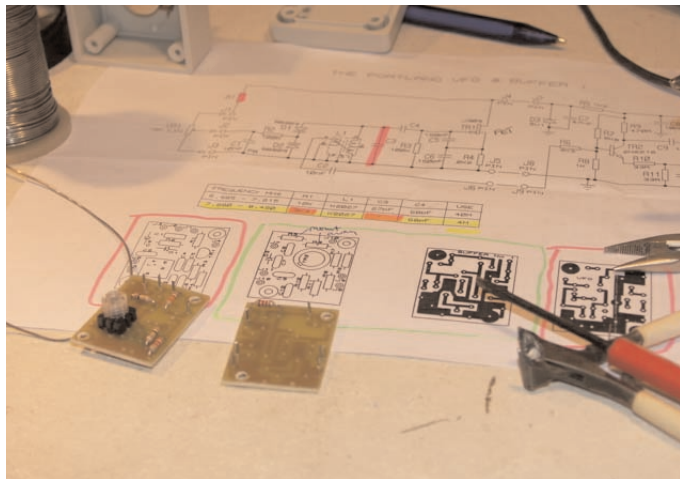
Check list for parts and components



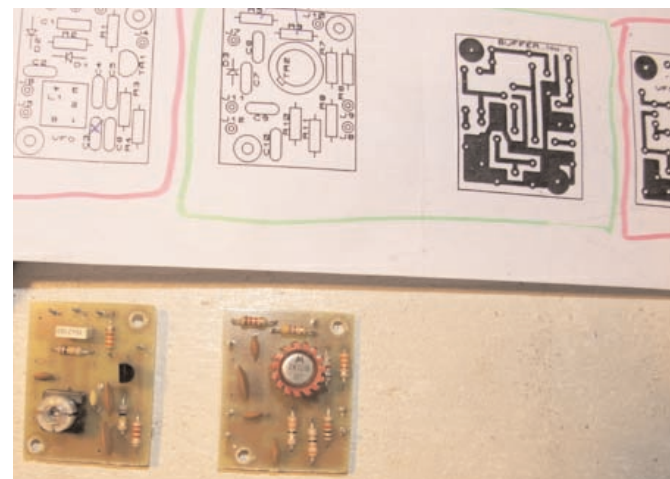
What's inside



What's inside 2



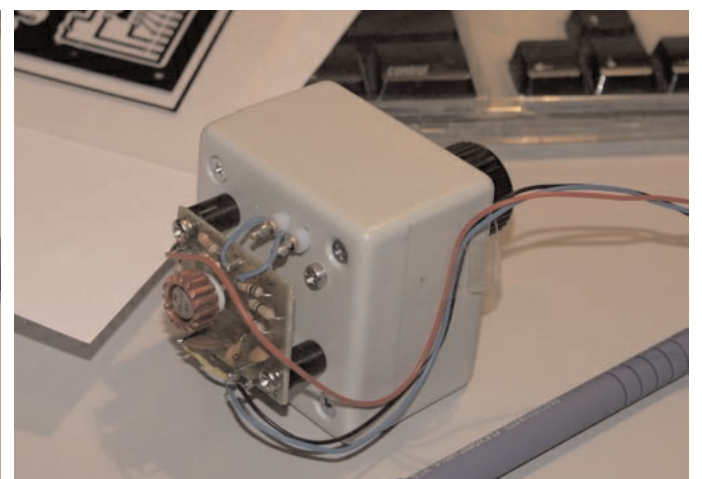
Circuit diagram



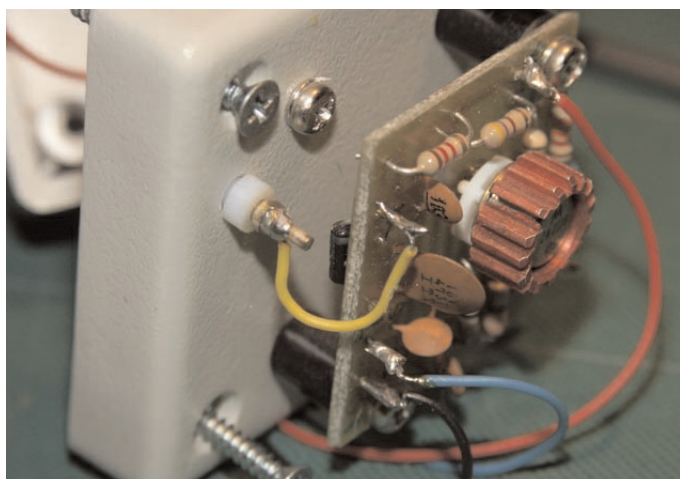
Both boards nearly built



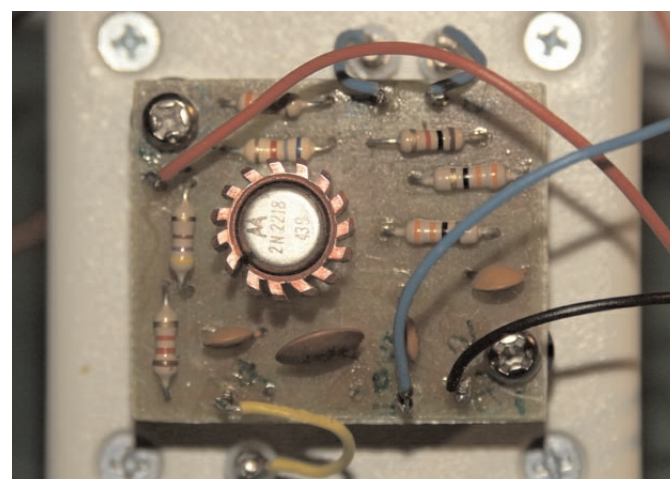
Finished - front view



Finished - back view



Fitting the back out



The back PC board in detail

The PW Portland A Rock steady VFO

You lucky readers! We've got a double helping of Tony Nailer G4CFY's designs this month! In this article Tony describes the Portland v.f.o. - for use with h.f. and v.h.f. designs, it will feature in forthcoming projects - including a 70MHz s.s.b. transceiver.

The local oscillator is a vital part of any receiver or transceiver. If it's a crystal controlled stage it's easy to make, has very low noise but only provides one frequency.

The variable crystal oscillator (VXO) trades some of its stability and noise performance for a limited frequency swing.

The digital synthesiser it is complex and moves across the band in a series of small jumps, it is also very noisy. Finally the variable frequency oscillator v.f.o. is wide range and low noise but is prone to drift.

In a future issue of *PW* it is my intention to publish a complete 70MHz s.s.b. transceiver, and several parts of this are already in existence. My choice of local oscillator is to mix a low frequency v.f.o. with a crystal oscillator, thereby gaining good frequency stability at 70MHz.

To this end I have determined that a v.f.o. with a range of 7.1 - 7.6MHz would tune the required 500kHz without a harmonic passing through the band. This, together with a 52.2MHz crystal oscillator,

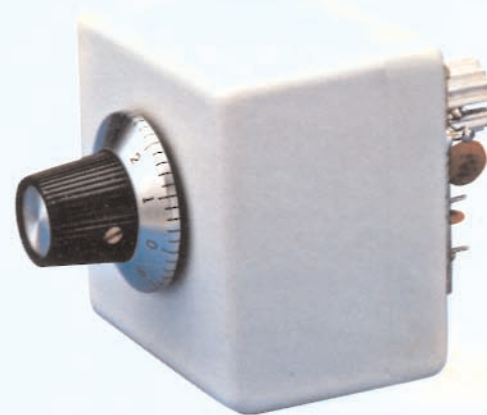
will produce a signal 59.3 - 59.8MHz for use with a 10.7MHz intermediate frequency (i.f.). Otherwise, the same v.f.o. mixed with 53.9MHz crystal oscillator will give 60 - 61.5MHz for a 9MHz i.f.

Recently the 40 metre band has been extended to cover 7-7.2MHz, so a v.f.o. operating directly over this range could also be useful for direct conversion receivers and transmitters. The v.f.o. described here includes component values for both these frequencies.

Design Considerations

Let's now look at the design considerations. Varicap tuning was chosen to keep down size and cost. Use of a good buffer amplifier to enable a signal level of 2V p-p to be delivered into a 50Ω load.

Also important is isolation of the resonant components from sources of heat to minimise drift. Rigid construction has to be adopted so that the board and resonant components would not be mechanically stressed when the tuning control was



Deceptively simple looking - the Portland v.f.o. is designed to be extremely stable for use in h.f. and v.h.f. projects (see text).

adjusted. There also has to be a 'clean' and stable supply rail to minimise amplitude and phase modulation and drift.

The requirement for the v.f.o. is that it be as 'pure' as possible and the target will be for all harmonics to be 40dB below the fundamental. This necessitates the use of a pair of Varicap diodes connected in anti-phase across the tuning coil. The reason for this is that a single diode would be forward biased by the oscillator voltage, and this would clip the wave and cause large amounts of second harmonic.

Circuit Design

An experimental Colpitts v.f.o. circuit is shown in Fig. 1. The frequency control components are R1 and VR1 where the value of R1 determines the voltage range over which the control voltage swings. Initially, I chose to use a 6V2 Zener diode stabilised supply together with a log law 10kΩ potentiometer in series with 10kΩ for R1. This would give a 3V swing for the Varicaps.

The capacitor C1 and resistor R2 allow the d.c. voltage to be applied to the Varicaps but filter the radio frequency (r.f.) signal from passing back to the potentiometer. The Varicaps being reverse biased diodes have leakage currents of around 10 nanoamps (nA). This means that if R2 was 1MΩ the voltage drop across it would be just 10mV.

Being wary of using very high value resistors because of p.c.b. losses and humidity effects, I chose R2 to be 100kΩ. The value of C1 was selected to be a 10nF

poly block which has a reactance of 2Ω at 7MHz and has low noise.

A sinewave signal on the junction of the Varicaps - which is probably in excess of 1V p-p - will see a resistance of 100kΩ then followed by 2Ω to ground. The amount of signal escaping to the wiper of the potentiometer will be $1V/50000 = 20\mu V$. (This will not radiate far, especially if connections to the potentiometer are short and if the 6.2V supply is also decoupled).

The frequency determining components are D1, D2, L1, C3, C4, C5, and C6. The design procedure follows that of the Colpitts Crystal oscillators explained in *Doing It By Design* in *PW* September 2004.

The capacitors C5 and 6 are chosen to be fairly high values, from experience 120pF is a good starting value. These two in series total 60pF. Then C4 is chosen to be close to this, and I chose 56pF. The total value of C4 in series with C5 is close to 29pF.

Inductor L1 is chosen initially as the 5.5μH TOKO coil 3334R. At 7.1MHz this resonates with 91.3pF and at 7.6MHz with 79.6pF. There is about 12pF change in capacitance between these frequencies.

The BB809 or BB109G Varicaps, with 0V bias are about 50pF and at 3V they're at about 26-32pF. Two diodes back-to-back will give a range of around 13 - 25pF (just right!). At 7.6MHz the total of 79.6pF will be made up from 13pF of the feedback capacitors plus 25pF of the Varicaps and 37.6pF in C3.

The d.c. bias components are R3, 4, and R5. Resistors R3 and 4 are equal value - and fairly high resistance - so their parallel equivalent value will not damp the resonant circuit. Values of 15kΩ were chosen. With a 6V2 supply the base of the transistor will be close to 3V, and the emitter will then be 2.3V. I also chose the starting value of emitter current to be 4mA. So $2.3V/4mA = 575\Omega$, (standard value 560Ω was used).

The d.c. input was chosen to be a standard 13.5V. If the Zener D1 current is set at 10mA and the transistor 4mA, the bias and tuning components draw 0.5mA then R6 will drop 7.3V for 14.5mA. $R6 = V/I = 7.3/14.5mA = 503\Omega$. (standard value 560Ω was used). Supply decoupler C7 was chosen as 47nF.

Dead Bug Breadboard

A breadboard of the circuit was built 'dead bug' style on blank p.c.b. material. The frequency could be tuned into the range using the coil core but the frequency swing was not wide enough and I reduced the value of R1 from 10kΩ to 4.7kΩ.

The next test achieved the required signal swing. The 3334R coil fortunately had a low impedance secondary winding. Taking the signal from this winding was considerably cleaner than that at the

emitter when observed on the oscilloscope.

It soon became clear that the frequency swing was cramped at one end of the scale, so I tried a linear potentiometer. The swing was then fairly even with rotation of the control knob, although the amplitude of the signal fell slightly across the band. Signal purity as observed on an oscilloscope was quite good and its amplitude was around 1.5V p-p unloaded.

Buffer Amplifier

Let's now take a look at the buffer amplifier. The purpose of this is to prevent loading from later stages from pulling the oscillator off frequency. It's also used to provide amplification and an output impedance of 50Ω or lower would be ideal.

I tried several buffer circuits, including a single stage common emitter amplifier, a Darlington configuration, and finally, a two stage amplifier with feedback - as used back in the 1970s. Next, I designed a common emitter amplifier was then developed as Buffer 1 as shown in Fig. 2, and was added to the original breadboarded circuit.

Buffer 1 includes an isolating resistor at the input to avoid loading of the v.f.o. The amplification then just about makes up for the step down at the input. Output was 1.5V p-p into a 50Ω load. The second harmonic distortion observed on my Spectrum Analyser was -35dB.

I made d.c. voltage measurements on the v.f.o. and Buffer 1 and dissipation of resistors and transistors were calculated. The v.f.o. transistor and bias components dissipated 28.6mW, R6 dissipated 95mW and the Zener D3 52mW. Buffer 1 dissipated 208mW of which 87mW was in R10 and 84mW in Tr2.

To reduce dissipation in the v.f.o., I increased the value of the emitter resistor R5, until the signal started to distort. A good compromise between efficiency and distortion was with a value of 1.5kΩ. The dissipation of the v.f.o. transistor and emitter resistor

then dropped from 28.6mW to 9mW.

The value of R6 was then recalculated for a v.f.o. current of 2mA and Zener current of 4mA which gave 1.2kΩ. The dissipation in the Zener was then 25mW, and in R6 44mW. (This 69mW of heat should be kept away from the resonant components).

Stability From Switch-On

Stability of the breadboard from switch on was quite good with it settling within five minutes. The stability was improved by

moving warm components well away from the resonant components. It was even susceptible to heat from my table lamp, from my body heat and from my breathing! Variations of 250Hz occurred due to these sources of heat.

Next, a two stage amplifier with feedback was designed with a low power stage one giving voltage amplification and the second one as an emitter follower. The circuit is shown in Fig. 3, and is designated Buffer 2. Using this buffer the second harmonic distortion was -40dB.

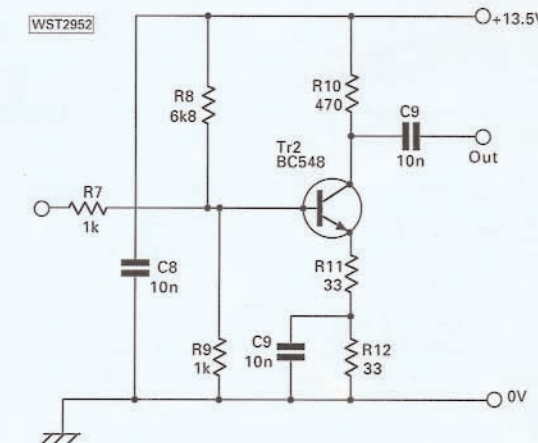


Fig. 2: Continuing the development process, G4CFY designed a common emitter amplifier, Buffer 1 as shown here and was added to the original breadboarded circuit (see text).

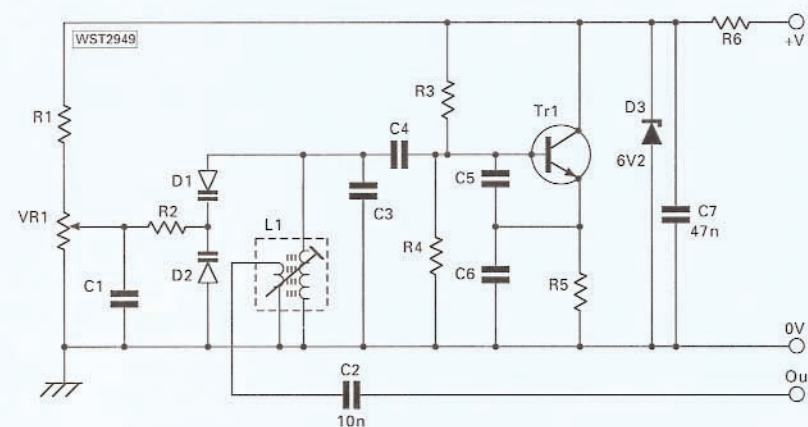


Fig. 1: An experimental Colpitts v.f.o. circuit. The frequency control components are R1 and VR1 (see text).

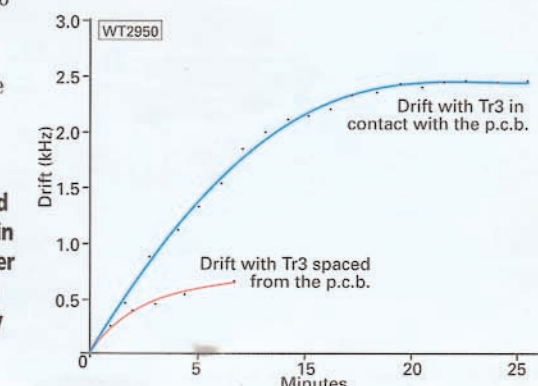


Fig. 3: Initially, from cold drift reached 2.5kHz, but a dramatic improvement in drift and time was achieved. The lower curve shown shows a total drift from switch on of only 500Hz with stability being reached within five minutes (see text).

The dissipation of components in Buffer 2 were 34.5mW in Tr2, 1.7mW in R8, 28mW R9, 8mW R10, 3.8mW R11, 204mW in Tr3 and 256mW in R12. This totals 535mW and together with R6 and D3 gives 604mW.

Measurements of output resistance revealed the v.f.o. section was 220Ω, Buffer 1 was 36Ω and Buffer 2 was 15Ω.

Prototype VFO

I decided that it would be best to incorporate the v.f.o. board within a box and have the buffer outside. (Use of a plastic box would reduce conduction of heat between buffer board and v.f.o. board. The box also had to include the tuning potentiometer to keep lead lengths down and to reduce mechanical stress between the potentiometer and the v.f.o. board.

A suitable plastic box was obtained from RS Components and I laid the p.c.b. boards out to fit in and on the box. The diodes D3 and R6 were fitted on the buffer board to keep the heat away from the v.f.o. components. Dissipation of the v.f.o. section within the box was then only 11mW.

Making a p.c.b. for the v.f.o. was the next job, and populated and fitted to the box. The unit worked first time - without any problems - and was easily tuned to the operating frequency range with a 15kHz

overlap at each band edge. Output level was 1.5V p-p.

Next, a p.c.b. for Buffer 2 was laid out and produced and populated and fitted to the outside of the v.f.o. box.

Stability Tests

At switch-on the frequency drifted upwards. This continued on for a long time and didn't reach a stable state as quickly as the prototype - "most strange" I thought to myself!

A proper test from cold was then undertaken and the results plotted as in Fig. 3. The stability was not reached until around 18 minutes from switch on - with the total drift reaching 2.5kHz.

It took me some thought to determine the cause of the problem - and I perhaps it was due to the Zener being heated by Tr3 nearby.

Removing D3 from the top of the board and fitting it to the track side greatly reduced the drift. In the end I concluded that Tr3 (being in close contact with the p.c.b.) was

also conducting heat to the Zener, so a new Tr3 was fitted on a plastic spacer.

Another thermal test was undertaken with a dramatic improvement in amount of drift and time to reach stability. The lower curve (Fig. 4) then showed a total drift from switch on of only 500Hz with stability being reached in five minutes.

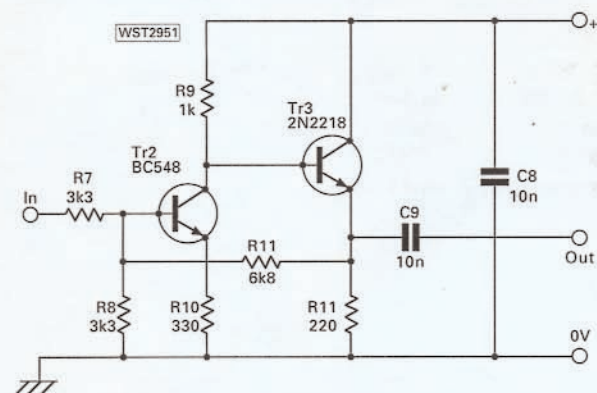


Fig. 4: A final circuit for the buffer stage (see text).

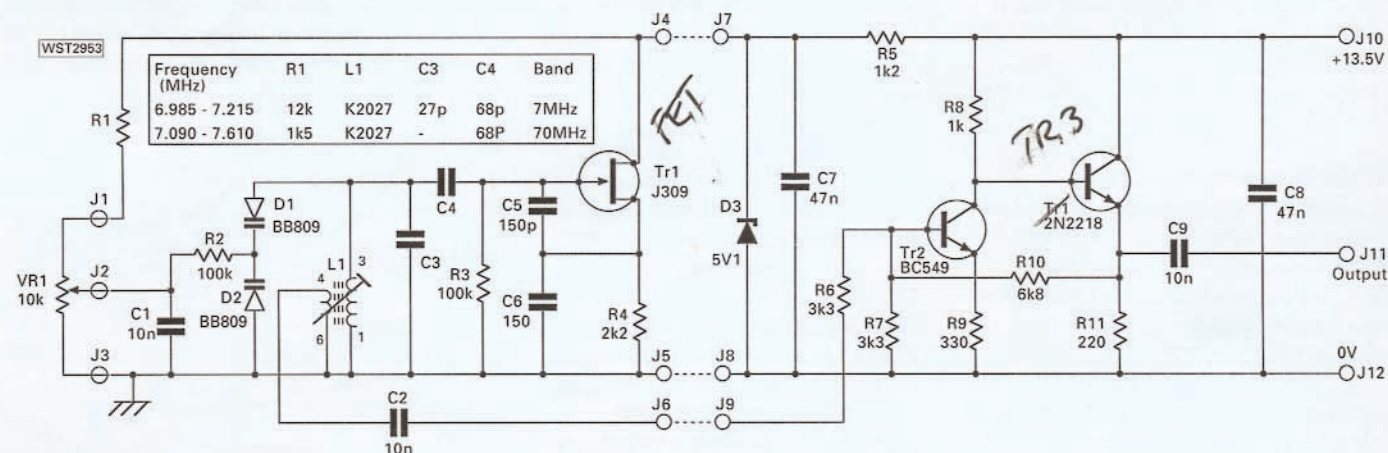


Fig. 5: The final circuit for the full oscillator project.

Variety Of Changes

A variety of changes were made in bias components and Zener current, with the box open and with the box closed. Sometimes the result was better and sometimes worse!

A change from 6V2 to 5V1 Zener was one positive move. Trying a 1N4148 in series with the Zener was not good. Eventually Tr1 was replaced with an f.e.t., R3 removed and R4 changed to 100kΩ.

During the design work over 30 stability test runs lasting from three minutes to 15 minutes were undertaken. Many of which gave suspiciously good results.

Starting tests in the morning always gave relatively high values of drift. Tests after component changes

often gave falsely good results. The problem here is that soldering applies 240°C to a localised part of the board. This permeates across the board material and along the tracks, and unless given sufficient time to dissipate masks the drift caused by d.c. and a.c. power dissipation.

Using an f.e.t. resulted in the drift going negative whereas before it had been positive (It's unlikely the base bias resistors were producing measurable heat). In fact it's more likely that the source-drain current reduces gradually as the device warms up, whereas it probably increased with the bipolar device.

With the transistor the frequency determining capacitors needed to be more 'positive' with temperature to negate the rising frequency they now need to be 'negative' going with temperature to negate the falling frequency.

The f.e.t. version of the v.f.o. used 150pF N150 capacitors for C5 and C6. Capacitor C4 was unfortunately a 68pF NPO type,

with zero temperature coefficient. At the time of going to press I had not changed this to an N150 type but had ordered some from RS. Hopefully, a negative coefficient component here might reduce the drift to a negligible level.

A final circuit is shown in Fig. 5 and new p.c.b. layouts and overlays are provided in Fig. 6.

Recommended Box

The recommended box is RS type 381 5120. You may purchase it from me ready drilled, or purchase it from RS Components and drill it yourself.

To start, I suggest you lay the v.f.o. p.c.b., copper side up, on the top of the lid, and adjust it to be equi-spaced from each edge. Next, mark the two fixing hole positions and lay the Buffer p.c.b., copper side down, on top of the lid and do the same.

Whilst the board is still in place mark a drill hole adjacent to J7 exactly mid way between the pin and the edge of the box. Do the same adjacent to J8 and J9. Mark the bottom of the box for a hole exactly in the centre.

Drill all holes using a 2.8mm (7/64th inch) drill bit. Drill out the holes adjacent to the pins with a 3.2mm (1/8th in) or 3.6mm (9/64th in) drill to suit Oxley p.t.f.e. feed-through capacitors. Drill the hole in the bottom of the box to 9mm (3/8th in) or drill it to 6.3mm (0.25in) and ream it out to size.

Assembling PCBs

Fit the pins to both boards from the copper side, ensure the heads are flush with the copper either by tapping them home with a small hammer or by using pliers. Fit the rest of the v.f.o. components as close to the board as possible.

The Buffer board will benefit from R5, 9, and 12 being slightly away from the board to aid heat dissipation. The diode D3 is intended to be fitted to the underside of the board to shield it from radiated heat from Tr3.

When the v.f.o. board is completed and has been thoroughly examined, connect it (using 7/0.2mm) wire to the potentiometer. Also, loosely wire the control to the buffer board.

Next, connect a 13.5V bench supply to the +13.5V and 0V of the buffer board and undertake measurements of the v.f.o. performance using an oscilloscope and frequency counter. Adjust the core of L1 to bring the operation into the correct frequency range. Check the swing of the v.f.o. is adequate. (If the range is too great or too small correct it by changing the value of R1 up or down to suit).

When you are satisfied the two boards are working much as they should - it will be time to secure them onto the lid of the box. Fit two 12.5mm (1/2in) screws for the v.f.o.,

Kits & Bits

As usual, kits and bits are available. Please specify which version is required, 6.99 - 7.21MHz or 7.09 - 7.61MHz.

The VFO & Buffer 2 p.c.b. £5.00.

PCB components & potentiometer £9.00.

Suitable box RS 381 5120 undrilled £4.50, drilled £8.00.

Ready built VFO and Buffer unit on chosen frequency £46.50.

Postage on PCB & components 50p. Postage on box kit or box built £1.50. Please make cheques payable to: A.J. & J.R. Nailer, and address to; Spectrum Communications, 12 Weatherbury Way, Dorchester, Dorset DT1 2EF.

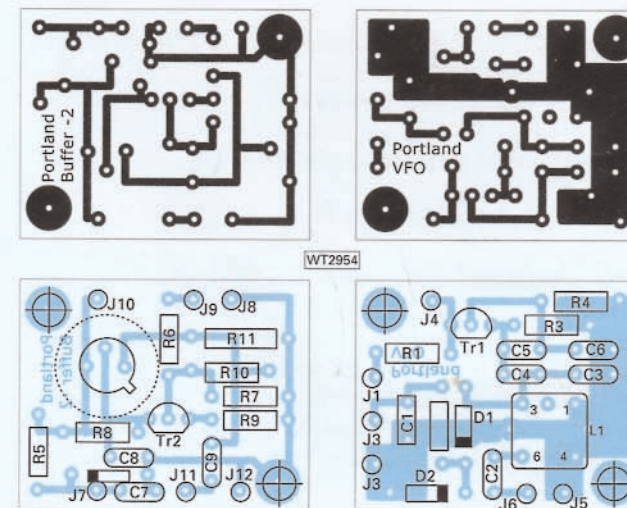


Fig. 6: The p.c.b. layout and overlay diagrams for both boards.

secure with double nuts, likewise for the Buffer. Then you can finally secure the boards in place with a further nut in each position. Push three Oxley feedthroughs into the holes provided and wire to the adjacent pins inside and out using fine tinned copper wire 26 - 30s.w.g.

Room Temperature

I suggest you leave the unit open for 10 minutes or so to assume room temperature. Then switch on and do final setting of L1 to put the frequency in the correct range with an even overlap at the band edges.

Try closing the lid and noting the frequency change then compensate by adjusting L1. Next, you secure the lid to the box and leave it for a few minutes more before undertaking a stability test.

Connect a frequency counter to the buffer output terminals. Write on a piece of paper a vertical column 0, 1, 2, 3, 4, and 5, representing minutes. When your watch is 5 to 10 seconds from the minute or half minute mark connect the bench supply and note the start frequency as the watch reaches the minute or half minute mark. Typical results are 120, 180, 100, 70, and 30Hz. Settling normally takes five minutes and the total drift is about 500Hz.

Performance Summary

Performance summary: The v.f.o. with Buffer 2 will deliver 1.6V p-p into a 50Ω

load with distortion products better than -40dB. This will suit f.e.t., m.o.s.f.e.t., and diode ring mixers in receivers and transmitters with negligible loading effects.

Knocking the side of the box also caused wild fluctuations in output frequency - presumably due to ceramic capacitors waving about. Additionally, when the p.c.b. prototype had its lid open, the pulling effect of the wires to the potentiometer caused significant changes in frequency. To obviate these effects requires embedding the frequency determining capacitors in bee's wax to hold them still. Making the connections from the p.c.b. to the potentiometer using extra-flex wire or fine enamelled copper wire - possibly adding vent holes to give limited air venting.

Another useful technique, which I haven't yet undertaken with this unit, is to heat cycle it. The process starts: First put it on a really hot central heating radiator for an hour. Then allow 30 minutes to cool to room temperature followed by 60 minutes in the freezer. Repeating this sequence twice should relieve all the stresses created by soldering and in the mechanical assembly.

Well, that's it! I hope you enjoy building the project - it's designed to be the foundation of a number of interesting larger projects. Watch this space! PW



P.W. Portland VFO & Buffer

The following 3 pages are a continuation of the previous article explaining:-

How to expand the project

Points arising from the first article and some suggested changes

doing it by design

This month Tony Nailer G4CFY delves deeper into the design of what will eventually form a 70MHz s.s.b. transmitter and describes the mixer v.f.o. for the project. As usual, for all you keen followers of DiBD - Tony has produced some helpful p.c.b. kits.



This month's project provides another step on the way to G4CFY's 70MHz s.s.b. transmitter.

In this column in March PWI published an article on the Portland - Rock Steady variable frequency oscillator (v.f.o.) with component values as a stand alone v.f.o. for 7MHz or as part of a local oscillator for 70MHz. This month I will continue onwards to develop the circuit for the 4 metre (70MHz) band local oscillator using the Portland v.f.o.

The 4 metre Amateur Radio band allocation is from 70-70.500MHz and the first consideration is that during the tuning swing of the v.f.o. there are no harmonics, which fall in that band. Now 1/9th of 70-70.5 is 7.777-7.8333. Also 1/10th of 70-70.5 is 7-7.05. Now you've been reminded of this, you should see that my choice of 7.1 to 7.6MHz for the Portland v.f.o. does not have any harmonics falling within the wanted band!

Using 9MHz SSB Filter

In a receiver or transceiver using a 9MHz

single sideband (s.s.b.) filter the local oscillator needs to be 61 - 61.5MHz. Likewise, in a receiver or transceiver with a 10.7MHz s.s.b. filter the local oscillator needs to be 59.3 - 59.8MHz. To create these frequencies requires the v.f.o. to be mixed with another crystal controlled source. In the first case 53.9MHz and the second 52.2MHz. These can be generated directly using an impedance inverting Colpitts crystal oscillator, which was the subject of Doing It by Design in PW September 2004.

Mixing the two signals together requires one of the range of mixers described in DiBD in PW May, July, and September 2005. The products of mixing need to be properly filtered and the wanted signal then amplified by a tuned amplifier (as described in DiBD July 2004, January 2005, or March 2005 PW).

With the exception of the filtering components, which have not yet been dealt with yet, all the other building blocks have been considered and

highlights the value of this series. (I agree - the DiBD approach really does work - Editor).

Mixer Block Diagram

A block diagram of the mixer - v.f.o. system is shown in Fig. 1. So, let's now take a

detailed look at the design.

To start, it's important to know that the main requirement of any local oscillator is that all feed through signals, products, and harmonics are ideally 40dB or more below the wanted signal level. The principal outputs could be feed through of the v.f.o. at 7.35MHz and the crystal oscillator at 53.9MHz and the sum of these two at 61.25MHz and the difference at 46.55MHz.

The 53.9MHz signal is the closest to the wanted output so it would be necessary to balance this out. This immediately rules out the unbalanced diode, field effect transistor (f.e.t.), and m.o.s.f.e.t. mixers discussed in PW May 2005 DiBD.

The next possible choices of mixer are the two transistor push-pull type, or the single balanced diode type described in July 2005. Here, both the v.f.o. and crystal oscillator have large signal outputs, so the twin diode mixer could be used with the subsequent filtering getting rid of large amounts of feed through of the 7MHz v.f.o. as well as rejecting the unwanted difference signal.

Comparing the twin diode single balanced type with the double balanced ring type, there are only two more diodes and we can get rid of the v.f.o. feed through signal to acceptable levels. Alternatively, the SA602 integrated circuit (i.c.) mixer could be used and the signal levels and drive impedance of its inputs are much less demanding than the diode ring.

Note: Whilst I'm well conversant with the carrier suppression performance of the diode ring, which is of the order of -50dB from each port to output, I've not explored this yet with the SA602.

Two Buffers

My Portland v.f.o. was designed with two types of buffer. The diagram, Fig. 2, shows the circuit of Buffer 1 as previously designed to achieve 1.5V p-p into a 50Ω load. It could be used here in a modified form in conjunction with the SA602 mixer, which has an input impedance of about 1.5kΩ and requires only about 400mV p-p.

The original breadboard v.f.o. with Buffer 1 was still in existence, so I undertook a test with the emitter capacitor C11 removed. The output level was then 300mV p-p, just right for the SA602. A p.c.b. for Buffer 1 was laid out, and this is now available separately (or together with) the v.f.o. p.c.b. and parts. The layout and component identification illustrations are shown in Fig. 3.

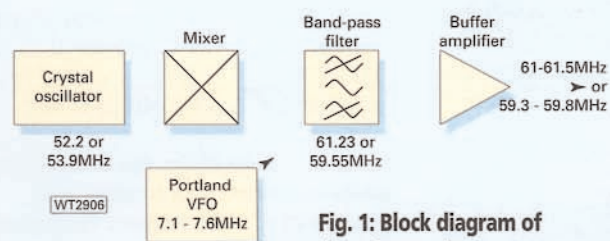


Fig. 1: Block diagram of the mixer - v.f.o. system (see text).

Moving on to the impedance inverting Colpitts oscillator, I used the board of September 2004 DiBD in the impedance inverting configuration with a surplus 52MHz crystal. On this occasion I chose to take the output from the collector where there are lower harmonics.

The circuit is shown in Fig. 4. I modified the board accordingly and when tested I found it fired on its series fundamental around 17MHz. No changes to the resistive loading seemed to alter this and I worked out that it was due to the drive level on the crystal. I altered the circuit so that the capacitor C2 coupling the crystal to the base became a low impedance only needed for d.c. blocking. The feedback capacitors C3 and C4 would then provide the resonating capacitance.

Using the 4028 TOKO coil with an inductance of 0.4μH the resonating capacitor was calculated as follows:-

$$\begin{aligned} C_t &= 1 / (39.5 * f^2 * L), \\ C_t &= 1 / (39.5 * 52^2 * 10^{-6} * 0.4 * 10^{-6}) \\ C_t &= 1 / (39.5 * 52^2 * 0.4 * 10^6) \\ C_t &= 1 / (42723.2 * 10^6) \\ C_t &= 0.0000234 * 10^{-6} \text{ or } 23.4 \text{ pF.} \end{aligned}$$

This could be achieved by two 46.8pF capacitors in series. So the feedback capacitors were changed to 47pF and the circuit re-tried. It then worked as it should on the third overtone.

A board for the SA602 as described in September 2005 was populated with the i.c. and just input and decoupling capacitors. It was then coupled to the v.f.o. and Buffer 1 and also to the Colpitts oscillator and the output viewed on the spectrum analyser. The two oscillator signals were at low level while the sum and difference signals were at high level.

Bandpass Filter

To select just the wanted product of mixing and to reject all other outputs I designed a

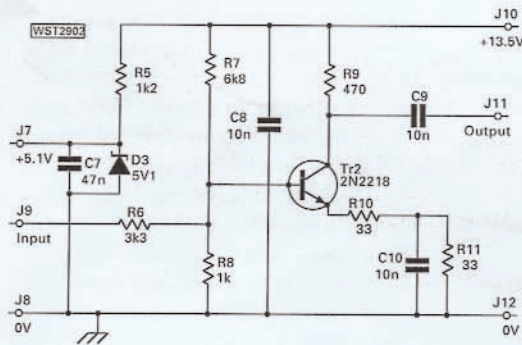


Fig. 2: The Portland v.f.o. was designed with two types of buffer. The diagram shows the circuit of Buffer 1 as previously designed to achieve 1.5V p-p into a 50Ω load (see text).

bandpass filter. The arrangement is as shown in Fig. 5, and comprises two top coupled parallel tuned circuits with a high impedance drive input and a low impedance output.

Using TOKO coils type 0764 has a Q of 80 and an inductance of 0.21μH the resonating capacitance is found as previously.

$$\begin{aligned} C_t &= 1 / (39.5 * f^2 * L), \\ C_t &= 1 / (39.5 * 60^2 * 10^{-6} * 0.21 * 10^{-6}), \\ C_t &= 1 / (39.5 * 60^2 * 0.21 * 10^6), \\ C_t &= 1 / (29862 * 10^6), \\ C_t &= 0.0000334 * 10^{-6} = 33.4 \text{ pF.} \end{aligned}$$

So L1 and L2 are 0.21μH, C1 is 33pF and C3 and C4 have a series total of 33pF.

The reactance of the coil is $XL = 2 * \pi * f * L$.

$$\begin{aligned} XL &= 2 * \pi * 60 * 10^6 * 0.21 * 10^{-6}, \\ XL &= 2 * \pi * 60 * 0.21 = 79.168 \Omega. \end{aligned}$$

The dynamic resistance $R_d = Q * XL$. $R_d = 80 * 79.168 = 6333 \Omega$.

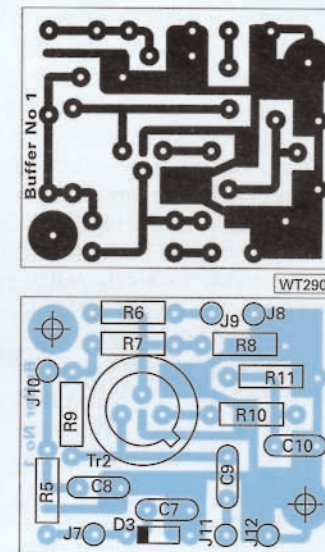
To feed the base of an amplifier transistor stage requires this to be reduced to about 300Ω, which represents a step down of 20:1. As a first approximation this will require a capacitance ratio for C4 to C3 of 4.5:1. If C3 is 39pF then C4 needs to be about 175pF.

Trying 39pF and 180pF in series gives:-

$$\begin{aligned} C_t &= C_3 * C_4 / (C_3 + C_4), \\ C_t &= 39 * 180 / (39 + 180), \\ C_t &= 32 \text{ pF. (Close enough!)} \end{aligned}$$

Incidentally, experience has shown me that critical coupling between TOKO coils occurs when the top coupling capacitor is around a fifteenth of the value of the resonating capacitance. In this case then C2 would be 2p2.

Tests with the bandpass filter connected to the output of the mixer did the job of selecting the required product and rejecting the unwanted signals. The signal level was about 250mV p-p and would need to be



amplified for use to drive a diode ring front end mixer.

Final Circuit

For the final circuit a common emitter amplifier was designed and added to the breadboard and is shown with the complete circuit in Fig. 6. Originally R7 was 10kΩ, R8 1kΩ, R10 was 47Ω and C15 was fitted. The gain was then far too high and the output stage would oscillate when input and output were in tune.

Next, capacitor C15 was removed so that the gain would be mainly set by the ratio of the dynamic resistance of the tuned circuit (L4 C17 C18) divided by R10, 47Ω. The output dropped significantly so a lower value of R10 was chosen and R7 and R8 recalculated to keep the emitter current about 5mA.

The process continued until the values shown were arrived at to achieve the maximum output before the onset of instability. With these values the coils L2 L3 and L4 could all be tuned or peaked without oscillation and with an output of around 1V rms, 2.8V p-p into the 50Ω load.

With any mixing system there will be the generation of intermodulation distortion products and these should be 40dB below the wanted output. This unit has these products generally better than 45dB down.

Portland Tuning

The Portland v.f.o. tuning 7.1-7.6MHz has its 8th harmonic sweeping across the range 56.8-60.8MHz. At the bottom end it is well away from the pass band of the wanted output, and is about 42dB down from the wanted signal.

Unfortunately, at the top end it's only 200kHz from the band edge and hence is only 32dB down. It means that when using a 9MHz i.f. and listening to an S3 signal on 70.5MHz, an S9 signal on 70.3MHz would produce an equal size interfering signal in the i.f. pass-band. It would create an even greater problem with the version for use with a 10.7MHz i.f. as the output band is then 59.3 to 59.8MHz and this harmonic of the v.f.o. on 7.1-7.6 sweeps across the range 56.8-60.8MHz.

The best solution to the problem is to use a different v.f.o. frequency and different mixer crystal. Having determined that 1/9th of 70-70.5 is 7.777-7.8333, and 1/8th of 70-70.5 is 8.75-8.8125MHz, so there's

Fig. 3: A p.c.b. for Buffer, this is now available separately (or together with) with the v.f.o. p.c.b. and parts. The layout and component identification illustrations are shown here (see text).

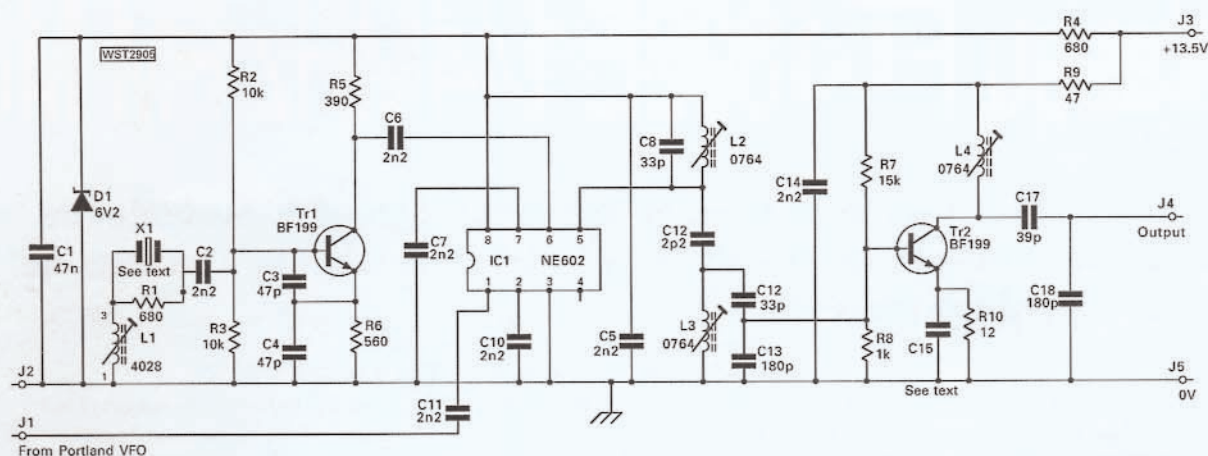


Fig. 6: A common emitter amplifier was designed and added to the breadboard and is shown with the complete circuit (see text).

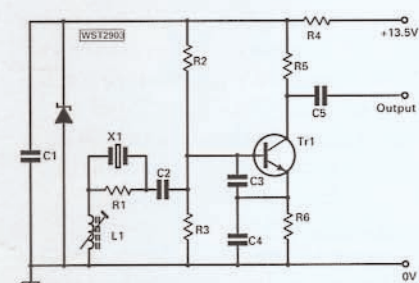


Fig. 4: The impedance inverting Colpitts oscillator, where G4CFY used the board of September 2004 DiBD in the impedance inverting configuration with a surplus 52MHz crystal (see text).

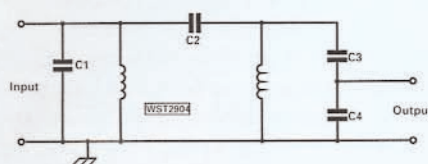


Fig. 5: The bandpass filter (see text).

another window between 7.8333 and 8.75MHz.

The 7th harmonic of 7.8333-8.75MHz is 54.8-61.25MHz, which overlaps with the local oscillator frequency of 59.3-59.8MHz of the 10.7MHz i.f. version. Trying 7.9-8.4MHz gives the 7th harmonic as 55.3-58.8MHz which avoids the 59.3MHz by 500kHz. This means that (contrary to my earlier work) that a v.f.o. on 7.1-7.6MHz would be ideal for a local oscillator for a 4 metre rig using a 9MHz or 10.7MHz i.f. I now find (as you do quite often) that there's a better solution.

The requirements for a 9MHz i.f. are a local oscillator on 61-61.5MHz produced by mixing a 53.1MHz crystal with a Portland v.f.o. running 7.9-8.4MHz.

The requirements for a 10.7MHz i.f. are a local oscillator on 59.3-59.8MHz produced by mixing a 51.4MHz crystal with a Portland v.f.o. running on 7.9-8.4MHz.

Being now confident that I've found the best v.f.o. and crystal oscillator frequencies to achieve the required result, a p.c.b. was laid out and is shown together with a component identity in Fig. 7.

Original Retuned

My original Portland v.f.o. which tuned 7.09-7.61MHz was then re-tuned to the required range and produced 7.855-8.455MHz. Note: This is a 90kHz excess swing which can be reduced by increasing the 1.5k Ω resistor in series with the tuning potentiometer. I changed this resistor to 3.3k Ω and achieved a swing of 7.88-8.42MHz.

That completes the work and it's really impressive to switch on the combination and observe a 60MHz local oscillator signal that just doesn't seem to drift. Oh how we would have loved this unit back in the crystal controlled days of a.m. on 144MHz in the early 1970s!

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