

**THIS
MONTH'S
LEADING
FEATURE**

Third-Method SSB HF Transceiver

*The first of five parts by
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THIS CONSTRUCTION project covers the theory and practice of building a 'Third Method' transceiver covering all nine HF bands. The basic theory of the Third Method and an outline of the project will be discussed in this part.

Details of the circuit of the RF side will be covered in Part 2, and Part 3 will describe construction practice in general and the RF board in particular.

Part 4 covers the AF side, employing the Third Method for sideband removal and Part 5 completes the transmit chain and details the alignment procedure.

IN THE BEGINNING . . .

THIS PROJECT WAS conceived from base motives. Mostly revenge against the home-brew spirits who, some thirty years ago, tempted a youthful G3XJP into his first Third Method effort using old fire elements as delay lines. With the benefit of hindsight, it could never have worked, but the thought was always there in the background that the method was basically elegant.

The original source of inspiration was the definitive article by D K Weaver [1]. The

treatment is somewhat mathematical, but rest assured that you can implement the Third Method using nothing more complicated than simple addition and subtraction. The topic is also covered in the *Radio Communication Handbook* [2] which concludes "... it provides the basis for high-performance receivers at relatively low cost, although suitable designs for amateur operation have yet to be developed."

My other base motive was a desire, with no technical justification, to do something different. Having always run a home-brew station, I have built several filter and phasing transceivers, mostly based on the ubiquitous Plessey SL600 ICs and most of them have worked very well. So although this project was started for all the wrong reasons, it has turned out to be easily the best transceiver I have ever built, comfortably outperforming all previous efforts in every department.

PROJECT PHILOSOPHY

WHY BOTHER with something as esoteric as a philosophy? Why not just build it? Home brew projects come in many forms - at one extreme there are 'cook-book' kits of parts complete with commercially produced PCBs. If they work first time and then stay working, there is never a need for any depth of understanding. At the other extreme one gets concepts or snippets of ideas which need develop-

ing and integrating to produce a useable end result.

This project is somewhere in the middle. Certainly, there is enough detail provided in order that you could build a close replica of my transceiver. But that would be to waste the opportunity of customising the design to your particular needs. All designs consist of trade-offs and compromises, and your particular operational needs might cause you to make radically different decisions to me.

The full project is aimed at the experienced constructor with much of the non-critical mechanical detail left to your discretion. Be warned, it requires some serious commitment. Alternatively, you could build and use it in phases or you may just want to treat it as a source for some ideas.

For these reasons (and because the Third Method is, at the least, unusual) significant space is devoted to the theory, so that even if in the end you decide to copy the design closely, you will understand why it works.

AS AN ASIDE

THE PROJECT has deliberately omitted features in my transceiver which are not germane to the Third Method. I mention this because the eagle-eyed amongst you will doubtless spot the evidence in the photographs. For the record, these are VOX, a receive audio notch filter, a frequency counter, servo control of my pump-up mast and the beginnings of some DSP work. The plug-in circuit boards were removed and ICs lifted from their sockets before photography to minimise any confusion.

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THE THREE METHODS COMPARED

THE OTHER TWO classic methods are of course the 'Filter Method' and the 'Phasing Method'. As a basis for comparison, the common amateur configuration of a single conversion transceiver is assumed, with the IF somewhere in the middle of the HF spectrum, say at 9MHz. To get the full picture, it is essential to compare the totality of the systems, not merely the differing mechanisms for removing the unwanted sideband; because the latter has important repercussions on the line-up.

Most attention is devoted to the receive side because by comparison, transmitters are easy. But note that the block diagrams which follow for all three methods can be turned round to fire the other way, simply by using the loudspeaker as the logical microphone.

A design goal of operation on most of the HF bands is also assumed, otherwise the intrinsic advantages and disadvantages of the different methods still apply, but take on different weights.

The 'Filter Method' relies on what is essentially a brute force mechanism, namely a crystal filter which passes the wanted spectrum and rejects the rest. Its advantages lie in simplicity of the signal path and ease of strong signal handling - given good gain distribution. The filter also provides some extra carrier suppression and its bi-directional nature lends itself readily to transceive operation. Fig 1 shows the basic idea and well-known designs by G3CLF, G3TXQ, G3ZVC and G3TSSO [3] follow this pattern.

These advantages come, literally at a price. Good crystal filters are increasingly hard to come by and are, against the general trend, getting more expensive. Even the Plessey ICs which in practice underpin these designs, although offering superb functionality and very

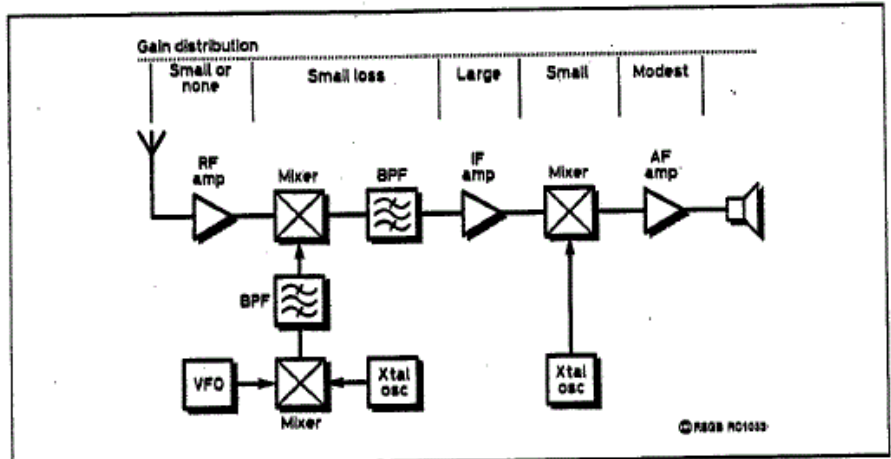


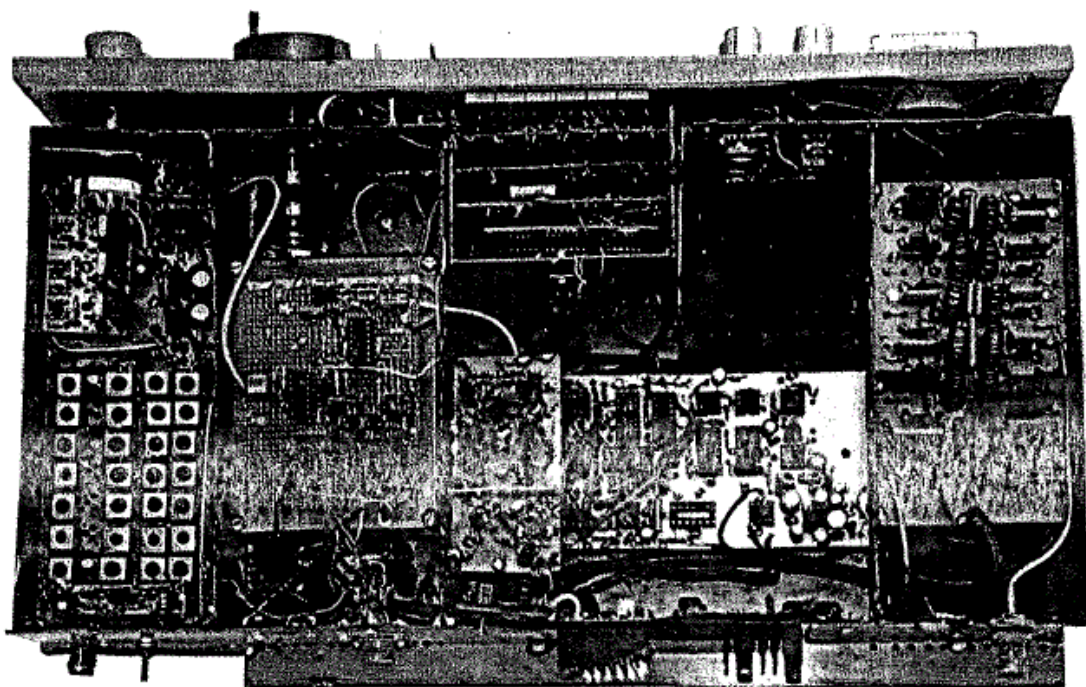
Fig 1: Filter Method receiver basic block diagram. This establishes a frame of reference to which the other methods may be compared. The desirable gain distribution is shown to emphasise that gain must be held to the minimum necessary to overcome internal noise until the bandwidth of the incoming signal has been restricted.

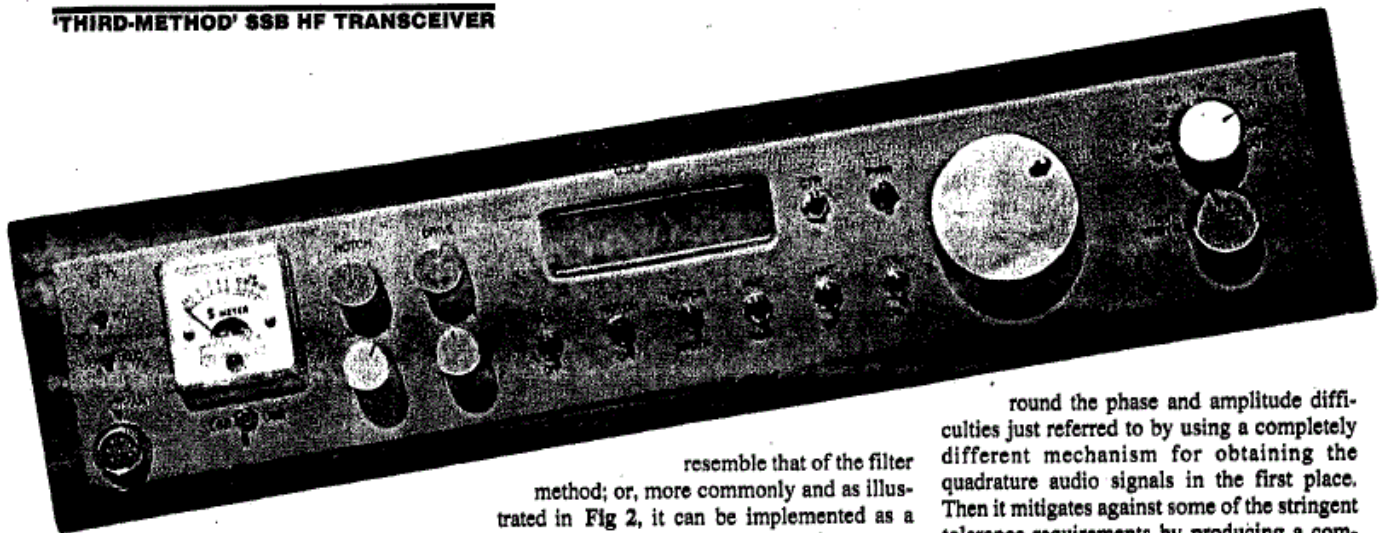
low associated component count, are now getting rather costly.

More fundamentally, the filter approach does have some intrinsic drawbacks which don't normally trouble us - given there is little we can do about them:-

- the bandwidth of the filter is fixed (or you need the inelegant solution of several filters and difficult associated switching)
- the pass band occurs at a fixed frequency. Imagine how easy life would be if a crystal filter could be tuned across any 500kHz of your choosing or even over the whole HF spectrum! So there has to be at least one mixer between the antenna and the filter which presents, so to speak, the chosen 3kHz (out of the 30MHz available) to the crystal filter for treatment. More literature has been devoted to the art of avoiding intermodulation distortion (IMD) products in and around this mixer than to any other topic in HF receiver design.

- for a single conversion configuration some mechanism for generating the multi-band injection frequencies to this mixer is required, usually at a significant power level. This typically involves either a VFO mixed with switched crystals, switched multi-band tuned circuits and a power driver - or the more esoteric realms of frequency synthesis. In any event, it is non-trivial to implement.
- at 9MHz, preventing leak around the filter requires a lot of care to achieve the theoretically possible ultimate stop band rejection.
- the IF amplifier following the filter on receive - whether broad band or relatively narrow band - invariably generates enough noise at the unwanted sideband that it appears in the output after product detection. This is a very real problem with some amateur designs which have a 90dB IF strip following the filter. The only real cure is yet





another filter before the product detector, albeit of more modest performance. Of all the problems, this is the most intractable.

So, given this apparent recipe for disaster, why is this easily the most popular form of amateur home brew construction? Mostly, I guess, because it is simple, requires a minimum of adjustment and is repeatable; and none of the above difficulties is insurmountable.

The 'Phasing Method' as the name implies, phases out (as opposed to filters out) the unwanted sideband. It employs both an RF and an audio phase shift network to achieve this. The G3TDZ design [4] is a recent implementation using modern components. With readily available high tolerance Rs and Cs together with cheap high performance op-amps it is relatively simple to build an audio phase shift network which will produce a 90 degree shift.

The problem is that it is difficult to hold the required tolerance except over a limited bandwidth of about 300Hz to 3kHz. This is fine on transmit where the speech can readily be constrained to this width. But on receive, unwanted signals more than 3kHz or less than about 300Hz on either side of zero beat will produce an output. The polyphase network [5] is one approach to overcoming the problem, but is somewhat cumbersome.

In principle, large signals tens or even hundreds of kilohertz off tune pass right through the receiver with all the usual IMD risks. The solution is a high grade audio filter and the requirement is for a shape factor no less good than a crystal filter. A low pass filter (LPF) is not sufficient because of the need to remove the low frequency responses. In practice, however, an LPF is usually pressed into service for expediency and the lower frequencies are rolled off with suitable coupling capacitors.

The net result is very acceptable but the fact that the receive bandwidth is only limited fairly late in the chain means that a lot of attention has to be paid to low noise and dynamic range in the front end - with a substantial part of the receiver gain placed after the filter in the audio chain.

The RF phase shift network is much less problematic. It can either be implemented at a fixed intermediate frequency - in which case the RF stages, mixer and injection circuitry

resemble that of the filter method; or, more commonly and as illustrated in Fig 2, it can be implemented as a tuneable IF. This implies the requirement to hold the phase shift over 500kHz or so. This can be done readily if the IF is high enough such that 500kHz is a relatively small percentage.

In any event, the overall requirement is to hold both phase quadrature between the two channels to within less than 1 degree and amplitude inequality to less than 1% to obtain a mere 45dB of sideband suppression. This may be just about acceptable for a transmitter, but is not exciting performance for a receiver. And remember, this has to be achieved both for a 500kHz change in VFO frequency - and over at least the full audio spectrum - and over a reasonable temperature range - and most importantly, sustained without constant adjustment over time! These figures can be obtained (or deviations at the extremes tolerated) but not easily.

The practical symptoms of any problems arise when all works well in the middle of the band, but falls off a bit near the edges. Then, when you switch sideband (in practice, by moving to a band which requires the other sideband), the whole thing turns out to be unacceptable and no amount of adjustment will give a realistic compromise on both USB and LSB.

The 'Third Method' (see Fig 3) is, at first glance, a mixture of the first two. It aims to get

round the phase and amplitude difficulties just referred to by using a completely different mechanism for obtaining the quadrature audio signals in the first place. Then it mitigates against some of the stringent tolerance requirements by producing a completely different effect to lack of sideband suppression as the penalty for non-conformance. This important 'completely different effect' is elaborated in a moment and is the secret of Third Method success. The benefits are bought at the price of somewhat increased circuit complexity; and it is somewhat more taxing intellectually also!

The front-end and RF phase shift approach is identical to the Phasing Method. So, they both share the advantage of not requiring a VFO/crystal oscillator/mixer or equivalent arrangement for injection purposes. This is a significant benefit since in practice this part of the circuit alone ends up much like a multi-band QRP CW transmitter. For a single band implementation, it is possible to dispense with the front-end conversion arrangement altogether, giving what is in effect a Direct Conversion format. In principle, it would be possible to extend this approach by having a separate front-end, VFO and RF phase shift network for each band, but this soon becomes a switching and adjustment nightmare. It is not considered realistic, but in rejecting this approach it does mean that there are two mixers in the signal path before the filter.

To overcome the problems associated with an audio phase shift network, a further mix with quadrature injection is used instead.

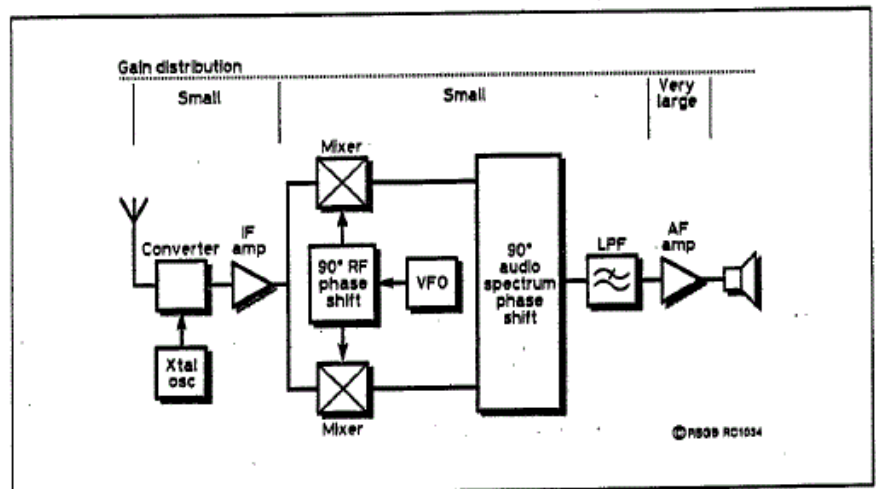


Fig 2: Typical Phasing Method receiver. It might be worth considering the insertion of a low pass filter in both channels after the mixers. This would significantly reduce the broadband assault on the phase shift network.

This approach has some important advantages.

The required injection is at a fixed frequency - known as the Pilot Carrier frequency. This frequency is chosen somewhere near the centre of the voice spectrum for SSB applications, ie at about 1.7kHz. At this low and fixed frequency, amplitude and phase balance are readily achieved. The mixers themselves can actually be switches - as in the text book classic description of a double balanced modulator. No balance adjustments are needed. I have even used reed relays to satisfy my curiosity. They worked fine except for the need to sound-proof them against the 1.7kHz buzz! These switches can be controlled digitally giving hard, fast switching. The required 90° phase difference is easily implemented digitally as well, giving a precise phase shift and exact sideband switching. The net result is a broad-band quadrature phase shift which holds tolerance from near DC up to tens of kilohertz.

A low pass filter is needed in each audio channel with a cut-off frequency around 1.4kHz. At this frequency it is possible to achieve excellent performance with little complexity. The MAX 293 8-pole elliptic filter chip is used which is 20dB down a mere 30Hz above cut-off and 75 dB down at 750Hz. The cut-off frequency can be effectively controlled by a front panel pot, giving variable receive width from near zero to several kilohertz as a free bonus. This represents a distinct improvement over old fire elements!

So much for comparisons, but how does it actually work? The transmit mode is discussed first, then the receive side; but note that almost all the elements are common to both.

ON TRANSMIT

MICROPHONE AUDIO is fed in parallel, essentially untailored, to two modest gain audio amplifiers - one for each channel.

Following one of these channels for a moment, the audio is mixed with the Pilot Carrier at about 1.7kHz in a double-balanced mixer. This mix produces the normal sum and difference frequencies as shown in Fig 4.

Assuming the original voice spectrum runs from, say, 30Hz to 10kHz, the sum output from the mix will be 1.73 - 11.7kHz and the difference will be 1.67 - 0kHz. This composite spectrum then goes to a low pass filter (LPF) with a cut-off at 1.4kHz. This strips off the sum and all the difference frequencies above 1.4kHz, leaving 0 - 1.4kHz (twice), which corresponds to the original voice spectrum components between 300Hz and 3.1kHz.

Thus the LPF has stripped off the unwanted low and high frequency components and confined the output to a nominal 2.8kHz band-

width. For the price of a low-pass filter, you actually get a band-pass filter!

Rather alarmingly, both high frequencies and low frequencies in the voice produce output in the range 0-1.4kHz, a process known as 'spectrum folding'. The nature of the mixing process is such that they are displaced by 180 degrees - which is the secret of sorting them out later.

The same process occurs in the other channel (as shown in the block diagram, Fig 5), but because the Pilot Carrier injections to the two channels are in quadrature, so are the resulting outputs. After buffering into the diode ring mixers, both channels are mixed with the VFO. This tunes 8.0 - 8.5MHz. In each channel, sidebands are produced which extend 1.4kHz either side of the VFO frequency. The two channels are combined in a quadrature hybrid which imparts a further mutual 90° phase shift. The various sideband components - there are 8

of them by now - reinforce (in phase) and cancel (out of phase) to leave a single sideband suppressed carrier signal centred on the VFO frequency. The nominal zero beat frequency - which we conventionally think of as the operating frequency - is displaced from the VFO frequency by the amount of the Pilot Carrier frequency, ie 1.7kHz.

Following a broad-band IF amplifier, the signal is now mixed with the output of a crystal oscillator to the required amateur band. The signal passes through the band-switched preselector (also receiver front-end) in order to suppress the image frequency from the mix. A broad-band RF driver and a commercial PA from Cirkit raise the power level to about 15W.

A low pass filter completes the line-up by removing any harmonic content.

In operation, any lack of carrier suppression will be received as a 1.7kHz tone. Tones at zero beat and at 3.4kHz result from any lack of Pilot carrier suppression. As implemented, they are better than 100dB down without adjustment and can be ignored.

The other consequence of the Third Method is that for all practical purposes, the conventional opposite sideband (ie on the other side of zero beat) does not exist; so I am reluctant to use the term 'sideband suppression' without qualification.

Imperfect reinforcement/cancellation of the sideband components will produce an inverted voice spectrum super-imposed on the wanted voice spectrum. That is, if you whistle a note of falling pitch,

this will be reproduced in a receiver along with a weaker whistle of rising pitch.

We all have a low tolerance of a different but inverted signal on the same channel ie interference from a station transmitting on the opposite sideband. However, when listening to normal speech, we are amazingly tolerant of a good deal of the same signal but inverted. In fact most people won't comment, even if asked, until the inverted signal is only about 20dB down.

This, in Third Method systems, is 'sideband suppression'. In practice, 40dB or more is easily achieved giving a characteristically smooth and rounded hi-fi feel from what is actually a tight and narrow signal.

This suppressed inverted signal is the 'completely different effect' referred to earlier. For the record, Reference [2] disputes the claimed benefits on the grounds that a modern narrow receiver monitoring your transmitter would remove any conventional unwanted sideband while providing no rejection of an inverted signal. True, but I find this argument wholly unconvincing. The idea is to generate a narrow signal for the benefit of those on adjacent channels who are not trying to listen to you. This is independent of the nature and quality of any device at the receiving end.

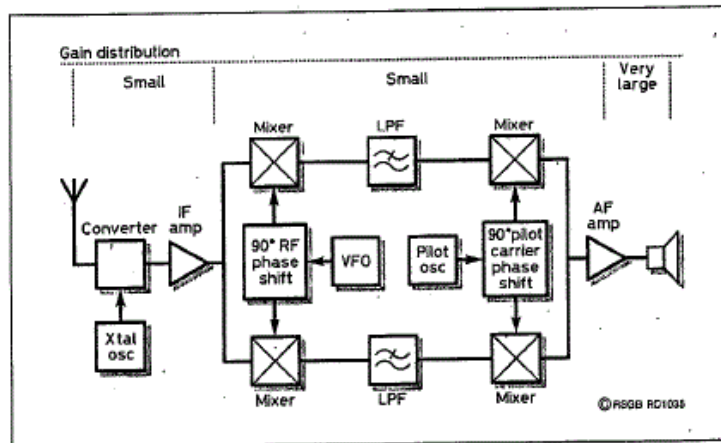


Fig 3: Third Method receiver basic block diagram. It is similar to the Phasing receiver but with a different method for generating the audio phase shift.

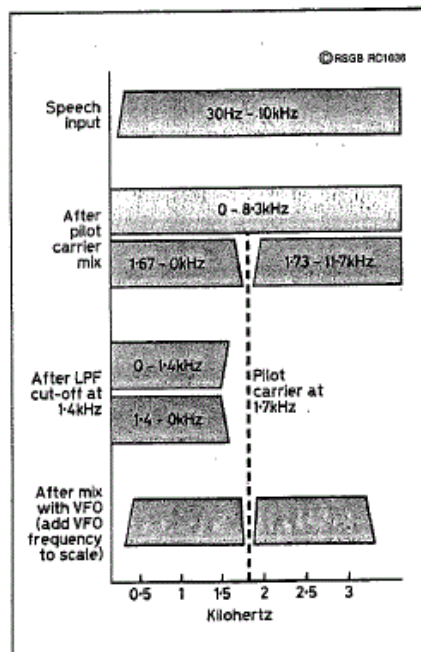


Fig 4: Third Method spectra at various stages in the transmit chain. Note a narrow notch in the output at the Pilot Carrier frequency. This arises in practice by not using DC (zero Hertz) coupling. It is of no practical consequence on either transmit or receive.

When it comes to receive, the importance of minimising responses to any signals on the other side of zero beat is even more critical. This is where the Third Method really scores over a conventional Phasing receiver.

ON RECEIVE

THE CRITICAL INSIGHT is to note that in tuning in a signal using the Third Method, you actually align the VFO to the centre of the spectrum of the wanted signal. This is as opposed to both the filter and phasing methods where you tune to zero beat with the suppressed carrier. If this sounds immediately off-putting, let me emphasise that this process is completely transparent to you. You simply tune, as usual, until the resolved incoming signal sounds right. But as a matter of fact, this puts the VFO frequency 1.7kHz away from zero beat.

Referring again to Fig 5, the signal from the antenna passes through the band-switched low pass filter, bandpass preselectors and pre-amplifier. The spectrum, now limited to a nominal 500kHz, is mixed to the tuneable IF of 8.0 - 8.5MHz. The post-mix IF amplifier provides some gain, but is there mostly to offer a broad-band termination to the mixer and further AGC action.

The spectrum is now split into two 90° mutually displaced components, each of which is mixed with the VFO output (8.0 - 8.5MHz) to give quadrature audio outputs.

It is conventional to impart the quadrature displacement to the VFO output as in Fig 3, not to the RF spectrum. The reasons for doing it the other way round will be covered later in the project.

At this stage, the net receiver gain is close to zero. Each of the outputs is first amplified by a low noise audio stage with significant roll-off at higher frequencies - and from there to a low pass filter. This strips off any unwanted outputs above 1.4kHz.

These unwanted outputs are produced by any signals either higher or lower in frequency than the edges of our wanted signal; as on transmit, we get a band-pass effect from the LPF.

Each channel is now mixed with the Pilot Carrier, introducing a further 90° phase shift. The mixing process 'unfolds' the audio spectrum and the results sum to reinforce one sideband and cancel the other. An audio amplifier completes the receive chain - not forgetting a good speaker.

On this transceiver, the latter makes all the difference. Listening to a good signal on a quiet band with the receiver band width opened up to about 4kHz reminds one of the days of 10m

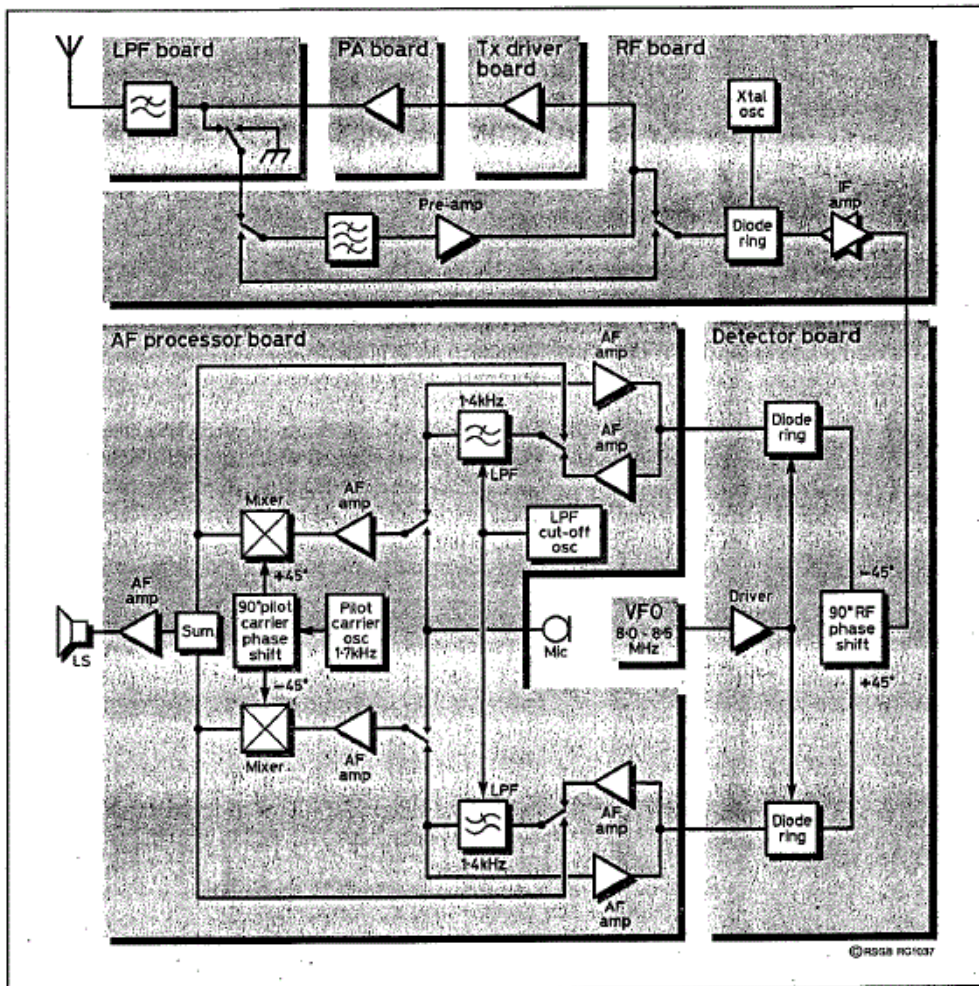


Fig 5: G3XJP Third Method transceiver showing both functional blocks and mechanical units. Not shown are band switching, power supplies, AGC and T/R switching control. Note that all the 'relays' shown, with the exception of the antenna c/o relay on the LPF board, are implemented by solid state switches.

AM in the late '60s. But equally, when the going gets tough, you can progressively restrict the bandwidth as necessary to maintain readability.

IN PRACTICE

I don't have access to the sophisticated test equipment needed to measure the absolute performance of this transceiver reliably. But let me give you some subjective observations against my FT200, an ancient but robust performer.

I can put a rock crushing signal into my amateur near neighbours on 80m. They cannot detect any residual carrier, pilot tone or conventional other sideband. They are not easily detectable for that matter on my own FT-200. The locals report the in-band audio quality, in brief, as excellent.

On the receive side, the obvious test is to tune across an S9+ carrier, using an SSB bandwidth. As you tune through zero beat, it literally disappears at about 200Hz and does not reappear out the other side. I would be happy to conduct an S2 SSB QSO with another S9 QSO on the opposite sideband. So, I make no claims for infinite sideband suppression, but there are certainly no problems in this department.

The only niggle is some trace of receive

cross-modulation if there are very strong local signals on the band. It has always been possible so far to hide the evidence by backing off the RF Gain - and it has given no operational problems. The up side of this issue is plenty of receiver sensitivity on all bands.

Receive quality is outstanding and certainly exceeds the transmit quality of many stations who are using the normal range of commercial equipment. It also beats the FT-200, which uses valves, and has a pleasant enough sound. The variable receive bandwidth is, for me, the winning feature on our more crowded bands.

REFERENCES

- [1] *Proc IERE*, December 1956.
- [2] *Radio Communication Handbook*, 5th edition, pages 4.3 and 6.82, RSGB.
- [3] *Radio Communication Handbook*, Chapter 7, RSGB.
- [4] 'Multi-Band Phasing Transceiver' by John R Hey, G3TDZ, *Radio Communication*, July 1993.
- [5] *ARRL Handbook*, 1994, page 18-9. ♦

... to be continued

Third-Method SSB HF Transceiver

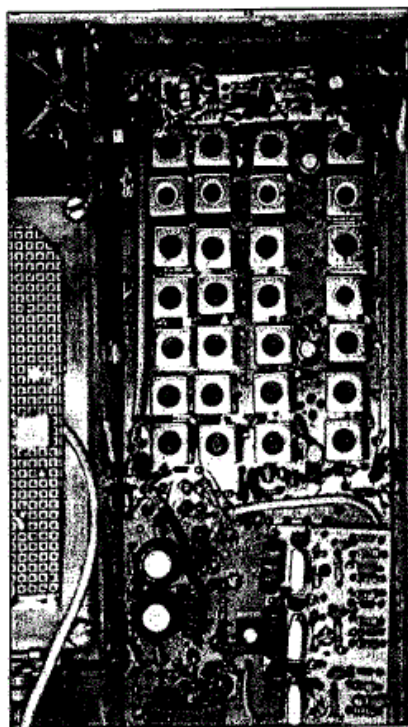
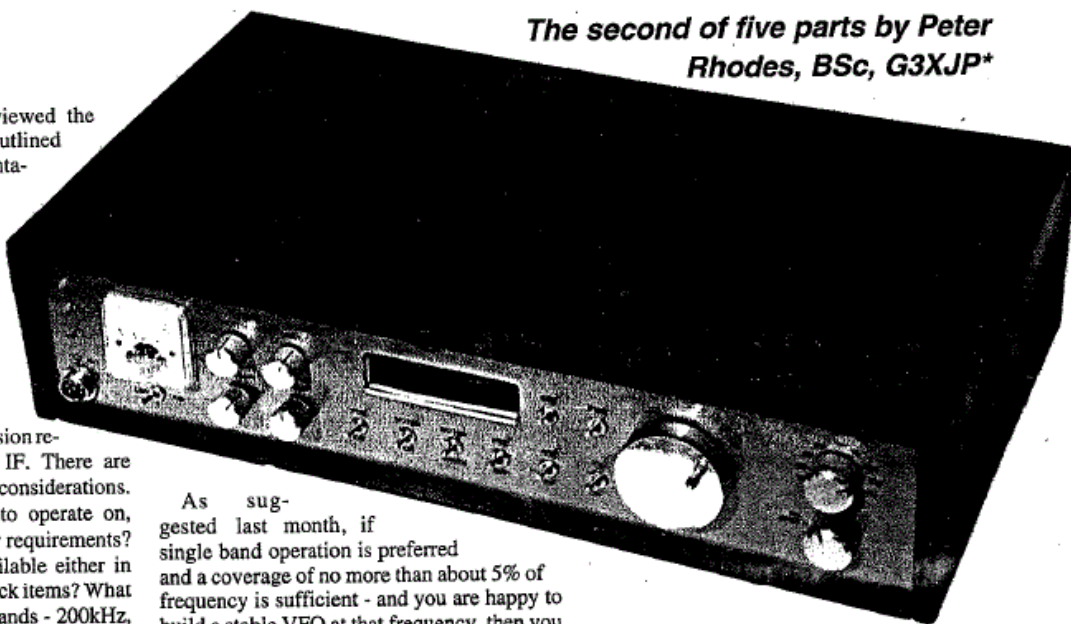
*The second of five parts by Peter
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LAST MONTH I reviewed the Third Method and outlined the G3XJP implementation. In this issue I will set out some design choices and criteria. The RF board which acts both as the front end for the receiver and as the low level driver for the transmitter, will also be described.

OVER TO YOU . . .

YOUR CRITICAL design decision revolves around the choice of IF. There are several potentially conflicting considerations.

What bands do you want to operate on, including any VHF transverter requirements? What crystals are readily available either in your junk-box or as cheaper stock items? What segments do you want in the bands - 200kHz, 500kHz or more? What sort of dial readout is preferred and does this impose any limitations on mechanical construction or on the direction of VFO tuning versus transceive frequency?



Detail of transceiver construction.

As suggested last month, if single band operation is preferred and a coverage of no more than about 5% of frequency is sufficient - and you are happy to build a stable VFO at that frequency, then you can dispense with an IF altogether. This would give about a 100kHz segment on Topband, 200kHz on 80m and full coverage of 40m.

You could plan to add converters for other bands to this later, but I would not recommend this approach in the long run because there are probably better choices of IF available for most applications.

40m is tempting arithmetically, but it is a disaster from IF-breakthrough considerations. Having said that, it is a fairly simple matter to change the IF subsequently anyway, so this might appeal as a quick way to get started.

For two band operation, you may prefer one of the pairs of bands which require the same injection, such as 20m and 80m with 5MHz VFO and 9MHz IF (or equally, 9MHz VFO and 5MHz IF). Actually, all you are saving is a crystal oscillator and some simple switching.

My personal requirement is for SSB operation on 80m, 17m and 10m. This is one of the trickier combinations. I chose an IF of 8.0 to 8.5MHz because this gives reasonable coverage of 10m with a 20MHz crystal and 17m with a 10MHz crystal. I initially used a 4.5MHz crystal for 80m. A 12MHz crystal would have been better, except that at the time, I had a digital readout which counted upwards only.

Then I sorted out my counter limitation and the net result is that I use eight crystals at 6, 10, 12, 13, 15, 17, 18 and 20MHz for nine band coverage with the lower edge of the IF extended to 7.8MHz on bands I rarely use. These

crystals are mostly not difficult to obtain.

One consequence of this particular choice is that you tune in an SSB signal by rotating the tuning knob in the same direction on all bands. Another, is that you always use the same sideband; so the sideband select switch is labelled Normal/Reverse and not USB/LSB and there is no additional band switching requirement. If necessary, you can optimise performance on this sideband, but you should not have to. On the down side, this means that a given direction of tuning knob rotation produces an increase of frequency on some bands and a decrease on others.

There are many other possibilities. 6, 8, 9 and 10MHz all have attractions for the IF. 5MHz is not bad, but a bit low for full 500kHz coverage whereas 11MHz is getting a bit near the high limit for a stable VFO. You need to do the arithmetic and make your own choice.

I am assuming here a major simplification of calibrating by counting the VFO only. This means trimming the converter injection crystals to exact frequency and ignoring them thereafter for read-out purposes. In practice, this constrains crystals to exact or half megahertz.

Do not be tempted to consider mixing the VFO with the crystal oscillator to give the required input to a counter, thus apparently circumventing these constraints. It will inevitably find its way into the receiver front end leading to desensing, AM detection and AGC instability at the least!

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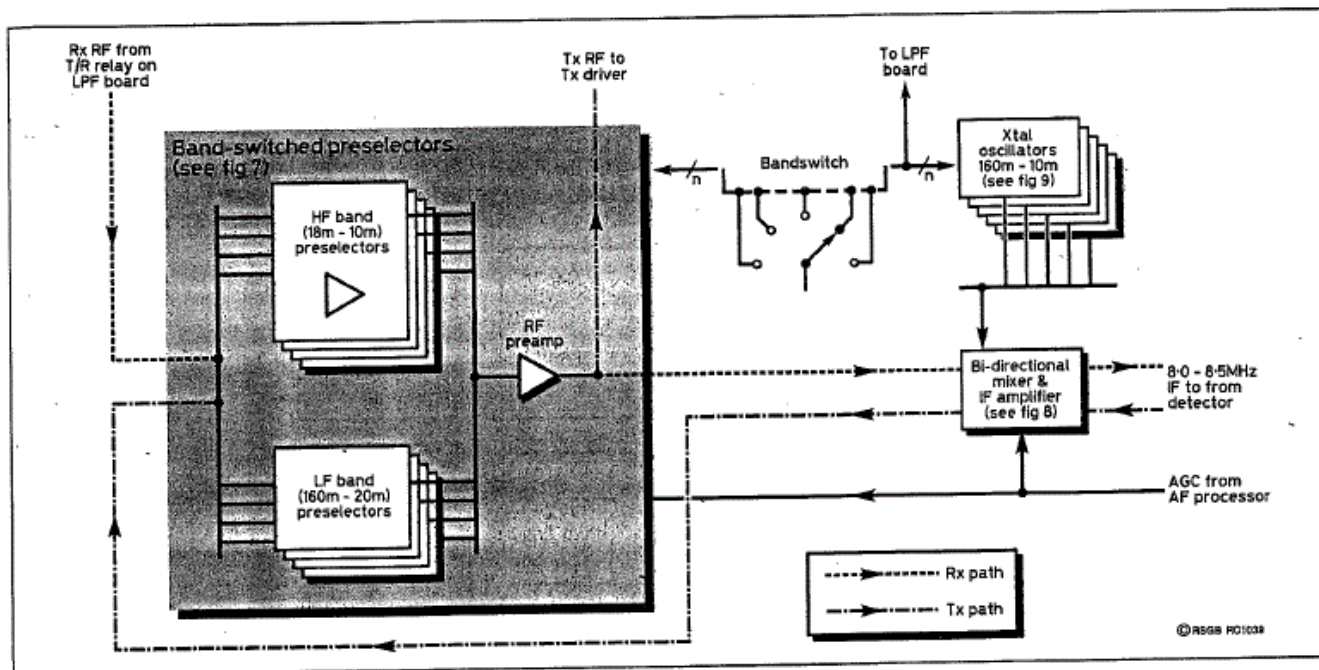


Fig 6: RF Board overview. The band switch is mounted on the board.

RF BOARD OVERVIEW

PRETTY WELL ALL THE circuitry on this board is used both on transmit and receive. Fig 6 outlines the basic concept. There is nothing about this board which is peculiar to the Third Method, but it does form part of the total system design.

On receive, the incoming signal from the low-pass filter (LPF) and the antenna c/o relay is routed to the appropriate preselector. This narrows the spectrum down to the amateur band of interest. From here it goes via an RF preamp to a mixer which translates it to the tuneable IF of 8.0 - 8.5MHz. Injection to the mixer comes from one of a set of band-switched crystal oscillators. The mixer is then followed by a 'strong' IF amplifier.

On transmit, the IF signal from the detector is amplified by the IF Amplifier, mixed to the chosen amateur band and routed to the appropriate preselector. It then leaves the board going to the transmitter driver and thence to the PA.

The preselectors come in two formats, one for the lower bands and one for the higher ones. All the lower bands (20m and below) use passive L/C filters, whereas each higher band (above 20m) has its own embedded amplifier. This is done to optimise gain and noise figure for each band. Note that these amplifiers are also used on transmit.

The design intent is that the receiver should never be 'deaf' - because we don't all have large antennas and there are quiet times on all bands - occasionally! This surplus gain does mean that you will have to adjust the RF gain control to suit prevailing operational circumstances.

The application of these formats for the different bands may be varied for extreme operational circumstances. For example, for mobile use on 20m, an 'HF' configuration may be preferable. Conversely, if you are operating a vast rhombic on 17m, the 'LF' configuration may be better.

Indeed, there is no objection to having both preselector formats available on different positions of the band switch.

One preselector is specified for coverage of the entire 10m band. If you are particularly interested in the top end of 10m or need it for transverter purposes, then you should consider including a further front end coil set.

Although this transceiver is aimed primarily at SSB operation, full details are provided for 9-band operation. The bandswitch is shown in Fig 6 with 'n' positions where the value of 'n' is your choice. The preselectors and crystal oscillators are modular in construction and it is a relatively trivial matter to remove or add further bands or 500kHz segments.

AGC is applied both to the preselector and to the IF amplifier. The same line is used for front panel control of transmitter drive - and for application of ALC from an external source.

PRESELECTORS

THIS PART OF THE RF board is shown in Fig 7. To illustrate the overall system, two HF and two LF filter sets are shown. In reality, the enclosed components are replicated as required to give the desired coverage. Component values not shown here and which are band dependant will be provided with the constructional detail next month. The basic design was freely adapted from the G3T50 transceiver [6].

Band selection is implemented by switching diodes which route the signal through the chosen filter set. Note that on both transmit and receive, RF passes through the filters from left to right.

The HF configuration is used on 17m and up. Each band has four poles of L/C filtering and each has its own embedded mosfet. It is unrealistic to attempt switching a single device between bands because of the high impedance levels on Gate1 and the Drain. A ferrite bead is used on Gate2 to suppress VHF parasitics. It is

essential that this be placed on the lead against the base of the mosfet and that all four leads be as short as possible.

Gain control on transmit and receive is applied to Gate2 of the mosfets. They do not object to having volts on G2 even when there is no voltage on the drain. This simplifies band switching considerably. TR2 is a simple inverting amplifier which provides a decreasing voltage with increasing signal at the correct levels for gain control of the mosfets.

The LF configuration is used on 160m through 20m. The L/C filtering is similar in principle to the HF configuration. Again, four poles of filtering give comprehensive protection against out of band signals.

All bands share the common broadband RF amp TR3. It is of conventional design providing a modest noise figure and modest gain.

Although the amplifier itself (TR4) is conventional, the circuit for applying gain control is less so. TR3 shunts the feedback resistors R20 and R21, progressively increasing the negative feedback around TR4 at increasing signal levels.

This has the effect of reducing the gain and increasing linearity for a strong signal. The disadvantage is that it lowers both the input and output impedance from the nominal design value of 50Ω - which in turn lowers the selectivity of the associated tuned circuits. No adverse effects have been noted in practice. On the contrary, with the J310 FET specified, some 40dB of AGC range is provided giving a net gain of 15dB with no AGC, down to some 25dB of attenuation at full AGC. RV1 is adjusted so that with no received signal, there is a barely perceptible reduction in gain. The circuit operates identically on transmit under ALC/Tx drive control.

The final gain control element is diode D7. This acts as a PIN diode attenuator on all bands on both transmit and receive. It is possible to purchase 'real' PIN diodes, but they are not cheap. None of the ones I tried offered as good

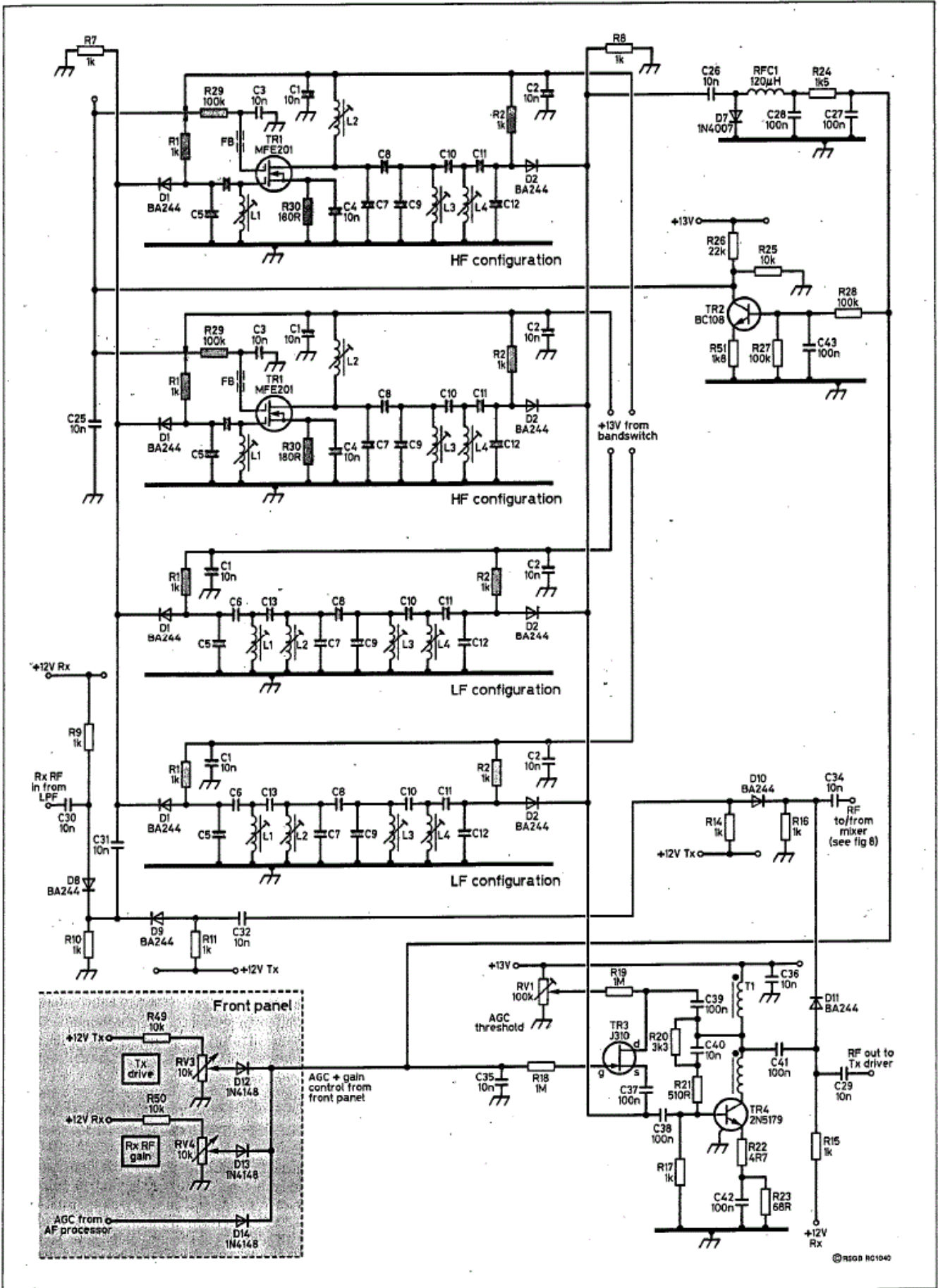


Fig 7: RF preselectors. The top two filter sets are representative of the HF configuration, the bottom two of the LF configuration.

performance as the 1N4007 in this application. This latter diode is sold, for a few pence, as a high-voltage rectifier.

Other more complex constant impedance attenuator circuits were tried out, but seemed to offer no practical advantage.

MIXER & IF AMPLIFIER

THE BI-DIRECTIONAL mixer and amplifier, which translates the selected 500kHz segment to/from the 8.0 - 8.5MHz tuneable IF, is shown in Fig 8.

A diode ring mixer, the SBL-1 was the device finally chosen. I tried both the Plessey SL640 and the Motorola MC1496 in this position as cheaper alternatives and an SRA-1H as a more expensive but higher performance diode ring alternative. The latter provided no great improvement and because it costs more and requires higher injection levels, it was discarded. The SL640 was acceptable but gave some cross-modulation effects under strong signal conditions.

The MC1496, although much cheaper, requires many more associated components and proved both excessively noisy and susceptible to IMD. This latter conclusion may be worth further investigation, because G3TDZ [4 - see last month] used the MC1496 in similar circumstances and [7] also found it unsatisfactory even following a crystal filter. In any event, the SBL-1 has proved the best solution, not least because a simple crystal oscillator provides sufficient drive.

TR5 on receive and TR6 on transmit provide the IF gain and offer broad-band termination to the mixer. The termination is further improved by a 6dB pad, R42 - R44 which isolates the amplifier and mixer from any reflections back from the bandpass filter. The temptation to realise an extra 6dB by omitting this pad should

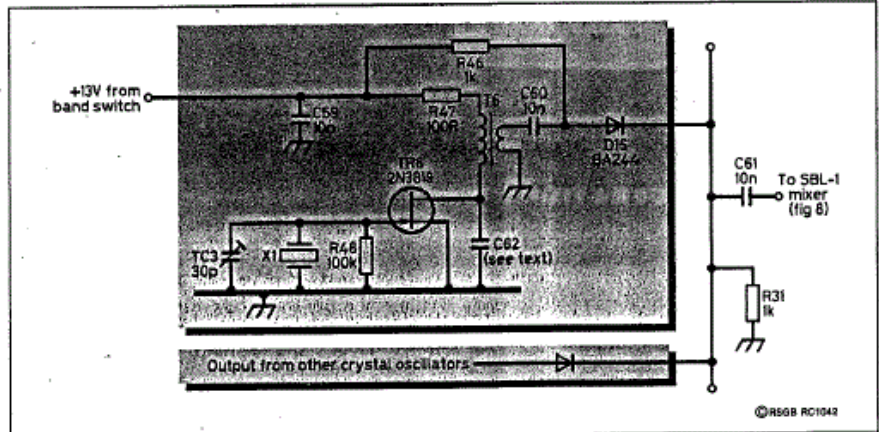


Fig 9: Crystal oscillator. T6 and X1 values will be provided with the construction detail next month.

be resisted since it is important to strong signal performance.

The bandpass filter itself is formed by T4/T5 and TC1/TC2, top-coupled by C58. It is stagger tuned to cover the IF. Some experimentation with the value of C58 may be useful to achieve a flat response.

Two forms of gain control are provided. D15 acts as a PIN diode attenuator and TR7 provides variable negative feedback around the IF amplifier on both transmit and receive. Listening to a quiet band, RV2 should be set to mid-range and then taken down to the earthy end until there is no perceptible increase in receiver output.

CRYSTAL OSCILLATORS

THE CIRCUIT FOR one oscillator is shown in Fig 9. This needs to be replicated for each band - except that bands which are image pairs use the same oscillator. 160m and 17m form such a pair with the chosen IF.

C62 tunes the primary of T6 to resonance. In practice, a trimmer was first used, measured and then replaced with a small ceramic capacitor.

TC3 is used to trim the oscillator frequency to the exact crystal frequency in order to adopt the expedient of ignoring the oscillator frequency for calibration purposes.

REFERENCES

- [6] 'Modular HF Transceiver' by Mike Grierson, G3TSO, *Radio Communication*, Oct, Nov 1988
- [7] *Solid State Design for The Radio Amateur*, by Wes Hayward, W7ZOI and Doug DeMaw, W1FB. ARRL.

In next month's issue, I will cover construction practices in general and the detail for this RF board in particular.

... to be continued

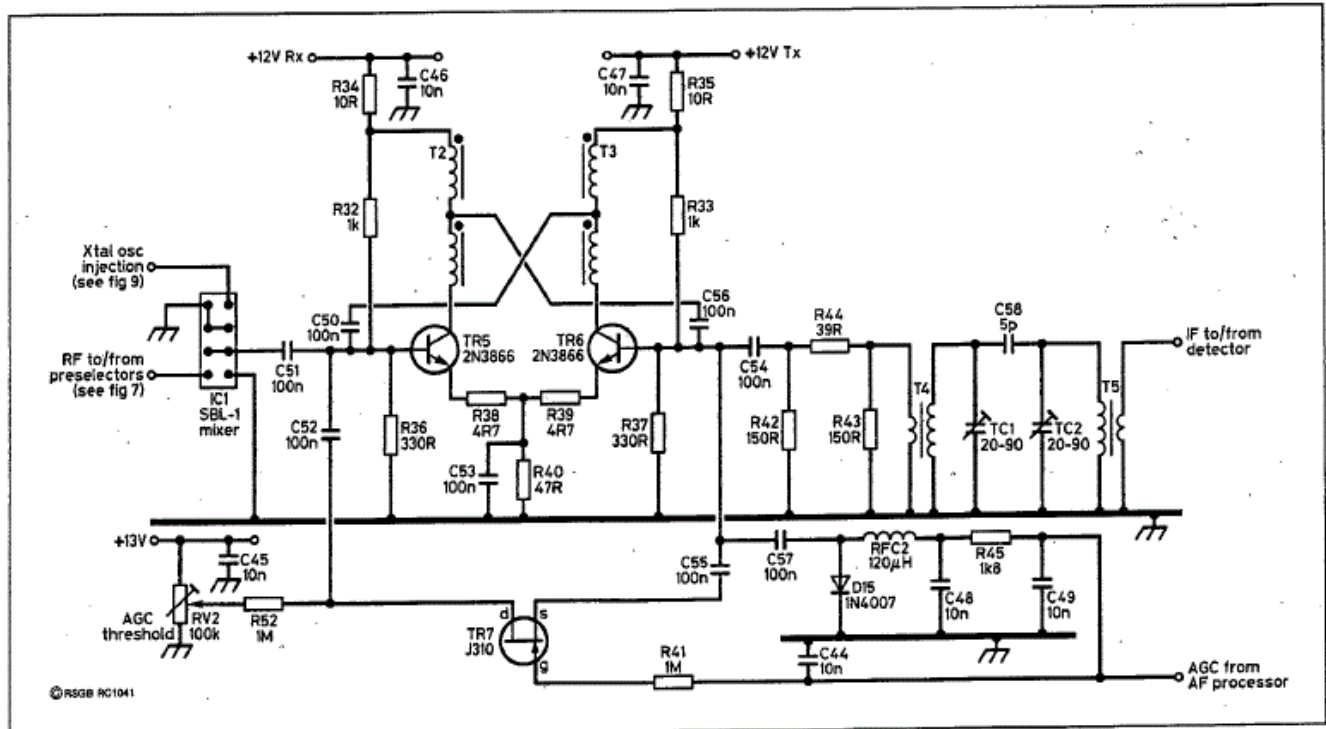
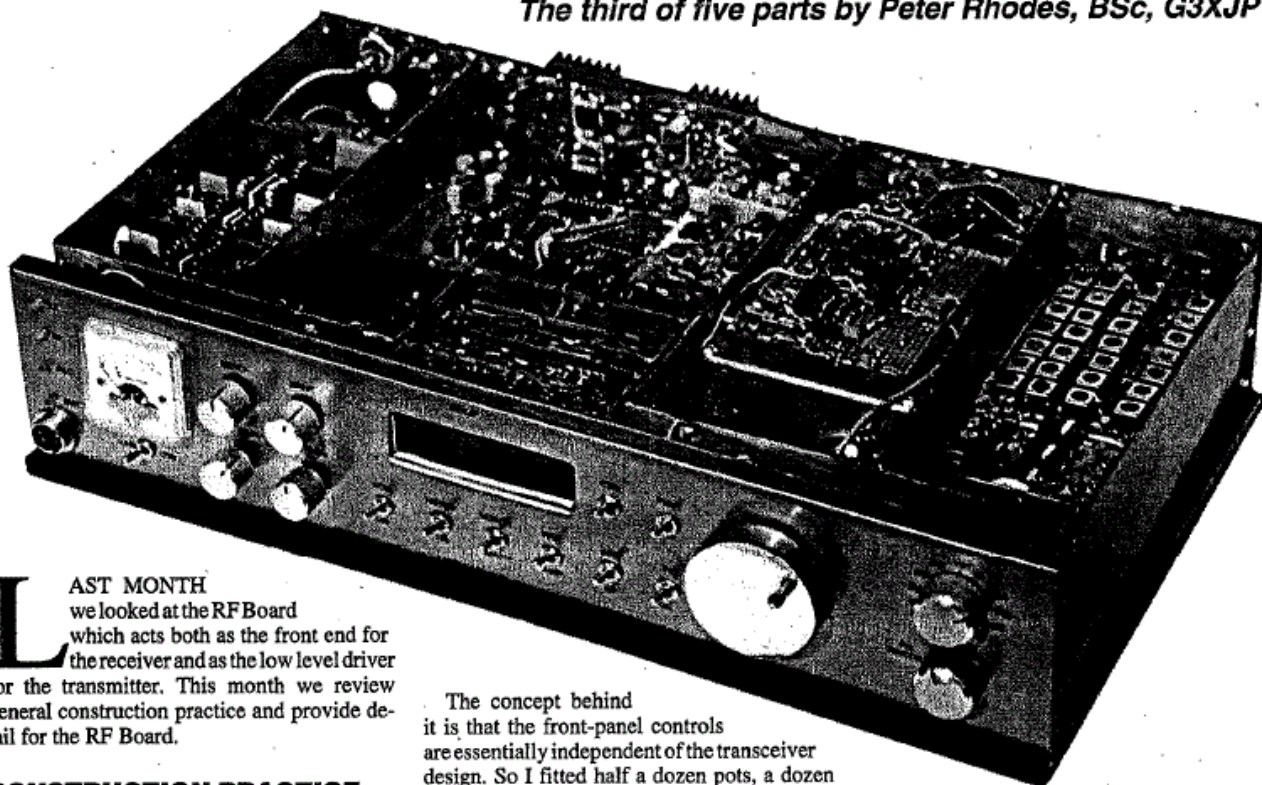


Fig 8: Bi-directional mixer and IF amplifier with gain control.

**RADCOM
TECHNICAL
FEATURE**

Third-Method SSB HF Transceiver

*The third of five parts by Peter Rhodes, BSc, G3XJP**



LAST MONTH we looked at the RF Board which acts both as the front end for the receiver and as the low level driver for the transmitter. This month we review general construction practice and provide detail for the RF Board.

CONSTRUCTION PRACTICE

BEFORE GETTING DOWN to specific board layouts, there are some general approaches to mechanical construction adopted throughout this transceiver which you may want to consider.

This is one area where I believe it is not always appropriate to emulate the professionals. There are techniques available to the amateur which are easier, quicker and cheaper than those used in professional construction - given exactly that our goal is to make a one-off. This is the critical distinction.

It is worth emphasising that these construction practices and the Third Method are both somewhat unconventional, but not related. You may choose to adopt either one without the other.

TRANSCIVER HOUSING

THIS HAS BEEN AROUND for about 20 years or so. I built it originally as a development platform and it has outlived about six generations of transceiver. I would commend it equally as a permanent approach but it is certainly not mandatory.

The concept behind it is that the front-panel controls are essentially independent of the transceiver design. So I fitted half a dozen pots, a dozen miniature toggle switches, a bandswitch, an 'S' meter, mic and 'phone jacks, a 6-digit counter and a tuning knob on a slow-motion drive - on the grounds that any transceiver I can conceive will require this user interface. They are generously spaced, since I am keen to avoid the modern trend of cramped front panels and space is not at a premium. The only thing that changes from one transceiver to the next - and that, rarely - is the lettering on the front panel or the value of a pot.

I also have a self-imposed rule that no signal wires are ever routed to the front panel for switching - only remote DC-controlled switching is allowed. Not only does this place the signal switch where it is wanted, but it means that the front panel can be ergonomically optimised with no electronic constraints. In particular, the band switch is a single pole multi-way switch and all band switching is done remotely. I am not keen on ganged wafer switches with long shafts.

There are three main and two small circuit boards as well as the VFO and PA to be housed. I also have a built-in frequency counter which is outside the scope of this project.

Construction of the housing is also a bit unconventional. PCB sheet is cut and soldered to make an open-top box chassis. The same

material is used to form internal partitions. This chassis is screwed down to a base of 5-ply wood to increase mechanical strength. This is immaculately finished in matt black where it shows on the front lip.

Fig 10 shows the general layout.

The lid is made from the same material as the base. It comprises two sides and a top, lined internally with kitchen foil using impact adhesive. The lid is connected electrically and mechanically to the box chassis using a piano hinge with a removable hinge pin. There are many alternatives.

The aspect ratio of the enclosure is such that it looks somewhat like a piece of hi-fi equipment or a video recorder in general shape. The front panel is 'long and low' and inside, there is enough height for one layer of electronics as opposed to the two layers - one each side of a chassis - frequently found. This general shape does not give the smallest possible volume but it does maximise flexibility and top access. I would only choose not to adopt it if I were either building a mobile rig or if I had an operating desk that was critically dimensioned to a few inches. On the contrary, when building home-brew designs I would always add 20% to the dimensions first con-

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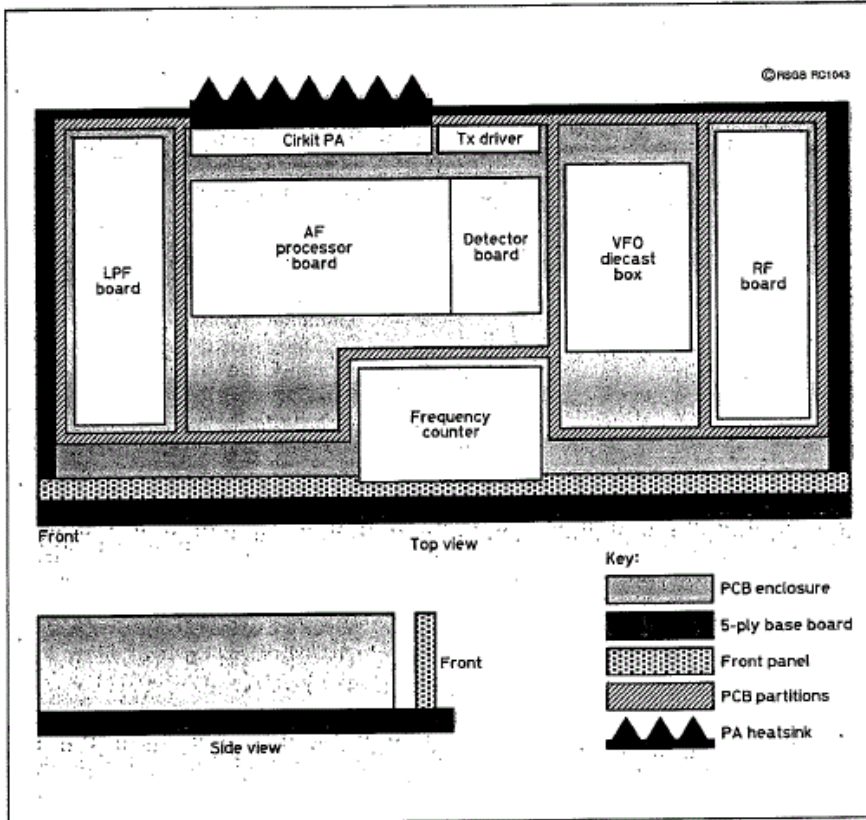


Fig 10: Overall concept of G3XJP transceiver enclosure, not to scale. The PA, mounted on its heatsink is let into the rear wall.

templated, if only to allow for the inclusion of accessories etc later. Nothing is worse than being space constrained. If the end result is a few empty compartments, so what? We don't face the same critical tests as the professionals in this respect.

Note that you have the opportunity to place the VFO tuning knob where you want it ergonomically, depending on whether you are right or left handed. The professionals tend to put it in the middle where it is not quite right for anyone!

The front of the PCB box forms a false front panel which is used to mount all potentiometers. About 2cm in front of this false front is the visible front panel. It is made from soft wood about 1cm thick, screwed and glued to the base board. It carries the 'S' meter, connectors for the microphone and headphones, LEDs, and all the toggle switches. The latter are countersunk into the thickness of the front panel until they just protrude by the correct amount.

The height of the front panel is determined by the 'S' meter, tuning knob or the PA heatsink at the rear wall, whichever is the taller. Mine is about 8 cm excluding the base board.

The minimum gap between the two front panels will probably be determined by the depth of your 'S' meter. The rotary shafts (pots and band switch) pass through the gap and protrude beyond the front panel where the knobs are attached. This gap also serves as a cable trough where unsightly interconnecting wires can be tucked out of the way. These wires pass through the false front panel on solder-in feedthroughs and are liberally sprinkled with

ferrite beads in order to retain the RF-tight enclosure. I tend to overdo these things, but better safe than sorry!

A three-sided enclosure is let into the false front panel. This allows my frequency counter module, otherwise known as a broad-band noise generator, to be retained on the front panel without compromising the screening integrity of the transceiver space. Its 7-segment display chips are mounted on high-profile DIL sockets (or sockets mounted on sockets if you prefer) so that they protrude from the front of the counter. They are then let into the thickness of the front panel so that the display ends up just short of flush with the front of the panel.

With the exception of the PA and its driver, all circuit boards are mounted horizontally, component side up for ease of access. All the inter-board wiring is underneath and out of sight. Because you can't get in through the

bottom of the box, the boards are mounted on 'hinges' made from short lengths of braid, soldered between the board and a vertical partition wall. A couple of solder bridges from the board to the chassis complete the mechanical mounting. These are crude but effective and can be broken with a touch of the soldering iron when necessary. I guess you could solder in some mounting gussets and use self-tapping screws if you were so inclined.

The idea behind this mechanical approach is that copper-clad sheet and wood are easier materials to obtain and to work with than say, aluminium. In particular, I have never managed to acquire the skill of bending the latter properly. PCB provides instant corners, instant brackets, instant shielding and instant earthing wherever you require it. Cable ties become a short length of wire soldered at one end to the chassis and then bent over the cables. Solder-in feedthrough capacitors become realistic. There are many benefits.

The first note of caution is that you must use fibreglass and not SRBP for making boxes. The latter distorts alarmingly under heat.

With minimal effort and little skill, the finish on wood can be made immaculate. I find spray putty as sold by car accessory shops best for filling in surface imperfections - either natural or man-made. A few coats of paint, some lettering and then a few coats of clear varnish complete the job.

The second note of caution is never to use both acrylic and cellulose paints and fillers. Either one or the other!

The final note of caution concerns board placement. The two critical considerations are: firstly the audio leads joining the Detector board to the AF Processor must be as short as possible - less than 2"; and secondly the VFO output lead to the Detector board must be kept very short and fully screened. In my case, the coax passes through a hole in the diecast box and the adjacent vertical partition, drilled exactly at the point where the VFO feed is needed on the Detector board.

CIRCUIT BOARD CONSTRUCTION

I AVOID ETCHING circuit board whenever possible. In this transceiver the only conventional PCB is the VFO - and this for mechanical stability.

The others are made using what can be classified as a 'semi-ugly' approach. The basic

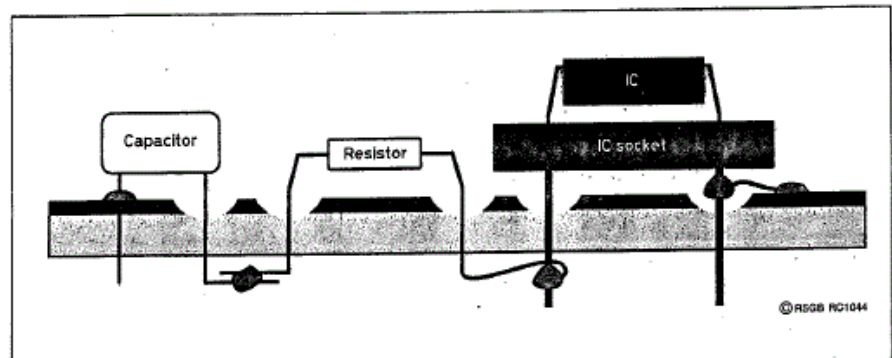


Fig 11: 'Semi-ugly' construction. Single sided board is used with the copper on top. Most components are mounted on the copper side as illustrated.

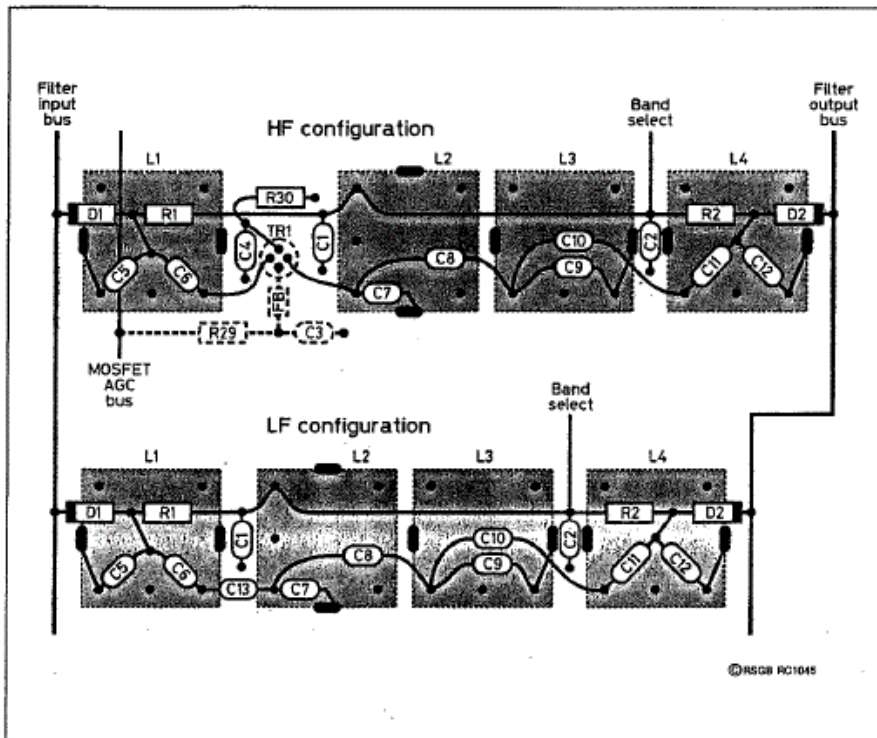


Fig 12: HF and LF configuration component layout, viewed from beneath (non-copper side). Components on top (copper side) are shown dotted.

material is single-sided copper clad board. I use SRBP to save on drill bits but fibreglass is marginally superior. The copper forms a continuous ground-plane with the components mounted on the copper side. Fig 11 shows the general idea.

The board also acts as the top of the screened compartment formed by the chassis base, vertical partition walls and the board itself.

Earthed component leads are soldered directly to the ground-plane, but I usually drill the hole and pass the lead though the board for the sake of appearance. Other leads are passed through a counter-sunk hole and then bent over and soldered to other component leads to form junctions on the back of the board. Thus the component leads themselves are used to form impromptu PC track. The approach is not dissimilar to valve construction, except that an IC socket is used for mechanical mounting rather than a valve base. On rare occasions this does not provide enough mechanical strength in which case a Vero pin is inserted in a counter-sunk hole to provide a fixing point.

All other point-to-point wiring is done using either subminiature coax or thin self-fluxing enamelled wire - depending on the nature of the signal and the length of the run. It goes in on either side of the board - with a preference for keeping it out of sight on the non-component side.

All DIL ICs are mounted on sockets. On this transceiver this is near mandatory because a significant proportion of CMOS is used. This, incidentally, pretty well rules out totally ugly or dead-bug construction.

The turned-pin style of socket is preferred over the cheaper variety since it provides firm anchor points for adjacent components. Before the socket is inserted, a small loop of wire is wrapped and then soldered to any grounded pin - and the adjacent board is tinned. The socket is then inserted fully and the earth connections soldered immediately adjacent to the socket. Don't try bending the socket pins instead. It doesn't work.

I prefer to use a trace of Super Glue to bond the socket to the board, but you have to be very

careful to avoid getting any in the sockets themselves. It travels far and fast by capillary action and if it gets into the pin insertion holes, the socket is a write-off. I use a 1mm drill as the default and with normal tolerances on the exact position of the holes, the sockets usually end up as a good push fit.

My general preference is not to use connectors for inter-board wiring. I tend to solder leads directly to Vero pins and occasionally regret my meanness.

Traditionalists will balk at most of the above, but the fact remains that it is cheaper, much faster, more flexible and more environmentally friendly than conventional etched PC construction.

Because in critical places you can achieve higher component density and shorter lead lengths, the end result is always at least as good electronically. The visible side of the board looks like the 'real thing'. The underside, let's face it, looks a bit of a mess - especially when some of the components get tacked on the back. That is the only price to be paid, always given that you are trying to build a one-off and have no aspirations either to go into volume production or to win construction beauty contests.

You may still prefer to etch the boards, in which case you must use double-sided PCB with an unetched ground plane. You will probably need to increase the size of the RF board and you will have to handle a large number of crossing tracks on the AF Processor board.

The choice is yours, but be aware that in this project, PCB artwork will be published only for the VFO board. The remainder will be provided as drilling templates and component location diagrams.

RF BOARD CONSTRUCTION

THE BOARD IS MADE from single sided SRBP with the majority of the components mounted on the top (copper) side.

Fig 12 shows the detailed under-board component detail for the tuned circuits associated with the Toko coils. This is distinctly 'ugly' construction. In practice, the leads should be as short as possible; somewhat shorter than as illustrated. Note that G2 of the mosfet TR1 is bent out on top of the board to minimise its lead length.

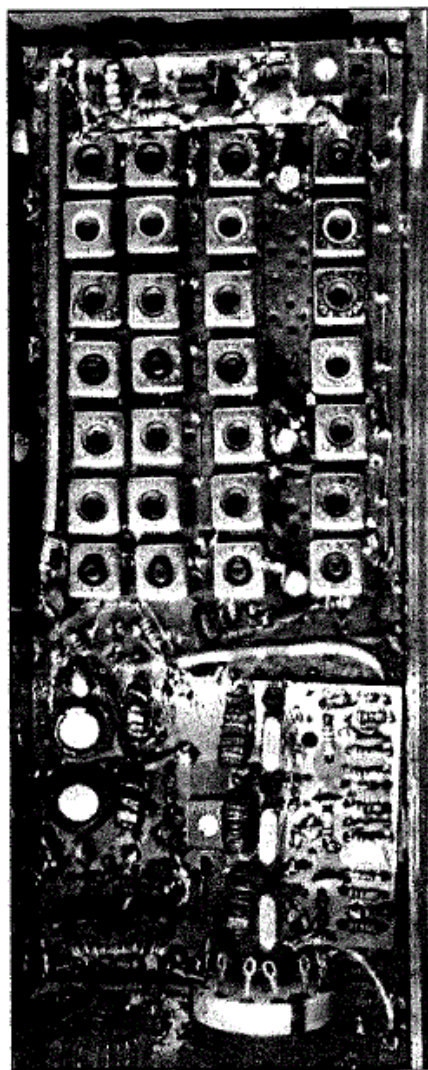
The shields for the coils are soldered to the board on the copper side and the tags on the underside are used as earth points. Note that the unused pins on some of the coils carry other windings, so they must be left clear. In particular, the holes are counter-sunk to prevent accidental earthing.

Fig 13 shows a suggested layout for a fully populated board. This diagram is intended for use as a drilling template. If you Xerox it first, make sure the dimensions (7.8in x 2.75in) still measure true on the copy. Note that the Toko coils are drilled on a metric grid, everything else being imperial. The best approach is to drill all the holes first and then lightly counter-sink the non-earthly ones on the copper side. This prevents any problems with swarf which can arise if you drill as you go.

You will need to consider access, particularly around L1 during assembly.

	L1, L2, L3, L4 Toko coils	C5 C12	C6 C11	C7 C9	C8 C10 C13	X1 MHz	T6
160m	BTKANS-9440HM, 8.2µH	2700	1800	750	180	See 17m	
80m	KANK3334R, 5.5µH	1000	390	270	33	12	30/4t
40m	KXNSK4173AO, 3.0µH	1000	220	150	10	15	20/4t
30m	KANS12354BM2, 2.5µH	470	120	100	5	18	16/4t
20m	KANK3335R, 1.2µH	560	120	100	5	6	45/5t
17m	KANK3335R, 1.2µH	390	100	82	3	10	34/4t
15m	BTKXNS-9453HM, 0.68µH	470	100	82	4	13	28/4t
12m	BTKXNS-9454HM, 0.56µH	470	82	75	5	17	18/4t
10m	BTKXNS-9455HM, 0.47µH	330	82	56	3	20	15/3t

Table 1: Component values for nine bands using an IF of 7.8 - 8.5MHz. All capacitor values are in pF and are sub-miniature ceramic plate types. 160m and 17m share a common oscillator. T6 is primary/secondary turns of 30SWG (0.315mm) on a T50-2 toroid. C13 is used on the HF configuration only.



The completed RF board with the components in place.

The main runs of actual wire are illustrated. All DC lines were run in using thin self-fluxing wire on the back of the board. RF lines of all but trivial length use subminiature coax. All other connections (the vast majority) are made using the component leads themselves, bent over and soldered together.

The board runs from front to back of the transceiver and so determines the overall depth. If you need less depth, you could split the board between HF and LF and mount the two pieces back to back. You would need to make provision for access to tune the coils.

Unless you are sure of your future requirements I would in any case advise cutting and drilling the RF board for a full complement of nine bands even if you do not populate the board at first. The incremental cost for each band is roughly four Toko coils (see Table 1) and a crystal oscillator, so there is no overwhelming cost constraint.

I have a personal aversion to plug-in coils, but they undeniably offer a cheaper solution at the expense of operating convenience.

The RF preamp is build on a small daughter board and mounted adjacent to the L4 bank for the LF bands. This implies that you leave space for at least four low frequency bands. An

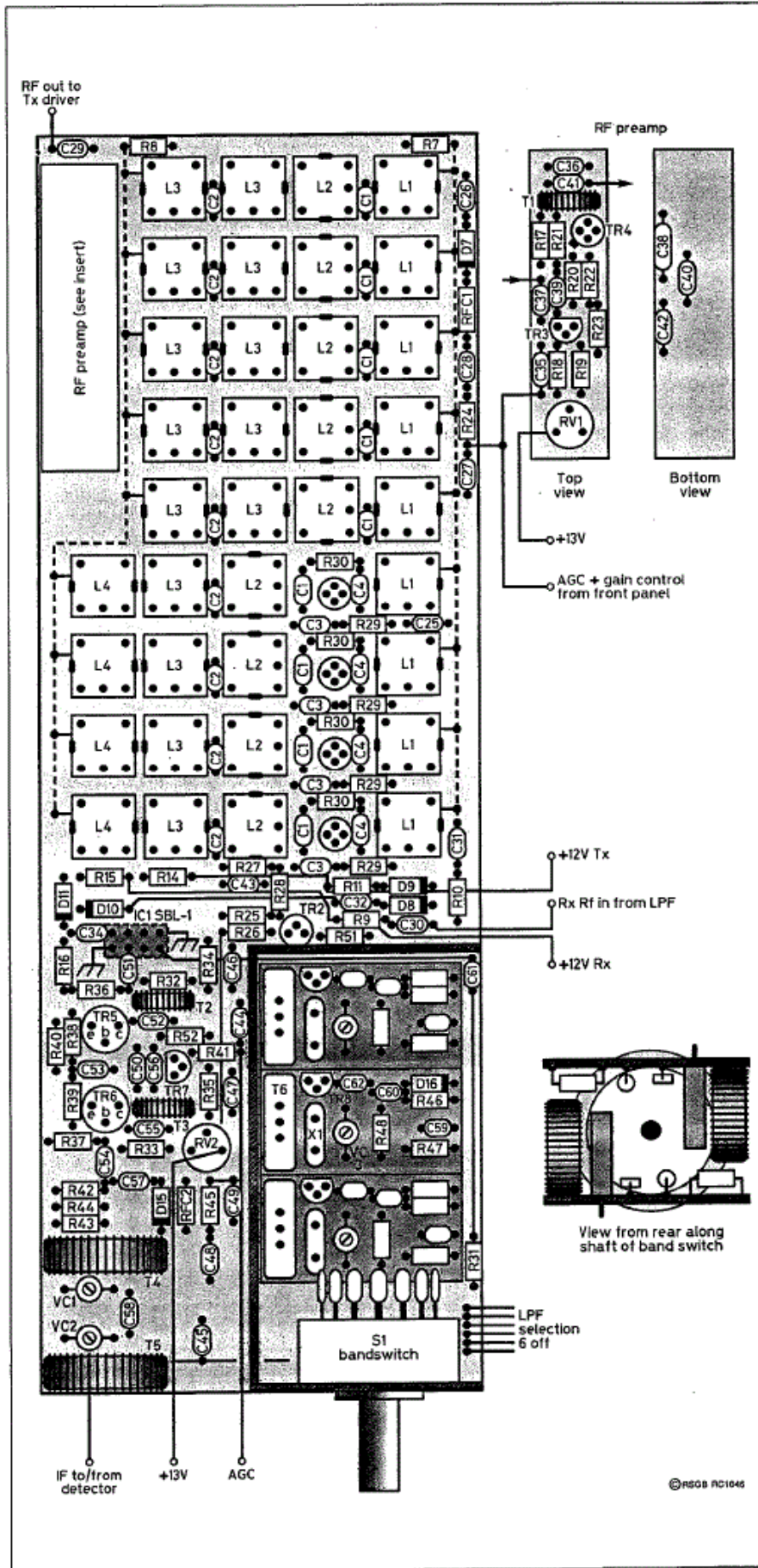


Fig 13: RF board component layout, viewed from top (copper side). The SBL-1 mixer is mounted under the board. See photograph above left for alternative layout.

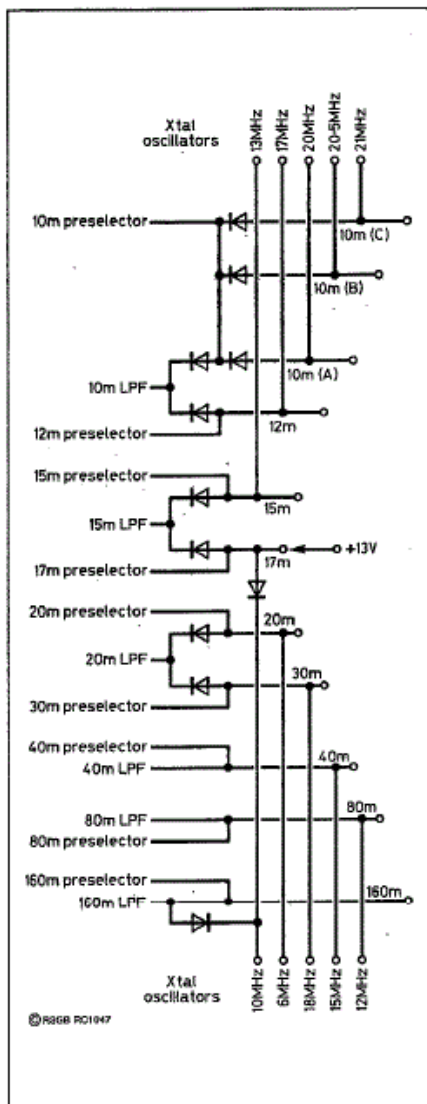


Fig 14: Band switch detail. All diodes are 1N4148 or equivalent and mounted on a small matrix board (not shown) next to the band switch.

alternative configuration would be to omit at least one band, turn the preamp through a right-angle and mount it in the space vacated by the four coils.

The gap between the L1 and L2 banks in the HF configuration is occupied by the separate mosfet amplifiers.

The two RF busbars, namely filter in and out, are made from stiff bare wire mounted below the board on Vero pins. They pick up the band selection steering diodes.

For nine bands, you will need eight crystal oscillators plus one more for each extra segment on 10m. These are built on child boards in groups of three.

Fig 13 shows one child mounted adjacent to the band switch and the inset shows how a further child is flipped over and mounted above it. Note that the crystal and transformer T6 are judiciously located to minimise the height requirements.

The top child is hinged on the short edge furthest from the band switch to allow access.

The 'hinges' are short lengths of braid soldered to the child and to the vertical screen. For up to a further six oscillators (twelve maximum

in total), this child pair arrangement is repeated below the mother board.

A screen is constructed from PC material around each child pair to keep any oscillator radiation out of the IF amplifier. The screen below the board is mounted on a few Vero pins (not shown) passed through the mother board.

If this whole mechanical arrangement for the oscillators does not appeal to you, they could be built in a quite separate enclosure in a more traditional two-dimensional fashion.

You may also be wondering why I did not use one oscillator with switched crystals. The basic reasoning was that such an untuned oscillator would not produce sufficient output and would therefore need a broad band driver - and perhaps some band-switched L/C filtering as well.

BAND SWITCH

THE BAND SWITCH is mounted on the vertical screen on the board. This saves trailing wires between the board and the front panel during commissioning when the LPF is not needed. The switching arrangements are shown in Fig 14.

The band select lines themselves are made by lightly twisting together nine strands of thin enamelled wire and Aralditing them to the underside of the board at intervals. This nine-core 'cable' has tapered down to single core by the time it gets to the far end of the board.

The same compact and flexible construction is used for the six-core cable run to the LPF board. In this case, the cores were screened by some coaxial braid.

RF BOARD SUMMARY

THIS IS A PRETTY comprehensive and rigorous RF board for any transceiver and represents by far the largest construction effort in this project. In order to get going quickly, my inclination would be to populate the board for one LF band only until such time as the whole transceiver is functioning correctly. The other approach would be temporarily to tap RF - or perhaps even broad band IF - from an existing (home-brew) transceiver. The G3T50 arrangement is particularly amenable, at least on receive.

● In next month's issue, we will cover the mixing of the tuneable IF to/from audio frequencies; and the removal of the unwanted spectral components, employing Third Method technology. ♦

... to be continued

COMPONENTS FOR RF BOARD

Resistors

- R1, R2, R7 - R11,
- R14 - R17,
- R31 - R33, R46 1k
- R18, R19, R41, R52 1M
- R20 3k3
- R21 510
- R22, R38, R39 4R7
- R23 68
- R24 1k5
- R25 10k
- R26 22k
- R27 - R29, R48 100k
- R30 180
- R34, R35 10
- R36, R37 330
- R40 47
- R42, R43 150
- R44 39
- R45, R51 1k8
- R47 100
- R49, R50 10k on front panel
- RV1, RV2 100k horiz preset
- RV3, RV4 10k lin on front panel

Capacitors

- C1 - C4, C25, C26, C29 - C32, C34 - C36
- C40, C44 - C49, C59 - C61 10n
- C5 - C13 see Table 1
- C27, C28, C37 - C39,
- C41 - C43, C50 - C57 100n
- C58 5pF see text, select on test
- C62 see text, select on test
- VC1, VC2 20 - 90 pf
- VC3 30 pf

Inductors

- T1 - T3 10 bifilar turns of 30SWG (0.315mm) on FT37 - 43
- T4, T5 35t/6t of 26SWG (0.45mm) on T68-2
- T6 see Table 1
- L1-L4 see Table 1
- RFC1, RFC2 120µH
- FB small ferrite bead, FX1115

Semiconductors

- D1, D2, D8, D9 - D11, D16 BA244
- D7, D15 1N4007
- D12 - D14 1N4148 on front panel
- IC1 SBL-1 diode ring mixer
- TR1 MFE 201 or equiv
- TR2 BC108 or equiv
- TR3, TR7 J310
- TR4 2N5179 with small heatsink
- TR5, TR6 2N3866 with small heatsinks
- TR8 2N3819 or equiv

Additional Items

- X1 see Table 1
- SW1 1pole 12way rotary bandswitch

Third-Method SSB HF Transceiver

The fourth of five parts by Peter Rhodes, BSc, G3XJP

THIS MONTH the heart of the Third Method transceiver is described, showing how on receive, the incoming IF is confined to a controllable audio bandwidth with one sideband eliminated. Conversely, how microphone input is translated to SSB at the IF.

First, a simple VFO is covered which produces a stable output across the tuneable IF range of 7.8 - 8.5 MHz. The output of this VFO is used to translate between audio and the IF.

VFO

THE VFO IS ADAPTED from one used in my previous G3TSO transceiver. Fig 15 shows the circuit diagram and Fig 16 illustrates the mechanical detail.

A single-sided printed circuit board is used for mechanical stability. The VFO assembly is mounted in a die-cast box which also carries a 30:1 reduction drive. A further 30:1 drive is mounted on the front panel, the two being connected by a flexible coupler.

The VFO coil, L1 comprises 17 turns of 20SWG (0.9mm) wire, close wound on some polypropylene tube of about 3/4in diameter. This is sold in plumbing shops as 'overflow pipe'. It is a tight push fit between two opposite walls of the die-cast box. To achieve maximum stability, no tuning slug is used and there is also no trimmer capacitor. This approach requires you to wind the coil with a few extra turns and then remove them one at a time until the desired coverage is achieved. Some adjustment to C10 may also be needed to optimise the band spread. Both adding the lid to the box and screwing it down tightly will significantly alter the VFO frequency.

The VFO output is fed by a short screened lead to the Detector board where the power is raised to a suitable level. The power driver was not included in the VFO housing because it does generate a fair amount of heat.

No provision has been made for IRT, but if you require this facility, you could add the usual varicap arrangement.

DETECTOR BOARD

THE CIRCUIT DIAGRAM is shown in Fig 17, and Fig 18 illustrates the tightly packed mechanical arrangement.

On receive, the IF spectrum from the RF board is split into two quadrature signals by T5

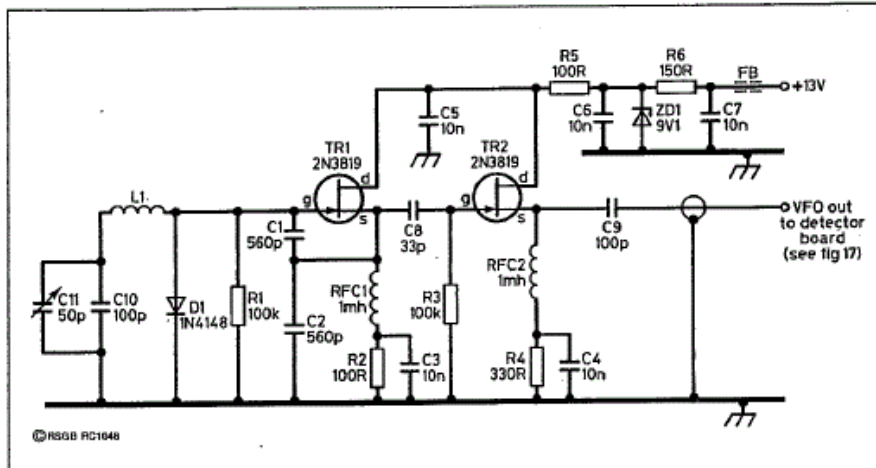


Fig 15: VFO tuning 7.8 to 8.5MHz.

and C12, C13. Provision is made separately to adjust the phase balance on receive (RV1) and transmit (RV2). This switched arrangement has turned out to be somewhat over-cautious, but the few extra components add some insurance against differential phase shift between transmit and receive. R1 is fitted on test. If the setting for both RV1 and RV2 turns out to be near zero resistance, R1 is added to give finer tuning control.

Each quadrature signal is fed to a conventional diode ring double balanced mixer. The VFO injection, amplified by TR1 and TR2 is fed in-phase to the two mixers, resulting in the two quadrature audio channels.

On transmit, the process is reversed. T5 and C12, C13 provide the required mutual phase shift of 90° and combine the two channels to give SSB output.

The VFO output is fed in phase to the two mixers but in series rather than in parallel. This is simply because it is much easier to get the requisite drive into a 100 ohm load than into 25 ohms.

It is traditional to place the RF phase shift network in the injection line from the VFO; and feed the IF signal in phase to the two mixers. The opposite approach has been adopted here for expediency. If the phase shift network were in the VFO feed, it would need to be at a low power level in order to avoid phase drift with temperature. So, it would need to go in the VFO output line

and be followed by two power drivers, one for each mixer. Some arrangement for combining/splitting the RF signal would still be needed.

I adopted the approach of including the 90° phase shift in the combiner/splitter and saving a power driver. The theoretical disadvantage of this approach is that it may lead to more amplitude imbalance between the two channels. The effects have not been noted in practice.

T1 - T4 are commercial Toko balun coils built around twin hole cores. T2 and T4 are modified by winding 5 turns of 40SWG (0.125mm) enamelled wire through the twin holes and over the existing winding. This extra

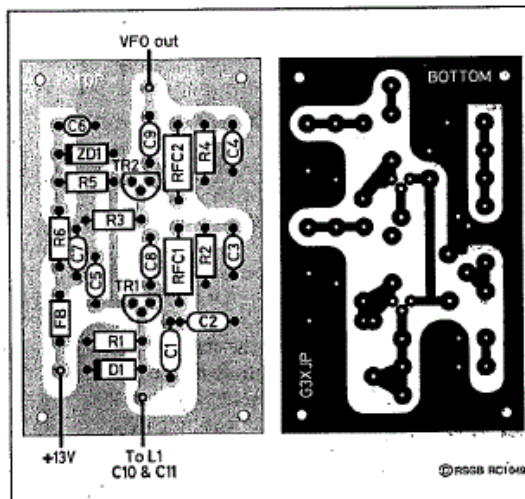


Fig 16: VFO PCB and component layout.

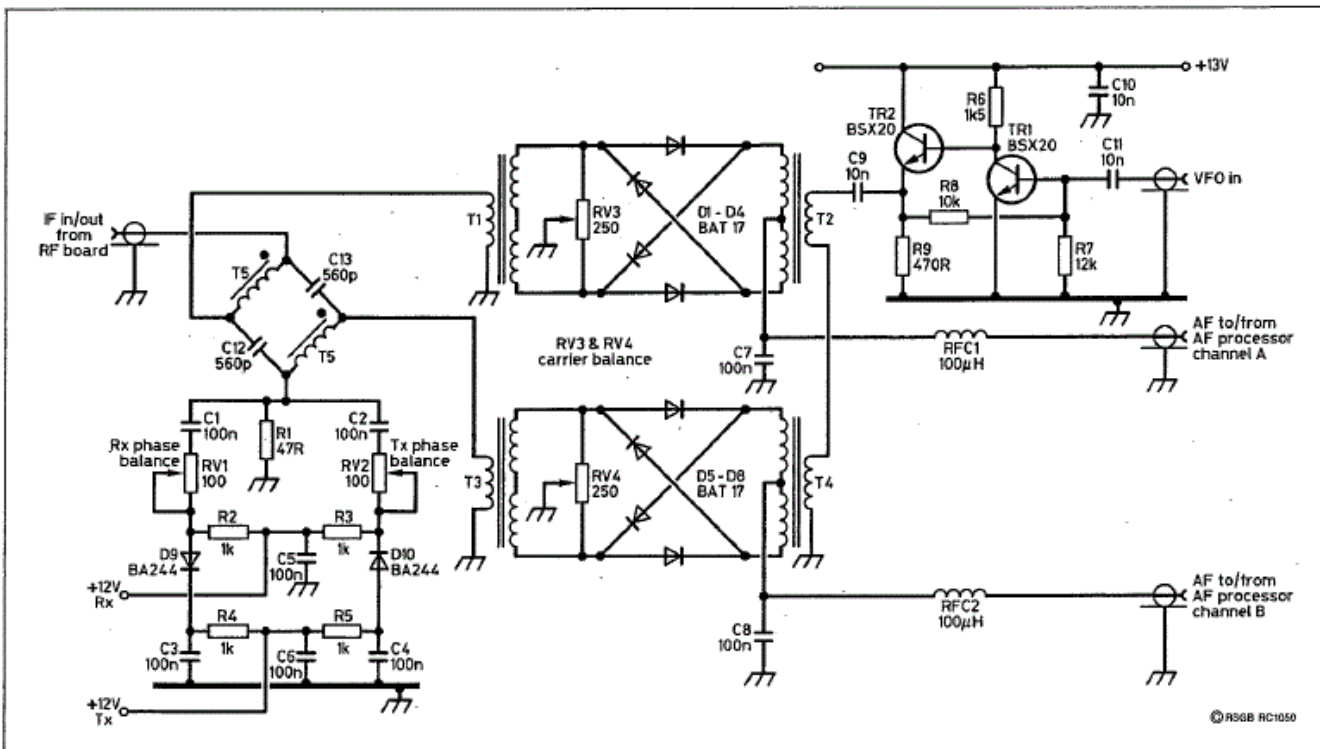


Fig 17: Detector board.

winding is used for the VFO feed. The existing winding terminals (the 2-pin primary as opposed to the 3-pin centre-tapped secondary) are unused and left floating. T1 and T3 are unmodified and the centre tap is not used. You could substitute conventional trifilar ferrite transformers, but the Toko baluns specified are not expensive, are small and offer good intrinsic balance.

The diodes used in the rings are cheap hot-carrier diodes. They all came on the same bandoleer, but no special efforts were made to match them.

Note that it is not possible to substitute commercial ring mixers because these make no provision for balancing out the carrier.

The construction sequence is critical in the area around the diode rings. Having drilled the board and countersunk the non-earthly holes, mount C7 and C8 on the underside. Then insert four Vero pins at the corners of each diode lattice. Then mount T4/T2 soldering the centre pin to C7/C8 and the outer pins to the Vero pins. Next solder a short vertical length of stiff wire to the ground plane. This will subsequently pick up the wiper of RV3/RV4.

Then attach the four diodes in each ring to the Vero pins. The diodes are mounted in a lattice configuration, as drawn in the circuit diagram. Note that the diodes all connect anode to cathode with the two 'diagonal' diodes passing either side of the vertical earth wire. Then add T3/T1.

Finally, mount RV3/RV4 over the top of the diodes, with the two ends of the track to the Vero pins and the wiper to the vertical earth wire. Fig 18 shows D1-D4 before RV3 has been mounted and D5-D8 with RV4 in place.

The remaining construction is not critical, except that RV1 and RV2 are added last. RFC1/RFC2 connect to the hot end of C7/C8 by soldering the shortest possible lead on the

choke to the end cap of the capacitor. Nasty, but effective!

The driver transistors TR1 and TR2 are fully enclosed in a box made from double-sided PC board.

It is essential to prevent any VFO leakage to the rest of the transceiver. A small strip of PCB also separates the two mixers. It would be good practice and not difficult to screen the whole assembly, but I haven't found it necessary.

AF PROCESSOR BOARD

THE MAIN CIRCUIT DIAGRAM is shown in Fig 19, with the additional power connections to the ICs shown in Fig 20 and the complete component layout in Fig 21.

The receive path from the Detector board - following Channel 'A' in Fig 19 for illustration

- starts with a low noise discrete transistor amplifier TR1. C3 and C11 provide significant high-frequency roll-off. IC1a provides just sufficient gain to overcome device noise in the MAX293 low-pass filter IC3.

The CMOS switches, IC2 and IC5 provide T/R switching. The cut-off frequency of the LPF, IC3 is determined by the clock oscillator IC10. This is controlled in turn by the front panel control RV3 (and on transmit, by the preset RV2).

R9 and C16 remove any residual clock noise from the signal before the audio amplifier IC6a and the unity gain inverter IC7a. These two provide 180° phase-splitting action and drive the product detector IC8a on pins 5 and 3. IC6a and IC7a are analogous to the transformer used to provide balanced anti-phase inputs to a conventional product detector.

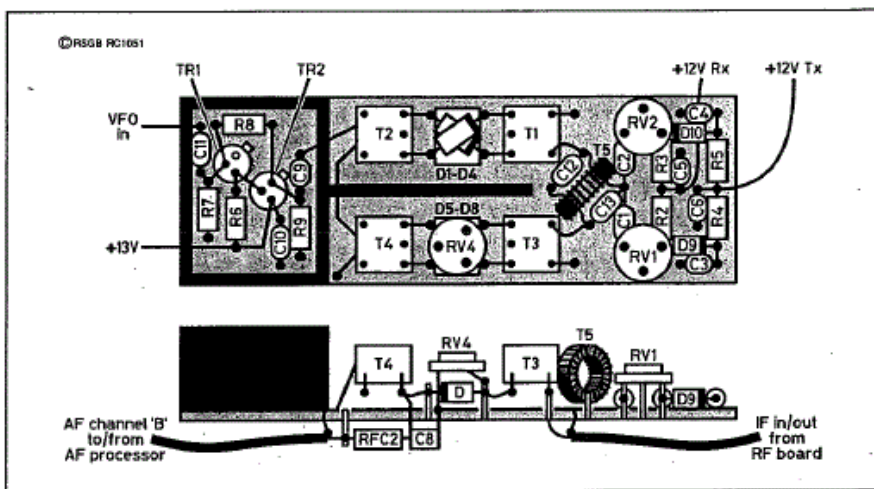


Fig 18: Detector board drilling template and component location. The board is single sided with the copper on the top (component) side. The diode rings, baluns, preset pots and phase shift network are mounted on Vero pins.

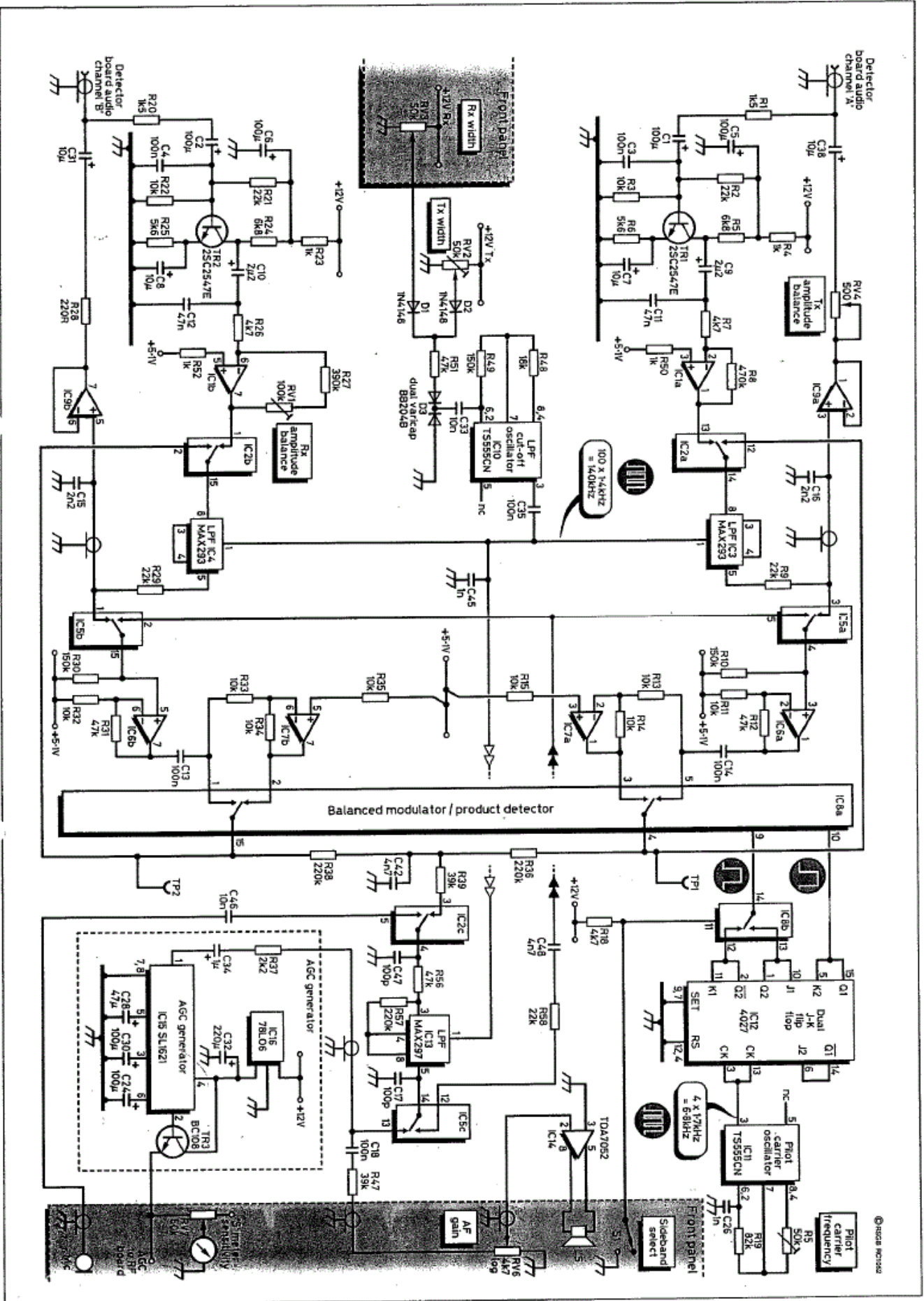


Fig 19: AF Processor board. See Fig 20 for power rail connections, Fig 21 for board layout.

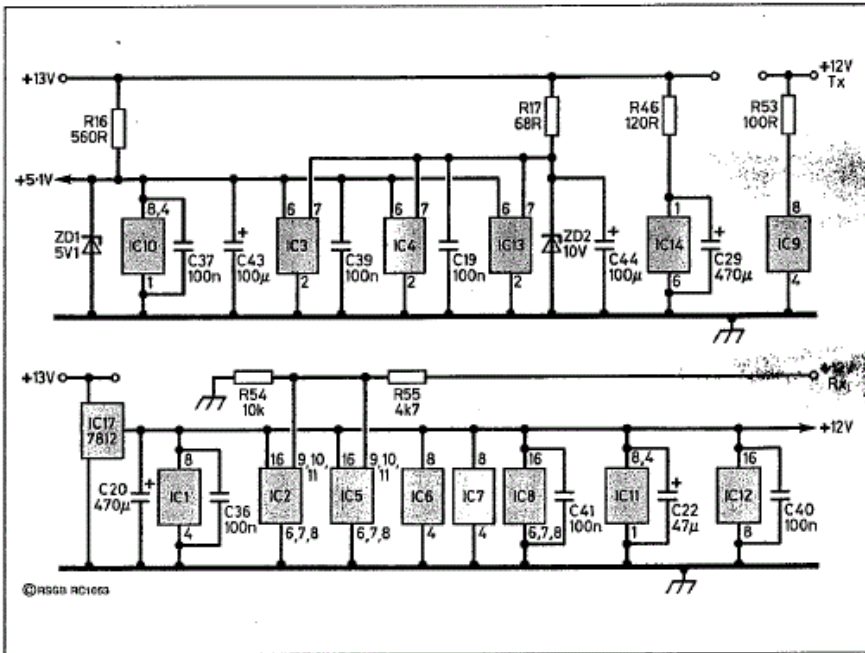


Fig 20: AF Processor board - power rail connections.

COMPONENTS FOR VFO

Resistors
 R1, R3 100k
 R2, R5 100R
 R4 330R
 R6 150R

Capacitors
 C1, C2 560p Polyester
 C3 - C7 10n disc ceramic
 C8 33p Silver mica
 C9, C10 100p Silver mica
 C11 50p air spaced variable

Semiconductors
 D1 1N4148
 TR1, TR2 2N3819
 ZD1 9V1 300mW Zener

Additional Items
 FB 4t 32SWG on ferrite bead FX1115
 L1 see text
 RFC1, RFC2 1mH

IC11 is a conventional oscillator running at four times the injection frequency required by the product detector. IC12 produces quadrature outputs and in the process, divides by four.

IC8b passes either the Q2 output or its inverse to provide sideband selection.

The product detector, IC8a is simply a single-pole 2-way CMOS switch toggled at the Pilot carrier frequency. Carrier leak is a function of the isolation between the switching line and the signal line and is for all practical purposes non-existent.

Audio channel 'B' operation is identical,

except that an amplitude trim pot is included in the feedback loop around IC1b. The two audio channels are summed at the junction of R36 and R38 to give sideband cancellation. For this to work well, it is essential that there be no differential phase shift at any frequency between the two channels. To this end, close tolerance R's and C's are specified and if you have the facility to match capacitors between channels, so much the better.

IC13, a MAX297 LPF chip removes any switching noise caused by the product detector and provides further frequency response tailoring.

IC13 is clocked from the same line as IC3 and IC4, but because the MAX297 has a 50:1 clock:cutoff ratio (as opposed to 100:1 for the MAX293), it has twice the bandwidth ie about 2.8kHz. This provides sharp top cut of the Rx audio.

IC14 is a minimum component count audio amplifier with more than enough output to drive a loudspeaker. An optional output to drive low impedance 'phones is not shown.

IC15 is a hang AGC generator and TR3 ensures enough output to drive the low impedance AGC load on the RF board as well as the

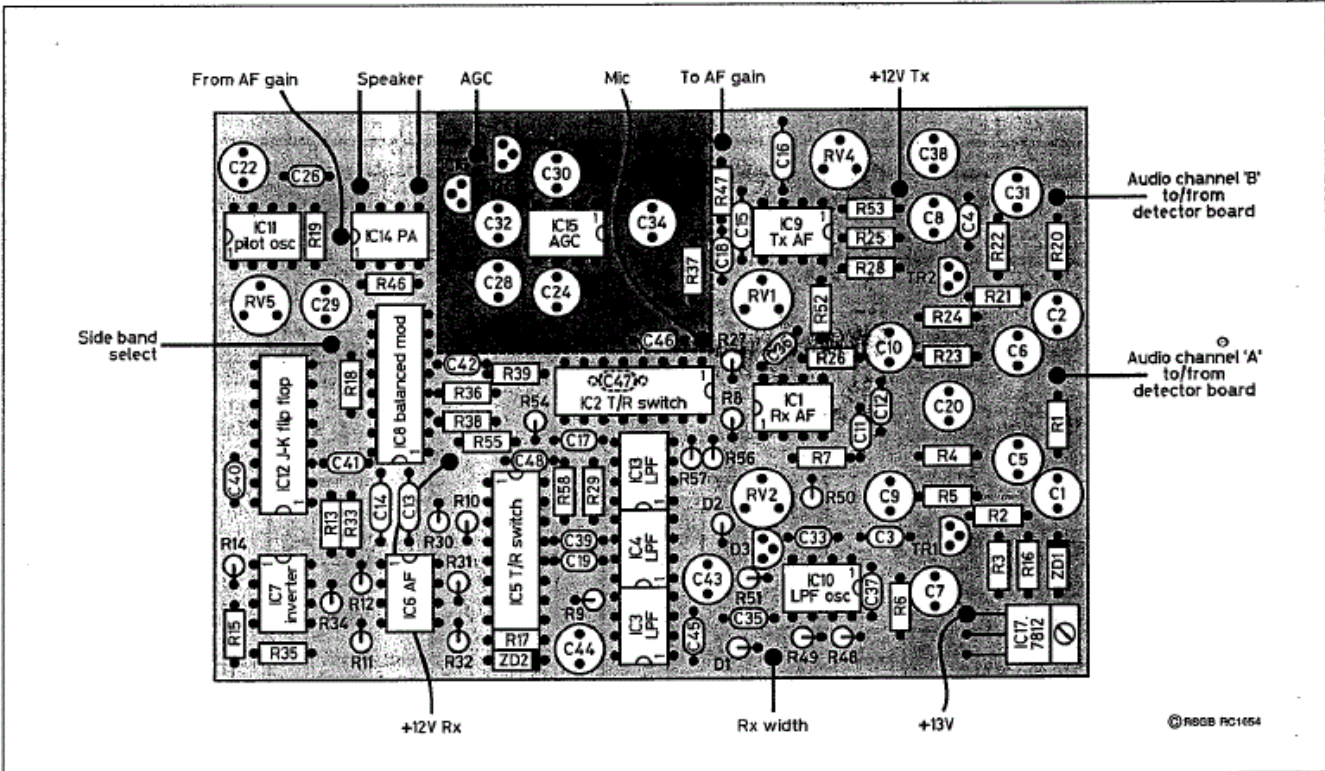


Fig 21: AF Processor Board drilling template and component layout (top view). The enclosed area shows the alternative AGC generator position.

'S' meter. In operation, this produces smooth AGC action with no sign of the thumps and bumps sometimes encountered. Since the original development work was completed, the SL1621 chip has become quite difficult to obtain. An alternative approach, which I first saw published by G3TXQ [8] will be described in the final part next month.

On transmit, the process is reversed. The high impedance microphone input is fed to the MAX297 LPF. This removes any risk of the higher voice frequencies mixing with the 3rd harmonic of the pilot oscillator to produce spurious outputs. The uncommitted op-amp on the chip is used to provide modest mic gain before filtering.

The tailored mic output feeds the two audio channels in parallel via the T/R switches IC5c and IC5a, IC5b. IC6a and IC6b raise the audio to a suitable level for the balanced modulators. The voice spectrum is mixed with the quadrature pilot tones in IC8a which now acts as a balanced modulator for each channel. The two quadrature outputs are routed to the low pass filters via IC2 and then to the transmit buffers IC9a and IC9b. RV4 trims amplitude balance and the two channels then route to the Detector board.

To avoid T/R switching transients, all the receive-only elements, including the AGC generator remain powered up on transmit. IC5c prevents any audio feed to the loudspeaker. However should there ever be excessive RF in the shack, strange noises emanate from the speaker as a warning!

If you want to bleed in a little sidetone, the simple addition of a high value resistor between IC5 pins 14 and 13 will achieve this and with careful adjustment of the bleed resistor, the AGC generator could also provide Tx drive limiting at moments of undue excitement.

CONSTRUCTION

ONE OF THE consequences of redesigning to accommodate the alternative AGC is a completely new layout of the AF Processor board since the photographs were taken. This is also a response to a desire to combine the AF Processor and the Detector board as a single assembly. This is an approach worth considering.

Construction is straightforward, but check the lead spacing of your capacitors before drilling the board. Having drilled the board and before mounting the sockets, my practice is to take a red felt-tip pen and circle round pin

COMPONENTS FOR DETECTOR BOARD

Resistors

R1	47R, fit on test, see text
R2 - R5	1k
R6	1k5
R7	12k
R8	10k
R9	470R
RV1, RV2	100R horizontal Cermet
RV3, RV4	250R horizontal Cermet

Capacitors

C1-C6	100n disc ceramic
C7, C8	100n 5% poly
C9-C11	10n disc ceramic
C12, C13	560 pf 5% poly
D1-D8	BAT17
D9, D10	BA244

Semiconductors

TR1, TR2	BSX20
RFC1, RFC2	100µH axial choke
T1-T4	Toko 456DB-1007, see text
T5	8 bifilar turns of 28SWG (0.375mm) on T50-6

1 of each IC position on the underside of the board. This greatly improves visualisation.

I would commend running in all the power lines to the IC sockets first and mounting the decoupling capacitors shown in Fig 20. Then just wire it all up as per the circuit diagram. Avoid trying to group wires into neat bundles. This increases the risk of cross-coupling. Straight and 'ugly' point-to-point wire runs are preferred. The only screened leads I used are as marked on the circuit diagram. The remainder are thin self-fluxing enamelled copper. My preference is to work backwards from the loudspeaker, ensuring that as each IC is inserted, a finger applied to the input pin produces a resounding hum.

● Next month, we conclude with the alternative AGC circuit and the alignment procedure.

REFERENCES

[8] 'Portability', G3TXQ, *RadCom*, Feb 1989. ♦

COMPONENTS FOR AF PROCESSOR BOARD

Resistors. All resistors are 1%, ¼watt

R1, R20	1k5
R2, R21, R9, R29, R58	22k
R3, R11, R13-R15, R22, R32-R35, R54	10k
R4, R23, R50, R52	1k
R5, R24	6k8
R6, R25	5k6
R7, R18, R26, R55	4k7
R8	470k
R10, R30	150k
R12, R31, R51, R56	47k
R16	560R
R17	68R
R19	82k
R27	390k
R28	220R
R36, R57, R38	220k
R37	2k2
R39, R47	39k
R46	120R
R48	18k
R49	150k
R53	100R
RV1	100k horizontal Cermet preset
RV2, RV5	50k horizontal Cermet preset
RV3	50k on front panel
RV4	500 horizontal Cermet preset
RV6	4k7 log on front panel
RV7	50k horizontal Cermet preset mounted on 'S' meter

Capacitors

C1, C2, C3, C6, C24, C30, C43, C44	100µF
C3, C4, C13, C14	100n 5% poly
C7, C8, C31, C38	10µF
C9, C10	2.2µF
C11, C12	47n 5% poly
C15, C16	2n2 5% poly
C17, C47	100p ceramic
C18, C19, C35-C37, C39-C41	100n disc ceramic
C20, C29	470µF
C22, C28	47µF
C26	1n 5% poly
C32	220µF
C33	10n 5% poly
C34	1µF
C42, C48	4n7 ceramic
C45	1n disc ceramic
C46	10n disc ceramic

Semiconductors

D1, D2	1N4148
D3	Dual Varicap BB204B
IC1, IC6, IC7, IC9	TL072
IC2, IC5, IC8	4053 3p 2w CMOS switch
IC3, IC4MAX293, both mounted in one 16-pin DIL socket	
IC10, IC11	TS555 CN
IC12	4027 CMOS dual J-K flip flop
IC13	MAX297
IC14	TDA 7052 AF amp
IC15	Plessey SL1621
IC16	78L06 regulator
IC17	7812 regulator
TR1, TR2	2SC2547E or equiv low noise transistor
TR3	BC108 or equiv
ZD1	5v1, 300 mW Zener
ZD2	10v, 300 mW Zener

Additional items

S1	1p 2w toggle on front panel
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Third-Method SSB HF Transceiver

The final part by Peter Rhodes, BSc, G3XJP*

IN THIS CONCLUDING part, the remaining details on the RF side are described and the overall system wiring is summarised. Finally, the full alignment procedure is elaborated.

ALTERNATIVE AGC

FIG 22 SHOWS the detail of an alternative AGC generator which I first saw published by G3TXQ [8]. The performance is indistinguishable in this application. The board drilling template is shown in **Fig 23** and this precisely fits the space outlined on the AF Processor board - except that C46 will need to be relocated under the board. Another viable alternative would be to build it as a separate board and mount it together with C18 and R47 near the AF Gain pot on the front panel.

Although a single TL084 appears to be the obvious implementation, the pin-out of this device is very hard to accommodate; two TL082s are therefore used instead.

In brief, IC1a amplifies the audio substantially with IC1b acting as a unity gain inverter. Together with D1-D4, they provide full-wave detection of the audio. Thus a rapid response occurs irrespective of whether the signal occurs on a positive or negative-going audio cycle. R13 slows the response up somewhat, but provides protection against the AGC hanging on a noise spike.

C7 and R15 provide AGC memory and IC2b buffers the stored voltage onto the AGC line. D6-D8 are included as necessary to reduce the no-signal AGC voltage to zero. You may not need them.

R14 via D5 and IC2a is another discharge path for C7. It occurs during any pauses in audio output, but only after the time-constant R10/C5. This is the 'hang' time and you may want to experiment with the value of C5. R14 determines the discharge time at the end of this hang time. R7 may also need some adjustment. Its value should cause TR1 to conduct on signals of about 'S4' or more.

None of the timing components has the values used originally by G3TXQ; and you in turn may want to customise them to suit your preferences.

LOW PASS FILTER BOARD

THE LPF IS OPERATIVE both on transmit and receive. On transmit, it provides signifi-

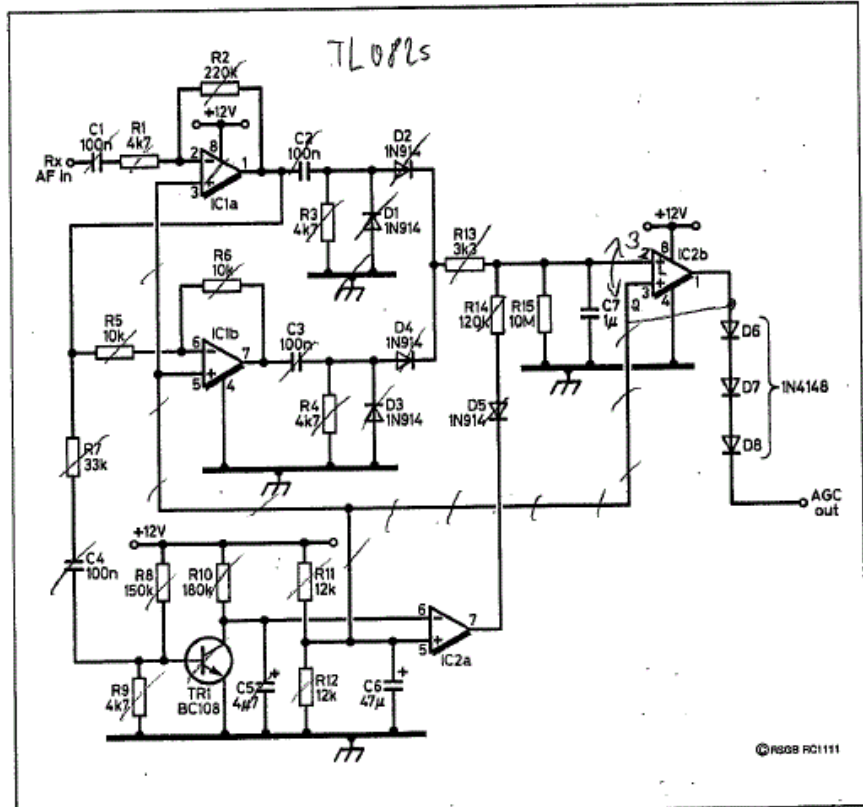


Fig 22: Alternative AGC generator.

cant attenuation of the harmonic content in the PA output. On receive, while having no obvious effect on in-band performance, it provides extra protection against strong out of band signals in general and VHF broadcast stations in particular.

The design is based on the original G3TSO

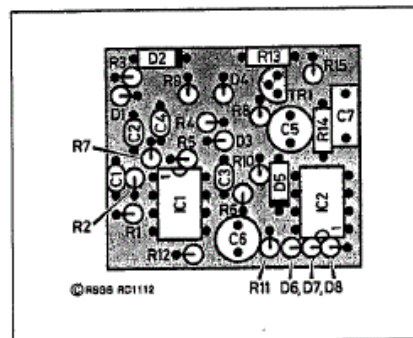


Fig 23: Alternate AGC board layout.

article with two significant differences. Firstly, I used somewhat larger relays for switching the filters. This is because I have some vague ideas about upgrading the PA later. Secondly, I have made no provision for internally generated ALC.

You may choose to add the G3TSO ALC circuitry and feed the result back onto the AGC/ALC line - or indeed you could build the entire board to the G3TSO design. It is fully compatible.

The circuit I used is shown in **Fig 24**, with the board layout in **Fig 25**. For the WARC bands, the next highest frequency filter is used. The associated switching is handled on the diode matrix board adjacent to the band switch. (See Part 3, Fig 5).

All the relays are mounted under the board and secured to it in most cases by arranging for diagonally opposite pins to be soldered to the ground plane. RL1 and RL2 were secured by soaking a thin layer of foam in rubber solution

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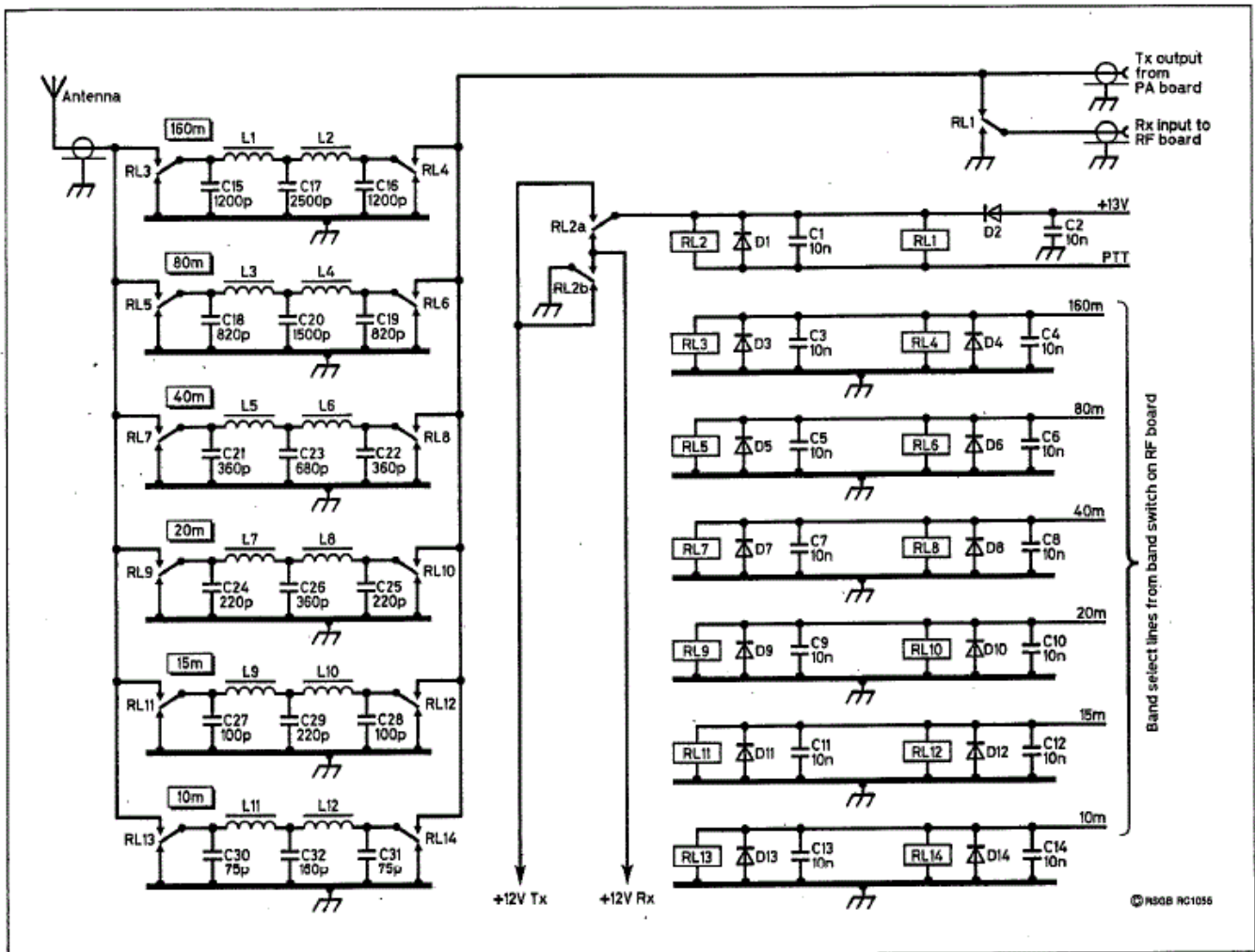


Fig 24: Low pass filter board. Antenna change-over and switching of the 12V rail between transmit and receive is also provided.

and then clamping them lightly to the board until set. This provides some useful sound deadening during T/R switching.

Before drilling the board, check the pins on your relays against the template. I have managed to acquire two slightly different spacings against the same type number.

The input/output busbars are made from stiff wire and mounted under the board on Vero pins.

No alignment is required beyond the normal checks for earth shorts and continuity before applying RF power.

When the transceiver is powered off, the antenna is disconnected from the PA. It might be a wise precaution during early testing, to connect a temporary jumper between the input and output busbars - and disconnect all the LPF select lines - to ensure that the transmitter PA is never looking at an open circuit load. This will, of course, render the LPF inoperative.

TRANSMIT DRIVER

A UTILITY broad band amplifier is shown in Fig 26. It has low input and output impedances and is used to raise the power level generated on the RF board to a suitable level for driving the PA. The result is a generous surplus of drive on all bands.

POWER AMPLIFIER

THE PA BOARD is supplied as a standard kit from Cirkit Holdings.

Mine has provided several years of reliable service. It comes complete with all components, PCB and full assembly and alignment instructions. Some modifications were suggested by G3TSO in his original design and these are well worth incorporating. I suggest you postpone carrying them out until you are confident the PA is performing as supplied.

To avoid switching the +13V power rail to the amplifier, the supply to the two bias regulators is switched instead. The main +13V rail is left permanently connected.

This involves cutting the main power track on the PA board where it feeds the bias regulators via Links 4 and 5. The regulators are then fed from the +12 volt Tx line instead.

The two 100Ω PA feedback resistors R13 and R14 should be changed to 1W rating to avoid overheating. R18 in the bias circuit should be changed from 10kΩ to 6k8 to improve the bias regulation.

The PA board is used as part of the rear wall of the transceiver enclosure, with its heatsink protruding from the back. Care needs to be taken in mounting it, as the power rails on the board run close to the edges.

SYSTEM WIRING

ALL THE INTER-CONNECTING wiring is shown in Fig 27.

The detail for the front panel pots can be found on the circuit diagrams for the board which they feed. The only new components introduced here are the Power On/Off switch and a generous reservoir capacitor across the supply rail.

COMPONENTS FOR TRANSMIT DRIVER

Resistors	
R1	56
R2	5R6
R3	470
R4	1k
R5	100
Capacitors	
C1-C4	10n disc ceramic
Inductors	
T1	12 bifilar turns of 30 swg (0.315mm) on FT37-43
Semiconductors	
TR1	2N3866 with small heatsink

TRANSCEIVER ALIGNMENT

THE BASIC SEQUENCE is to get the receiver roughly right, then the transmitter and, finally, complete the receiver alignment.

TEST EQUIPMENT REQUIREMENTS

You need to be able to measure the Pilot oscillator frequency and for this purpose an audio frequency counter is ideal. The other approach is comparison with a known tone. Suggested sources are: a reliable audio signal generator; a musical instrument; a tuning fork; a beat note from a calibrated receiver. You need to be able to set the Pilot oscillator as close as you can to 1700Hz, but between 1600Hz and 1800Hz will produce acceptable results.

The other frequency to be set is that of the LPF cut-off oscillator. This runs at 100 x the cut-off frequency. The cut-off should be set to 300Hz below the Pilot oscillator, ie 1400Hz giving 140kHz for the cut-off oscillator frequency. This can be measured either with a counter or by very loosely coupling a calibrated receiver near to the oscillator and measuring the separation between harmonics in the HF spectrum. So long as the cut-off is somewhat below the Pilot oscillator frequency, it can be refined later.

Some means of generating a stable carrier

with no modulation and with an output in one of the chosen amateur bands is required to adjust the receiver. An existing CW transmitter which uses a tone generator may well not be suitable since any residual carrier, opposite sideband or impurities in the tone may give misleading results. I would commend the small investment of building a simple crystal oscillator, battery powered, with an output in the 80m band.

To adjust the transmit side, you will need a single tone generator to feed into the microphone socket and another SSB/CW receiver in order to monitor and adjust the resultant transmitter output.

As well as the above, a twin or chopped beam 'scope is invaluable in providing visual confirmation that your adjustments are moving in the right direction and provides a sense check that all is well. This is by no means essential, but the trace produced by the receiver is captivating enough that I rarely switch it off!

BEFORE POWERING ON

You will doubtless have your own strategy for commissioning a project of this scale. At the least it would be good practice to check for shorts on the 13V and 12V Tx and 12V Rx power rails and on the PTT line.

Then set the following pots to reasonable starting positions:

- Tx drive level to minimum.
- Rx audio gain to mid travel.
- The two carrier balance presets on the Detector board to mid travel.
- The Rx and Tx phase balance pots on the Detector board (RV1 and RV2) to minimum resistance.
- The Rx and Tx amplitude balance pots on the AF Processor board (RV1 and RV4) to mid travel.
- The two AGC threshold level pots on the RF board (RV1 and RV2) to mid travel - and unplug the microphone.

RECEIVER ADJUSTMENT

The first step is to get to the point where the receiver is clearly producing noises which show that you are listening to a chosen amateur band. SSB signals may not be intelligible at this stage.

Roughly peak the four RF coils for the band in question and likewise the two trimmers for the IF bandpass filter.

Now tune in a constant low level signal - band noise is ideal - and set each AGC threshold pot on the RF Board to maximum receive output and then adjust to just before the point where the output starts to drop. Tuning across a loud signal should now produce definite hang AGC action and a reading on the 'S' meter.

Now set the Pilot oscillator to 1700Hz (AF Processor Board, RV5) and tune across a clean carrier. Ideally one should get an output from about 300Hz to 3kHz with a narrow notch at the Pilot oscillator frequency near the middle.

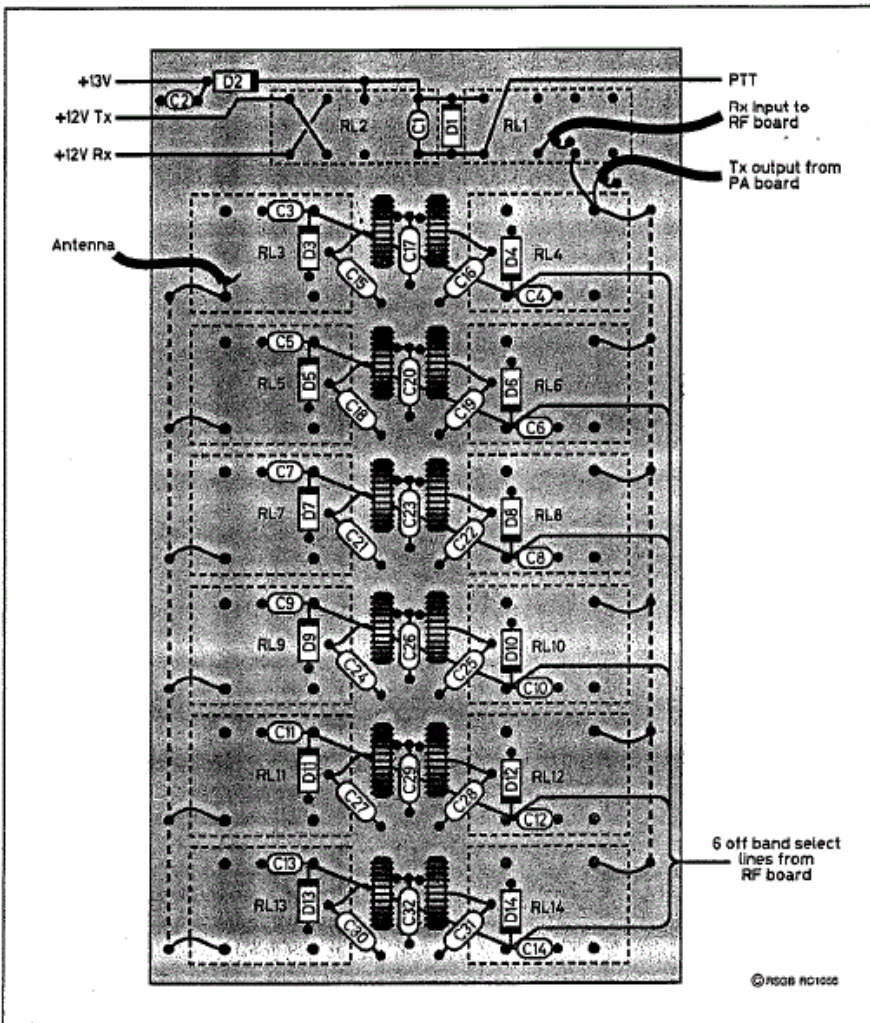


Fig 25: LPF board drilling template and component layout. Some of the hole positions will depend on the size of the filter capacitors.

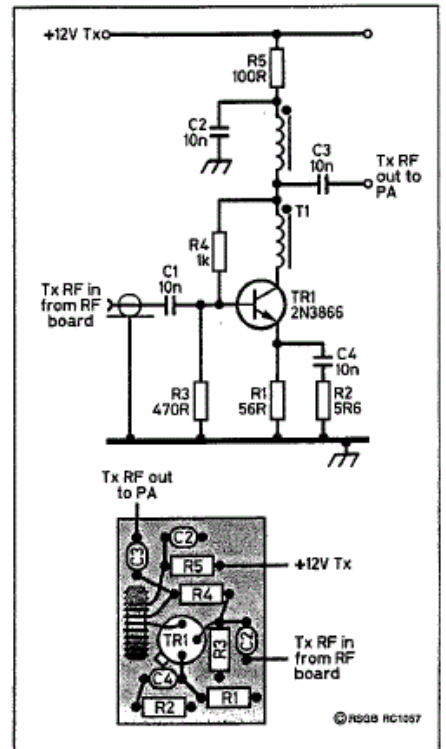


Fig 26: Transmit Driver circuit diagram, component layout (top view). The board is mounted immediately adjacent to the RF input of the PA and connected to it by about 2cm of plain wire.

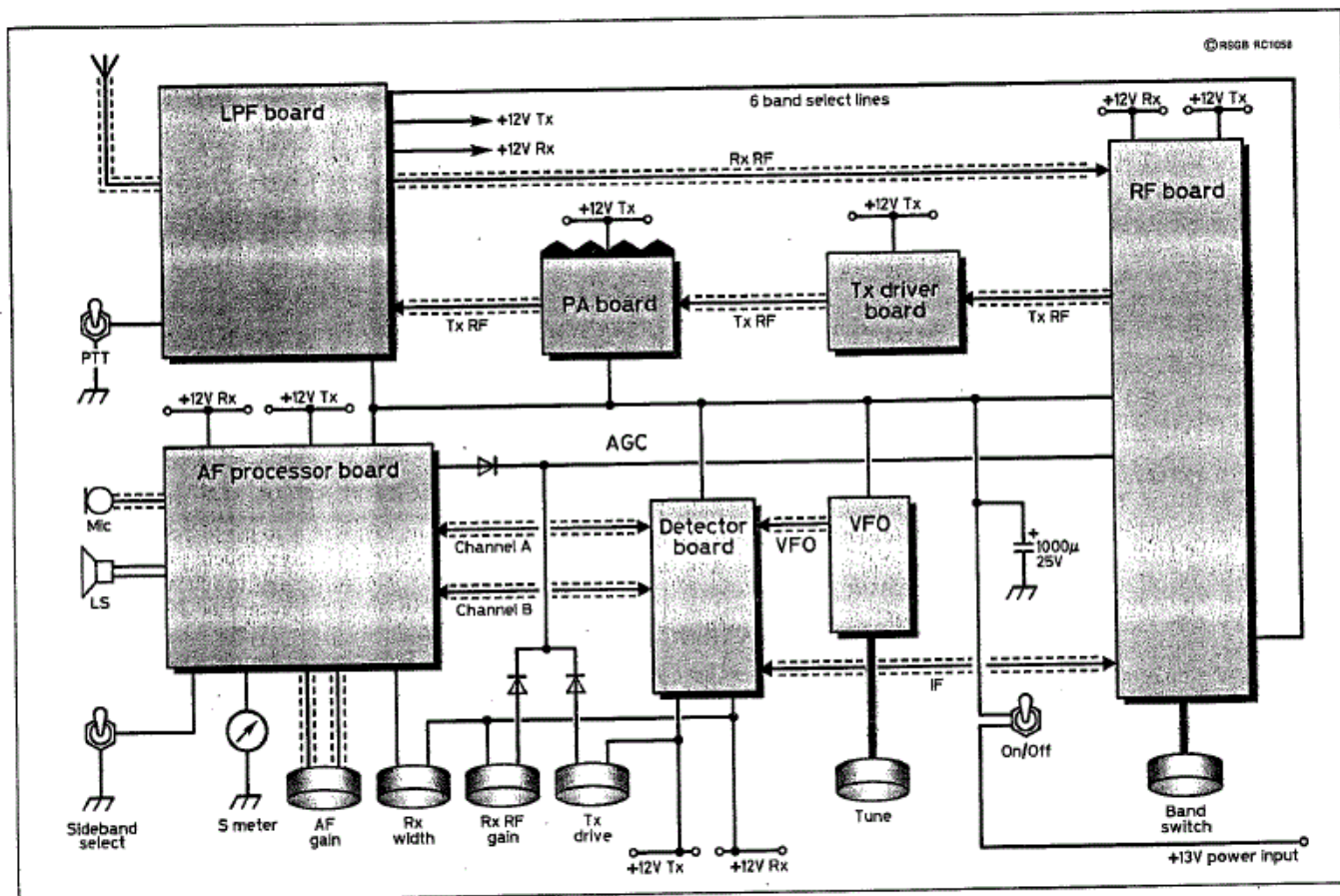


Fig 27: System wiring diagram. Although not to scale, it also shows approximate board location and physical cabling.

Fig 28 shows the worst case response when fully mistuned. The significant point to note is that there are various outputs going up and down in frequency, and that they include responses right down to zero and upward to very high audio frequencies.

Reduce the LPF cut-off frequency (Rx Width on front panel) to 1400Hz (recall that the cut-off oscillator runs at 100x this frequency) and this should essentially eliminate all the extreme low and high frequency components as you tune across the carrier. Specifically there should now be a noticeable gap as you tune through zero beat. Fig 29 shows the response with the cut-off frequency reduced but still not low enough. Note that two of the unwanted products have nearly been eliminated and the high frequency response has been significantly curtailed. But the low frequency response down to, and either side of, zero beat is still present.

Fig 30 shows the response with the LPF set correctly. As you tune across the carrier there are two responses; one going from high to low pitch, the other going from low to high. In practice, one of these will probably be stronger because there will be some accidental phase balancing at this stage. There are no very low or high frequency responses and the basic bandwidth of the receiver is now determined.

The next step is to remove one of these two responses - either one - by phasing it out. To this end, tune until you can hear both a high frequency note and a low frequency note at the same time. Then adjust in turn RV1 on the AF Processor (Rx amplitude balance) and RV1 on the Detector board (Rx phase balance) to minimise the low

frequency note. At this stage it may not disappear completely, but it should be noticeably attenuated.

If you have a double-beam scope connect the two probes to TP1 and TP2 on the AF Processor board, set the two 'scope channels to the same gain and set one channel as the time base for the other. (This is the standard settings for Lissajous figures). The desired trace is a circle. If the circle is squashed (ie an ellipse), adjust the amplitude balance until the trace is circular. If there are two ellipses, adjust the phase balance until they coincide. Fig 31 shows a typical trace with both phase and amplitude imbalance. In practice, tuning by ear will give more precise results than tuning by eye, but using both together will tell you if there is any fundamental problem such as a wiring error.

You now have a single sideband receiver! It may, however, be on the wrong sideband, so simply flick the sideband select switch at this stage. There may also be evidence of AM detection because the carrier balance has not yet been adjusted. This latter is best done on the transmit side.

The 'scope pattern when listening to an SSB signal should now be a series of concentric circles with a maximum diameter of about 400mV, subject to AGC action.

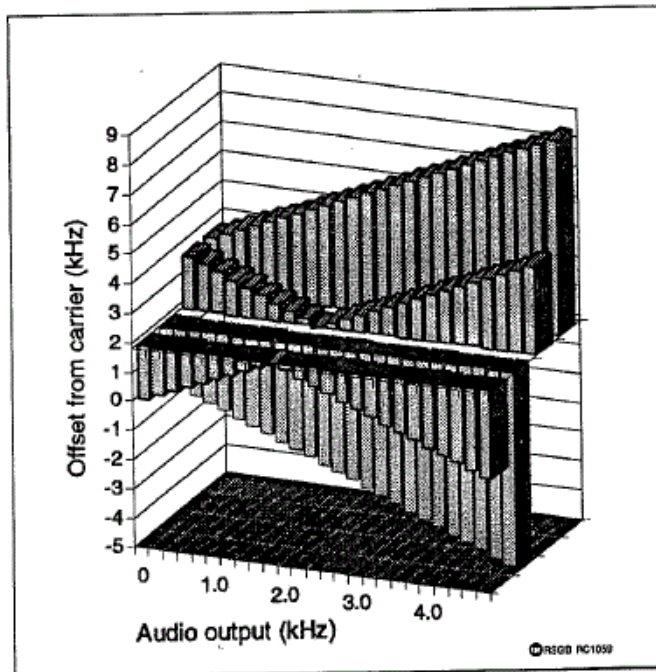


Fig 28: Worst case 'Third Method' response with no phase or amplitude balance and no effective LPF.

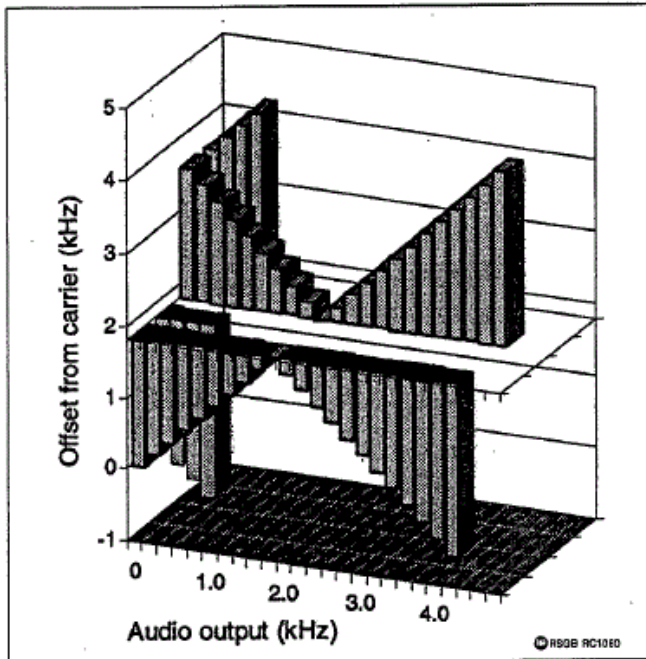


Fig 29: Response with no phase or amplitude balance and LPF cut-off improved from Fig 28, but still too high.

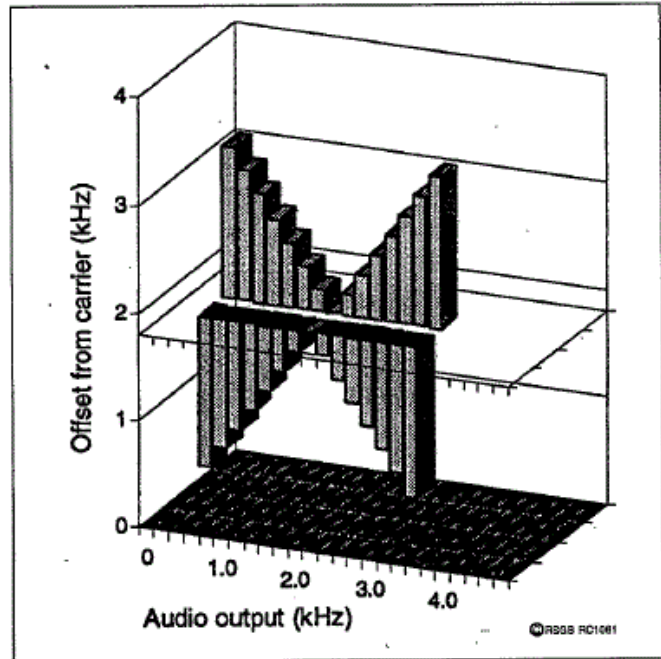


Fig 30: Response with no phase or amplitude balance and LPF set correctly. This is Third Method DSB.

TRANSMITTER ADJUSTMENT

Start off by adjusting the Cirkit PA bias in accordance with the instructions supplied. Connect the transmitter to a dummy load and establish that you can generate some sort of RF output in the chosen amateur band and that the Tx Drive level control operates smoothly.

Connect a pure 1kHz (or thereabouts) tone generator to the microphone socket and monitor the result on another receiver. Set the tone generator output to the lowest level which gives a sensible output. It is essential to avoid overdriving the audio stages and generating harmonics. With a continuous single tone input, also avoid turning the drive level up as the PA will quickly overheat. Verify that T/R switching is clean, and that there is no sign of instability.

If you have a 'scope, connect it to monitor the transmitter output and set the timebase in the audio range. In all probability you will see the conventional 'stripe' trace associated with a carrier but with one or more audio frequency ripples. The general idea is to eliminate the ripple.

The first adjustment is to carrier balance. With nothing connected to the microphone socket, locate the residual carrier on the monitor receiver. Adjust in turn RV3 and RV4 on the Detector board to minimise the carrier leak. You will find the two adjustments interact somewhat but you should be able to reduce the leak to a comparable level to the broad band noise output. At some stage you should also establish that any residual leak goes up and down in amplitude with the Tx Drive control. If not, this is evidence that there is some coupling between the VFO output and the RF or IF stages which is bypassing the balanced modulators and this must be eliminated or reduced to an insignificant level before proceeding.

Once the carrier has been suppressed, recon-

nect the tone generator and tune across the transmitter output. Exactly the same outputs will be found as on the receive side and in the worst case, all four tones shown in Fig 28 will be audible at similar magnitudes. First adjust RV2 on the AF Processor to set the LPF cut-off on transmit to 1400Hz. This will eliminate two of the tones. Now locate the weaker of the two remaining tones and tune the monitor receiver to place it centrally in the passband. Adjust in turn RV4 on the AF Processor and RV2 on the Detector board to minimise this weaker tone. Repeat this adjustment and the carrier leak adjustment - which interact somewhat - until one tone remains and everything else is buried in the broad band noise.

The 'scope should now show a clean trace and almost any residual ripple is a cause for concern. Again, monitoring the output on another receiver is a more sensitive test than measuring percentage ripple on the scope, but the latter is a quicker indication that you are moving in the right direction.

Now connect a high impedance microphone and before going any further, confirm you have SSB output on the correct sideband, ie the same sideband as produced by the receiver adjustment. If not, swap over the leads on any part of the AF Processor which is unique to either Tx or Rx - but not common to both. The inputs to IC 9, pins 3 & 5 are probably the easiest. If you have to do this, repeat the transmitter adjust-

COMPONENTS FOR LPF

Capacitors	
C1 - C14	10n disc ceramic
C15, C16	1200pF SM 350v
C17	2500pF SM 350v
C18, C19	820pF SM 350v
C20	1500pF SM 350v
C21, C22	360pF SM 350v
C23	680pF SM 350v
C24, C25	220pF SM 350v
C26	360pF SM 350v
C27, C28	100pF SM 350v
C29	220pF SM 350v
C30, C31	75pF SM 350v
C32	160pF SM 350v

Semiconductors	
D1 - D14	1N4001

Additional Items

Note: 26 SWG = 0.45mm, 20 SWG = 0.9mm

L1, L2	31t of 26 SWG on T50-2
L3, L4	22t of 20 SWG on T50-2
L5, L6	18t of 20 SWG on T50-6
L7, L8	12t of 20 SWG on T50-6
L9, L10	10t of 20 SWG on T50-6
L11, L12	9t of 20 SWG on T50-6
RL1, RL2	SMR12D (16-pin DIL)
RL3-RL14	SPCO Type KUIT with 12V coil

COMPONENTS FOR ALTERNATIVE AGC

Resistors	
R1, R3, R4, R9	4k7
R2	220k
R5, R6	10k
R7	33k
R8	150k
R10	180k
R11, R12	12k
R13	3k3
R14	120k
R15	10M

Capacitors	
C1 - C4	100n disc ceramic
C5	4.7µF
C6	47µF
C7	1µF poly, not electrolytic

Semiconductors	
D1 - D5	1N914
D6 - D8	1N4148 or equiv (as needed, see text)
IC1, IC2	TL082
TR1	BC108 or equiv

ment in full on the correct sideband. You may also want to experiment with both the pilot tone and cut-off frequencies to suit your microphone and voice characteristics.

FINAL ADJUSTMENTS

Having adjusted the transmitter carrier balance, this will have altered the receive amplitude balance, so this adjustment needs repeating. The basic interaction rule is that no adjustment you make on the receive side will alter anything on transmit, but any change to carrier balance will effect both.

When you are happy with the operation of the 'Third Method' aspects you can now proceed to align the RF stages for each band. For all practical purposes this means peaking one RF coil near the top of the segment of interest, one near the bottom and the other two either side of centre. The IF trimmers should be peaked to give a flat response across the tuneable IF passband and some experimentation with C58 may be needed to achieve this. Absolutely avoid using a large value here just be-

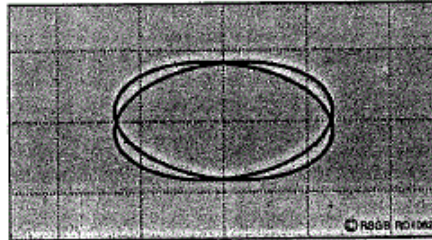


Fig 31: 'Scope trace showing both phase and amplitude imbalance.

cause it gives more output. The idea is to span the 500kHz, but no more.

OPERATION

SHOULD BE A DELIGHT! The quality on receive should be smooth and round with the Rx Width adjusted to suit prevailing conditions. The receiver is deliberately designed to have excess gain, so under noisy conditions you must reduce the RF Gain to eliminate any overload. On transmit, you should receive reports of outstanding audio quality in a tight

bandwidth. Again, there will probably be a surplus of drive and you must avoid using it.

I hope you gain as much pride and pleasure from building and using this transceiver as I do! I extend my thanks to the home-brew net on 80m for much patient help and on-air testing. G3DKJ, G3LUB, G3GMN, G3WIB, G3XKF, G4HMC and G4HUV deserve special mention for their discerning ears and patience. Thanks to Steve Largent, G4ZBV, for the photography. Last, but not least, to my wife Fran for proof reading and tolerating two years of development time and mess.

CORRECTION

IN PART 3, Fig 12, the band select line for the LF Configuration is shown routed to a pin on L2, which should be earthed. The adjacent unused pin on L1 should be used as a tie point instead. My thanks to G0VBJ for spotting this one, and apologies.

In part 4, Figs 19, pin 3 and 4 of IC 13 were transposed. My thanks to G3PTN. ♦

THIRD METHOD TRANSCEIVER

RADCOM JUNE - OCTOBER 1996.

PART 2, FIG 7, R49 and R50 should be changed to 1k to improve manual gain.

On Part 4, Fig 16 there is a small piece of PC track missing which should connect C8 to R3 as per circuit diagram in Fig 15.

In Part 5, Fig 22 there is an error in the circuit diagram for the alternate AGC generator around IC2b. Pins 2 and 3 should be transposed and then pin 2 (inverting input) should be connected only to the output, Pin 1.

Some constructors have reported the microphone gain to be somewhat lacking. If you find this to be the case, then I suggest adding a simple microphone preamp with a high