11 Practical Microwave Receivers and Transmitters

Many of the techniques for generating and receiving microwave frequencies were investigated and developed more than 70 years ago, in the 1930s. Microwave usage was given added impetus by the development of radar and the advent of the Second World War. Before 1940, the definition of the higher parts of the radio frequency spectrum [1, 2] read like this:

- *30 to 300Mc/s* *Very high frequencies (VH/F)*
- *300 to 3000Mc/s* *Decimetre waves (dc/W)*
- *3000 to 30.000Mc/s* *Centimetre, waves (cm/W)*

Radio frequencies above 30,000Mc/s (now 30GHz) apparently did not exist! Various definitions have appeared in the intervening years. These have included terms such as super-high frequencies (SHF) and extra-high frequencies (EHF).

In the course of time, the unit of frequency cycles per second (c/s), its decimal multiples, kilocycles per second (kc/s) and megacycles per second (Mc/s), have been replaced by the unit hertz (Hz), its decimal multiples kilohertz = 10^3Hz (kHz), megahertz = 10^6Hz (MHz), gigahertz = 10^9Hz (GHz) and terahertz = 10^{12}Hz (THz).

Today, the term microwave has come to mean all radio frequencies above 1000MHz (1GHz). The division between radio frequencies and other electromagnetic frequencies, such as infra-red, visible (light) frequencies, ultra-violet and X-rays, is still not well defined since many of the techniques overlap, just as they do in the transition between HF and VHF or UHF and microwaves. There has been keen interest in amateur communications using infra-red and visible light, over the last few years, so a section on it has been included in this chapter. To a large extent the divisions are artificial insofar as the electromagnetic spectrum is a frequency continuum, although there are several good reasons for these divisions.

Around 1GHz (30cm wavelength) the lumped circuit techniques used at lower frequencies are replaced by distributed circuit techniques such as resonators and microstrip. Conventional components, such as resistors and capacitors, become a significant fraction of a wavelength in size so surface-mount devices (SMDs) are used which are very small and leadless. These require special techniques for constructors; these are described later.

Conventional valves (vacuum tubes) and silicon bipolar solid state devices are usable beyond 1GHz - perhaps to about 3.5GHz - and, as frequencies increase, these devices are replaced by special valves such as klystrons, magnetrons and travelling-wave tubes. The first semiconductor devices to be used at the higher microwave frequencies were varactor diodes, PIN diodes and Gunn diodes. Because of the massive development in semiconductors for use by commercial telecommunications systems there are now a wide range of transistors available to radio amateurs including gallium arsenide field effect transistors (GaAsFETs), metal epitaxial semiconductor field effect transistors (MESFETs), pseudomorphic high electron mobility transistors (pHEMTs) and many more.

The other device that has revolutionised microwave designs is the microwave monolithic integrated circuit (MMIC); these are usually designed to work in matched 50-ohm networks, making them easy to use and (usually) free from instability problems. They are used as building blocks, in a similar way that the operational amplifier (OP-amp) is used in DC and audio designs, but are extremely wide band and often require bandpass filtering to obtain the desired results.

Many of the more exotic semiconductor devices are now appearing on the surplus market making them more acceptable to the amateur's pocket. These can be used to build new amateur equipment or whole surplus units can be modified to work on amateur bands. Quite a few articles have appeared, describing the modification of commercial equipment, so there is an example of such an article later in this chapter.

Other advantages of operation in the microwave spectrum are compact, high-gain antennas and available bandwidth. None of these advantages are attainable in an amateur station operating on the lower frequency amateur bands. High gain antennas are impossibly large below VHF, and the levels of spectral pollution from man-made and natural noise are such that low noise receivers, needed to handle weak signals, cannot now be effectively used, even at VHF. Communication on the many available microwave bands over distances of hundreds of kilometres is now quite common (sometimes over thousands of kilometres, given favourable tropospheric propagation conditions or the use of amateur satellites or moonbounce). This destroys the perception that microwaves are useful only over limited line-of-sight paths! A good place to find details of the latest records and operating conditions is *Dubus Magazine*; all of the main amateur bands are reported with details of activity using various types of propagation.

There is an increasing amount of commercially produced equipment available from amateur radio retailers for all of the microwave bands and plenty of designs for the constructor. It is true that attaining really high transmitter power output above about 3 or 4GHz is still difficult and expensive for most amateurs. Many successful amateur operators settle for comparatively low power output, ranging from perhaps 50 to 100W in the lower-frequency bands, to milliwatts in the 'centimetre' bands, or even microwatts in the 'millimetre' bands. This is compensated for by using very high antenna gain and, as already mentioned, receivers with very low noise figures.

Since the first essential requirement of microwave construction is easy availability of designs and components, many leading microwave amateurs have launched small-quantity component sources or have designed and can supply either kits of parts for home construction or ready-made equipment to these designs. Most microwave equipment now uses printed circuit boards (PCBs) and surface-mount components. This avoids the use of the precision engineering usually associated with older, waveguide based designs and means that construction of microwave equipment is not restricted to the amateur who has his own mechanical workshop. Conventional tools can be used together with some fairly simple test equipment to construct some very sophisticated equipment that produces excellent results.

There are many examples of designs that can be purchased in kit form or ready built; suppliers are listed in the Bibliography at the end of this chapter. The most widely used designs in the UK come from Michael Kuhne (DB6NT) who has equipment for all bands up to 241GHz; most have been described in the pages of *Dubus Magazine* or the *Dubus Technik* publications.

Free and easy access to practical information is important to the microwave amateur enthusiast. The question most often asked by amateurs new to microwaves is "Where do I get reliable information and (possibly) help?" Many microwave designs have appeared in the amateur press, in published books or magazines and in the various national amateur radio societies' journals. Some of the more prolific or rewarding titles are given in the bibliography. In addition to these sources, obtaining up-to-date designs, component information and design tools is extremely easy using the internet. The number of suppliers of microwave components has mushroomed with the expansion of the mobile phone networks, so has the sophistication of the design tools available.

Many suppliers of design tools have student or 'Lite' versions of their software free to download from their websites. These generally have reduced functionality compared with the full versions of the software, which may cost several thousand pounds, but are more than adequate for most amateur use. A selection of website addresses for the more significant, current sites is given at the web page associated with this book (the address is given at the end of this chapter in the bibliography). Please note that website addresses (URLs) change quite frequently, if the links supplied do not lead to the expected web pages a quick search with one of the popular search engines, using the relevant key words, should find the new location.

The range of current amateur microwave allocations offers scope to try out all of the modes and techniques available to amateurs. All amateurs are encouraged to try out some of these which will help retain our allocations. The lowest microwave frequency amateur allocation, the so called 23cm band, (1240MHz to 1325MHz in the UK), can be regarded as the transition point from 'conventional' radio techniques and components to the 'special' microwave techniques and components to be reviewed here.

In the space of a single chapter it will only be possible to give a flavour of some of the practical techniques involved, by outlining a few representative designs for most of the bands currently used by amateurs. If you need more detail, there are plenty of pointers to other sources of information shown in the bibliography.

The microwave bands support a wide range of activities such as:

- All narrow band modes
- Amateur TV, including wide band colour transmission
- Moon bounce (EME)
- Amateur satellite operation
- Meteor scatter

Since a significant amount of amateur microwave interest centres on the use of narrow band modes to achieve long distance, weak signal communication, the majority of the designs outlined here will concentrate on such equipment. More details of components and techniques (including wide band modes) are available in other publications [3-5].

In some instances construction and alignment procedures are described in some detail, again to illustrate the techniques used by amateurs in the absence of elaborate or costly test equipment, such as microwave noise sources, power meters, frequency counters or spectrum analysers. Most of the designs described are capable of being home constructed without elaborate workshop facilities (most can be constructed using hand tools, a generous helping of patience and some basic knowledge and skills!) and aligned with quite ordinary test equipment such as matched loads, directional couplers, attenuators, detectors, multimeters and calibrated absorption wavemeters.

AMATEUR MICROWAVE ALLOCATIONS

Most countries in the world have amateur microwave allocations extending far into the millimetre wave region, ie above 30GHz. Many of these allocations are both 'common' and 'shared Secondary', ie they are similar in frequency in many countries but are shared with professional (in this case 'Primary') users who take precedence. Amateur usage must, therefore, be such that interference to Primary users is avoided and amateurs must be prepared to accept interference from the Primary services, especially in those parts of the spectrum designated as Industrial, Scientific and Medical (ISM) bands. The UK Amateur Service allocations are summarised in **Table 11.1** and the UK Amateur Satellite Service allocations are shown in **Table 11.2**.

All the familiar transmission modes are allowed under the terms of the amateur licence: in contrast to the lower frequency bands, most of the microwave bands are sufficiently wide to support such modes as full-definition fast-scan TV (FSTV) or very high speed data transmissions as well as the more conventional amateur narrow-band modes, such as CW, NBFM and SSB.

Many of the bands are so wide (even though they may be Secondary allocations) that it may be impracticable for amateurs to produce equipment, particularly receivers that cover a whole allocation without deterioration of performance over some part of the band. Most amateur operators do possess a high-performance multimode receiver (or transceiver) as part of their station equipment and this will frequently form the 'tuneable IF' for a microwave receiver or transverter: commonly used intermediate frequencies are 144-146MHz or 432-434MHz. either of which are spaced far enough away from the signal frequency to simplify the design of good image and local oscillator carrier sideband noise rejection filters. An intermediate fre-

Allocation (MHz)	Amateur Status	Preferred (alternative) narrow band segment
1,240 - 1,325	Secondary	1,296 - 1,298
2,310 - 2,450	Secondary	2,320 - 2,322
3,400 - 3,475	Secondary	3,400 - 3,402 (3,456 - 3,458)
5,650 - 5,680	Secondary	5,668 - 5,670
5,755 - 5,765	Secondary	5,760 - 5,762
5,820 - 5,850	Secondary	
10,000 - 10,125	Secondary	All modes
10,225 - 10,475	Secondary	10,368 - 10,370 (10,450 - 10-452)
10,475 - 10,500	Secondary	Space only
24,000 - 24,050	Primary	24,048 - 24,050
24,150 - 24,250	Secondary	24,192 - 24,194
47,000 - 47,200	Primary	47,088 centre of activity
75,500 - 76,000	Primary	75,976 centre of activity (until December 2006)
76,000 - 77,500	Secondary	
77,500 - 78,000	Primary	77,500 - 77,502 (after January 2007)
78,000 - 81,000	Secondary	
122,250 - 123,000	Secondary	
134,000 - 136,000	Primary	
136,000 - 141,000	Secondary	
142,000 - 144,000	Primary (until December 2006)	
241,000 - 248,000	Secondary	
248,000 - 250,000	Primary	

Table 11.1: UK Amateur Service Allocations at Spring 2005

Allocation (MHz)	Amateur Status	Comments
1,260 - 1,270	Secondary	ETS only
2,400 - 2,450	Secondary	ETS/STE. *(Note 1)*
5,650 - 5,568	Secondary	ETS only
5,830 - 5,850	Secondary	STE only
10,475 - 10,500	Secondary	ETS/STE
24,000 - 24,050	Primary	ETS/STE
47,000 - 47,200	Primary	ETS/STE
75,500 - 76,000	Primary	ETS/STE
76,000 - 77,500	Secondary	ETS/STE
77,500 - 78,000	Primary	ETS/STE
79,000 - 81,000	Secondary	ETS/STE
122,250 - 123,000	Secondary	ETS/STE
134,000 - 136,000	Primary	ETS/STE
136,000 - 141,000	Secondary	ETS/STE
142,000 - 144,000	Primary	
	(until Dec 2006)	ETS/STE
241,000 - 248,000	Secondary	ETS/STE
248,000 - 250,000	Primary	ETS/STE

ETS = Earth to space, STE = Space to earth,
ISM = Industrial, scientific and medical applications

Note 1: Users must accept interference from ISM users

Table 11.2: UK Amateur Satellite Service Allocations at Spring 2005

Table 11.3: Some harmonic relationships for the microwave bands

Starting frequency	Multiplication	Output frequency
144MHz	x3	432MHz
	x9	1296MHz
	x16	2304MHz
	x24	3456MHz
	x46	5760MHz
	x72	10,368MHz
	x108	24,192MHz
432MHz	x3	1296MHz
	x8	3456MHz
	x24	10,368MHz
	x56	24,192MHz
1152MHz	+144*	1296MHz
	x2	2304MHz
	x3	3456MHz
	x5	5760MHz
	x9	10,368MHz
	x21	24,192MHz

* Note: additive mixing, not multiplication

quency of 1296-1298MHz is often used for the millimetre bands, ie 24GHz and higher.

There are 'preferred' sub-bands in virtually all of the amateur allocations where the majority of narrow band (especially weak signal DX) operation takes place. Typically 2MHz wide sub bands, often harmonically related to 144MHz as shown in **Table 11.3**, were originally adopted for this purpose.

Some of these harmonic relationships are no longer universally available or usable because the lower microwave bands are rapidly filling up with Primary user applications. Indeed, the position is changing particularly rapidly at the time of publication and the reader should refer to the latest ITU/IARU band plans (see RSGB web site) to get up-to-date information on current amateur usage, even though the current narrow band segments are indicated in Table 11.1.

MODERN MICROWAVE COMPONENTS AND CONSTRUCTION TECHNIQUES

Static Precautions

Some types of microwave components, for example Schottky diodes (mixers and detectors), microwave bipolar transistors and GaAsFETs can be damaged or destroyed by static charges induced by handling, and thus certain precautions should be taken to minimise the risk of damage.

Such sensitive devices are delivered in foil lined, sealed envelopes in conductive (carbon filled) foam plastic or wrapped in metal foil. The first precaution to be taken is to leave the device in its wrapping until actually used. The second precaution is to ensure that the device is always the last component to be soldered in place in the circuit. Once in circuit the risk is minimised since other components associated with the device will usually provide a 'leakage' path of low impedance to earth that will give protection against static build up.

Before handling such devices the constructor should be aware of the usual sources of static. Walking across nylon or polyester carpets and the wearing of clothes made from the same materials are potent sources of static, especially under cold dry conditions. The body may carry static to a potential of several thousand volts although much lower leakage potentials existing on improperly earthed mains voltage soldering irons are still sufficient to cause damage. Some precautions are listed below:

- Avoid walking across synthetic fibre carpets immediately before handling sensitive devices.
- Avoid wearing clothes of similar materials.
- Ensure that the soldering iron is properly earthed whilst it is connected to its power supply. This is a common sense precaution in any case. Preferably use a low voltage soldering iron.
- Use a pair of crocodile clips and a flexible jumper wire to connect the body of the soldering iron to the earth plane of the equipment into which the device is being soldered.
- If the component lead configuration allows (and the usual flat pack will), place a small metal washer over the device before removing it from its packing in such a position that all leads are shorted together before and during handling. Alternatively, it might be possible to use a small piece of aluminium foil to perform the same function, removing the foil once the device has been soldered in place.
- A useful precaution that will minimise the risk of heat damage, rather than static damage, is to ensure that the surfaces to be soldered are very clean and preferably pre tinned.
- Immediately before handling the device, touch the earth plane of the equipment and the protective foil to ensure that both are essentially at the same potential.
- Place the device in position, handling as little as possible.
- Disconnect the soldering iron from its power supply and quickly solder the device in place. It may be necessary to repeat some of the operations if the soldering iron has little heat capacity.

Finally, when assembling items of equipment to form a complete operating system, for instance when installing a masthead pre-amplifier and associated transmit/receive switching, it is important to keep leads carrying supply voltages to the sensitive devices well away from other leads carrying appreciable RF levels

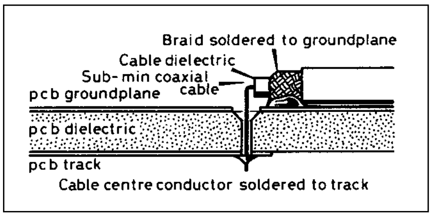


Fig 11.1: Correct coaxial cable connection technique

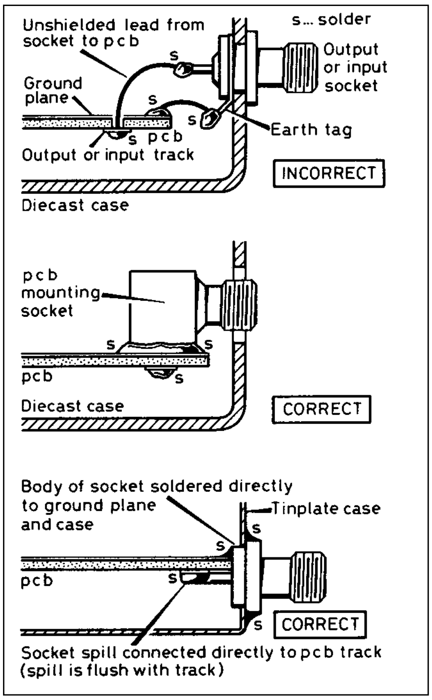


Fig 11.2: Correct and incorrect methods of connecting coaxial sockets to microwave PCBs

or those leads which might carry voltage transients arising from inductive (relay) switching. Such supply lines should be well screened and decoupled in any case, but physical separation can minimise pick up, thus making the task of decoupling easier.

PCB Materials

A printed circuit board (PCB) in a microwave design is not like the one you find in HF equipment. The printed tracks are an integral part of the circuit not just there to interconnect the components. The tracks are microstrip transmission lines, used to form matching circuits, tuned circuits and filters. The design process takes into account the base material of the board used, the thickness of the copper deposit and the dimensions of the track etched. It is therefore important to use the material specified in the design that you are using otherwise the circuit may not perform as expected. Conventional Epoxy/glass PCB board is usable, with care, up to about 3GHz. Most designs use special PCB materials such as Rogers RO 4003 or RT/duroid 5870.

If possible try to use PCBs that have been professionally produced using the correct PCB material and good artwork. If that is not possible you can produce your own PCBs using conventional etching techniques. Microwave PCBs always have an earth plane on one side and etched tracks on the other side; the earth connections from one side to the other are important and are formed by plating through the appropriate holes on professionally manufactured boards. For home made PCBs the best technique is to use small rivets to make these connections, they are fitted and soldered in place forming a good, low inductance, interconnection.

Earthing and Interconnections

Earthing is a very important topic for the microwave constructor. As mentioned above, the earth connections from one side of a PCB to the other must be as good as possible to reduce the effects of stray inductance. The same is true for all other earth connections; they must be as short and solid as possible. It is important to house finished PCBs properly in order to screen the circuitry from stray pick-up and provide a good earth. Small die-cast boxes can be used but they are quite expensive and difficult to use. Piper Communications stock tinplate boxes of various sizes that are an acceptable substitute for the die-cast boxes. These are widely used in Europe for housing such PCBs and are much less expensive. They consist of two L-shaped side pieces, and top and bottom lids. It is intended that the PCB be put into the box, joining the edges of the ground plane to the sides of the box.

To interconnect circuits it is necessary to use RF connectors. N-type connectors are too large and BNC connectors can be unreliable. UHF sockets *must not* be used (they are absolutely useless at UHF, despite their name, and significantly mismatched even at 144MHz) and the only really reliable types are SMA, SMB or SMC, all of which are expensive. You might like to consider taking the outputs away by directly connecting miniature 50Ω coaxial cable as shown in **Fig 11.1**. Do not take the output away as shown in **Fig 11.2**, this is disas-

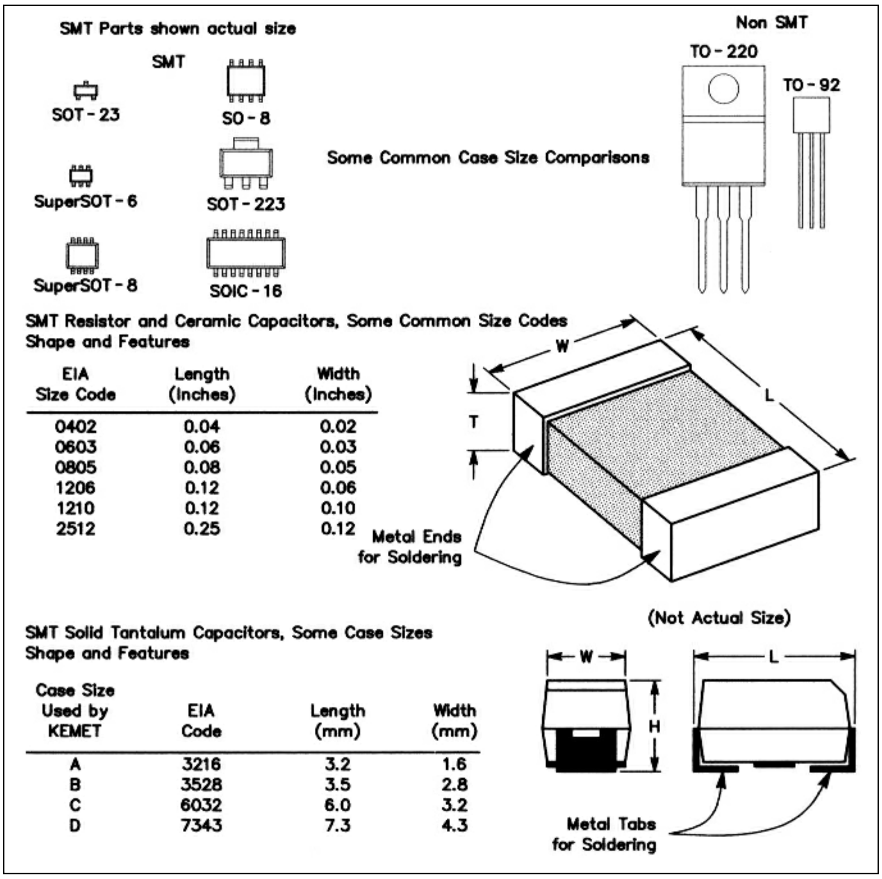


Fig 11.3: Size comparison of some surface mount devices and their dimensions

Product Family	Min Freq (GHz)	Max Freq. (GHz)	Pout (dBm)	Gain (dB)	Noise Figure (dB)	Isolation (dB)	Bias current (mA)	@ Vdd (V)
MGA-641	1.2	10	12	12	7.5	35	50	10
MGA-725	0.1	6	17	14	1.4	23	20	3
MGA-815	0.5	6	14	12	2.8	24	42	3
MGA-825	0.5	6	17	13	2.2	22	84	3
MGA-835	0.5	6	22	22	6	32	142	3
MGA-855	0.5	8	9	18	1.6	41	15	3
MGA-865	0.8	8	6	20	2	46	16	5

Table 11.4: Data for Agilent MMICs. (Copyright Agilent Technologies. Reproduced with permission)

Model, Mini Circuits	Equivalent MAR/MAV	Equivalent Avantek	Alphanumeric	Dot Colour
MAR-1	MAV-1	MSA0185	A01	Brown
MAR-2	MAV-2	MSA0285	A02	Red
MAR-3	MAV-3	MSA0385	A03	Orange
MAR-4	MAV-4	MSA0485	A04	Yellow
MAR-6		MSA0685	A06	White
		MSA0735		
MAR-7			A07	Violet
MAR-8		MSA0885	A08	Blue
		MSA0835		
MAV-1	MAR-1	MSA0104	1	
MAV-2	MAR-2	MSA0204	2	
MAV-3	MAR-3	MSA0304	3	
MAV-4	MAR-4	MSA0404	4	
		MSA0504	5	
		MSA0604	6	
		MSA0704	7	
		MSA0804	8	
MAV-11		MSAO 1 104	A	
ERA-1			E1	
ERA-2			E2	
ERA-3			E3	
ERA-4			E4	
ERA-5			E5	
ERA-6			E6	

	Gain typical dB at Frequency (GHz)						Max power out	Noise	IP3 dBm
Model	0.1	0.5	1.0	2.0	3.0	4.0	6.0	1dB comp. 1GHz	figure	
MAR-1	18.5	17.5	15.5					+1.5dBm	5.5	+14.0
MAR-2	12.5	12.3	12.0	11.0				+4.5dBm	6.6	+17.0
MAR-3	12.5	12.2	12.0	11.5				+10dBm	6.0	+23.0
MAR-4	8.3	8.2	8.0					+12.5dBm	6.5	+25.5
MAR-6	20.0	18.5	16.0	11.0				+2.0dBm	3.0	+14.5
MAR-7	13.5	13.1	12.5	11.0				+5.5dBm	5.0	+19.0
MAR-8	32.5	28.0	22.5					+12.5DbM	3.3	+27.0
MAV-11	12.7	12.0	10.5					+17.5dBm	3.6	+30.0
ERA-1				11.6	11.2		10.5	+13dBm (2GHz)	7.0	+26.0
ERA-2	16.0			14.9	13.9		11.8	+14dBm (2GHz)	6.0	+27.0
ERA-3	22.2			20.2	18.2			+11dBm (2GHz)	4.5	+23.0
ERA-4	13.8		14.0	13.9	13.9	13.4		+19.1dBm	5.2	+36.0
ERA-5	20.4		20.0	19.0	17.6	15.8		+19.6dBm	4.0	+36.0
ERA-6	11.1		11.1	11.8	11.5	11.3		+18.5dBm	8.4	+36.5

Application	Model	Application	Model
High frequency gain	ERA-1, usable to 10GHz	Stable high gain	MAR-1, ERA-3
Low noise amplifier	MAR-6, MAR-8, MAV-11	Medium output	MAV-11, MAR-3, MAR-4
Medium noise	ERA-3, ERA-5	High output	MAV-11, ERA-4, ERA-5
High dynamic range	MAV-11	Multiplier	ERA-3 (clean harmonics)

Table 11.5: MMIC data, an extract from data on the Mini Circuits web site

trous as it will almost certainly cause mismatch, stray inductive losses and may detune the output lines so that they will not resonate properly.

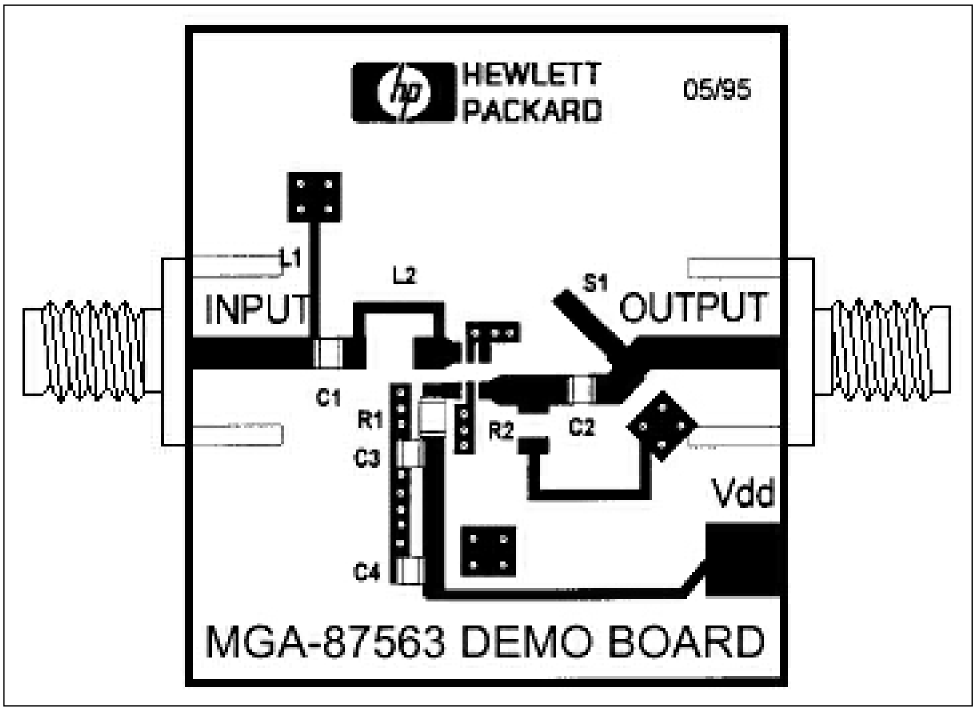


Fig 11.4: Agilent PCB for 2,400MHz LNA. For more details see [6]

Surface Mount Components

Surface mount components are ideally suited for microwave construction because there are no leads to introduce extra inductance in series with the actual component being used. This means that circuit performance can be reproduced more easily. It does introduce a new construction challenge for the newcomer to microwaves. Dealing with tiny components and soldering them in place can seem a daunting task but after some practice it is easy.

Fig 11.3 shows some of the more common SMD component sizes, obviously some special tools are needed to cope with these small components. The essential tools are:

- An illuminated magnifying glass. For instance, a bench-mounted magnifier with a five inch glass.
- A low power temperature-controlled soldering iron. This should have its tip earthed to reduce problems with static. A fine conical tip is best.
- Thin flux cored solder, preferably 26SWG (0.5mm). Larger diameter solder is difficult to use because it floods the solder pads and tends to create short circuits between solder pads.
- De-soldering braid, for use when too much solder has been applied.
- A flux pen to apply a small amount of flux before components are mounted.
- A good pair of non-magnetic tweezers.
- A PCB frame or some other method to hold your printed circuit board whilst soldering. If you don't hold the PCB down you will run out of hands to hold everything else!

Before you start to mount components, the PCB should be lightly tinned, just enough solder for the solder to flow onto the component but not too much otherwise short-circuits will be a problem. As with wired components the assembly sequence should start with the low value parts such as resistors and capacitors.

Position the component in position and apply heat just long enough for the solder to flow; you will need to hold the component in place otherwise it will move as the solder flows and may well land up on the end of your soldering iron rather than on the PCB!

Multi-leaded devices, like ICs, should be tacked in place by soldering leads at opposite corners and then flowing solder to all the other legs. Some of the latest ICs have an earthing pad underneath which makes life very difficult. The only successful technique that I have heard of is to mount such components first by heating the complete PCB over an electric cooker to flow the solder under the IC.

Monolithic Microwave Integrated Circuit (MMIC) Amplifiers

MMICs are now widely used in amateur radio designs and are available from several manufacturers including Mini-Circuits and Agilent (Hewlett Packard) at a price that makes them very attractive for many applications.

Keeping track of the devices that are available can be a problem; two useful sources of information are the Agilent (Hewlett Packard) website [6] and the Mini-Kits website [7]. Some useful information from these suppliers is reproduced in **Tables 11.4 and 11.5**.

The Agilent website has a wide range of application notes to show how to use their MMICs. They also supply evaluation boards so that the complete circuit can be built and tested. One such application note is for a 2,400MHz LNA, designed to provide an optimum noise match from 2,400 through 2,500MHz, making it useful for applications that operate in the 2,400 to

Frequency (MHz)	Gain (dB)	Noise Figure (dB)
1,700	10.4	2.60
1,800	13.8	2.57
1,900	11.3	2.45
2,000	11.9	2.38
2,100	13.3	2.06
2,200	12.8	2.02
2,300	12.9	2.12
2,400	11.5	2.05
2,500	11.5	2.14
2,600	10.5	2.25
2,700	10.9	2.29
2,800	10.3	2.33
2,900	9.8	2.35
3,000	9.6	3.42

Table 11.6: 2,400MHz LNA noise figure and gain with Vd = 3V

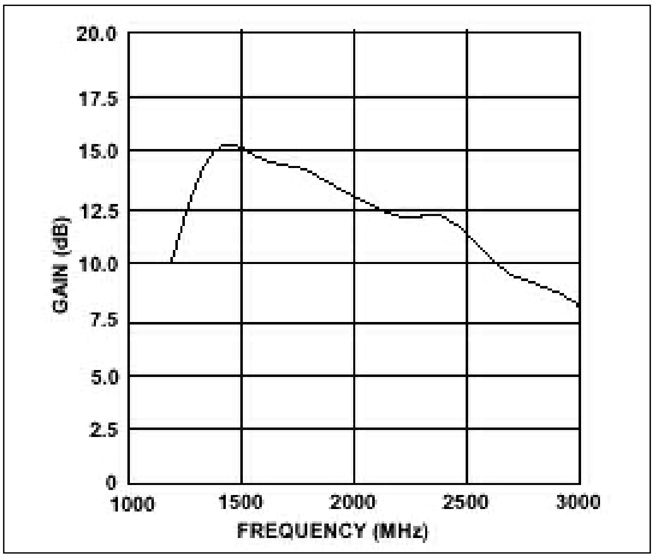


Fig 11.5: 2,400MHz LNA associated gain at maximum noise figure. With Vd = 5V

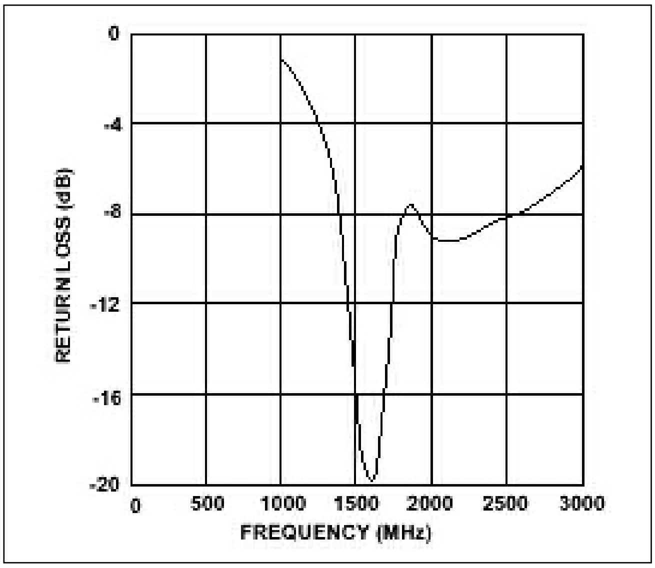


Fig 11.7: 2,400MHz LNA input return loss. With Vd = 5V

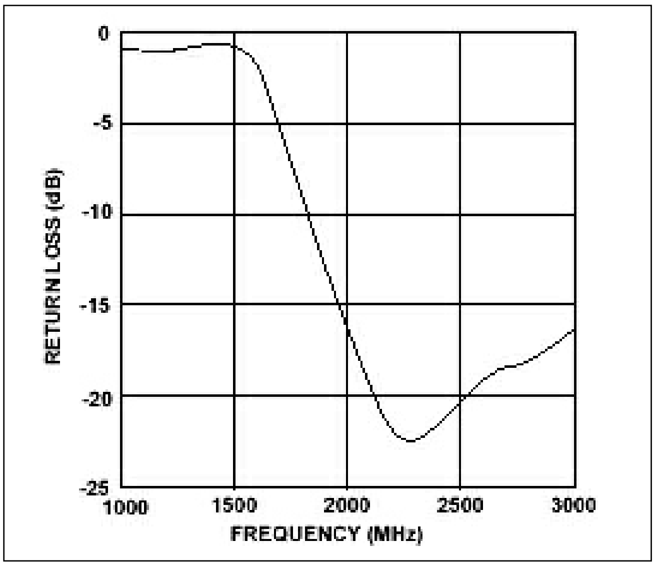


Fig 11.6: 2,400MHz LNA output return loss. With Vd = 5V

2,483MHz ISM band. The component labels appearing in the following paragraphs refer to positions shown in **Fig 11.4**. The input match consists of a shunt inductor at L1 and a series inductor at L2. Both of these inductors use the tracks as originally etched on the circuit board without modification. The output is matched with a simple shunt open circuited stub (S1) on the output 50-ohm microstripline. 22pF capacitors were used for both the input (C1) and output (C2) blocking capacitors.

A 16-ohm chip resistor placed at R1 and decoupled by a 100pF capacitor at C3 provides a proper termination for the device power terminal. An additional bypass capacitor (100 to 1000pF) placed further down the power supply line at location C4 may be required to further decouple the supply terminal, especially if this stage is to be cascaded with an additional one. Proper decoupling of device VCC terminals of cascaded amplifier stages is required if stable operation is to be obtained. If desired, a 50-ohm resistor placed at R2 will provide low frequency loading of the device. This termination reduces low frequency gain and enhances low frequency stability. The MGA-87563 has three ground leads, all of which need to be well grounded for proper RF performance. This can be especially critical at 2.4GHz where common lead inductance can significantly decrease gain.

The performance of the LNA as measured on the HP 8970 Noise Figure Meter is shown in **Table 6**. At 2.4GHz, the loss of the FR-4/G-10 epoxy glass material can add several tenths of a dB to noise figure and lower gain by double the amount. The swept plots, **Figs 5 – 7** were taken on a scalar analyser and show the performance of the amplifier.

Another simple design that is usable up to 5.7GHz, using MMIC amplifiers as gain blocks, is shown here. Two designs are presented, a two and a three stage amplifier using similar circuitry [8]. The first amplifier is shown in **Fig 11.8**. It uses two MSA0835 devices, in cascade, with a gain that varies from 27dB at 2GHz down to 7dB at 6GHz. Output power is only a few milliwatts at the upper end of the range but the amplifier is not intended to form the output stage of a transmitter. This function is better provided by a

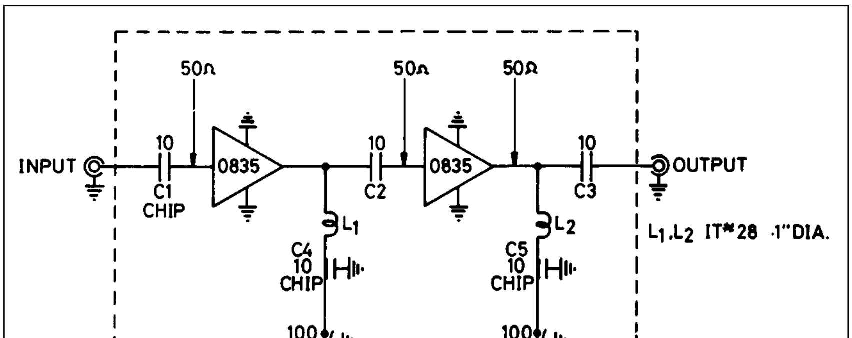


Fig 11.8: Circuit of the two stage MSA0835 wide band amplifier

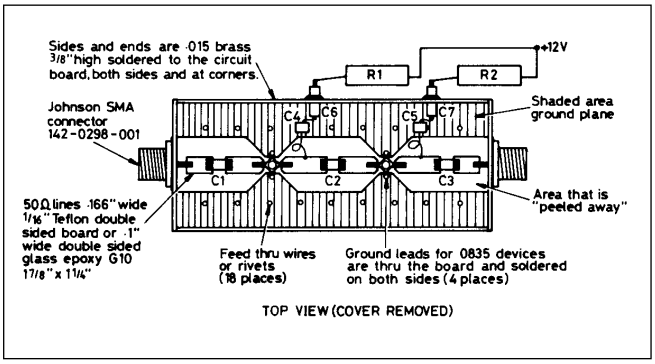


Fig 11.9: Layout of components in the two stage MSA0835 wide band amplifier

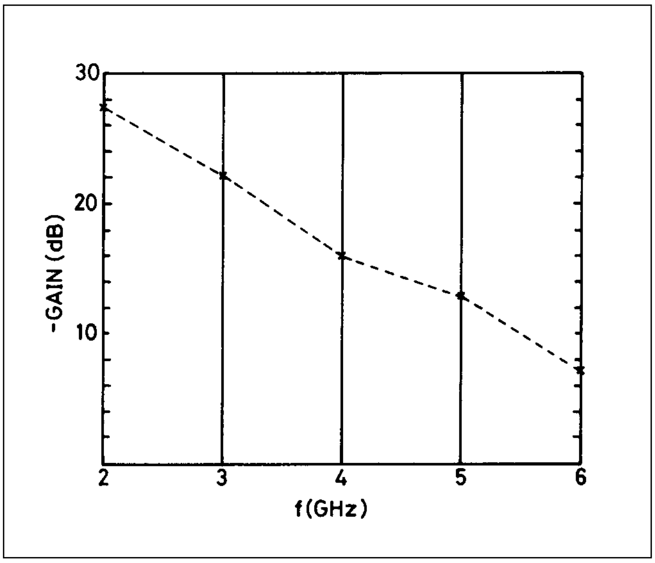


Fig 11.10: The frequency response of the two stage MSA0835 wide band amplifier

GaAsFET power amplifier above 3GHz. and by a bipolar amplifier below 3GHz.

Construction of the two stage amplifier is straightforward using a simple microstrip line track of 50-ohm impedance, on either glass fibre or PTFE double sided board, as shown in Fig 11.9. PTFE board material is preferred if operation above 3GHz is required.

Small l0pF ceramic chip capacitors prevent the bias supply to the MSA0835 devices being shorted out by the source and load. Another capacitor prevents the collector supply to the first stage shorting out the second stage base bias. Simple resistor current limiting from the amplifier 12V supply provides bias. Single-turn chokes in the collector leads of the MMICs prevent the bias resistors shunting the output signal to ground. The frequency response of the two stage amplifier is shown in Fig 11.10.

The three stage amplifier shown in Fig 11.11 offers more gain than the two stage amplifier right across the frequency range. Construction is similar to the latter but uses a slightly longer board as shown in Fig 11.12. Its frequency response is shown in Fig 11.13.

More output can be achieved in this design if a Siemens CGY40 MMIC is used in the output stage in preference to the MSA0835. Gain will be slightly less and the maximum frequency of operation will fall to 3.4GHz. With this modification the output power at 3.4GHz can be as high as 50mW. This power level should be satisfactory for short links or longer line-of-sight paths.

These amplifiers have very high gain and are only conditionally stable when used between good 50-ohm load and source impedances. In practice, the insertion of a 3 or 6dB attenuator at the input and output of the amplifier should ensure stable operation, although in many cases the attenuators will not be required.

Replacing the MSA0835 MMIC with the lower gain 0735 device results in unconditional stability and much less gain variation across the frequency range. The penalty for

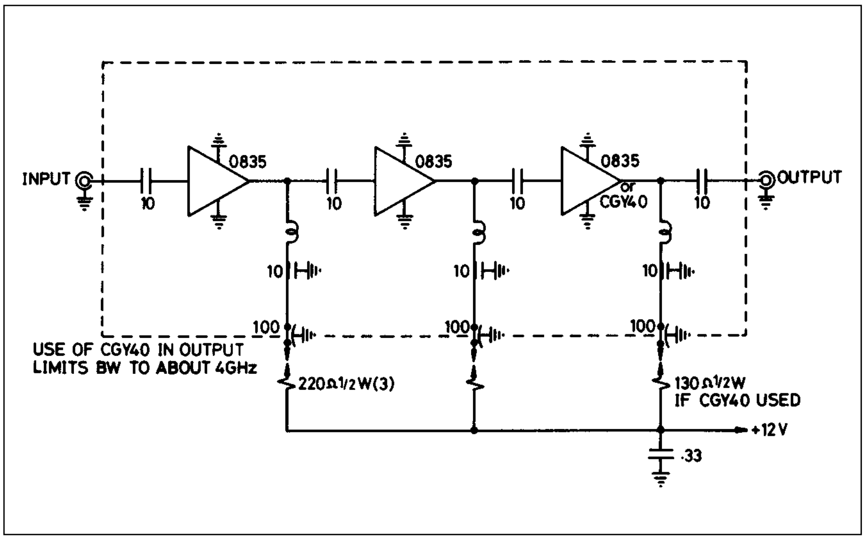


Fig 11.11: Circuit of the three stage MSA0835 wide band amplifier

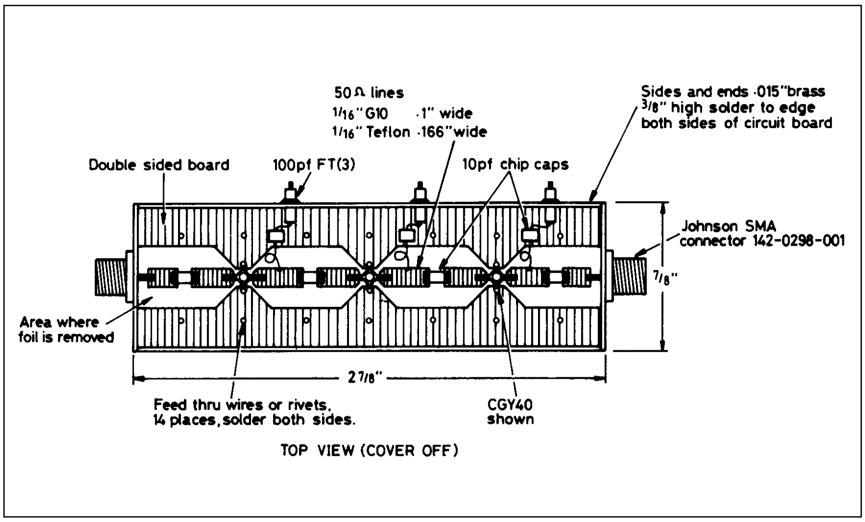


Fig 11.12: Layout of the components in the three stage MSA0835 wide band amplifier

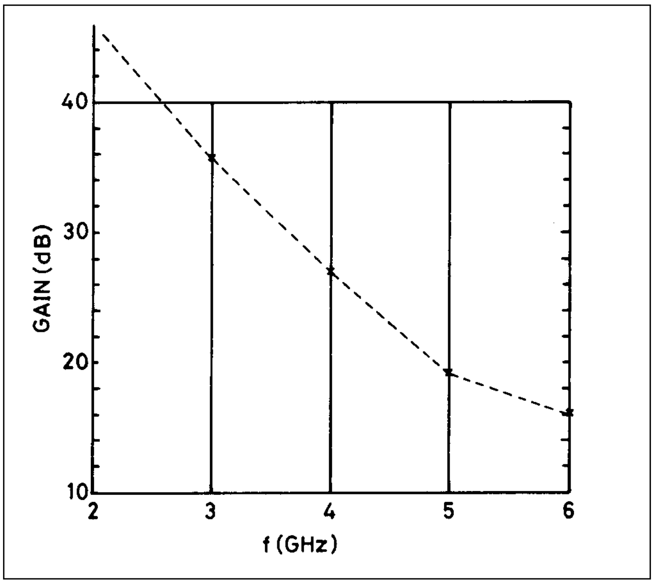


Fig 11.13: The frequency response of the three stage MSA0835 wide band amplifier

this is lower gain and output power. About 32dB gain at 2GHz, falling to 22dB at 3.4GHz can be expected with this design.

It cannot be stressed enough how important the grounding is around the emitter leads of the MMIC devices on this board. In the absence of a plated through-hole board, through-board wire links must be provided where shown, and especially under the emitter lead connections to the top ground plane of the board. Soldering copper foil along the edges of the board is just not sufficient to ensure good ground integrity and the amplifier will self oscillate if links are not used.

MICROWAVE LOCAL OSCILLATOR SOURCES

Early experimenters on the amateur microwave bands used wide band modes, so frequency accuracy was not that important. Many designs in the 1960s used free running oscillators at the operating frequency using Gunn diodes. Then improvements were made and the free running oscillators were locked to a known stable source. The quest for communications using narrow band modes, such as SSB, necessitated a different approach. A stable crystal controlled oscillator, followed by a multiplier chain to produce the required local oscillator frequency became commonplace. The local oscillator is mixed with the output of a commercial transceiver to produce the required signal on the amateur band to be used (see **Fig 11.14**). The local oscillator and mixer are usually combined into a single unit – a transverter. This still remains the technique of choice used by serious microwave operators. Any 144MHz transceiver can be used but the IC202 is still often used because of its clean and stable output. The main design criteria for a good local oscillator source for microwave use are:

- Good short term frequency stability. Short term frequency variations may be caused by such things as the type of crystal used, the type of oscillator circuit, stability of supply voltages and temperature changes. Small changes in the frequency of the crystal oscillator are multiplied, eg if a local oscillator is used to generate an output on the 10GHz band, a 106.5MHz crystal frequency will be multiplied by 96 to give a local oscillator frequency of 10,244MHz. Thus a change of 25Hz in crystal frequency will change the frequency at 10GHz by about 2.5kHz.
- Good long term frequency stability. Long term frequency variations may be caused by ageing of the crystal used. Also crystals suffer from a hysteresis effect that causes them to operate on a slightly different frequency each time they are started.
- Good signal purity. This is governed by the design of the oscillator. It is important to keep phase noise and spurious outputs of the crystal oscillator to a minimum because the multiplier chain magnifies these.

Other techniques, such as Phase Locked Loops (PLL) and Direct Digital Synthesisers (DDS) are used to generate the local oscillator signal, but these can suffer from poor signal purity.

A High Quality Microwave Source for 1.0 to 1.3GHz

This high quality microwave source, with output in the range 1.0 to 1.3GHz, was originally designed by G4DDK [9] to provide two +10dBm (10mW) outputs at 1152MHz for use in a 1296MHz transverter with a 144MHz IF. It is known as the G4DDK-001 microwave source. Several modified versions have appeared since the original design, but this one remains the 'standard'. Later work showed it was possible to use the same board, with a suitable crystal, anywhere in the range 1000 to 1400MHz with only minor changes in component values and output spectral purity. Versions of the board have since been produced to provide outputs between 700 and 1500MHz. At these two extremes it has been found necessary to change the length of some of the tuned microstrip lines in order to maintain resonance. PCBs are available from the Microwave Component Service [10].

The circuit of the oscillator source is shown in **Fig 11.15** and the component values in **Table 11.7**. The crystal oscillator section uses the Butler design. In this design the oscillator operates with a crystal frequency between 90-110MHz, although by changing the value of C3 the circuit can be made to operate reliably between about 84-120MHz. Operation outside this range may require changes in the value of other components. A 9V integrated circuit regulator (78L09) stabilises the supply to the oscillator and limiter stages.

The third stage is a times two multiplier. Input, in the frequency range 250-330MHz, is taken from the high imped-

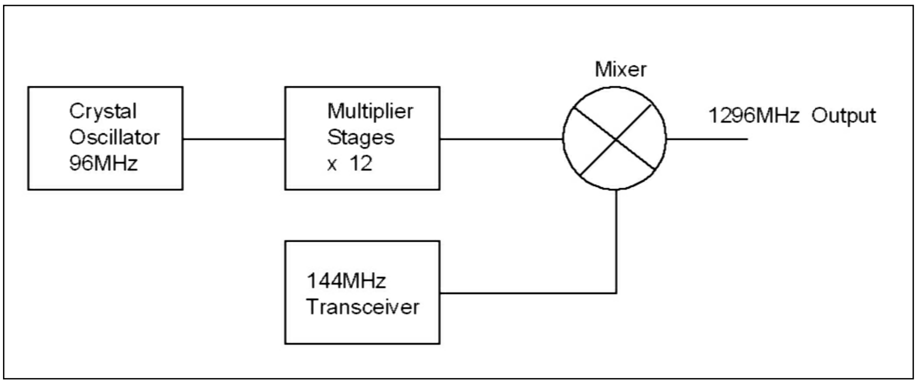


Fig 11.14: Local oscillator mixed with transceiver to give microwave output

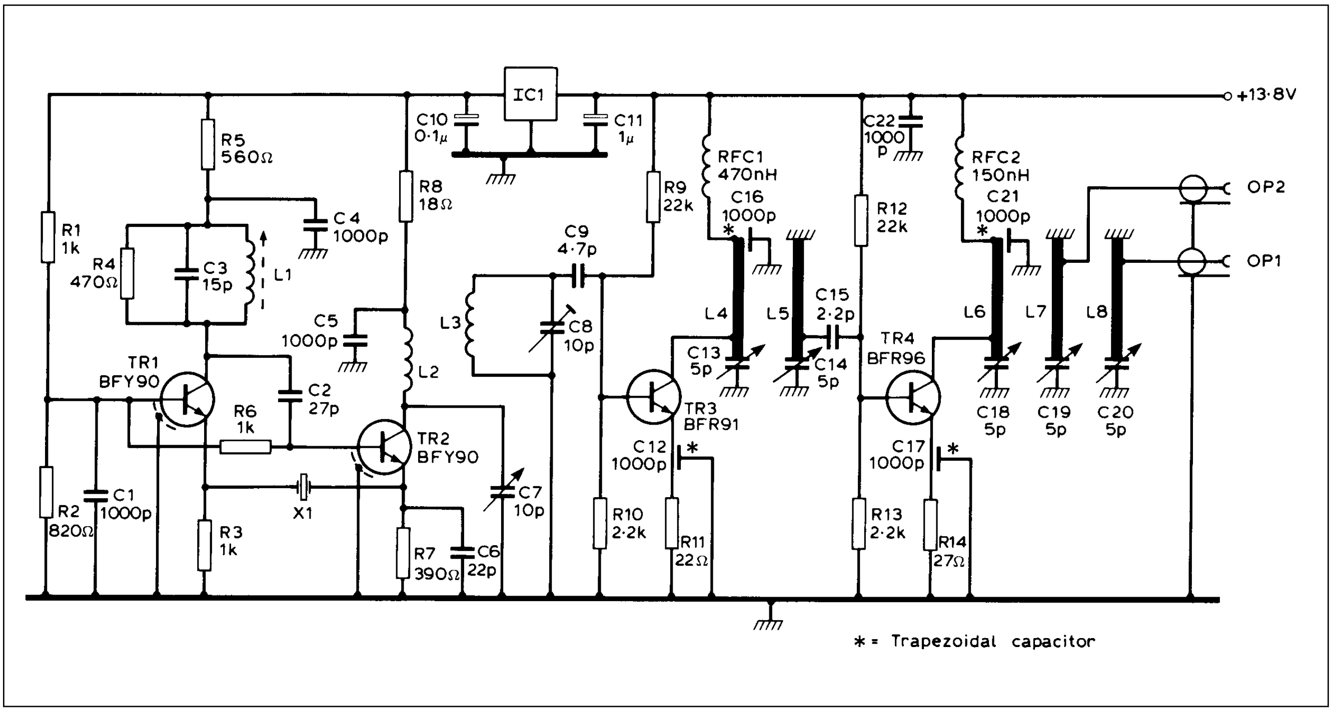


Fig 11.15; Schematic diagram of the microwave source G4DDK-001

ance end of the tuned circuit formed by L3 and C8 via C9. The coupled microstripline tuned circuits L4/C13 and L5/C14 tune the output of TR3 to the range 500-660MHz.

A high impedance tap on L5 couples the doubled signal to the base of the final multiplier stage, TR4. The output of this stage is tuned by three coupled tuned microstriplines; L6/C18, L7/C19 and L8/C20.

The tuning range of this filter is highly dependent upon the type of trimmer capacitors chosen for C18, 19 and 20. By using the specified capacitors, the filter will tune from 980 to 1400MHz, encompassing the range of second harmonics from the previous stage (1000 to 1320MHz).

Three stages of filtering are used to achieve a very clean spectrum at output l. A slightly less clean output is available at output 2, since this output is taken from the second stage of the filter. Even so, the output here is more than adequate for use in a receive converter with output 1 being reserved for the transmit converter. Each output is at a level of +10dBm (10mW), but if only one output is required then this should be taken from output 1 and the track to output 2 cut where it leaves L7. A single output of +13dBm (20mW) should be available in this configuration. Fig 11.16(a) and (b) are analyser plots of the outputs of the board.

PCB artwork for the UHF source is shown full size in Fig 11.17 (in Appendix

Resistors

R1, 3, 6	1k0	R9,12	22k0
R2	820R	R10, 13	2k2
R4	470R	R11	22R
R5	560R	R14	27R
R7	390R	0.25W miniature carbon film or metal film	
R8	18R		

Capacitors

C1,4,5,22	1000pF high-K ceramic plate, e.g. Philips 629 series
C12, 16, 17, 21	1nF trapezoidal capacitor from Microwave Component Service or Piper
C10	0.1μF tantalum bead, 16V working
C11	1μF tantalum bead, 16V working
C3	15pF low-K ceramic plate, e.g. Philips 632 Series
C7, 8	5mm trimmer, 10pF maximum
C13,14	5mm trimmer, 5pF maximum
C18.19,20	5mm trimmer, 5pF maximum. Must be able to reach 0.9pF minimum, e.g. SKY (green) or Murata TZ03 (black)

Inductors

L1	TOKO S18 5.5t (green) with aluminium core
L2, 3	3t of 22swg tinned or enamelled copper wire, 3mm ID. Turns spaced 1 wire diameter, height of coil 5mm above board.
L4-8	Printed on PCB
RFC1	470nH
RFC2	150nH

Semiconductors

TR1, 2	BFY90, available from Bonex, Piper etc.
TR3	BFR91A, available from Bonex, Piper etc
TR4	BFR96, available from Bonex, Piper etc
IC1	78L08, Piper, STC Components etc

Miscellaneous

X1	5th overtone crystal in HC18/U case. Frequency of crystal = Fout/12

Table 11.7: Component list for G4DDK-001 oscillator/multiplier

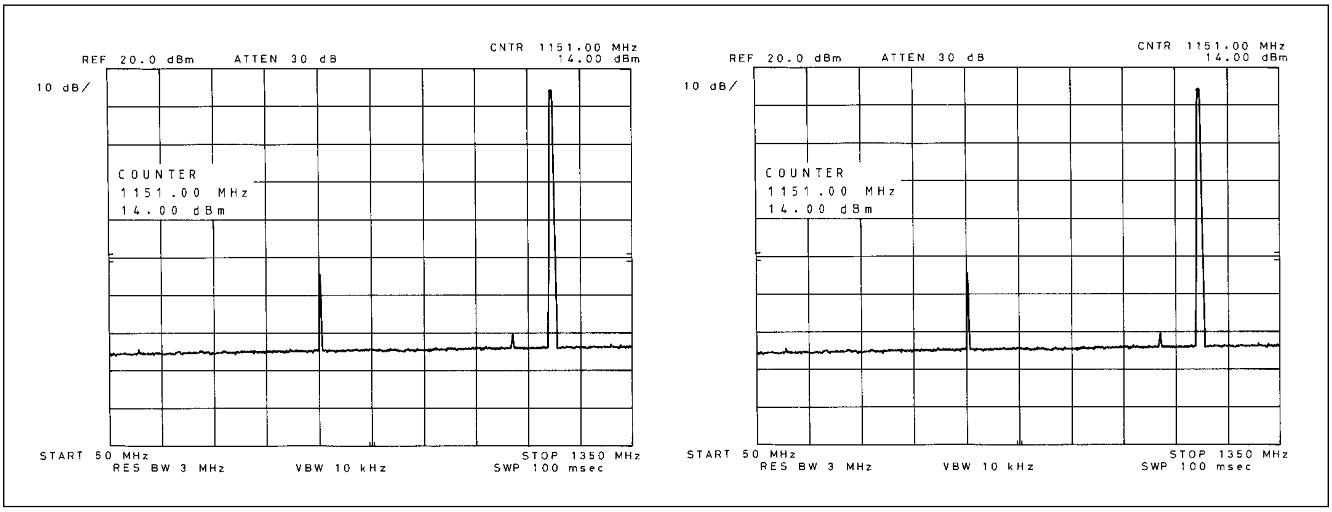


Fig 11.16: A spectrum analyser plot of the output from the microwave source G4DDK-001. (a) Output 1, output 2 terminated in 50Ω. (b) Output 2, output 1 terminated in 50Ω.. If output 2 is not required, cut the track near the line. Output 1 should then look like (a), but be at +13dBm

B). **Fig 11.18** is a photograph of a finished unit. Construction is generally straightforward, with the trapezoidal capacitors being carefully soldered into the slots in the board as shown in the component overlay diagram **Fig 11.19**. Capacitors C18, 19 and 20 must be miniature 5mm diameter types such as SKY, Murata or Oxley. The use of larger 7mm trimmers such as the popular Dau or Philips types will inevitably lead to tuning problems. The circuit was designed to take the small types. This also applies to C7, 8, 13 and 14, where overcoupling and consequently tuning problems can also occur.

Alignment is straightforward. Connect DC power and check that the current drawn is no more than about 150mA. If significantly more, check for short-circuits or wrongly placed components.

When all is well, proceed with the alignment. Place an absorption wavemeter pick-up coil close to L1 and tune to the crystal frequency. A strong reading should be indicated on the meter. Peak the indication by turning the core of L1. Turn the oscillator on and off to check that it re-starts satisfactorily. If it doesn't, turn the core of L1 about a quarter turn and try again. Too close coupling of the wavemeter (ie too much power absorption) may also inhibit easy re-starting, so use the loosest coupling possible to give adequate indication.

Set an analogue (moving coil) multimeter to the 2.5V range (or nearest equivalent) and measure the voltage across R11. This should be no more than a few hundred millivolts. Peak the reading by tuning C7 and C8. Confirm that the frequency selected is three times the crystal frequency by placing the coil of the wavemeter close to L3. The reason for using an analogue meter is that digital meters do not measure in real time and therefore tend to show what you have just done, rather than what you are doing, whilst adjusting the

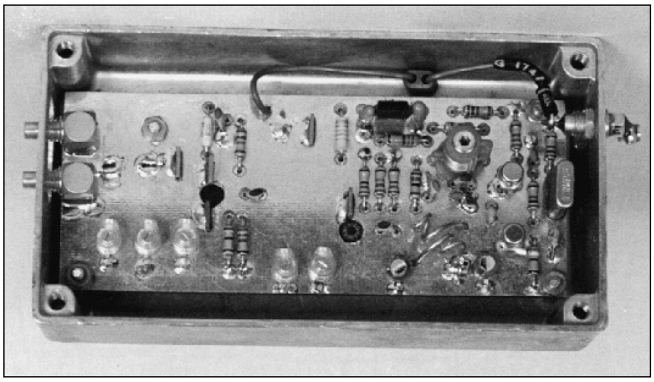


Fig 11.18: Photograph of the G4DDK-001 microwave source in the recommended size diecast box. The picture shows an early version of the source. L1 and L2 were laid out slightly differently in later versions

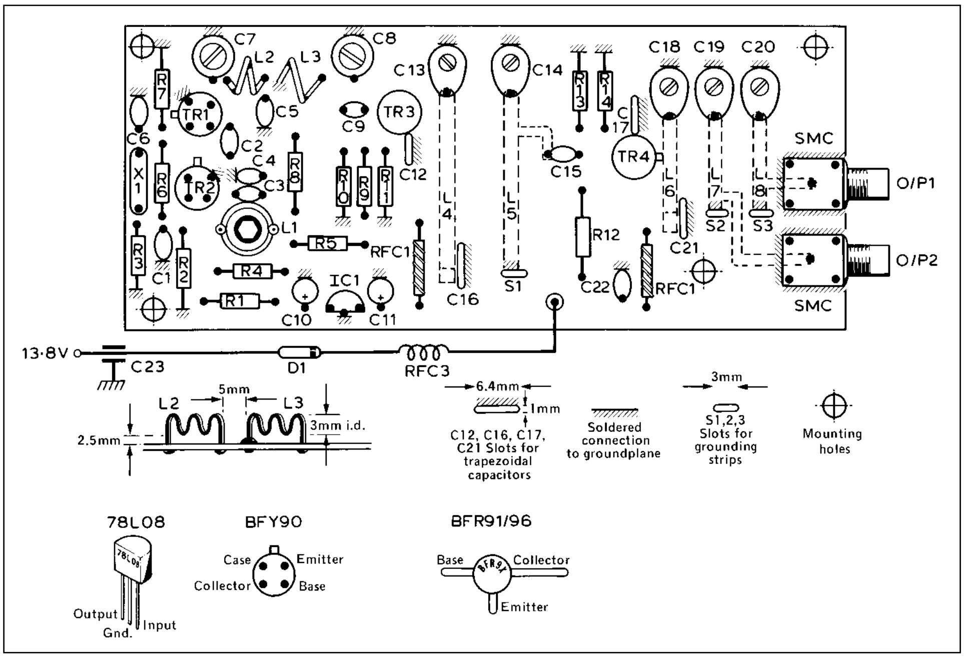


Fig 11.19: Component layout diagram of the G4DDK-001 microwave source

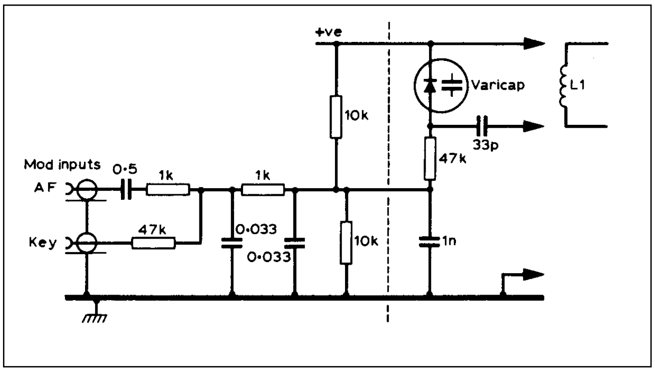


Fig 11.20: A simple FM/PSK modulator circuit for the G4DDK-001 microwave source

trimmers: indeed, if a digital meter is used, it is possible to miss the increase in measured voltage which occurs as each circuit is brought to resonance - the tuning is sharp!

Transfer the meter leads across R14 and peak the reading by tuning C13 and C14. Again check that the correct harmonic (twice the preceding stage) has been selected by using the wavemeter.

Finally, connect a low power wattmeter (+10 to +20dBm full scale, ie 10 to 100mW) to the output and tune C18, 19 and 20 for a maximum reading.

By using the wavemeter, confirm that the correct harmonic (now the output frequency and twice the preceding stage) has been selected. It may now be necessary to go back and slightly re-peak the trimmers for an absolute maximum reading at the final output frequency.

Final setting of the frequency of the crystal oscillator can now be done by either using a known high accuracy frequency counter or by connecting the source as the local oscillator of your 1296MHz converter and listening for a beacon whose frequency is known. L1 can then be adjusted to bring the signal onto the correct receiver dial calibration.

If difficulty is experienced in pulling the frequency to that marked on the crystal, it is very likely you have a non standard crystal. Pulling the frequency too far can result in the oscillator failing to restart after switching off and then on. The cure is to put a small value ceramic plate capacitor, say 10 - 33pF, in series with the crystal by cutting the PCB track near the latter. If you have to do this modification, keep the leads of the new capacitor short and use a zero temperature coefficient (NPO) capacitor, or frequency may drift unacceptably as the crystal oscillator warms up.

The source may be frequency modulated, although if it is to be used as a transmitter in the 1.3GHz band, greater deviation will be required than when multiplying to 10GHz. The modulator circuit shown in **Fig 11.20** can be used.

If the source is to be used as a 1.3GHz beacon source, the required 800Hz frequency shift can easily be obtained using this same circuit. Note, though, that the 'sense' of the deviation should be such that the marker signal (carrier only) should be at the nominated beacon frequency and 'space' keys the beacon low in frequency by 800Hz, returning to 'mark' for each character element. With the circuit given, this means the keying voltage should be low for mark and high for space (conventional), ie mark corresponds to an earth condition on the keying lead.

High Precision Frequency 10MHz Standard

A high stability frequency standard for 10MHz can be created using only three system components, a voltage controlled crystal oscillator (VCXO), a counter controlled by the signal from the GPS satellite system and a D/A converter for fine control of the VXCO. The short term and long term frequency stability that can be obtained by this simple means far exceed the requirements for practical amateur radio operations. The 10MHz standard can thus be used as the basis of a highly accurate method of generating local oscillator signals for microwave use.

The most commonly used method for precision time comparisons nowadays makes use of the satellites of the Global Positioning System (GPS). The GPS satellites carry atomic clocks of the highest accuracy, the operation of which is carefully monitored by the ground stations. A stable quartz oscillator regulated with the aid of the GPS ensures that its maximum frequency deviation always remains better than 1×10^{-11}. This is a precision of 0.0001Hz in 10MHz! Or, for the microwave amateur, 1Hz in 100GHz.

The frequency control of a 10MHz oscillator using GPS, shown in **Fig 11.21**, was designed by Wolfgang Schneider, DJ8ES, and Frank-Peter Richter, DL5HAT, and published in *VHF Communications* [11]. It uses an HP10544A VXCO (**Fig 11.22**), these are now available on the surplus market and often found for sale on Ebay. In practice an accuracy of approximately 4×10^{-10} can be achieved or, in other words, 4Hz in 10GHz. This value results from the inaccuracies of the counting process built into the system. In all frequency counters, the last bit should be taken with a pinch of salt. Depending on the phase position of the gate time to the counting signal, an error occurs of ±1 bit (phase error ±100ns). For a gate time of 1s, that would be 1Hz for the measuring frequency 10MHz (±1 x 10-7).

The first practical measurements were based on a gate time of 8s, which corresponds to a resolution of 0.125Hz. Together with the phase jitter of the GPS signal, there should have been

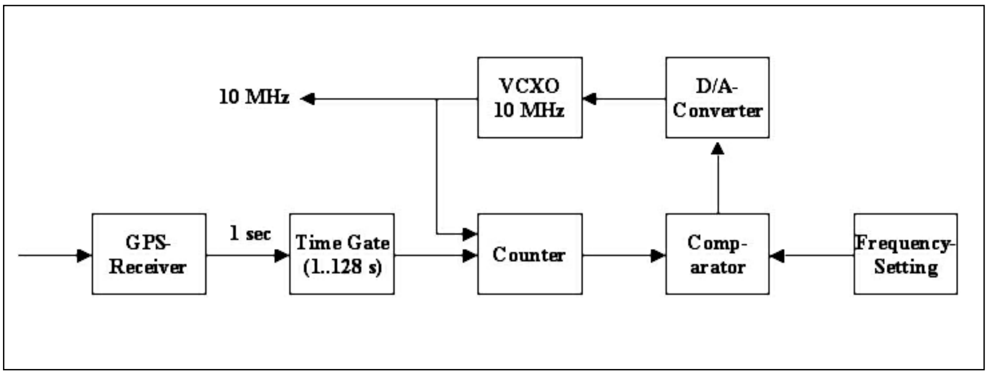


Fig 11.21: Block diagram of frequency control via GPS

Fig 11.22: HP10544A VXCO

uniform distribution and thus a levelling off of the reading over a relatively long period of time (max. 64 measurements). This turned out to be wishful thinking. On investigation, it was established that the oscillator frequency varied very slowly around the required value of 10MHz. The absolute frequency here was 10.0MHz ±0.0305Hz.

If the gate time is increased to 128s, in theory the reading improves to at least ±0.0078125Hz. However, the influence of the GPS phase jitter is now reduced. This results in an effective usable precision for the 10MHz signal of approximately 4×10^{-10} or 4Hz at 10GHz.

The control stage (**Fig 11.23**) operates like a frequency counter with an additional numerical comparator. The 10MHz output of the HP10544A oscillator is counted. The gate time of the counter is generated from the 1pps signal of the GPS receiver with a 74LS393. For control operation a gate time of 128 seconds is used, and 8s is used in the comparison mode for the OCXO.

The 74HC590 8-bit counter can be used, with a gate time of 8s, to measure the input frequency 10MHz ±16Hz. The minimum resolution here is 0.125Hz. In control operation (gate time 128s), this is improved by a factor of 16, resulting in the system determined precision of 0.78×10^{-9}, based on the 10MHz frequency oscillator.

The frequency of the HP oscillator can be finely adjusted using a tuning voltage of ±5V. This is done by the digital to analogue converter (AD 1851). It has a resolution of 16 bits for a control voltage range of ±3V. This gives a setting range for the OCXO of approximately ±0.5Hz.

The AT89C52 micro-controller controls all the functions including the D/A converter and the status in the LC display. The software in the micro-controller performs two tasks. Firstly, it enables a rough comparison operation to be carried out, and secondly it will continuously carry out the final fine adjustment using the GPS signal.

If pin 4 of K13 is open, then when the voltage is applied to the control circuit board the LC display shows "Warming Up". In order to eliminate any artificial jitter in the GPS signal, a mean value is formed and displayed from 64 readings from the 74HC590 counter. A change in the oscillator frequency using its frequency adjustment control will thus not display any effect for some time. So after a change of the control we must just wait for approximately 64 x 8s until the next adjustment takes place. If a value of ≥ ±0.250Hz is attained, we can switch over to the basic control by earthing Pin 4 of K13 and selecting the long gate time of 128s by means of a bridge between pins 15 and 16 at K14. The first message on the display is "Warming Up", the first value is displayed after approximately 15 minutes, it is not the deviation in Hertz but the value that is written in the AD 1851 D/A converter. This value can reach a maximum of ±32767, which means approximately ±3V. The software assesses the output of the counter and calculates the value for the AD 1851 D/A converter. From the present value of the counter and the mean of the last 64 counter results, the figure is determined which is to be added to or subtracted from the current D/A converter value. This can be seen on the display.

Theoretically, with the gate time of 128s, and with a mean value formed over 64 readings, the precision of 4×10^{-10} Hz (ie 4Hz in 10GHz) is achieved after 4.5 hours. However, it has been demonstrated in practice that this value has already been reached after approximately two hours.

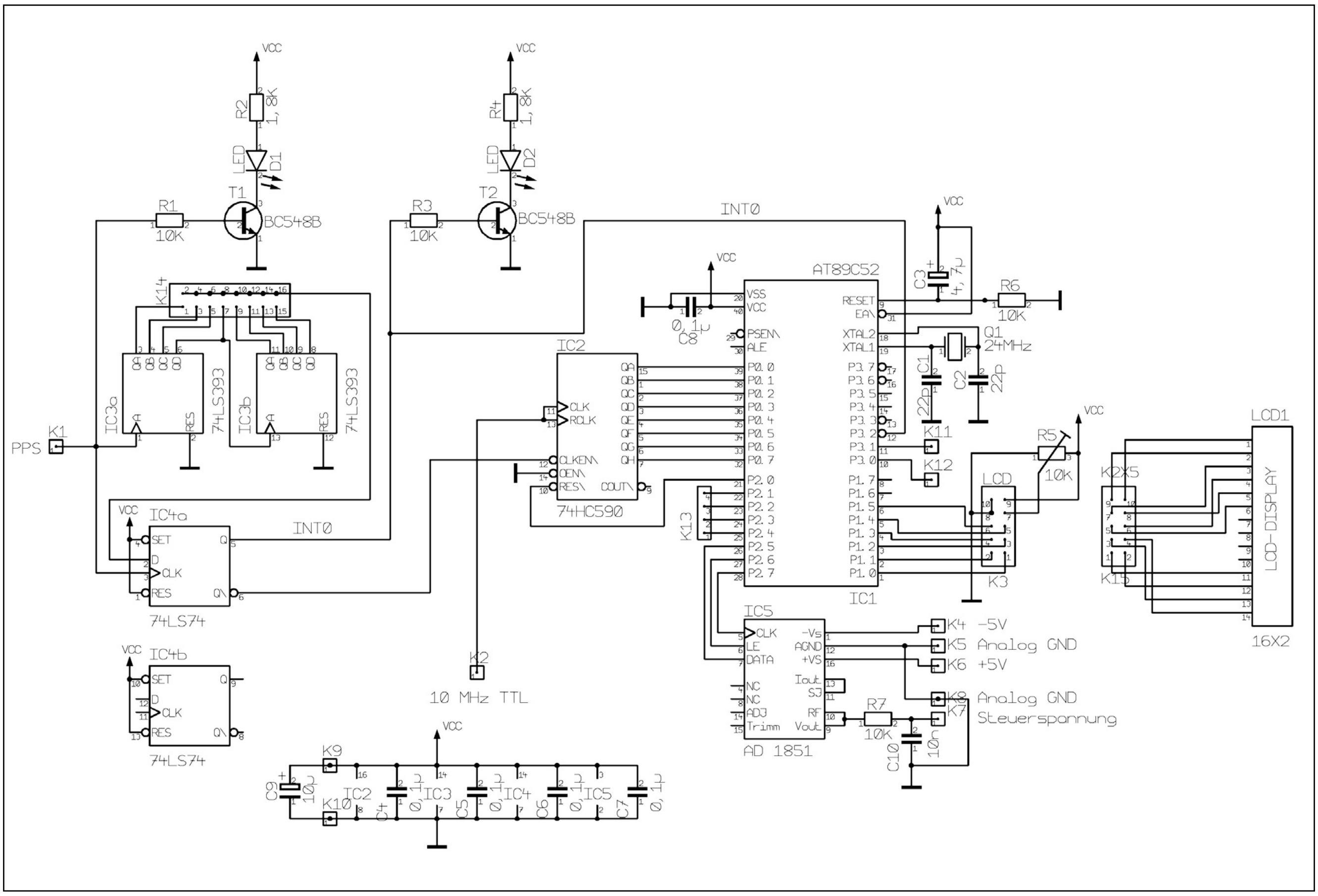


Fig 11.23: Circuit diagram of GPS control stage of high precision frequency standard for 10MHz

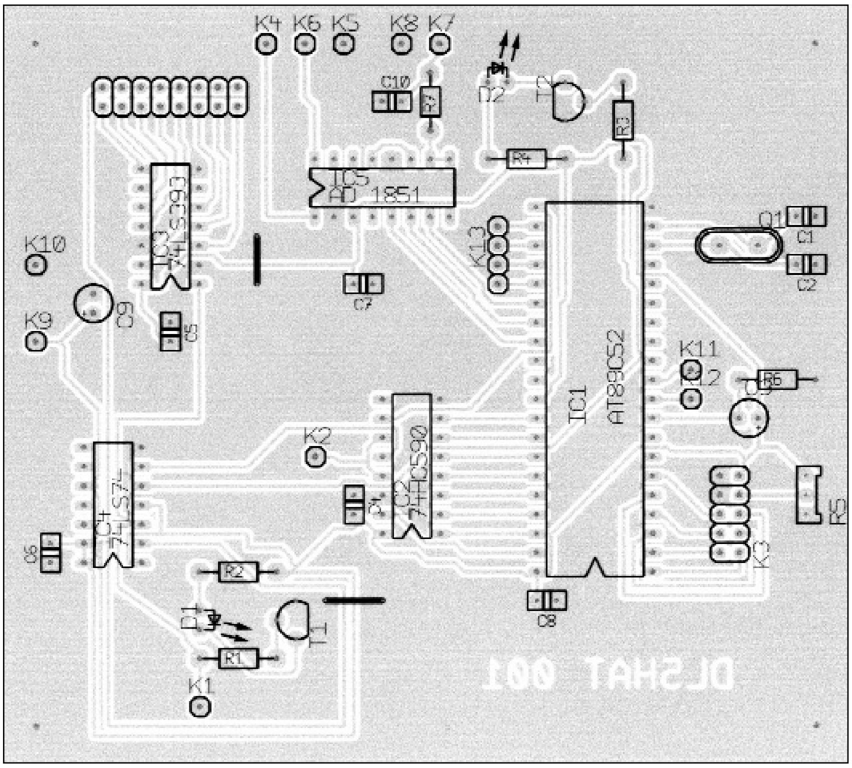


Fig 11.25. Component layout for GPS control stage of high precision frequency standard for 10MHz

The frequency controller circuit is assembled on 100mm x 100mm double sided PCB (**Fig 11.24** in Appendix B), the component layout can be see in **Fig 11.25** and the component list is shown in **Table 11.8**.

In the author's prototype (**Fig 11.26**) the HP10544A oscillator was used to drive a buffer stage designed by DJ8ES [12]. This gives TTL outputs at 1, 5 and 10MHz and 3 separate 10MHz sine wave outputs. The 1pps signal was generated from a GPS receiver manufactured by Garmin (GPS 25-LVS receiver board). The control assembly output supplies the control voltage for the HP oscillator. The tuning voltage, ±5V, *must* be separately generated in the frequency controller power supply. The following power supplies are required:

- +24V HP oscillator
- +5V GPS receiver
- +5V control assembly
- ±5V control voltage

Long-term observations of the 10MHz frequency standard over approximately four weeks confirmed the design criteria.

Miscellaneous components	
1 x micro-controller	AT89C52
1 x A/D converter	AD1851
1 x TTL-IC	74LS74
1 x TTL-IC	74LS393
1 x TTL-IC	74HC590
2 x Transistor	BC848B
2 x LED	green, low current
1 x crystal	24 MHz
1 x potentiometer	10k
1 x socket terminal strip, 10-pin	
1 x plug strip, 10-pin	
1 x stud strip, 14-pin	
1 x jumper	
1 x circuit board	DL5HAT 001
Resistors	
2 x	1.8k
4 x	10k
Ceramic capacitors	
5 x	0.1μF
2 x	22pF
1 x	10nF
Tantalum capacitors	
1 x	4.7μF/25V
1 x	10μF/25V

Table 11.8: Component list for GPS controller for 10MHz frequency standard.

RECEIVE PREAMPLIFIERS

Receive preamplifiers are an important part of the microwave station. They are often used as masthead preamplifiers to overcome feeder loss that can be quite considerable for the higher frequency bands. There are many suitable semiconductors available to produce very low noise preamplifiers for all the amateur bands up to 10GHz. Above that, the devices are available but they are fairly expensive. Two designs are shown here to illustrate the technology available.

13cm Preamplifier

This design is by Rainer Bertelsmeier, DJ9BV, and originally appeared in *Dubus Magazine*. A preamp equipped with a PHEMT provides a top-notch performance in noise figure and gain as well as unconditional stability for the 13cm amateur band. The noise figure is 0.35dB at a gain of 15dB. It utilises the C band PHEMT, NEC NE42484A and provides a facility for an optional second stage on board. The second stage with the HP GaAs MMIC MGA86576 can boost the gain to about 40dB in one

Fig 11.26: Picture of the prototype high precision frequency standard for 10MHz controlled by GPS

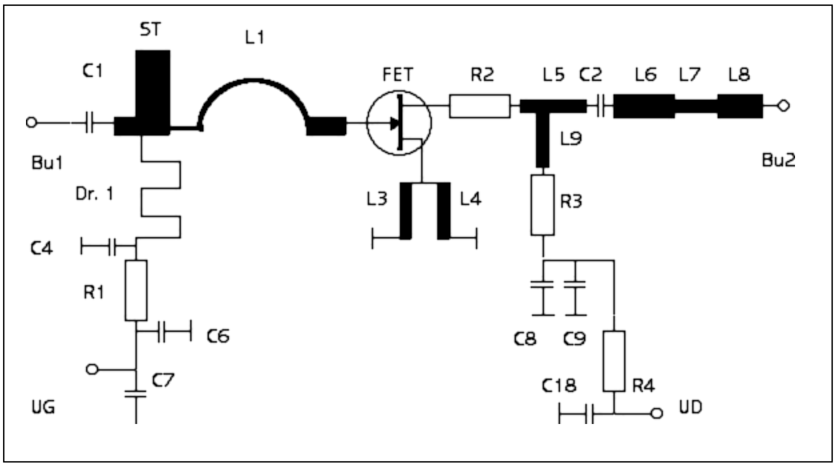


Fig 11.27: Circuit diagram of single stage 13cm PHEMT preamplifier

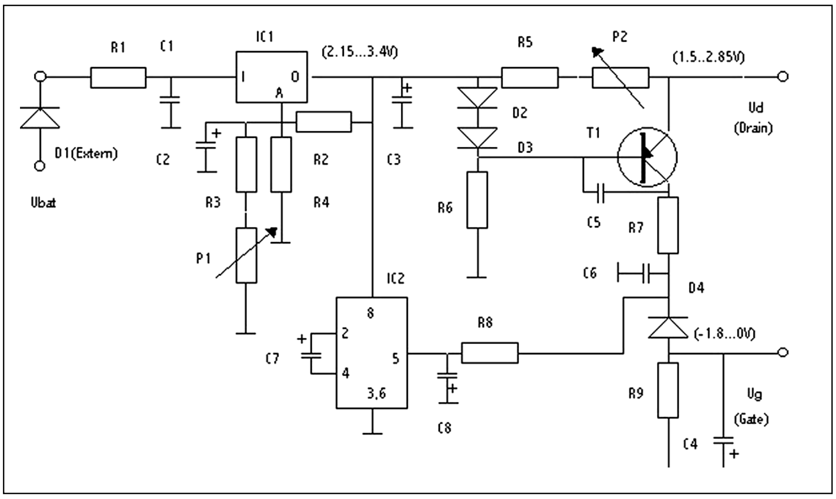


Fig 11.28: Circuit diagram of bias circuit for 13cm PHEMT preamplifier

Capacitors	
C1	4.7pF Chip-C 50mil (500 DHA 4R7 JG)
C2, 3, 6	100pF SMD-C, size 0805
C4	5.6pF SMD-C, size 0805
C7, 8, 19	1000pF SMD, size 0805
C9, 18, 20	10nF SMD-C, size 0805
C10, 12, 17	0.1µF SMD-C, size 1206
C11, 14, 15	10µF SMD-Electro, size 1210
C13	1µF SMD-Electro, size 1206
C16	1000pF Feedthrough
Resistors	
R1, 3	470 SMD-R size 1206
R2, 14	390 SMD-R size 1206
R4, 5	100 SMD-R size 1206
R12	6.8kO SMD-R size 1206
P1	1000 SMD-Pot, Murata 4310
Miscellaneous	
Dr.1	Printed λ/4
L1	Wire loop, 0.5mm gold plated copper wire 18mm long, 1mm above board
L2	Wire loop, 0.5mm gold plated copper wire 8mm long, on PCB
D1	IN4007
FET	NE42484A, NEC
MMIC	MGA-86576, HP
T1	PNP eg BC807, BC856, BC857, BC858, BC859, SOT-23
IC1	LTC1044SN8
Bu1, 2	N small flange or SMA
PCB	Taconix TLX, 35 x 72mm, 0.79mm er = 2.55
Box	Tinplate 35 x 74 x 30mm

Table 11.9: Components for 13cm PHEMT preamp

enclosure. The preamplifier is rather broadband and usable from 2300 to 2450MHz.

The construction of this LNA follows the proven design of the 23cm HEMT LNA [13] by using a wire loop with an open stub as an input circuit (**Fig 11.27**). The FET's grounded source requires a bias circuit to provide the negative voltage for the gate. A special active bias circuit (**Fig 11.28**) is integrated into the RF board that provides regulation of voltage and current for the FET. The component list is shown in **Table 11.9**.

Stub ST and inductance L1 provide a match for optimum source impedance for minimum noise figure. L1 is as a dielectric transmission line above a PTFE board and has somewhat lower loss than a microstripline. L3 and L4 provide inductive feedback to increase the stability factor and input return loss. R1, R2, L9 and R3 increase the stability factor. The system of C2/L5/C6/L7 and L8 is specially designed to match the output of the single stage version to 50 ohms and to allow easy insertion of the GaAs MMIC for the two stage version.

In the two stage version (**Fig 11.29**) it provides the appropriate input and output match to the MMIC. This solution was found by doing some hours of design work with the software design package *Microwave Harmonica*. It allows the two versions to have the same PCB. C4 provides a short on 2.3GHz, because it is in series resonance at this frequency. On all frequencies outside the operating band the gate structure is terminated by R1. Dr1 is a printed λ/4 choke to decouple the gate bias supply.

The two stage version utilises a HP GaAs MMIC, MGA86576 in the second stage. It provides about 2dB noise figure and 24dB gain. Input is matched by a wire loop for optimum noise figure. Output is terminated by a resistor R5 and a short transmission line L10. Together with L7/L8 and C3 a good output return loss

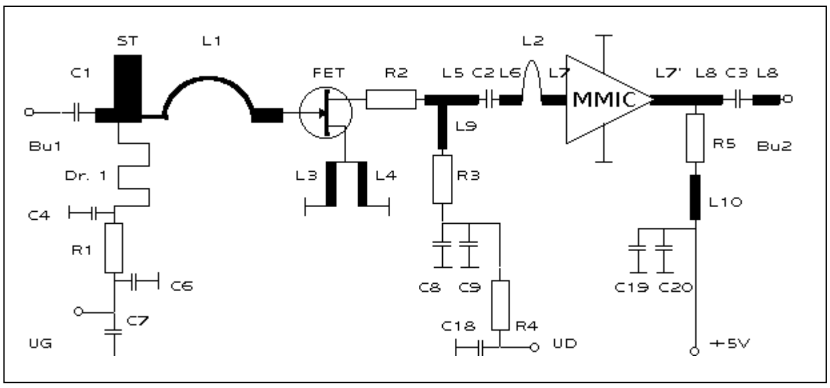


Fig 11.29: Circuit diagram of two stage 13cm PHEMT preamplifier

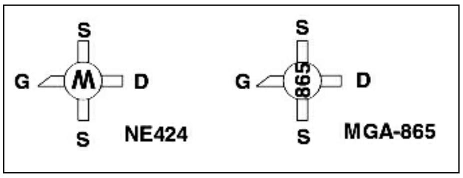


Fig 11.31: Top view of FET and MMIC used in 13cm preamplifier

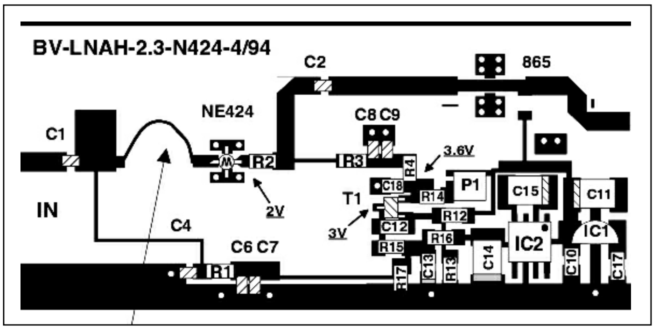


Fig 11.32: Component layout for single stage 13cm PHEMT preamplifier

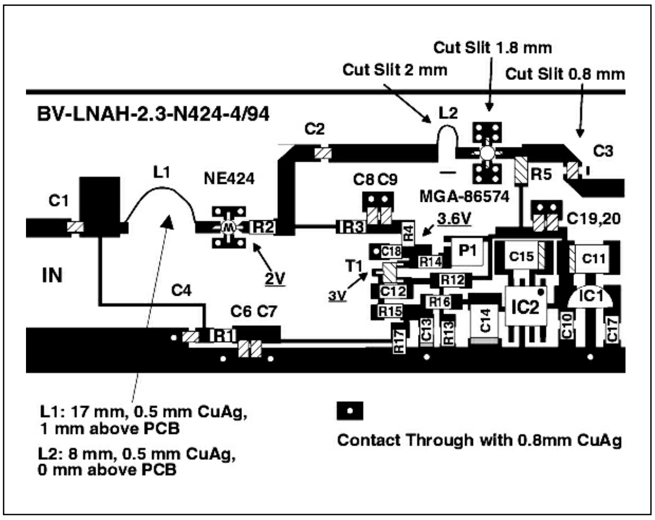


Fig 11.33: Component layout for two stage 13cm PHEMT preamplifier

is measured. The source pads have to provide a very low inductance path to the ground plane, to preserve the MMIC's inherent unconditional stability.

To achieve unconditional stability four ground connections are needed on each source. Appropriate source pads are provided on the PCB. Simulation indicates a minimum K factor of 1.2 in this arrangement on a 0.79mm thick substrate. A thicker substrate is prohibitive. The MMIC typically adds 0.07dB to the noise figure of the first stage. This is somewhat difficult to measure, because most converters will exhibit gain compression, when the noise power of the source, amplified by more than 40dB, will enter the converter.

The construction uses microstripline techniques on glass PTFE substrate Taconix TLX with 0.79mm thickness with the PCB having dimensions of 34 x 72mm. An active bias circuit, which provides constant voltage and current, is integrated into the PCB (**Fig 30** in Appendix B). **Fig 31** shows a top view of FETs and **Fig 11.32** shows the component layout. The construction process is:

- Prepare tinned box (solder side walls).
- Prepare PCB to fit into box.
- Prepare holes for N connectors. Note, Input and output connector are asymmetrical. Use PCB to do the markings.
- Drill holes for through connections (0.9mm diameter) and use 0.8mm gold plated copper wire at the positions indicated.
- Solder all resistors onto PCB.
- Solder all capacitors onto PCB.
- For L1, cut a 17mm length of gold plated copper 0.5mm diameter wire. Bend down the ends at 1mm length to 45 degrees. Form wire into a half circle loop and solder into the circuit with 1mm clearance from the PCB. The wire loop has to be flush with the end of the gate stripline and should be soldered a right angles to it. The wire loop has to be oriented flat and parallel to the PCB.
- Verify the open circuit function of bias circuit. Adjust P1 to 45 ohms. Solder a 100-ohm test resistor from the drain terminal on the PCB to ground. Apply +12V to IC1 and measure +5V at output of IC1, -5V at IC2 Pin5, -2.5V> at collector of T1, +3.6V at emitter of T1, +3V at base of T1, -2.5V at R17 and +2.0V across the 100 ohms. If OK, remove 100-ohm test resistor.
- Solder PHEMT onto PCB. Ground the PCB, your body and the power supply of the soldering iron. Never touch the PHEMT on the gate, only on the source or the drain, when applying it to the PCB and solder fast (much less than 5 seconds).
- Solder N Connectors into sides of the box.
- Solder the finished PCB into the box, solder both sides of the PCB at the sides of the box and solder centre pins of the connectors to the microstriplines.
- Solder feed-through capacitors into box.
- Connect D1 between feed-through capacitor and PCB.
- Connect 12V and adjust P1 for 16mA drain current (measure 160mV across R4 on RF Board). Voltages should be around +2.0V at the drain terminal, -0.4V at the gate, +3.6V at emitter of T1.
- Connect LNA to a noise figure meter, if you have one, and adjust input wire loop, adjust the clearance to PCB as well as drain current by adjusting P1 for minimum noise figure. Even without tuning, the noise figure should be within 0.1dB of minimum because of the limited tuning range of the wire loop.
- Glue conducting foam inside the top cover and slip into the top of the box.

To add the MMIC Amplifier, refer to **Fig 11.33** for Construction.

- Prepare PCB by cutting slits into the microstriplines around the MGA865. These are a 2mm slit for L2, a 1.8mm slit for the MMIC and a 0.8mm slit for C3.
- For L2 cut an 8mm length of gold plated copper 0.5mm diameter wire. Form wire into a half circle loop and solder wire loop into the circuit. The wire loop has to lie flat on the PCB, flush with the end of the gate stripline and should be soldered in a right angle to it.
- Follow other instructions given above

Measurements were taken using an HP8510 network analyser and HP8970B/HP346A noise figure analyser, transferred to a PC and plotted. **Figs 11.34 and 11.35** show the measurement results for gain and noise figure for the one stage and two stage version respectively. Using a special PHEMT, NEC NE42484A optimised for C Band, a typical noise figure of 0.35dB at a gain of 15dB can be measured on 2.32GHz. An optional second stage on the same PCB using the GaAs MMIC MGA86576 from HP will boost the gain from 15db

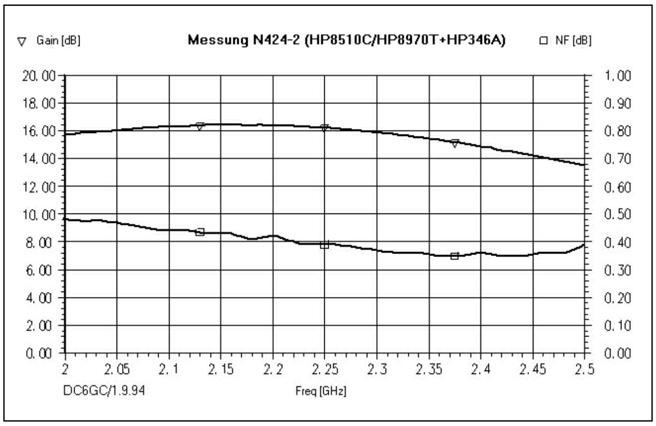


Fig 11.34: Noise figure and gain measurements for single stage 13cm preamplifier

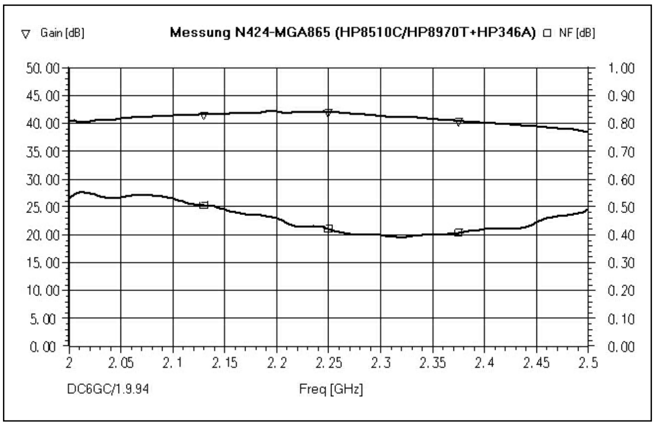


Fig 11.35: Noise figure and gain measurements for two stage 13cm preamplifier

to 41dB. The two stage version measures with a noise figure of 0.45dB. This version can be used for satellite operation. For EME, where lowest noise figure is at premium, a cascade of two identical one stage LNAs may be more appropriate. Both versions are rather broadband. They can cover the various portions of the 13cms amateur allocation from 2300 to 2450MHz without re-tuning. The real surprise is the performance of the C band PHEMT NE424. It performs better than several other HEMTs (FHX35, FHX06, NE324, and NE326) tried in this circuit, and it measures 0.15dB better than its published noise figure. In fact the *Microwave Harmonica* simulation predicts a 0.5dB noise figure based on the data sheet value. The lower noise figure measured seems to be due to a special bias current and the lower value of gamma at approximately 0.75 which is due to the gate length of 0.35µm. This provides optimum properties for application in 2 - 4GHz LNAs. Stability is excellent. This has been achieved by a carefully controlled combination of inductive source feedback, resistive loading in the drain and non resonant DC feed structures for drain and gate. A broadband sweep from 0.2 to 20GHz showed a stability factor K of not less than 1.2 and the B1 measure was always greater than zero. These two properties indicate unconditional stability. At the operating frequency of 2.3GHz, stability factor is about 1.6. The two-stage version with the MGA865 measures with K>>4 at all frequencies.

The preamplifier provides quantum leap towards the perfect noiseless preamplifier. It uses a low cost and rugged C band PHEMT instead of relying on expensive X band HEMTs. An improvement of about 0.2dB in noise figure has been achieved in comparison to the no tune HEMT preamplifier described in [14]. This improvement provides roughly 1.5dB more S/N in EME or satellite operation but is not noticeable in terrestrial links. However, the new preamplifier has to be tuned. This requires a noise figure meter for alignment. For those who like a no tune device, the HEMT preamplifier in [14] provides adequate performance with a typical NF of 0.55dB.

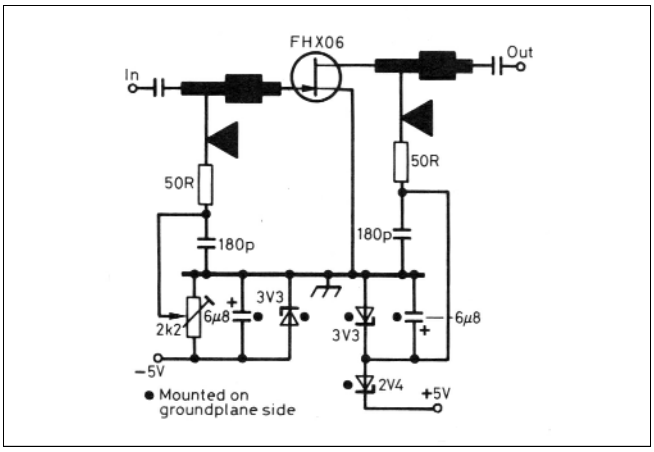


Fig 11.36: Circuit diagram of the G3WDG-004 3cm HEMT preamplifier

3cm Preamplifier

Much of the equipment used by amateurs in the UK comes from designs by G3WDG. This preamplifier design, G3WDG-004, is one of the full range of designs for 10GHz. Two versions of the -004 amplifier are available, one with SMA input and one with WG16 waveguide input. All prototypes averaged 1dB NF or slightly under. Where the noise figure was fractionally above 1dB there was some evidence of degradation during construction, possibly by static. The circuit for either amplifier is identical and is shown in Fig 11.36. A short kit is available from the Microwave Components Service [10], it contains both the HEMT and a detailed construction sheets. The advantage of the waveguide input version is that a probe inputs signal directly from the waveguide to the HEMT. The SMA version will almost certainly require a least one more SMA connector and a short length of semi-rigid 'cable' to the amplifier input. Thus the noise figure will be degraded by a few tenths of a dB, it is vital to win that little loss back if you are searching for the ultimate performance, for instance for moonbounce!

The layout details of the board of either version are given in Fig 11.37 to give you some idea of the work involved. There is also a universal regulated power supply that is used on many of the microwave designs available, the circuit diagram is shown in

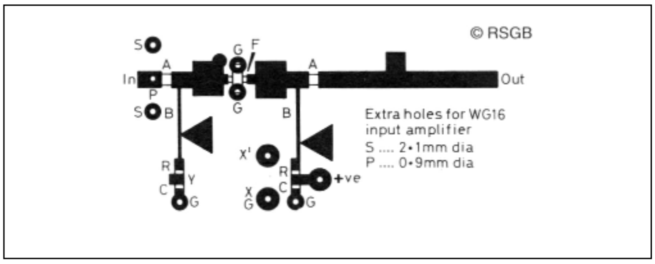


Fig 11.37: PCB and component layout of the G3WDG-004 HEMT preamplifier. Key to components: A- 2.2pF ATC capacitor, B - bias wires, C - 180pF cip capacitors, F - HEMT, G - grounding vero pins, R - 47O chip resistors, X and Y positions of connection to bias pot (X is -ve input)

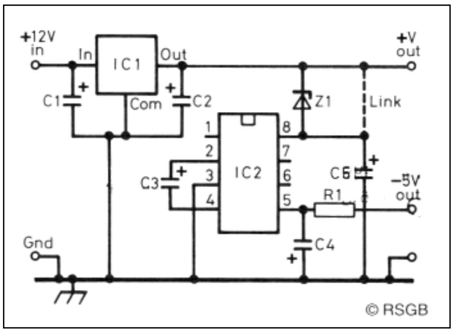


Fig 11.38: Basic circuit of the G4FRE-023 dual output power regulator module. As used on the G3WDG-004 3cm preamplifier

R1	2k2O
C1, 2	1µF SMD tantalum
C3, 4	22µF SMD tantalum
C5	10µF SMD tantalum
Z1	Not fitted - replace with a link
IC1	78L05
IC2	ICL7660SCPA

Table 11.10: Component list for dual output power regulator used on G3WDG-004 HEMT pre-amplifier

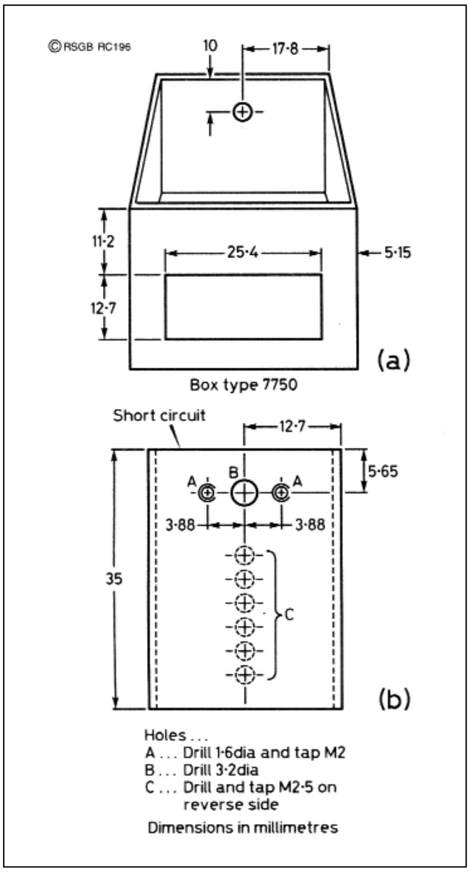


Fig 11.39: (a) Preparing a tin plate box for the G3WDG-004 HEMT pre-amplifier (waveguide version). (b) Waveguide 16 dimensions and drilling sizes

Fig 11.38 . Details of the components for this LNA are given in Table 11.10. The preamplifier gives a gain of about 10dB with a noise figure of 1dB or a little less. The additional holes shown on the layout are needed for the waveguide 16 (WG16) version of the amplifier which is more difficult to construct than the SMA input version. This is because more 'mechanical' work is needed for this version in order to gain those few valuable tenths of a dB noise figure where ultimate performance is needed. Fig 11.39(a) shows the engineering work needed to adapt a Type 7750 tin plate box (Piper Communications) to take the short length of WG16 (input), the PCB, and output SMA connector. Construction should follow the order suggested by G3WDG:

- Connect 12V and adjust P1 for 16mA drain current (measure 160mV across R4 on RF Board). Voltages should be around +2.0V at the drain terminal, -0.4V at the gate, +3.6<N>V at emitter of T1.Solder the tin plate box together in the usual fashion. Then cut out the hole needed for the WG16 by drilling a lot of small holes inside the outline of the large hole and then join them and file out to be a tight fit on the WG16.
- Connect 12V and adjust P1 for 16mA drain current (measure 160mV across R4 on RF Board). Voltages should be around +2.0V at the drain terminal, -0.4V at the gate, +3.6V at emitter of T1. Cut the piece of WG16 to length, square off the ends and drill/tap all the holes shown in Fig 11.39(b) .
- Connect 12V and adjust P1 for 16mA drain current (measure 160mV across R4 on RF Board). Voltages should be around +2.0V at the drain terminal, -0.4V at the gate, +3.6V at emitter of T1. Cut a short circuit plate (thin copper sheet or double sided FR4/GI0 PCB material can be used) slightly larger than the waveguide and solder it to the end of the waveguide where shown.
- Connect 12V and adjust P1 for 16mA drain current (measure 160mV across R4 on RF Board). Voltages should be around +2.0V at the drain terminal, -0.4V at the gate, +3.6V at emitter of T1. File off any excess, especially where the PCB abuts the WG.

Now refer to Fig 11.40 and prepare the PCB by very carefully cutting away the copper around the hole P with a Vero cutter, to about 3mm diameter. Trim the PCB to be a neat fit into the box, be sure to leave enough on the input end of the board to allow 2 x 4mm screws to fit inside the box! These screws serve to clamp the PCB to the waveguide (holes S on the PCB to holes A on the WG). Cut off part of the input line, leaving 1mm to the left of the centre line of hole P. Fit and solder all the grounding pins. Fit the Veropin probe and solder it in place in hole P. Cut and fit the PTFE washer, referring to Fig 11.40(b), on to the pin on the ground plane side of the PCB, pushing it firmly down to the PCB, but taking care not to damage the pin or board.

Fit the piece of WG into the box, through the hole, offer up the PCB, check alignment and loosely fit the 4mm screws. If the board does not align with the probe pin in the centre of the WG probe hole, Fig 11.40(b), then file the hole in the box as required for proper alignment. Mark the centre of the hole for the SMA output connector (the 17.8mm dimension shown in Fig 11.39(a) may need to be altered!). Remove the screws and dismantle PCB and WG, drill the holes in the box for the SMA connector and open the centre clearance hole to 3.3mm. Refit the WG, PCB and 4mm

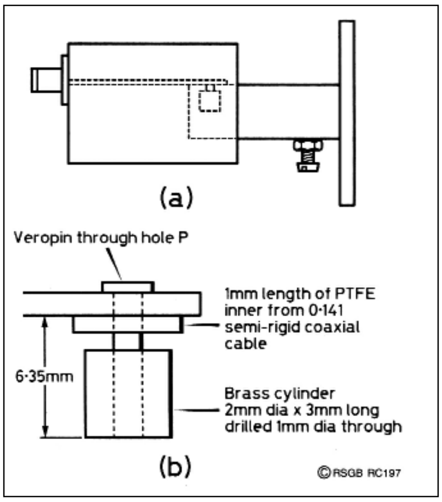


Fig 11.40. (a) General side view of G3WDG-004 WG16 HEMT pre-amplifier assembly. (b) Detail of waveguide to PCB transition probe

screws. Then cut the spill of the SMA connector to 1.5mm and fit it to the box. Fit the PCB into place so that the output track touches the spill, and solder it to the spill. Solder three sides of the WG to the tinplate box, but not to the side to which the PCB mates. Remove the SMA fixing screws, unsolder the spill and remove the connector. Remove the 4mm screws and the PCB. Carefully solder the last face of the waveguide, but do not melt the solder on the WG shorting plate or you'll have to start again! After soldering, make sure the mating surface of both the waveguide and the PCB are clean, bright and free from oxide and flux. Fit the brass collar, as shown in Fig 11.40((b), to the probe pin, the length shown is critical to 0.05mm. Complete the WG assembly by soldering a WG16 flange to the input end, again being careful not to melt the solder on the rest of the assembly. Apply a small amount of conductive epoxy to the waveguide where the PCB mates, especially around the probe hole, but not so much that a short circuit occurs when the PCB and WG are squeezed together! Reassemble the PCB and SMA connector. Solder the output track to the SMA spill and tighten the 4mm screws. Check there is not a short between the WG and input probe caused by 'squeezing' of excess epoxy. Cure the epoxy by heating at about 150 degrees C for an hour or so. After curing, recheck for absence of short circuit on the input probe - if there is a short, remove the PCB and try again! This may damage the PCB, so do it with care and try to get it right the first time! Finally, solder all around the ground plane / box junction. Finish off the amplifier by mounting all the components as for the SMA version. Incidentally, the holes C on the WG16 are used for matching screws to literally "screw the last tenth of a dB" out of the amplifier!

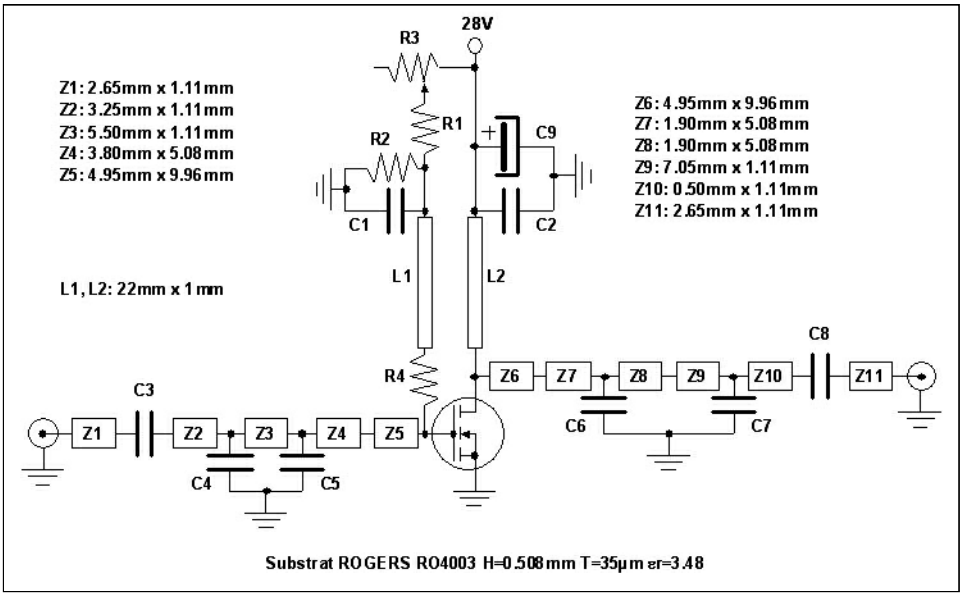


Fig 11.41: Circuit diagram for MRF284 23cm power amplifier designed by F4CIB

TRANSMIT POWER AMPLIFIERS

Generating any reasonable amount of power on the microwave bands was the speciality of the valve. On the lower bands the trusty 2C39A was well used in anything up to eight valve designs. On the higher bands the travelling wave tube was used but this needed special powers supplies. These are being replaced with semiconductor devices, with their compact design and more manageable power requirements. These can also be masthead mounted.

On 23cm, the Mitsubishi M57762 hybrid amplifier has been used for many years but this is now going out of production and other semiconductors are replacing them, an example shown here. As the frequency increases it becomes more difficult to find, or afford, semiconductors for power amplifiers. Fortunately the gain from the antennas used comes to the rescue reducing the input power needed to achieve the required radiated power. On the higher bands special techniques are still required, such a direct bonding to semiconductor substrates, and example of these techniques is shown here.

LDMOSFET Power Amplifier for 23cm

This amplifier design was researched by Franck Rousseau, F4CIB, [15] and published in *Dubus Magazine*. Conversion of Motorola devices on 23cm and 13cm was the subject of articles by Maurice Niquel, F5EFD, [16] and [17] and Jean-Pierre Lecarpentier, F1ANH, [18]. These devices provide a really good opportunity to replace expensive power bricks like the M57762, and the 2C39 cavity amplifiers. The transistor characteristics, extracted from product datasheet [19] are really interesting for ham applications:

- Broadband (1GHz-2.6GHz), class A & AB specified

Specified Two–Tone Performance @ 2000 MHz, 26V

- Output Power = 30 Watts PEP
- Power Gain = 9dB
- Efficiency = 30%
- Intermodulation Distortion = –29dBc

Typical Single–Tone Performance at 2000MHz, 26V

- Output Power = 30 Watts CW
- Power Gain = 9.5dB
- Efficiency = 45%
- Capable of Handling 10:1 VSWR, @ 26V DC, 2000MHz, 30 Watts CW Output Power

The circuit diagram (**Fig 11.41**) is an adaptation from the original demonstration board with some component values changed. The PCB layout is shown in **Fig 11.42**.

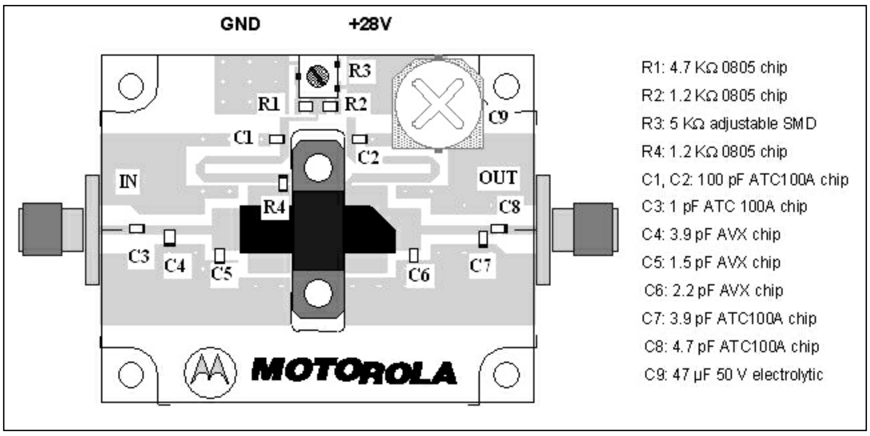


Fig 11.42: PCB layout for MRF284 23cm power amplifier

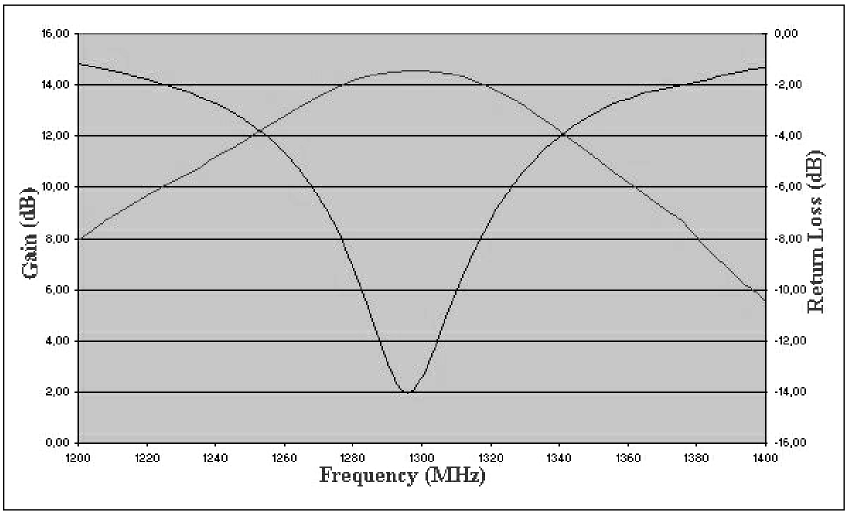


Fig 11.43: Gain and return loss for MRF284 23cm power amplifier

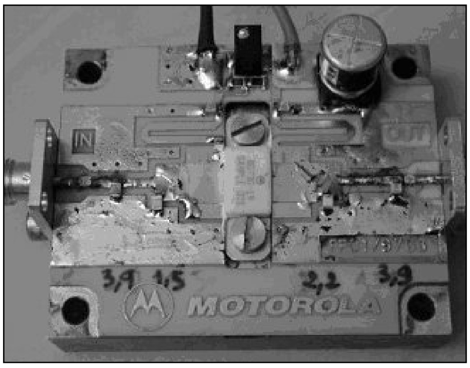


Fig 11.44: Photograph of the completed MRF284 23cm power amplifier

All measurements have been done on professional calibrated equipment:

- Network Analyser HP8753: Forward Gain S21 and Input Return Loss S11.
- Power Meter HP438A: RF Output Power.
- DC Power Supply HP6632: DC Sources and Measures.

A single DC supply is required, so the first step is to set bias current with the help of the adjustable resistor. Set bias to 200mA and forget it! Input matching was achieved with a 3.9pF capacitor close to the DC blocking capacitor (C3) and a 1.5pF capacitor on the 30-ohm line. The position of the 3.9pF capacitor is the most critical, the 1.5pF capacitor along 30-ohm line helps to centre the tuned frequency at 1296MHz. Once the input circuit is set up like this, the gain is around 9-10dB, and the Input Return Loss (IRL) is –13 to 14dB. The output network should be a 2.2pF capacitor soldered at the centre of the 30-ohm line and a 3.9pF as close as possible to the DC blocking capacitor. This will increase the gain to 14dB, while IRL remains at –14dB (Fig 11.43). The 1dB compressed power, P1dB, (Output power level for a gain drop of 1dB) has been measured at 45.5dBm (35W) in accordance with device specifications. For the following conditions:

- VCC = 28 V
- Input Power = 31.6dBm
- Output power measured at 45.2dBm
- DC current 2.41A.

Computing the Power Added Efficiency (PAE) gives 47% which exceeds specifications, but that's not a surprise as the device is able to work up to 2 GHz. Fig 11.44 shows a picture of the completed amplifier.

As expected, the MRF282, MRF286 and MRF284 work without any problems on 1296MHz. Output power is twice that of a power brick, and gain should be a good alternative to these expensive power modules. The trick is to find a good and cheap source of these transistors; as a clue they are used in cellular base stations so they should be available on the second hand market. The design does need a 2.5A 28V DC power supply, a suitable design is shown in [20]

GaAsFET Amplifier for 3cm

The power amplifier for 3cm described here was developed by Peter Vogl, DL1RQ, [21] with a view to ease of copying and reliable long term operation. Numerous examples of the 1W power amplifier have been running for some years (some of them installed on masts) and none has yet given rise to any problems. This circuit was expanded using a TIM 0910-4 from Toshiba to give 5W output, the design for the 5W amplifier is shown in [21] and [3]. In spite of all efforts at reproducibility, the cumulative total of the small tolerances in the component values and the assembly can eventually lead to significant individual deviations (-3 dB is normal) in the amplification and output power. But there is some comfort in the fact that, with patience, experience and good measurement facilities, a power amplifier can be trimmed to the rated values with a fine calibration using the "small disc method" (see below).

The circuit diagram of the 1W amplifier, shown in Fig 11.45, is similar to DL1RQ's two stage 5.7GHz power amplifier, published in [22] and [23], including the additional voltage inverter for the negative gate voltage. Good experiences with the reliable FSX52WF transistors (drive) and FLC103WG transistors (high level stage) from Fitjitsu led to trials at 10GHz, which were immediately successful, although the FLC103WG is only specified for use up to 8GHz by the manufacturer. 0805 model 1pF SMD capacitors were used as high frequency coupling elements

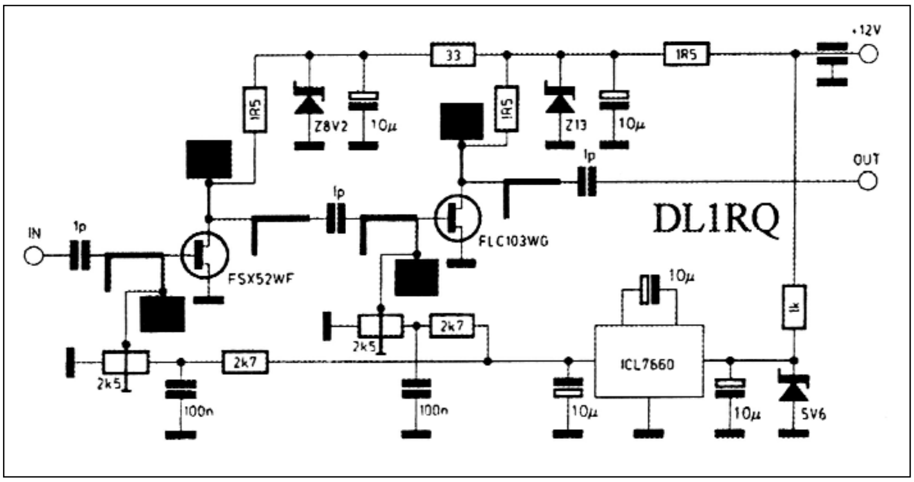


Fig 11.45: Circuit diagram of the 1W GaAsFET amplifier for 3cm

(2.0 mm x 1.25 mm). Research carried out only recently as part of a specialist project [24] showed that the SMD capacitors used by the author, with a series inductance of L_{series} = 0.66 ±0.01nH, differ from those components available by normal mail order, with L_{series} = 0.72 ±0.02nH. Unfortunately, DL1RQ was not in a position to identify the individual manufacturer. In any case, a rough calculation quickly shows that the series resonance frequency of these SMD capacitors lies around 6.0GHz. So, at 10GHz the coupling “capacitors” should be considered more as

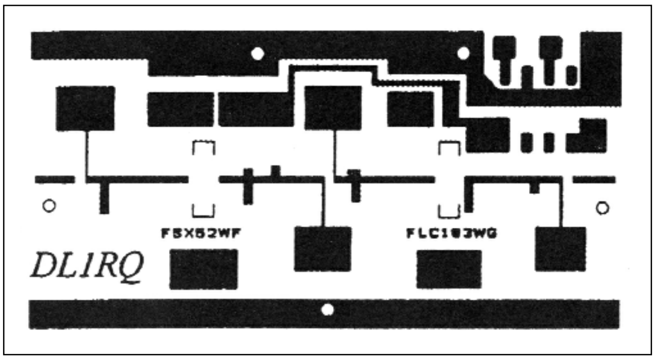


Fig 11.46: PCB layout for the 1W GaAsFET amplifier for 3cm

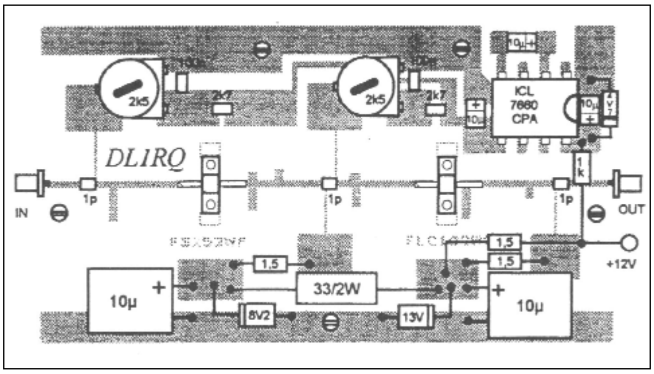
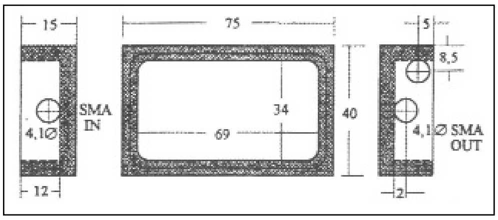


Fig 11.47: Component layout for the 1W GaAsFET amplifier for 3cm

Fig 11.48: Dimensions of a milled housing for the 1W GaAsFET amplifier for 3cm



DC disconnecting components with inductive behaviour. Naturally, this inductance affects the matching by the striplines (at the cost of narrowing the band!).

The power supply was deliberately made simple. Stabilisation was provided through 1.3W zener diodes, which provide protection against over voltage and reverse polarity. The 1.5-ohm/0.25W axial carbon film resistors act as both isolation resistances and safety resistances. A circuit for protection if the negative power supply failed was dispensed with following an involuntary 24 hour test without any negative supply voltage, which did not damage the semiconductors.

The assembly and board layout of an amplifier for microwaves are determined by two essential requirements:

- The high frequency transition from the earth surface of the board to the source flange of the transistor must be as close to ideal and as smooth as possible
- The extraction of the transistor’s lost heat must be as close to ideal as possible

A board with a sandwich construction has proved to be a way of being able to fulfil both requirements. The PCB layout (**Fig 11.46**) is etched onto an RT/Ditroid D-5870 board measuring 68.5mm x 34mm x 0.25mm. After pre-tinning the earth surface, the board is soldered onto a 1.0mm thick copper plate under high pressure. Next, two oval grooves 0.75mm deep are milled, using a 2.5mm diameter bore groove milling cutter, for the source flanges of the transistors. When the five 2.1mm diameter holes have been made for the board to be screwed into the housing and for the contacts to be connected up, and when the tracks have been tin plated (or silver plated), the board is assembled as far as the two 10µF capacitors (on the drain side) and the transistors (**Fig 11.47**). In order to guarantee good heat transfer between the copper plate and the milled aluminium housing (**Fig 11.48**), some heat conducting paste is smeared over the aluminium base in the vicinity of the transistors. To ensure good connection between the high frequency section and the earth in the input and output areas, silver conducting lacquer can be smeared there (very sparingly, of course). The partly assembled board is now fitted into the suitably prepared aluminium housing and screwed down by five M2 brass screws. When the connections to the feed-through capacitor have been completed, it is possible to check the DC function. For this purpose, the two trimmers are pre set to a gate voltage of about -1.5V. The trickiest stage in the procedure is the soldering of the GaAsFET into the milled grooves. To this end, the aluminium housing is first heated, with the board inside, to precisely 150 degrees C. Each milled groove is then pre tinned, using low temperature solder with a melting temperature of 140 degrees C.

Excess tin is then removed using a de-soldering pump. The transistors are next placed in the grooves; all the relevant safety measures known must be taken. Normally, the tin binds very well with the gold plated flanged base, something that can easily be tested by a visual check of the flanged holes. Naturally, this soldering process should be carried out as rapidly as possible. The housing is then immediately placed on a cold copper block or a large cooling body, so that the temperature quickly falls. Drain and gate connections are soldered onto the striplines; all the relevant safety measures must be taken. The two 10µF capacitors on the drain side are fitted and the SMA flanged bushes are screwed on. The power amplifier is ready for tuning (**Fig 11.49**).

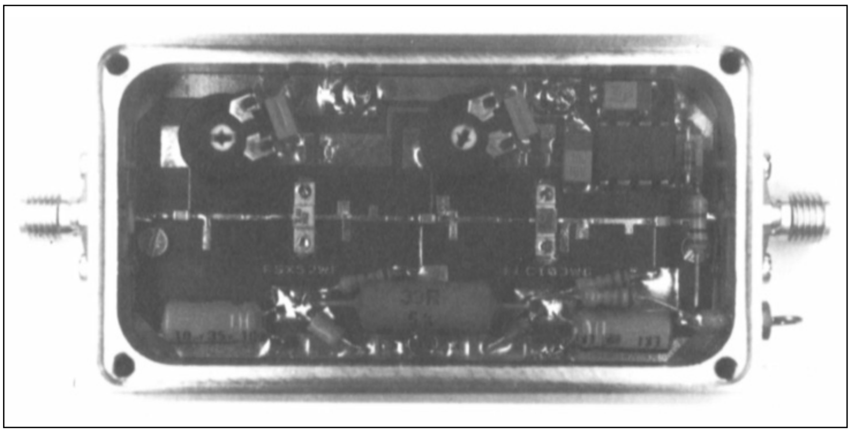


Fig 11.49: Photograph of the 1W GaAsFET amplifier for 3cm

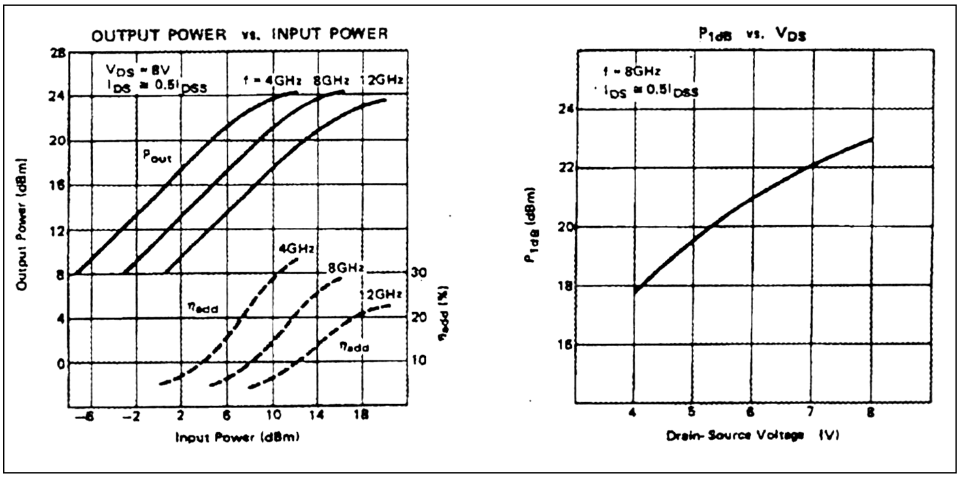


Fig 11.50: Data for the Fujitsu FSX53WF used in the 1W GaAsFET amplifier for 3cm

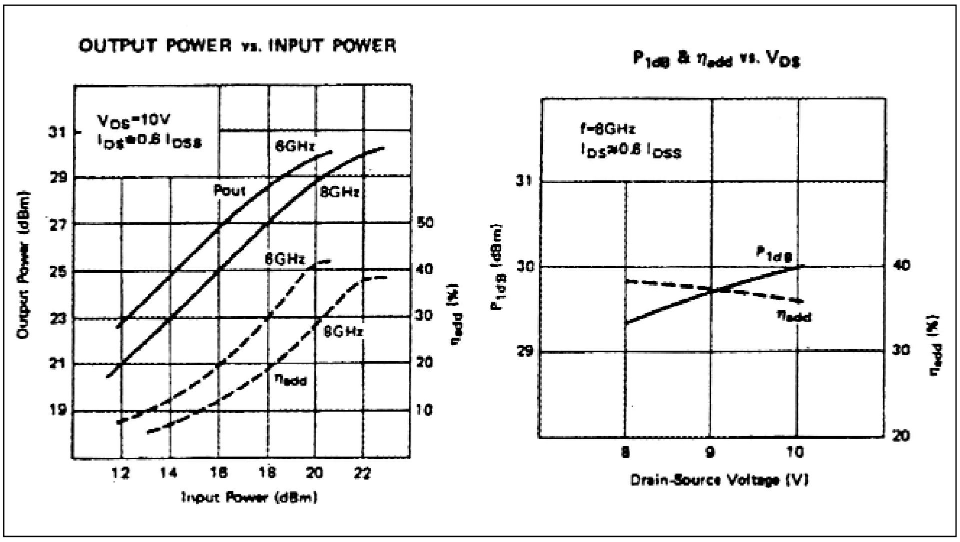


Fig 11.51: Data for the Fujitsu FSX53WF used in the 1W GaAsFET amplifier for 3cm

First, the no-signal currents are set as follows:

- For the FSX52WF at approximately 70mA, this corresponds to a voltage drop of 105mV, across a 1.5-ohm protective resistor. For the FLC103WG, at approximately 240mA, it corresponds to a voltage drop of 360mV, across a 1.5-ohm protective resistor.
- With 30mW drive at the desired frequency, an output of approximately 400mW (in the worst case) and of 1W (in the ideal case) should be measurable.

The "small disc method" is normally of assistance when tuning the amplifier.

You will need small discs, measuring about 2 - 4mm², a few toothpicks to press down and push and a lot of patience. Above all, the greatest care in watching out for short circuits will (hopefully) soon lead you to achieve full output power. After tuning, an aluminium cover plate 1mm thick can be fitted.

In the overwhelming majority of the power amplifiers measured, almost no influence from the cover could be detected.

Of course, there were just a few cases in which minimal self excitation were detected when the cover was fitted. This is caused by astonishingly stable housing resonance, slightly above the calibration frequency, with a few milliwatts of power at the output.

Even this undesirable oscillation disappeared with a low powered drive. A strip of absorbent material about 5mm wide and about 10mm long, glued to the inside of the cover in the area above the FSX52WF, provided a reliable remedy here.

A comparison of the output data from the semiconductors (**Figs 11.50 and 11.51**) with the readings from a typical power amplifier (**Figs 11.52 - 54**) makes clear how successful the project is in practice.

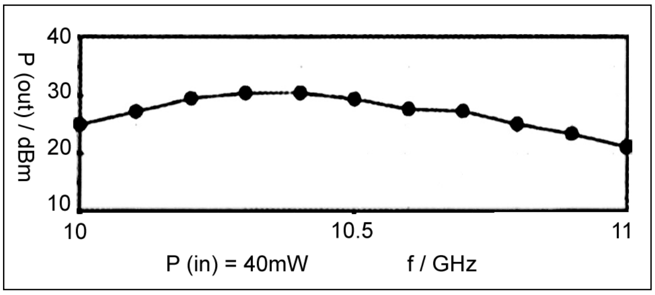


Fig 11.52: Power bandwidth of the 1W GaAsFET amplifier for 3cm

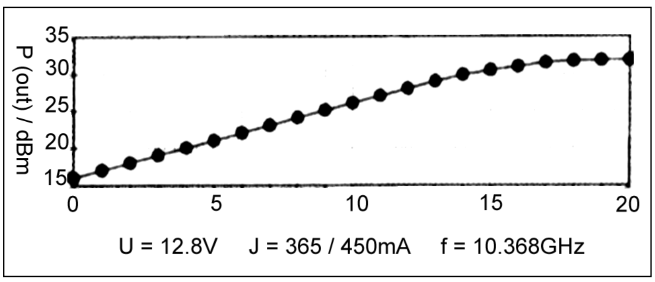


Fig 11.53: Input matching of the 1W GaAsFET amplifier for 3cm

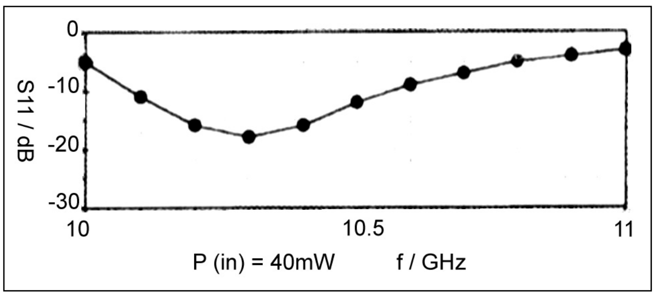


Fig 11.54: Linearity of the 1W GaAsFET amplifier for 3cm.

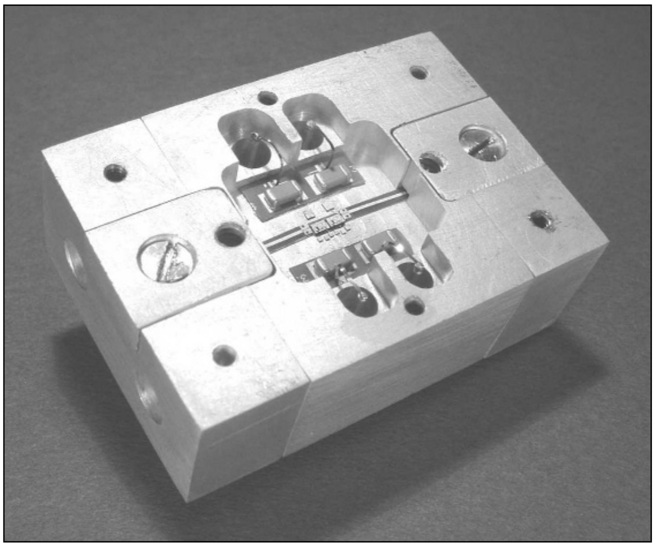


Fig 11.55: Picture of the completed 76GHz amplifier

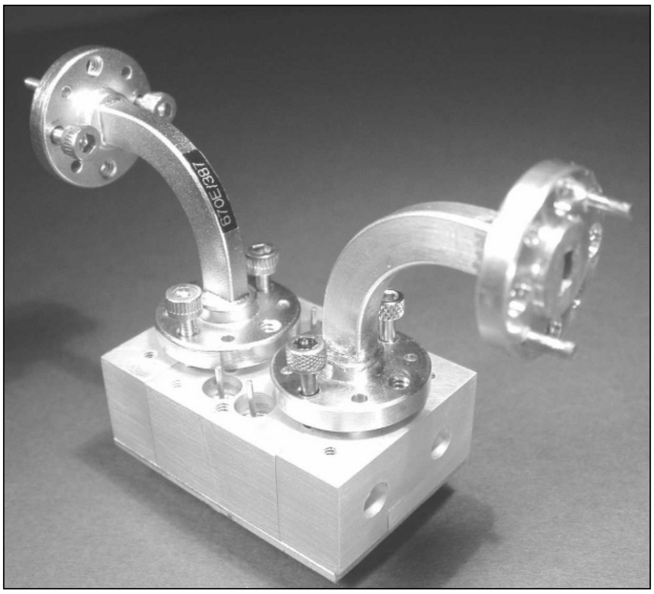


Fig 11.56: Picture of the completed 76GHz amplifier with WR-12 waveguide connections fitted.

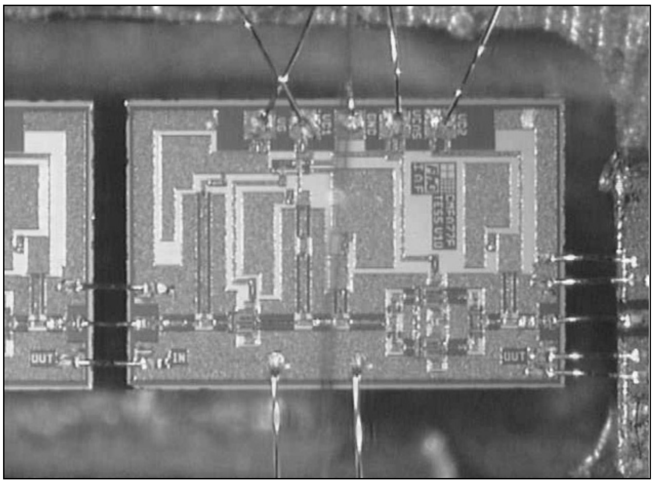


Fig 11.57: The 1AF-MPA7710 chip, used in the 76GHz amplifier, magnified about 120 times

76GHz Amplifier

This design [25] is part of a series of articles by Sigurd Werner, DL9MFV, published in *VHF Communications* magazine. It describes an amplifier that uses two MMIC's (IAF-MPA7710) connected in series, and originates from development work by the Fraunhofer Institute for Applied Solid-State Physics in Freiburg. The gain of the amplifier is 24dB at 76,088MHz.

The first difficulty lies in the procurement (selection would be something of an exaggeration) of suitable MMIC's for this frequency. Siemens manufacture a two-stage GaAs amplifier chip (T602B-MPA-2) for use in its car collision radar, which amplifies small signals by approximately 9.5dB [27]. One genuine alternative to this is an MMIC that has been developed at the Fraunhofer Institute. This is another two-stage amplifier chip (1.5mm x 1mm x 0.635mm), with the designation IAF-MPA7710. It has the following characteristics:

- Frequency range: 73 - 80GHz
- Gain: >11dB
- Output >14dBm at 1dB compression, and with a power consumption of approximately 800mW [28] and [29].

Since gate 1 is earthed, only two positive voltages of approximately 1.5V (for Gate 2) and approximately 4V each for all drain connections are required. In order to attain a usable level of amplification, two MMIC's are wired in series, without any additional matching circuit between the two chips.

The housing with dimensions of 27.5mm x 39.8mm x 13.5mm was milled from brass and subsequently gold-plated (**Fig 11.55**). The two chips were mounted in a 0.5mm hollow in the 4.1mm machined cavity. This balances out the difference in height between the MMIC's (0.635mm!) and the connection substrates (0.254mm). Strict attention was paid to ensure that the distance between the two MMIC's and their distance from the substrates were as small as possible (approximately 60 or 75μm). The connection substrates that link the chips with the two WR-12 waveguides (amplifier input and output) are made from aluminium nitride (approximately 1.1mm wide). They were sawn out of an existing ceramic, a substrate designed for other purposes by R&S. Since the chips have co-planar RF connections, the substrates likewise have a short co-planar section (0.5mm), which then goes into a 50-ohm stripline (8mm). This line projects approximately $\lambda/8$ into the waveguide. This construction technique is described in more detail in [30]. The ground plane of the ceramic projecting into the waveguide was milled off. The power supplies for the chips are initially blocked with 100pF single layer capacitors, and subsequently with 100nF ceramic capacitors, and are then fed out via feedthrough filters through the housing base (see **Fig 11.56**). The connections between chips and substrate (or capacitors) were created using wedge-wedge bond technology [30].

The following problem arose here: The RF connection pads of the MMICs, which were actually designed for flip-chip installation, are extremely small (see **Fig 11.57**). Directly behind these pads there are air bridges running to the chip circuit. These fragile structures are very easily caught and destroyed during the bonding by the tool that feeds the 17μm gold thread. The chip is then naturally unusable. This difficulty was avoided through the use of a still thinner needle and a correspondingly finer gold thread of 12μm. However, all other connections were created using a conventional 17μm thread.

The amplifier was initially operated at low power levels (approximately 50μW) at 76,032MHz. The gain observed was initially very disappointing (in the region of 8dB). Even after the fine adjustment of the waveguide short circuit screws the gain reading was scarcely 10dB. On the basis of experience of mis-

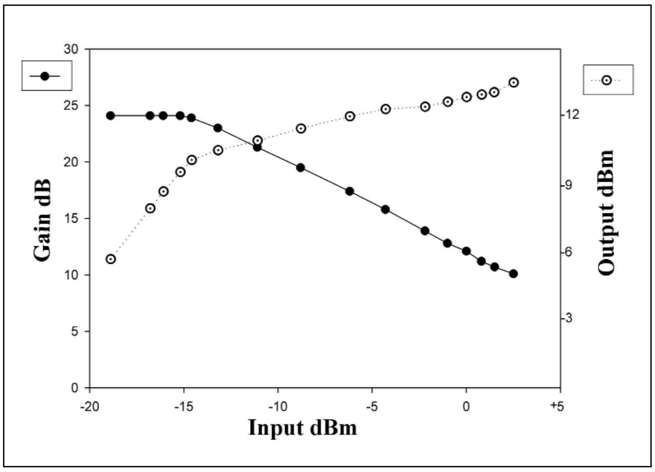


Fig 11.58: Transfer characteristics of the 76GHz amplifier at 76,032MHz.

matching of the waveguide couplings obtained during the transverter project, another series of gold threads was attached and fastened to the striplines using a UV activated adhesive (see [26]). The input power during matching amounted to -13dBm. After a laborious sequence of 9 pennants, the work was rewarded by an amplification of 24dB. That means 12dB per amplifier stage, a value which tallies well with the specifications in the data sheet [28]. A value of > 9dB (SWR < 2.1) was measured at the amplifier input, with > 25dB (SWR < 1.1) at the output. The gain and the output power for various input levels (f = 76,032 MHz) is shown in **Fig 11.58**. It can be seen that for an input power of > -15dBm the amplification is already decreasing, a behaviour to be expected. At an input power of 100μW, there is still 10mW measured at the output anyway (20dB gain). The saturation power of 12.6dBm remains unsatisfactory (approximately 18mW). Approximately 15dBm would have been expected! There could be several reasons for this behaviour. The large number of threads certainly increased the matching and thus the amplification, but at the same time a lot of energy is lost with each stub attached. Secondly, the length of the striplines at these frequencies leads to additional losses. The power supply voltages were varied in a further attempt to attain a higher saturation power on the individual MMIC's. The optimal setting yielded the following values:

- The driver chip, gate 2 +1.4V, drains 3.0V (190mA);
- The output chip, gate 2 + 1.2V, drains 3.7V (220mA).

The result was an increase in the power of only approximately 20%. The MMIC (IAF-MPA7710) can provide good service in the manufacture of a power amplifier. Better performance could be achieved by using at least two chips operated in parallel. Possible solutions for the addition of the outputs are "magic tee" or a 3/4λ Wilkinson coupler on a quartz substrate (0.127mm!). A project showing these techniques was published in [31].

MICROWAVE SSB TRANSCEIVER DESIGN [32]

When discussing SSB transceivers, the first question to be answered is probably the following - does it make sense to develop and build new SSB radios? Today SSB transceivers are mass produced items for VHF and UHF. Most radio amateurs are therefore using a base SSB transceiver (usually a commercial product) operating on a lower frequency and suitable receive and transmit converters or transverters to operate on 1296MHz or higher frequencies. The most popular base transceiver was the IC-202. All narrow band (SSB/CW) microwave activity is therefore historically concentrated in the first 200kHz of amateur microwave segments, eg 1296.000-1296.200, 2304.000-2304.200 etc, due to the limited frequency coverage of the IC-202.

Transverters should always be considered a poor technical solution for many reasons. Receive converters usually degrade the dynamic range of the receiver while transmit converters dissipate most of the RF power generated in the base SSB transceiver. Both receive and transmit converters generate a number of spurious mixing products that are very difficult to filter out, due to the harmonic relationships among the amateur frequency bands 144/432/1296MHz. However, the worst problem of most transverters is the breakthrough of strong signals in or out of the base transceiver intermediate frequency band. This prob-

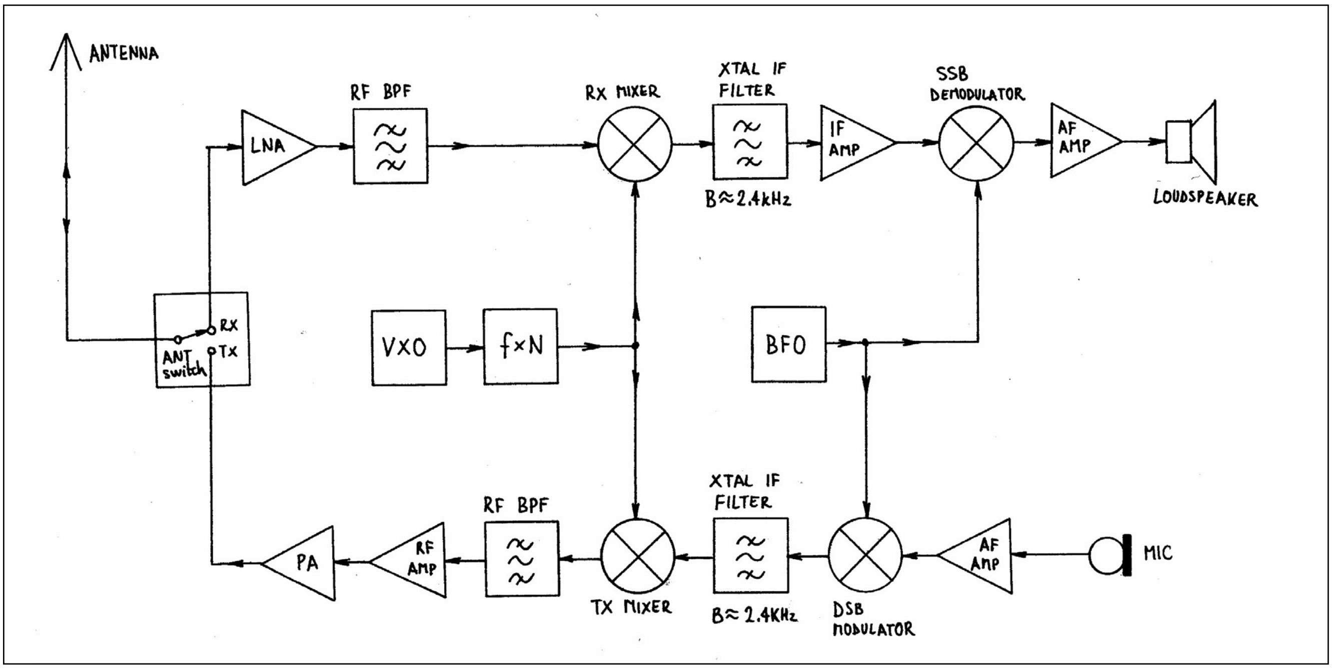


Fig 11.59: Block diagram of a conventional SSB transceiver

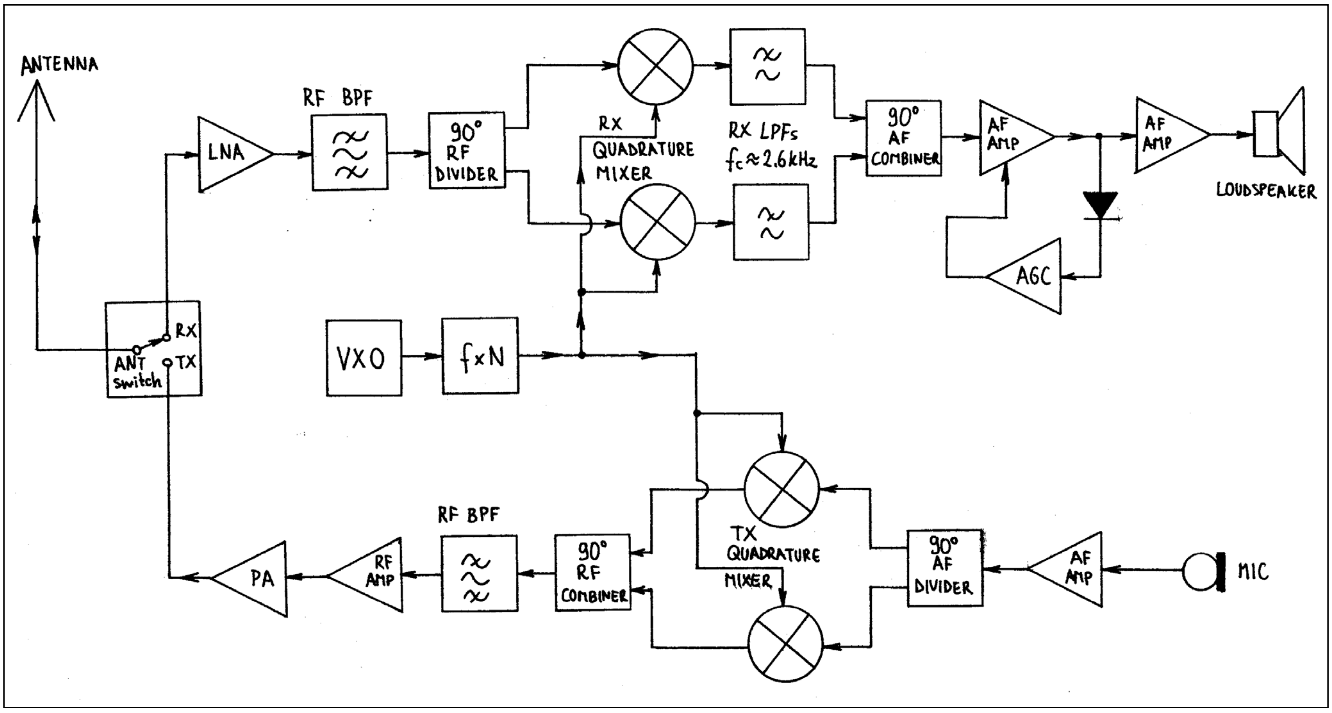


Fig 11.60: Block diagram of a direct conversion SSB transceiver

lem seems to be worst when using a 144MHz first IF. Strong 144MHz stations with big antenna arrays may break into the first IF, even at distances of 50 or 100km. Since the problem is reciprocal, a careless microwave operator may even establish two-way contacts on 144MHz although using a transverter and antenna for 1296MHz or higher frequencies.

Some microwave operators solved the above problem by installing a different crystal in the transverter; so that for example 1296.000MHz is converted to a less used segment around 144.700MHz. Serious microwave contesters use transverters with a first IF of 28MHz, 50MHz or even 70MHz to avoid this problem. Neither solution is cheap. The biggest problem is carrying a large 144MHz or HF all mode transceiver, together with a suitable power supply, on a mountaintop.

Even the IC-202 has its problems;. this radio has not been manufactured for several decades. However, there is now a growing number of lightweight transportable multimode transceivers.

As a conclusion, today it can still make sense to develop and build SSB radios for 1296MHz and higher frequencies. Since the problems of the transverters are well known and are not really new, different designers have already considered many technical solutions. Most solutions were discarded simply because they are, too complex, too expensive and too difficult to build, even when compared to the already complex combination of a base transceiver and transverter. Most commercial SSB transceivers include a modulator and a demodulator operating on a high IF, as shown in **Fig 11.59**. The resulting SSB signal is converted to the RF operating frequency in the transmitter and back to the IF in the receiver. Both the transmitter and the receiver use expensive components like crystal filters. Besides crystal filters, additional filtering is required in the RF section to attenuate image responses and spurious products of both receiving and transmitting mixers. The design of conventional (high IF) SSB transceivers dates back to the valve age, when active components (valves) were expensive and unreliable. Passive components like filters were not so critical. Complicated tuning procedures only represented a small fraction of the overall cost of a valve SSB transceiver. SSB crystal filters usually operate in the frequency range around 10MHz. A double or even triple up-conversion is required to reach microwave frequencies in the transmitter. On the other hand, a double or triple down-conversion is required in the receiver to get back to the crystal filter frequency. Commercial VHF/UHF SSB transceivers therefore save some expensive components by sharing some stages between the transmitter and the receiver. A conventional microwave SSB transceiver is therefore complicated and expensive. Building such a transceiver in amateur conditions is difficult at best. Lots of work and some microwave test equipment are required. The final result is certainly not cheaper and may not perform better than the familiar transverter plus base transceiver combination.

Fortunately, expensive crystal filters and complicated conversions are not essential components of a SSB transceiver. There are other SSB transceiver designs that are both cheaper and easier to build in amateur conditions. The most popular seems to be the direct conversion SSB transceiver design shown in **Fig 11.60**. A direct conversion SSB receiver achieves most of its gain in a simple audio frequency amplifier, while simple RC low pass filters achieve the selectivity.

The most important feature of a direct conversion SSB transceiver is that there are neither complicated conversions nor image frequencies to be filtered out. The RF section of a direct conversion SSB transceiver only requires simple LC filters to attenuate distant spurious responses like harmonics and sub harmonics. In a well-designed direct conversion SSB transceiver, the RF section may not require any tuning at all. The most important drawback of a direct conversion SSB transceiver is a rather poor unwanted sideband rejection. The transmitter includes two identical mixers operating at 90 degrees phase shift (quadrature mixer) to obtain only one sideband. The receiver also includes two identical mixers operating at 90 degrees phase shift to receive just one sideband and suppress the other sideband. A direct conversion SSB transceiver operates correctly only if the gain of both mixers is the same and the phase shift is exactly 90 degrees. It therefore includes some critical compo-

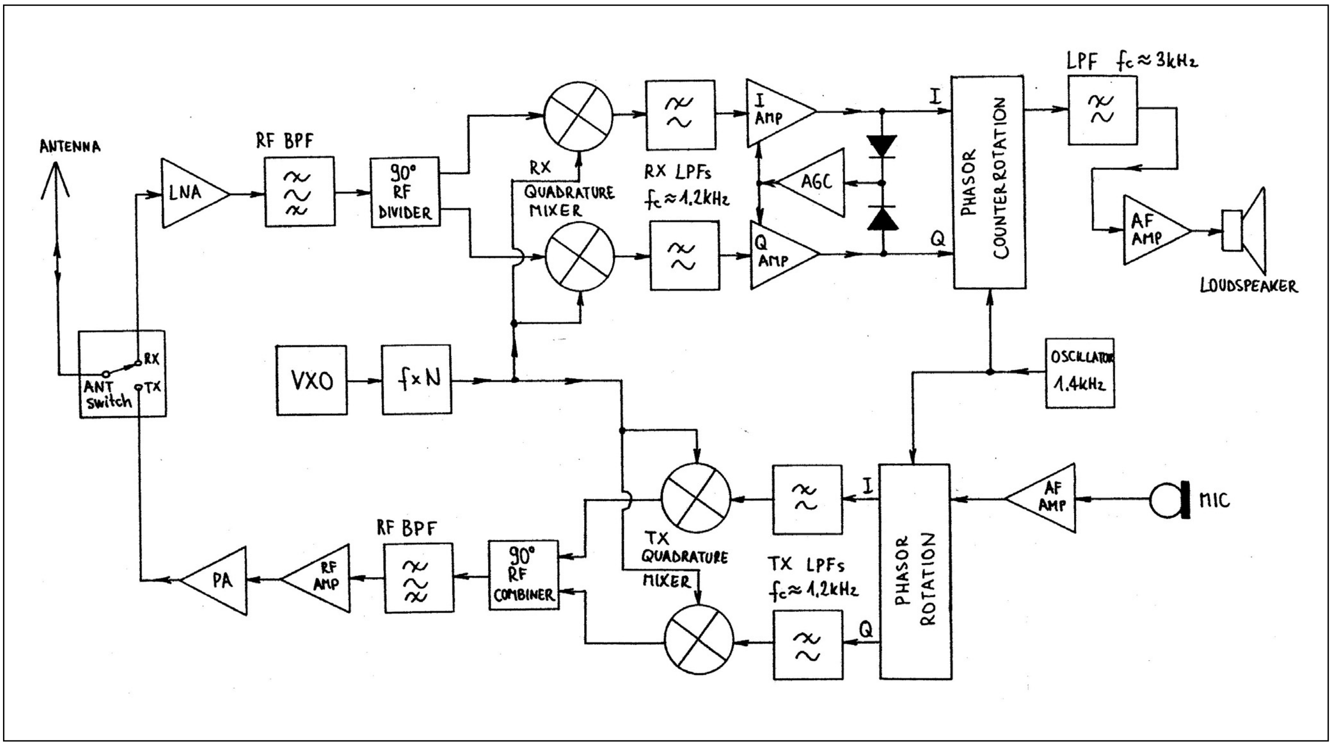


Fig 11.61: Block diagram of a Zero-IF SSB transceiver

nents like precision (1%) resistors, precision (2%) capacitors, selected or "paired" semiconductors in the mixers and complicated phase shifting networks. The most complicated part is usually the audio frequency 90 degree divider or combiner including several operational amplifiers, precision resistors and capacitors. Although using precision components, the unwanted sideband rejection will seldom be better than -40dB. This is certainly not enough for serious work on HF. In spite of these difficulties, direct conversion designs are quite popular among the builders of QRP HF transceivers. At frequencies above 30MHz it is increasingly more difficult to obtain accurate phase shifts. Due to the low natural (antenna) noise above 30MHz, a low noise RF amplifier is usually used to improve the mixer noise figure. An LNA may cause direct AM detection in the mixers. It may also corrupt the amplitude balance and phase offset of the two mixers, if the antenna picks up the local oscillator signal. A VHF direct conversion SSB transceiver is therefore not as simple as its HF counterpart.

On the other hand, a direct conversion SSB design has important advantages over conventional SSB transceivers with crystal filters, since there are no image frequencies and fewer spurious responses. Professional (military) SSB transceivers therefore use direct conversion, but the AF phase shifts are obtained by digital signal processing. The DSP uses an adaptive algorithm to measure and compensate any errors like amplitude unbalance or phase offset of the two mixers, to obtain a perfect unwanted sideband rejection. Additional AF signal processing also allows a different SSB transceiver design, for example a SSB transceiver with a zero IF as shown in **Fig 11.61**. The latter is very similar to a direct conversion transceiver except that the local oscillator is operating in the centre of the SSB signal spectrum, in other words at an offset of about 1.4kHz with respect to the SSB suppressed carrier frequency. In a zero IF SSB transceiver, the audio frequency band from 200Hz to 2600Hz is converted in two bands from 0 to 1200Hz. The low pass filters therefore have a cut-off frequency of 1200Hz, thus allowing a high rejection of the unwanted sideband. A zero IF SSB transceiver retains all of the advantages of a direct conversion design and solves the problem of the unwanted sideband rejection.

The quadrature IF amplifier of a zero IF SSB transceiver includes two conventional AF amplifiers. Since the latter are usually AC coupled, the missing DC component will be converted in the demodulator as a hole in the AF response around 1.4kHz. Fortunately this hole is not harmful at all for voice communications, since it coincides with a hole in the spectrum of the human voice. In fact, some voice communication equipment includes notch filters to create an artificial hole around 1.4kHz to improve the signal to noise ratio and/or to add a low baud rate telemetry channel to the voice channel. Thus, for voice communications, a potential drawback of a zero IF design is actually an advantage. Like a direct conversion transceiver, a zero IF SSB transceiver also requires quadrature transmit and receive mixers. However, amplitude unbalance or phase errors are much less critical, since they only cause distortion of the recovered audio signal. Conventional components, like 5% resistors, 10% capacitors and unselected semiconductors may be used anywhere in the transceiver.

Finally, a zero IF SSB transceiver does not require complicated phase shifting networks. Both the quadrature modulator in the transmitter and the quadrature demodulator in the receiver (phasor rotation and counter rotation with 1.4kHz) are made by simple rotating switches and fixed resistor / op-amp networks. CMOS analogue switches like the 4051 are ideal for this purpose, rotated by digital signals coming from a 1.4kHz oscillator. Although the block diagram of a zero IF SSB transceiver looks complicated, such a transceiver is relatively easy to build. In particular, very little (if any) tuning is required, since there are no critical components used anywhere in the transceiver. In particular, the RF section only includes relatively wideband (10%) band pass filters that require no tuning. The IF/AF section also accepts wide component tolerances and thus requires no tuning. The only remaining circuit is the RF local oscillator. The latter may need some tuning to bring the radio to the desired operating frequency.

Microwave SSB Transceiver Implementation

The described zero IF concept should allow the design of simple and efficient SSB transceivers for an arbitrary frequency band. Four successful designs of zero IF SSB transceivers covering the amateur microwave bands of 1.3, 2.3, 5.7 and 10GHz have been published by Matjaz Vidmar, S53MV, originally in *CQ ZRS* but also in [3], the basic design is given in this chapter.

Of course several requirements and technology issues need to be considered before a theoretical concept can materialise in a real world transceiver. Fortunately, the requirements are not severe for the lower amateur microwave bands. In this frequency range no very strong signals are expected, so there are no special requirements on the dynamic range of the receiver. Only a relatively limited frequency range needs to be covered (200 to 400kHz in each band) and this can be easily achieved using a VXO and multipliers as the local oscillator. From the technology point of view it is certainly convenient to use up to date components.

High performance and inexpensive microwave semiconductors were developed, first for satellite TV receivers and then for mobile communications like GSM or DECT telephones. These new devices provide up to 25dB of gain per stage up to 2.3GHz and up to 14dB of gain per stage up to 10GHz. Many other functions, like schottky mixer diodes or antenna switching PIN diodes are also available. Using obsolete components makes designs complicated. For example, the familiar transistors BFR34A and BFR91 were introduced almost 25 years ago. At that time they were great devices providing almost 5dB of gain at 2.3GHz. Today it makes more sense to use an INA-03184 MMIC to get 25dB of gain at 2.3GHz or in other words replace a chain of 5 obsolete transistor amplifier stages.

The availability of active components also influences the selection of passive components. Many years ago, all microwave circuits were built in waveguide technology. Waveguides allow very low circuit losses and high Q resonators. Semiconductor microwave devices introduced microstrip circuits built on low loss substrates like alumina (Al203) ceramic or glassfibre-teflon laminates. Conventional glassfibre-epoxy laminates like FR4 were not used above 2GHz due to the high losses and poor Q of microstrip resonators. However, a zero IF SSB transceiver design does not require a very high selectivity in the RF section. If the circuit losses can be compensated for by high gain semiconductor devices, cheaper substrates like the conventional glassfibre-epoxy FR4 can be used at frequencies up to at least 10GHz. The FR4 laminate has excellent mechanical properties. Unlike soft teflon laminates, cutting, drilling and hole plating in FR4 is well known. Even more important, most SMD component packages are designed for installation on a FR4 substrate and may break or develop intermittent contacts if installed on a soft teflon board. Therefore, losses in FR4 microstrip transmission lines and filters were investigated. Surprisingly, the losses were found inversely proportional to board thickness and rather slowly increasing with frequency. This simply means that the FR4 RF losses are mainly copper losses, while dielectric losses are still rather low. FR4 RF copper losses are high since the copper surface is made very rough to ensure good mechanical bonding to the dielectric substrate. In fact, if the copper foil is peeled off a piece of FR4 laminate, the lower foil surface is rather dark. On the other hand, if the copper foil is peeled off a piece of microwave teflon laminate, the colours of both foil surfaces are similar. Since different manufacturers use different methods for bonding the copper foil, RF losses are different in different FR4 laminates. On the other hand, the dielectric constant of FR4 was found quite stable. Finally, silver or gold plating of microstrip lines etched on FR4 laminate really makes no sense, since most of the RF losses are caused by the (inaccessible) rough foil surface bonded to the dielectric.

A practical FR4 laminate thickness for microwave circuits with SMD components is probably 0.8mm. A 50-ohm microstrip line has a width of about 1.5mm and about 0.2dB/cm of loss at 5.76GHz. Therefore microstrip lines have to be kept short if etched on FR4 laminate. For comparison, the FR4 microstrip losses are about three times larger than the microstrip losses of a glassfibre-teflon board and about ten times larger than the losses of teflon semi rigid coax cables.

Although FR4 laminate losses are high, resonators and filters can still be implemented as microstrip circuits. Considering PCB etching tolerances and especially under etching, both transmission lines and gaps in between them should not be made too narrow. A practical lower limit is 0.4mm width for the transmission lines and 0.3mm for the gaps.

As already mentioned, modern semiconductor devices are really easy to use even at microwave frequencies. Silicon MMIC amplifiers provide 25dB of gain (limited by package parasitics) up to 2.3GHz. If less gain is required, conventional silicon bipolar transistors can be used, since their input and output impedances are also close to 50 ohms. GaAs semiconductors are more practical above about 5GHz. In particular, high performance devices like HEMTs have become inexpensive since they have been mass produced for satellite TV receivers. HEMTs operate at lower voltages and higher currents than conventional GaAsFETs, so their input and output impedance are very close to 50 ohms at frequencies above 5GHz. Serious microwave engineers are afraid of using HEMTs since these devices have enough gain to oscillate at frequencies above 50GHz or even 100GHz. In this case it is actually an advantage to build the circuit on a lossy laminate like FR4,

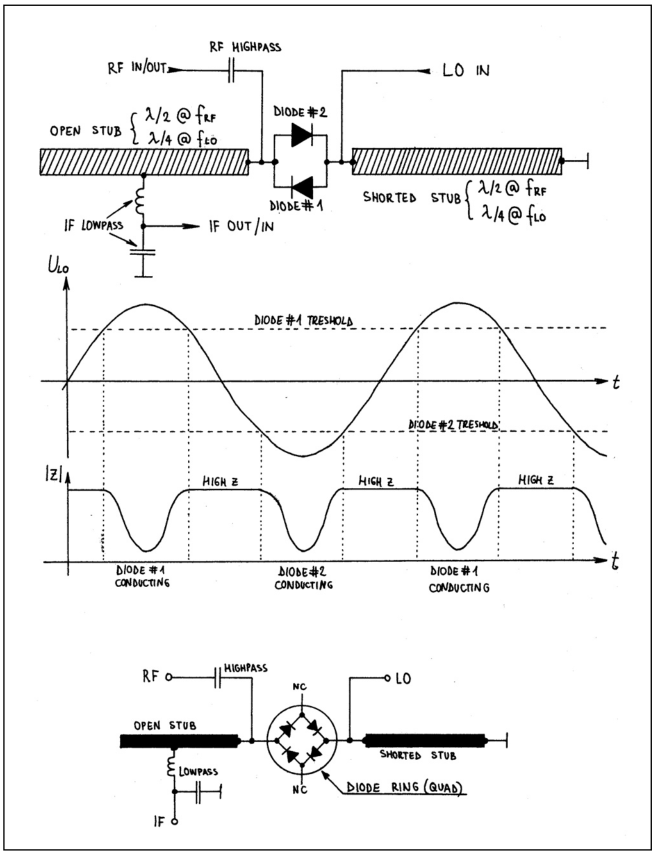


Fig 11.62: Sub harmonic mixer design

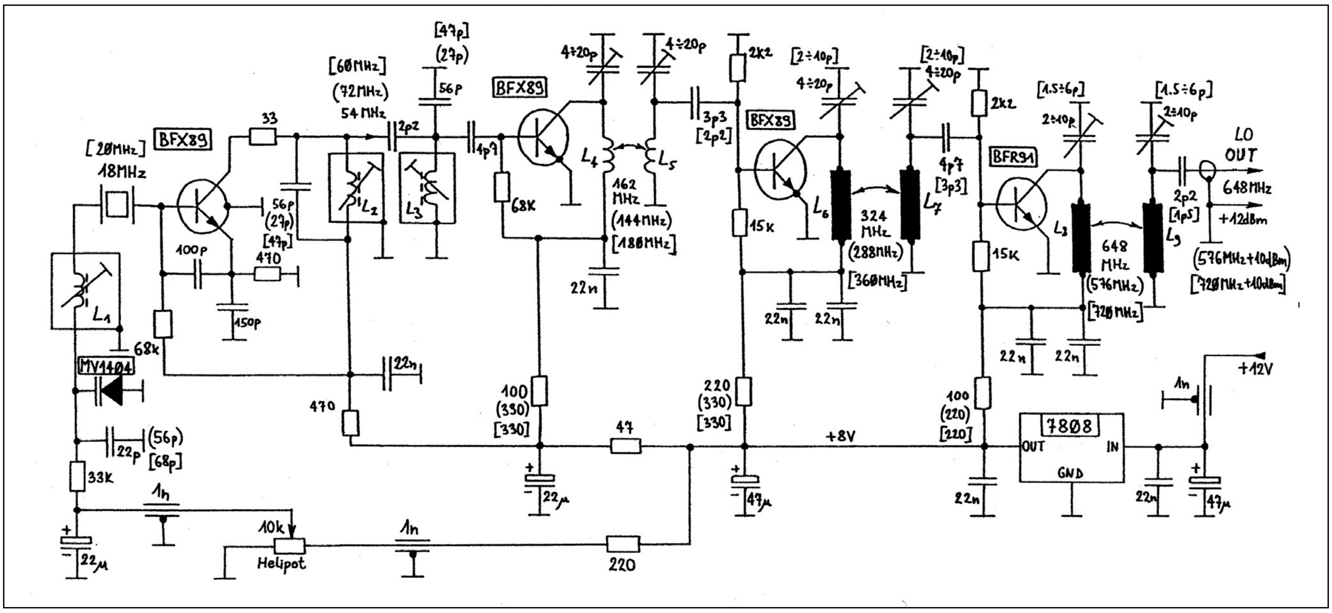


Fig 11.63: Circuit diagram of VXCO for Zero-IF transceiver

since the latter will efficiently suppress any oscillations in the millimetre frequency range. Having the ability to control the loss in a circuit therefore may represent an advantage! The availability of inexpensive power GaAsFETs greatly simplifies the construction of transmitter output stages. In particular, the high gain of power GaAsFETs in the 23cm and 13cm bands greatly reduces the number of stages when compared to silicon bipolar solutions.

Zero IF and direct conversion transceivers have some additional requirements for mixers. Mixer balancing is very important, both to suppress the unwanted residual carrier in the transmitter and to suppress the unwanted AM detection in the receiver.

At microwave frequencies, the simplest way of achieving good mixer balancing is to use a sub harmonic mixer with two anti parallel diodes as shown in **Fig 11.62**. Such a mixer requires a local oscillator at half frequency. Frequency doubling is achieved internally in the mixer circuit. A disadvantage of this mixer is a higher noise figure in the range 10 to 15dB and sensitivity to the LO signal level. Both a too low LO drive or a too high LO drive will further increase the mixer insertion loss and noise figure. On the other hand, the sub harmonic mixer only requires two non critical microstrip resonators that do not influence the balancing of the mixer. The best performances were obtained using schottky quads with the four diodes internally connected in a ring. The schottky quad BAT14-099R provides about -35dB of carrier suppression at 1296MHz and about -25dB of carrier suppression at 5760MHz with no tuning. A very important advantage of the sub harmonic mixer is that the local oscillator operates at half of the RF frequency. This reduces the RF LO crosstalk and therefore the shielding requirements in zero IF or direct conversion transceivers. A side advantage is that the half frequency LO chain requires fewer multiplier stages.

The common modules for the whole range of transceivers are described in this chapter. The modules for different frequency ranges can be found in [3]. Finally, an overview of the construction techniques is given in this chapter as well as shielding of the modules and integration of the complete transceivers.

VCXO and Multipliers

Since a relatively narrow frequency range needs to be covered, a VXO followed by multiplier stages is an efficient solution for the local oscillator. The VXO is built as a varactor tuned VCXO with a fundamental resonance crystal, since the frequency-pulling range of overtone crystals is not sufficient for this application. A fundamental resonance crystal has a lower Q and is less stable than overtone crystals, but for this application the performance is sufficient. Fundamental resonance crystals can be manufactured for frequencies up to about 25MHz. Therefore the output of the VCXO needs to be multiplied to obtain microwave frequencies.

Frequency multiplication can be obtained by a chain of conventional multipliers, including class C amplifiers and band pass filters or by a phase locked loop. Although the PLL requires almost no tuning and is easily reproducible, this solution was discarded for other reasons. A SSB transceiver requires a very clean LO signal, therefore the PLL requires buffer stages to avoid pulling the VCXO and/or the microwave VCO. Shielding and power supply regulation is also critical, making the whole PLL multiplier more complicated than a conventional multiplier chain. The circuit diagram of the VCXO and multiplier stages is shown in **Fig 11.63**. The VCXO is operating at around 18MHz in the transceivers for 1296MHz and 2304MHz, and at around 20MHz in the transceiver for 5760MHz.

All multiplier stages use silicon bipolar transistors BFX89 (BFY90) except the last stage with a BFR91. The module already supplies the required frequency of 648MHz for the 1296MHz version of the transceiver. In the 2304MHz version, the module supplies 576MHz by using different multiplication factors. The latter frequency is doubled to 1152MHz inside the transmit and receive mixer modules. In the 5760MHz version, the module supplies 720MHz and this frequency is further multiplied to 2880MHz in an additional multiplier module. Of course, the values of a few components need to be adjusted according to the exact operating frequency, shown in ( ) brackets for 2304MHz and in [ ] brackets for 5760MHz. The VCXO and multiplier chain are built on a single sided FR4 board with the dimensions of 40mm x 120mm as shown in **Fig 11.64**. (in Appendix B) The corresponding component location (for the 648MHz version) is shown in **Fig 11.65** (also in Appendix B). The exact value of L1 depends on the crystal used. Some parallel resonance crystals may even require the replacing of L1 with a capacitor. L2 and L3

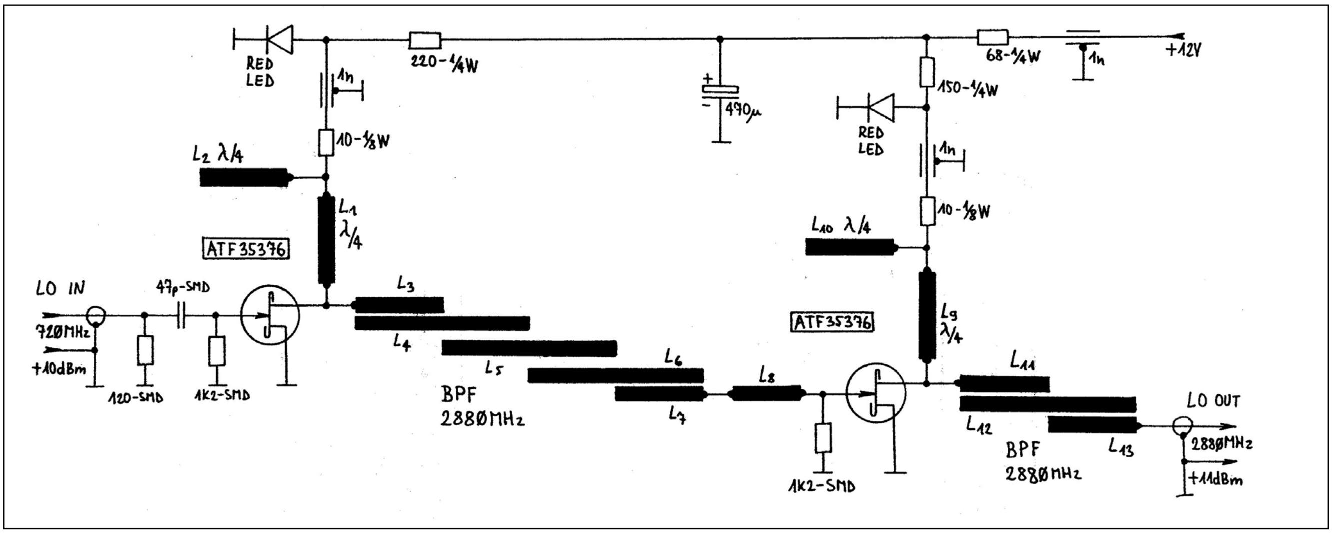


Fig 11.66: Circuit diagram of x4 multiplier to 2880MHz for Zero-IF transceiver

have about 150nH each or 4 turns of 0.25mm copper enamelled wire on a 10 x 10mm IF transformer coil former. L4 and L5 are self-supporting coils of 4 turns of 1mm copper enamelled wire each, wound on an internal diameter of 4mm. L6, L7, L8 and L9 are etched on the PCB. The VCXO module is the only part of the whole transceiver that requires tuning. L2, L3 and the capacitors in parallel with L4, L5, L6, L7, L8 and L9 should simply be tuned for the maximum output at the desired frequencies. In a multiplier chain, measuring the DC voltages over the base emitter junctions of the multiplier transistors can easily check RF signal levels.

When the multiplier chain is providing the specified output power, L1 and the capacitor in parallel with the MV1404 varactor should be set for the desired frequency coverage of the VCXO. If standard 'computer grade' 18.000MHz or 20.000MHz crystals are used, it is recommended to select the crystal with the smallest temperature coefficient. Unfortunately not all amateurs are allowed to use the international segment around 2304MHz on 13cm. It is a little bit more difficult to find a crystal for 18.125MHz for the German segment around 2320MHz. The 5760MHz transceiver requires an additional multiplier from 720MHz to 2880MHz as shown in **Fig 11.66**. The first HEMT ATF35376 operates as a quadrupler while the second HEMT ATF35376 operates as a selective amplifier for the output frequency of 2880MHz. The additional multiplier for 2880MHz is built on a double sided microstrip FR4 board with the dimensions of 20mm x 120mm as shown in **Fig 11.67** with the corresponding component location shown in **Fig 11.68** (both in Appendix B). The 2880MHz multiplier should provide the rated output power of +11dBm without any tuning. On the other hand,

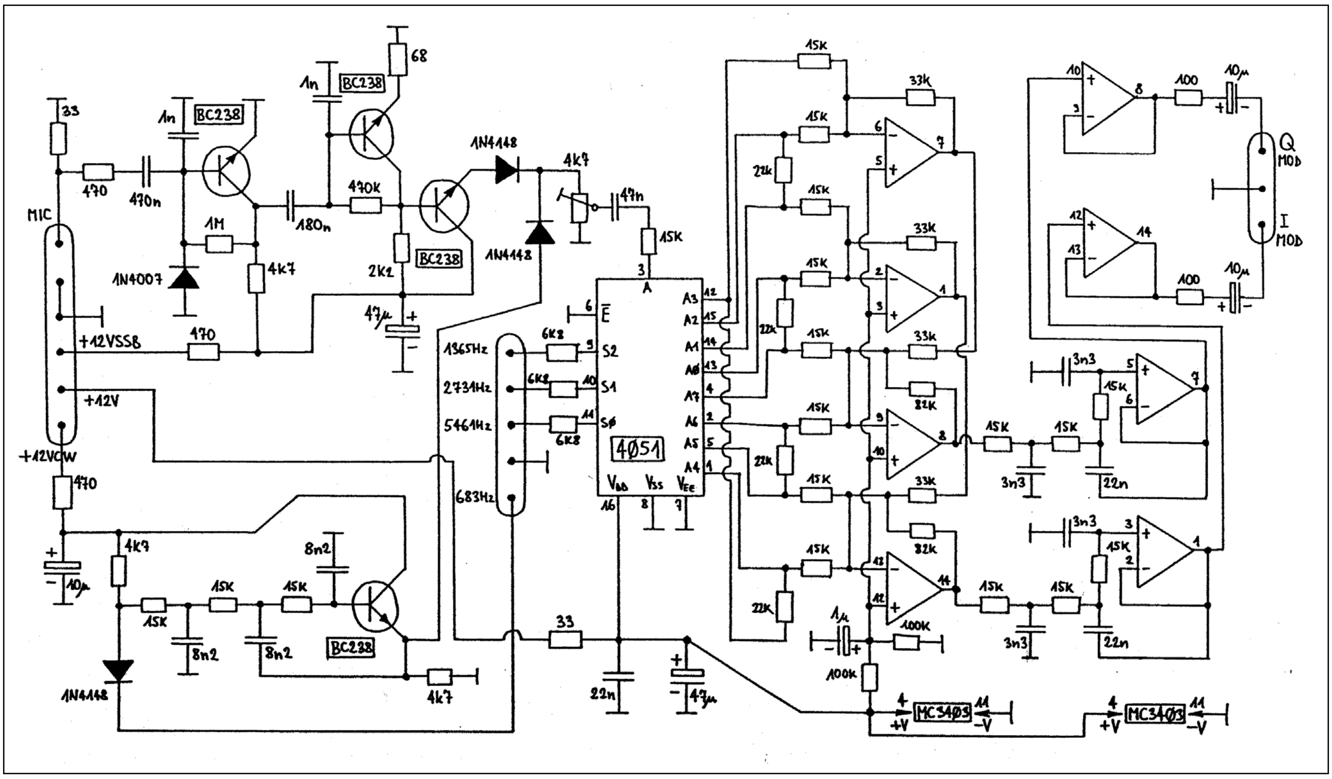


Fig 11.69: Circuit diagram of SSB/CW quadrature modulator for Zero-IF transceiver

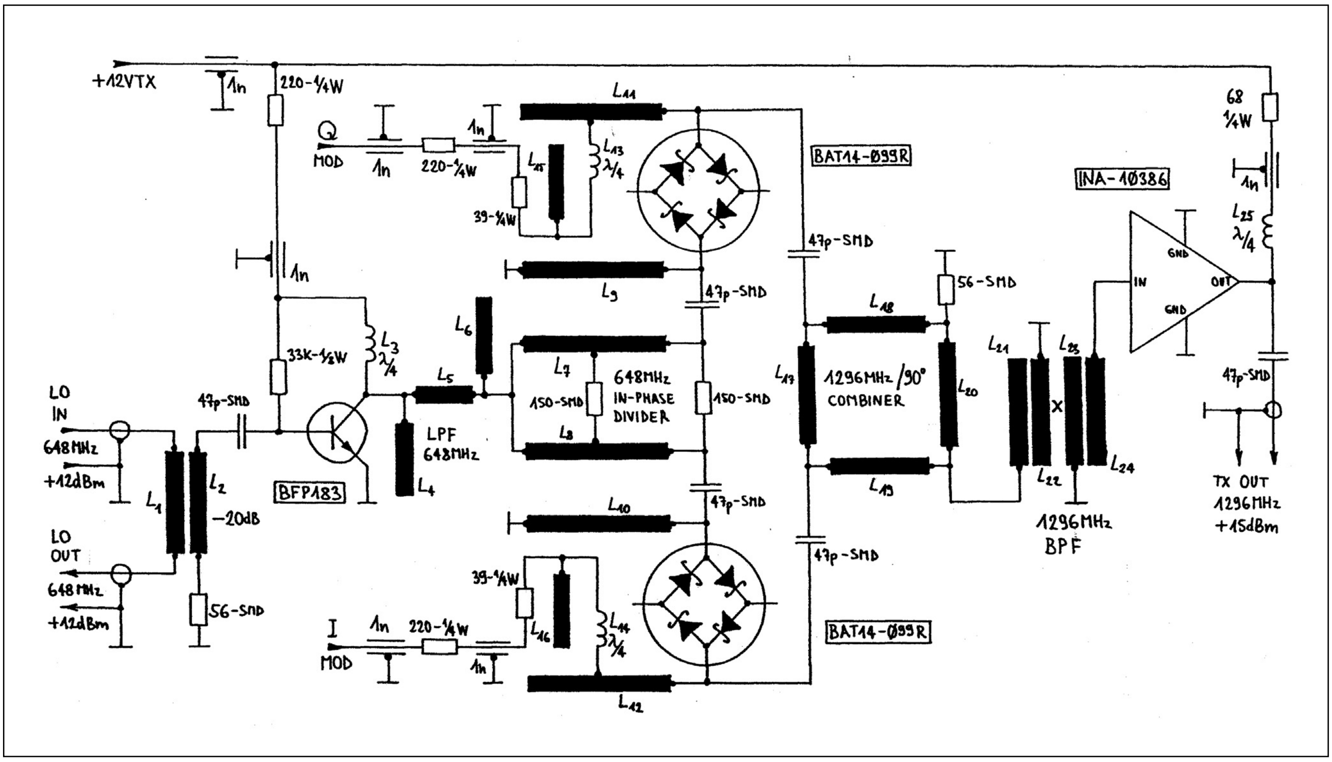


Fig 11.72: Circuit diagram of quadrature transmit modulator for 1296MHz Zero-IF transceiver

the tuning of L8 and L9 to 720MHz in the VCXO module can be optimised for the minimum DC drain current (max DC voltage) of the first HEMT. The two red LEDs are used as 2V Zeners. LEDs are in fact better than real Zeners, since they have a sharper knee and do not produce any avalanche noise.

SSB/CW Quadrature Modulator

The main purpose of the SSB/CW quadrature modulator is to convert the input audio frequency band from 200Hz to 2600Hz into two bands 0 to 1200Hz to drive the quadrature transmit mixer. Additionally the module includes a microphone amplifier and a circuit to generate the CW signal. The circuit diagram of the modulator module is shown in **Fig 11.69**. The microphone amplifier includes two stages with BC238 transistors. The input is matched to a low impedance dynamic mic with the 33-ohm resistor. The 1N4007 diode protects the input in the case the microphone input is simply connected in parallel to the loudspeaker output. Finally the output drives an emitter follower with another BC238.

The CW carrier is generated in the same way as the SSB transmission. The 683Hz square wave, coming from the demodulator module, is first cleaned in a low pass audio filter and then processed in the same way as a SSB signal. Both AF modulation sources are simply switched by 1N4148 diodes. The main component of the modulator is the 4051 CMOS analogue switch. The switch is rotated with the 1365Hz, 2731Hz and 5461Hz clocks coming from the demodulator. The input audio signal is alternatively fed to the I and Q chains. The I and Q signals are obtained with a resistor network and the first four op-amps (first MC3403). Then both I and Q signals go through low pass filters to remove unwanted mixing products. Finally there are two voltage followers to drive the quadrature transmit mixer.

The SSB/CW quadrature modulator is built on a single sided 40mm x 120mm FR4 board as shown in **Fig 11.70**, with the corresponding component location shown in **Fig 11.71** (both are in Appendix B). Most components are installed vertically to save board space. The SSB/CW quadrature modulator does not require any alignment. The 4.7k-ohm trimmer is provided to check the overall transmitter. Full power (in CW mode) should be obtained with the trimmer in the central position.

Quadrature Transmit Mixers

All three transmit mixer modules for 1296MHz, 2304MHz and 5760MHz include similar stages: an LO signal switching, an in-phase LO divider, two balanced sub harmonic mixers, a quadrature combiner and a selective RF amplifier. LO signal switching between the transmit and receive mixers is performed in the following way:

Most of the LO signal is always fed to the receive mixer. A small fraction of the LO signal is obtained from a coupler and amplified to drive the transmit mixer. During reception the power supply of the LO amplifier stage is simply turned off. This solution may look complicated, but in practice it allows an excellent isolation between the transmit and receive mixers. The practical circuit is simple and the component count is low as well. The circuit diagram of the quadrature transmit mixer for 1296MHz is shown in **Fig 11.72**. The 648MHz LO signal is taken from a -20dB coupler and the LO signal level is restored by the BFP183 amplifier stage, feeding two sub harmonic mixers equipped with BAT14-099R schottky quads. The 648MHz low pass attenuates the second harmonic at 1296MHz to avoid corrupting the symmetry of the mixers. The two 1296MHz signals are combined in a quadrature hybrid, followed by a 1296MHz band pass filter. The latter removes the 648MHz LO as well as other unwanted mixing products. After filtering, the 1296MHz SSB signal level is rather low (around -10dBm), so an INA-10386 MMIC is used to boost the output signal level to about +15dBm. The quadrature transmit mixer for 1296MHz is built on a double-sided microstrip FR4 board with dimensions 40mm x 120mm as shown in **Fig 11.73**,.with the corresponding component location shown in **Fig 11.74** (both are in Appendix B). The circuit does not require any tuning for operation at 1296MHz or 1270MHz.

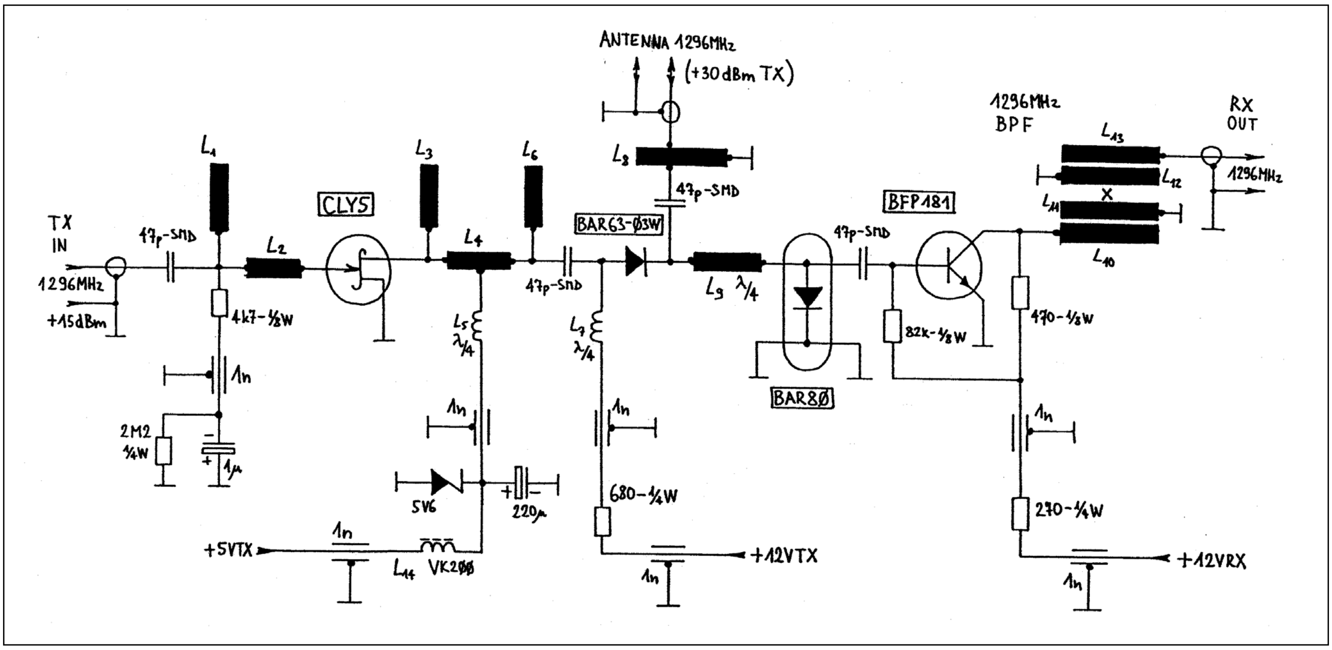


Fig 11.75: Circuit diagram of RF front end for 1296MHz Zero-IF transceiver

RF Front Ends

The RF front ends include the transmitter power amplifiers, the receiver low noise amplifiers and the antenna switching circuits. Of course there are major differences among different power amplifier designs, depending not just on the frequency, but also on the technology used and the output power desired. It no longer makes sense to use expensive coaxial relays, since PIN diodes can provide the same insertion loss and isolation at lower cost with better reliability and much shorter switching times.

The circuit diagram of the RF front-end for 1296MHz is shown in **Fig 75**. The transmitter power amplifier includes a single stage with a CLY5 power GaAsFET, providing a gain of 15dB and an output power of about 1W (+30dBm). The CLY5 is a low-voltage transistor operating at about 5V.

The negative gate bias is generated by rectification of the driving RF signal in the GS junction inside the CLY5 during modulation peaks. The gate is then held negative for a few seconds thanks to the 1µF capacitor. To prevent overheating and destruction of the CLY5, the +5VTx voltage is obtained through a current-limiting resistor. This arrangement may look strange, but it is very simple, requires no adjustments, allows a reasonably linear operation and most important of all, it proved very reliable in PSK packet-radio transceivers operating 24 hours per day in a packet-radio network.

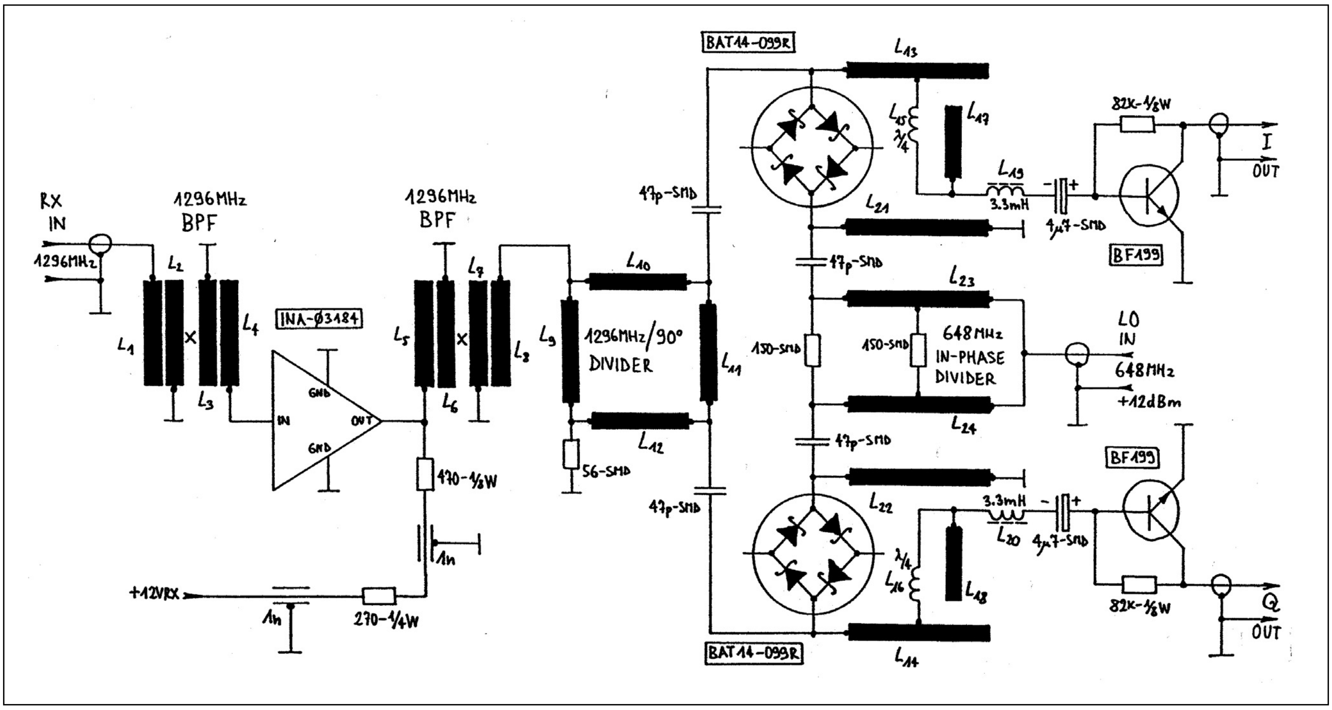


Fig 11.78: Circuit diagram of quadrature receive mixer for 1296MHz Zero-IF transceiver

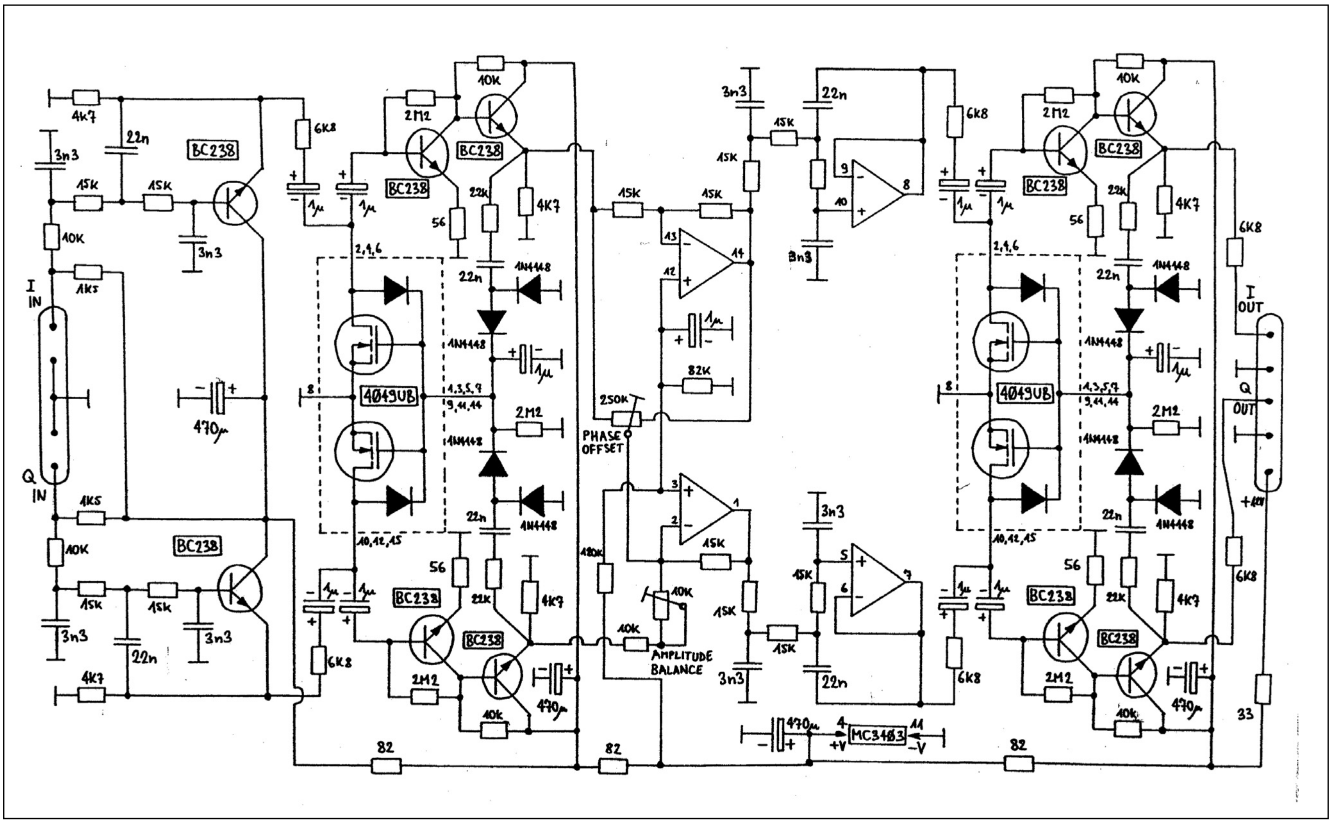


Fig 11.81: Circuit diagram of quadrature receive SSB IF amplifier for 1296MHz Zero-IF transceiver

The antenna switch includes a series diode BAR63-03W and a shunt diode BAR80. Both diodes are turned on while transmitting. L9 is a quarter wavelength line that transforms the BAR80 short circuit into an open for the transmitter. The receiving preamplifier includes a single BFP181 transistor (15dB gain) followed by a 1296MHz band pass filter (-3dB loss).

In the 1296MHz RF front-end, the LNA gain should be limited to avoid interference from powerful non-amateur users of this band (radars and other radio navigation aids). The RF front-end for 1296MHz is built on a double-sided microstrip 40mm x 80mm FR4 board as shown in Fig 11.76, with the corresponding component location shown in Fig 11.77 (both are in Appendix B). The RF front end for 1296MHz requires no tuning. However, since the output impedance of the INA-10386 inside the transmit mixer is not exactly 50 ohms, the cable length between the transmit mixer and the RF front-end is critical. Therefore L1 may need adjustments if the teflon-dielectric cable length is different from 12.5cm.

Quadrature Receive Mixers

All receiver mixer modules include similar stages, an additional RF signal amplifier, a quadrature hybrid divider, two sub harmonic mixers, an in phase LO divider and two IF preamplifiers. The mixers, in phase and quadrature dividers and RF band pass filters are very similar to those used in the transmitting mixer modules. The circuit diagram of the quadrature receiving mixer for 1296MHz is shown in Fig 11.78.

The incoming RF signal is first fed through a microstrip band pass filter, then amplified with an INA-03184 MMIC and further filtered by another, identical microstrip band pass. The total gain of the chain of the two filters and the MMIC is about 20dB. A high gain in the RF section is required to cover the relatively high noise figure of the two sub harmonic mixers and the additional losses in the quadrature hybrid. The two receiving sub harmonic mixers are also using BAT14-099R schottky quads. The mixer outputs are fed through low pass filters to the IF preamplifiers. The IF preamplifiers use BF199, HF transistors. These were found to perform better than their BCxxx counterparts in spite of the very low frequencies involved (less than 1200Hz). HF transistors have a smaller current gain, their input impedance is therefore smaller and better matches the output impedance of the mixers. Both IF preamplifiers receive their supply voltages from the IF amplifier module. The quadrature receiving mixer for 1296MHz is built on a double-sided microstrip FR4 board, 40mm x 120mm, as shown in Fig 11.79 with the corresponding component location shown in Fig 11.80 (both are in Appendix B). The receiving mixer for 1296MHz requires no tuning.

SSB Zero IF Amplifier with AGC

The basic feature of direct conversion and zero IF receivers is to achieve most of the signal gain with simple and inexpensive AF amplifiers. Further, the selectivity is achieved with simple RC low pass filters that require no tuning. The circuit diagram of such an IF amplifier equipped with AGC is therefore necessarily different from conventional high IF amplifiers. A zero IF receiver requires a two channel IF amplifier, since both I and Q channels need to be amplified independently before demodulation. The two IF channels should be as near identical as possible to preserve the amplitude ratio and phase offset between the I and Q signals. Therefore, both channels should have a common AGC so that the amplitude ratio remains unchanged. The circuit diagram of the quadrature SSB IF amplifier with AGC is shown in Fig 11.81.

The IF amplifier module includes two identical low pass filters on the input, followed by a dual amplifier stage with a common AGC. An amplitude/phase correction is performed after the first amplifier stage, followed by another pair of low pass filters and another dual amplifier stage with a common AGC. The two input low pass filters are active RC filters using BC238 emitter follow-

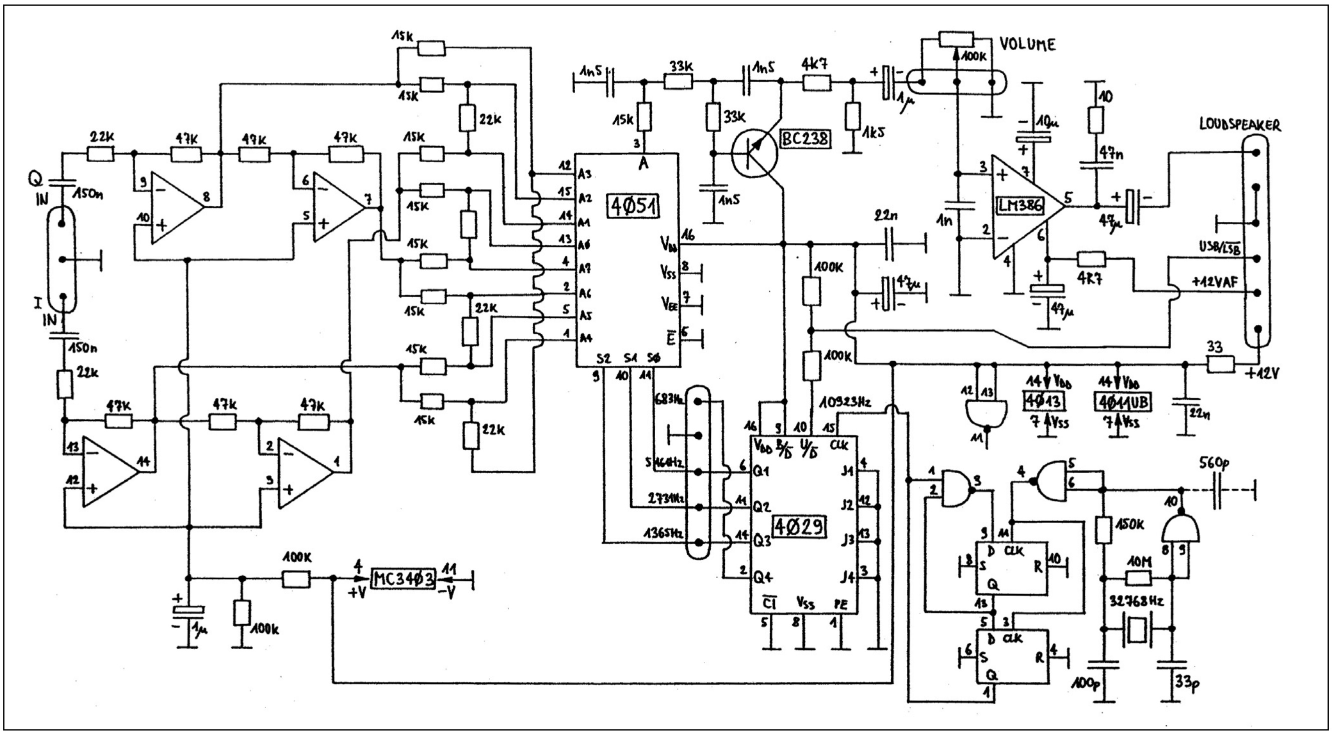


Fig 11.84: Circuit diagram of quadrature SSB demodulator and AF amplifier for 1296MHz Zero-IF transceiver

ers. Discrete bipolar transistors are used because they are much less noisy than operational amplifiers. The input circuit also provides the supply voltage to the IF preamplifiers inside the receiving mixer module through the 1.5k-ohm resistors.

The dual amplifier stages are also built with discrete BC238 bipolar transistors. Each stage includes a voltage amplifier (first BC238) followed by an emitter follower (second BC238) essentially to avoid mutual interactions when the amplifiers are chained with other circuits in the IF strip. The AGC uses MOS transistors as variable resistors on the inputs of the dual amplifier stages. To keep the gain of both I and Q channels identical, both MOS transistors are part of a single integrated circuit 4049UB. The digital CMOS integrated circuit 4049UB is used in a rather uncommon way; however the remaining components inside the 4049UB act just as diodes and do not disturb the operation of the AGC.

The IF amplifier module includes two trimmers for small corrections of the amplitude balance (10kΩ) and phase offset (250kΩ) between the two channels. The correction stage is followed by two active RC low pass filters. These use MC3403 operational amplifiers since the signals are already large enough and the op-amp noise is no longer a problem. Finally there is another, identical dual amplifier stage with its own AGC. The quadrature SSB IF amplifier is built on a single sided 50mm x 120mm FR4 board, as shown in **Fig 11.82**. with the corresponding component location shown in **Fig 11.83** (both are in Appendix B).

In order to keep the differences between the I and Q channels small, good quality components should be used in the IF amplifier. Using 5% resistors, 10% foil type capacitors and conventional BC238B transistors should keep the differences between the two channels small enough for normal operation. Most components are installed vertically to save board space. The amplitude balance (10kΩ) and phase offset (250kΩ) trimmers are initially set to their neutral (central) position. These are only used while testing the complete receiver to obtain the minimum distortion of the reproduced audio signal.

Quadrature SSB Demodulator and AF Amplifier

The main function of the quadrature SSB demodulator is the conversion of both I and Q IF signals (frequency range 0 to 1200Hz) back to the original audio frequency range: 200-2600Hz. The same module includes a power AF amplifier and a clock generator for both the phasor rotation in the transmitter and the phasor counter rotation in the receiver. The circuit diagram of the module is shown in **Fig 11.84**. The quadrature SSB demodulator includes four operational amplifiers (MC3403) to produce an 8-phase system from the I and Q signals, using a resistor network similar to that used in the modulator. The CMOS analogue switch, 4051, performs the signal demodulation or phasor counter rotation, rotating with a frequency of 1365Hz. The I and Q signals are alternatively fed to the output, or in other words the circuit performs exactly the opposite operation of the modulator. Unwanted mixing products of the phasor counter rotation are removed by an active RC low pass (BC238). The demodulated audio signal is fed to the 100kΩ volume control. A LM386 is used as the audio power amplifier due to its low current drain and small external component count.

The three clocks required to rotate both 4051 switches in the modulator and in the demodulator are supplied by a binary counter 4029. The 4029 includes an up/down input that allows the generation/demodulation of USB or LSB in this application. The up/down input has a 100kΩ pull up resistor for USB operation. LSB is obtained when the up/down input is grounded through a front panel switch. USB/LSB switching is usually not required for terrestrial microwave work; it is needed when operating through satellites or terrestrial linear transponders, or when using inverting converters or transverters for other frequency bands. Finally, USB/LSB switching may be useful to attenuate interference during CW reception.

An alternative way to switch sidebands is interchanging the I and Q channels. When assembling the transceiver modules together, it is therefore important to check the wiring so that the

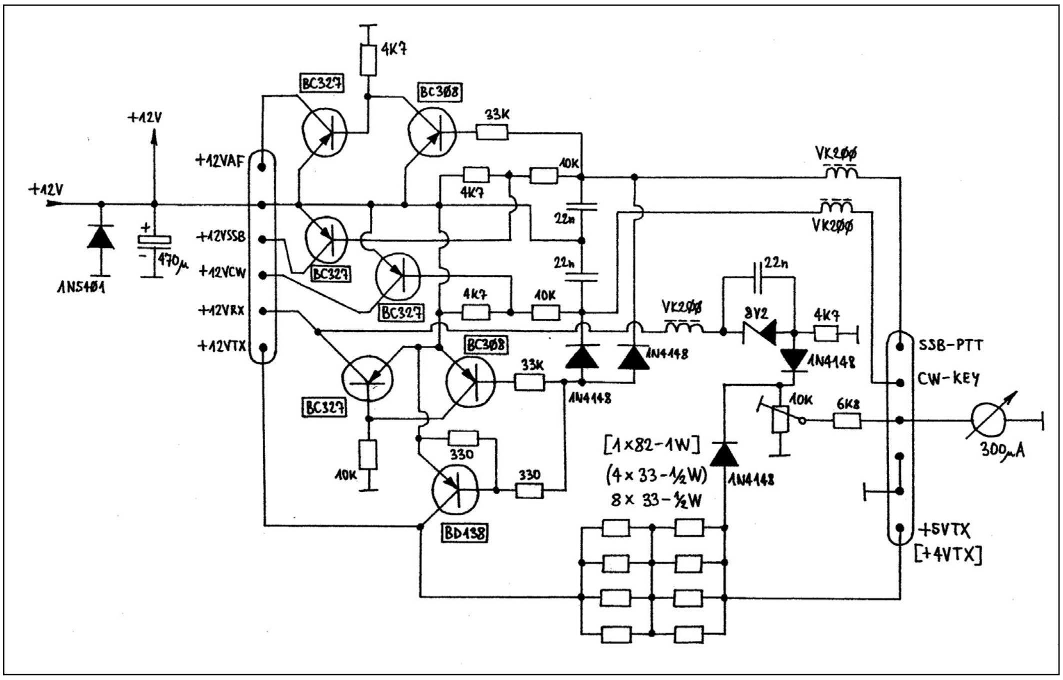


Fig 11.87: Circuit diagram of SSB/CW transmit/receive switching for Zero-IF transceiver

transmitter and the receiver operate on the same sideband at the same time.

The 4029 counter requires an input clock around 11kHz. This clock does not need to be particularly stable and an RC oscillator could be sufficient. In the transceivers described, a crystal source was preferred to avoid any tuning. In addition, if all transceivers use the same rotation or counter rotation frequency, the mutual interference is reduced. The oscillator uses a clock crystal, operating on a relatively low frequency of 32768Hz. The dual D flip-flop 4013 divides this frequency by 3 to obtain a 10923Hz clock for the 4029 binary counter. The resulting rotation frequency for the 4051 switches is 1365Hz. This almost perfectly matches the hole in the frequency spectrum of human voice. The same 4029 counter also supplies the CW tone, 683Hz, to reduce unwanted mixing products in the transmitter.

The quadrature SSB demodulator and AF amplifier are built on a single sided 40mm x 120mm FR4 board, as shown in **Fig 11.85** with the corresponding component location shown in **Fig 11.86**. (both are in Appendix B). Most of the components are installed vertically to save board space. The 32768Hz crystal oscillator will only operate reliably with a 4011UB (or 4001UB) integrated circuit. The commonly available "B" series CMOS integrated circuits (4011B in this case) have a too high a gain for this application. In the latter case, a 560pF capacitor may help to stabilise the oscillator. On the other hand, the oscillator circuit usually works reliably with old 4011 or 4001 circuits with an "A" suffix or no suffix letter at all.

SSB/CW Switching Rx/Tx

A SSB/CW transceiver requires different switching functions. Fortunately both SSB and CW modes of operation require the same functions in the receiver. Of course, two different operating modes are required for the transmitter: SSB Voice and CW Keying.

The Rx/Tx changeover is controlled by the PTT switch on the microphone in the SSB mode. On CW, most transceivers use an automatic delay circuit to keep the transmitter enabled during keying. This delay was perhaps required in old radios using several mechanical relays. In modern transceivers with all electronic switching it makes no sense, since the Rx/Tx switching can be performed in less than one millisecond. It therefore makes sense that SSB transmit is enabled by simply pressing the PTT switch, while CW transmit is enabled by pressing the CW key. No special (and useless) controls are required on the front panel of the transceiver. On CW, no delays are required and the receiver is enabled immediately after the CW key is released (full break-in). The circuit diagram of the switching is shown in **Fig 11.87**.

In the transceivers described, most modules are enabled at all times with a continuous +12V supply to VCXO and multipliers, receiving mixer, IF amplifier, demodulator and modulator.

When enabling the transmitter, either by pressing the PTT or Morse key, the Rx LNA is turned off (+12VRX) and the Tx PA is turned on (+12VTX and +5VTX or +4VTX). During SSB transmission the receiver AF amplifier is turned off (+12VAF), to avoid disturbing the microphone amplifier (+12VSSB). On the other hand, during CW transmission the AF amplifier and most receiver stages remain on, so that the keying can be monitored in the loudspeaker or 'phones. The +12VCW supply connects the 683Hz signal to the modulator input.

The supply voltages +12VAF, +12VSSB, +12VCW and +12VRX are switched by BC327 PNP transistors. Due to the higher current drain, the +12VTX supply voltage requires a more powerful PNP transistor BD138. The transmitter PA receives its supply voltage through a current limiting resistor from the +12VTX line. Since the latter dissipates a considerable amount of power, it is built from several smaller resistors and located in the switching unit to prevent heating the PA transistor(s). The value of the current limiting resistor depends on the version of the transceiver. The 1296MHz PA with a CLY5 requires eight 33-ohm half-watt resistors for a total value of 16.5 ohms. The 2304MHz PA with a CLY2 requires four 33-ohm half-watt resistors, for a total value of 33 ohms. Finally, the 5760MHz PA with two ATF35376s requires a single 82-ohm one watt resistor.

The switching module also includes the circuits to drive the front panel meter. The latter is a moving coil type with a full scale sensitivity of about 300µA. The meter has two functions. During reception it is used to check the battery voltage. The 8V2 Zener extends the full scale of the meter to the interesting range from about 9V to about 15V. During transmission the meter is used to check the supply voltage of the PA transistor(s). Due to the self biasing operation, the PA voltage will be only 0.5 - 1V without modulation and will rise to its full value, limited by the Zener diode inside the PA, only when full drive is applied. The operation of the PA and the output RF power level can therefore be simply estimated from the PA voltage.

An S-meter is probably totally useless in small portable transceivers as those described here. If desired, the AGC voltage can

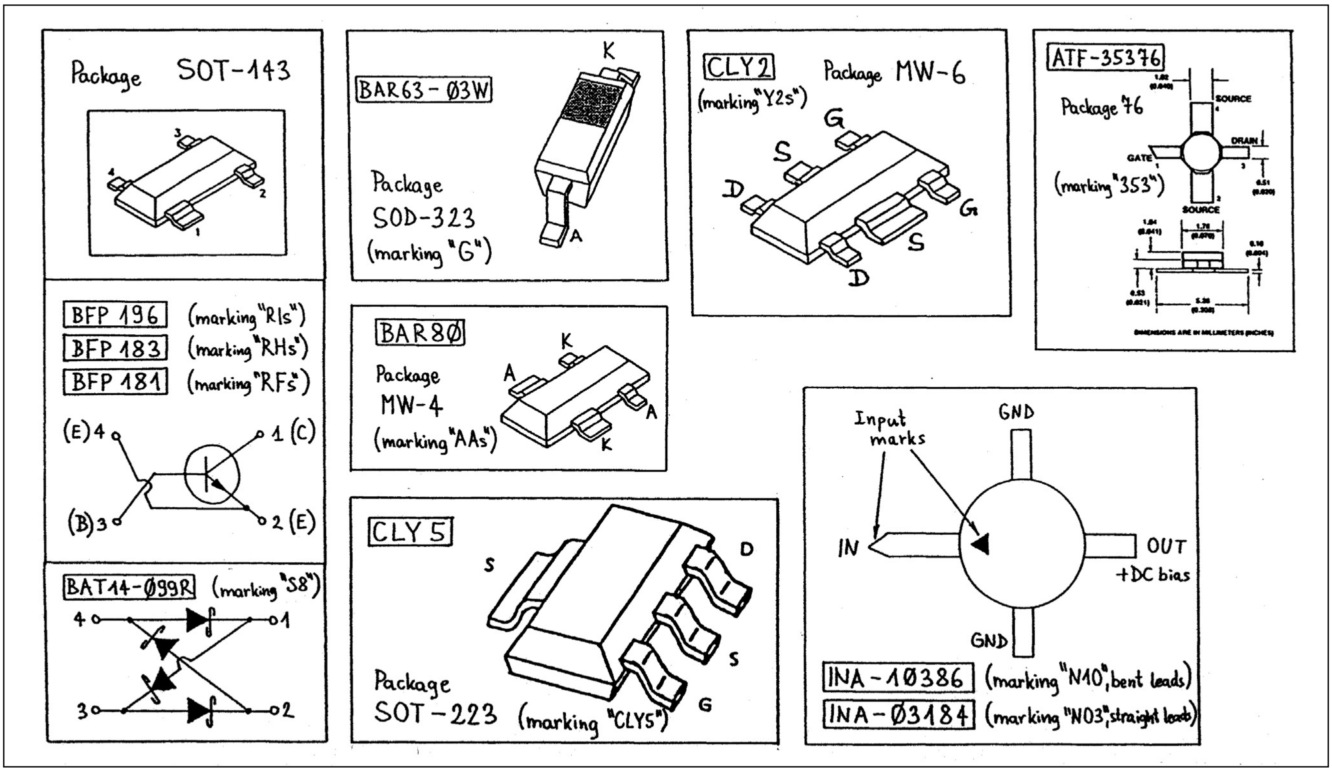


Fig 11.90: SMD semiconductor packages and pin-outs

be amplified and brought to a front panel meter. However, LED indicators are not visible in full sunshine on a mountaintop, so the choice is limited to moving-coil and LCD meters.

Most components of the SSB/CW switching are installed on a single sided FR4 board, 30mm x 80mm, as shown in **Fig 11.88** with their location (1296MHz version) shown in **Fig 11.89** (both diagrams are in Appendix B). Only the reverse polarity protection diode 1N5401 and the 470µF electrolytic capacitor are installed directly on the 12V supply connector. The 10k-ohm trimmer is used to adjust the meter sensitivity.

Construction of Zero IF SSB Transceivers

The described zero IF SSB transceivers use many SMD parts in the RF section. SMD resistors usually do not cause any problems, since they have low parasitic inductance up to at least 10GHz. On the other hand, there are big differences among SMD capacitors. For this reason, a single value (47pF) was used everywhere. The 47pF capacitors used in the prototypes are NPO type, rather large (size 1206), have a self resonance around 10GHz and introduce an insertion loss of about 0.5dB at 5.76GHz. Finally, the 4.7µF SMD tantalum capacitors can be replaced by the more popular tantalum 'drops'.

Quarter wavelength chokes are used elsewhere in the RF circuits. In the 5760MHz transceiver all quarter wavelength chokes are made as high impedance microstrips. On the other hand, to save board space in the 1296MHz and 2304MHz versions, the quarter wavelength chokes are made as small coils of 0.25mm thick copper enamelled wire of the correct length, chosen according to the frequency: 12cm for 648MHz, 9cm for 23cm mixers (648/1296MHz), 7cm for 1296MHz; and for L3, 5.5cm for 13cm mixers (1152/2304MHz) and 4cm for 2304MHz. The wire is tinned for about 5mm on each end and the remaining length is wound on an internal diameter of about 1mm. The SMD semiconductor packages and pin outs are shown in **Fig 11.90**.

Please note that due to lack of space, the SMD semiconductor markings are different from their type names. Only the relatively large CLY5 transistor in a SOT-223 package has enough space to carry the full marking "CLY5". The remaining components only carry one, two or three letter marking codes.

All of the microstrip circuits are built on double sided 0.8mm thick FR4 glassfibre-epoxy laminate. Only the top side is shown here, since the bottom side is left un-etched to act as a ground plane for the microstrips. The copper surface should not be tinned, nor silver or gold plated. The copper foil thickness should be preferably 35µm. Since the microstrip boards are not designed for plated through holes, care should be taken to ground all components properly. Microstrip lines are grounded

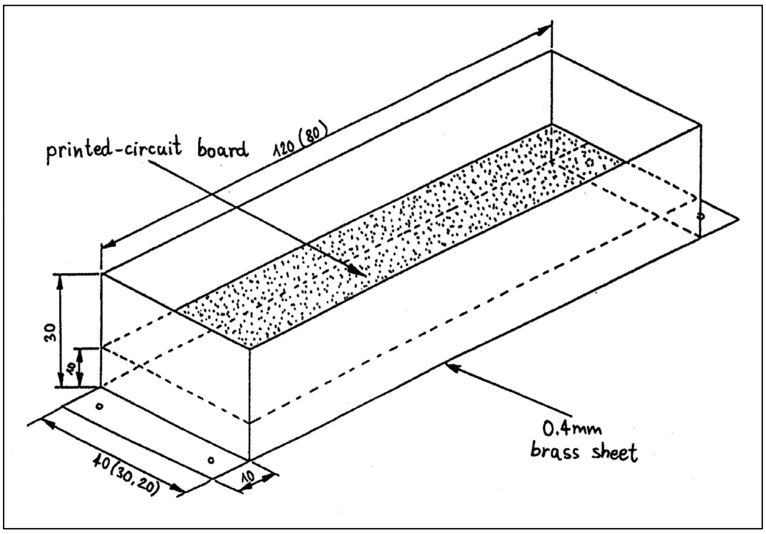


Fig 11.91: Shielded RF module enclosure for Zero-IF transceiver

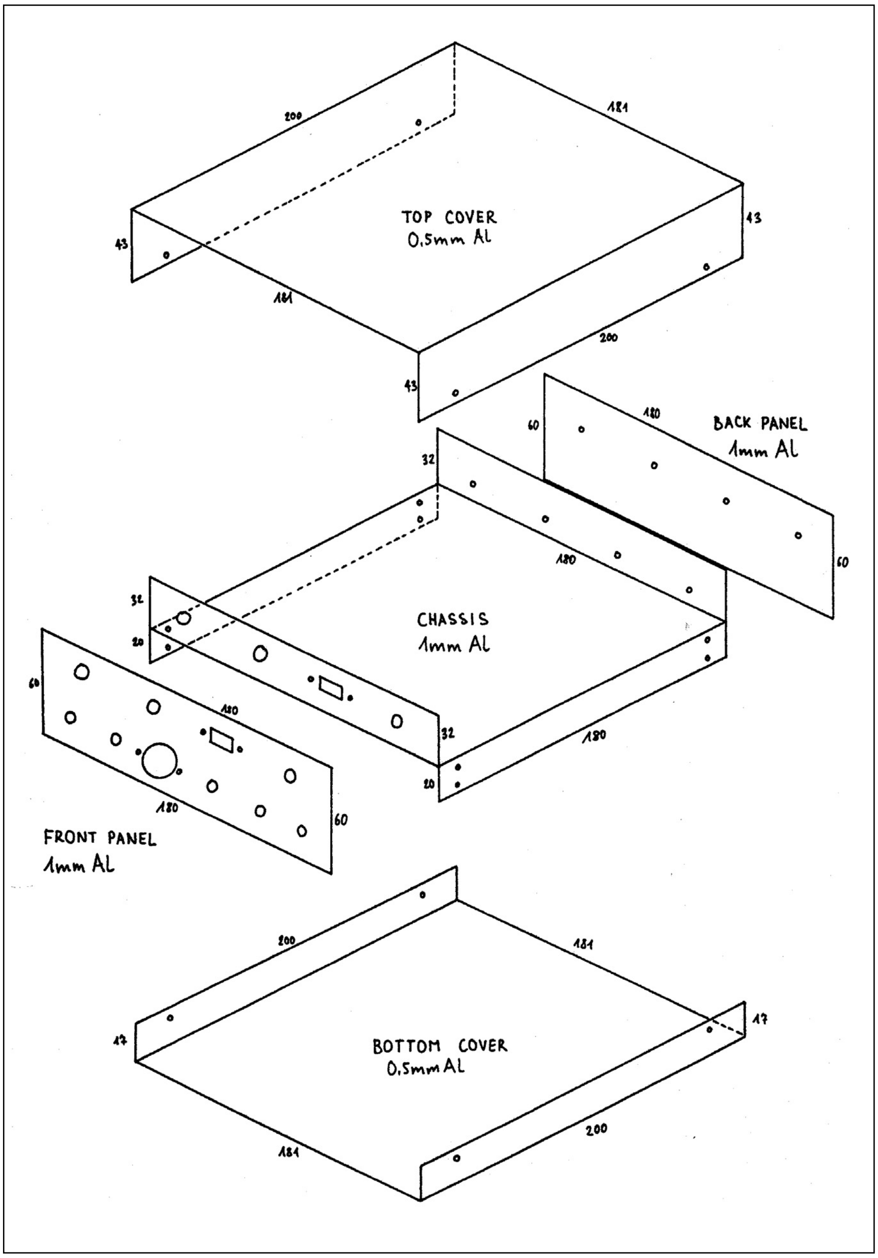


Fig 11.92: Enclosure for Zero-IF transceiver

height of about 10mm from the bottom. Additional feed through capacitors are required in the brass walls. RF signal connections are made using thin teflon dielectric coax like RG-188. The coax braid should be well soldered to the brass frame all around the entrance hole. Finally the frame is screwed on the chassis using sheet metal screws. The covers are kept in place thanks to the elastic brass sheet, so they need not be soldered. An inspection of the content is therefore possible at any time.

The VCXO/multiplier module is built on a single sided board and therefore requires both a top and a bottom cover. The remaining modules are all built as microstrip circuits, so the microstrip ground plane acts as the bottom cover and only the top cover is required. The sizes and shapes of the microstrip circuit boards are selected so that no resonances occur up to and including 2880MHz. Microwave absorber foam is therefore only required in the three modules operating at 5760MHz. To avoid disturbing the microstrip circuit, the foam is installed just below the top cover. The modules of a zero IF SSB transceiver are installed in a custom-made enclosure as shown in **Fig 11.92**. The most important component is the chassis. This must be made of a single piece of 1mm thick aluminium sheet to provide a common ground for all modules. If a common ground is not available, the receiver will probably self-oscillate in the from of 'ringing' or 'whistling' in the loudspeaker, especially at higher volume settings. The chassis carries both the front and back panels as well as the top and bottom covers.

All of the connectors and commands are available on the front panel which is screwed to the chassis using the components installed, CW pushbutton, SMA connector, meter and tuning helipot. To save weight, the top and bottom cover are made of 0.5mm thick aluminium sheet. They are screwed to the chassis using sheet metal screws. The shielded RF modules are installed on the top side of the chassis where a height up to 32mm is allowed. The audio frequency bare printed circuit boards are installed on the bottom side of the chassis. Interconnections between both sides of the chassis are made through five large diameter holes. The location of the modules as well as the location of the connectors and commands on the front panel are shown in **Fig 11.93** for both sides of the chassis.

Since the RF receiving modules are quite sensitive to vibrations, the loudspeaker should not be installed inside the transceiver. The same loudspeaker may also be used as a dynamic microphone for the transmitter. The circuit is designed so that it allows a simple parallel connection of the loudspeaker output and the microphone input. The PTT and CW keys are simple switches to ground. **Fig 11.94** shows a picture of the completed transceiver.

using 0.6mm thick silver plated wire (RG214 central conductors) inserted in 1mm diameter holes at the marked positions and soldered on both sides to the copper foil. Resistors and semiconductors are grounded through 2mm, 3.2mm and 5mm diameter holes at the marked positions. These holes are first covered on the ground plane side with pieces of thin copper foil (0.1mm). Then the holes are filled with solder and finally the SMD component is soldered into the circuit. Feed-through capacitors are also installed in 3.2mm diameter holes in the microstrip boards and soldered to the ground plane. Feed through capacitors are used for supply voltages and low frequency signals. Some components like bias resistors, Zeners and electrolytic capacitors are installed on the bottom side (as shown on the component location drawings) and connected to the feed-through capacitors.

The VCXO/multiplier module and all of the microstrip circuits are installed in shielded enclosures as shown in **Fig 11.91**. Both the frame and the cover are made of 0.4mm thick brass sheet. The printed circuit board is soldered in the frame at a

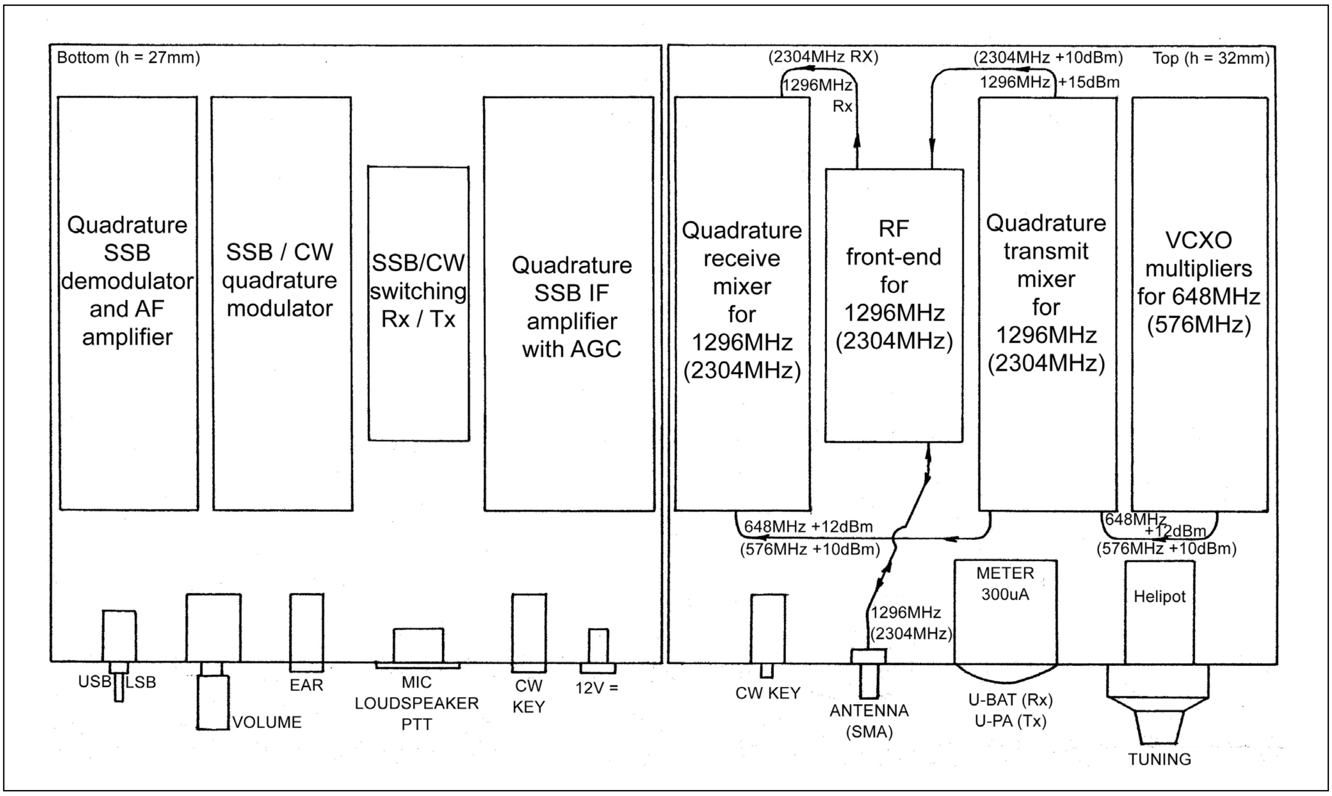


Fig 11.93: 1296MHz Zero-IF transceiver module locations

Testing the Zero IF SSB Transceivers

These SSB/CW transceivers do not require much tuning. The only module that really needs tuning is the VCXO/multiplier which is simply tuned for the maximum output on the desired frequency. Of course, the desired coverage of the VCXO has to be set with a frequency counter. After the VCXO/multiplier module is adjusted, the remaining parts of the transceiver require a check to locate defective components, soldering errors or insufficient shielding. The receiver should already work and some noise should be heard in the loudspeaker. The noise intensity should drop when the power supply to the LNA is removed. The noise should completely disappear when the receiving mixer module is disconnected from the IF amplifier. A similar noise should be heard if only one (I or Q) IF channel is connected.

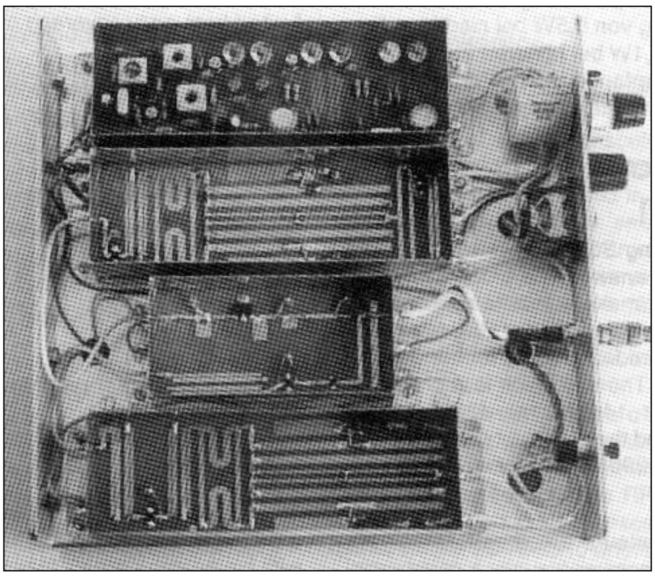


Fig 11.94: Picture of 1296MHz Zero-IF transceiver

For the next test, the receiver is connected to an outdoor antenna far away from the receiver and tuned to a weak unmodulated carrier (a beacon transmitter or another VCXO/multiplier module at a distance of a few tens of metres). Tuning the receiver around the unmodulated signal should produce both the desired tone and its much weaker mirror image changing its frequency in the opposite direction. The two trimmers in the IF amplifier should be set so that the mirror tone disappears. The correct function of the USB/LSB switch can also be checked.

Finally, the shielding of the receiver should be checked. A small, handheld antenna (10-15dBi) is connected to the receiver and the main beam of the antenna is directed into the transceiver. If the noise coming from the loudspeaker changes, the shielding of the local oscillator multiplier chain is insufficient. Next a mains operated fluorescent tube (20W or 40W) is turned on in the same room. A weak mains hum should only be heard when the handheld antenna is pointed towards the tube at 2 - 3m distance. If a clean hum without noise is heard regardless of the antenna direction, the shielding of the local oscillator multiplier chain is insufficient.

The transmitter should be checked for output power. Full power should be achieved with the modulator trimmer in the middle position whilst in CW mode. The DC voltage across the PA transistor should rise to the full value allowed by the 5.6V or 4.7V Zener diode. The output power should drop by an equal amount if only I or only Q modulation is connected to the transmit mixer. Finally the SSB modulation must be checked with another receiver for the same frequency band, or preferably in a contact with another amateur station at a distance of a few kilometres. This is the simplest way to find out the correct sideband, USB or LSB, of the transmitter, since the I and Q channels can be easily interchanged by mistake in the wiring. The residual carrier level of the transmit mixer should also be checked. Due to the conversion principle the carrier results in a 1365Hz tone in a correctly tuned SSB receiver. The carrier suppression may range from -35dB in the 1296MHz transceiver down to only -20dB in the 5760MHz transceiver. Poor carrier suppression may be caused by a too high LO signal level or by a careless installation of the BAT14-099R mixer diodes. Note that the residual carrier cannot be monitored on another correctly tuned zero IF receiver, since it falls in the AF response hole of the zero-IF receiver.

The current drain of the described transceivers should be as follows:

Receive

- For 1296MHz: 105mA
- For 2304MHz: 175mA
- For 5760MHz: 300mA.

The current drain of the transmitters is inversely proportional to the output power due to the self biasing of the PA. The minimum current drain corresponds to SSB modulation peaks or CW transmission.

Transmitters:

- For 1296MHz: 650-870mA
- For 2304MHz: 490-640mA
- For 5760MHz: 410-440mA.

All figures are given for a typical sample at a supply voltage of 12.6V.

Finally, it should be appreciated that zero IF transceivers also have some limitations. In particular, the dynamic range of the receiver is limited by the direct AM detection in the receiving mixer. If very strong signals are expected, the LNA gain has to be reduced to avoid the above problem. This is already done in the 1296MHz receiver, since strong radar signals are quite common in the 23cm band. The sensitivity to radar interference of the 1296MHz transceiver was found comparable to the conventional transverter + 2m transceiver combination. On the other hand, the 2304MHz and 5760MHz transceivers have a higher gain LNA. If the dynamic range needs to be improved, the second LNA stage can simply be replaced with a wire bridge in both transceivers. Of course, the internal LNA gain must be reduced or the LNA completely eliminated if an external LNA is used.

TRANSVERTERS

Despite the arguments listed in the preceeding section, the most popular method of becoming active on the microwave bands today is still to use a commercial transceiver and a transverter. The transverter performs two functions, it down-converts the incoming signal, on the microwave band being used, to the chosen input frequency of the commercial transceiver (tuneable IF) and up-converts the output of the transceiver to the microwave band. The most common tuneable IF is 144MHz; using lower frequencies makes it difficult to filter out the unwanted signals. The transverter will use a common local oscillator chain so that transmit and receive frequencies on the microwave band will be the same, making it possible to use narrow band modes such as SSB.

Designs of transistorised transverters for 23cm started to appear in the mid 1970s. As semiconductor technology improved, designs for the higher amateur bands became available. Most of these were quite tricky to build and persuade to work properly. In the mid 1980s "No Tune" transverter designs started to appear. These overcame many of the construction and set-up problems. Some of the amateurs who designed

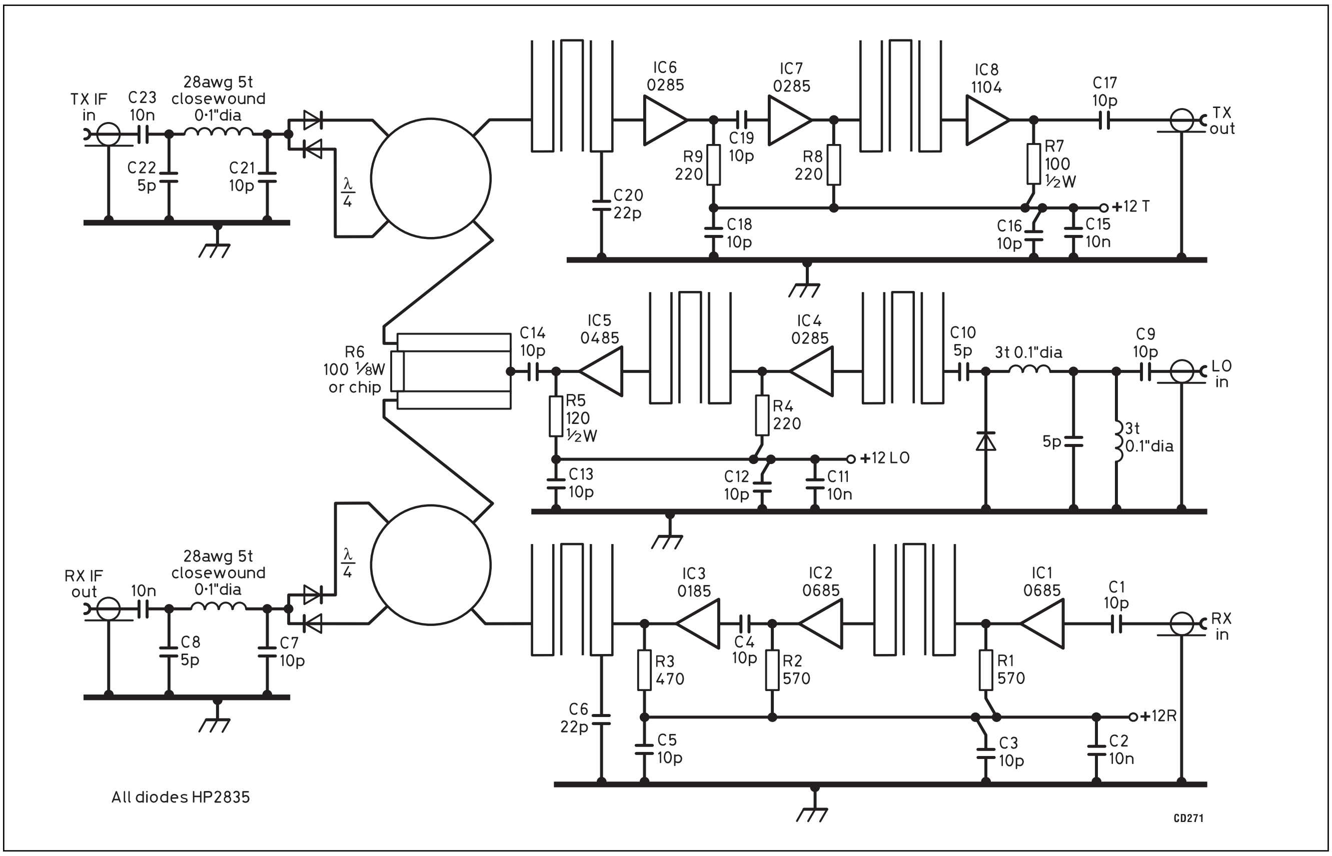
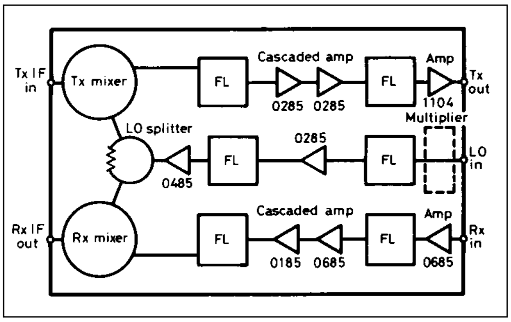


Fig 11.95: Circuit diagram of the KK7B transverter. BA481 Schottky diodes can be used instead of the HP5082-2835 diodes. Other, more modern MMICs can be substituted for the specified types, provided that the supply resistors are calculated and adjusted in value to suit the devices chosen. See reference [33]

Fig 11.96: Component layout for the KK7B transverter



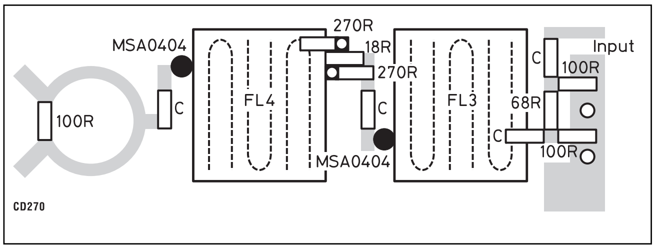


Fig 11.99: Layout of components for the modified LO chain in the KK7B transverter

these transverters now produce them as kits and ready built units, these are all listed in the bibliography.

Amateurs have always enjoyed modifying commercial equipment to work on the amateur bands, this is now possible for most of the microwave bands because there is a lot of surplus microwave link equipment on the market. The problem with this route is to recognise the potential of the equipment that we all see at rallies and equipment sales. The Internet is a wonderful source of knowledge about this, including pictures of the equipment and modification details, one such modification is shown in this chapter.

A Single-Board, No-Tune 144MHz/1296MHz Transverter

The comparatively recent development of economically priced and readily available microwave monolithic integrated circuits (MMICs) has allowed the development of a number of broadband (no-tune) low-power transverters from the 144MHz amateur band to the lower microwave bands, typically 1.3, 2.3 and 3.4GHz. Such designs use microstrip technology, including no-tune inter-stage band pass filtering in the LO, receive and transmit chains.

A 144MHz to 1296MHz transverter circuit was described by KK7B in [33]. This circuit and layout, although it does not give 'ultimate' performance in terms of either receive noise figure or transmit output power (nor is it particularly compact in terms of board size), is probably one of the simplest and most cost-effective designs available. Its simplicity also makes it suitable for novice constructors. It is also flexible enough to allow the constructor to substitute new, improved MMICs as these become available, without major re-engineering.

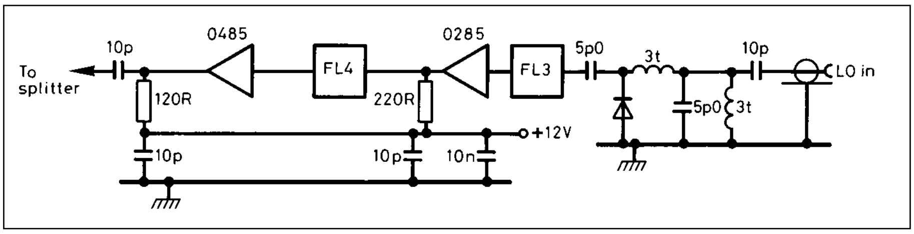


Fig 11.97: Original KK7B LO circuit used with his no-tune transverter

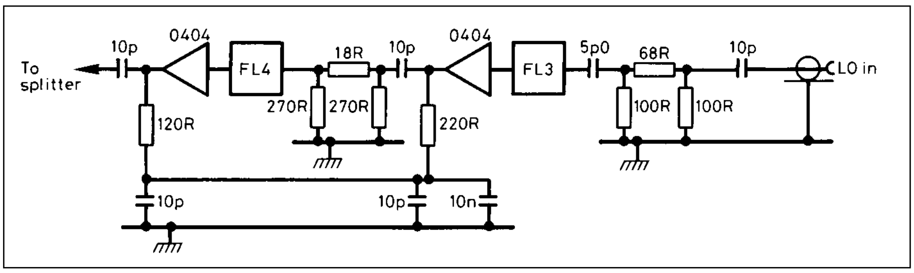


Fig 11.98: Modification to the KK7B LO circuit to allow use of the G4DDK-001 1152MHz source and the higher output level available from this circuit

The receive performance can be enhanced by means of an external (possibly mast-head) low-noise amplifier (LNA), such a high-performance GaAsFET or PHEMT design. The transmit output level, at l3dBm (20mW), is ideal for driving a linear PA module such as the G4DDK-002 design [34]. This, in turn, could drive either a solid-state power block amplifier or a valve linear amplifier.

Precision printed circuit boards for this design and a similar design for the 2.3GHz band [35] have been available for some time, produced and marketed by Down East Microwave [36] in the USA. They are also available from a number of sources in the UK. A similar design concept was adopted for a transverter for the US 900MHz amateur band [37]. The Down East Microwave website [36] is worth visiting, it has a number of useful designs and application notes that can be freely downloaded

The original circuit diagram of the 1296MHz version is given in **Fig 11.95** and the physical layout of the circuit is shown, not to scale, in **Fig 11.96**. Wide use is made of hairpin-shaped, self resonant, printed microstripline filters in the LO, receive (RX) and transmit (TX) chains, together with printed microstripline transmit and receive balanced mixers and 3dB power splitter for the LO chain. An external LO source at any sub-harmonic frequency of the required injection frequency (1152MHz for the 1296-1298MHz narrow-band communications segment of the 23cm band when using an IF of 144-146MHz) was used and a simple, on-board diode multiplier produced the required injection frequency from the LO input. Although direct injection of 1152MHz was mentioned in the original description, little guidance was given as to how to achieve this. With a simple modification to the LO chain and a few changes to circuit values and devices, without need for PCB changes, it is easily possible to use the G4DDK-001 1152MHz source, already described, as the LO for this design.

Fig 11.97 gives the circuit and component values for the original LO chain, while **Fig 11.98** shows the modifications to allow the correct mixer injection levels to be attained when using the single +13dBm output option of the G4DDK-001 1152MHz source. **Fig 11.99** shows the layout of the modified circuit using the existing PCB pads and tracks.

Construction is straightforward, using surface-mount techniques, ie all components, whether SMD or conventional, are mounted on the track side of the board unlike conventional construction. All non-semiconductor components - connectors, resistors, capacitors and inductors - should be mounted first, the MMICs and mixer diodes last, taking adequate precautions to avoid both heat and static damage. Note that the

3216 PLL Calculations for X Band Transverter with 144MHz 2nd IF; 1st LO=11,360MHz; 1st IF=992MHz								
Ref MHz	2	Ref MHz can be 10MHz divided by any integer from 1 - 16						
VCO MHz	2,272							
PLL MHz	1,136	PLL in MHz is VCO/2 and must be an integer multiple of Ref MHz						
N	568							
		M6(Pin15)	M5(Pin14)	M4(Pin13)	M3(Pin10)	M2(Pin9)	M1(Pin8)	M0(Pin7)
M	55	0	1	1	0	1	1	1
Board as is		0	0	0	0	0	0	0
		A3(Pin21)	A2(Pin20)	A1(Pin19)	A0(Pin18)			
A	8	1	0	0	0			
Board as is		0	0	0	0			
		R2(Pin5)	R2(Pin4)	R1(Pin3)	R0(Pin2)			
R	4	0	1	0	0			
Board as is		0	0	0	0			
Lift pin22								
Reference suppression filter modifications, parallel these capacitors with the following values								
Ref MHz	C1	C2,C3		Add 1pF to VCO				
5	None	None						
2	1000pF	3000pF						
1	4700pF	6800pF						

Table 11.11: Synthesiser calculations for Qualcomm OmniTracks unit. 3216 PLL Calculations for X Band Transverter with 144MHz 2nd IF; 1st LO=11,360MHz; 1st IF=992MHz

values of the bias resistors, which set the working points of the MMICs, were chosen for a supply rail of +12V DC. Higher supply voltages will require recalculation of these values and the constructor should refer to either the maker's data sheets for the particular devices used or to the more general information given in reference [38]. When construction is complete and the circuit checked out for correct values and placing of components, there is no alignment as such, assuming that the LO source has already been aligned! It should simply be a matter of connecting the transverter to a suitable 144MHz (multimode) transceiver via a suitable attenuator and switching interface such as that by G3SEK [39] or the G4JNT design available from reference [40].

A 10GHz Transverter from Surplus Qualcomm OmniTracks Units

These modifications were produced by Kerry Banke, N6IZW, of the San Diego Microwave Group and presented at The Microwave Update in 1999. The project offers an economical route to 10GHz, the unmodified transceiver, 10MHz TXCO and unmodified 1W PA can, at the time of writing, be ordered from Chuck Houghton for about £100 [41].

An earlier Qualcomm X-Band conversion project required considerable mechanical changes as well as electrical modifications and was based on replacing the original stripline filters with pipecap filters. These filters were required to provide sufficient LO and image rejection at 10GHz that the original stripline filters could not provide for a two meter IF. This version uses a somewhat smaller, more recent OmniTracks unit that contains the power supply and synthesiser on the same assembly as the RF board, and utilises dual conversion high side LO to allow use of the stripline filters. The filter modification has been proven to work well by extending the filter elements to specified lengths. Some additional tuning of the transmit output stages appears to be required for maximum output.

The synthesiser VCO operates at 2,272MHz, and when multiplied by five it becomes 11,360MHz for the first LO. The first IF frequency is 992MHz which is near the original internal IF frequency of 1GHz. The second LO is derived from the synthesiser

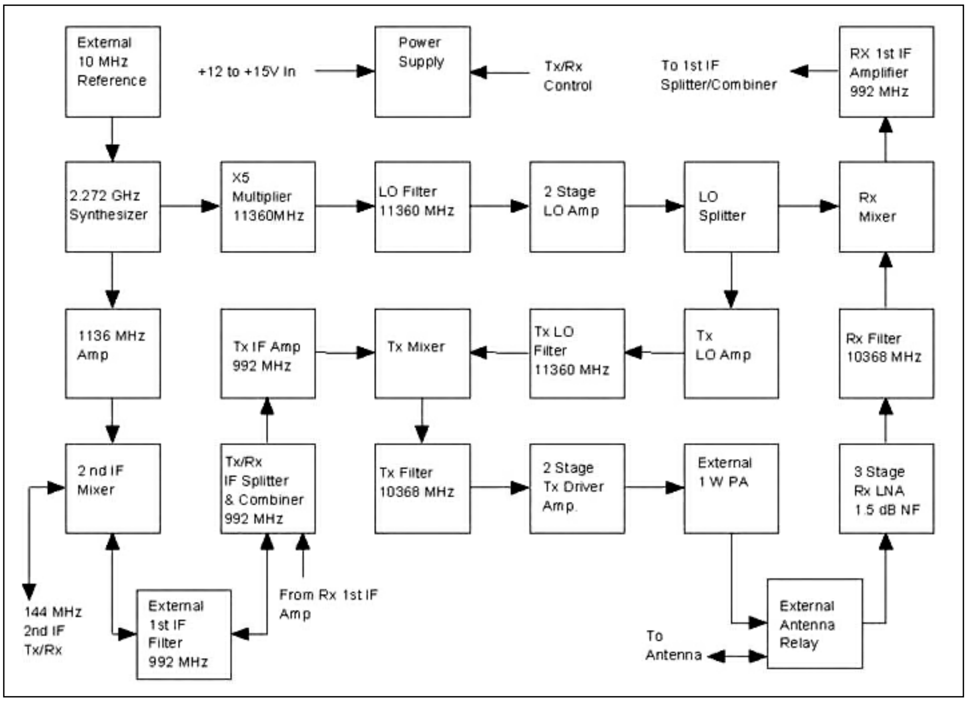


Fig 11.100: Block diagram of Qualcomm X band transverter conversion

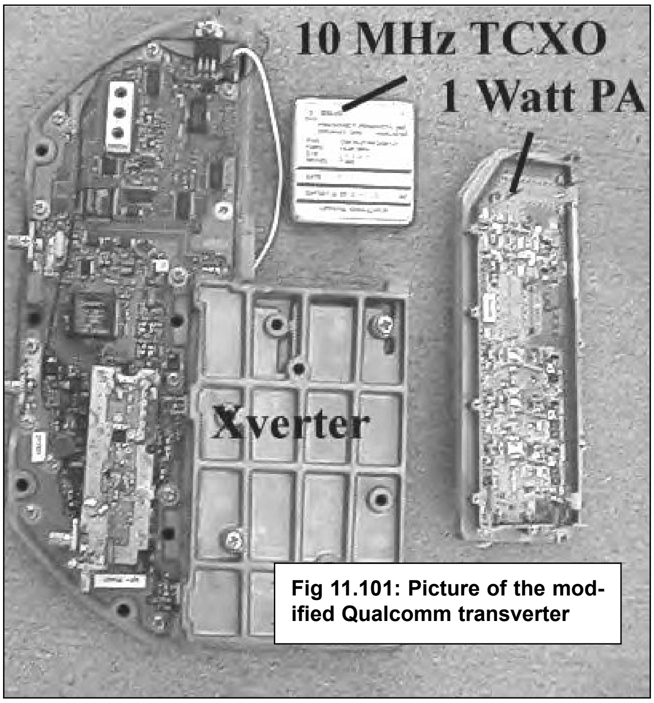


Fig 11.101: Picture of the modified Qualcomm transverter

pre-scaler, this divides the VCO frequency by two to produce 1,136MHz. Other second IF frequencies may be calculated using the relationship (RF-IF2)/0.9 = LO1 where RF is the 10GHz operating frequency (10,368MHz), IF2 is the second IF frequency, and LO1 is the first LO frequency. The synthesiser output frequency is then LO1 divided by five. **Table 11.11** shows the Excel spread sheet used to calculate the synthesiser programming.

The second conversion stage consists of a second LO amplifier (1,136MHz) and SRA-11 mixer converting the 992MHz 1st IF to the 144MHz 2nd IF. A 992MHz filter is required between the two conversion stages. Both Evanescent Mode and Coaxial Ceramic filters have been used. The conversion yields a reasonably high performance transverter with a noise figure of about 1.5dB and a power output of +8dBm, frequency locked to a stable 10MHz reference. Power required is +12VDC with a current consumption of about 0.5A on receive and 0.6A on transmit (about 1.5A total on transmit when including the 1W PA). **Fig 11.100** is a block diagram of the modified unit.

The unmodified circuit has a synthesiser output of 2,620MHz providing an LO of 13.1GHz. The original transmit frequency was around 14.5GHz with one watt output, and the receiver was near 12GHz. Unfortunately, the integrated PA in the original configuration provides no useful output below 12GHz and is not modifiable, so it has been removed for the 10GHz conversion. The transmit and receive IF preamplifiers make the transmit input requirement low (-10dBm) and provides high overall transverter receive gain.

Fig 11.101 shows a picture of the modified transverter, 1W amplifier and 10MHz TCXO. **Fig 11.102** shows a picture indicating the locations of the various functions. The following is an outline of the conversion procedure [42]:

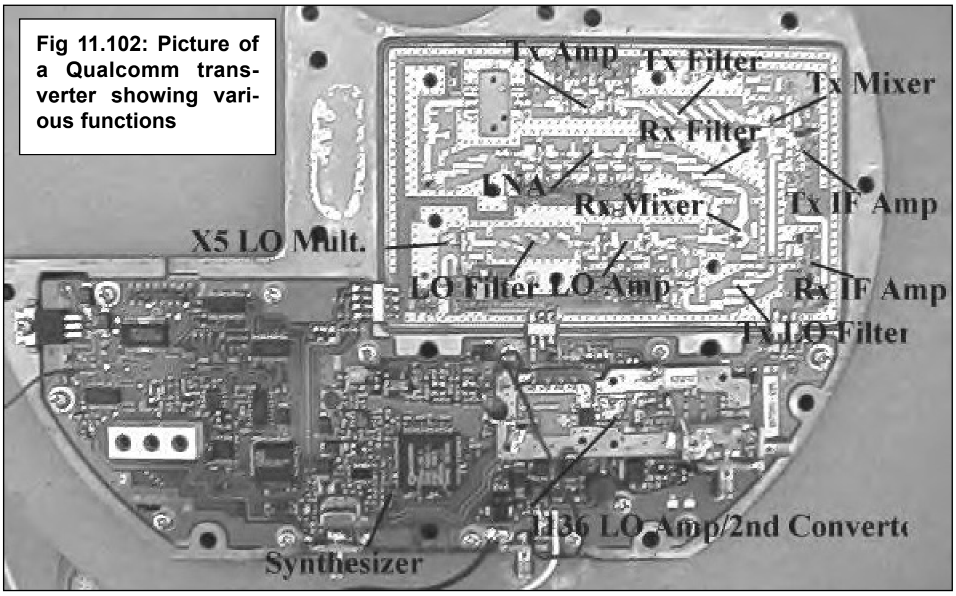


Fig 11.102: Picture of a Qualcomm transverter showing various functions

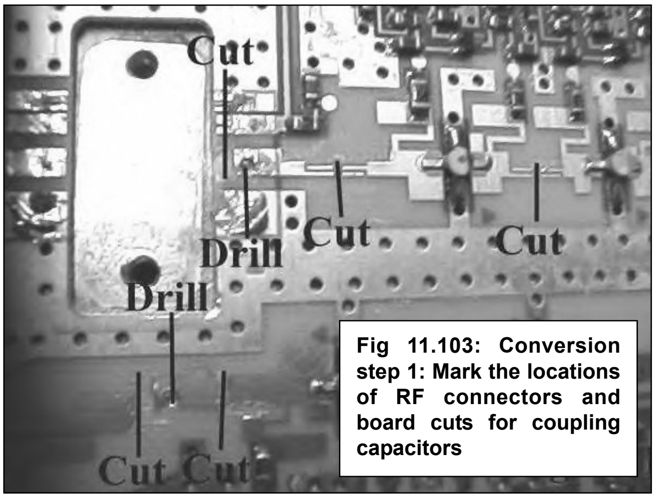


Fig 11.103: Conversion step 1: Mark the locations of RF connectors and board cuts for coupling capacitors

1 Marking location of RF connectors and removal of circuit boards.
2 Base plate modification for mounting two SMA connectors (10GHz receive and transmit) plus four SMA connectors installed (2 RF + 1 IF and 10MHz Reference input).
3 Clearing of SMA connector pin areas in PCB ground plane.
4 Remounting of PCBs.
5 Cuts made to PCB and coupling capacitors installed.
6 Stripline filter elements extended and tuning stubs added.
7 Synthesiser reprogrammed and 4 capacitors added.
8 Add tuning stubs to the x5 Multiplier stage
9 2nd LO amplifier, mixer and 1st IF filter added.
10 Power and transmit/receive control wires added.
11 Test of all biasing.
12 Synthesiser and receiver test.
13 Transmitter test and output stage tuning.

Step 1. Mark the location of RF connectors and board cuts for coupling capacitors: Before removing the boards from the base plate, carefully drill through the board in the two places shown using a 0.050 inch diameter drill just deep enough to mark the base plate. These are the locations for receive and transmit RF SMA connectors. The upper connector hole (transmit) is located 0.5 inches to the left of the transistor case edge. The lower hole (receive) is located 0.4 inches to the left of the transistor case edge. Make the cuts as shown in **Fig 11.103** using a sharp knife.

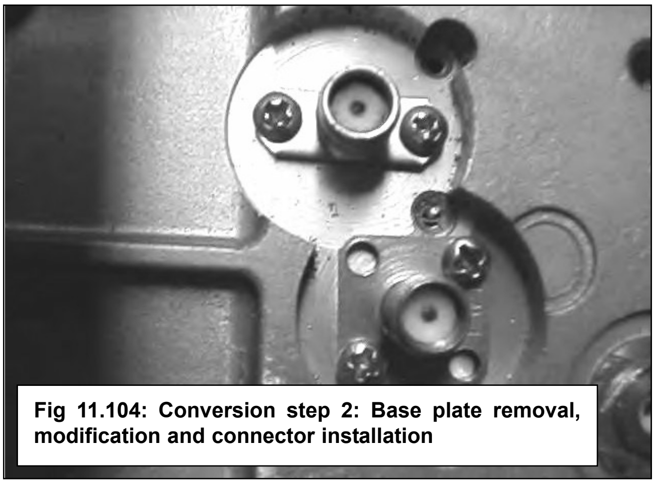


Fig 11.104: Conversion step 2: Base plate removal, modification and connector installation

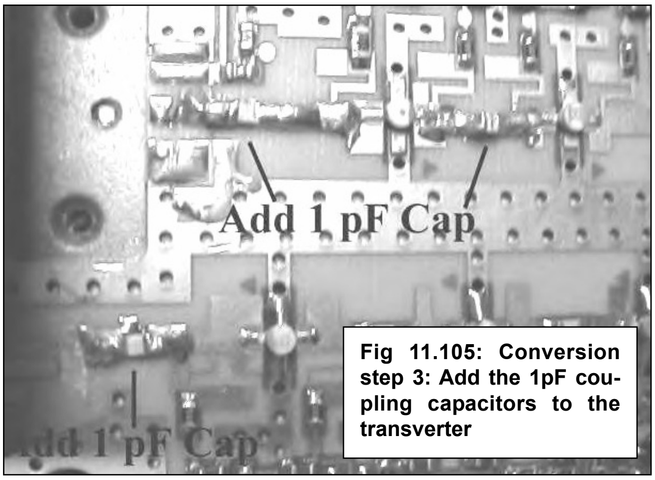


Fig 11.105: Conversion step 3: Add the 1pF coupling capacitors to the transverter

Step 2. Base plate removal, modification, and connector installation: After making the holes and cuts, remove all screws and lift the boards off of the base plate. (Note: the original antenna connector pin must be de-soldered to remove the board. Once the boards are removed, drill through the plate in the 2 locations marked using a 0.161 inch drill to clear the teflon insulator of the SMA connectors. Use a milling tool to remove enough material on the back side of the base plate (see **Fig 11.104**) to clear the two SMA connector locations, taking the thickness down to about 0.125 inches (may vary depending on available SMA connector pin length). Locate, drill and tap the base plate for two 2-56 mounting screws at each connector. Mount the SMA connectors on the base plate and cut the Teflon insulator flush with the top side of the base plate (circuit board side). Carefully clear the ground plane around the two connector holes on the bottom side of the circuit board to prevent the SMA probe from being shorted (using about a 0.125 inch drill rotated between your fingers). Reinstall the circuit boards onto the base plate.

Step 3. Add coupling capacitors: Add the 3 capacitors along with the additional microstrip pieces to modify as shown in **Fig 11.105**.

Step 4. Extend the transmit LO filter elements to the total length shown in **Fig 11.106**: Filter extensions are made by cutting 0.003 - 0.005 inch copper shim stock into strips about 0.07 inches wide and tinning both sides of the strip, shaking off excess solder. No additional solder is normally needed when attaching the extensions as the tinning re-flows when touched by the soldering iron. The length of the top element (0.21 inches) is measured between the marks as shown.

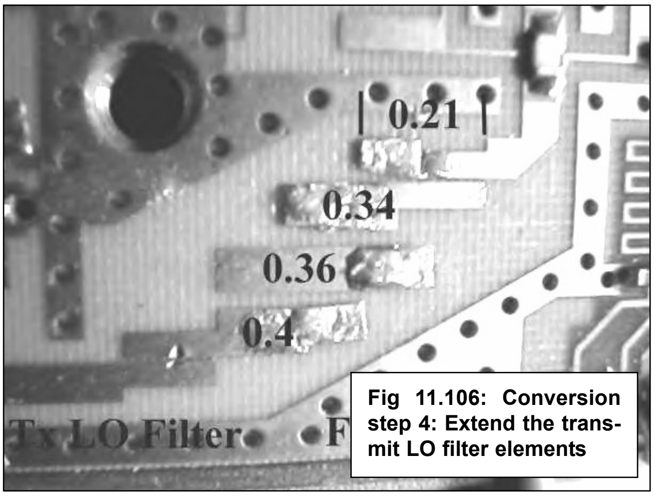


Fig 11.106: Conversion step 4: Extend the transmit LO filter elements

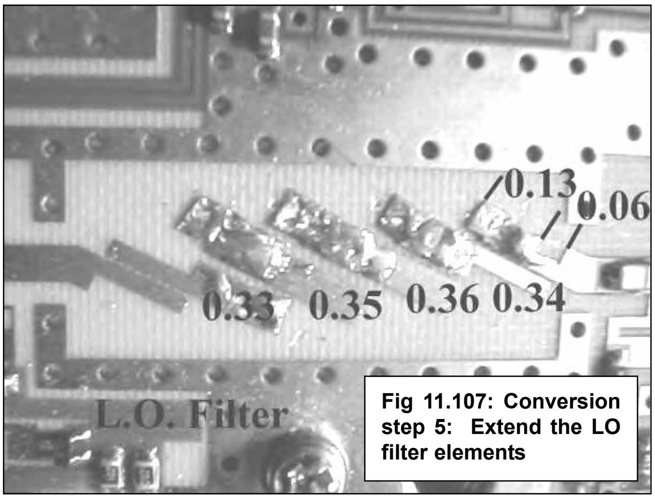


Fig 11.107: Conversion step 5: Extend the LO filter elements

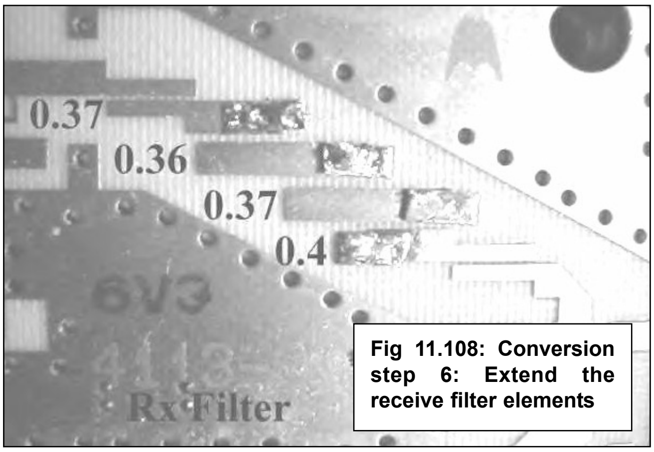


Fig 11.108: Conversion step 6: Extend the receive filter elements

Step 5. Extend the LO filter elements as shown in **Fig 11.107**: Again, total element lengths are shown except for the right-most element that has additional dimensions.

Step 6. Extend the receive filter elements as shown in **Fig 11.108**: Dimensions shown are total element length.

Step 7. Extend the transmit filter elements as shown in **Fig 11.109**: Dimensions shown are total element length.

Step 8. Add the tuning stubs to the x5 Multiplier stage: This stage is located directly to the left of the LO filter which is shown in Fig 11.107. The gate of the x5 Multiplier stage requires addition of two stripline stubs, as shown in **Fig 11.110**.

Step 9. Modify the 2nd LO amplifier board, mount onto transverter and connect 1,136MHz LO input through 1pF coupling

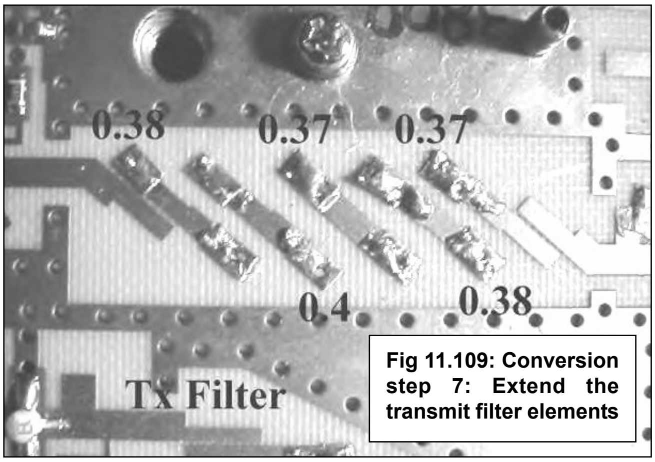


Fig 11.109: Conversion step 7: Extend the transmit filter elements

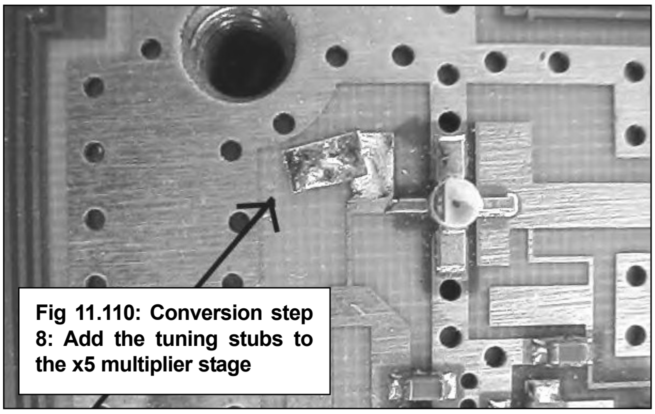


Fig 11.110: Conversion step 8: Add the tuning stubs to the x5 multiplier stage

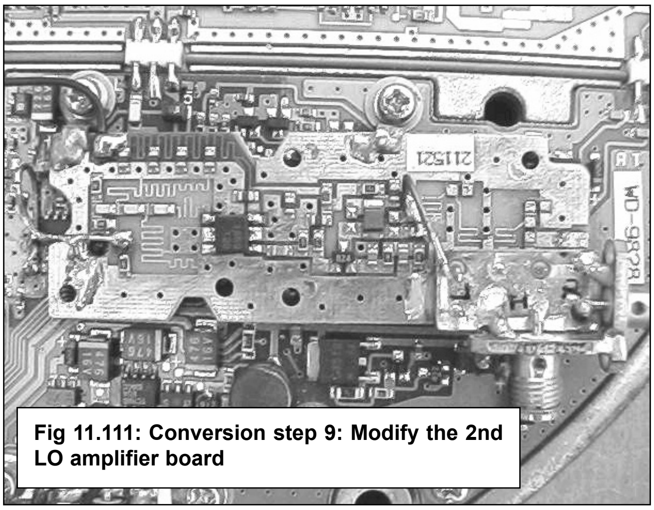


Fig 11.111: Conversion step 9: Modify the 2nd LO amplifier board

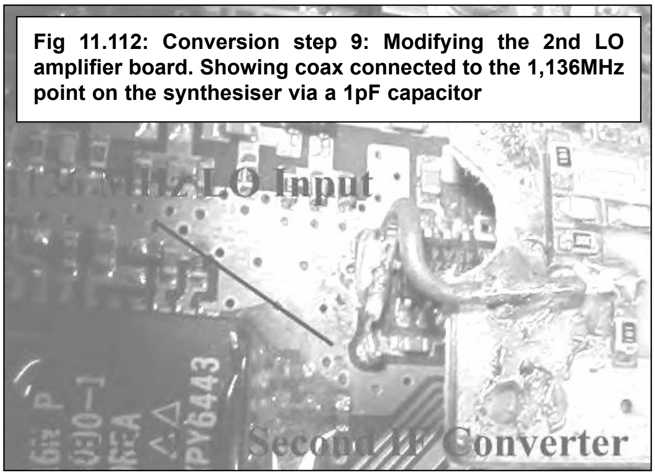


Fig 11.112: Conversion step 9: Modifying the 2nd LO amplifier board. Showing coax connected to the 1,136MHz point on the synthesiser via a 1pF capacitor

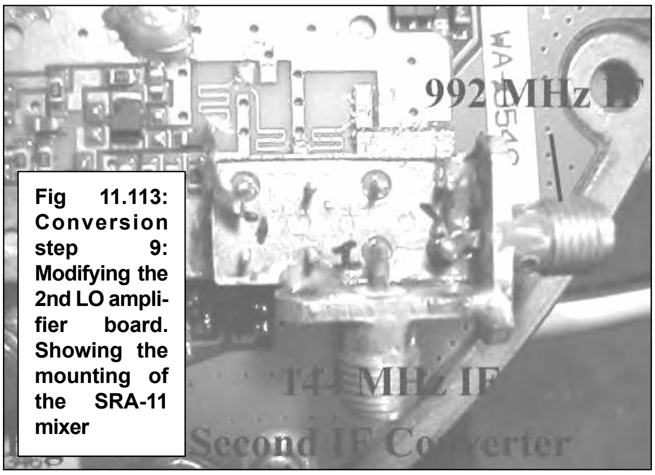


Fig 11.113: Conversion step 9: Modifying the 2nd LO amplifier board. Showing the mounting of the SRA-11 mixer

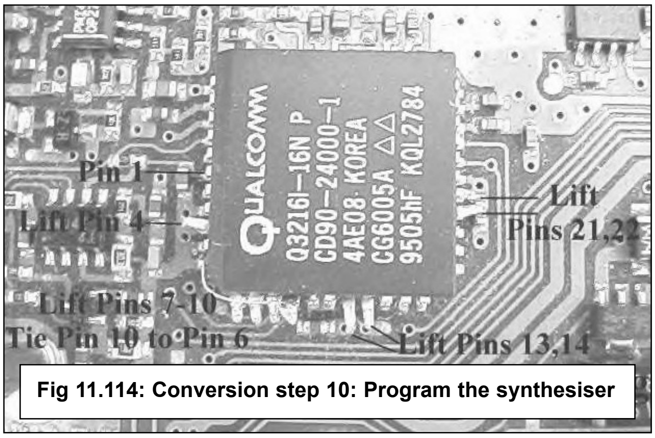


Fig 11.114: Conversion step 10: Program the synthesiser

capacitor as shown in **Figs 11.111 - 11.113**. Fig 11.111 shows the overall second IF converter which is mounted using two grounding lugs soldered to the top edge of the LO amplifier board and secured by two of the screws which mount the main transverter board. Fig 11.112 shows the coax connected to the 1,136MHz point on the synthesiser through a series 1pF capacitor. Fig 11.113 shows the mounting and wiring of the SRA-11 mixer onto the LO amplifier board. Note the cut on the original amplifier output track after the connecting point to the mixer. The mixer case is carefully soldered directly to the LO amplifier board ground plane. The IF SMA connectors are mounted by carefully soldering them directly to the top of the mixer case.

Step 10. Program the synthesiser as shown in **Fig 11.114** by carefully lifting the pins shown with a knife. Ground pin 10, connecting it to pin 6 that is ground. Add the two 3000pF and 1000pF in parallel with the existing reference filter capacitors as shown in **Fig 11.115**.

Step 11. Add a 1pF capacitor as shown in **Fig 11.116** to lower the VCO frequency

Step 12. Add three transmit mixer tuning stubs as shown in **Fig 11.117**.

Step 13. The transmit/receive control is connected as shown in **Fig 11.118**. Grounding the control line places the transverter in transmit mode. The control can be open or taken to +5V to place the transverter in receive mode.

Step 14. The +12VDC power input is connected to the point shown in **Fig 11.119**. The original air core coil, with one end connected to that point, has been removed from the board. (This choke was originally used to supply +12V to the transverter through the 1st IF port).

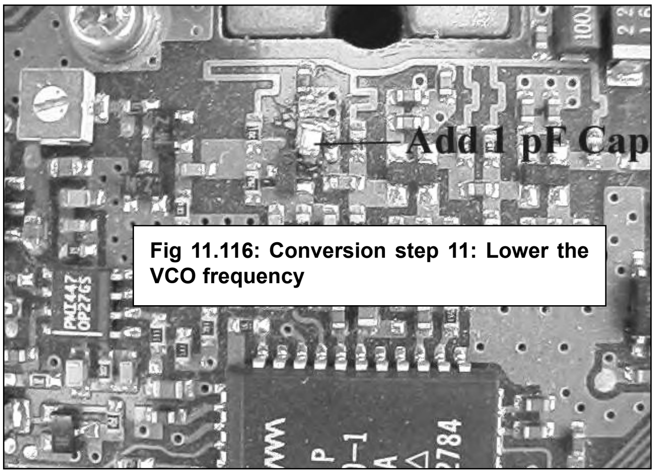


Fig 11.116: Conversion step 11: Lower the VCO frequency

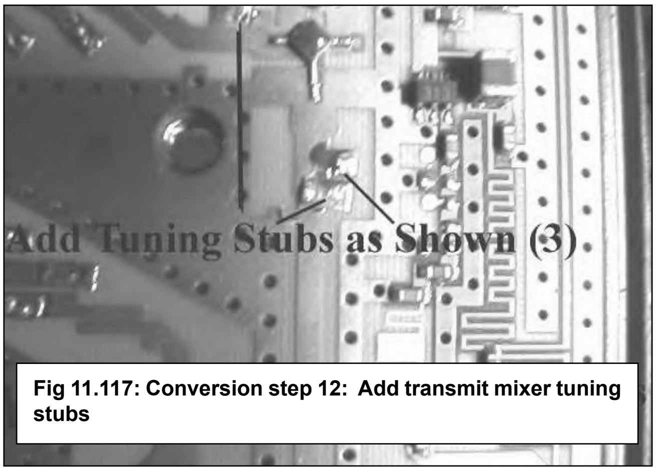


Fig 11.117: Conversion step 12: Add transmit mixer tuning stubs

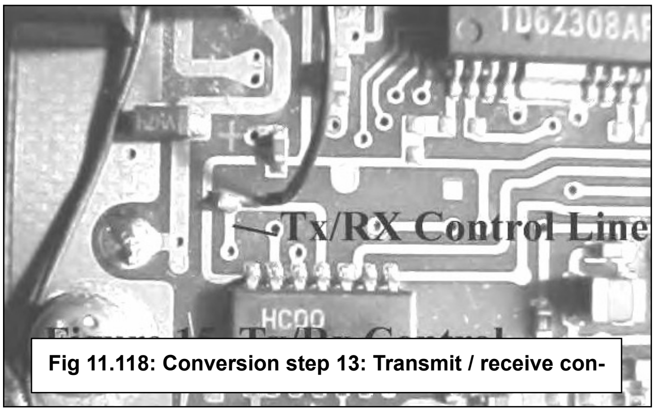


Fig 11.118: Conversion step 13: Transmit / receive con-

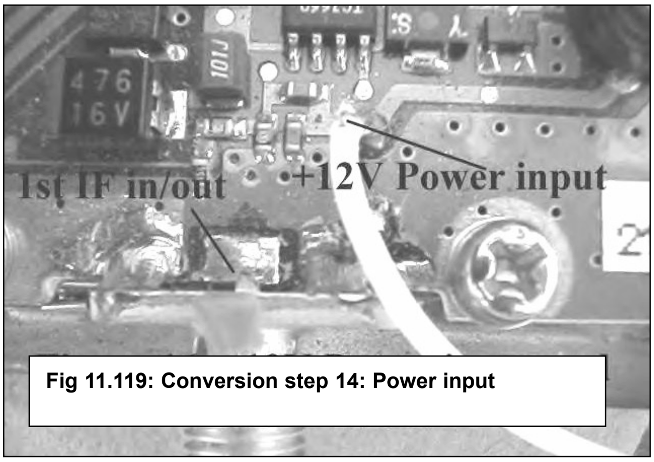


Fig 11.119: Conversion step 14: Power input

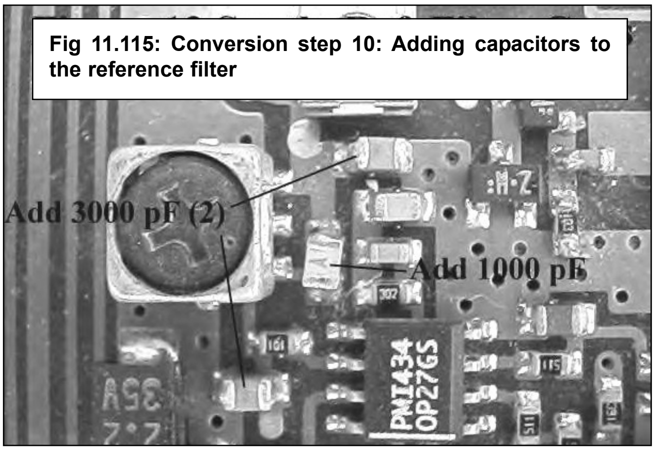


Fig 11.115: Conversion step 10: Adding capacitors to the reference filter

Step 15. Powering up the Transverter: Apply +12V to the power connector and verify that the current drawn in receive mode is about 0.5A. Connect the 10MHz reference to the transverter board. Pin 43 of the synthesiser IC should be high when locked. If available, use a spectrum analyser to check (sniff using a short probe connected by coax) the synthesiser output frequency and spectrum. The synthesiser should be operating on 2,272MHz and no 2MHz or other spurs should be visible. Carefully probe the drain of each FET in the LO multiplier, LO amplifier, and LNA to verify biases are approximately +2 to +3VDC. A drain voltage of near 0V or 5V probably indicates a problem with that stage. Place the transverter in transmit mode and verify the biasing on the transmit LO amplifier and transmit output amp stages. Tune the 992MHz 1st IF filter (not part of the transverter board) and connect it between the 1st IF ports on the transverter board and second IF converter. The receiver noise level at the 2nd IF port on the 2nd converter should be very noticeable on a 2 meter SSB receiver. A weak 10,368MHz signal can then be connected to the receiver RF input connector and monitored on the 2m receiver. The overall gain from receiver RF input to 2nd IF output should be roughly 35 to 45 dB. Place the transverter into transmit mode and connect about –10dBm at 144MHz to the 2nd IF port. Monitor the power level at the transmit RF output port and add/move the transmit amplifier tuning stubs shown in Fig 11.120 as required for maximum output. Typical transmit output will be about + 8dBm. This is considerably more than required to drive the one watt amp to full power.

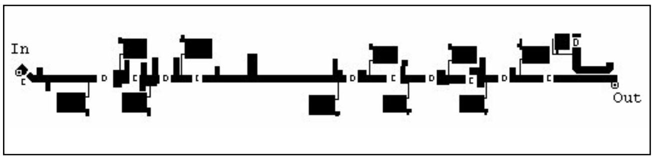


Fig 11.120: 1W PA board prior to tuning. -15dBm input gives +5 to +10dBm output with 10V at approximately 1A

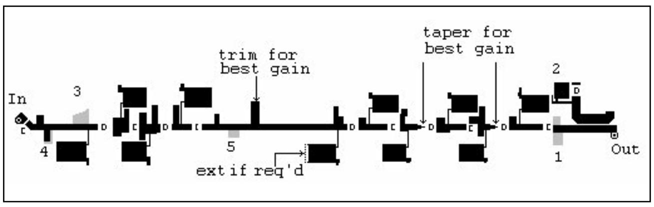


Fig 11.121: 1W PA board after tuning. The shaded tabs were added and tuned in the sequence shown. Results vary slightly from board to board. Key: [= coupling capacitor, D = devices, Input coupled with 2pF

To Convert the 1W PA

These conversion notes were produced by Ken Schofield, W1RIL [43]: Many PA boards have been successfully re-tuned for 10GHz operation. No two boards are exactly alike and each will tune a little differently from its apparent twin. The numbered steps in **Fig 11.121** will in many cases get your PA up into the gain range stated. You will find that numbered step 3 to be the most sensitive to gain increase. Unfortunately it is also one of the 'busiest' areas on the board.

Be careful! A few dos and don'ts are shown to help you bypass some of the many pitfalls that can be encountered - many are obvious and have been stated before, but bear repeating.

Do:

- Use a low voltage grounded soldering iron, and work in a static-free area.
- Check for negative bias on all stages prior to connecting Vcc voltage.
- Use good quality 50 mil chip caps - in and out approximately 1 to 2pF.
- Remove all voltages prior to soldering on board.

Don't:

- Work on board tracks when tired, shaky or after just losing an argument with your wife.
- Touch device inputs with anything that hasn't been just previously grounded.
- Apply Vcc to any stage lacking bias voltage.
- Shoot for 45dB gain - you won't get it! Be happy with 25 to 30dB

LASER DXING

Communication by light has been used for many centuries, from beacons being used to warn of advancing invaders through the use of the Aldis Lamp to transmit Morse code messages, to advanced laser communications used in today's high speed telecommunications links. For amateurs, the use of light is an extension of the frequency spectrum into the Terahertz (THz) region. Visible light is in the range 380 – 750THz and infra red from 100THz – 380THz. Use of these frequencies introduces some new challenges, not least being much more dependency on weather conditions. Amateurs in Germany and America have been actively operating in the light spectrum for many years.

In the UK there is an active group of amateurs [44] who have established the UK records for laser DX contacts. The account of this contact, by Allan Wyatt, G8LSD, (see box overleaf) emphasises the difficulties encountered in establishing a contact at these frequencies and will hopefully encourage more amateurs to try this interesting avenue of the hobby.

OPT301 Laser Receiver

This receiver was produced by David Bowman, G0MRF [45]. The OPT301 is in a TO-99 eight lead package and has good sensitivity but a reduced bandwidth of 4kHz. The peak response is just into the Infra Red region at 750nm but its sensitivity in the visible red spectrum at 670nm is only a few percent down. The circuit is shown in **Fig 11.123.** It uses a single supply at 12 - 13.8V. The detector has a 10M-ohm feedback resistor soldered directly between pins 2 and 5. The output is followed by a NE5534 low noise op-amp (3.4nV/Hz) configured as a band pass filter. The final NE5534 is an inverting buffer amplifier with a gain of 20. The complete detector assembly is designed to be mounted in a small metal box at the focal point of a lens. **Fig 11.124** (in Appendix B) shows the PCB layout viewed from the top or component side and **Fig 11.125** shows the component layout. The detector is fitted to the other side of the board which has a ground plane in un-etched copper but has holes counter-

Fig 11.122: Picture of G0MRF's laser transmitter and receiver

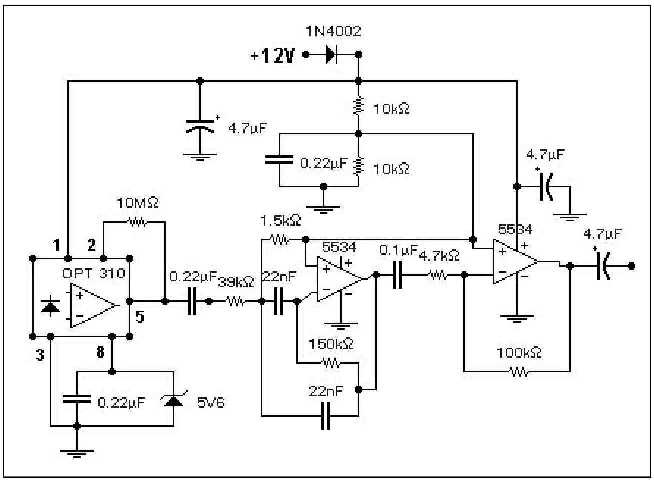


Fig 11.123: Circuit diagram of laser receiver

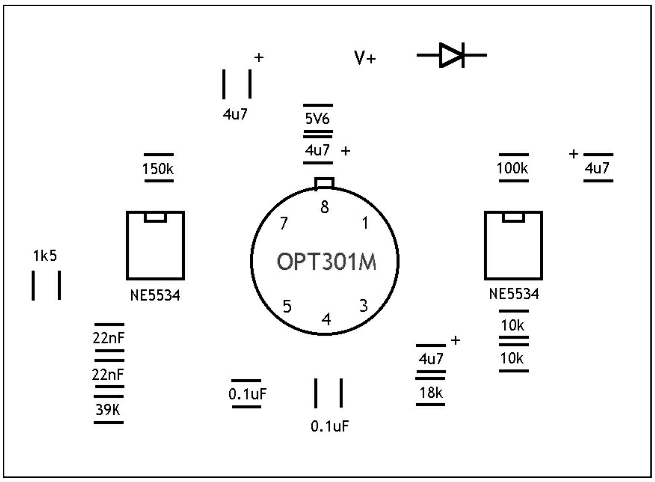


Fig 11.125: PCB layout for laser receiver

Account of setting the UK Laser DX Record – 76.1km on 8 October 2003

"It all started with a trip to Tunbridge Wells. On the return leg the air seemed very clear so I went home via the Ashdown Forest. In the binoculars I could see the towers on Truleigh Hill, the trees at Chanctonbury, then quite easily I found the twin masts at Bignor (50km+) next I could see the television mast north of Midhurst and finally I saw the stubby mast at Butser. From here I had only ever seen Butser with the aid of the telescope. Everything then moved at great speed!

"After a call to David Bowman, G0MRF [whose laser equipme nt is shown in **Fig 11.122**], I was back on Ashdown Forest with all the gear by 16:45 local time, but the conditions had deteriorated and to the left of the Midhurst television mast was only mid-distance hills. The cloud was very thick and black and rain looked very possible. The 70cm talkback (433.400 FM) was difficult to establish as the wind kept blowing my nine element beam round on the lighting stand that I used as a mast.

"The telescope was set-up and I confirmed that Butser was not visible. However an orange tinted light was appearing above the South Downs as a thin slit against the cloud darkened sky. The effect was to make the top of the hills glow and be exceptionally clearly defined. On Chanctonbury, the cloud covered the top of the Downs and it was less easy to make out details. Bignor Hill merged into the sky as though rain were falling. I could still see the mast at Midhurst and it was obvious that visibility was extending only just past 50km; still exceptional but less so than earlier.

"I spoke to G3JMB on 70cm to ask if he would listen-in and verify the contact that David and I hoped to achieve and then waited for G0MRF to arrive on site at Butser Hill. The weather was slowly deteriorating while I was standing by the road and I had gone from just a shirt and jumper to adding my fleece and cagoule, all in the space of 20 minutes! David arrived at Butser and had to carry all his equipment to the top of the hill wearing no wind protection. When we finally established talkback the visibility had dropped to a disappointing degree. My telescope was aimed at the dip on the horizon where Butser should have been and the horizon was a uniform grey.

"The wind was now gusting and making the image in the eyepiece jump in spite of the very heavy mounting base of the solidly made 4.5 inch Russian reflector telescope. David was unable to operate from his planned location as the local vegetation was higher than the top of his tripod. He had to move site and make a second long trek to the car for the second box of equipment. Due to the back packing required at Butser, David did not have the power for a beacon light. Any location aids had to come from my end at Ashdown. I had 250Ah of 13.8V from the batteries under the back seat of the vehicle and the associated switch mode converter, as well as 240 Volts AC from the inverter run from the starting battery. By the time David had found a suitable location the visibility had dropped even further and the wind had not abated.

"We then set about converting our usual operating plan into something to suit the conditions. David swept his laser across an angle that included Ashdown, the beam width was usually sufficient for this to work, however this time the telescope showed no sign of even the briefest of flashes of the characteristic red. Plan B time.

"As clouds were passing over the Downs to the south of me and the wind dropped. I set up the laser transmitter, the tripod being held in place by the weight of an 80Ah leisure battery, and wired up both transmitter and receiver. A second set of scans with the laser at the Butser end also failed to give any tone in the receiver. Even a strobe light operated from Ashdown found no path to Butser. Time was passing. Amid much head scratching I went back to serious scanning of the horizon with the telescope. By now a beautiful sunset was developing and the sky was a vivid orange through the broken cloud.

"For one moment I thought I could see as a very faint change in the uniform colour of the horizon, the inverted bathtub shape of Butser. I aimed the telescope at the northern end where the mast is to be found. Again no contact. Having taken note of the local terrain I had a good idea as to the aiming point and so set up my laser onto the target. David saw nothing. Light was fading fast and I expected the horizon to go black before I had a clear fix. As David had no beacon this would have ended the contact.

"Returning to the telescope, a point of light from a mercury vapour lamp at probably 45 to 50km away was revealed. This was not visible with binoculars or through the rifle sight on the transmitter. David still could see nothing. A sweep of his laser showed no flashes in the telescope. The sunset was fading into a beautiful burnt orange.

"The wind had dropped even more. Once again, I closely scanned the horizon. First I saw the faint outline of Butser Hill and then for a moment I imagined that the mast was visible against the darkening grey orange. A few minutes later after talking on the 70cm link I returned to the telescope. Several points of mercury vapour lamps acted as a horizontal grid and slightly above them was the mast on the top of Butser. The binoculars and rifle sight still showed nothing but the points of light, almost blurred into a single point. This gave just enough information to aim the laser. David still saw nothing.

"Next, as it was now quite dark, I switched on my torch. David saw it almost at once. He still could not see the laser, even though it was directly aimed at him. He swept his laser at the point where the torch had been and I saw a brief flash of red. It was momentarily bright but did not stay visible. I moved the talkback from the van closer to the receiver and was able to give a running commentary, but we could not get more than brief flashes of the laser.

"I returned to my laser and started a systematic scan of the horizon. At one point I accidentally turned the vertical control too far, and so I was seeing into the hill in the rifle sight. David saw the beam. Once again, the rifle sight prism had moved in transit and the sight had lost vertical accuracy. However, the transmitter alignment sight was still spot on in the horizontal plane. David now relayed the laser signal back over our 70cm FM link. With this feedback I could optimise the pointing and we established one way contact.

"The CW signal was clear and lacked the usual scintillation. The laser's brightness was such that David reported that I was the brightest light on the horizon. We then concentrated on receiving David's signal. This required critical adjustment but once I got a faint signal in the receiver he was able to optimise by our usual feedback route. By now I had spent 135 minutes standing by the road. We exchanged callsigns, reports and random characters in just a few minutes. David had an active filter on receive tuned to 488Hz, our tone frequency. I suggested that we should try my audio. In the past this has not worked, but as scintillation was at an all time low it seemed worth a second try. Once the microphone had made proper contact in its socket the audio came out loud and clear; a 5 by 7 report on the audio was received.

"To get further evidence of this result we contacted Jack, G3JMB, again and he pointed his beam towards Butser to receive David's 70cm transmission. While Jack set-up, David and I kept the conversation going in a cross-band full duplex contact. Jack made contact and David relayed my audio to him. He was then hearing Pulse Width Modulation to a 3mW laser diode being received on a CW optimised receiver and re-transmitted the 50km to his QTH via a hand held rig from a windy hill. It was a surreal moment. Then we reversed the process and I re-transmitted David's CW to Jack. Finally we closed the laser link that had been stable for 30 minutes and packed up."

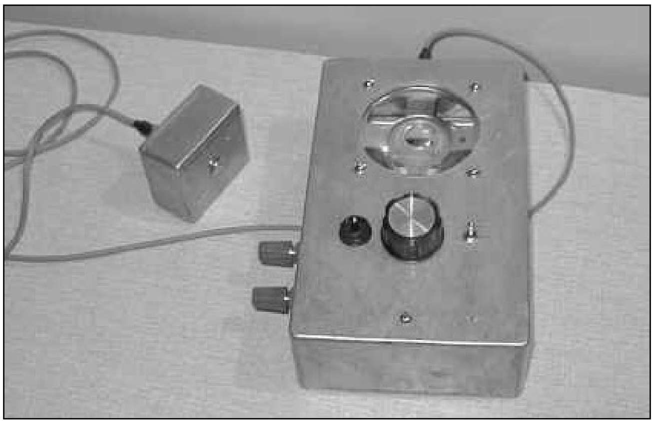


Fig 11.126: Picture of completed laser receiver

sunk for the TO-99 package. The 1N4002 diode protects against reverse polarity. It is soldered to the un-drilled pads on the PCB.

Circuit notes:

- A 10kΩ resistor is needed between Pins 1 and 8 on the OPT301. This is shown on the PCB artwork and on the overlay.
- The capacitor coupling the signal out of the detector to the 39kΩ was changed from 0.1 to 0.22µF The capacitor across the zener diode was also changed to 0.22µF
- With a single supply line the body of the detector is held at the Zener voltage. Therefore it should be isolated from the ground plane and from any metal enclosure.

Four earth connections are required through the board:

- NE5534 pin 4 (on the left above)
- Supply decoupling 4.7µF tantalum (above and right of 5534 pin 8)
- Detector Pin 3
- Pad in lower right of PCB connected to the 10k-ohm resistor and output ground.

A picture of the completed laser receiver is shown in **Fig 11.126**. The performance of the band pass filter and buffer amplifier were evaluated by connecting a 600-ohm signal generator to the input of the band pass filter and measured the response at the output of the buffer. The nput was adjusted from 50Hz to 3kHz with a constant level of 200mV peak to peak. **Table 11.12** shows that the 6dB bandwidth is just 200Hz. This makes the detector ideal for transmitters using modulated CW on a fixed frequency. An OPT301, using 10MΩ feedback with a bandwidth of 200Hz gives a Noise Equivalent Power of 3 x 10^{-11} Watts. The transmitter uses a 4MHz crystal and a CMOS 4060 oscillator / divider to generate 488Hz. This degree of accuracy gives the option of using a laptop computer and modern DSP software (eg Argo [46] or Spectran by IK2PHD) to receive signals 20dB below normal noise level. The 40dB rejection at 50Hz gives a high tolerance to interference from street lighting. Voltage gain at the design frequency is x11 or 20.8dB. The band pass filter centre frequency is selected by two capacitors and two resistors. It can be changed to any audio frequency of your choice. Design equations are published in the *ARRL handbook*.

Frequency Hz	Output mV	Relative -dB
50	20	40.80
100	80	28.80
200	215	20.20
300	430	14.20
350	630	10.86
400	980	7.02
450	1700	2.24
500	2200	0.00
550	1500	3.32
600	1050	6.42
700	620	11.00
800	445	13.90
900	360	15.70
1000	295	17.50
1500	170	22.20
2000	120	25.30
2500	95	27.30
3000	80	28.80

Table 11.12: Band pass filter / buffer amplifier characteristics for laser receiver

Laser Transmitter Circuits

Here are two circuits for laser transmitters designed by David Bowman, G0MRF. The first, shown in **Fig 11.127**, uses a simple laser pointer module. Any of the normal 3 or 5mW devices will function well. Imports from the USA or Hong Kong are frequently a higher power than the UK-approved units. The modules normally contain two or three 1.5V mercury cells.

The external modulated supply should duplicate the voltage used in your particular module.

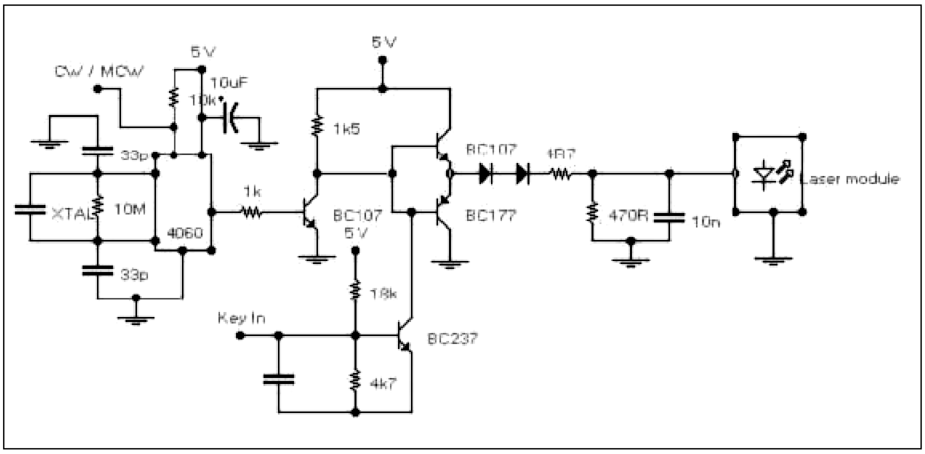
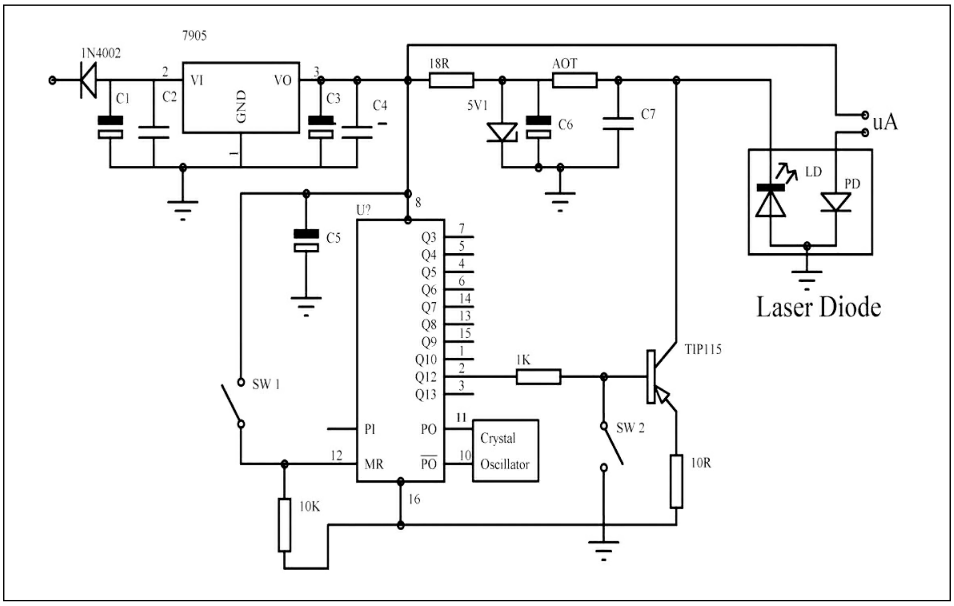


Fig 11.127: Circuit diagram of laser transmitter using a laser pointer. 4060 pins: 16; supply, 8; ground, 2; output, 10+11; crystal, 12; reset

The electronics contained in the module will include the semiconductor laser diode and a constant current source. The ability of the constant current source to be switched on and off at audio frequency will change between manufacturers, but below 1KHz, this is not normally a problem.

However, there are applications where the supply can be switched at 20 to 200kHz and voice or high-speed data can be modulated onto this subcarrier. For these high-speed applications the circuit should be tested carefully with an oscilloscope. If unsuitable, the constant current source can be modified or replaced with a circuit that has a faster response.

Fig 11.128: Circuit diagram of laser transmitter using a laser diode



The two diodes are 1N4001 etc. The capacitor on the key input is 10nF. This circuit uses a crystal and a 4060 oscillator / divider to produce a square wave output in the audio range; in this case a 4MHz crystal giving 488Hz on pin 2. The 5Vp-p square wave is buffered by three transistors which include a totem pole driver. Two diodes have been used to reduce the voltage by 1.2V which then supplies the laser module. A switch is attached to the reset pin of the 4060. This is used to select CW (constant laser output) or Modulated CW. Another switch, wired between the key input and ground, allows the laser to be keyed via a phono socket or it can be switched permanently on.

WARNING: Lasers are dangerous

- The dangers from lasers are essentially from the amount of energy contained in a very small area. If a narrow beam is used, then all of the power can be directed into the typical 5mm diameter of the human eye causing significant damage.
- Whenever possible, ensure the beam created in home constructed or converted equipment has its beam expanded. If the beam is expanded to larger diameter then its energy is contained over a larger area and the danger of accidental eye damage is much reduced. Including a beam expander to your laser system also has the advantage of reducing the divergence of the beam, increasing the distance potential of the transmitter.
- Never leave a laser transmitter on and unattended as you are not in control of where the laser is pointing. If you move away from the transmitter always turn the laser off.
- Consider any beam, even that generated by a cheap laser pointer, as potentially dangerous until its beam has diverged to a minimum of 50mm diameter.
- Night-time eye response: At night and in dark conditions, the pupil in the human eye will dilate to increase sensitivity to light. This significantly increases the danger from lasers.
- Visible wavelength lasers can be seen by the operators and others, and their danger anticipated. However, special care should be taken when using invisible wavelengths. Infra red lasers are generally more powerful but the risks from accidentally leaving a laser source on are significantly increased. By the time you realise the source is on, it may already be too late. Consider a beam expander as mandatory with non visible wavelengths.

Further information on laser safety can be obtained from the Health and Safety Executive we site http://www.hse.gov.uk

This is very useful when aligning the beam onto a distant receiver. Some circuits have been published using 555 timers but it was felt that a crystal source would give greater potential for weak signal working.

Caution:

- Always use talkback to coordinate transmissions with the other station. Looking at a laser over long distances through your receive optics can be dangerous.
- Most laser modules use positive earth.
- Heating the module to solder wires can damage the optics and increase the beam divergence.
- A laser diode can fail but still appear to be working. However, what you'll have is an expensive LED with a light output that is a fraction of the laser and a beam that is not a coherent light source.
- As the laser cools down on a cold winter's night, the light output power will increase. This is not the advantage it first appears as the laser can fail because of excessive output power.

If you want to increase your laser output power, you will probably have to move on from laser pointers and use a laser diode. Then it's down to "how much you want to spend", but at least you get to choose the output power and wavelength. The circuit shown in **Fig 11.128** is adapted from an original idea suggested by K3PGP. It uses a 7905 negative voltage regulator so that the body of the laser diode can be connected to ground. The audio is generated by a 4060 as above and a PNP Darlington transistor is wired across the laser diode as a shunt modulator. You will need optics to collimate the output from a semiconductor laser. The angle of radiation of a prototype was 9 degrees high by 32 degrees wide. The circuit uses two resistors to limit the laser diode current. The resistor labelled AOT (adjust on test) should be selected *very* carefully. It should be the correct value to provide just less than the maximum laser power from the diode. The original transmitter output was set to 10mW, well inside the max ratings of the diode. Audio is generated by a 4060. The 4MHz crystal, 10MΩ resistor and two 33pF capacitors have been omitted from the diagram for clarity but are the same as used in the other circuit described earlier. C6 is 47µF, all other electrolytics are 10µF. Other capacitors are 10nF. Modulation is applied to the laser via a shunt modulator. When the Darlington transistor conducts, the current supplied from the 7905 and the two resistors is shunted to ground through the transistor and the 10Ω resistor in its emitter. This circuit can be used with lasers up to 100mW output.

The completed PCB (**Fig 11.129** in Appendix B) is mounted on the side wall of a diecast box. The optics are contained in a small metal tube simply glued onto the PCB. A coarse adjustment of the laser can be made by carefully adjusting the position of the PCB on the four mounting bolts. The final adjustment is made with the cross hairs on the telescopic sight which is a cheap 4 x 28 (times 4 magnification and 28mm lens diameter) bought new for £19.95 from sources on the Internet.

BIBLIOGRAPHY

Books

ARRL Handbook for Radio Amateurs, ARRL, revised / re-published annually

DubusTechnik, available in the UK from G4PMK (QTHR). E-mail: dubus@marsport.demon.co.uk

International Microwave Handbook, editor Andy Barter, G8ATD, RSGB, 2002

Microwave Projects, editor Andy Barter, G8ATD, RSGB, 2003

UHF/ Microwave Experimenter's Handbook, ARRL

UHF/Microwave Projects Manual, Volumes 1 & 2, ARRL

UHF Compendium, Parts I, II, III and IV, editor K Weiner, DJ9HO, available from UKW Berichte. E-mail: info@ukwberichte.com, or from KM Publications, E-mail: andy@vhfcomm.co.uk

Magazines

CQ-TV, BATC Publications, Fern House, Church Road, Harby, Notts, NG23 7ED. E-mail: publications@batc.org.uk. Web: http://www.batc.org.uk.

Dubus magazine, available in the UK from G4PMK (QTHR). E-mail: dubus@marsport.demon.co.uk.

Proceedings of the Microwave Update, ARRL, annual publication

RadCom, RSGB members' monthly journal; it has a bi-monthly microwave column and carries occasional microwave articles. Web: http://www.rsgb.org.

Scatterpoint (formerly *Microwave Newsletter*), UK Microwave Group, 10 issues per year, contact: Martyn Kinder, G0CZD, Secretary UK Microwave Group, 12 Jessop Way, Haslington, Crewe, Cheshire, CW1 5FU. E-mail: ukug@czd.org.uk. Web: http://www.microwavers.org.

VHF Communications magazine, KM Publications, 63 Ringwood Road, Luton, Beds, LU2 7BG. E-mail: andy@vhfcomm.co.uk. Web: http://www.vhfcomm.co.uk.

Websites

A list of amateur microwave websites is given on the RSGB site at http://www.rsgb.org/books/extra/handbook.htm instead of here as URLs may change. Web users should use their preferred search engine to find the required website if the links given have changed.

Component or Kit Suppliers

Down East Microwave Inc, 954 Rt. 519 Frenchtown, NJ 08825, USA. Web: http://www.downeastmicrowave.com.

Farnell Electronic Components Limited, Canal Road, Leeds, LS12 2TU. Tel: 0870 1200 200. Fax: 0870 1200 296. Web: http://www.farnellinone.co.uk.

GH Engineering, The Forge, West End, Sherborne St. John, Hants, RG24 9LD. Tel: 01256 889295. Fax: 01256 889294. E-mail: sales@ghengineering.co.uk. Web: http://www.ghengineering.co.uk

Kuhne electronics GmbH, Scheibenacker 3, D-95180 Berg / Oberfranken, Germany. E-mail: info@kuhne-electronic.de. Web: http://www.kuhne-electronic.de or http://www.db6nt.com.

Maplin Electronics, P.O. Box 777, Wombwell, S73 0ZR. Tel: 0870 429 6000. Fax: 0870 429 6001. Web: http://www.maplin.co.uk.

Microwave Committee Component Service, P Suckling, 314A Newton Road, Rushden, Northants, NN10 0SY. Tel: 01933 411446. Fax: 01933 411446. Web: http://www.g3wdg.free-online.co.uk.

Mini-Circuits Europe, Dale House, Wharf Road, Frimley Green, Camberley, Surrey, GU16 6LF. Tel: 01252 832600. Fax: 01252 837010. Web: http://www.minicircuits.com.

R F Elettronica, Via Dante 5, 20030 Senago, MI, Italy. E-mail: info@rfmicrowave.it. Web: http://www.rfmicrowave.it.

UKW Berichte, Postfach 80, D-91081 Baiersdorf, Germany. E-mail: info@ukwberichte.com. Web: http://www.ukwberichte.com.

Some kits available in the UK from KM Publications, 63 Ringwood Road, Luton, Beds, LU2 7BG. Tel 01582 581051. Fax: 01582 581051. E-mail: andy@vhfcomm.co.uk. Web: http://www.vhfcomm.co.uk.

REFERENCES

[1] CCIR, 1937.

[2] *Admiralty Handbook of Wireless Telegraphy*, HMSO, London, 1938.

[3] *International Microwave Handbook*, editor - Andy Barter, G8ATD, RSGB 2002.

[4] *Microwave Projects*, editor - Andy Barter, G8ATD, RSGB 2003.

[5] *UHF/Microwave Experimenter's Manual*, 'Antennas, components and design', ARRL.

[6] Agilent (Hewlett Packard). Web: http://www.semiconductor.agilent.com.

[7] Mini Kits. Web: http://www.minikits.com.au.

[8] *ARRL Proceedings of Microwave Update* 1987.

[9] 'A local oscillator source for 1153MHz', S T Jewell, G4DDK, *Radio Communication*, February 1987, p218 and March 1987, pp191–201.

[10] Microwave Component Service, P Suckling, 314A Newton Road, Rushden, Northants, NN10 0SY, Tel: 01933 411446. Fax: 01933 411446. Web: http://www.g3wdg.free-online.co.uk.

[11] 'High precision frequency standard for 10MHz. Part II: Frequency control via GPS', Wolfgang Schneider, DJ8ES and Frank-Peter Richter, DL5HAT, *VHF Communications* 1/2001 pp2–8

[12] 'High precision frequency standard for 10MHz', Wolfgang Schneider, DJ8ES, *VHF Communications* 4/2000 pp177–190.

[13] 'HMET LNAs for 23cm', R Bertelsmeier, DJ9BV, *Dubus Technik IV*, pp191–197.

[14] 'No tune HEMT preamp for 13cm', R Bertelsmeier, DJ9BV, *Dubus Technik IV*, pp191–197.

[15] Franck Rousseau, F4CIB, 18 rue Colbert, Porte B AppT 31, 31400 Toulouse, France. E-mail: f4cib@ref-union.org.

[16] 'Amplificateur 4W 13cm à transistor LDMOS', Maurice Niquel, F5EFD, *Hyper* n°56, February 2001.

[17] 'Amplificateur 70W sur 1296MHz', Maurice Niquel, F5EFD, *Hyper* n°65, November 2001.

[18] 'Amplificateur 2300MHz à MRF282; 12 ou 24VDC', Jean-

Pierre Lecarpentier, F1ANH, *Hyper* n°73, July/August 2002.
[19] http://e-www.motorola.com/brdata/PDFDB/docs/MRF284.pdf
[20] '28V switched powers supply for LDMOS-PAs', Achim Volthardt, DH2VA / HB9DUN, *Dubus*, 1/2004 pp38–40.
[21] 'GaAsFET power amplifier stage up to 5W for 10GHz', Paeter Volg, DL1RQ, *VHF Communications Magazine*, 1/1995, pp.52–63.
[22] '6cm transverters in modern stripline technology', Peter Vogl, DL1RQ, *VHF – UHF* Munich 1990, pp49–66.
[23] '6cm transverters in stripline technology, part 2', Peter Vogl, DL1RQ, *VHF Communication Magazine* 2/1991, pp69–73.
[24] 'Determination of parasitic inductances on capacitors', Roland Richer, Special project at Dominicus vo Linprun Grammar School Viechtach, 1994.
[25] '76GHz amplifier', Sigurd Werner, DL9MFV, *VHF Communications Magazine*, 3/2003, pp163-169.
[26] 'A simple concept for a 76GHz transverter', Sigurd Werner, DL9MFV, *VHF Communications Magazine*, 2/2003, pp77-83.
[27] 'Medium Power Amplifier @ 77GHz', Infineon Technologies Data Sheet, T 602B_MPA_2; 05.05.99.
[28] Data Sheet IAF-MAP7710, Fraunhofer Institut for Applied Solid-State Physics, Freiburg.
[29] 'Frequency multiplier for 76GHz with an integrated amplifier', Sigurd Werner, DL9MFV, *VHF Communications Magazine*. 1/2002, pp35-41.
[30] 'Amplifier for 47GHz using chip technology', Sigurd Werner, DL9MFV, *VHF Communications Magazine*, 3/2002, pp160–164.
[31] 'Combining power at 76GHz: Three possible solutions discussed', Sigurd Werner, DL9MFV, *VHF Communications Magazine*, 1/2004, pp13–19.
[32] From article by Matjaz Vidmar, S53MV, *CQ ZRS*
[33] 'A single-board, no-tuning 23cm transverter', Richard L Campbell, KK7B, 23rd Conference of the Central States VHF Society, Rolling Meadows, Illinois, 1989, pp44–52 and 'Engineering Notes', ibid, pp53–55. Subsequently republished in *The ARRL Handbook for the Radio Amateur*, 69th edition, 1992.
[34] 'A 1W linear amplifier for 1152MHz', Sam Jewell, G4DDK, *RSGB Microwave Handbook Volume 2*, pp8.21–8.23.
[35] Richard L Campbell, KK7B, *Proceedings of the Microwave Update*, ARRL, 1988.
[36] Down East Microwave, 954 Rt. 519 Frenchtown, NJ 08825, USA. Web: http://www.downeastmicrowave.com.
[37] Richard L Campbell, KK7B, *Proceedings of the Microwave Update*, ARRL, 1989.
[38] 'VHF and microwave applications of monolithic microwave integrated circuits', Al Ward, WB5LUA, *ARRL UHF/Microwave Experimenter's Manual*, ARRL, 1990. pp7.32–7.47.
[39] *Microwave Handbook, Vol 3*, edited by M W Dixon, G3PFR, RSGB, 1992, pp18.116–18.118.
[40] Microwave Components Service, c/o Mrs P Suckling, G4KGC, 132A Newton Road, Rushden, Northants NN10 0SY, UK.
[41] For further conversion and materials availability information contact Chuck Houghton, WB6IGP, E-mail: clhough@pacbell.net, or Kerry Banke N6IZW of the San Diego Microwave Group, E-mail: kbanke@qualcomm.com.
[42] Additional conversion information articles and sources: 'Microwave GaAs FET Amps for Modification to 10GHz', C Houghton, WB6IGP & Kerry Banke, N6IZW, *Nts Feedpoint Newsletter*, December, 1993. 'UP, UP & Away to 10GHz Semi-Commercial Style', Bruce Wood, N2LIV, *Proceedings of the 20th Eastern VHF/UHF Conference*, August, 1994, p133. '10GHz Qualcomm Modification Notes' by Dale Clement, AF1T.
[43] 'Suggestions for Modifications of Qualcomm LNA Board for 10GHz', Ken Shofield, W1RIL, *Proceedings of the 21st Eastern VHF/UHF Conference*, August, 1995, p63 and 'Modification Update of Omnitrack PA Board for 10GHz'. p65. Also at http://www.uhavax.hartford.edu/disk$user-data/faculty/newsvhf/www/w1ril.html
[44] Laser communications. Web http://www.lasercomms.org.uk/basics.htm
[45] 'A simple laser communications system', David Bowman, G0MRF. Web http://www.g0mrf.freeserve.co.uk/laser.htm
[46] Argo. Web http://www.weaksignals.com