11 Practical Microwave Receivers & Transmitters

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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cycles per second (c/s), its decimal multiples, kilocycles per second (kc/s) and megacycles per second (Mc/s), have been replaced by the unit hertz (Hz), its decimal multiples kilohertz = 10^3Hz (kHz), megahertz = 10^6Hz (MHz), gigahertz = 10^9Hz (GHz) and terahertz = 10^{12}Hz (THz).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 is 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 are example articles later in this chapter.

Other advantages of operation in the microwave spectrum are compact, high-gain antennas and available bandwidth. None of these advantages is 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 quick search with one of the popular search engines, using the relevant key words, should find information about the software.

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, 4, 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

Allocation (MHz)	Amateur Status	Narrow band segment centre of activity (MHz)
1,240 - 1,325	Secondary	1,296.200
2,310 - 2,450	Secondary	2,320.200
3,400 - 3,475	Secondary	3,410 (EME)
5,650 - 5,680	Secondary	5,668.200
5,755 - 5,765	Secondary	5,760.100
5,820 - 5,850	Secondary	
10,000 - 10,125	Secondary	
10,225 - 10,475	Secondary	10,368.100
10,475 - 10,500	Secondary	Amateur Satellite Service only
24,000 - 24,050	Primary	24,048.200
24,050 - 24,250	Secondary	
47,000 - 47,200	Primary	47,088.200
75,500 - 75,875	Secondary	
75,875 - 76,000	Primary	75,976.200
76,000 - 77,500	Secondary	
77,500 - 78,000	Primary	77,500.200
78,000 - 81,00	Secondary	
122,250 - 123,000	Secondary	
134,000 - 136,000	Primary	
136,000 - 141,000	Secondary	
241,000 - 248,000	Secondary	
248,000 - 250,000	Primary	

Table 11.1: UK Amateur Service allocations 2009

Allocation (MHz)	Amateur Status	Comments
1,260 - 1,270	Secondary	ETS
2,400 - 2.450	Secondary	ETS/STE
5,650 - 5,668	Secondary	ETS
5,830 - 5,850	Secondary	STE
10,475 - 10,500	Secondary	ETS/STE
24,000 - 24,050	Primary	ETS/STE
47,000 - 47,200	Primary	ETS/STE
75,500 - 75,875	Secondary	ETS/STE
75,875 - 76,000	Primary	ETS/STE
76,000 - 77,500	Secondary	ETS/STE
77,500 - 78,000	Primary	ETS/STE
78,000 - 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
241,000 - 248,000	Secondary	ETS/STE
248,000 - 250,000	Primary	ETS/STE

ETS = Earth to Space, STE = Space to Earth

Table 11.2: UK Amateur Satellite Service allocations 2009

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

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 frequency 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 segment centres of activity 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 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

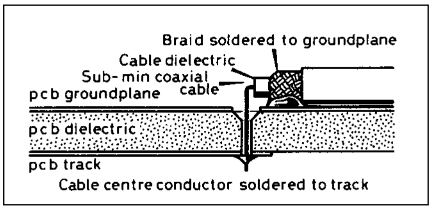


Fig 11.1: Correct coaxial cable connection technique

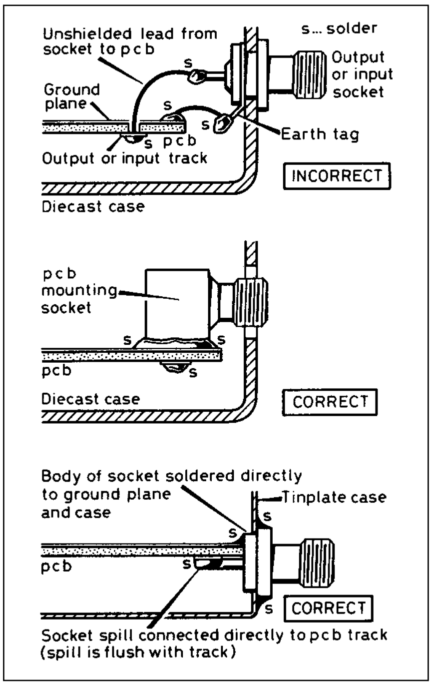


Fig 11.2: Correct and incorrect methods of connecting coaxial sockets to microwave PCBs

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

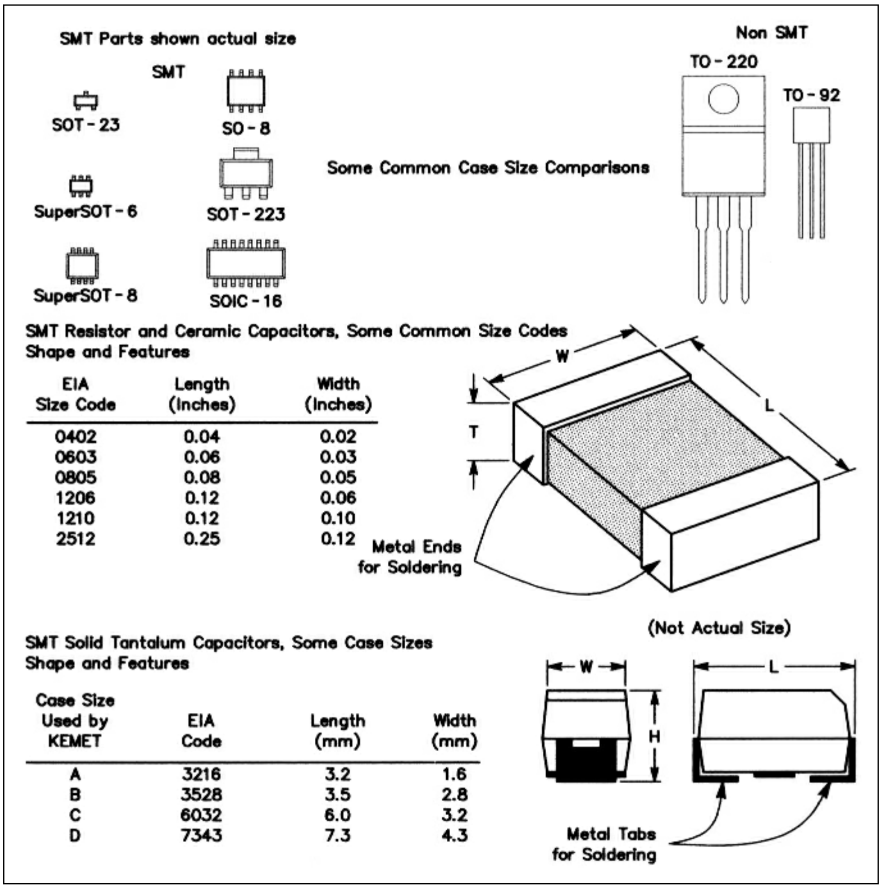
Earthing is a very important topic for the microwave constructor. As mentioned above, the earth connections from one side of a PCB to the other must be as good as possible to reduce the effects of stray inductance. The same is true for all other earth connections; they must be as short and solid as possible. 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 50-ohm coaxial cable as shown in Fig 11.1. Do not take the output away as shown in Fig 11.2, this is disastrous as it will almost certainly cause mismatch, stray inductive losses and may detune the output lines so that they will not resonate properly.

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



EIA Size Code	Length (Inches)	Width (Inches)
0402	0.04	0.02
0603	0.06	0.03
0805	0.08	0.05
1206	0.12	0.06
1210	0.12	0.10
2512	0.25	0.12

Case Size Used by KEMET	EIA Code	Length (mm)	Width (mm)
A	3216	3.2	1.6
B	3528	3.5	2.8
C	6032	6.0	3.2
D	7343	7.3	4.3

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

Part number	Description	Bias voltage (V)	Bias current (mA)	NF (dB)
MGA-13516	High gain, high linearity, active bias, low noise amplifier	5	54	0.66
MGA-14516	High gain, high linearity, active bias, low noise amplifier	5	45	0.66
MGA-30116	750MHz - 1GHz 1/2W high linearity amplifier	5	202.8	2.0
MGA-30216	1.7 - 2.7GHz 1/2W high linearity amplifier	5	206	2.8
MGA-30316	3.3 - 3.9GHz 1/2W high linearity amplifier	5	198	2.7
MGA-52543	5V LNA, 32dBm OIP3, 0.4 - 6GHz, SOT343(SC-70)	5	53	1.9
MGA-53543	5V high linearity LNA, 39dBm OIP3, 0.44 - 6GHz, SOT343(SC-70)	5	54	1.5
MGA-53589	50MHz to 6GHz high linearity amplifier	5	54	1.66
MGA-545P8	Low current 22dBm medium power amplifier in LPCC2x2 for 5 - 6GHz	3.3	92	4.4
MGA-565P8	20dBm Psat high isolation buffer amplifier	5	67	
MGA-61563	Current adjustable low noise amplifier	3	41	1.2
MGA-62563	Current adjustable low noise amplifier	3	60	0.9
MGA-631P8	Low noise, high linearity amplifier with active bias	4	60	0.5
MGA-632P8	Low noise, high linearity amplifier with active bias	4	60	0.6
MGA-665P8	0.5 - 6GHz low noise amplifier	3	21	1.5
MGA68563	Current adjustable low noise amplifier	3	11	1.0
MGA-685T6	Current adjustable low noise amplifier	3	10	0.9
MGA-71543	3V LNA with bypass switch, 0 to 9dBm adjustable IIP3, SOT343(SC-70)	2.7	10	0.8
MGA-72543	3V LNA with bypass switch, 2 to 14dBm adjustable IIP3, SOT343(SC-70)	2.7	20	1.4
MGA-725M4	3V LNA with bypass switch, 2 to 14dBm adjustable IIP3, SOT343(SC-70)	2.7	20	1.3
MGA-785T6	Low noise amplifier with bypass switch	3	7	1.1
MGA-81563	3V driver amplifier. 14dBm P1dB, low noise, 0.1 - 6GHz, SOT363(SC-70)	3	42	2.8
MGA-82563	3V driver amplifier. 14dBm P1dB, low noise, 0.1 - 6GHz, SOT363(SC-70)	3	84	2.2
MGA-83563	3V PA/driver amplifier. 22dBm PSAT, 0.5 - 6GHz, SOT363(SC-70)	3	152	
MGA-85563	3V LNA, 12 to 17dBm adjustable OIP3, 0.8 - 6GHz, SOT363(SC-70)	3	15 - 30	1.9
MGA-86563	5V LNA. 20dB high gain, 0.5 - 6GHz, SOT363(SC-70)	5	14	1.5
MGA-86576	5V LNA. 23dB high gain, 1.5 - 8GHz, SOT363(SC-70)	5	16	1.6
MGA-87563	3V LNA, 4.5mA low current, 0.5 - 4GHz, SOT363(SC-70)	3	4.5	1.6

Table 11.4: Data for a selection of Avago GaAs MMICs (Copyright Avago, reproduced with their permission)

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 Avago (formerly Agilent or 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 Avago website [6] and the Minicircuits website [7]. Some useful information from these suppliers is reproduced in **Tables 11.4 and 11.5**.

The Avago 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 - 2,500MHz, making it useful for applications that operate in the 2,400 to 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

Model Number	Frequency range (MHz)		Gain (dB). Typ	Max. Power output @ 1dB comp. (dBm) Typ.	N.F. (dB) Typ.	IP3 (dBm) Typ.	VSWR (:1) Typ.		Device DC operating power	
	Low	High					In	Out	Voltage (V)	Current (mA)
ERA-1+	DC	8000	10.9	12.0	4.3	26.0	1.5	1.5	3.4	40
ERA-2+	DC	6000	14.4	13.0	4.0	26.0	1.3	1.2	3.4	40
ERA-3+	DC	3000	18.7	12.5	3.5	25.0	1.5	1.4	3.2	35
ERA-4+	DC	4000	13.4	17.3	4.2	34.0	1.2	1.3	4.5	65
ERA-5+	DC	4000	18.5	18.4	4.3	32.5	1.3	1.2	4.9	65
ERA-6+	DC	4000	12.2	17.9	4.5	36.0	1.3	1.6	5.0	70
ERA-1SM+	DC	8000	10.9	12.0	4.3	26.0	1.5	1.5	3.4	40
ERA-2SM+	DC	6000	14.4	13.0	4.0	26.0	1.3	1.2	3.4	40
ERA-21SM+	DC	8000	12.2	12.6	4.7	26.0	1.1	1.3	3.5	40
ERA-3SM+	DC	3000	18.7	12.5	3.5	25.0	1.5	1.4	3.2	35
ERA-33SM+	DC	3000	17.4	13.5	3.9	28.5	1.6	1.25	4.3	40
ERA-4SM+	DC	4000	13.4	17.3	4.2	34.0	1.2	1.3	4.5	65
ERA-4XSM+	DC	4000	13.5	17.0	4.2	35.0	1.3	1.3	4.5	65
ERA-5SM+	DC	4000	17.6	18.4	4.3	32.5	1.3	1.2	4.9	65
ERA-5XSM+	DC	4000	17.6	17.8	3.5	33.0	1.3	1.3	4.9	65
ERA-50SM+	DC	2000	19.4	17.2	3.5	32.5	1.3	1.2	4.4	60
ERA-51SM+	0C	4000	16.1	18.1	4.1	33.0	1.1	1.2	4.5	65
ERA-6SM+	DC	4000	12.2	17.9	4.5	36.0	1.3	1.6	5.0	70
ERA-8SM+	DC	2000	19.0	12.5	3.1	25.0	1.4	1.8	3.7	36
GALI-1+	DC	8000	11.8	10.5	4.5	27.0	1.3	1.4	3.4	40
GALI-19+	DC	7000	11.6	9.0	6.5	23.7	1.6	1.5	3.6	40
GALI-2+	DC	8000	14.8	11.0	4.6	27.0	1.6	1.6	3.5	40
GALI-21+	DC	8000	13.1	10.5	4.0	27.0	1.1	1.3	3.5	40
GALI-24+	DC	6000	16.6	19.3	4.3	35 3	1.4	2.0	5.8	80
GALI-29+	DC	7000	19.7	10.0	6.0	24.7	1.5	1.5	3.6	40
GALI-3+	DC	3000	19.1	10.5	3.5	25.0	1.5	1.2	3.3	35
GALI-33+	DC	4000	17.5	11.4	3.9	28.0	1.6	1.2	4.3	40
GALI-39+	DC	7000	19.7	9.0	2.4	22.9	1.6	1.5	3.5	35
GALI-4+	DC	4000	13.5	16.0	4.0	34.0	1.2	1.4	4.6	65
GALI-4F+	DC	4000	13.4	13.8	4.0	32.0	1.2	1.5	4.4	50
GALI-49+	DC	5000	13.6	15.0	5.5	33.3	1.7	1.5	5.0	65
GALI-5+	DC	4000	17.5	16.0	3.5	35.0	1.2	1.4	4.4	65
GALI-5F+	DC	4000	17.4	14.2	3.5	31.5	1.2	1.4	4.3	50
GALI-51+	DC	4000	16.1	16.5	3.5	35.0	1.3	1.5	4.5	65
GALI-51F+	DC	4000	15.9	14.4	3.5	32.0	1.2	1.5	4.4	50
GALI-52+	DC	2000	17.8	15.5	2.7	32.0	1.35	1.4	4.4	50
GALI-55+	DC	4000	18.5	15.5	3.3	28.5	1 25	1.3	4.3	50
GALI-59+	DC	5000	18.3	17.6	4.3	33.3	1.6	1.5	4.8	65
GALI-6+	DC	4000	11.3	18.2	4.5	35.5	1.5	1.8	5.0	70
GALI-6F+	DC	4000	11.6	15.8	4.5	35.5	1.5	1.9	4.8	50
GALI-74+	DC	1000	21.8	18.3	2.7	38.0	1.2	1.6	4.8	80
GALI-84+	DC	6000	16.7	21.0	4.4	37.4	1.4	2.1	5.8	100
GALI-S66+	DC	3000	17.3	2.8	2.7	18.0	1.25	1.7	3.5	16
MAR-1+	DC	1000	16.5	2.5	3.5	14.0	1.3	1.2	5.0	17
MAR-1SM+	DC	1000	16.5	2.5	3.3	14.0	1.3	1.2	5.0	17
MAR-2SM+	DC	2000	12.0	7.0	3.7	22.0	1.3	1.3	5.0	25
MAR-3+	DC	2000	12.0	10.0	6.0	23.0	1.5	1.7	5.0	35
MAR-3SM+	DC	2000	12.0	10.0	3.7	28.0	1.3	1.3	5.0	35
MAR-4+	DC	1000	8.0	12.5	7.0	25.5	1.5	1.9	5.25	50
MAR-4SM+	DC	1000	8.0	12.5	6.0	25.5	1.6	2.0	5.3	50
MAR-6+	DC	2000	20.0	3.0	3.0	14.5	1.7	1.7	3.5	16
MAR-6SM+	DC	2000	20.0	3.0	3.0	14.5	1.3	1.3	3.5	16
MAR-7SM+	DC	2000	12.5	3.5	5.0	19.0	1.3	1.3	4.0	22
MAR-8A+	DC	1000	25.0	12.5	3.1	25.0	1.4	1.8	3.7	36
MAR-8ASM+	DC	1000	25.0	12.5	3.1	25.0	1.4	1.8	3.7	36
MAR-8SM+	DC	1000	22.5	12.5	3.3	27.0	3.1	3.1	7.8	36

Table 11.5: Extract from Mini Circuits [7] web page showing MMIC characteristics

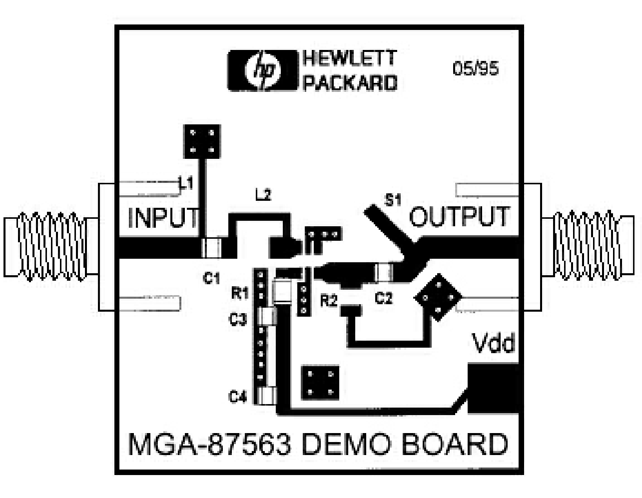


Fig 11.4: Agilent PCB for 2,400MHz LNA. For more details see [6]**

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 11.6**. 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 11.5** - **11.7** were taken on a scalar analyser and show the performance of the amplifier.

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

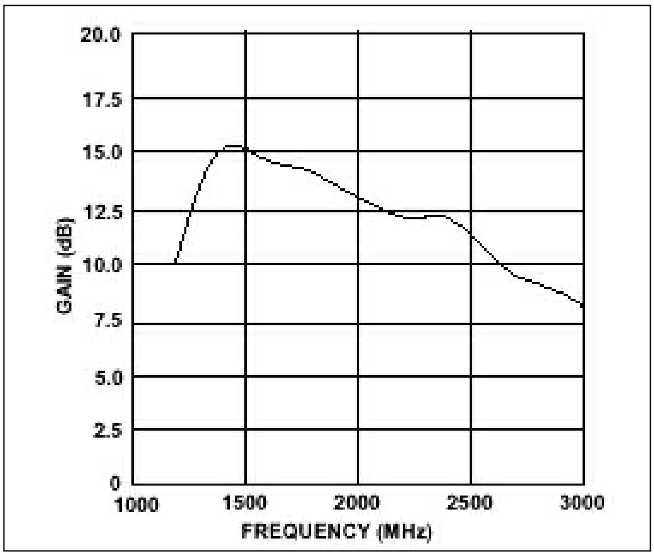


Fig 11.5: 2,400MHz LNA associated gain at maximum noise figure. With Vd = 5V

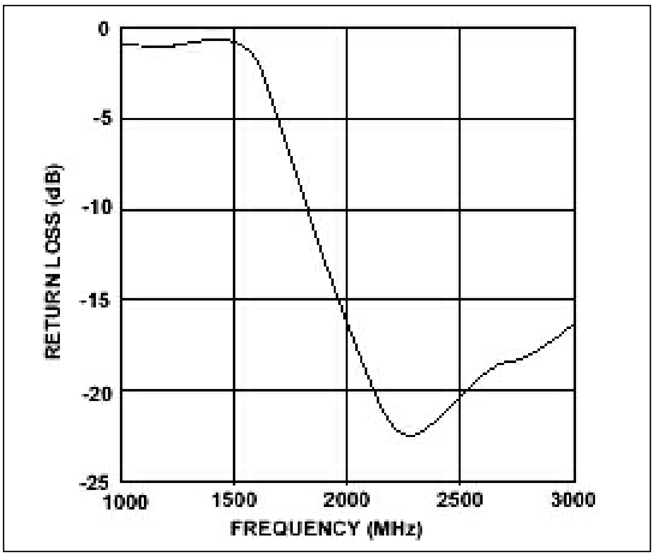


Fig 11.6: 2,400MHz LNA output return loss. With Vd = 5V

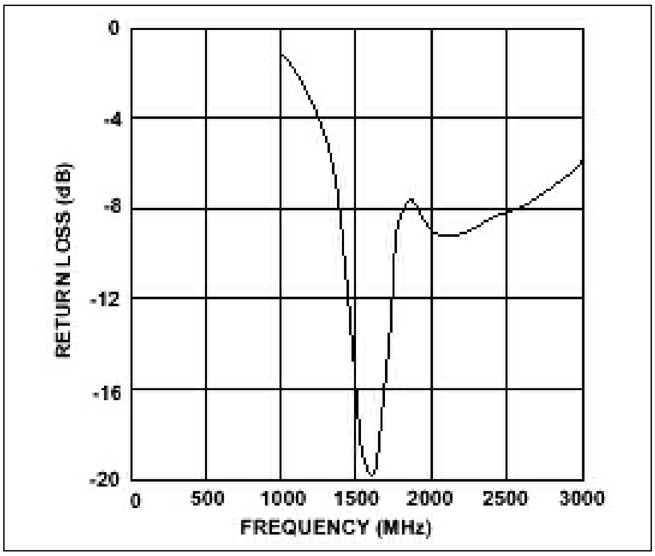


Fig 11.7: 2,400MHz LNA input return loss. With Vd = 5V

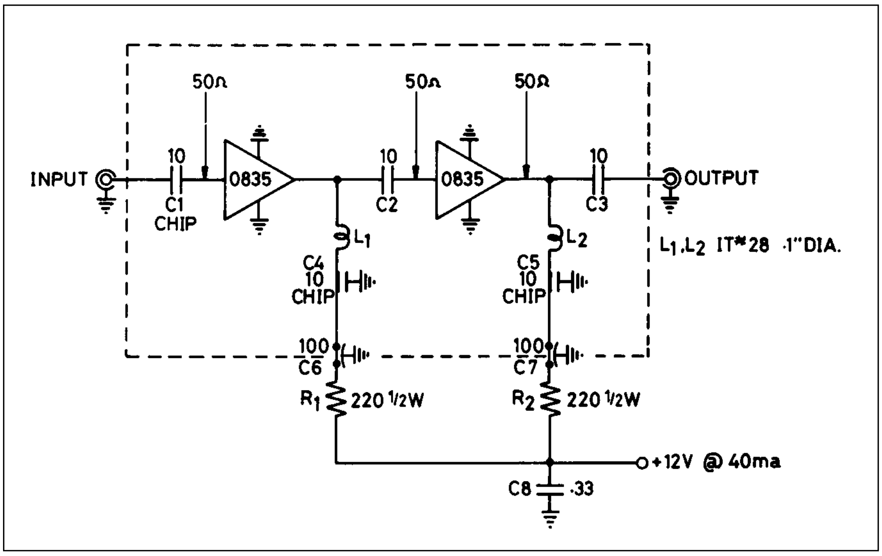


Fig 11.8: Circuit of the two stage MSA0835 wide band amplifier

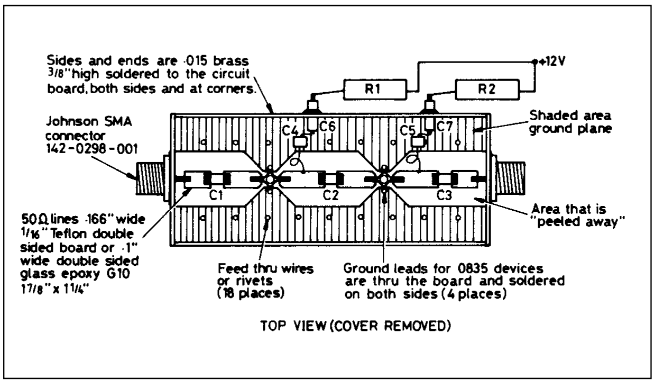


Fig 11.9: Layout of components in the two stage MSA0835 wide band amplifier

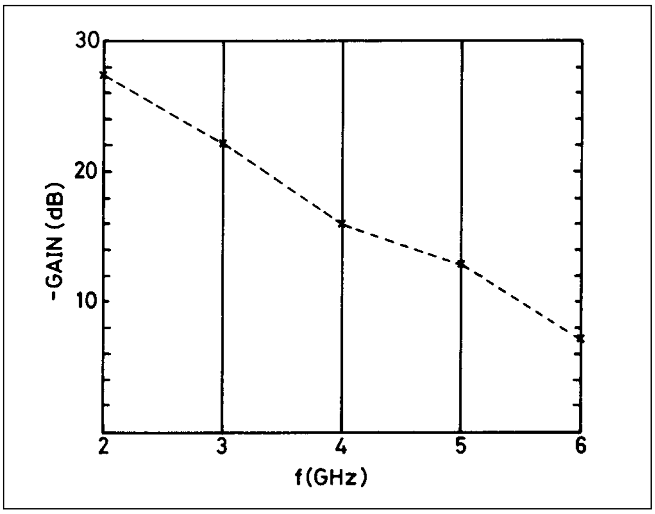


Fig 11.10: The frequency response of the two stage MSA0835 wide band amplifier

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 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 10pF 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 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.

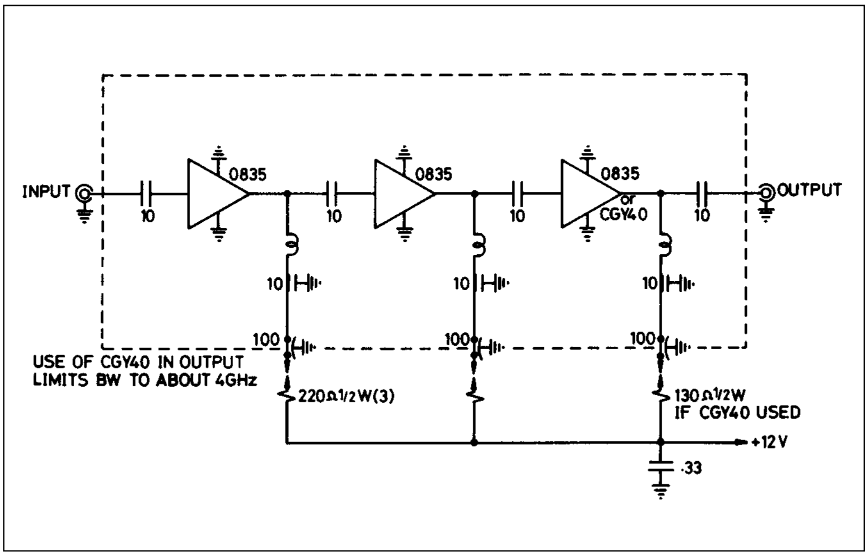


Fig 11.11: Circuit of the three stage MSA0835 wide band amplifier

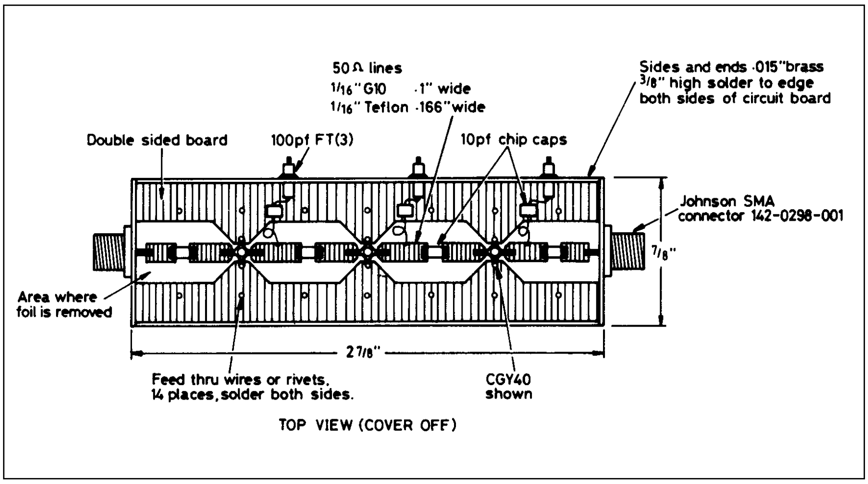


Fig 11.12: Layout of the components in the three stage MSA0835 wide band amplifier

DETERMINING THE PERFORMANCE OF A TERRESTRIAL MICROWAVE RECEIVER SYSTEM

What constitutes acceptable performance for an amateur microwave terrestrial receiver system? This can be a very difficult question to answer, as it will depend on location and operating aspirations.

The usual amateur radio approach is to use a high-gain, low-noise preamplifier, preferably at masthead, into the station transverter or transceiver. While this will undoubtedly provide a very sensitive receiver system, an ever-growing number of commercial digital radio-based systems means that our receivers are now being bombarded with many high level out-of-band signals as well as strong in-band signals from other amateurs and primary users. These signals may cause the appearance of intermodulation products as well as odd noises due to reciprocal mixing right on top of a weak station you are trying to work. No longer is it advisable to "pile on the pre-amplifiers" as we have done in the past, to achieve an arbitrarily low system noise figure. A good receive system design must now take account of the strong signal performance of the entire receiver as well as its noise figure. But how do you decide what is a good strong signal performance?

All amplifiers are non-linear. Low-noise pre-amplifiers are especially poor in this respect. The onset of non-linear operation is normally associated with the output signal level no longer increasing at the same rate as the input signal level increases. When the amount of increase is 1dB less than it should be, this is known as the 1dB compression point and is a common way of expressing the linearity of an amplifier, especially a power amplifier. While the compression point can be very useful, a better measure is something called the intercept point (IP). All amplifiers have an infinite number of intercept points, depending on the number of input signals and the order of the product generated from these. In general, only the second- (IP2) and third-order (IP3) are required to characterise strong signal performance. In this section, we will be looking mainly at IP3, as this is the parameter used in the *AppCAD* software programme. It is also important to realise that intercept points can refer to both input (IIP2, IIP3) and output (OIP2, OIP3).

Intercept points are theoretical values, which cannot be directly measured, but which can be extrapolated from measurements of the input or output signal level and the level of the products generated within the amplifier from these signals. There are many variations on this depending on the number of input signals, and the order of the harmonic mix. The one of interest here is the mixing of the second harmonic of one signal plus or minus the fundamental of a second signal. This is the classic third order product. See **Fig 11.14** for a graphical explanation of third-order product.

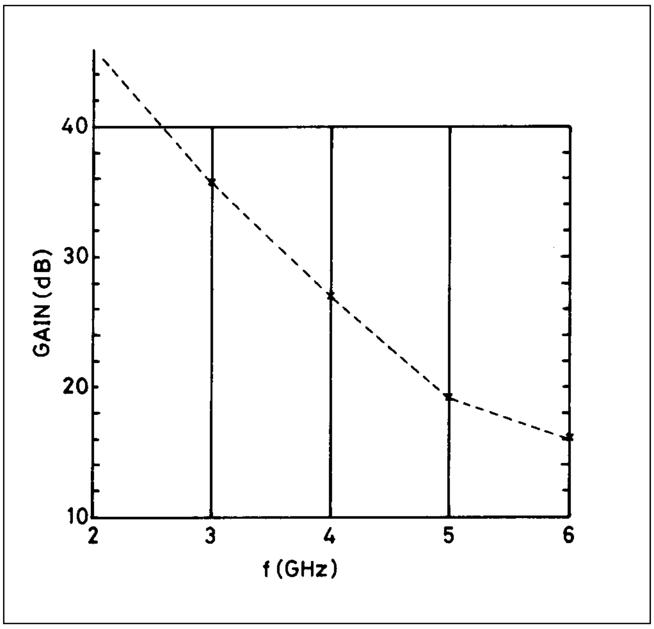


Fig 11.13: The frequency response of the three stage MSA0835 wide band amplifier

f_1 = frequency of signal 1
f_2 = frequency of signal 2

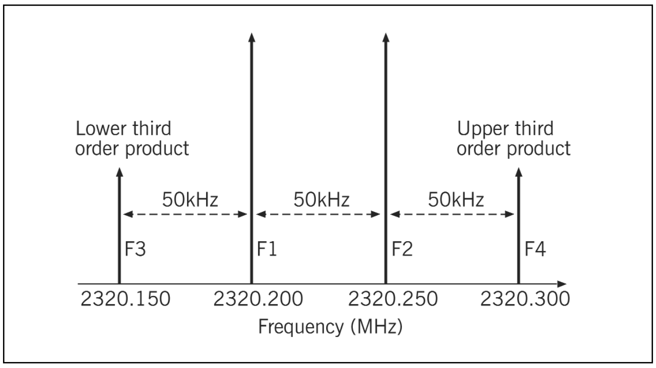


Fig 11.14: Third order products generated from f_1 and f_2

Due to the non-linear nature of the amplifier, a pair of unwanted signals at frequency f_3 and f_4 are generated at the amplifier output. Their frequencies are given by:

$$f_3 = 2 \times f_1 - f_2 \text{ and } f_4 = 2 \times f_2 - f_1$$

For example if

f_1 = 2320.200MHz, f_2 = 2320.250MHz,

then

f_3 = (2 x 2320.200) - 2320.250 = 2320.150MHz

and

f_4 = (2 x 2320.250) - 2320.2 = 2320.300MHz

These two products are symmetrically spaced from f_1 and f_2 by the difference frequency (50kHz) between f_1 and f_2. The significance of these unwanted signals is that they are within the frequency band of interest, and cannot practically be removed by filtering. Because they are third-order products, their levels will increase at three times the rate of the level of the wanted signal. That is, for every 1dB increase in the level of the wanted signal f_1 and f_2, the unwanted third-order products f_3 and f_4 will increase by 3dB. Clearly, if the unwanted products increase at three times the rate at which the wanted signals increase there will be a point at which the levels of $f_1 = f_2 = f_3 = f_4$. In practice, the amplifier will saturate long before this point, so it cannot practically be measured. However, with the knowledge that the unwanted signal levels increase at three times the input rate, it is possible to extrapolate the point at which the unwanted signals, f_3 and f_4, equal f_1 and f_2. This theoretical point is called the intercept point and, when extrapolated at the output, is called the output third-order intercept (OIP3) and if extrapolated at the amplifier input is called the input third-order intercept (IIP3).

It is usually easier to measure the level of the wanted signals f_1 and f_2 and the unwanted products f_3 and f_4 at the output of an amplifier rather than the input. The output intercept point can be extrapolated from these values. The input intercept point is then the output intercept minus the amplifier gain.

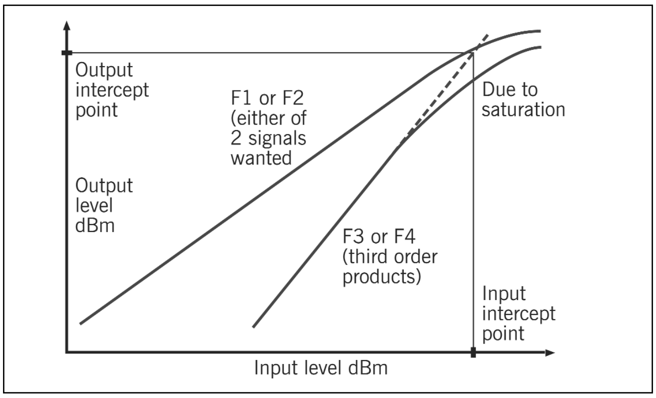


Fig 11.15: Extrapolation of the third order intercept point

The third order intercept can be calculated from:

$$IIP3 = P_{in} + 0.5\,(P_{in} - IM_{3in})$$

A graphical extrapolation of the intercept point usually requires that the output level of f_1 or f_2 is plotted on linear graph paper and then the unwanted product at f_3 or f_4 is plotted on the same graph. The slope of f_3 or f_4 should be three. Extrapolating the f_3 or f_4 lines to the point where either crosses the f_1 or f_2 lines gives the intercept point. This is shown in **Fig 11.15**.

Practically, it can be difficult to determine the actual line to extrapolate, and there may be a few dB of uncertainty in the resulting intercept point determination using this technique. This can be avoided by using a useful piece of software called *Intercept Point* within the AppCAD programme suite, which is available to download from Avago Technologies [6].

The great value of knowing the intercept point is that the level of unwanted products can be determined accurately from the level of the wanted signals and the intercept point. When the two equal-level input signals just produce unwanted products at the amplifier noise floor, the difference in the level between the noise floor and either of the two input tones (in dB) gives the spurious-free dynamic range (SFDR) of the amplifier. This is a good measure of the amplifier (and receiver system) strong-signal performance.

Before moving on to use this information to determine system performance, it is necessary to understand a bit about noise figure (NF) and noise temperature.

Noise figure is the most common way to express the sensitivity of amplifiers and receivers. It is a measure of the excess noise (always) added by a circuit or component, such that at the output of a circuit there is always a worse signal to noise ratio compared with the signal going into the circuit. Noise figure is normally measured and quoted with a source temperature of 290 Kelvin (290K) or 17 degrees C. In determining the performance of a microwave radio system it is often more convenient to work with system noise temperature rather than NF To convert from noise figure to noise temperature the formula is:

$$T = 290 \times (10^{NF/10} - 1)$$

where T= noise temperature in Kelvin, and NF = noise figure in dB. For example, 1dB noise figure = 70K

Conversely, to convert noise temperature to noise figure:

$$NF = 101og\,(1 + (T/290))$$

A terrestrial microwave band antenna would normally be orientated towards the horizon. In this position, the antenna would "see" a complex noise temperature consisting of approximately half cold sky at perhaps 10 - 30K and half warm earth at near 290K. See **Fig 11.16** for a graphical explanation. In practice, the resulting mean noise temperature will be somewhere between 150 and 200K due to a less-than-perfect

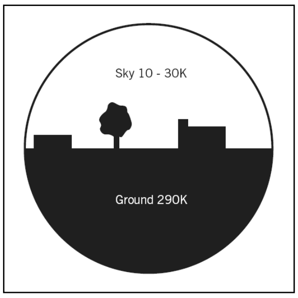


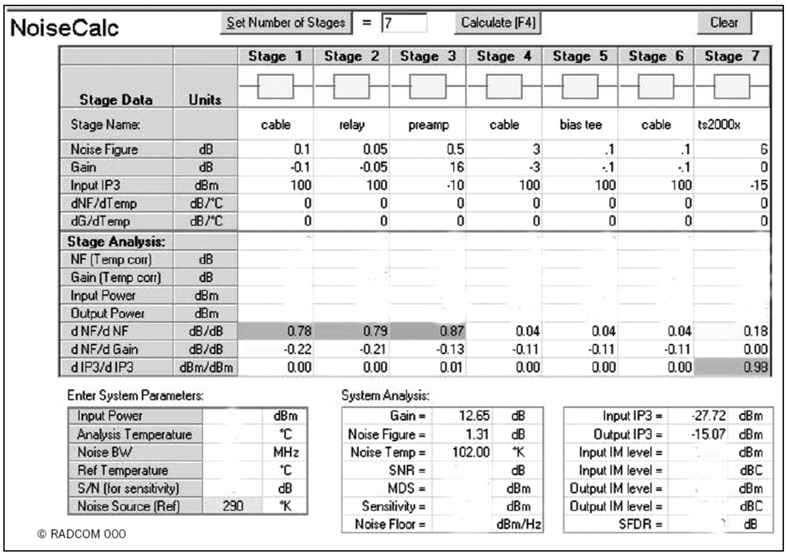
Fig 11.16: Possible antenna noise temperatures as seen by a microwave antenna

horizon line, often consisting of trees and buildings protruding into the cold sky area, and unwanted antenna lobe noise pick-up.

The noise temperature of the complete system, consisting of antenna and receiver noise temperatures limits receiver system sensitivity. The lower T_{sys}, the more sensitive the receiver system.

$$T_{sys}\ (K) = T_{antenna} + T_{receiver}$$

Clearly, if the contribution from the receiver (including feeders etc) can be minimised, the overall sensitivity can be increased. Playing with *AppCAD* will show that attempting to reduce the receiver contribution too far may well result in a very sensitive system, but with poor dynamic range. Compromise is usually necessary.


NoiseCalc

Set Number of Stages = 7 | Calculate [F4] | Clear

Stage Data	Units	Stage 1	Stage 2	Stage 3	Stage 4	Stage 5	Stage 6	Stage 7
Stage Name:		cable	relay	preamp	cable	bias tee	cable	ts2000x
Noise Figure	dB	0.1	0.05	0.5	3	.1	.1	6
Gain	dB	-0.1	-0.05	16	-3	-.1	-.1	0
Input IP3	dBm	100	100	-10	100	100	100	-15
dNF/dTemp	dB/°C	0	0	0	0	0	0	0
dG/dTemp	dB/°C	0	0	0	0	0	0	0
Stage Analysis:								
NF (Temp corr)	dB							
Gain (Temp corr)	dB							
Input Power	dBm							
Output Power	dBm							
d NF/d NF	dB/dB	0.78	0.79	0.87	0.04	0.04	0.04	0.18
d NF/d Gain	dB/dB	-0.22	-0.21	-0.13	-0.11	-0.11	-0.11	0.00
d IP3/d IP3	dBm/dBm	0.00	0.00	0.01	0.00	0.00	0.00	0.98

Enter System Parameters:

Input Power		dBm
Analysis Temperature		°C
Noise BW		MHz
Ref Temperature		°C
S/N (for sensitivity)		dB
Noise Source (Ref)	290	°K

System Analysis:

Gain =	12.65	dB	Input IP3 =	-27.72	dBm
Noise Figure =	1.31	dB	Output IP3 =	-15.07	dBm
Noise Temp =	102.00	°K	Input IM level =		dBm
SNR =		dB	Input IM level =		dBC
MDS =		dBm	Output IM level =		dBm
Sensitivity =		dBm	Output IM level =		dBC
Noise Floor =		dBm/Hz	SFDR =		dB

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Fig 11.17: AppCAD worksheet for the 1.3GHz example receiver system. Unused numbers have been removed for clarity

Using *AppCAD*

- Download the executable file AppCAD-302.exe from the website and install it on your PC.
- Open AppCAD. Select Signals-systems from the menu on the left-hand side of the window.
- Select NoiseCalc from the menu on the left hand side of the window.
- The open page displays an example worksheet for a CDMA handset receiver.
- Select and press Clear (next to Main Menu [F8]). The graphical boxes will clear of information. Don't worry; the graphics will re-appear next time you open AppCAD. You are now ready to begin entering data.

The following paragraphs will work through an example of a 1296MHz receive system using a masthead preamplifier and a Kenwood TS-2000X, but it could equally be any equipment and band of choice.

Under Options, at the top of the window, select Automatically Calculate after Edit and Intercept Point specified at Input.

It is necessary to set the number of stages you need. This should include one each for antenna cable, transmit/receive relay, preamplifier and receive feeder, bias-T (if used), second connection cable (if used) and radio. That is seven stages. It is better to have too many than too few. Type "7" to the Set Number of Stages box and press the similarly marked button. The worksheet will show seven stages. You cannot enter any graphics into these boxes now, so use the Stage Name worksheet cells to enter "ant cable", "relay" etc. If you are using, say, four antennas, enter the antenna cable to combiner cable once including the excess loss for the combiner. Don't forget the combiner to preamplifier cable loss, in this configuration. This would now be eight boxes.

Starting with the antenna to preamplifier cable, enter its loss in negative dB in the Gain cell. Enter its noise figure as the loss, eg if the loss of this cable is 0.1dB, enter -0.1 as the gain and +0.1 as the noise figure. Do the same for the relay and following stages. You only need enter the noise figure for the final (TS-2000X) stage. Leave the gain as "0". You will note that we have not entered the Input IP3 yet. These should be left as "100" for now.

The program has been calculating the system noise figure as you have been entering the data. Look at the System Analysis box at the bottom of the window. You will see that the cumulative gain is shown as 12.65dB up to the TS-2000X input, whilst the system NF is 1.31dB. Under NF, the Noise Temp cell shows the equivalent noise temperature of 102K.

Now go back and add the Input IP3 data as shown in **Fig 11.17** The passive components are entered as "100" as it is assumed they contribute little intermodulation. The preamplifier is entered as -10dBm, but obviously you would enter the data for your particular preamplifier. The average is about -10dBm for a small-signal HEMT device. The TS-2000X is entered as -15dBm from the ARRL measurements on this radio.

In the System Analysis boxes, look at the box on the lower right. The Input IP3 is shown as -27.72dBm. This is the Third Order Input Intercept (IIP3) for the complete receiver system. It is not particularly good. Below the Input IP3 is the Output IP3. This is the figure at the input to the TS-2000X and is shown as 15.07dBm. This is about the same as the TS-2000X on its own. Try increasing the preamplifier Input IP3 to -3dBm. The system Input IP3 doesn't improve very much so there is little to be gained by fitting this higher IIP3 preamplifier to the system as shown. A much higher preamplifier IIP3 is needed to have much impact. The TS-2000X is the limitation in this system.

You will notice that the Stage Analysis box has the stage 7 (TS-2000X) cell highlighted in red and showing 0.98. This is a sensitivity analysis where the closer the number is to "1" the weaker the stage is in terms of the parameter shown. In this case, d(IP3)/d(IP3) which is the sensitivity of the system IIP3 to the stage IIP3, is shown as 0.98 for the rig. The lower the better.

The d(NF)/d(NF) line shows some numbers in blue. Here, the sensitivity of the system NF to the NF of the stage is shown. The loss of the relay and cable are shown to be sensitive but less so than the preamplifier (0.87).

You will learn a lot by playing with the numbers and hopefully gain a better understanding of how your system works and to what you should pay most attention and what to leave until later. The worksheet shows lots of other numbers to aid system analysis. As a guide, for the middle microwave bands including 1.3GHz and probably down to 430MHz, a good target for a terrestrial system noise figure is about 1dB, as the antenna noise temperature is about 170K (2dB NF) for a good terrestrial antenna in the clear. Try entering 19dB as the gain for the preamplifier in the worksheet, holding the NF at 0.5dB. Notice the system IIP3 reduces to around -30dBm. A 1dB system NF means that antenna noise will be the main contributor to system sensitivity. If you strive for much better than about 1dB, it may result in poor system dynamic range, but this is very dependent on other system parameters.

Ideally, the IIP3 for the system should be 0dBm or better. An acceptable figure is -10dBm, while -27dBm probably means some strong-signal problems will be experienced.

Receiver system design usually involves a trade-off between strong-signal performance and noise figure. AppCAD (and other system analysis programs such as TCALC) are essential tools in analysing and designing better systems. AppCAD does not deal with parameters such as Blocking Dynamic Range (BDR) or the effects of local oscillator phase noise on reciprocal mixing.

GETTING STARTED ON 3CM

Brian Coleman, G4NNS, and Ian Lamb, G8KQW, describe how to become operational on the 10GHz band with minimum effort.

10GHz is the most popular of the higher amateur radio microwave bands and provides abundant opportunity for challenging operation. The current UK (terrestrial) distance record of 1347km was made between two home stations. This might be a surprise for those who don't know the band or who perhaps thought DX could only be worked between mountaintops. There are, of course, many opportunities for those who want to work from portable locations and the compact size and lightweight of the system described here makes this not only possible but also very enjoyable and rewarding. This section describes how to become operational on the 10GHz band with minimum effort, but with a system that provides a sound basis for future development. Even in its basic form hilltop contacts to 300 or 400km should be possible, and even from the average home station location such distances are possible during enhanced propagation conditions that include rain scatter and thermal inversions.

The system consists of a DB6NT transverter and a suitable relay with associated drive circuit, all housed in a weatherproof plastic enclosure. The transverter is small and light enough to be mounted, with the appropriate feed horn, on the support arm of a small, surplus, TVRO (Receive Only) offset dish. The feed horn is mounted in place of the original LNB (**Fig 11.18**). Thus the need for waveguide (plumbing) is minimised.

While 200mW output might sound like low power (QRP), a small TVRO dish will provide about 30dBi of gain and thus achieve an Effective isotropic Radiated Power (EiRP) of about 200W.

The Transverter

Kuhne Electronic (DB6NT) offers the modern MKU 10 G2, high performance, 10GHz transverter, either as a ready-built module or as a kit. Whilst the kit is easy to build, this route should only

Fig 11.18: Quickstarter for 3cm mounted on the feed arm of a surplus TVRO offset dish

Fig 11.19: A typical SMA relay suitable for use at 10GHz

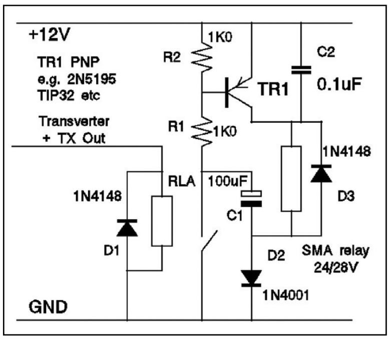


Fig 11.20: A circuit to drive a 24V or 28V relay from a 12V supply

be taken by experienced home constructors who are well versed in working with surface mount components. To help you decide if this approach is for you, you can download, free, the manual from the Kuhne Electronic web site [9]. While the kit comes with adequate instructions, John Hazell, G8ACE, has written some additional notes, a link to which can be found in the hardware section of the UK Microwave Group web site [10].

Any 144MHz band transceiver that can provide between 200mW and 2W output, can be used to drive the transverter. A multimode transceiver is preferred as most communications on 10GHz takes place using either SSB or CW. A transceiver used for this purpose is known as an Intermediate Frequency (IF) rig. The actual transceiver output power is not too critical as an adjustable attenuator inside the transverter allows the user to set full 10GHz output for any input in this range. The small, battery powered, Yaesu FT290 or FT817 are commonly used as if transceivers by microwave operators. The transverter has separate connectors (SMA) for RF in, RF out and IF.

The Changeover Relay

Whilst there is no requirement for an rf relay between the transceiver and the transverter, an external coaxial change over relay is required between the transverter and the antenna.

Coaxial relays that have reasonably low loss (<1dB) and sufficient port to port isolation at 10GHz may sound expensive. They are, in fact, quite common and appear frequently at mobile radio rallies and on eBay. Prices are normally in the range £10 to £25 depending on type and condition. Make sure that the relay you buy is useable at 10GHz before you part with your money. These relays invariably have SMA connectors.

An example is shown in **Fig 11.19** but they come in a variety of shapes, sizes and colours and most commonly have a 28V coil. These will operate satisfactorily down to about 20V, but the simple circuit shown in **Fig 11.20** can be used to operate them from a 12V supply. If, at a later date, you add a power amplifier you will need to think about sequencing the relay to avoid damage to the PA or transverter front end. You should check the relay for correct operation at DC before connecting it to the transverter. C1 may need some adjustment. Too small a value and the relay won't close, too large a value and it has an excessive hang time when switching from transmit to receive the feed horn.

Coaxial cable is lossy at 10GHz so its use is kept to an absolute minimum. Waveguide is much less lossy but can put some constructors off. In the Quickstarter, the transverter is purposely located very close to the dish feed point to minimise the feeder length. The feed system consists of two components, combined into one assembly. The components are a coax to waveguide transition and a feed horn designed to provide efficient illumination of the dish. An SMA to waveguide transition is

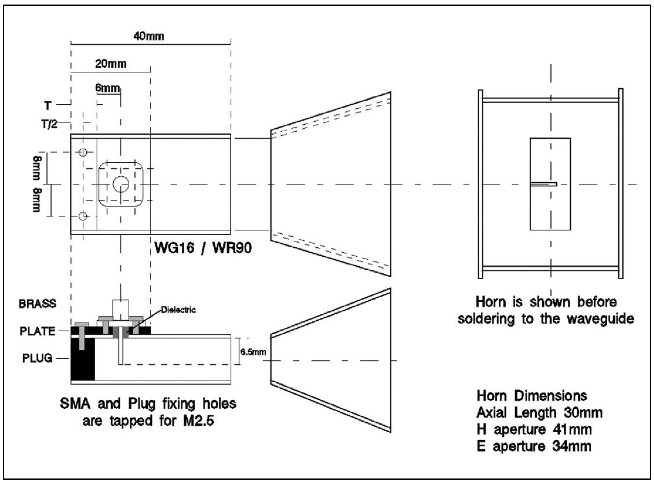


Fig 11.21: SMA to waveguide transition and feed horn

shown in Fig 11.21. These are easier to make than you might think, but for those unwilling to try, the complete transition can be supplied by, for example, the UK Microwave Group or a number of commercial suppliers. The critical dimensions are the centring of the SMA connector, the length of probe within the waveguide and the distance from the probe to the waveguide plug (short circuit) to the rear. If you have access to the necessary test equipment the probe can be made over sized and trimmed down for minimum return loss (best SWR). If you follow the dimensions shown you will have a satisfactory transition. Construction details and a template for making the 10GHz feed horn from copper laminate can be found on the G4NNS webpage [11].

Metalwork

Many of us find metalwork a challenge so the metalwork for this project has been kept very simple. It should be possible to fashion the various parts from surplus aluminium sheet of 1.0 to 1.5mm thickness, using basic tools such as a hacksaw, drill and vice.

Essentially there are six basic components: A pair of right angle brackets to clamp the horn in place (Fig 11.22), a pair of right angle brackets to clamp the unit onto the feed support arm of the dish (Fig 11.23), and two plates to clamp the transverter in place (Fig 11.24). The pairs of clamps are folded in opposite directions to form mirror images of each other. The plastic enclosure is available from Maplin [12], type MB6, Code YN39. For portable operation the enclosure needs only a few drilled holes. For a more permanent installation, it should be painted with a reflective paint such as

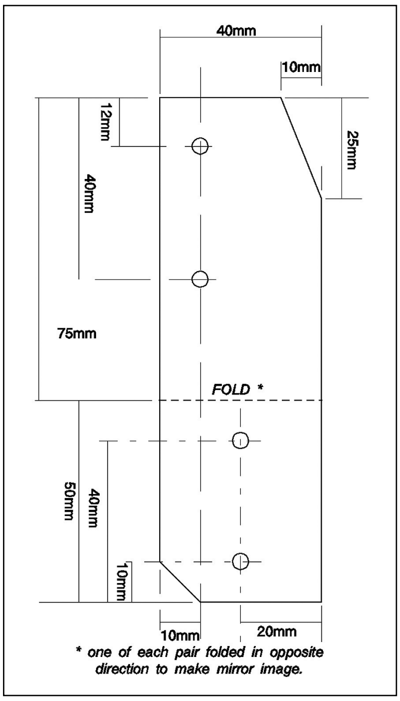


Fig 11.22: Feed horn bracket

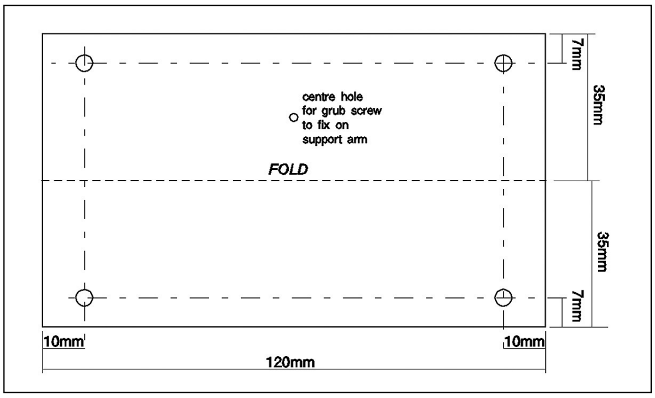


Fig 11.23: Support arm bracket

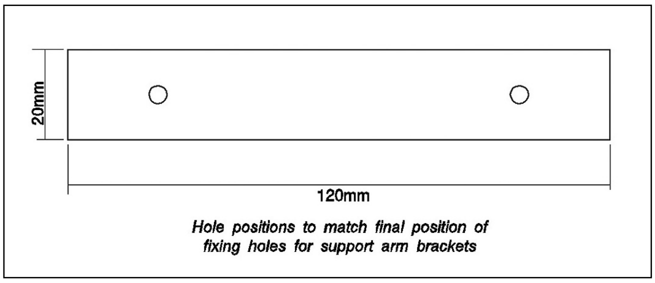


Fig 11.24: Clamping plates for the transverter

silver, and thoroughly sealed against rain ingress. Some people prefer to add a small drain / breather hole at the lowest point of the enclosure.

I have not shown the hole positions as they will vary according to the dish used and its feed support arm in particular. What is important is that the centre line of the feed support arm and the feed horn are aligned. The brackets allow adjustment of the whole assembly on the feed support arm and alignment of the feed horn to place it at the focus and pointing at the centre of the dish.

The Antenna

Surplus TVRO dishes often appear at mobile rallies and can often be found as scrap. It is important that the profile is in good condition and that the feed support arm is included. If it still has the Low Noise Block (LNB) converter, or its fixing clamp attached, so much the better as this will give a clear idea of where the focus is located. If this is unknown it can be calculated using the W1GHZ program *HDL_Ant* [13]. Use the "Offset dish calculation" option, enter the frequency (10368MHz), the large "diameter" of the dish, the small "diameter" of the dish, the depth at the deepest point and the distance of the deepest point from what is described as the "bottom" edge of the dish. As we will be using the dish mounted on its side, this point is perhaps best described as the edge nearest the feed arm and I will refer to it as the "feed point edge". The software will then provide the location of the focus in terms of distance from the feed point edge and the opposite edge of the dish. It also provides a calculated gain.

For terrestrial operation I prefer to use offset dishes mounted on their side as this makes mounting them on a vertical mast much simpler and avoids the critical adjustment of elevation that is otherwise needed. The exact method of mounting depends upon the construction of the dish. Usually there is a bracket at the rear of the dish that fixes the feed support arm and the mounting brackets. The screws fixing this to the dish can be replaced by studding (usually M6) and a plate or angle

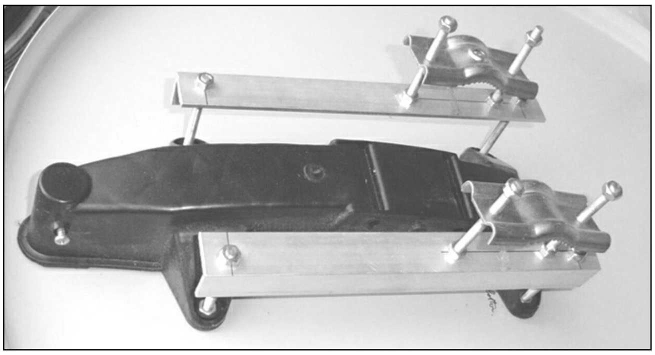


Fig 11.25: Mounting arrangement for a surplus TVRO dish using angle sections

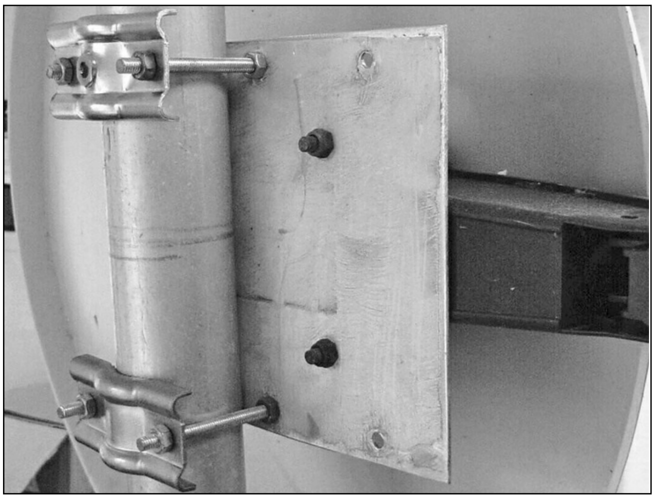


Fig 11.26: Mounting arrangement for a surplus TVRO dish using a plate

sections fixed to the rear see Figs 11.25 and 11.26. The plate or angle sections then have holes drilled for standard mast clamps. These holes should be positioned to give the best balance point for the dish, transverter and cables.

The direction in which the dish points in terms of azimuth, when mounted on its side, may appear a bit of a mystery but in fact is usually described by the line from the feed point edge of the dish to the centre of the horn, see Fig 11.27. This is because offset dishes are usually sections of a normal parabola that includes the centre at the feed point edge. Calibration can be confirmed by listening to a signal on a known bearing and unlike elevation this can be easily compensated using a rotator or by hand. With the offset dish mounted on its side like this it is also easy to add an additional feed horn for another band (such as 5.7GHz), but both feed horns cannot occupy the same place so there will be some offset in azimuth between the two (or more) bands.

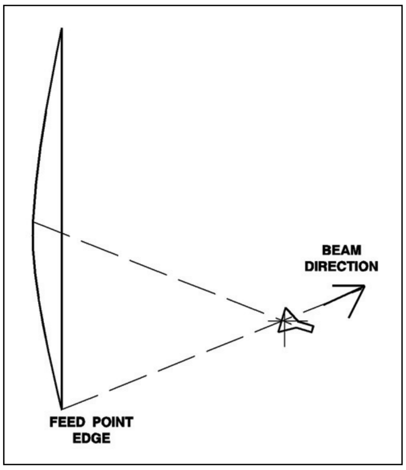


Fig 11.27: Dish beam pattern geometry

Antenna Support

The arrangement described is ideal for fixing the dish to a mast either for home station or portable operation. For portable operation a tripod, if a sufficiently sturdy one is available, is useful. Many potential sites have hedges or small trees that cannot be cleared with a tripod so a short mast is a good choice. With a 144MHz talk back antenna at the top and the dish mounted below the guys even a small mast can provide good support for a dish at a greater height than with a tripod. Some means of locking the azimuth will be necessary.

Coaxial Connections

As mentioned previously any coaxial cable will be lossy at these frequencies, as are most connectors. The preferred cable is RG402 (also available as UT141) semi rigid or the more flexible and easier to use equivalents such as Sucoflex and Quickform (available from Farnell [14] stock code 157 995). All these cables have an outside diameter (OD) of about 3.5mm and are used with direct solder SMA connectors such as Farnell 105-6352. These cables frequently appear at mobile rallies with the connectors already fitted. They are unlikely to be the right length but can often be cut so that only one additional connector is needed. Avoid right angle connectors as these can present a mismatch and can be very lossy. Although few of us have the correct tools to work with these cables, they can be worked quite satisfactorily with standard hand tools. Cut the cable to length using a junior hacksaw. Remove the outer conductor, where necessary, by heavily scoring it all the way around with a sharp knife. Flex the cable holding the piece of outer you wish to remove, using pliers so that it breaks away quite easily. Then cut the inner conductor and insulation in accordance with the instructions for the connector you are using. The total amount of coaxial cable you need for this project is 150mm or less.

IF Transceiver

The most popular choices of if transceiver are the FT290 Mk1 and the FT817 but other multi mode transceivers that can provide a suitably low level output can be used.

The FT290 Mk1 can be found second hand for around £100. It provides a DC output on the antenna line when in transmit mode and this is used to set the transverter to transmit.

It is also possible to do this by grounding the PTT line but if the DC on the IF line is used an additional connection is not needed. Without internal modification this is slightly harder to do for the FT817. However the circuit shown in Fig 11.28 can be used. It has the slight disadvantage that if the FT817 is switched off

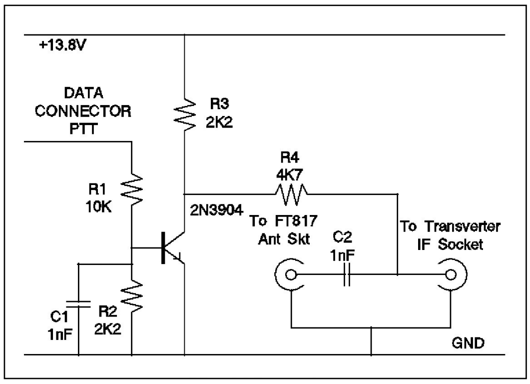


Fig 11.28: A Bias T for the FT817

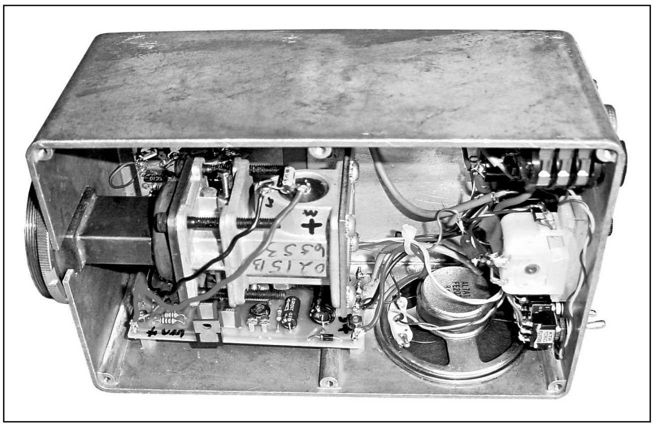


Fig 11.29: A typical Solfan 10GHz transceiver

while power is supplied to the transverter it will be switched to the transmit state. This is not likely to result in damage and as most people will turn off power to the transverter when the IF transceiver is off this is not likely to be a problem. The miniature DIN DATA plug is the same as on some old computer mice so you probably don't need to buy one! Whichever transceiver you choose, set up your transverter to be driven with the minimum power.

Conclusion

The system described above will provide a quick and easy means of getting going on 10GHz and although not the cheapest route will provide good results and a sound basis for further development. With the growing popularity of microwave operation it is likely to retain its value. If you need help with the project contact a member of the UK Microwave group Technical Support Team [10]. See you on 10GHz soon I hope.

10GHZ WIDEBAND ACTIVITY REVIVAL

Sam Jewell, G4DDK noticed in 2006 that the UK Microwave Group [10] was receiving a lot of correspondence indicating renewed interest in wideband frequency modulation (WBFM) equipment and operation within the 10GHz amateur band allocation. During the 1960s and 1970s, WBFM was virtually the only way to become active on the band, with klystrons (723A/B and similar) providing an easy way to generate a few tens of milliwatts just inside the lower edge of the band.

Perhaps more popular, from the mid 1970s, were Gunn diode oscillators. These used surplus, low-power, low-voltage, Gunn effect semiconductor devices to generate 10mW or so at 10GHz. Early Gunn oscillators were often home-constructed using waveguide-based cavities and, as I recollect, were not particularly stable and were often highly temperamental.

The introduction of the commercial Gunnplexer enabled more stable transceivers to be assembled although, for many of us, the Gunnplexer was just too expensive and we sought an alternative. This turned out to be the commercial Doppler intruder alarm module, the best known was the Solfan unit. These were (and still are) also available with other markings, such as C&K, but are otherwise identical. The Solfan unit could easily be retuned from its normal operating frequency of around 10.6GHz down to typically 10.2 to 10.3GHz. The Solfan unit contained not only the Gunn oscillator but also a mixer diode as part of the Doppler receiver.

Solfan head Gunn transceivers use a single oscillator for the transmitter and the receiver. This means that a pair of these transceivers needs to operate with an agreed oscillator frequency offset such as 10.7, 30 or 100MHz, this being the intermediate frequency or IF. Transceivers with different IFs are still able to communicate, but one operator has to change frequency between transmit and receive in order to keep signals tuned-in. This is a rather hit-or-miss affair at the best of times and tends to lead to more complex solutions using separate transmitter and receiver Gunn oscillators. The almost unique feature of the common IF, single -source transceiver is that full-duplex (simultaneous transmit and receive) operation is possible. **Fig 11.29** shows a typical Solfan 10GHz transceiver, using an old Ambit 10.7MHz IF strip.

Fig 11.30 shows a pair of full-duplex Gunn transceivers together with an example of some typical full-duplex frequencies. When used with a wideband FM receiver (such as a scanner or broadcast FM receiver) as the IF, a simple but effective wideband 10GHz transceiver can still be put together for a few tens of pounds. Solfan heads are still regularly seen on the surplus market and a few are still in amateur use in 10GHz transceivers.

The growth in popularity in the late 1980s of stable, narrowband, systems based on GaAsFET amplifiers, mixers and multipliers, used in 144MHz to 10GHz transverters, meant that system performance improved dramatically. It became possible to work from home station to home station at distances of over 200km under normal conditions and over 1000km under enhanced propagation conditions.

Faced with the reality of what could be achieved on the higher bands with more advanced equipment, wideband equipment operation has fallen away in recent years, except for the dedicated few who enjoyed the challenge of working long distances with modest equipment. Whilst this is not necessarily a bad thing from the aspect of equipment development and having sufficient performance to investigate new propagation phenomena, it does mean that many prospective newcomers to the microwave bands no longer go through the worthwhile learning process that simple and low-cost equipment can provide. I wonder how many potential newcomers have been put off by the relatively high cost of the modern 10GHz transverter and therefore never give microwaves a go?

Sams own interest in simple wideband FM was again sparked when he read the microwave group messages and later was given a pair of Solfan Doppler intruder alarm heads.

Andrew, MOBXT, had measured the sensitivity of the

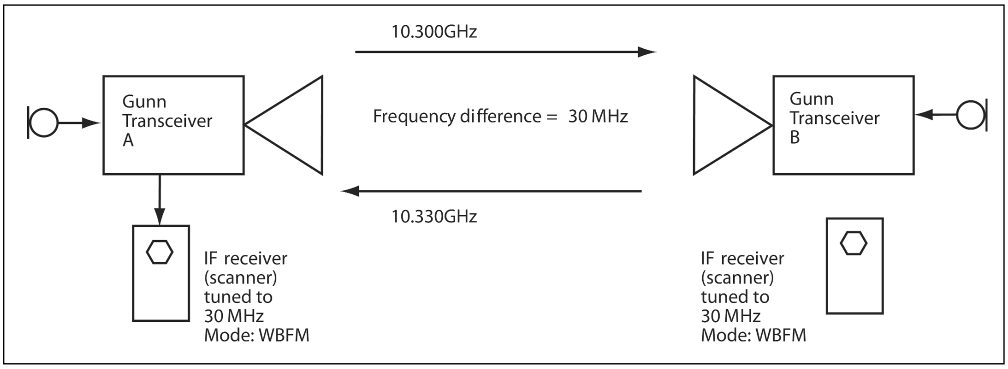


Fig 11.30: A simple Gunn device based duplex transceiver pair with 30MHz IF

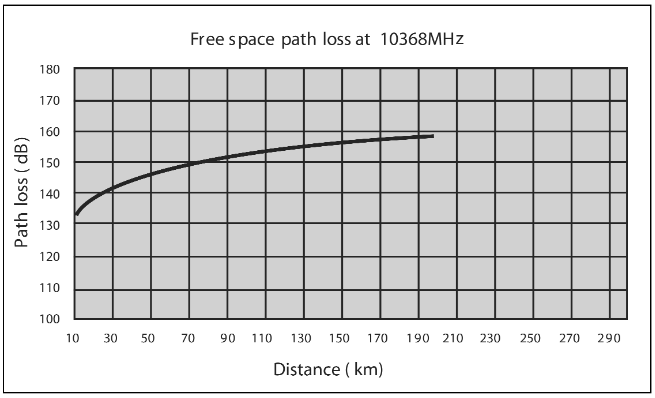


Fig 11.31: A plot of free space path loss at 10.368GHz

Solfan-based transceiver shown in Fig 11.29. It was found possible to detect a usable signal (tone modulated FM) down to about -87dBm. A typical WBFM scanner (Fair Mate HP100) was used as a WBFM IF, has a measured sensitivity of about -100dBm at both 10.7MHz and 30MHz. The conversion loss of a typical Solfan head, when optimally adjusted for mixer diode current, is about 13dB, giving a usable overall sensitivity of about -87dBm.

Putting this into perspective; a 10mW (+10dBm) output Gunn oscillator transmitter, Solfan based receiver, directly connected to dishes with +30dBi gain (typical of measured 60cm diameter dishes with penny feeds) at each end the path, will have a path loss capability (plc) given by:

plc(dB) = eirp (dBm) - ers (dBm),

where:
Transmit effective isotropic radiated power (eirp) = +10dB + 30dB = +40dB
Effective receive sensitivity (ers) = -87dB - (+30dB) = -117dB
Path loss capability (plc) = +40 - (-117) = 157dB
Free-space loss is given by Path loss (dB) = 32.45 + 201og(f) + 201og (d),
Where f is in MHz and d is in km.

Transposing for distance, a free-space path loss of 157dB gives a working distance of approximately 150km as can be seen in **Fig 11.31** This is also the distance that must be worked on the 10GHz band in order to achieve the coveted RSGB 10GHz minimum distance award. This simple equipment must be working well to achieve this award.

10GHz WBFM system performance can be improved quite dramatically by adding GaAs FET preamplifiers or moving from the use of Gunn oscillators to more stable Dielectric Resonant Oscillator (DRO) sources as used in satellite TV Low-Noise Block (LNB) converters.

An LNB based receive converter can be stable enough to receive quasi-narrowband FM modulation (20 - 30kHz) instead of the typical 50 - 100kHz deviation used with Gunn-only systems. The reduction in bandwidth in the receiver will give a small but worthwhile receive system improvement while the low noise figure (<1dB) of a modern LNB front-end, even at 10GHz, would further improve the sensitivity by 12dB or so. Of course we are starting to get more complex than is possible with the simple Gunn transceiver, and it then becomes a question of where to put the work in order to get the best return for that effort.

The Boomerang

A simple but effective way to test a Solfan or similar transceiver is called the boomerang and requires that, in addition to the transceiver to be tested, you need a signal generator or simple crystal oscillator at the wanted IF frequency and a suitable waveguide mixer. It is possible to use the mixer part of a second Solfan waveguide head.

The signal generator is set to the receiver IF frequency and connected to the boomerang waveguide mixer. A signal generator output of between 1 and 10mW is required at IF. If the Gunn transceiver has a built-in modulation source, the signal generator need not be modulated, otherwise set the modulation to FM and about 50kHz deviation at, say, 1kHz modulating frequency. With the Gunn transceiver aimed at the mixer, the modulated IF signal applied to the mixer will be heard in the transceiver receiver.

If the transceiver is self modulated with a tone, this should be switched on and will be heard when the mixer and transceiver antenna are lined up. A good transceiver will detect a good mixer at up to several hundred metres, although the exact distance will depend on several factors including the gain of the antenna on the transceiver and if any antenna is used on the mixer waveguide.

The boomerang works by the radiated RF from the transmitter mixing in the boomerang mixer diode with the locally generated IF signal and then being reradiated by the mixer diode. If the transceiver uses a 30MHz IF and the signal generator is set to 30MHz, then the reradiated 10,300MHz signal will contain 30MHz sidebands and these will appear at 10,270 and 10,330MHz as a received signal at the transceiver receive frequency . See **Fig 11.32** for a diagrammatic explanation.

Care must be taken to ensure that the IF receiver does not receive the signal generator by direct pick-up. This is most easily arranged by aiming either the mixer or transceiver antenna away from the other unit and checking that the received signal disappears.

A timely reminder of the temperamental nature of the Solfan (and other) Gunn transceivers was experienced when measuring the sensitivity of the 10GHz transceiver. As the Gunn oscillator voltage was varied, in order to fine-tune the Gunn frequency, numerous spurious signals were heard. Investigation with a spectrum analyser showed these to be due to unwanted moding or spurious oscillations, in the Gunn at certain critical tuning voltages. Simple Gunn transceivers are not always problem-free, but can offer a simple route to 10GHz operation.

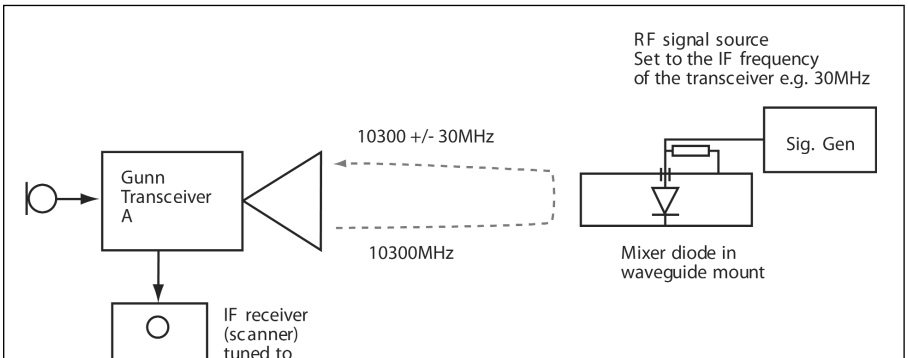


Fig 11.32: A simple "Boomerang" system for testing a full duplex transceiver

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 11.33**). 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.

1.0 - 1.3GHz High Quality Microwave Source

Sam Jewell's design for the DDK001 L band source was first published in the *RadCom* ‘Microwaves’ column in 1987 and in volume two of the *RSGB Microwave Handbook*. Supplies of the PCBs dried up many years ago after a key component became unavailable, effectively making the design obsolete.

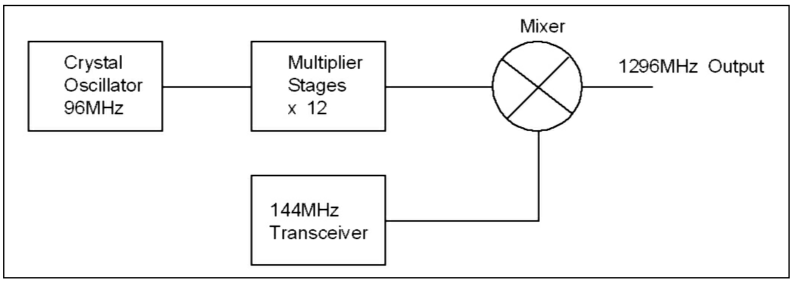


Fig 11.33: Local oscillator mixed with transceiver to give microwave output

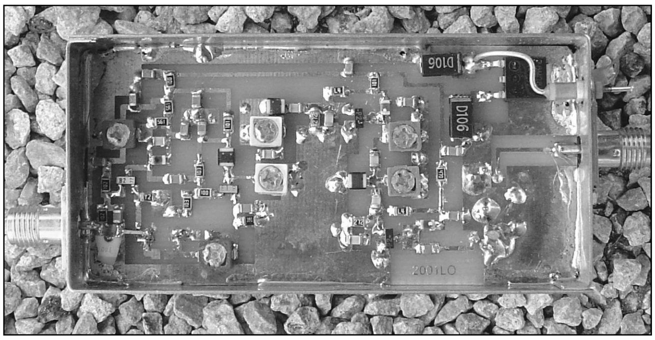


Fig 11.34: One of the prototype microwave signal sources. This version uses different trimmer capacitors to those specified. The trimmers shown are red and now replaced by black bodied 4 to 25pf trimmers

Ongoing demand for sources for use as local oscillators, test sources and small 1.3GHz transmitters encouraged Sam to update the original design. The new, compact, version 2 DDK001 local oscillator source (2001 LO) covers the frequency range from about 1150MHz to 1305MHz, determined by the available Toko helical filters used in the final multiplier stage. Output level is typically +7dBm at 1152MHz and the output spectrum is cleaner than the original 001 source. The following description assumes an output frequency of 1152MHz (LO frequency for 2m IF at 1296MHz, among other applications).

Although the new 2001 LO follows the same architecture as the older 001 design, there are several significant changes. The side coupled stripline filters have been replaced by discrete component lumped element band pass filters in the first two positions and by a Toko 2-pole helical filter in the output stage. This final filter determines the frequency range of the source. An external high stability source can be connected in place of the internal crystal oscillator if required.

The two-stage Butler bipolar transistor overtone crystal oscillator has been used again. This design has been endlessly analysed with respect to its phase noise performance and stability. Whilst there can be little doubt that the Driscoll and some other Butler variants can produce better phase noise performance, it is still a versatile and low noise oscillator design that is easy to align, forgiving of component variation and reliable in operation. The oscillator has proven itself over the years and even when multiplied up to 10GHz the phase noise performance is still good. Butler oscillators have been operating in the GB3MHL and GB3MHX (J002PB) beacons for over 20 years without failure. A finished 2001 LO source is shown in **Fig 11.34**.

Circuit description

An overtone crystal oscillator drives a series of frequency multiplier stages with inter-stage filtering to define the output frequency. The overall multiplication is 12, implemented as x3 in the Butler oscillator stage and then two frequency doubler stages. The circuit is shown in **Fig 11.35**.

Common base amplifier, TR1, is the oscillator maintaining stage. Its collector tank circuit is tuned to the crystal overtone frequency (96MHz). The tuned circuit is heavily damped by R6. Trimmer C5 resonates the circuit. This stage is also used as the input buffer amplifier when an external source is to be connected.

TR2 is an emitter follower with its output feeding the overtone crystal. Its collector is tuned to 288MHz, the third harmonic of the crystal overtone

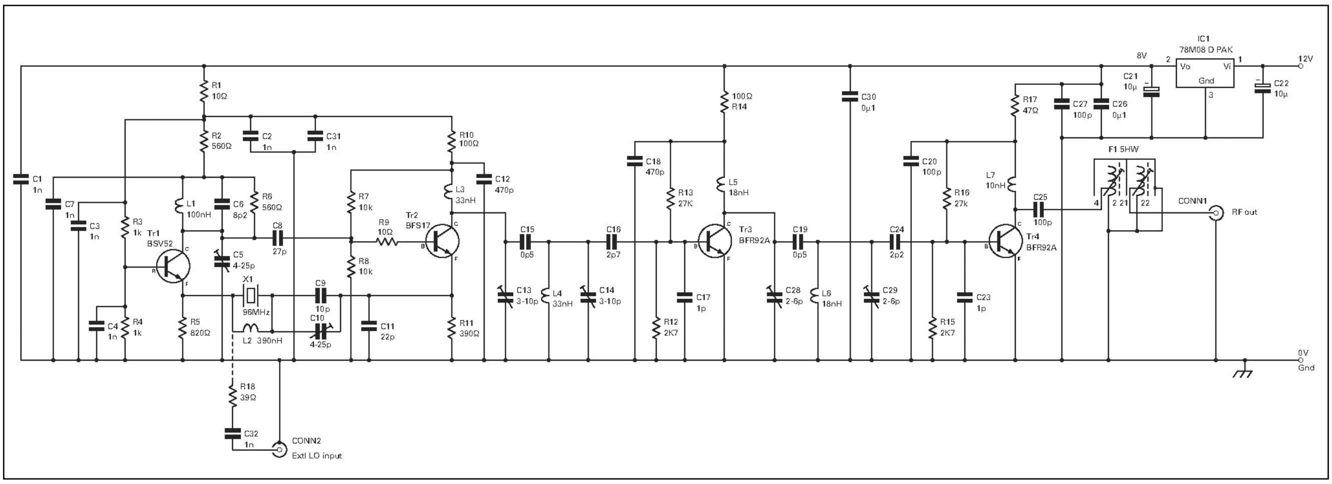


Fig 11.35: Circuit diagram of the 20011.0 microwave source

frequency. Soft limiting is used to reduce phase noise degradation, but this results in low harmonic output levels. A compromise has had to be made here.

Since the crystal is connected between the emitters of TR1 and TR2, the feedback is in phase, which allows oscillation. Capacitors C9 and C10 allow the crystal to be pulled over a few hundred Hz at 96MHz. L2 may not be necessary but can be used if required in order to allow the crystal to be pulled onto frequency. In practice, C10 should be used to set the oscillator frequency and C5 used to peak the output, although there will be strong interaction between the settings of these capacitors. The quality of the crystal oscillator is largely determined by the quality of the crystal used. Cheap crystals may not prove to be so economical in terms of performance.

L3, C13, C15, L4 and C14 form a 288MHz band pass filter. This is sharply tuned and only the recommended trimmer capacitors should be used to ensure resonance is achieved at 288MHz. Top capacitive coupling is provided by C15.

TR3 is the frequency doubler from 288MHz to 576MHz with L5, C28, C19, L8 and C29 forming the band pass filter at this frequency. Bias stabilisation is provided by returning R13 to the collector side of R14.

TR4 is the frequency doubler from 576 to 1152MHz and uses a similar bias arrangement to that of TR3. A Toko 5HW series helical filter selects the second harmonic of the 576MHz drive. The PCB connections are for the Toko F pin-out configuration filters (5HW 115045A-1195, 5HW 120050F-1225 and 5HW 125055F-1305). A 78M08 surface mount voltage regulator provides a stabilised 8V to drive the source.

No attempt has been made to temperature stabilise the crystal oscillator as it is felt that where this is required, an external, low noise, stabilised crystal oscillator (OCXO), direct frequency synthesiser (DFS) or PLL could be used. However, the small housing used for the 2001 LO source means that it is practical to enclose the entire unit within a temperature stabilised oven to control the whole of the unit and not just the crystal.

Where an external source is to be used, remove crystal X1 and add R18 and C32 as shown in Fig 11.35. Since the input impedance of the common base stage is low at about 7 ohms it is necessary to increase the impedance, to achieve a decent match, by adding series 39 ohm resistor, R18. The required drive is in the range -10 to +6dBm, with optimum drive at about -3dBm.

With the component values shown in Fig 11.35 the oscillator will operate satisfactorily between approximately 95 and 109MHz, giving an output of between 1150 and 1305MHz respectively. One of the three versions of the helical filter will be required to cover the whole frequency range.

Construction

The source is constructed on a 37 x 74 x 1.6mm, FR4, double sided PCB, shown in **Fig 11.36**. All the components used are surface mount devices (SMD) with the exception of the crystal and the helical filter. 0805 size passive components are used, with SOT23 packaged transistors, D-Pak voltage regulator and B case size tantalum polarised decoupling capacitors. The trimmer capacitors are Murata TZB4 series.

The source will fit into a Schubert 37 x 74 x 30mm size tin plate box, or it can be fitted into convenient die-cast aluminium or other housing as required. The Schubert tin plate box is available from G3NYK [15]. If the tin plate box is used then it should be marked with a line around the inside 16.5mm down from the rim. This marks the position of the ground plane (copper side) of the PCB, with the component side of the board now 11.3mm below the rim of the box. The area above the PCB ground plane side is sufficient to clear the helical filter and crystal, which are both mounted on the ground plane side of the PCB.

Next, mark the position where the two or 4 hole gold plated SMA RF output connector will be soldered to the outside of the box. Drill a 4mm diameter hole through the box in a position where the connector spill will lie flat to the PCB RF output track. If an external source is to be used, an SMA connector can be fitted at the other end of the box, near the crystal location.

Drill a hole in the end of the box for the supply feedthrough capacitor. This should be above the voltage regulator and on the short edge of the box near the RF output SMA connector.

It is preferable to solder the PCB into the box before soldering the SMD parts into place. This avoids accidental damage to these small and sometimes fragile parts during the seam soldering process.

If you are making your own PCB for the source, it will be necessary to use thin wire through-board links to

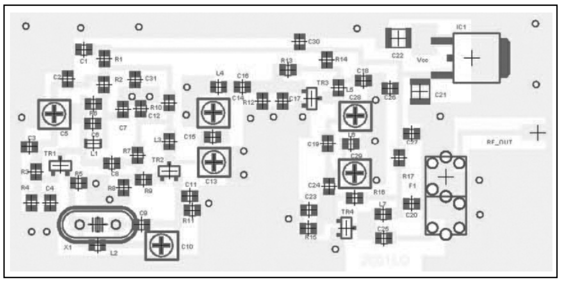


Fig 11.36: Component overlay, slightly smaller than full size (71.6 x 34.6mm)

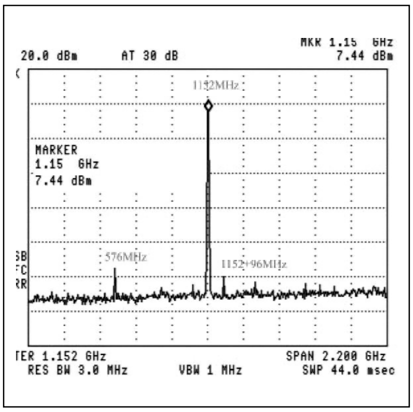


Fig 11.37: Output spectrum of the 20011.0 microwave source from 50 to 2250MHz. The measured output is +9dBm at 1152MHz on narrow span. Half frequency (576MHz) is >45dB down on the 1152MHz output. All other non-harmonic outputs are over 50dB down on the 1152MHz output

connect the ground plane to areas of the component side of the PCB. Normally these are plated through holes (PTH). Use 0.3 - 0.4mm diameter tinned wire through 0.5mm diameter holes. Use only 28SWG solder, both for the through board links and for soldering the SMD parts. Standard size 22SWG solder is unacceptable and will make a mess of the PCB. The 2001 LO PCB component overlay is shown in Fig 11.36.

Alignment

Apply +12V and check that the current draw is no more than about 50mA. If it is significantly higher, check for faults. Check for +8 volts at the output of IC1.

The source can be aligned with little more than a multimeter. However, this is not recommended, as it can be difficult to ascertain if intermediate stages have been tuned to the correct frequency. A far better technique is to use a spectrum analyser with the probe described in [16], see **Fig 11.37**. Tune the analyser to 96MHz with 1MHz span and set to the amplitude reference to 0dBm. Place the probe on the base of TR2 and adjust C5 for a strong 96MHz output indication (approximately -15dBm with the -20dB probe). C5 should be set around mid capacitance. Re-tune the analyser to 288MHz and place the probe on the base of TR3. Tune C13 and C14 for maximum 288MHz output (approximately -15dBm). C13 and C14 should be set to near maximum capacitance.

Re-tune the analyser to 576MHz. Place the probe on the base connection of TR4 and adjust C28 and C29 for maximum 576MHz output (approximately -15dBm). C28 and C29 should be set around mid capacitance.

Finally, re-tune the analyser to 1152MHz and set the amplitude reference to +20dBm. Connect the analyser to the output connector and adjust the cores of Fl for maximum output. This should be in the range +7 to +11dBm.

With a suitable means to accurately measure the output frequency, adjust C10 for 96.000000MHz. It may be necessary to remove C9 with some crystals. L2 is not normally required, but provision is made to add this inductor for use with some crystals that will not otherwise adjust onto frequency.

There will be some frequency interaction with the adjustment of C13 and C14 as the TR2 frequency tripler will tend to pull the frequency. To a lesser extent, so will adjustment of C28 and C29.

Work in progress

By the time this is in print, small changes may have been made to the circuit. It is advisable to check Sam Jewell's web page [17] for the latest information before commencing construction.

RSGB members can download the PCB foil pattern and comprehensive components list from the RSGB *RadCom Plus* website [18]

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 11.38**, was designed by Wolfgang Schneider, DJ8ES, and Frank-Peter Richter, DL5HAT, and published in *VHF Communications* [19]. It uses an HP10544A VXCO (**Fig 11.39**), 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 ($\pm 1 \times 10^{-7}$).

The first practical measurements were based on a gate time of 8s, which corresponds to a resolution of 0.125Hz. Together

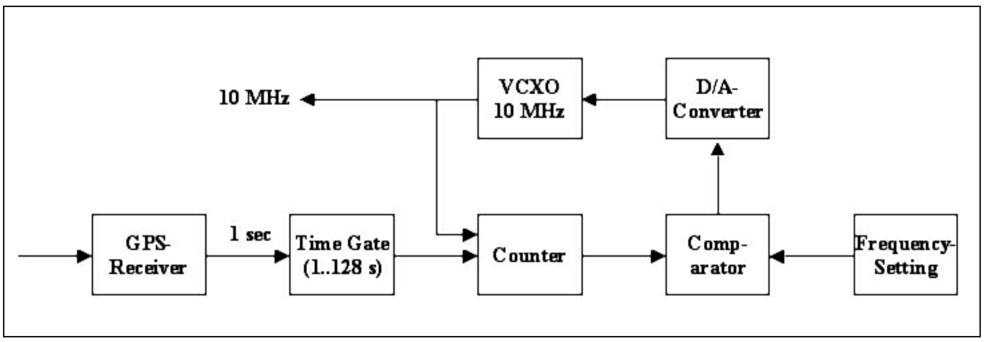


Fig 11.38: Block diagram of frequency control via GPS

Fig 11.39: HP10544A VXCO

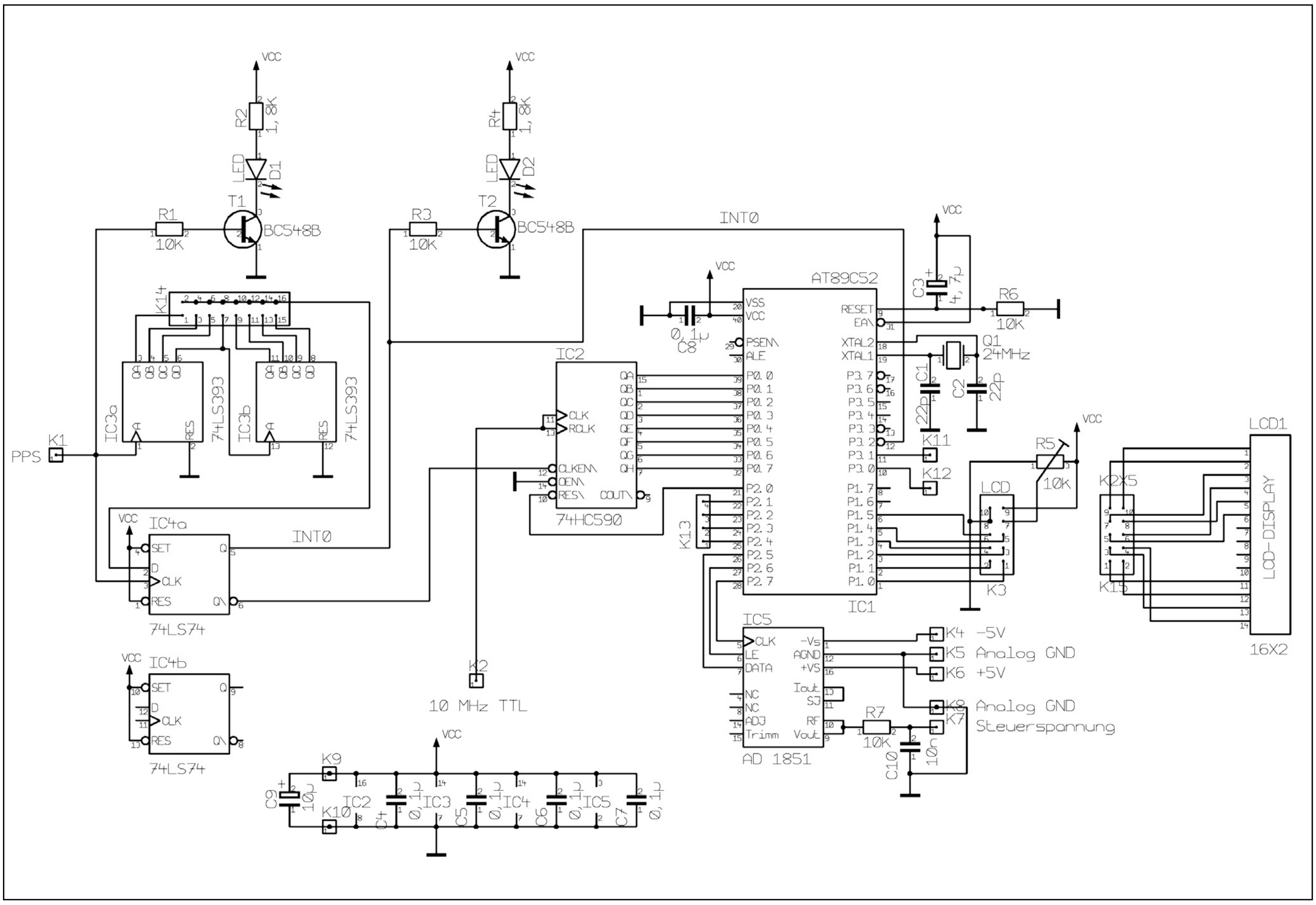


Fig 11.40: Circuit diagram of GPS control stage of high precision frequency standard for 10MHz

with the phase jitter of the GPS signal, there should have been 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 11.40) 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.

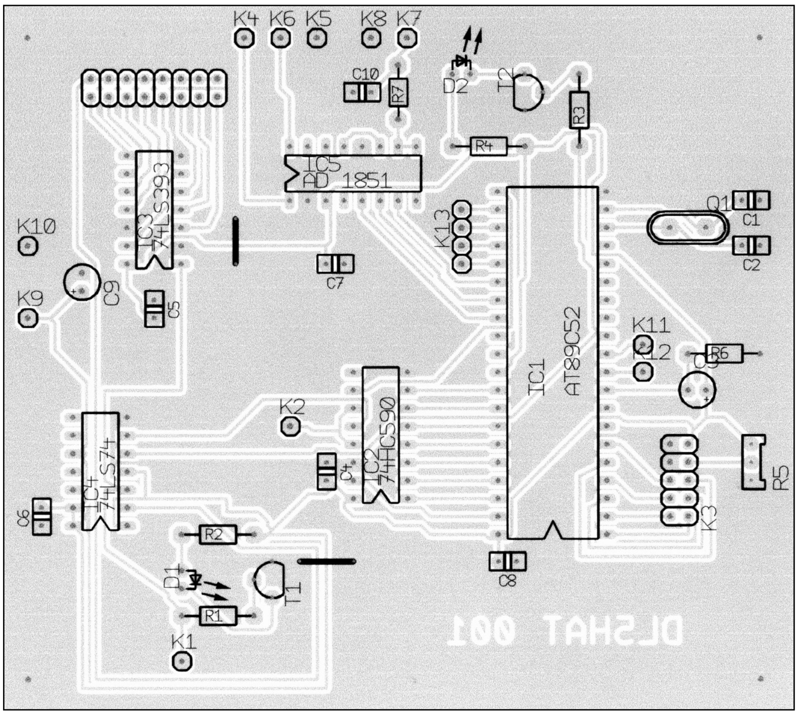


Fig 11.42. Component layout for GPS control stage of high precision frequency standard for 10MHz

Table 11.7: Component list for GPS controller for 10MHz frequency standard

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

The frequency controller circuit is assembled on 100mm x 100mm double sided PCB (**Fig 11.41** in Appendix B), the component layout can be see in **Fig 11.42** and the component list is shown in **Table 11.7**.

In DL5HAT's prototype (**Fig 11.43**) the HP10544A oscillator was used to drive a buffer stage designed by DJ8ES [20]. 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

Fig 11.43: Picture of the prototype high precision frequency standard for 10MHz controlled by GPS

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

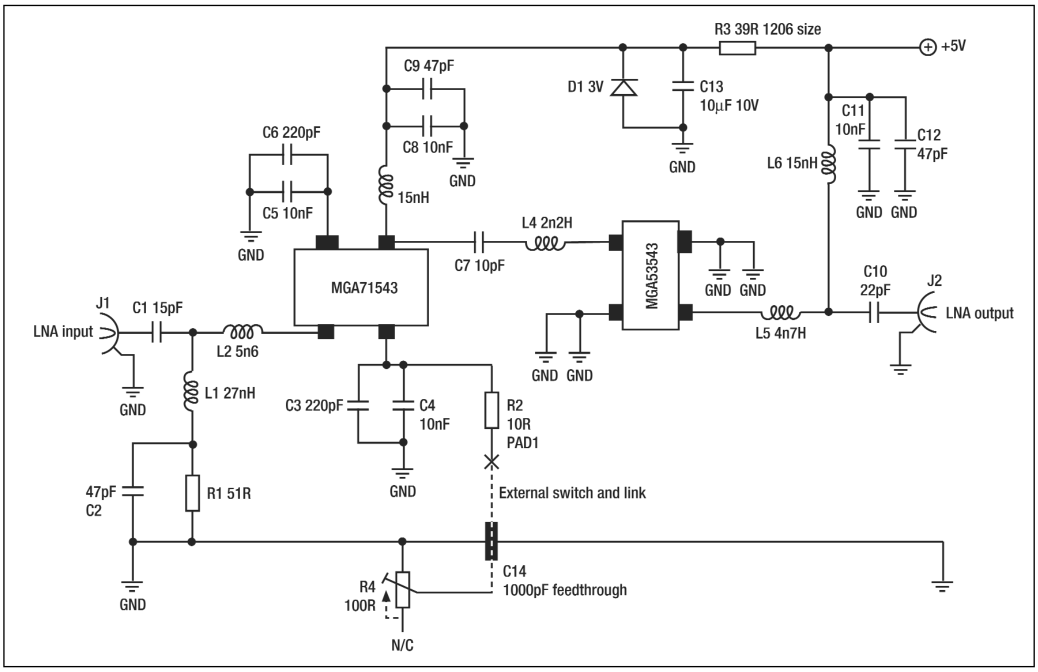
RECEIVE PREAMPLIFIERS

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. Three designs are shown here to illustrate the technology available.

23cm Preamplifier

Developments in the early 2000s of cellular radio component technology have provided the VHF, UHF and microwave enthusiast with access to low-cost, low-noise, high-dynamic-range parts such as the Agilent ATF54143 PHEMT FET. (Pseudomorphic High Electron Mobility Transistor) As a result, if you are looking for an effective low-noise 23cm preamplifier there have been several designs published in the amateur press. While there have been many such designs optimised for the low-noise amplifier stage, there has been a dearth of designs for the equally-important receive converter second-stage amplifier. Using some of this new technology, a switchable, low-noise, second-stage amplifier has been developed for use with a masthead preamplifier.

Sam Jewell, G4DDK had been developing a new 23cm transverter and one of his design objectives was to enable the receive side to maintain high dynamic range when using a masthead preamplifier, yet still have adequate sensitivity if used as a stand-alone transverter. The key to achieving this was discovered when the data sheet for the Agilent MGA71543 was studied. This GaAs Microwave Monolithic Integrated Circuit (MMIC) amplifier device is designed for use in Code Division Multiple Access (CDMA) and W-CDMA cellular receivers, where



11.44: Circuit diagram of the 23cm switchable amplifier

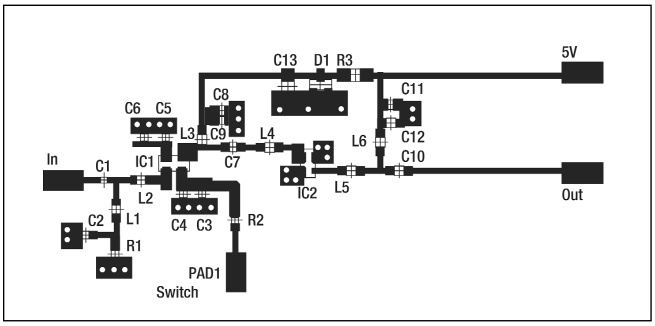


Fig 11.45: PCB layout of the 23cm switchable amplifier (not exact size - see text)

the level of received signal is critical to maintaining traffic capacity. This device incorporates a simple, internal, switching system where the amplifier can be bypassed whilst maintaining a good match at the input and output. This type of device is called a low noise amplifier mitigated bypass switch. When used with a good masthead preamplifier, the system receive noise figure can be maintained below 1dB, with relatively good dynamic range, but when a very strong local signal appears, the MGA71543 bypass switch can be operated to reduce system gain, so that operation can continue, although with a higher system noise figure, and with reduced blocking or intermodulation problems.

The transverter design required 30dB of pre-mixer gain without the masthead preamplifier. An MGA71543, on its own, can provide 16dB gain with a noise figure below 1dB at 23cm. A second amplifier stage is obviously required to achieve 30dB gain. A good second-stage candidate is the MGA53543. Using the MGA71543 as the first stage and the MGA53543 as the second stage, the two-stage amplifier measured 29.6dB gain at a noise figure of 1.1dB and an input intercept (IIP3) of -3dBm (calculated from two-tone measurements). This is a slightly lower IIP3 than expected, although it is adequate for current requirements. Switching out the MGA71543 reduces the overall amplifier gain to 8.7dB, with 8.2dB noise figure, but provides a measured 11P3 of +11.5dBm. Depending on the gain and noise figure of the masthead preamplifier and receive feeder loss, the overall system noise figure can still be maintained below 2dB with the MGA71543 switched out and below 1dB with it in circuit.

Component	Value	Type	Supplier
R1	51R	SMD 0603	Rapid Electronics
R2	10R	SMD 0603	Rapid Electronics
R3	39R	SMD 1206	Rapid Electronics
R4	100R	10 Turn trimmer	Rapid Electronics
C1	15pF	SMD 0603 NPO	Farnell InOne
C2	47pF	SMD 0603 NPO	Farnell InOne
C3	220pF	SMD 0805 NPO	Farnell InOne
C4	10nF	SMD 0805 NPO	Farnell InOne
C5	10nF	SMD 0805 NPO	Farnell InOne
C6	220pF	SMD 0805 NPO	Farnell InOne
C7	10pF	SMD 0603 NPO	Farnell InOne
C8	10nF	SMD 0805 NPO	Farnell InOne
C9	47pF	SMD 0805 NPO	Farnell InOne
C10	22pF	SMD 0603 NPO	Farnell InOne
C11	10nF	SMD 0805 NPO	Farnell InOne
C12	47pF	SMD 0805 NPO	Farnell InOne
C13	10ìF	SMD 10V wkg Tantalum	Farnell InOne
C14	1000pF	Screw- or solder-in feedthrough	Mainline Electronics
L1	27nH	SMD 0603	Farnell InOne
L2	5n6H	SMD 0603	Farnell InOne
L3	15nH	SMD 0603	Farnell InOne
L4	2n2H	SMD 0603	Farnell InOne
L5	4n7H	SMD 0603	Farnell InOne
L6	15nH	SMD 0603	Farnell InOne
D1	3V	SOT23 Zener	Rapid Electronics
IC1	MGA71543		Agilent BFI Optilas
IC2	MGA83543		Agilent BFI Optilas, Farnell InOne
Tinplate box		E-Kafka 1000102	Eisch-Kafka Electronics
J1		SMA 4 hole mounting socket	Farnell InOne
J2		SMA 4 hole mounting socket	Farnell InOne

Table 11.8: Components list for the 23cm switchable amplifier

Fig 11.44 shows the circuit schematic of the two-stage amplifier and the components list is in **Table 11.8**. The layout (**Fig 11.45**) has been derived from the application notes for each device, with matching for 1.3GHz rather than the published

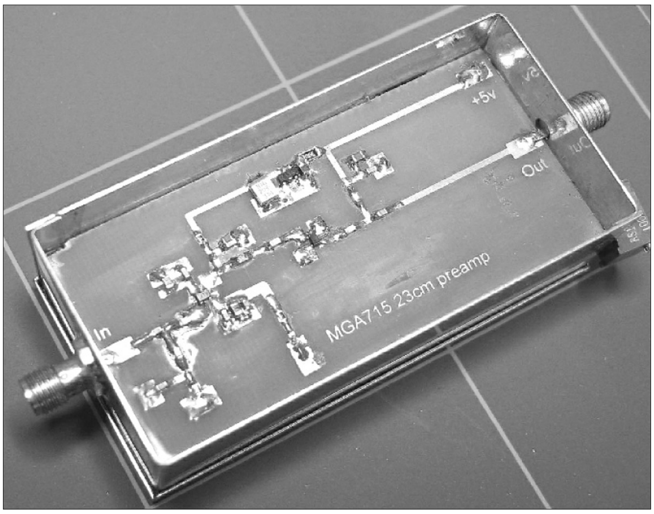


Fig 11.46: The prototype 23cm switchable amplifier in its tinplate box. This view shows the component side of the PCB

900MHz or 1800MHz cellular allocations. The MGA71543 first stage device is noise-matched with a "T" circuit using a series capacitor and inductor and a shunt inductor. It is important that the indicated values are used or noise figure will suffer, as will gain.

The input device has two source connections. One of these is double decoupled with two capacitors. The second source connection is also decoupled with a similar arrangement, but here the source connection is connected through a 10 ohm resistor and then a feedthrough capacitor to an external 100 ohm potentiometer and toggle switch, used to set the bias and bypass the stage respectively. The value of the potentiometer should be adjusted for best noise figure or best intercept, depending on requirements. Opening the switch removes the bias and causes the MGA71543 to bypass itself internally. A further improvement in noise figure can be achieved with the use of better (lower loss) decoupling capacitors on the source connections of the MGA71543.

The first stage 3V drain voltage is derived from the 5V supply by using a well-bypassed 3V Zener diode and dropping resistor. The resistor value is chosen to pass 10mA Zener current when the MGA71543 is set to draw 40mA. If the first stage is operated at much less than 30mA, to improve noise figure, the series resistor R3 must be increased accordingly to ensure that the Zener diode current does not exceed about 10mA.

An MGA53543 is capable of a +38dBm output intercept at 15.4dB gain, if the output match is optimised. In this amplifier, the output match is not fully optimised, resulting in a lower than expected IIP3.

The second stage is matched at its input using surface mount inductors and capacitors. The MGA53543 requires 5V drain-to-source bias and therefore operates directly from the +5V supply.

Fig 11.46 shows the amplifier built into in a standard Schuberth tinplate box, with dimensions 37 x 74 x 30mm, obtainable from Eisch-Kafka Electronics in Germany [21]. The amplifier is built on FR4 double-sided 1.6mm thick fibreglass PCB. A 1:1 PCB foil for the amplifier can be obtained from G4DDK on request [22].

Home produced double-sided PCBs with good quality through board connections can be difficult to engineer. G4DDK's solution to this problem is to drill 0.5 - 0.6mm diameter holes, where through board ground connections are required, using a sharp high-speed drill bit. A single strand of thin tinned or silver plated wire is then threaded though the holes and soldered both sides using very thin (28SWG) solder to make a small size soldered joint over the wire. The small diameter hole seems to produce a capillary action, sucking the solder down into the hole and effectively making an excellent through board connection. This technique avoids the unsightly and often-ineffective wire `worms' formed by off cuts of old 0.25W resistors.

The two MGA devices are mounted as shown, taking care to ensure that the larger (wider) source lead is orientated as shown in the circuit schematic. SMA connectors are used at the input and output.

The 5V supply should be connected into the box through a second feedthrough capacitor.

G4DDK has become a devotee of surface mount inductors and capacitors in place of the more popular, but ungainly, microstrip matching arrangements. He uses 0603 size surface mount components where possible although, at 23cm, the larger 0805 size is quite usable. 0603 size inductors and capacitors are often cheaper than the 0805 size alternatives.

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 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 [23] by using a wire loop with an open stub as an input circuit (Fig 11.47). The FET's grounded source requires a bias circuit to provide the negative voltage for the gate. A special active bias circuit (Fig 11.48) is integrated into the RF board that provides regulation of voltage and current for the FET. The component list is 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

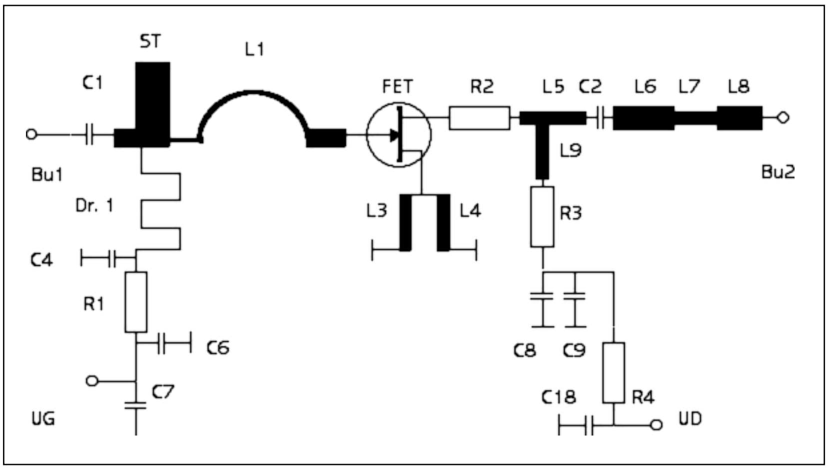


Fig 11.47: Circuit diagram of single stage 13cm PHEMT preamplifier

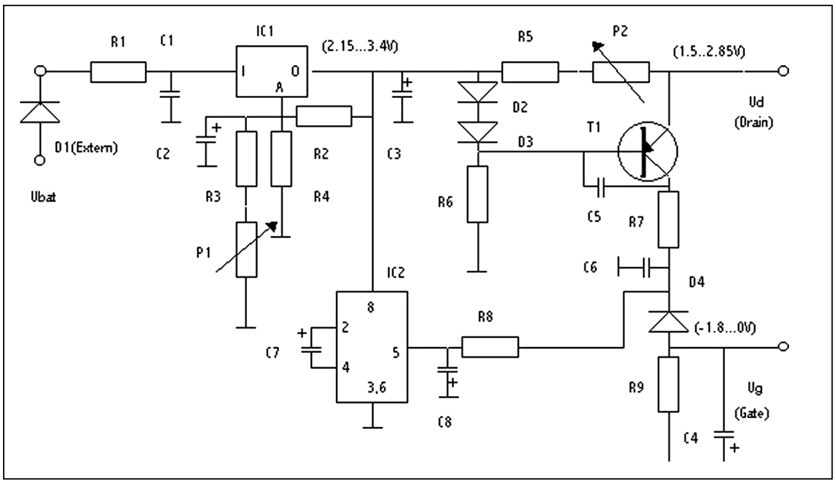


Fig 11.48: Circuit diagram of bias circuit for 13cm PHEMT preamplifier

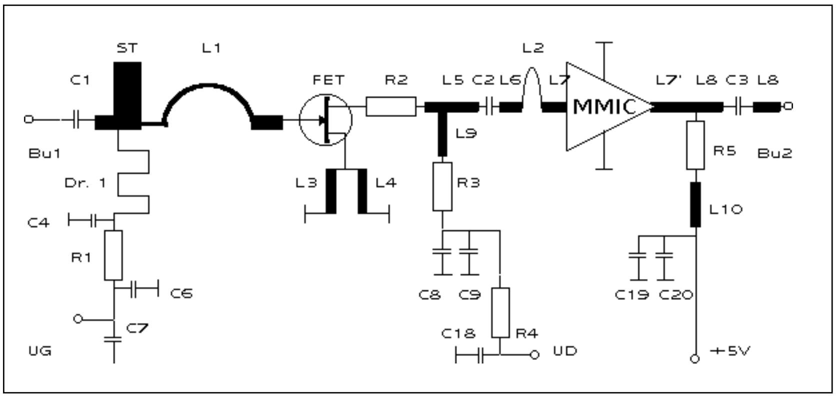


Fig 11.49: Circuit of the two stage 13cm PHEMT preamplifier

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 11.49**) 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 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.

Table 11.9: Components for 13cm PHEMT preamp

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	47O SMD-R size 1206
R2, 14	39O SMD-R size 1206
R4, 5	10O SMD-R size 1206
R12	6.8kO SMD-R size 1206
P1	100O 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

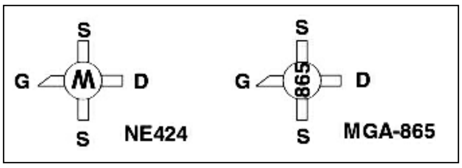


Fig 11.51: Top view of FET and MMIC used in 13cm preamplifier

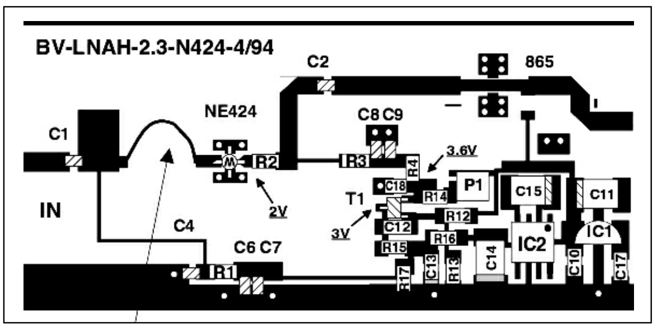


Fig 11.52: Component layout for single stage 13cm PHEMT preamplifier

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 11.50** in Appendix B). **Fig 11.51** shows a top view of FETs and **Fig 11.52** 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 at 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 11.53 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 11.54 and 11.55 show the results for gain and noise figure for the one stage and two stage version respectively.

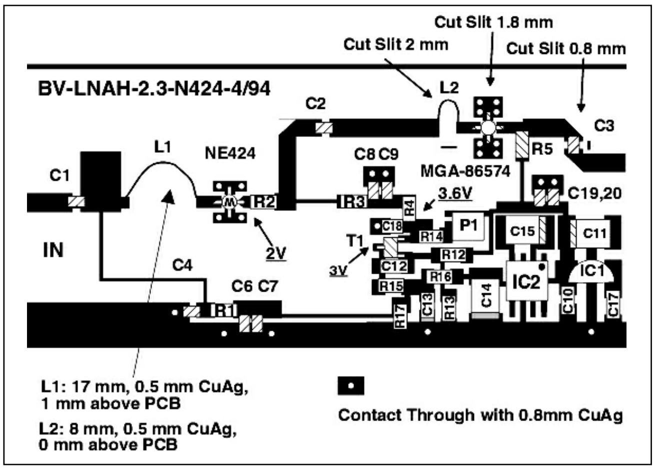


Fig 11.53: Component layout for two stage 13cm PHEMT preamplifier

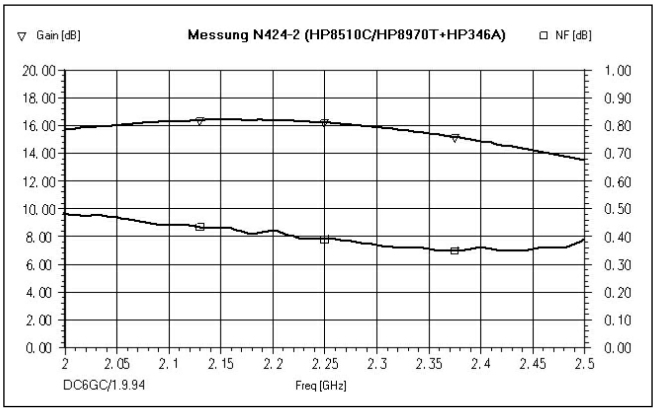


Fig 11.54: Noise figure and gain measurements for single stage 13cm preamplifier

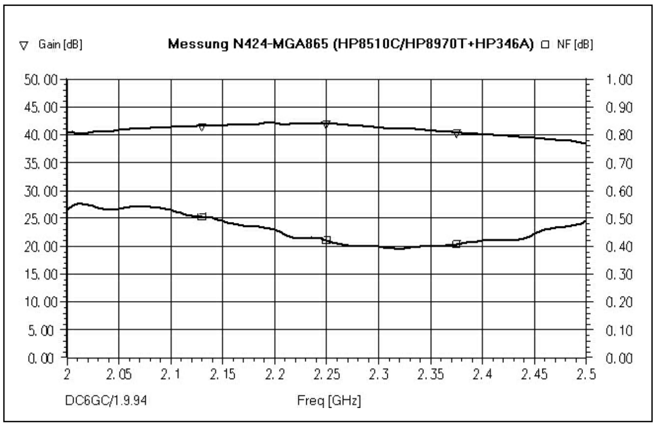


Fig 11.55: Noise figure and gain measurements for two stage 13cm preamplifier

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 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 broadband. They can cover the various portions of the 13cm 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 0.35 micrometres. 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 [24]. 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 [24] provides adequate performance with a typical NF of 0.55dB.

Low Priced 10GHz Preamplifiers

Described by Gerard Galve, F6CXO

Franco Rota runs an RF component supply company in Italy called R F Elettronica [25]. His main objective is to sell bulk components such as SMD parts to the electronics industry. He attends some radio rallies in Europe and often has interesting items for sale that can be used or adapted by radio amateurs for use on the amateur bands. Franco sells the PCB described in this article (**Fig 11.56**) for 3 Euros. The board was initially purchased to salvage the 4 NE 32584s. After examination, I realised it was fairly rare to find satellite boards with preamplifiers that are so well aligned and so suitable for modifications.

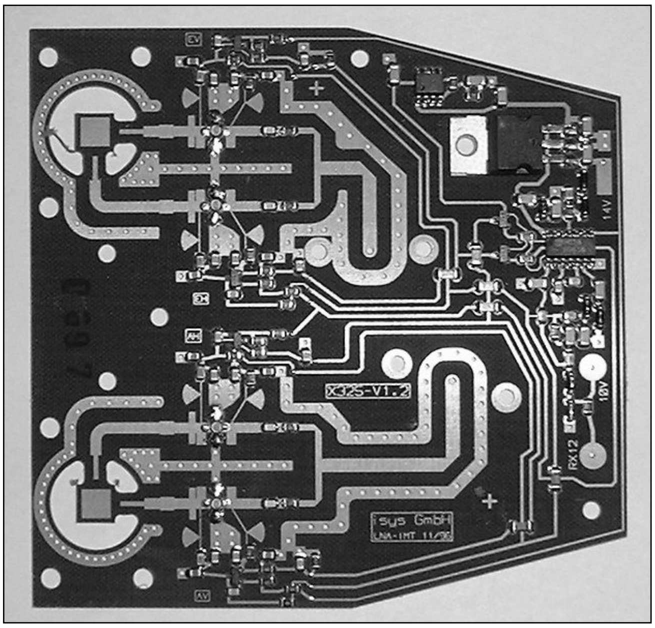


Fig 11.56: The amplifier PCB available from Franco Rota

Fig 11.57 shows a close up of the two preamplifier circuits. Preamplifier No. 1 can be adapted to a wider range of enclosures. The input track can be cut at any point throughout the black area.

Fig 11.58 shows that if you are careful, you can even salvage two preamplifier circuits on this board; use the output capacitor of one as the input for the other. You could just salvage transistors (that was the original idea).

Fig 11.59 shows how the output capacitor of the right hand preamplifier is connected to the input of the left hand one. The

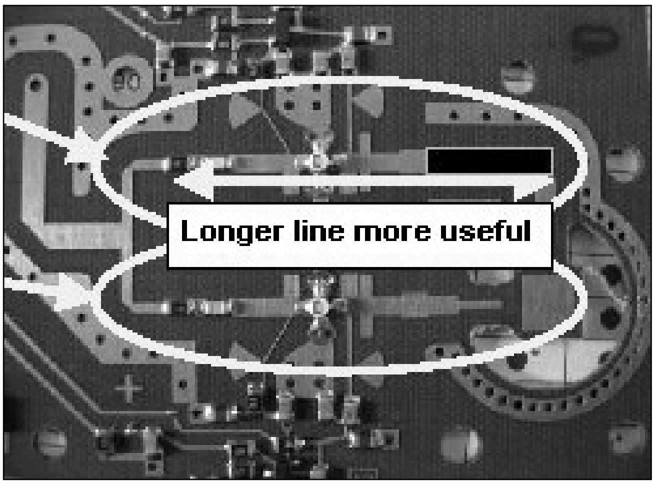


Fig 11.57: One pair of amplifiers. The top circuit is referred to as Preamplifier No 1; the bottom one is amplifier No 2

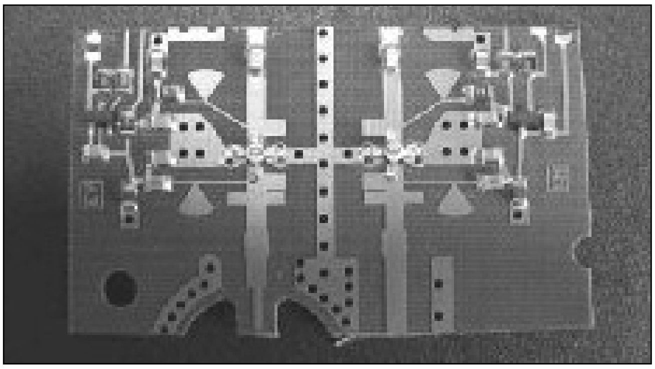


Fig 11.58: How to use two amplifier circuits

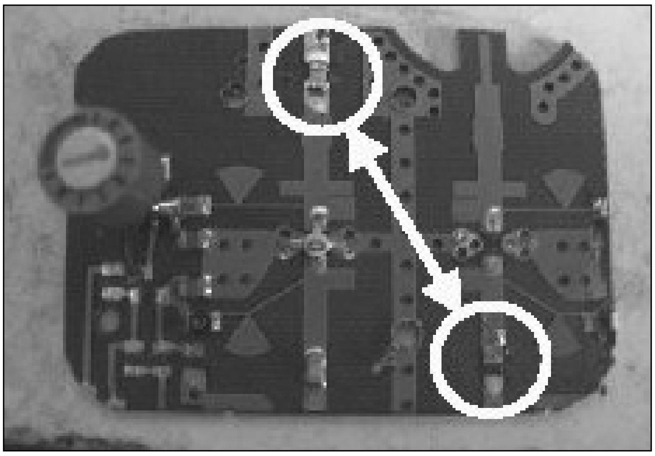
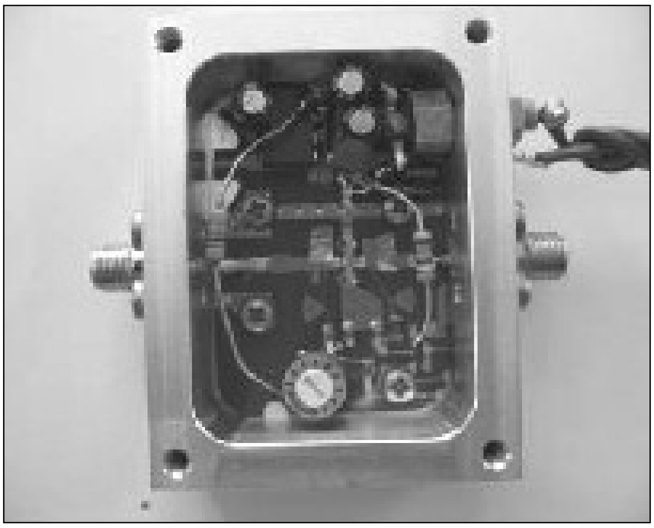
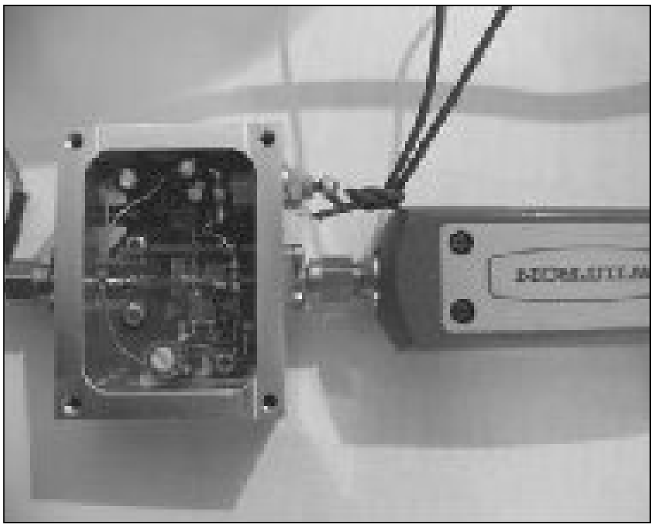
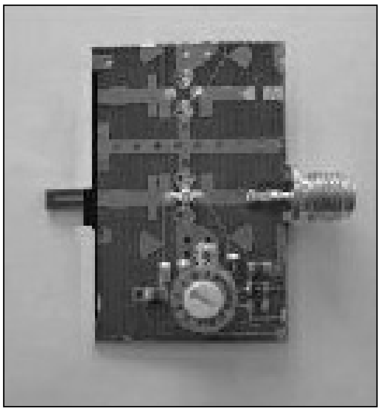


Fig 11.59: Connecting two amplifiers together







(Above right and left) Figs 11.60 and 11.61: Example housing for the 10GHz amplifier

(Left) Fig 11.62: Using Preamplifier No 2 with direct waveguide input

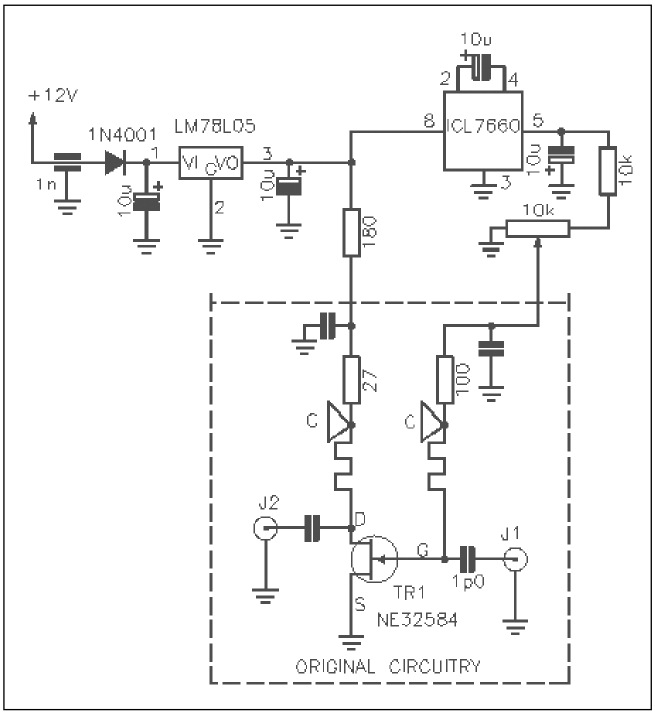


Fig 11.63: Circuit diagram of the preamplifier and power supply used

negative supply potentiometer can be fitted on the existing printed circuit board.

It is that easy, the hardest part is to find a housing that suits the length of the printed circuit board. Just use your imagination and see what might be available at the back of some old drawer. Figs 11.60 and 11.61 show the author's example.

Performance

This box of tricks produced a noise figure of 0.7dB without the lid, 0.8dB with the lid, and a gain of 14dB from 10,368 to 10,450MHz. That's pretty fantastic for 1.5 Euros.

Alternative use

Circuit No. 2 is a little too short to position a capacitor at the input. It can be used very easily with the input line directly in the waveguide as used in the original design (Fig 11.62).

Circuit diagram

The original diagram is in the box shown in Fig 11.63. The power supply was made very simple with wiring "in the air". The ICL 7660 is upside down and the components are soldered directly onto the pins. The drain current should be set to 10mA.

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 being replaced by the MOSFET hybrid amplifier RA18F1213G. Discrete semiconductors are also replacing hybrid modules, an example is 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 as direct bonding to semiconductor substrates, and an example of these techniques is shown below.

L Band Power Amplifier for AO-40 Uplink

Amateur radio satellites offer many options for experimentation as can be seen in the chapter on satellites and space.

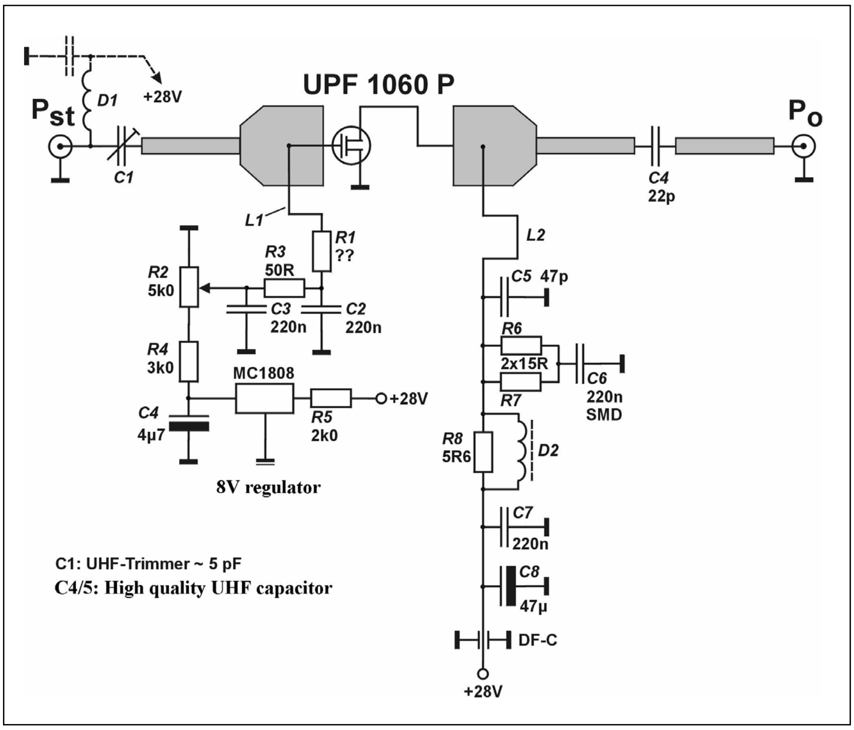


Fig 11.64: Circuit of L band amplifier

The AMSAT-OSCAR 40 (AO-40) satellite is currently in orbit after surviving an explosion on board when the orbit was being established. In addition to operating on other bands, it has an uplink in the 1296MHz L Band. However, the power required for satisfactory radio contact is above the output of a normal transceiver, which makes it necessary to use a power amplifier.

A suitable RF power amplifier has been designed by Konrad Hupfer, DJ1EE. The experiences of many AO-40 users have demonstrated that a PEP of approximately 50W at the input of a circular radiating antenna with approximately 20dB gain is sufficient for the L Band uplink, even using squint angles. Looked at from the point of view of cable losses it makes most sense to generate the power at the point where it is used: directly at the antenna power feed. The amplifier described here has the following characteristics:

- P_{opep} = 50W
- U_{ds} = 28V
- G = 12dB
- I_q = 300mA

Directly mounting an L Band helix or a patch antenna on the rear of the reflector offers the ideal solution. The reflector plate of an antenna (eg made from 3mm aluminium) having the normal area of approximately 400cm², is not quite adequate for a heatsink at ambient temperatures of >25°C. One remedy can be an additional 'chimney', consisting of approximately 1mm thick aluminium plate, with an associated cover. This additional cooling makes it possible to obtain thermally stable functioning, even in summer, for normal SSB mode with a PEP of approximately 55 watts.

The amplifier

The L Band amplifier uses a fairly standard circuit (**Fig 11.64**). Printed line transformers are used to transform the relatively low complex input and output impedances of the L-DMOS transistor used (UPF 1060P from ULTRA RF). These transformers are made on the familiar material RO 4003, substrate thickness 0.79mm, r = 3.35, using stripline technology.

The general calculations for transformation networks of this type can be found, among other places, in [26]. The printed matching networks are shown in **Figs 11.65 and 11.66**. Since the most important dimensions of the lines are shown, you can easily construct them yourselves.

The first specimens of the amplifier circuit boards were designed with Indian ink, using the old technology, and then etched in the usual way. The undersides of the circuit boards naturally have a copper coating.

In the present circuit, for the sake of simplicity, no stabilisation is provided for the 300mA quiescent current. It would also be expedient to incorporate temperature compensation, to be prepared for 'extreme cases'. A cut out when excessive temperatures arise is also recommended.

The DC wiring can be laid out as you wish in accordance with the mechanical size of the components used. It can be seen from **Figs 11.67 and 11.68** that, in addition to the transformer lines, some small areas are provided. These are the so-called trim elements, used to fine-tune the amplifier for the best input matching or for maximum output power and linearity. This is needed to compensate for, for example, when similar transistors are used, differences in the input and output impedances occur or there are tolerances in the transistor mounting.

Mechanical and electrical assembly

The semiconductor type selected, UPF 1060 P, is a flangeless model, because of price. Fig 11.67 and **Figs 11.69 and 11.70** show one possible assembly, using a 3mm thick carrier plate made from copper as a heat spreader. This mounting plate should be as level as possible on both sides; it should preferably be finished by surface milling on both sides. The individual circuit boards are soldered onto the copper plate using a hotplate. The distance of 6.5mm between the circuit boards should be adhered to; the copper surface must not be tinned in this area.

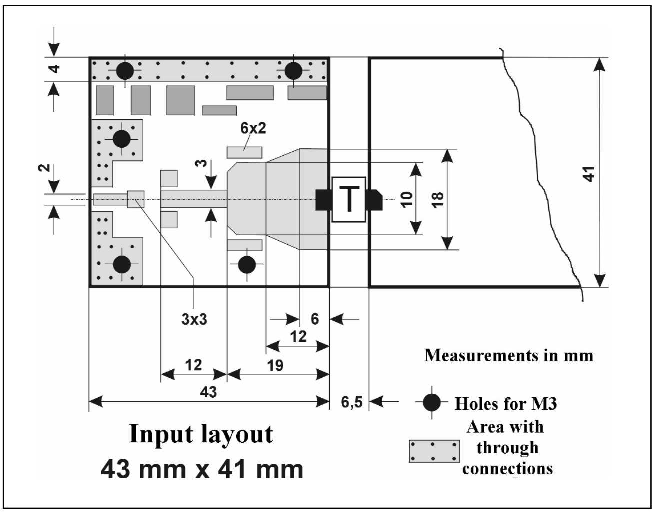


Fig 11.65: The input circuit layout showing measurements of matching networks

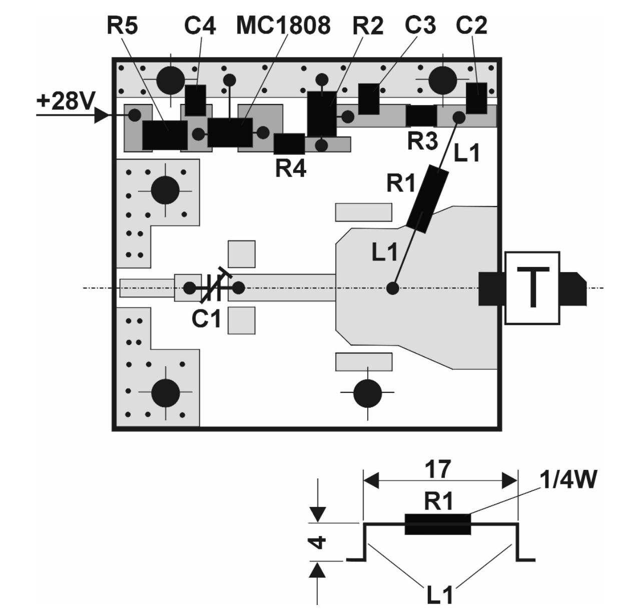


Fig 11.66: Component layout for input circuit showing details of L1 and R1

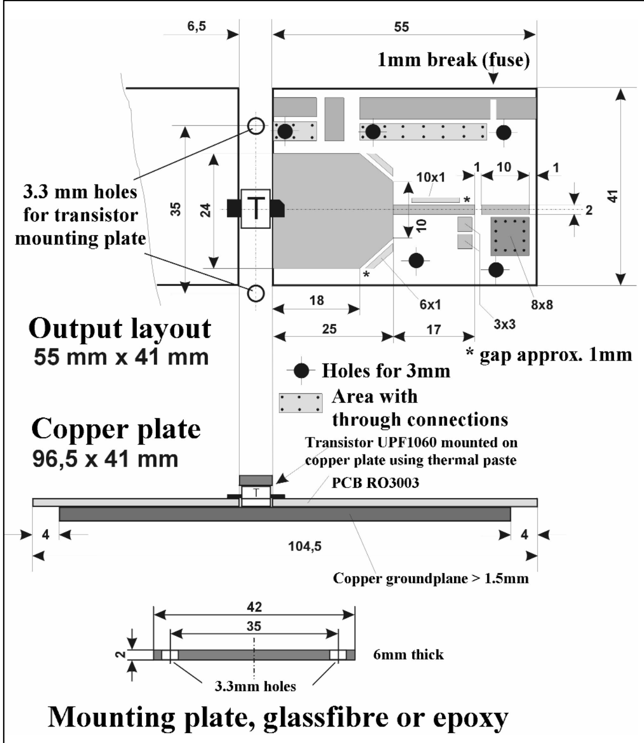


Fig 11.67: Output circuit layout showing measurements of matching networks and details of transistor mounting

When the circuit boards have been assembled, the transistor is put in place, but initially without its connections soldered to the striplines.

The module (copper plate with soldered-on circuit boards) is screwed into the housing, the base of which has been finished by surface milling. A uniform application of heat conducting paste should be used between the transistor, the copper plate and the housing base. The transistor is now pressed onto the copper plate using a flexible implement (un-coated printed circuit material, fibreglass etc). You must treat the equipment

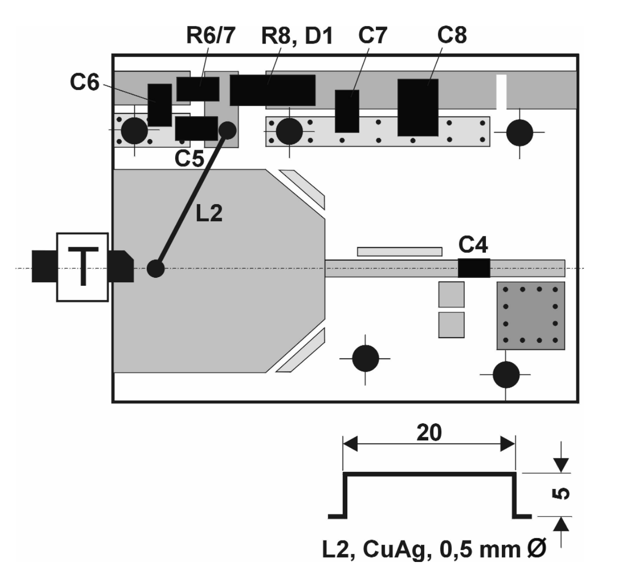


Fig 11.68: Component layout for output circuit showing details of L2

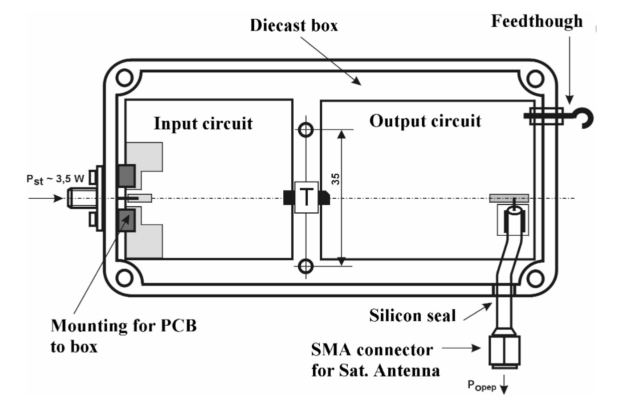


Fig 11.69: Installing the amplifier in an Eddystone diecast box measuring 111mm x 61mm x 30mm

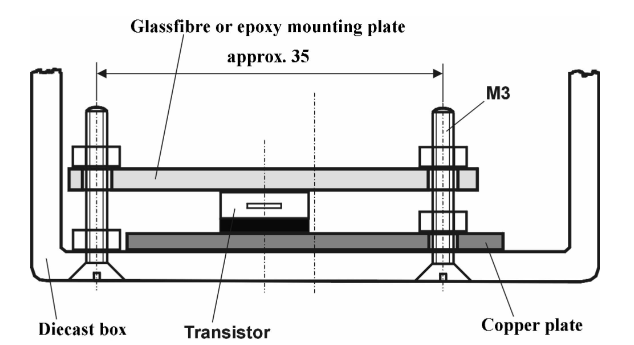


Fig 11.70: Details of transistor mounting using mounting plate

carefully. Finally the connecting lugs can be soldered (Figs 11.69 and 11.70).

The amplifier must now be mounted flat onto the chimney/radiation reflector installation in its housing for best heat dissipation. The holes required for M3 countersunk screws can be made in the reflector. Another solution is to tap M3 threads into the reflector and then screw the entire unit onto the reflector from the amplifier side (Figs 11.71 and 11.72).

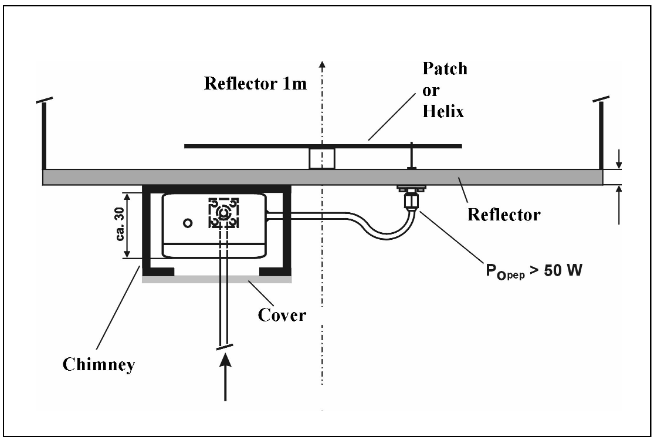


Fig 11.71: Mounting the amplifier on an antenna with cooling chimney approximately 150 x 200mm

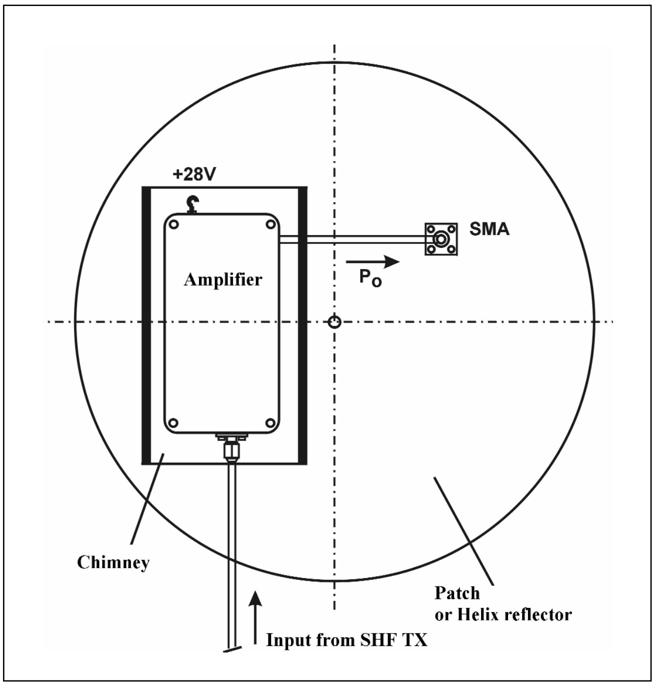


Fig 11.72: Rear view of mounting the amplifier on an antenna with cooling chimney

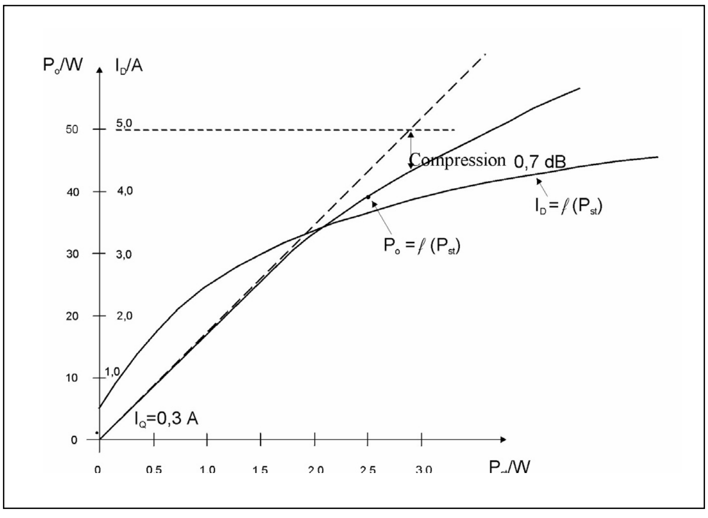


Fig 11.73: Transfer characteristics of the L band amplifier

Various mechanical solutions can be used depending on the application. For simplicity's sake, the power supply in this layout is provided using a feed-through filter mounted in the housing wall. The supply voltage of 28V can be fed through the coaxial feeder cable. The entire structure should be painted white to keep the temperature as low as possible, even when the sun is shining!

The circuit has been assembled and tested many times. The characteristics, interpolated from the worst values of three amplifiers are shown in **Fig 11.73**. The typical measurements of the RF power amplifier for single-tone test are as follows:

- P_O = 50W (with compression approximately 0.6dB)
- J_D = 4.5A
- J_{DQ} = 300mA
- P_{st} = 3.2W
- G = 12dB
- Efficiency ≥ 49 %

G H Engineering PA1.3-100 23cm Power Amp

For the last 10 years I have been using a 23cm power amplifier for contesting using two Mitsubishi M57762 power modules. The amplifier is built into a diecast box with a preamplifier and mounted at the top of the tower to reduce the feeder loss. The power is fed to the amplifier from a 12V gel battery mounted near the amplifier, the battery is float charged from the shack 60 feet below. This arrangement has worked very well but I decided that it was time to try and increase the power output and chose the new PA1.3-100 amplifier from G H Engineering [27]. This is available as a mini kit or a ready built unit; I decided to go for the mini kit and placed an order as soon as the kits became available early in 2008. The following article describes my construction and testing efforts.

Circuit description (from the GH Engineering documentation)

The PA1.3-100 is a linear amplifier for the 1.3GHz band. It is capable of producing at least 100W in the frequency range 1240 - 1300MHz with a drive level of 3W. The amplifier operates in linear mode, which makes it suitable for SSB operation as well as FM. For best linearity, it is recommended that the input power does not exceed 2W PEP, which will give an output power of approximately 80W at 1296MHz. The amplifier uses four Mitsubishi RA18H1213G PA modules that require no tuning being internally matched to 50 ohms at the input and output. An external 13.8V DC power supply is required that is able to supply at least 45A. The amplifier is rated for continuous operation with an input power of 3W and a supply voltage of 13.8V

The PA1.3-100 is intended for use with a transverter of 3W output or similar. It can be used with a 10W transceiver with the use of an external 6dB attenuator. The amplifier incorporates a 7.5dB input attenuator on the PCB; this can be changed or removed if required, thus allowing the amplifier to be used with an input signal down to approximately 750mW.

Fig 11.74 shows a photograph of the circuit diagram. The original is printed on an A3 sheet so it is not really suitable for reproduction on an A4 page in this handbook but the main elements can be seen to help understand the following description of the RF distribution.

The RF input signal is split with a Wilkinson splitter and each of the two signals is then split again with two more splitters. This gives four signals of equal amplitude and phase that are attenuated by approximately 7.5dB by pi-networks.

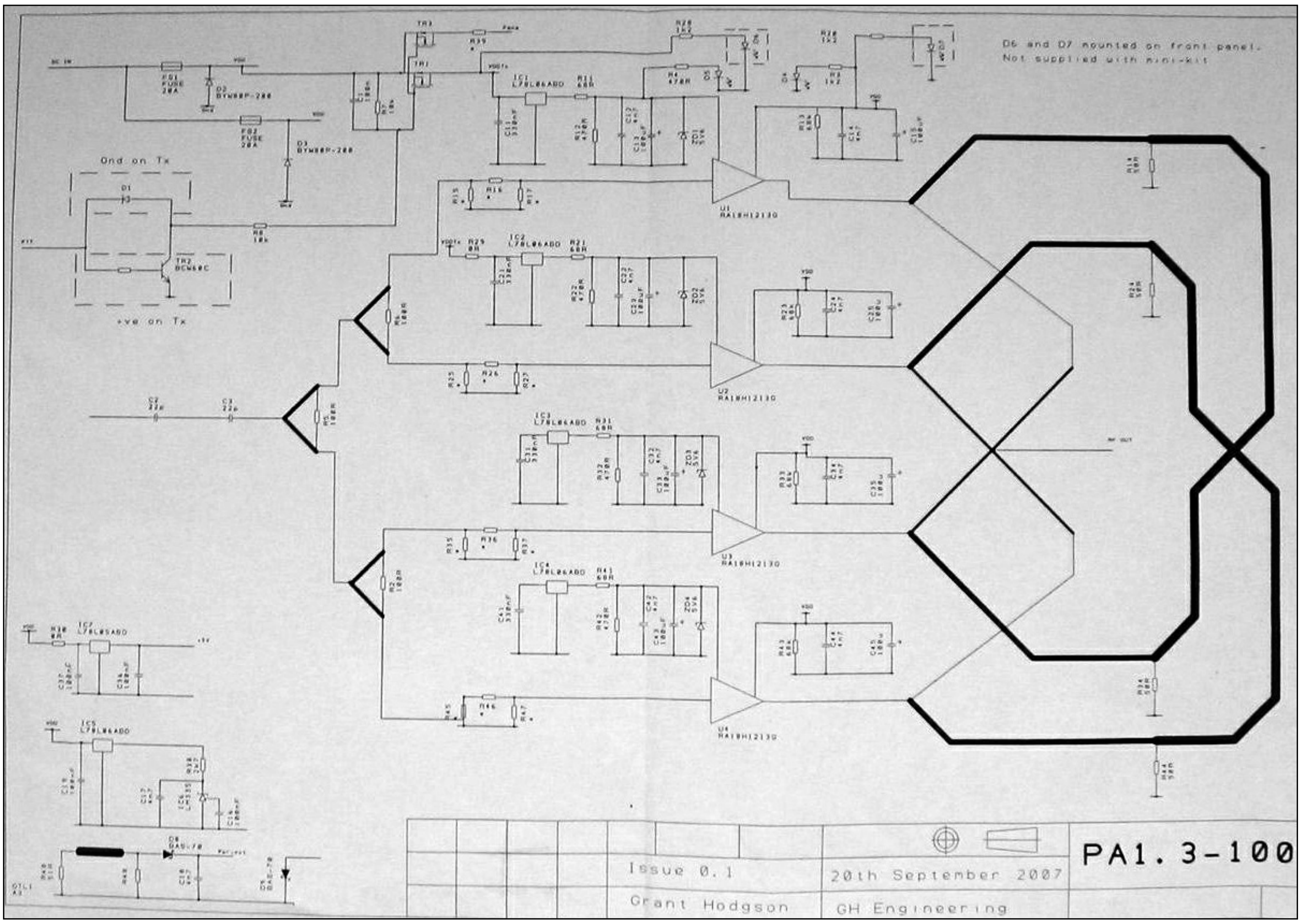


Fig 11.74: Photograph of the A3-sized circuit diagram for the PA1.3-100 23cm power amplifier (see text)

Each RA18F1213G module provides a small signal gain of approximately 23 - 25dB depending on frequency. The input and output microstrip lines have been designed to ensure that the overall amplifier gain is as flat as possible over the 1240 - 1300MHz band.

The output signal is fed to a 4-way Gysel combiner, which is a derivative of the N-way Wilkinson combiner. It consists of four quarter-wave transmission lines, each 100 ohm impedance. Therefore the output signal is transformed to 200 ohms, which gives an output of 50 ohms when all four are connected together in parallel, The isolating network is somewhat more complex than for the input splitters. three-quarter wavelength transmission lines are connected to each of the junctions of the 50 ohm and 100 ohm lines. The ends of these lines are terminated with 50 ohm high power terminating resistors. This is also connected to the end of another quarter wavelength line, of 25 ohm impedance. These four 25 ohm lines are connected together. At this point there is a virtual ground, in that no current flows across the junction, but the impedance is actually very high, and approaches an open circuit. Therefore the impedance at the termination resistor is very low (approaching a short circuit) and no power flows into the resistors. Therefore, the impedance looking into the end of the three-quarter wavelength lines is very high. Consequently, the three-quarter wavelength line places no load on the output, and no current flows into the isolating network.

The PA1.3-100 is fitted with a PCB temperature sensor giving an output voltage that is proportional to temperature. The voltage gradient is 10mV/ C and the voltage is directly proportional to absolute temperature in Kelvin. Therefore at 20 degrees C the output voltage would be 2.93V and at 50 degrees C would be 3.23V.

There is an output power detector using a sampling line and a BAS-70 diode. The output voltage developed is dependant on temperature, to compensate for this dependency a second diode is fitted. A simple power meter can be made using a operational amplifier that compares the output voltage of the two diodes.

Both the temperature sensor and power detector are provided as unsupported features for the constructor to use as they see fit.

The kit

The kit comprises all the electronic components, a pre drilled heatsink and a set of documentation. Fig 11.75 shows the kit of parts and Fig 11.76 shows the documentation and parts list. The first task was to check that everything was present. Thoughtfully the SMD were split into several self-seal bags with the values on a list stuck to the bag. Because some components are difficult, or impossible, to identify the bags are split so that the components can be identified by the quantity of each in the bag. I made a note to keep the components in their bags until I used them so they did not get mixed up.

The next thing to do was to read the assembly instruction to make sure that I knew of any special requirements during assembly. There were a few modifications to do to the PCB and a decision to be made about the sense of the PTT switching. Fortunately I wanted PTT switched to ground for transmit since one of the PCB modifications made it impossible to fit the additional components for PTT going high for transmit. The remaining instructions looked straightforward SMD assembly followed by some care required when mounting the four power modules.

Construction

The heatsink is a hefty piece of aluminium and made a suitable place to mount the PCB and hold it still for assembly of the SMD components (Fig 11.77). Because there were a few errors on the circuit diagram and the component layout care had to be taken to double-check each part as it was assembled but assembly of all the SMD parts was straightforward. The only point to note is that any component connected to the ground surface needed an extra bit of heat to make a good solder joint, so I used two different bits on my old Weller soldering iron. If you have a more modern soldering station this may not even be noticeable.

Once all the SMD components are fitted the instructions require some voltage checks before the power modules are fitted. In my case the voltage measurement of the top power module bias supply was not correct. This was quickly traced to the monitor LED that placed an extra load on the circuit. A quick wire link to a different supply point and all was well.

The next step was to fit the two three-quarter wavelength lines that are made from semi rigid cable. Some care was required to make these and shape them as shown on the diagram in the instructions. My attempt at this is shown in Fig 11.78.

The final step was to bolt the PCB to the heatsink and fit the power modules. Static precautions must be taken when fitting the power modules because they are sensitive to static electricity. The instructions mention that the mounting holes for the power modules must be as clean as possible to provide a good DC and RF path to the power modules, some cotton buds are provided to clean the tapped holes in the heatsink. Thermal transfer paste was spread on the centre (slightly recessed) section on the base of each power module before it was bolted down to the heatsink Because I wanted to mount the amplifier on the top of the mast I wanted the finished unit to be weather proof and did not want to mount the two fans on the heatsink. After a quick conversation with Grant Hodgeson at G H Engineering we decided that the fans are needed if the power amplifier is going to be used continuously for FM or ATV, but because my use was intermittent SSB and the amplifier will be in the open air the fans could be

Fig 11.75: The kit for the PA1.3-100 23cm power amplifier

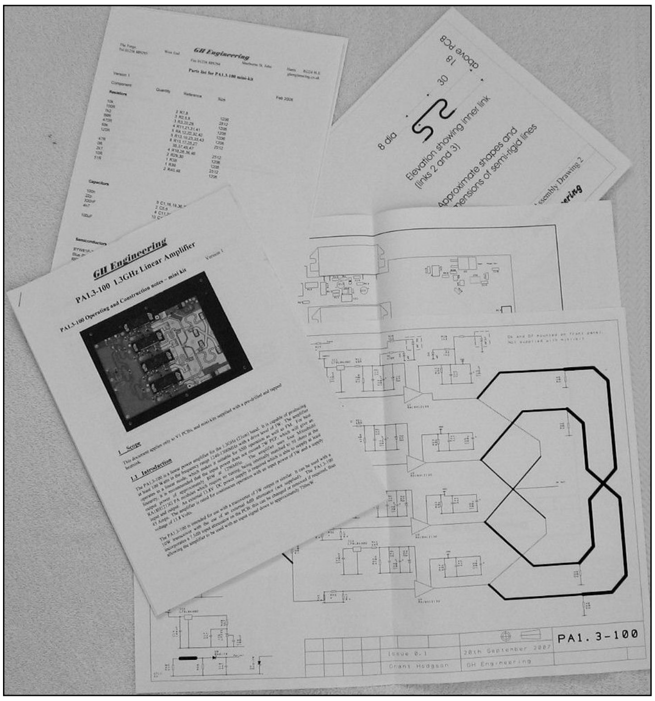


Fig 11.76: The documentation for the PA1.3-100 23cm power amplifier

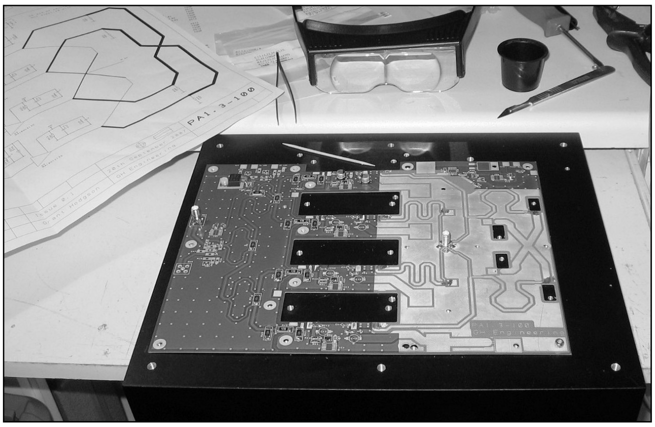


Fig 11.77: The PCB mounted on the heatsink as a workbench to hold the PCB still while fitting the SMD components

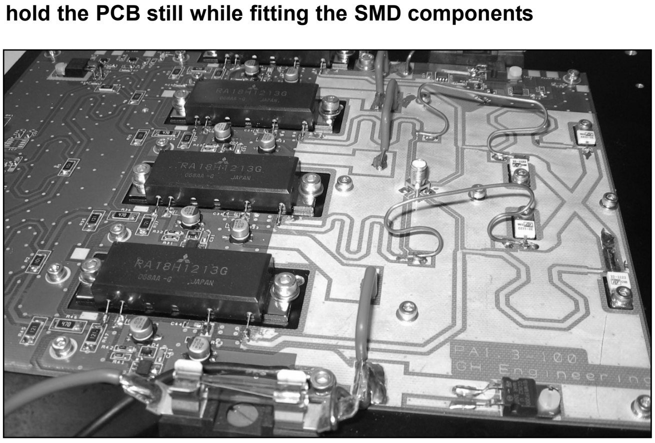


Fig 11.78: The three-quarter wavelength lines fitted to the output circuit

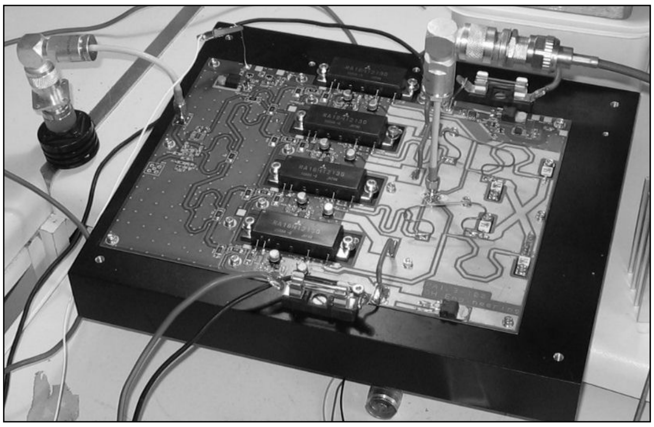


Fig 11.79: The completed PCB ready for initial testing

dispensed with. This meant blocking up the tapped holes in the heatsink used to mount the fans; this was done using Allen screws coated with silicon sealant. The completed power amplifier is shown in Fig 11.79, ready for initial testing.

Testing and fault finding

To make an initial test the amplifier was fitted with a dummy load on the input and output. The instructions state that the current drawn by the power amplifier with no RF input should be approximately 15A so I intended to use my 25A bench power supply for the initial test. When I switched on, the power supply went into current overload, so I knew there was a problem.

The final method of powering the amplifier was to be a 38AH sealed lead acid battery so that was connected up and the amplifier was actually drawing 26A with no RF input. A quick check showed that the amplifier was producing about 10W output on random frequencies (Fig 11.80); it was self-oscillating. An email to G H Engineering indicated that the mounting holes for the power amplifiers were probably not clean enough or some thermal paste had oozed under the mounting flanges.

The modules were removed, holes cleaned and a small spot of thermal paste removed from one of the modules. The method that I used to clean the holes was to shave down a cotton bud and spray it with degreasing cleaner then 'screw' it into the tapped hole. When the amplifier was switched on again it still drew 26A but the output was now on a single frequency. By grounding the control, pin (Pin 2) on each power module in turn the oscillation was isolated to power modules 3 and 4. Again the problem was thought to be poor grounding of the power module so some brass strips were fitted to improve the grounding of modules 3 and 4 (Fig 11.81) but this did not solve the problem.

After carrying out more checks it was noticed that pressing the PCB around the general area of power module 3 and 4

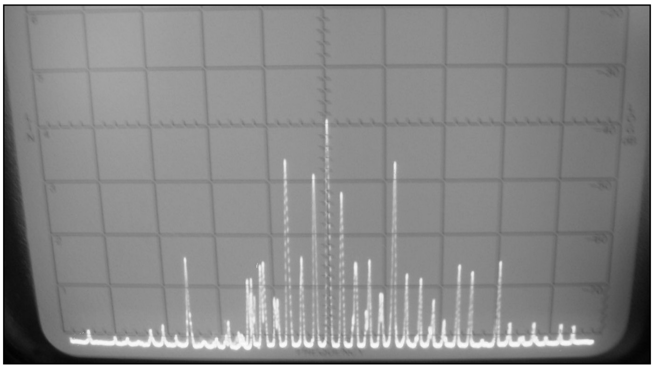


Fig 11.80: The output of the amplifier when it was self-oscillating

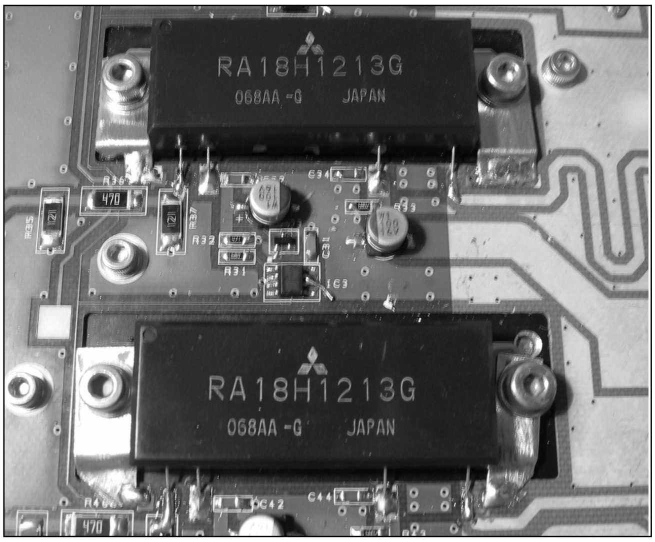


Fig 11.81: The additional grounding on modules 3 and 4

stopped the oscillation; this was the clue. The mounting screws for the PCB to the heatsink were loosened on one side and re-tightened with the PCB pressed firmly against the heatsink. The amplifier then drew 15A and no oscillation.

Next the amplifier was fed with 23cm RF and the output measured, with 12.5V at the battery terminals 87W output was measured. To get full output the full 13.8V supply would be required. To do this, the power supply wiring had to be tidied up; it is surprising how much volt drop there can be across a thick piece of cable when it is carrying the 40 or so amps drawn at full power.

The amplifier was then fitted into its box to tidy up the wiring and fit the final input and output connections to the changeover relays. I spent some time researching the possibilities for an enclosure; my first thought was diecast aluminium with a fully sealed lid. However, having seen the price and weight of suitably sized enclosures, I realised there had to be a better option. While searching the Farnell web site I found that there is a range of steel terminal enclosures [28] that are fully sealed at a reasonable price and weighing less than the equivalent aluminium enclosure (the aluminium enclosures are built to support a tank). The only problem with this enclosure is cutting out a 200mm x 250mm hole to fit the power amplifier. The result is shown in Fig 11.82. Final testing with everything tidied up gave nearly 100W output.

In conclusion, the PA1.3-100 kit is well designed and easy to build by anyone with a reasonable level of experience. The amplifier performs well and does exactly what it says on the tin.

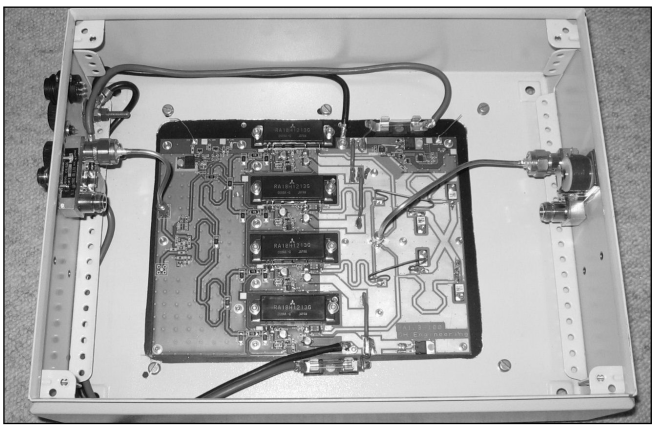


Fig 11.82: The amplifier mounted in its metal enclosure with a sequencer and preamplifier

The main thing to note when using the amplifier is that a 45A 13.8V supply is difficult to provide at the top of a mast but I think the advantages of mounting the amplifier near the antenna for contest working outweigh the problems involved. If I were using the amplifier for a permanent installation this would need careful thought.

Surplus High Power Amplifiers for 2.3GHz

Activity on the 2.3GHz band was at a low level until mobile phone and MMDS technology provided amateurs with a source of surplus solid state power amplifiers that can be modified for use in the amateur allocation at 2.3GHz and producing output power levels of 50W to more than 200W. Many of these amplifiers were originally used in base stations in the digital cellular radio bands at 1.8GHz, 1.9GHz and 2.1GHz. In addition to the surplus cellular amplifiers, there has recently been a quantity of high power Spectrian 2.3 to 2.4GHz MMDS solid state amplifiers and individual power amplifier modules offered on eBay.

GSM digital cellular radio systems use a form of digital signal modulation known as Gaussian Medium frequency Shift Keying (GMSK). This form of modulation does not require high amplifier linearity. A basic GSM base station consists of a radio transceiver providing a transmit radio carrier with minimum eight timeslots, each of which provides a switched voice circuit. Therefore each base station provides for up to eight voice calls on a single radio frequency. If a base station needs more capacity than eight voice channels, one (or more) transceivers are added on another radio frequency.

GSM base station transmitters are sometimes combined at a moderate power level, using passive combiners, in order to use a common antenna. Alternatively, combining can be performed at much lower power levels, with the combined multiplex of two or more radio channels then amplified up to the required power level. This then requires very linear solid state amplifiers. First generation combining amplifiers used bipolar power transistors, operating from 26V, where they would provide a peak output power of 50 to 100W (at 960MHz) and perhaps 50W at 1800MHz. American PCS1900 (GSM) systems operate at 1.9GHz, but are otherwise similar in output power.

Radio Frequency Metal Oxide Semiconductor Field Effect Transistors (RF MOSFETs), or similar device technology, has largely replaced the bipolar transistor amplifier technology, with the advantage of better linearity, efficiency and reliability. Consequently, large numbers of the earlier bipolar amplifiers have started to appear on the surplus market. While being less than perfect for the cellular phone operators, if they can be converted to 2.3GHz, they can provide a useful boost in transmit signal levels for the amateur operator.

Several years ago Günter Köllner, DL4MEA, published details on his web page showing how to modify a Siemens GSM 1800 or PCS 1900 amplifier module to produce up to 50W output on 2.3GHz. Siemens amplifiers were available in GSM 1800 and PCS1900 versions, but were basically to a similar design, consisting of a Motorola or Philips hybrid driver module, providing typically 15W drive to a MA/COM PHI 1920-60 bipolar output stage from about 10mW input. Unfortunately, the hybrid modules have proved unusable at 2.3GHz but can be modified to give up to 15W output at 1.3GHz. See [29] for details of modifying these amplifier modules for 1.3GHz operation.

Modifying these particular Siemens bipolar amplifiers requires no more than a minor change to the input matching arrangement and the addition of a larger overlaid matching section at the transistor output. This can be seen quite clearly in Fig 11.83. It is necessary to re-position the input and output coaxial leads and connectors with this arrangement. Full details of the modifications and tuning are shown on Gunter's web page [30].

My own Siemens bipolar amplifier provides about 25W output for 8W drive at 2320MHz. A similar amplifier, modified by WW2R/G4FRE, provides 40W output for 10W input at 2304MHz. Both of these amplifiers are gain limited by their respective driver amplifiers. Greater output power can be achieved with more drive as well as by combing several similar amplifiers.

The introduction of Third Generation (3G) cellular systems, operating around 2.1GHz, has required base station transmitters with excellent linearity. This high linearity has been achieved by using RF MOSFET transistors operating from a 26V supply. 3G systems use wideband spread-spectrum modulation (WCDMA), which, when several channels are in use, has a very large peak-to-average signal level ratio (known as a large crest factor), necessitating base station amplifiers with a peak power rating of up to 10 times the mean power rating. Consequently, many 3G base stations use amplifiers with a CW rating of over 100W for 10W input.

Motorola was one of several semiconductor manufacturers producing RF MOSFET devices. They recently sold their semiconductor business, but many of the surplus amplifiers use Motorola parts. Motorola produced a very useful MRF21120 RF MOSFET power amplifier reference design for W-CDMA use that has been incorporated into some 3G base station designs that are capable of 120W CW output at 2.17GHz. The good news is that similar modules have recently appeared on the surplus market. Even better news is that they will work at 2.3GHz with little or no modification and they can provide over 100W output with about 10dB gain, which is more than enough for most amateur terrestrial

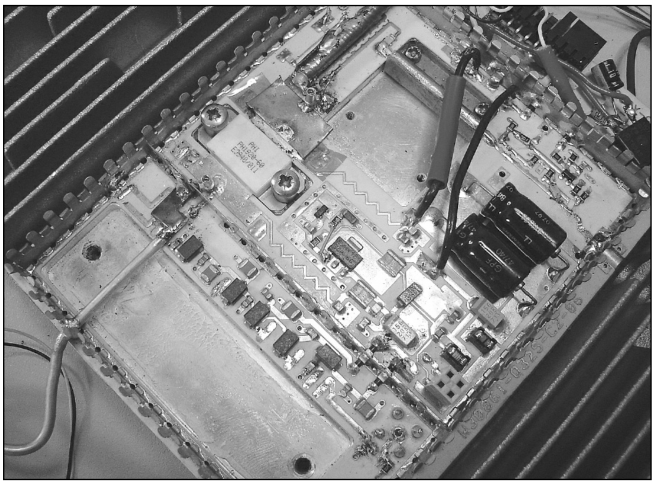


Fig 11.83: A Seimens GSM1900 bipolar amplifier modified to give 40W output at 2.3GHz

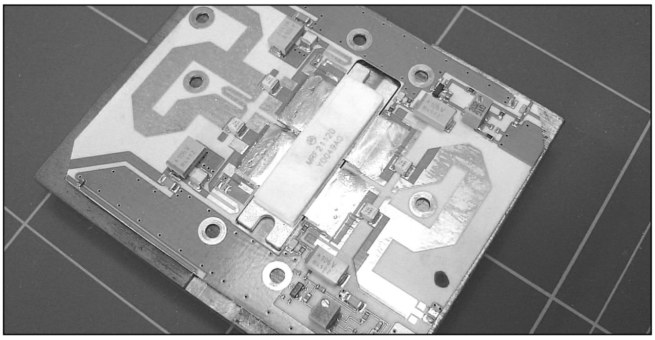


Fig 11.84: A W-CDMA 120W module that can be used at 2.3GHz with minimum modification

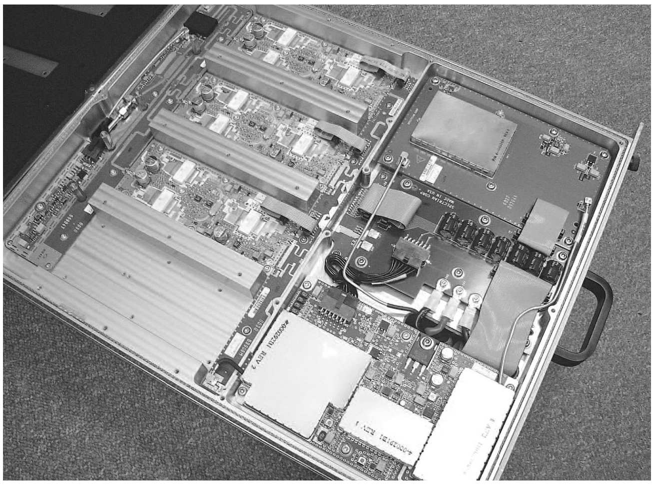


Fig 11.85: A Spectrian 200W MMDS amplifier suitable for use at 2.3GHz

operation, and probably enough for EME with a 3m, or larger, dish. Combing two or more such amplifier modules will allow in excess of 200W output. **Fig 11.84** shows a typical RF MOSFET W-CDMA module.

A number of Spectrian SCPA1063W 200W Multi-point Multi-channel Distribution System (MMDS) amplifiers and modules have recently appeared on eBay. These amplifiers are, apparently, designed to work in the USA MMDS allocations between 2.3 and 2.4GHz, and have excellent linearity due to the use of X(MRF) 286 RF MOSFETS. An individual module consists of a single X286 driving a push-pull pair of the same devices. Output is usually between 70 and 80W with about 18dB gain at 1dB saturated output. A complete amplifier includes a driver amplifier with about 15W output and requiring about 10mW drive. This is ideal for connecting to the transmit output of a low-level transverter. **Fig 11.85** shows the inside of one of these amplifier units where three individual output modules can be seen together with the splitter and combiner arrangement.

The NR6CA web page [31] is the usual first stop for details of how to use the Spectrian amplifiers on 2.3GHz. Several of the Spectrian amplifiers and modules are known to be in use in the UK and elsewhere in Europe. Christophe, ON4IY [32], show details of his unit and mounting arrangements on his web page.

GaAsFET Amplifier for 3cm

The power amplifier for 3cm described here was developed by Peter Vogl, DL1RQ, [33] 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 [33] 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 11.86**, is similar to DL1RQ's two stage 5.7GHz power amplifier, published in [34] and [35], 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 (2.0 mm x 1.25 mm).

Research carried out only recently as part of a specialist project [36] 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 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 printed circuit board layout (**Fig 11.87**) is etched onto an RT/Duroid 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

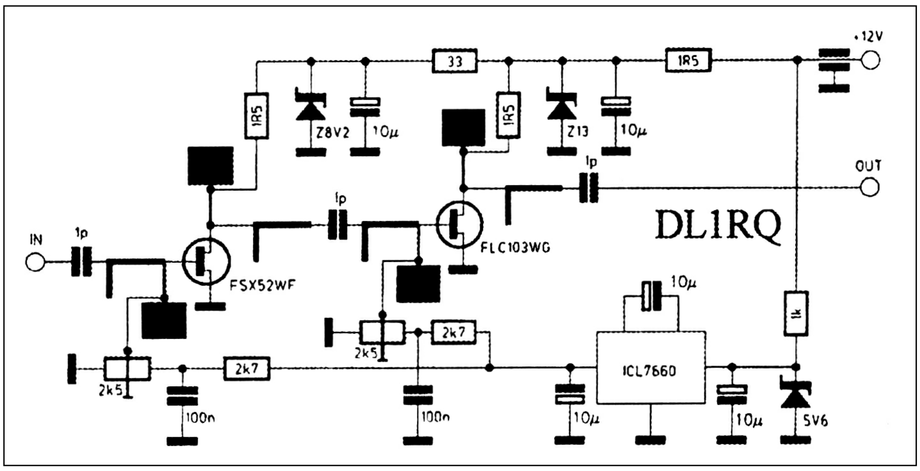


Fig 11.86: Circuit diagram of the 1W GaAsFET amplifier for 3cm

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 11.88**). In order to guarantee good heat transfer between the copper plate and the milled aluminium housing (**Fig 11.89**), 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).

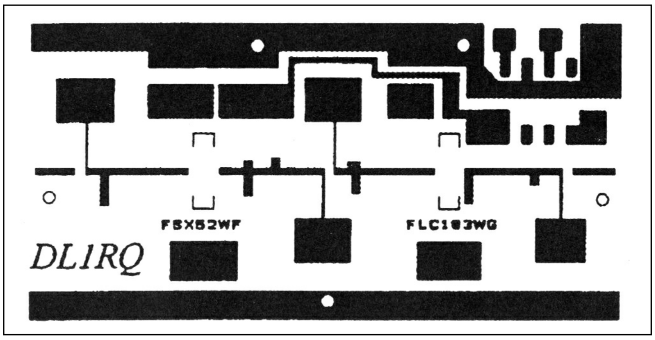


Fig 11.87: PCB layout for the 1W GaAsFET amplifier for 3cm

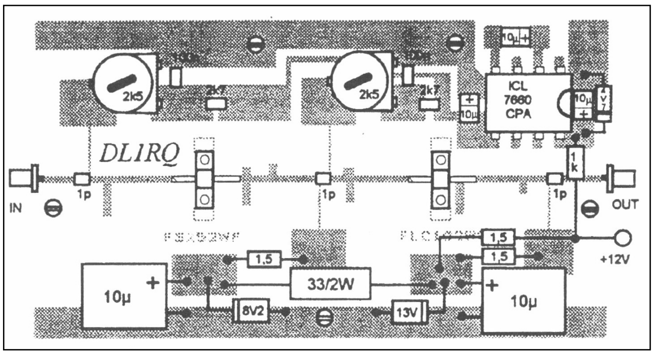


Fig 11.88: Component layout for the 1W GaAsFET amplifier for 3cm

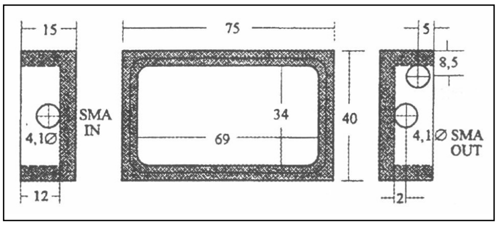


Fig 11.89: Dimensions of a milled housing for the 1W, 3cm GaAsFET amplifier

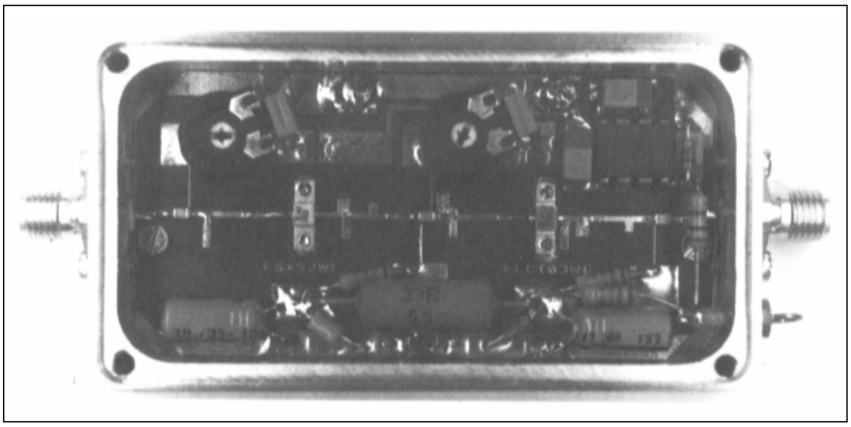


Fig 11.90: The 1W GaAsFET amplifier for 3cm

The partly assembled board is now fitted into the suitably prepared aluminium housing and screwed down by five M2 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°C. Each milled groove is then pre tinned, using low temperature solder with a melting temperature of 140°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 11.90**).

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 - 4mm^2, 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 11.91 and 11.92**) with the readings from a typical power amplifier (**Figs 11.93 - 95**) makes clear how successful the project is in practice.

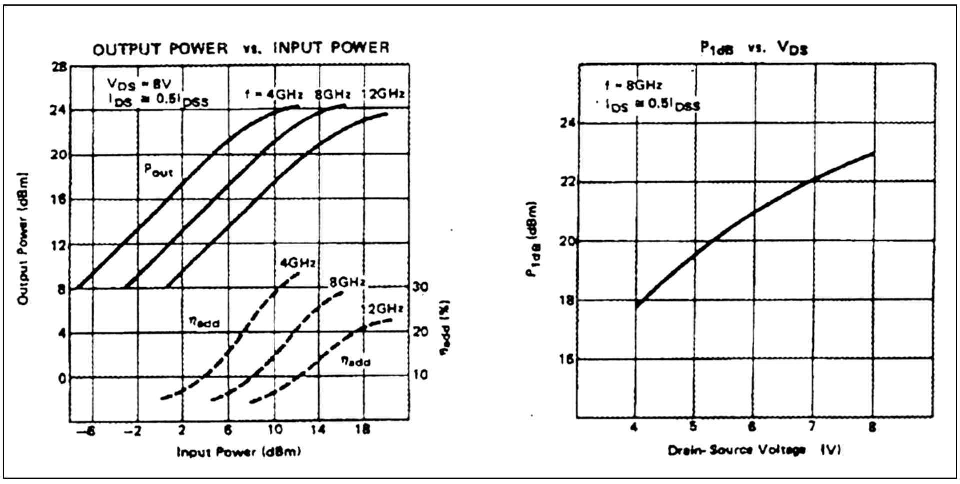


Fig 11.91: Data for the Fujitsu FSX53WF used in the 1W GaAsFET amplifier for 3cm

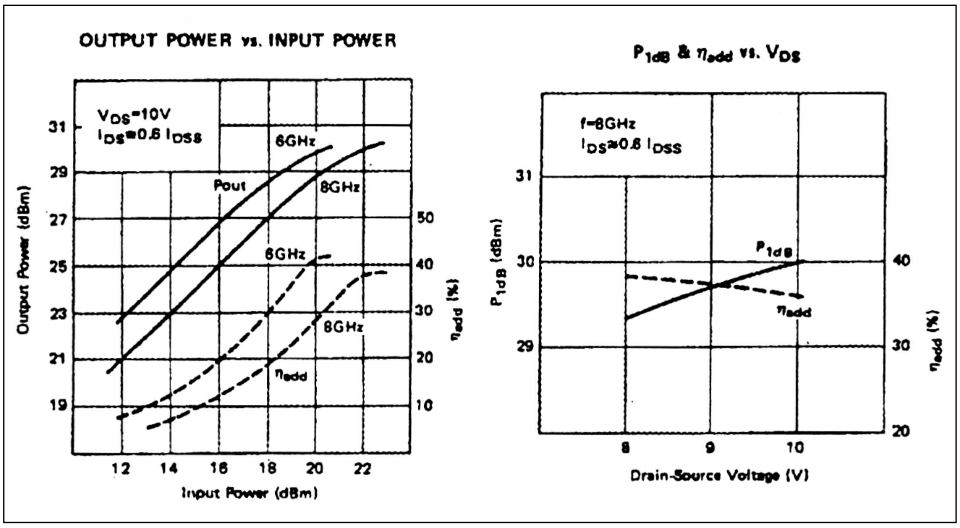


Fig 11.92: Data for the Fujitsu FSX53WF used in the 1W GaAsFET amplifier for 3cm

76GHz Amplifier

This design [37] is part of a series of articles by Sigurd Werner, DL9MFV, published in *VHF Communications* magazine. It describes an amplifier that uses two MMICs (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 MMICs for this frequency band.

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 [38].

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 [39] and [40].

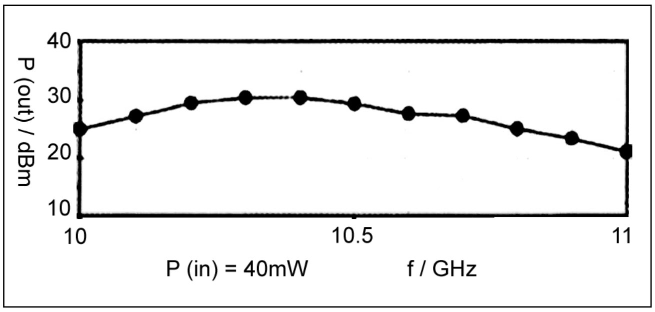


Fig 11.93: Power bandwidth of the 1W GaAsFET amplifier for 3cm

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 11.96**). 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

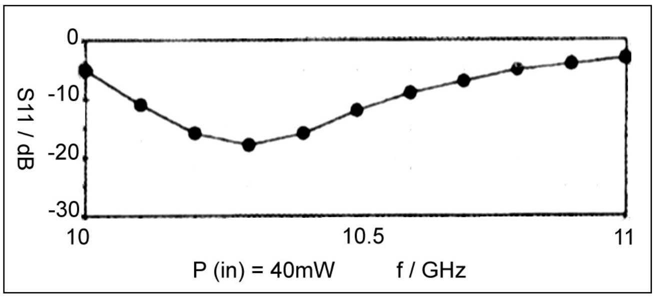


Fig 11.94: Input matching of the 1W GaAsFET amplifier for 3cm

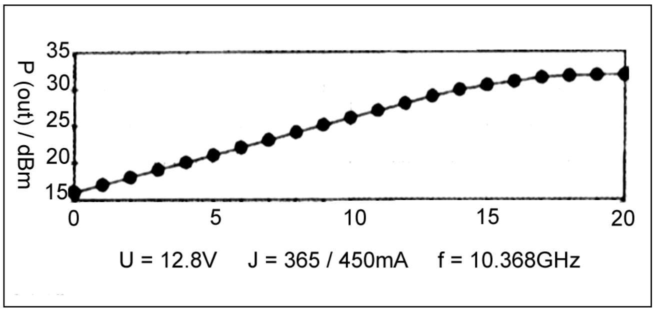


Fig 11.95: Linearity of the 1W GaAsFET amplifier for 3cm

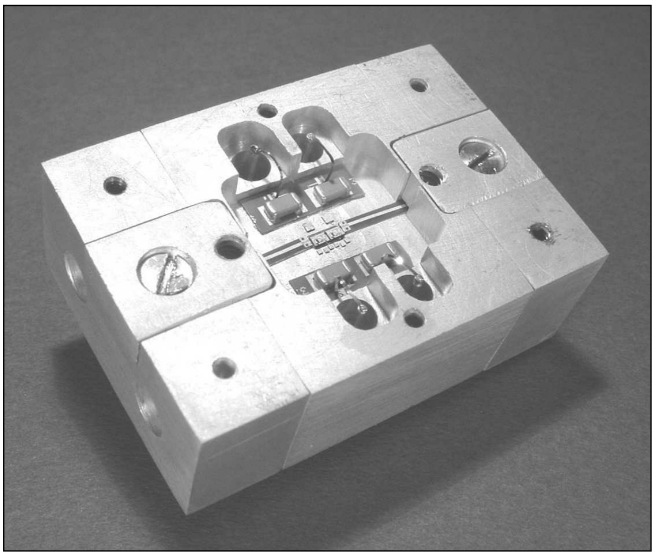


Fig 11.96: Picture of the completed 76GHz amplifier

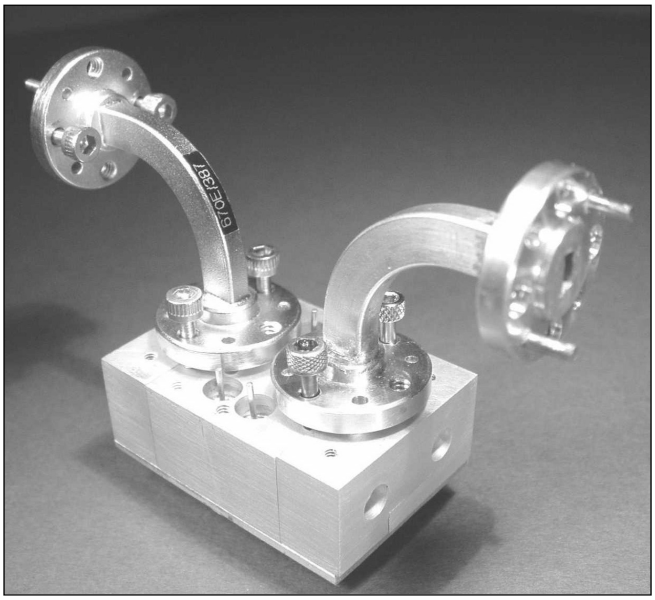


Fig 11.97: Picture of the completed 76GHz amplifier with WR-12 waveguide connections fitted

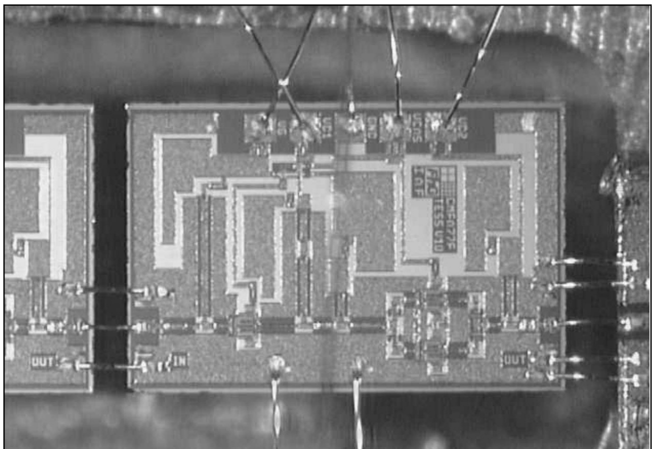


Fig 11.98: The 1AF-MPA7710 chip, used in the 76GHz amplifier, magnified about 120 times

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 λ/8 into the waveguide. This construction technique is described in more detail in [41]. 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 11.97**). The connections between chips and substrate (or capacitors) were created using wedge-wedge bond technology [41].

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 11.98**). 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 mismatching 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 [42]). The input power during matching amounted to -13dBm.

After a laborious sequence of nine 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 [39]. 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 = 76,032GHz) is shown in **Fig 11.99**. 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,

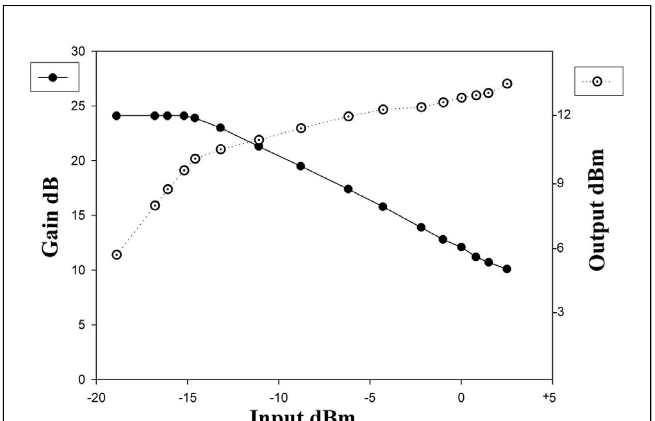


Fig 11.99: Transfer characteristics of the 76GHz amplifier at 76,032MHz

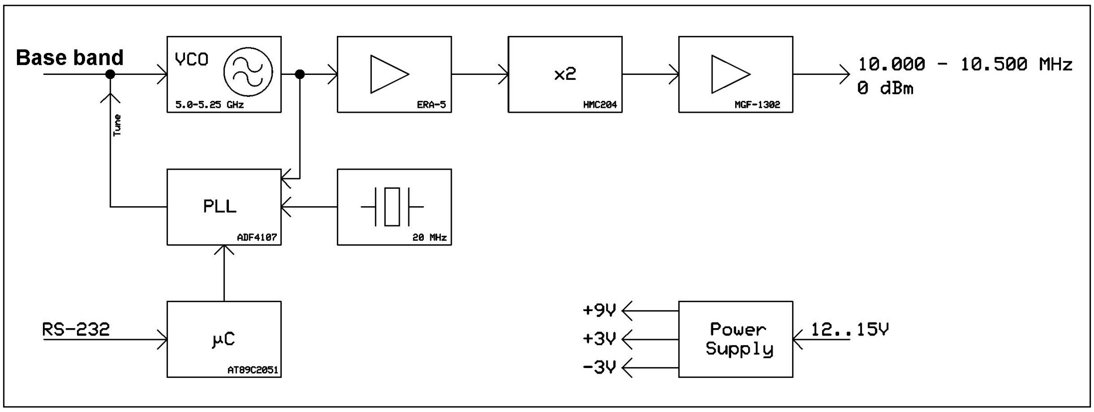


Fig 11.100: Block diagram of the 10GHz ATV transmitter

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 [43].

ATV TRANSMITTER WITH PLL FOR 10GHZ

The ATV mode is very popular because many ATV repeaters have become active in recent years. This has meant that the 3cm band is becoming more and more important, in addition to links; it is increasingly used for operator input and output. For more on ATV, see the chapter on imaging techniques (on the free CD).

With this in mind, a modern, frequency-stabilised ATV transmitter has been developed by Alexander Meier, DG6RBP, the transmit frequency can be simply and accurately selected at any point on the entire 10GHz amateur radio band. This makes it suitable for use as a transmitter in a repeater or for direct operation.

A transmitter in an ATV repeater, eg as operator output or for a link, is subjected to temperature variations and is in operation for long periods of time. So the transmit frequency should change as little as possible. For direct operation, on the other hand, simple frequency adjustment is important, particularly if it is also used to work with repeaters. The PLL stabilised transmitter has a few advantages here: the transmit frequency can be selected simply and easily anywhere in the 3cm. band, by means of a microcontroller, and frequency deviations are prevented by the Phase Locked Loop (PLL).

The transmitter with a PLL is based on Voltage Controlled Oscillators (VCOs), that have recently become available at reasonable prices up to and beyond 10GHz [44]. You won't find a suitable VCO that will cover the entire 3cm band with an acceptable tuning rate. But there is an option to use a VCO of 5.00 to 5.25GHz then doubling and amplifying the output signal. There is, however, a disadvantage of this approach; compared with free running oscillator or DRO transmitters, the circuit complexity is considerably greater.

Fig 11.100 shows the block diagram of the project. A VCO of 5.00 to 5.25GHz is stabilised using a PLL and after a passive doubler (x2) covers the entire 3cm band from 10.00 to 10.50GHz. The losses that this generates are balanced out again using two amplifier stages with an output of approximately 1mW at 10GHz. This is sufficient to drive an additional power amplifier, for example a small TWT or a 200mW FET amplifier.

The PLL is controlled by means of an 8051 microcontroller from ATMEL. The transmit frequency can thus be conveniently selected anywhere on the 3cm band via the RS232 interface from a PC or through a separate control module. It is not obligatory to use the interface for fixed frequency operation in a repeater because the transmitter starts at a fixed frequency when it is switched on (pre-programmed in the microcontroller). An LED signals when the PLL is locked. If this is not the case, the microcontroller switches the amplifier stages off and sends an acknowledgment via the RS232 interface.

As an option, the ATV transmitter can also monitor an externally connected power amplifier (PA); if the voltage at the monitor

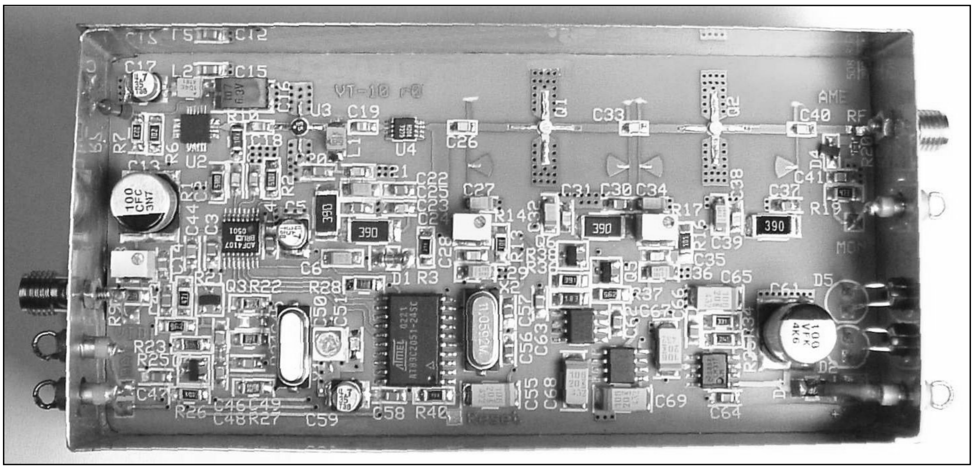


Fig 11.101: The completed 10GHz ATV transmitter

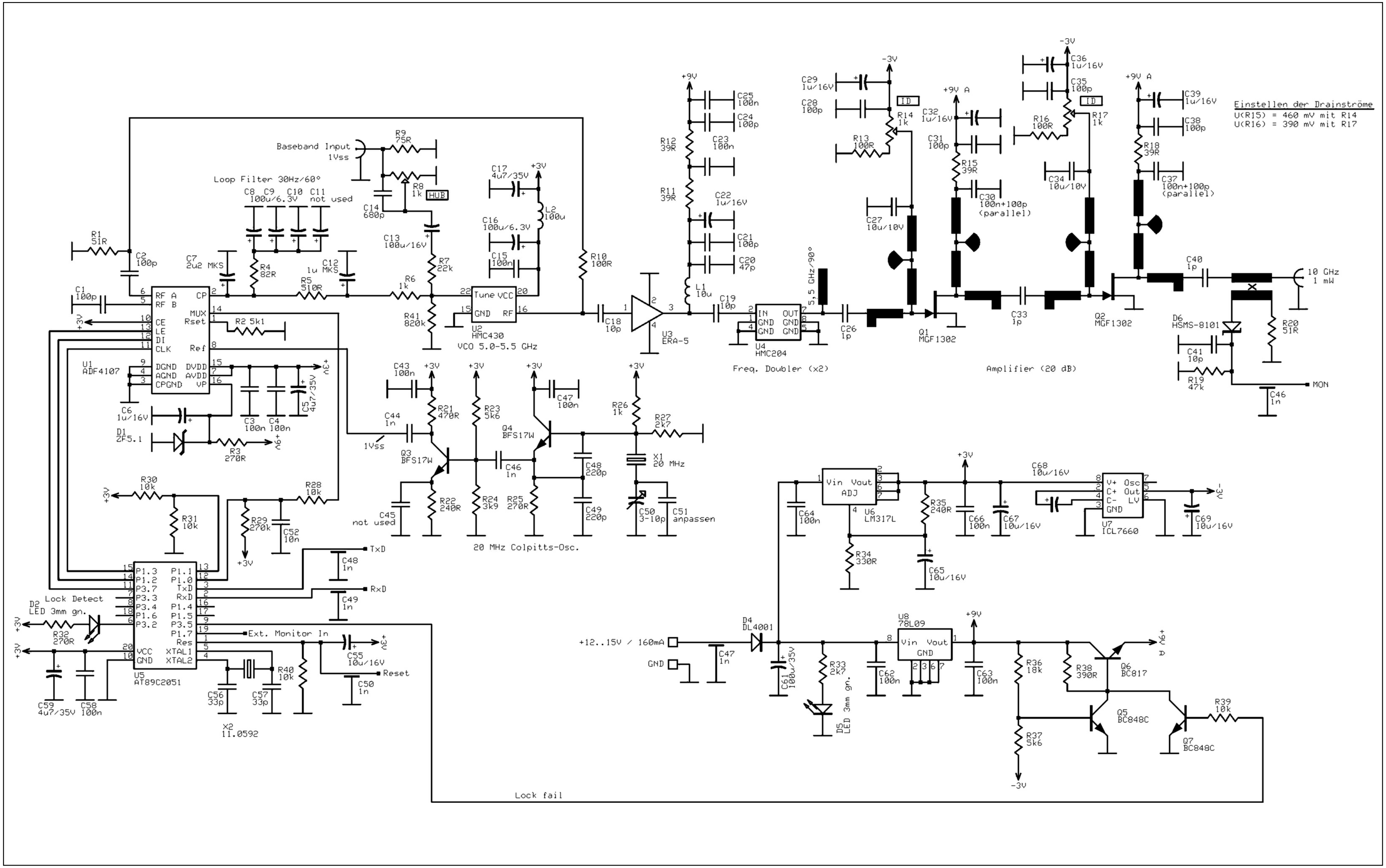


Fig 11.102: Circuit diagram of the 10GHz ATV transmitter

Resistors		
R1, R20	51R	1206
R9	75R	1206
R4	82R	1206
R10, R13, R16	100R	1206
R22, R35	240R	1206
R3, R25, R32	270R	1206
R34	330R	1206
R38	390R	1206
R21	470R	1206
R5 5	10R	1206
R6, R26	lk	1206
R27, R33	2k7	1206
R24	3k9	1206
R2	5k1	1206
R23, R37	5k6	1206
R28, R30, R31, R39, R40	10k	1206
R36	18k	1206
R7	22k	1206
R19	47k	1206
R29	270k	1206
R41	820k	1/4W
R11, R12, R15, R18	39R	2510
R8, R14, R17	1k	SMD Trimmer
Capacitors		
C26, C33, C40	IpF	0805
C18, C19, C41	10pF	0805
C51	18pF	0805
C56, C57	33pF	0805
C20	47pF	0805
C1, C2, C21, C24, C28, C30, C31, C35, C37, C38	100pF	0805
C48, C49	220pF	0805
C14	680pF	0805
C44, C46	1 nF	0805
C52	10nF	0805
C3, C4, C15, C23, C25, C30, C37,C43, C47, C58, C62, C63, C64, C66	100nF	0805

C6, C22, C29, C32, C36, C39	lμF/16V	SMD Tantalum
C12	1 μF	MKS-2
C7	2.2μF	MKS-2
C5, C17, C59	4.7μF/35V	SMD Electrolytic
C27, C34	10μF/10V	SMD Tantalum
C55, C65, C67-C69	10μF/16V	SMD Tantalum
C8-C10, C16	100μF/6.3V	SMD Tantalum
C13	100μF/16V	SMD Electrolytic
C61	100μF35V	SMD Electrolytic
C50	C-Trimmer 3 - 10pF	SMD
C46-C50	1nF	Feedthrough
Inductors		
L1	10μH SIMID	1210
L2	100μH, SIMID	1210
Crystals		
Xl	20MHz,	SMD CL = 16pF
X2	11.0592MHz	SMD
Semiconductors		
D1	ZF5.1	
D4	DL4001	
D2, D5	3mm green LED	low current
D6	HSMS-8101	
Ql, Q2	MGF 1302	
Q3, Q4	BFS 17W	
Q5, Q7	BC 848C	
Q6	BC 817	
U1	ADF4107 BRU	
U2	HMC 430 LP4	
U3	ERA-5	
U4	HMC 204 MS8G	
U5	AT 89C2051	SMD, programmed
U6	LM 317L	SMD
U7	ICL 7660	SMD
U8	78L05	SMD
Hardware		
1 x	tinplate housing	111 x 55 x 30mm
1 x	PCB DG RBP-VT 10	
2 x	SMA socket	

Table 11.10: Components list for the 10GHz ATV transmitter

output of the external PA falls below a selected threshold value, the microcontroller reports this via the RS-232 interface. A picture of the completed ATV transmitter is shown in **Fig 11.101**.

Circuit description

The circuit diagram of the transmitter is shown in **Fig 11.102** and **Table 11.10** shows the parts list.. The VCO, U2, can be tuned at pin 22, using a tuning voltage of between 0 and 10V giving between 5.0 and 5.5GHz at Pin 22. Since only a frequency range of 5.0 to 5.25GHz is required before doubling, a tuning voltage of between 0 and 5V is satisfactory. The operating voltage is fed to the VCO via a low pass filter (L2 and C15-C17) in order to suppress noise and other residual frequencies in the supply voltage.

The VCO output is approximately 1mW, one part of this is fed to the PLL via the resistor, R10. The majority goes to the amplifier, U3. This gain block increases the power to a few mW to feed the passive doubler, U4. A stub cable at the output of the doubler suppresses the first harmonic at 5GHz. The second harmonic at 10GHz (approximately -20dBm) corresponds to the desired frequency and is amplified, using the Mitsubishi FETs Q1 and Q2, to an output of approximately 0dBm (1mW). The quiescent currents of the amplifier are set by means of the trimmer potentiometers, R14 and R17. If required, the printed circuit board provides the option of a monitor circuit with a directional coupler and a Schottky diode (R19, R20, C41, C46, D4).

The PLL module (U1) from Analog Devices has the output frequency of the VCO (5.00 - 5.25GHz) fed into it via pin 6, and the reference frequency (20MHz) via pin 8. The phase comparison is carried out at 100kHz, so that there is a more than satisfactorily small step size of 0.2MHz in the 10GHz band. The loop filter of the PLL (C7, R4, C8 to C10, R5, C12) has been designed for the transmission of video signals. The loop bandwidth is 30Hz. The low frequency of the phase comparison makes it possible to use practical component values. High-quality foil capacitors must be used for C7 and C12, and good tantalum capacitors for C8 to C10.

The tuning voltage is combined with the modulation signal (base band) via R6 and R7 before the VCO. The trimmer, R8, serves to set the frequency deviation. The capacitor, C14, is used to provide for a slight increase in the high modulation frequencies.

A simple quartz oscillator in a Colpitts circuit (Q4) is used as a reference oscillator. The frequency can be varied somewhat using the trimmer capacitor, C50. A subsequent amplifier stage (Q3) serves mainly as a buffer. If required, the capacitor, C45, could be used to increase the AC voltage amplification of the stage.

The PLL's programming is carried out by an 8051 microcontroller (U5) from ATMEL via pins 11, 14 and 15. The controller operates with a quartz crystal frequency of 11.0592MHz to generate a baud rate of 9600 Baud. The lock-detect signal of the PLL is compared with a pre-set voltage, using an integrated comparator (Pin 12, 13). If the current at pin 12 exceeds the reference voltage of 2.5V, the controller recognises that the PLL has locked and switches the amplifier stages on via pin 9. In addition, an LED signals to pin 6 that the PLL has locked.

For the optional supervision of an external output amplifier stage, the corresponding monitor signal is compared using a comparator (external circuit as and when required). The output signal of the comparator is then fed to the microcontroller via pin 19. A voltage level of 0V means the PA has failed, while a level of 5V means it is operating. The status of pin 19 is continuously monitored in the controller, and should any change occur a message is transmitted via the RS232 interface.

The supply voltage for the module (12 to 15 Volts) is fed in through a reverse battery protection diode (D4). The fixed voltage regulator (U8) uses it to generate the operating voltage for the amplifier stages (9V). This is applied, through the protective circuit of Q5-Q7, only if the PLL is locked and the negative gate voltage is correct.

An easily adjustable voltage regulator, type LM 317 (U6) is used for the power supply for both the PLL and the VCO (3V). The charge pump (U7) uses this to generate a negative voltage for the gates of the amplifier stages.

Assembly and tuning

The entire circuit is built on a 109 mm x 54 mm x 0.51 mm printed circuit board. The inexpensive RO-40003C from Rogers is used as the printed circuit board material. To make assembly easier, particularly for the VCO (U2) and the doubler (U4), the printed circuit board is through-hole plated and coated with a solder resist.

After a coating of solder paste has been applied, the VCO is melted on, using a reflow soldering unit or a hot air station. This component cannot be mounted using a soldering iron because of the VCO package connections are on the underside. This is the same for the doubler (U4), it has an earth side on its underside, which must be soldered correctly.

Finally, all other components apart from Q6 are mounted, with no particular instructions. A soldering iron can be used here. Good eyes and steady hands are required for the PLL (U1) when the components are mounted by hand. The wired resistor, R41 is soldered ‘quick as a flash’ by hand, depending on the earth (eg housing wall). The PCB layout is shown in **Figs 11.103** and **11.104** and the component layout is in **Figs 11.105 and 11.106** (all four of these drawings are in the Appendix).

Prepare the tinplate housing by drilling the appropriate holes. Clean the flux residue off of the printed circuit board and fit it into the housing, soldering it all around the underside. Then the printed circuit board is cleaned again.

Once the printed circuit board has been connected to the feedthrough capacitors, the first function test can be carried out. When the supply voltage has been applied, the voltage on the gates of, Q1 and Q2, is initially set to -3V, using the trimmers, R14 and R17. Switch off and fit transistor Q6. Switch on again and check the important parts of the circuit. The most important elements are the operating voltages (3V, -3V, 9V, 5.1V at pin 16 of U1) and the 20MHz quartz oscillator. Next the tuning begins.

When the supply voltage is applied, the module starts at the default, pre-programmed, frequency. Now the quiescent currents of the amplifier stages are set. First, the voltage is measured across resistor R15 and adjusted to 460mV using the trimmer R14. The voltage across R18 is set to 390mV using R17. A spectrum analyser is used to check whether the PLL has locked correctly. The output will still be somewhat below 1mW. Finally, the module is connected to a PC via the RS232 interface and the frequency is altered from 10.000 to 10.500GHz in steps of 50MHz and check that the PLL has locked correctly. The stubs at the input and output of the transistors, Q1 and Q2, are shortened slightly with a scalpel to find the maximum output in the middle of the frequency range or at any preferred frequency. The frequency shift of the transmitter is set to ±3.5MHz (as per DL2CH ATV standard [45]) using a spectrum analyser and the carrier minimum method. A 15MHz signal is applied at the base band input, with an amplitude of 1Vpp. Increase the shift until the first time that the carrier frequency reaches a minimum. The frequency deviation can also be set to a higher value if required.

After a final function test with a test image and the transmitter is ready for operation.

Programming

The programming of the default frequency of the module, ie the frequency at which the module starts when the supply voltage is applied, is pre-set once and for all in the software of the microcontroller (U5). If the transmitter were used in an ATV repeater, this would be the output frequency. The transmitter could then be used without utilising the RS232 interface.

Otherwise, the output frequency can be changed at any time via the RS232 interface in steps of 1MHz. The PLL would allow for smaller steps, but these are not expedient for ATV applications. To set the frequency, we can use a PC with a USB programming adapter [46], another microcontroller with an RS232 interface, or a frequency input module developed individually for the transmitter [47]. Direct connection to the RS232 interface of a PC is possible, using a level converter (0/5V, e.g. MAX232).

Controller to ATV transmitter	F10xxx	New transmit frequency 10xxx = frequency in MHz
Controller to ATV transmitter	S	Request transmitter status
ATV transmitter to Controller	P	Power on
ATV transmitter to Controller	L	PLL locked
ATV transmitter to Controller	U	PLL un-locked
ATV transmitter to Controller	A0	External monitor signal fail
ATV transmitter to Controller	A1	External monitor signal OK
ATV transmitter to Controller	E	Error
ATV transmitter to Controller	ERR	Error
ATV transmitter to Controller	OK	New frequency confirmed
ATV transmitter to Controller	F10xxxyAz	Transmitter status 10xxx = frequency in MHz; y = L,U (locked, un-locked); z = 0,1 (external monitor 0,1)

Table 11.11: Programming control by the RS-232 port on the ATV transmitter

Table 11.12: Technical data for ATV transmitter

Frequency range	10.000 to 10.500GHz
Steps	1MHz
Frequency stabilisation	PLL
Long term stability	±5ppm / year
Temperature stability	±50ppm (+10 - +40°C)
Modulation input	Baseband IV_{pp} (positive input gives positive modulation)
Modulation type	FM
Modulation depth	±3.5MHz for I V_{pp} (internally adjusted with reserve)
Modulation freq. range	30Hz to >7MHz
Interface	RS-232, 0/3V signalling, 9600 baud, 8 data bits, 1 stop bit, no parity
Interface cable length	20m
RF output impedance	50 ohms
Typical output	1mW / 0dBm (±3dB)
Typical harmonic output	<-40dBc (5 - 5.25GHz), <-40dBc (20 - 21GHz)
Other harmonics	<-50dBc
Supply voltage	12 to 15V DC
Supply current	Approximately 160mA
Temperature range	+10 to +40°C
Connectors	SMA socket for RF output and baseband input. Feedthrough capacitors for RS-232
Indicators	Power - green LED, Locked - green LED
Case	Tinplate housing
Size	111 x 55 x 30mm approximately

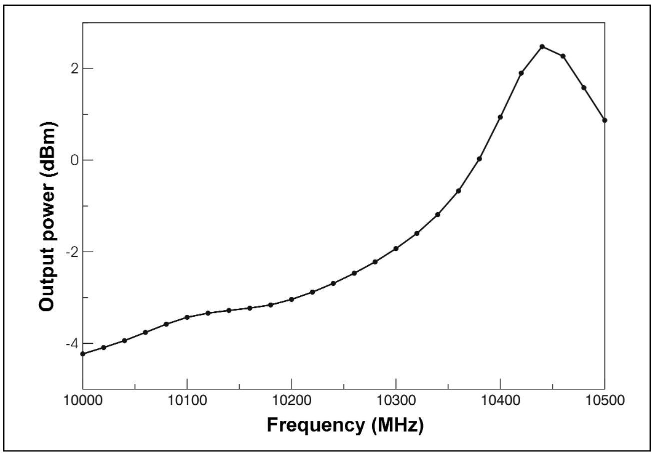


Fig 11.107: Graph of output power against frequency for the ATV transmitter

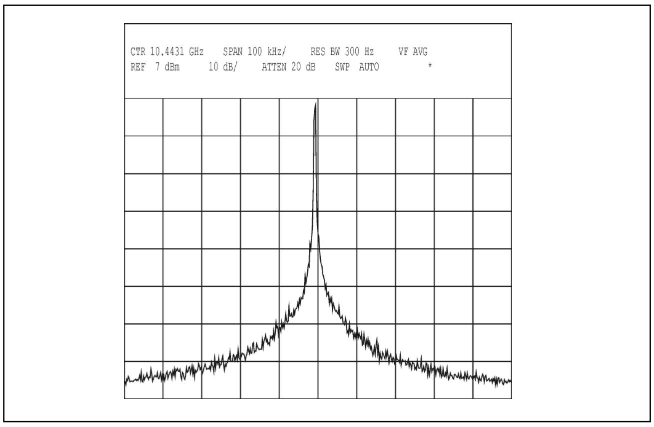


Fig 11.108: Output spectrum of the ATV Tx with no modulation

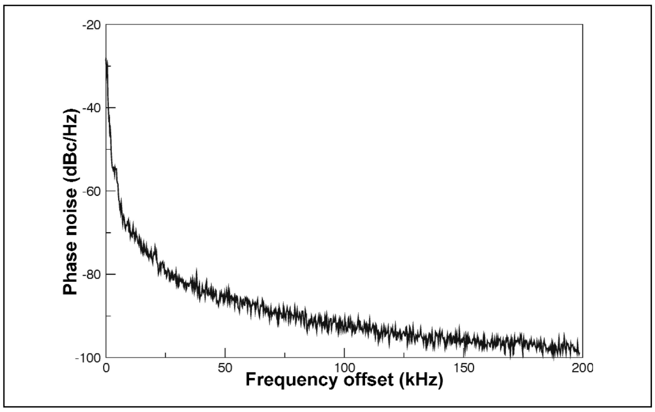


Fig 11.109: Phase noise performance of the ATV transmitter

Table 11.11 lists the interface commands. Thus, when the supply voltage is applied, the transmitter transmits a 'P' for 'Power On'. As soon as the PLL is locked, the transmitter reports this with an 'L'. If we now wish, for example, to set the frequency to 10.450MHz, we transmit the character string 'F10450' to the transmitter. The new frequency is confirmed and selected using 'OK'. As the PLL is not locked during the frequency change, the transmitter reports this with 'U'. As soon as the PLL is locked, we receive an 'L' as a response.

We can check the transmitter status by sending the character 'S' to the transmitter. For example, if the transmitter answers 'F10200LA1', this means: frequency 10200MHz, PLL locked (L) and monitor signal for external amplifier stage OK (A1).

The technical data of the ATV transmitter are summarised in Table 11.12. As the transmitter modulates the output frequency towards higher frequencies for positive voltages on the base band input, the video signal must be inverted before being applied to the base band input to obtain the inverted image usually required for ATV repeater outputs.

The output power at different frequencies can be seen in Fig 11.107. The frequency of the maximum power is determined during tuning. Tuning was carried out here at 10.440GHz. Fig 11.108 shows the output spectrum without modulation and the phase noise of the transmitter is shown in Fig 11.109.

TRANSVERTERS

The most popular method of becoming active on the microwave bands today is 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 these transverters now produce them as kits and ready built units, these are all listed in the bibliography. Sam Jewell has reviewed the DB6NT 23cm and 3cm transverters.

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 later in this chapter.

Kuhne Electronics MKU13 G3 1.3GHz Transverter Review by Sam Jewell, G4DDK

DB6NT, in conjunction with DF9LN, published a 1.3GHz transverter design in the European magazine *Dubus* [48] 4/1992. This design was very popular and when DB6NT formed Kuhne Electronics the design was made available commercially as the MKU 13 G. It was later upgraded to the G2. The obsolescence of several key parts, together with the usual commercial imperative to update designs, has resulted in the release of the new G3 transverter. An external photograph of the new G3 is shown in **Fig 11.110**.

Although the G2 was sold as both a kit and a ready-built module, the G3 is only being made available in ready-built form. Many of the small surface mount components used in the G3 now make it impractical for most home constructors to build.

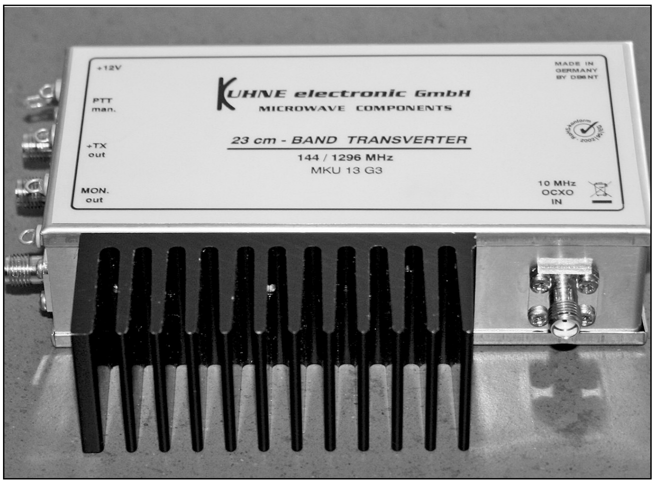


Fig 11.110: Outside of the G3 1.3GHz transverter

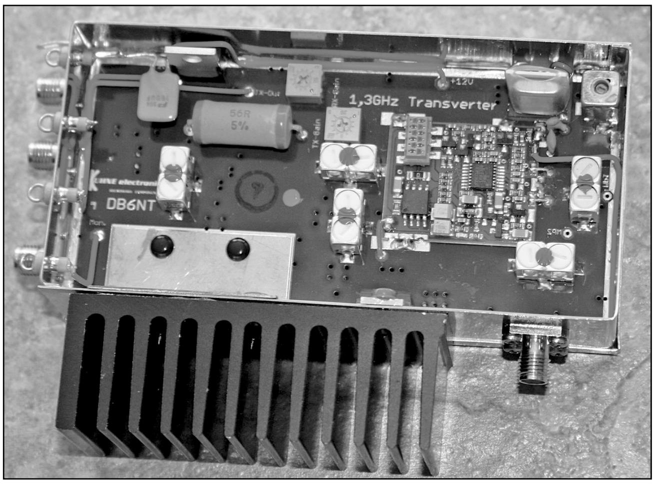


Fig 11.111: Top side of the G3 unit. The frequency locking board, helical filter, crystal (with heater) and the 50-ohm IF load can all be seen

The most significant features of the new transverter are more output power and the addition of an external 10MHz local oscillator locking input. This latter feature will be much appreciated by those who fought the thermally-induced frequency drift that the older G2 was noted for. It is not necessary to use the locking facility if it is not required. **Fig 11.111** shows the top side of the PCB with the frequency locking board, helical filters and gain setting controls.

Circuit description

In common with the previous design from DB6NT, the G3 uses GaAsFETs in many of the critical RF stages. An NE32584 GaAsFET low noise amplifier (LNA) provides a claimed noise figure of under 1dB. This is followed by a two-pole helical filter into an integrated circuit gain block ERA3. A second helical filter feeds into a high level (+17dBm) ADE 17 diode ring mixer. Following the 144MHz diplexer a miniature relay switches the receiver IF output via an adjustable attenuator to the common IF connector. The gain is adjustable over a large range to allow optimisation of the receive system according to cable lengths etc.

DB6NT told me, at Weinheim, that they had decided to revert to using miniature relays rather than PIN diodes because of greater isolation and reliability at the higher 144MHz IF transmit level that can now be used. This has become necessary because many newer VHF multimode rigs cannot be adjusted to below 5W RF output at 144MHz.

On transmit the 144MHz IF signal is connected, via the relay contacts, to the mixer. A transmit variable attenuator allows the transmit drive to the mixer to be set to the correct level so as not to over drive the mixer (a common problem).

The wanted 1296MHz product, at the mixer output, is selected by a helical filter and then PIN diode switched to the first amplifier stage, again filtered and then amplified to 3.4W output. This is a linear process, up to the point where the transmit chain starts to compress, and therefore is suitable for use with all modes including SSB, CW, digital (eg JT65C) and FM.

The G3 uses a Watkins Johnson FP31 HFET [49], rated at 2W for 1dB compression, in the transmitter output. This device is in a plastic 28 pin QFN package and this must be one of the reasons why Kuhne Electronics believe it is no longer practical for most radio amateurs to solder assemble units like the G3. **Fig 11.112** shows the component side of the G3 PCB. The FP31 can be seen clearly.

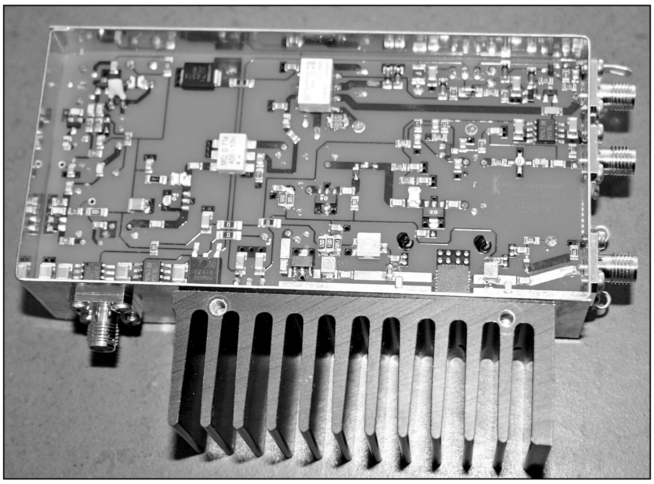


Fig 11.112: Surface mount component side of the PCB. The mixer, IF relay and FP31 output are readily identified

The transmitter and receiver chains are separate and unless separate transmit and receive antennas are used, an external antenna changeover relay must be used with a single antenna or array system. Most operators will want to use an external masthead preamplifier on receive and an external high power amplifier on transmit. If this is the case then the antenna changeover relay must be situated between these components and the antenna. It is also desirable to use an external low pass filter to reduce harmonics from the transmitter and this should be situated after the power amplifier.

Changeover, from receive to transmit, can be initiated by either applying a ground connection to the manual PTT input or, with a suitable IF rig, applying a small positive voltage on the IF coaxial cable connection. Details of where to find how to modify a number of popular rigs are given in the accompanying instruction sheet.

Frequency locking

Transverter local oscillators usually consist of an overtone crystal oscillator followed by a chain of frequency multipliers to the required injection frequency. Whilst many overtone oscillators are capable of better than 1kHz accuracy (at the final frequency) over a temperature range of around 0°C to +40°C, this is no longer regarded as adequate with the growing use of digital modes like JT65C and the desire to monitor the transmission from amateur beacons to an accuracy adequate to determine the amount of Doppler shift being introduced by the propagation path.

There are several ways to achieve the desired accuracy, including locking to GPS, an Oven Controlled Crystal Oscillator (OCXO) or a high stability atomic source (such as Rubidium) as the reference. Possibly the most popular way is to use what is known as a Reflock [50] board to lock the existing on-board overtone oscillator. This method is capable of giving very low phase noise because it uses the inherently high Q overtone crystal oscillator rather than a low Q voltage controlled oscillator (VCO). An external, high stability, low phase noise, 10MHz signal connected to the Reference input port will cause the in-built 96MHz crystal oscillator to phase lock to the 10MHz reference. This is subsequently multiplied to the required 1152MHz local oscillator injection frequency.

The technique used is limited by the very small frequency range over which an overtone oscillator can be pulled. If the oscillator drifts more than a few tens of Hz from the nominal frequency, the Reflock cannot maintain control. This drift may be caused by local heating of the crystal, for example when the transverter is in transmit mode, and the output amplifier is therefore generating considerable local heat inside the transverter box. To reduce this effect Kuhne Electronic has also used one of their small, proportional control, crystal heater boards attached to the crystal by a short length of heat shrink tubing. This is most effective.

Kuhne Electronics have used a National Semiconductor LMX2306 synthesiser chip in a conventional Phase Locked Loop (PLL) circuit to control the 96MHz crystal oscillator. The external 10MHz input acts as the reference frequency for the PLL. An Atmel eight-bit microcontroller ensures that the synthesiser is loaded with the correct frequency data each time the G3 is powered up.

Construction and interfaces

The transverter is built on a double sided PCB contained in folded metal box plated with what Kuhne Electronics describes as German Silver (typically 12.5% nickel, 50-65% copper and the balance zinc). This should not rust like the more common tin plate boxes, but it is more expensive to produce.

A small heat sink is attached to the side of the case and heat from the power amplifier device is conducted to the heat sink through an L shaped bracket. I do wonder if the copper sheet bracket will prove adequate in conducting away the heat. To be fair, the instructions do specify that the small heat sink must be attached to a larger heat sink. The heat sink is suitably drilled and tapped and two screws are provided to facilitate this.

SMA connectors are provided for the 144MHz IF, 1.3GHz transmitter output, 1.3GHz receiver input and the 10MHz external oscillator input. The power supply connection is via a feed-through capacitor. Feed-through capacitors are also provided to bring out an active connection on transmit, to drive the antenna changeover relay, a DC output representing the RF level and a manual press to talk (PTT) input.

On-air impressions

I was able to test the G3 in the heat of battle during the October 2007 IARU multiband contest. Used with my FT-817 as the IF and my two 44-element Wimo Yagis at 10m, this was an ideal opportunity to see how the transverter would perform with a mixture of strong local signals and a number of much weaker DX signals. I used an FXLabs 1500MHz low pass filter and CX540 antenna relay. With this arrangement I was not going to be a very big signal, so I allowed myself the luxury of using the ON4KST chat system to advertise my presence.

During the 1.5 hours I operated with the G3 I worked seven stations on the 1.3GHz band. The best DX was G3CKR/P at 252km and P14Z at 186km. Reports from local stations were complimentary when I told them what I was using.

Conditions were pretty average, but a good number of German and Dutch stations were heard, although with just 3W it was difficult to attract attention. The G3 acquitted itself well in the contest and the few receive problems I experienced were probably due to the performance of the FT-817. With about 20dB gain from the receive converter in front of the FT-817 this is not too surprising with strong local contest stations on the band. To get the best from the G3 a high performance 144MHz IF receiver must be used.

Comments

The G3 is a well constructed unit and should provide many years of reliable service. It meets or exceeds its claimed performance, allowing for the accuracy of the measurements.

One side effect of the use of the HFET output device, with its very high gain at high frequencies, is the relatively high level of harmonic output. This transverter should not be used directly

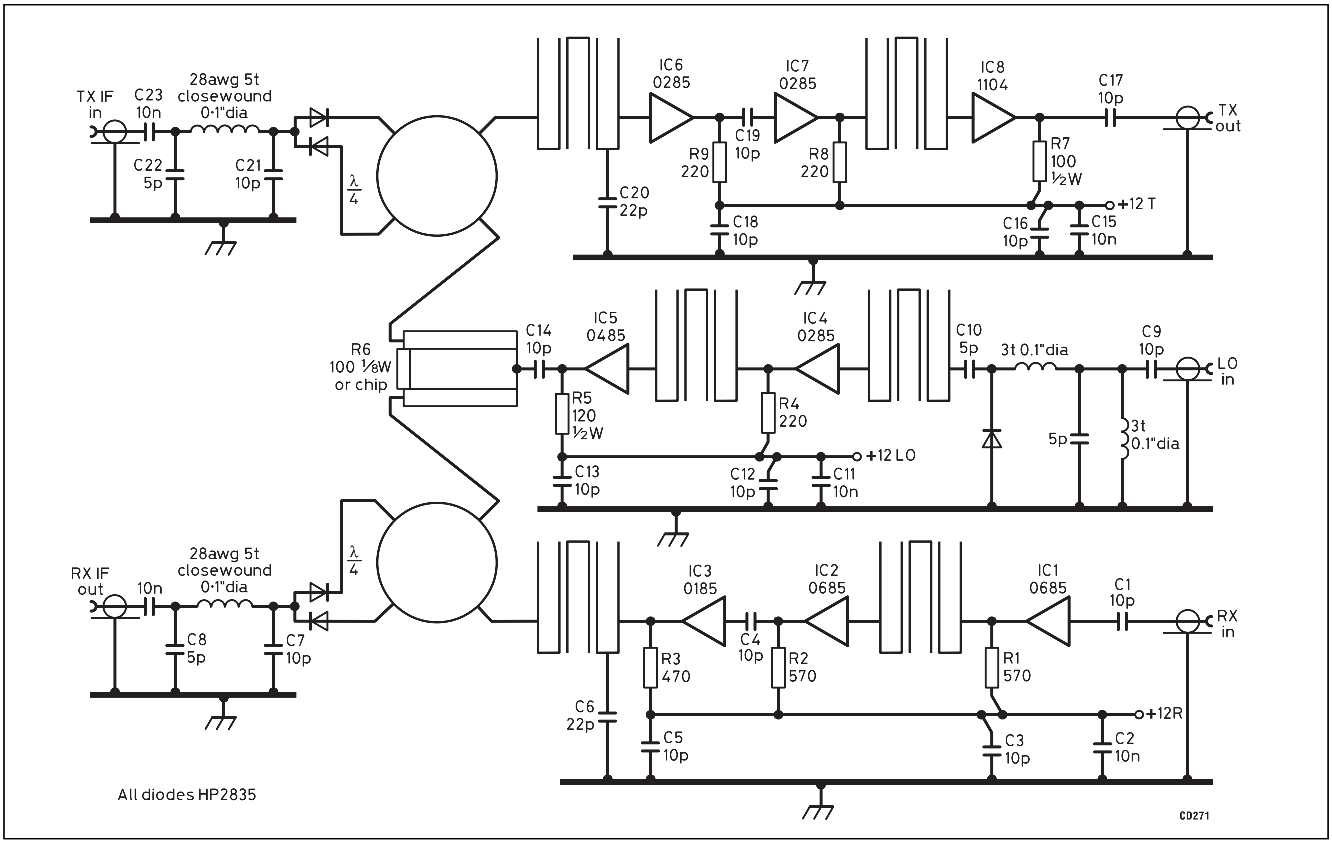


Fig 11.113: Circuit diagram of the KK7B transverter. BA481 Schottky diodes can be used instead of the HP5082-2835 diodes. Other, more modern MMICs can be substituted for the specified types, provided that the supply resistors are calculated and adjusted in value to suit the devices chosen. See reference [52]

into an antenna without a transmitter low pass filter to reduce the level of the 2nd and higher harmonics. Mention is made of this in the accompanying instruction sheet.

Now that transverters have a microcontroller inside I wonder how long it will be before we can expect a USB socket on the side of the unit, and the ability not only to interrogate the transverter for operating conditions and failures but to be able to programme operating parameters such as LO offsets and system gains?

My thanks to Kuhne Electronics for the loan of the test unit.

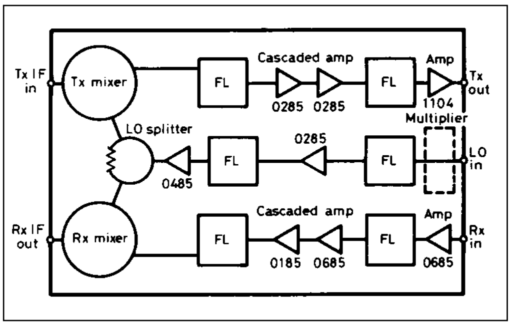


Fig 11.114: Layout for the KK7B transverter

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 [51]. 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.

The receive performance can be enhanced by means of an external (possibly mast-head) low-noise amplifier (LNA), such as a high-performance GaAsFET or PHEMT design. The transmit output level, at +13dBm (20mW), is ideal for driving a linear PA module such as the G4DDK-002 design [52]. 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 [53] have been available for some time, produced and marketed by Down East Microwave [54] 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 [55]. The Down East Microwave website [56] 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 11.113** and the physical layout of the circuit is shown, not to scale, in **Fig 11.114**. 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

Fig 11.115: Original KK7B LO circuit used with his no-tune transverter

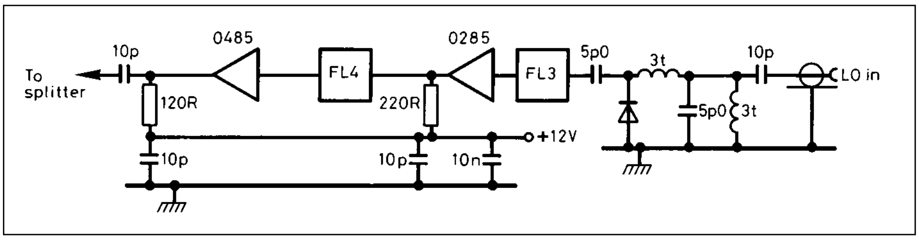
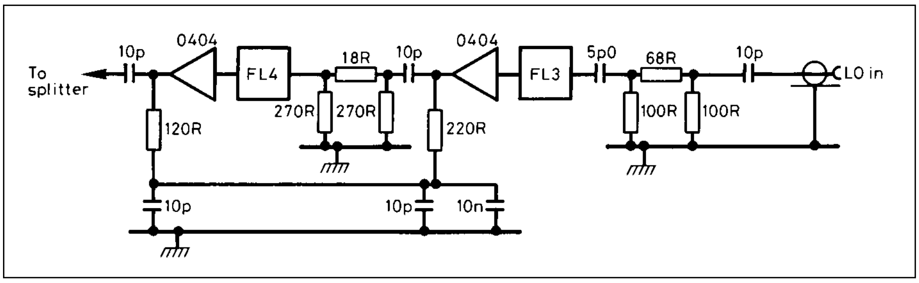


Fig 11.116: Modification to the KK7B LO circuit to allow use of the G4DDK-001 1152MHz source and the higher output level available from this circuit



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 11.115 gives the circuit and component values for the original LO chain, while **Fig 11.116** 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 11.117** 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 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 [57]. 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 [58].

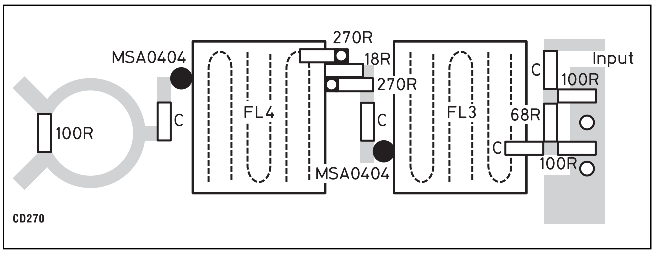


Fig 11.117: Layout of components for the modified LO chain in the KK7B transverter

Kuhne Electronics MKU10 G3 10GHz Transverter Review by Sam Jewell, G4DDK

Kuhne Electronic GmbH (DB6NT) produces a range of transverters, frequency converters, preamplifiers, power amplifiers and accessories for the amateur radio market. This is a review of the new MKU 10 G3 10GHz transverter. The 23cm MKU13 G3 1296MHz transverter, reviewed above has many circuit block similarities, so some of the following description may sound familiar if you read the MKU 13 G3 review above.

New features

Possibly the most significant feature of the new transverter is the addition of an external 10MHz local oscillator reference input. With the transverter local oscillator locked to a high quality 10MHz reference source, frequency stability is exceptional and an accuracy of better than 10Hz becomes practical on the 10GHz amateur band.

Fig 11.118 shows the top side of the PCB with the frequency locking board, helical filters and gain setting controls. The other side, with all the surface mount components, is shown in Fig **11.119**.

Construction

The transverter is constructed on a double sided printed circuit board soldered into a folded metal box. The main PCB carries surface mount components (SMD) on one side of the board and larger, leaded, parts such as the regulators and helical filters on

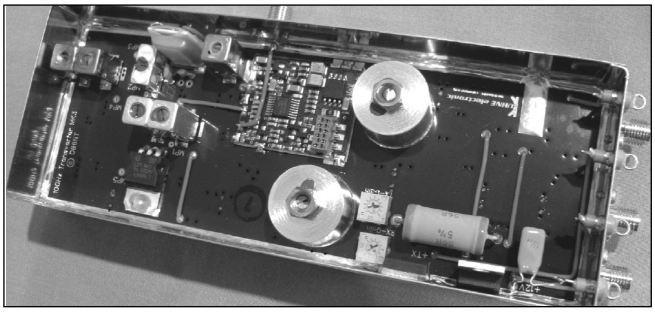


Fig 11.118: The top side of the G3 10GHz transverter, showing the frequency locking board, helical filters, crystal (with heater) and 50 ohm IF load

the top side. A small daughter board, with SMD parts, carries the reference frequency locking circuit. A second daughter board carries the crystal heater controller. GaAsFETs are used in many of the critical RF stages. The remainder of the stages use bipolar transistors.

The RF inputs and outputs all use SMA connectors. Power supplies, Press to Talk (PTT), TX supply and Monitor outputs connect via feed through capacitors. Press fit top and bottom box lids allow access to both sides of the main PCB. The lower lid carries a piece of electric field absorber foam to aid stability. Two short lengths of similar absorber foam are glued to the SMD side of the main PCB in the vicinity of the transmitter and receiver amplifier chains.

Receive path

An NEC NE32584 GaAsFET front end provides a noise figure of under 1.5dB. This is followed by two further GaAsFET amplifier stages using quarter wave length coupled microstrip capacitors. A microstrip Wilkinson splitter (combiner) separates the transmitter and receiver signal path. A substantial silver plated cavity filter is located between the splitter and the dual diode balanced mixer. This filter provides the sole RF selectivity on receive. A single mixer is used for both transmit and receive and is followed by a miniature relay that switches the mixer IF output to an adjustable attenuator (RX Gain) which is used to set the receive converter gain.

Transmit path

On transmit, the relay switches the 144MHz IF to a power attenuator consisting of a 5W (+37dBm) rated resistive load in a Pi configuration, with a series resistor and a variable resistor (TX Gain) to allow a range of adjustment. It is compatible with many of the modern 144MHz transceivers that cannot be adjusted to below 5W RF output.

With the TX Gain control set to give maximum 10GHz output (saturation), with 5W of 144MHz transmit input, the 10GHz power output was slightly in excess of the claimed 180mW output.

Following the mixer, the wanted 10GHz transmit product is filtered in the cavity filter, separated from the receive path in the Wilkinson divider and then amplified in three GaAsFET stages to around 280mW (+25dBm), at saturation output. There is a second cavity filter located between the first and second stage transmit amplifier to improve spectral purity. As the spectrum plot of **Fig 11.120** shows, this is just adequate to reduce the local oscillator feed-through and the image frequency to acceptable levels, although an extra stage of filtering would be desirable.

Both the receiver and transmitter chains are linear (up to the point where they start to compress) and therefore the transverter is suitable for use with all modulation modes including SSB, CW and FM. Most of the GaAsFET amplifier stages use active bias to maintain the bias operating point over temperature and device spread.

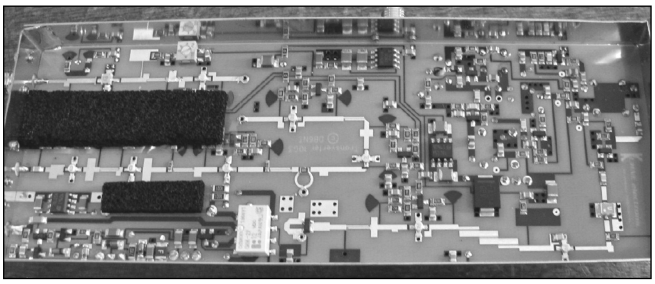


Fig 11.119: Component side of the G3 10GHz transverter PCB. The black absorber foam can clearly be seen

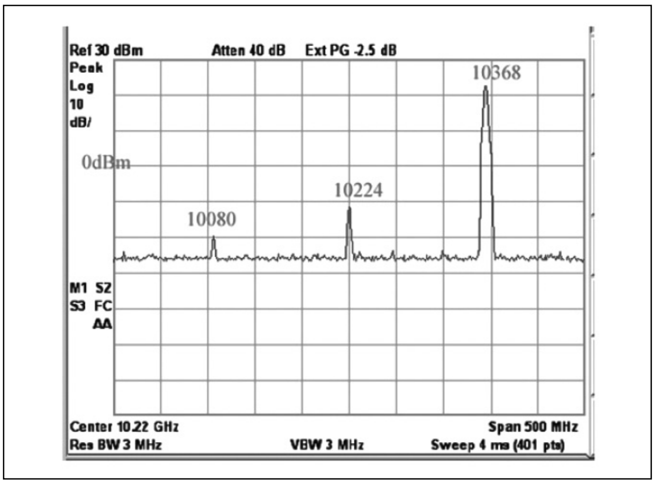


Fig 11.120: Plot of the output spectrum showing the level of the local oscillator and the image frequencies

The transmitter and receiver paths are separate and an external antenna changeover relay must be used to connect to the antenna.

Changeover from receive to transmit can be initiated by either applying a ground connection to the manual PTT input or, with a suitable IF rig, applying a small positive voltage on the IF coaxial cable connection. The transverter incorporates a simple but effective sequencer circuit for the internal transmit and receiver timing.

Local oscillator

In common with most other 10GHz transverters the MKU10 uses a crystal controlled overtone oscillator at 106.500MHz followed by a chain of frequency multipliers to the final injection frequency of 10224MHz. Bipolar transistor multipliers are used up to the final 2556MHz multiplier stage. From 2556MHz to 10224MHz a GaAsFET multiplier, followed by an LO amplifier is used. Intermediate stages use commercial helical filters whilst the final 10224MHz selection filter is a multistage side coupled microstrip circuit. The 2256MHz stage uses a miniature ceramic block filter.

Frequency locking

In order to provide exceptional frequency stability the MKU 10 G3 uses an optional phase locking circuit. Whilst most overtone oscillators are capable of better than 1kHz accuracy (at the final 10GHz frequency) over a temperature range of around 0°C to +40°C, when suitably conditioned, this is no longer regarded as adequate, even for 10GHz.

There are several ways to achieve better accuracy using a reference source derived from GPS satellite, an Oven Controlled Crystal Oscillator (OCXO) or a high stability atomic source such as Rubidium. Possibly the most popular way to provide the desired locking is to use what is known as a Reflock [50] board overtone oscillator to lock one of these reference sources. This method also gives exceptionally low phase noise because it uses the inherently high Q overtone crystal oscillator rather than a low Q voltage controlled oscillator (VCO).

An external, high stability, low phase noise, 10MHz signal, connected to the Reference input connector, will cause the inbuilt 106.5MHz crystal oscillator to phase lock to the 10MHz reference.

The Reflock technique used is limited by the very small frequency range over which an overtone oscillator can be pulled. If the oscillator drifts more than a few tens of Hz from the nominal frequency, the Reflock cannot maintain control. To improve basic stability Kuhne Electronic has used one of their small, proportional control, crystal heater boards attached to the crystal by a short length of heat shrink tubing. The heater operates at 40°C and is very effective at holding the crystal at that temperature over a large range of ambient temperature up to 40°C.

Kuhne Electronic has used a National Semiconductor LMX2306 synthesiser chip in a conventional Phase Locked Loop (PLL) circuit to control the 106.5MHz crystal oscillator. The external 10MHz input acts as the reference frequency for the PLL. An Atmel 8 bit microcontroller ensures that the synthesiser is loaded with the correct frequency data each time the G3 is powered up. The PLL circuit is assembled on a small daughter PCB that is soldered onto the top of the main PCB.

On-air impressions

I was only able to try the transverter on air in receive mode due to time constraints. I used a 1 metre length of low loss coaxial cable to connect a 17dBi gain horn antenna to the receive connector of the transverter. With the horn pointed out of the shack window, towards the dock cranes at Felixstowe, I was able to hear the 10GHz Martlesham GB3MHX beacon, by reflection, at strengths of up to 539. This is a remarkably good result and by comparison a barefoot G3JVL transverter (a popular design 20 years ago) gives a barely detectable signal when similarly connected.

I used my HP Z3801 10MHz GPS disciplined oscillator (+9dBm output) as the reference input and this worked extremely well, providing a frequency accuracy on 10GHz that was ultimately limited by my TS2000 IF transceiver and not by the transverter local oscillator chain.

Comments

The MKU10 G3 is a well constructed unit and should give many years of service. It may appear to be an expensive way to get onto the amateur 10GHz band. However, many amateurs think nothing of spending over £1000 on an HF transceiver and often hundreds of pounds on a commercial antenna. By comparison, the 10GHz band is also capable of providing a great deal of fun and access to propagation modes such as rain scatter, tropospheric ducting and aircraft reflection that are not experienced on HF. My thanks to Kuhne Electronics for the loan of the test unit.

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 [59].

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 2.272GHz, and when multiplied by five it becomes 11.360GHz 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 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 11.13** 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

3216 PLL Calculations for X Band Transverter with 144MHz 2nd IF; 1st LO=11,360MHz; 1st IF=992MHz

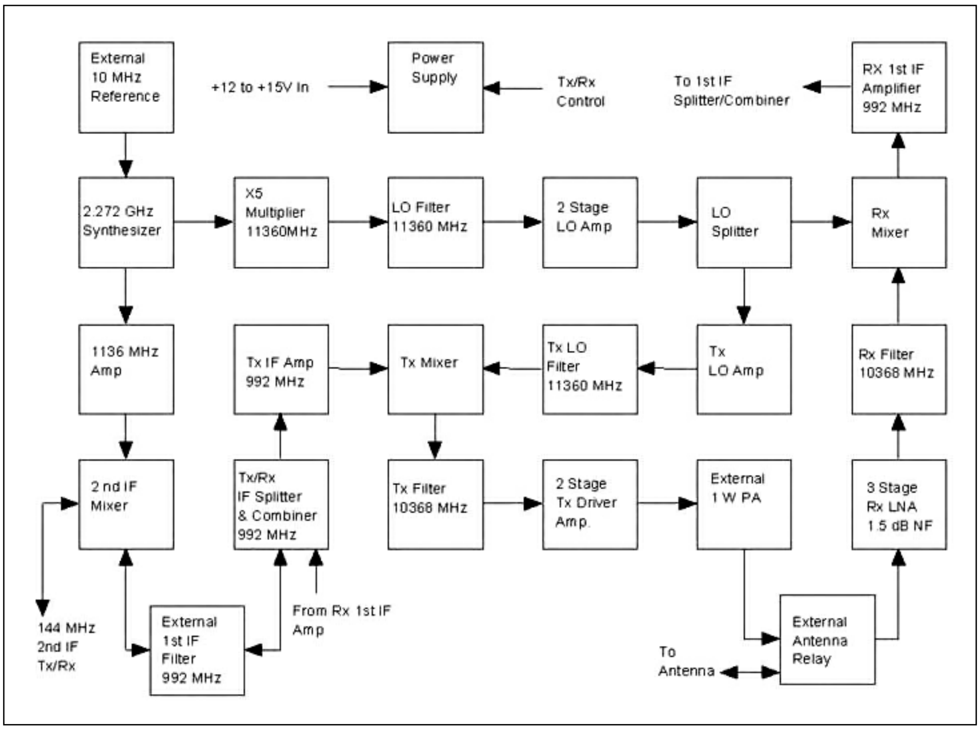
Ref MHz	2	Ref MHz can be 10MHz divided by any integer from 1 - 16						
VCO MHz	2,272							
PLL MHz	1,136	PLL in MHz is VCO/2 and must be an integer multiple of Ref MHz						
N	568							
		M6(Pin15)	M5(Pin14)	M4(Pin13)	M3(Pin10)	M2(Pin9)	M1(Pin8)	M0(Pin7)
M	55	0	1	1	0	1	1	1
Board as is		0	0	0	0	0	0	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 11.13: Synthesiser calculations for Qualcomm OmniTracks unit. 3216 PLL calculations for X Band Transverter with 144MHz 2nd IF; 1st LO=11,360MHz; 1st IF=992MHz

Fig 11.121: Block diagram of Qualcomm X band transverter conversion



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 11.121** 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 11.122 shows a picture of the modified transverter, 1W amplifier and 10MHz TCXO. **Fig 11.123** shows a picture indicating the locations of the various functions. The following is an outline of the conversion procedure [60]:

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.

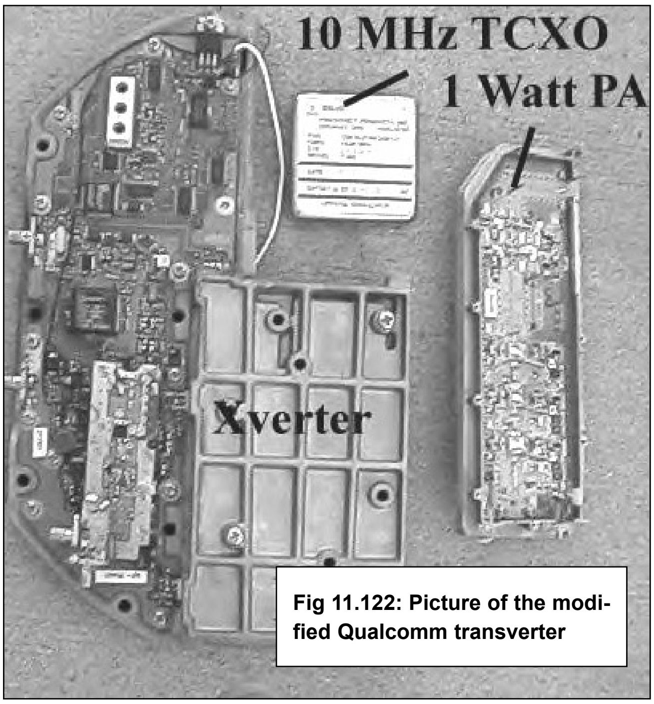


Fig 11.122: Picture of the modified Qualcomm transverter

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 11.124** using a sharp knife.

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 11.125**) 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 three capacitors along with the additional microstrip pieces to modify as shown in **Fig 11.126**.

Step 4. Extend the transmit LO filter elements to the total length shown in **Fig 11.127**: 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.

Step 5. Extend the LO filter elements as shown in **Fig 11.128**: 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 11.129**: Dimensions shown are total element length.

Step 7. Extend the transmit filter elements as shown in **Fig 11.130**: 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 11.128**. The gate of the x5 Multiplier stage requires addition of two stripline stubs, as shown in Fig 11.131.

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

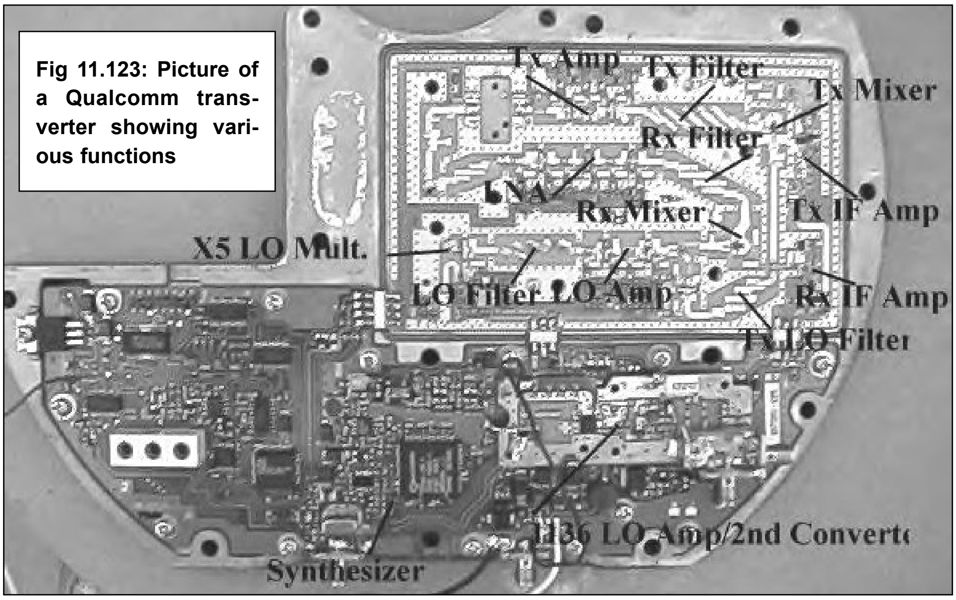


Fig 11.123: Picture of a Qualcomm transverter showing various functions

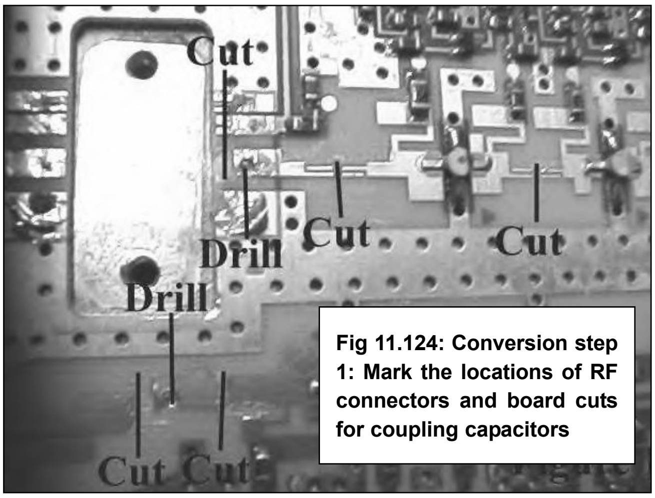


Fig 11.124: Conversion step 1: Mark the locations of RF connectors and board cuts for coupling capacitors

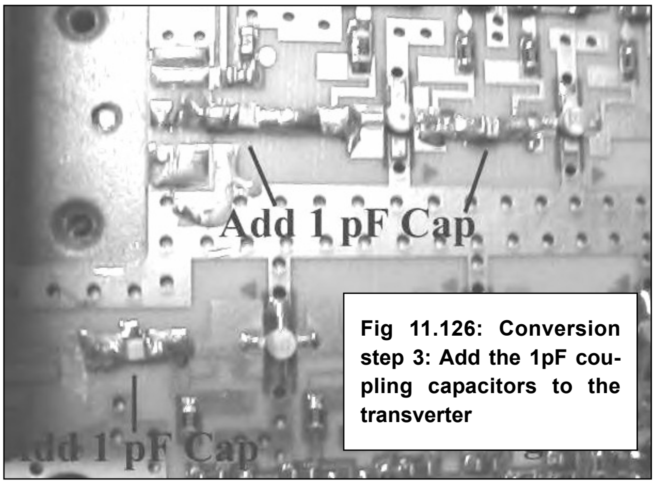


Fig 11.126: Conversion step 3: Add the 1pF coupling capacitors to the transverter

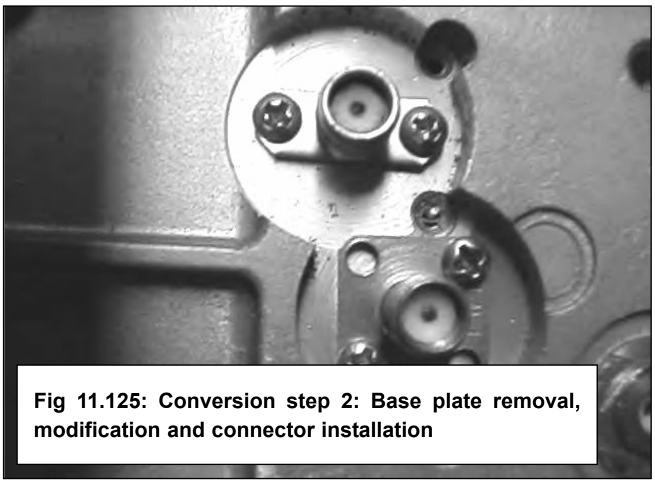


Fig 11.125: Conversion step 2: Base plate removal, modification and connector installation

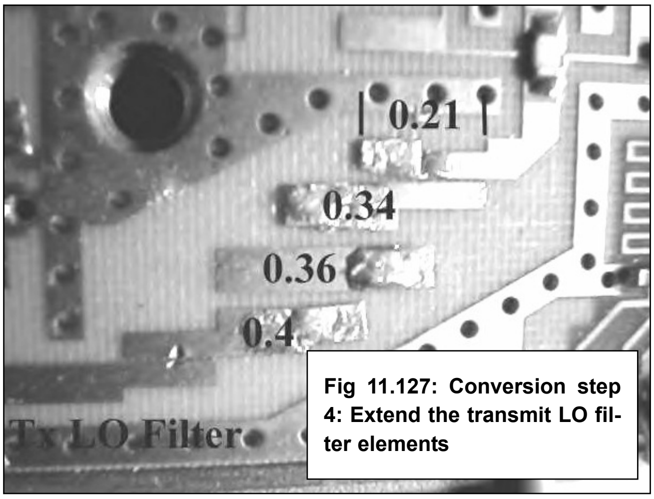


Fig 11.127: Conversion step 4: Extend the transmit LO filter elements

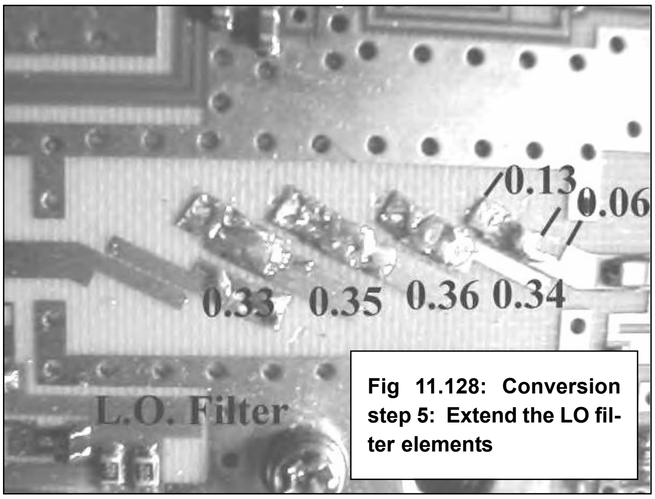


Fig 11.128: Conversion step 5: Extend the LO filter elements

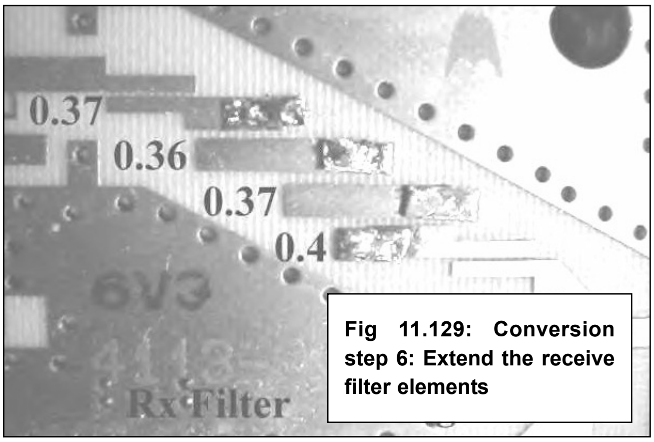


Fig 11.129: Conversion step 6: Extend the receive filter elements

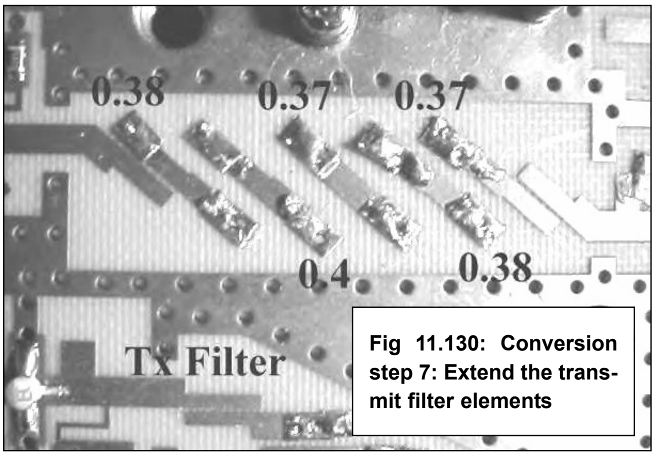


Fig 11.130: Conversion step 7: Extend the transmit filter elements

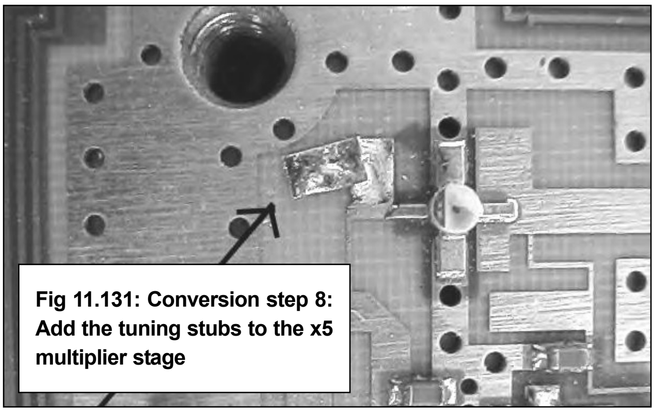


Fig 11.131: Conversion step 8: Add the tuning stubs to the x5 multiplier stage

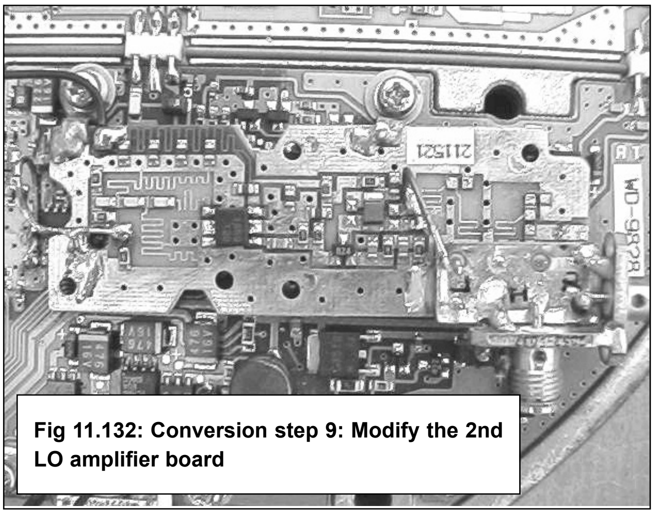


Fig 11.132: Conversion step 9: Modify the 2nd LO amplifier board

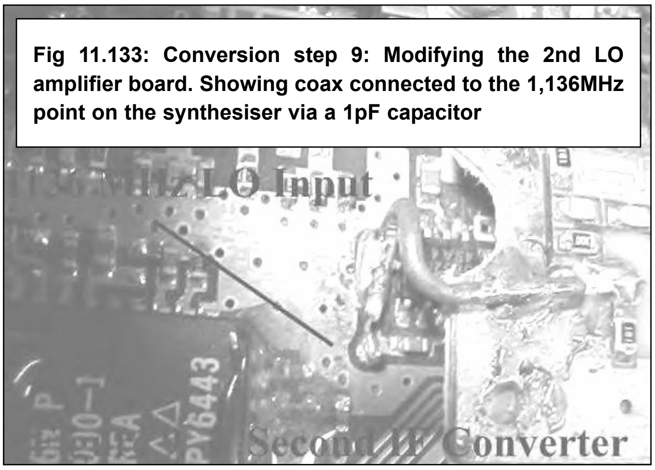


Fig 11.133: Conversion step 9: Modifying the 2nd LO amplifier board. Showing coax connected to the 1,136MHz point on the synthesiser via a 1pF capacitor

through 1pF coupling capacitor as shown in **Figs 11.132 - 11.134**. Fig 11.132 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 11.133 shows the coax connected to the 1,136MHz point on the synthesiser through a series 1pF capacitor. Fig 11.134 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 11.135** 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 11.136**.

Step 11. Add a 1pF capacitor as shown in **Fig 11.137** to lower the VCO frequency

Step 12. Add three transmit mixer tuning stubs as shown in **Fig 11.138**.

Step 13. The transmit/receive control is connected as shown in **Fig 11.139**. 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 11.140**. 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).

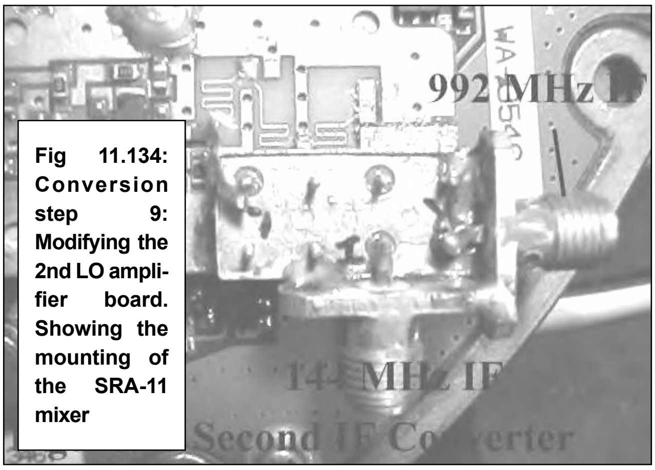


Fig 11.134: Conversion step 9: Modifying the 2nd LO amplifier board. Showing the mounting of the SRA-11 mixer

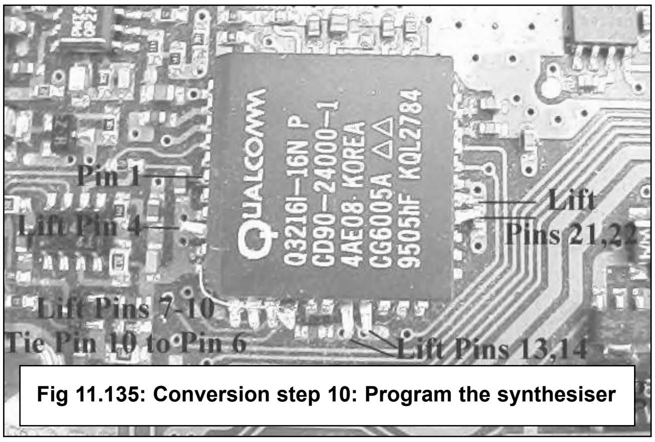


Fig 11.135: Conversion step 10: Program the synthesiser

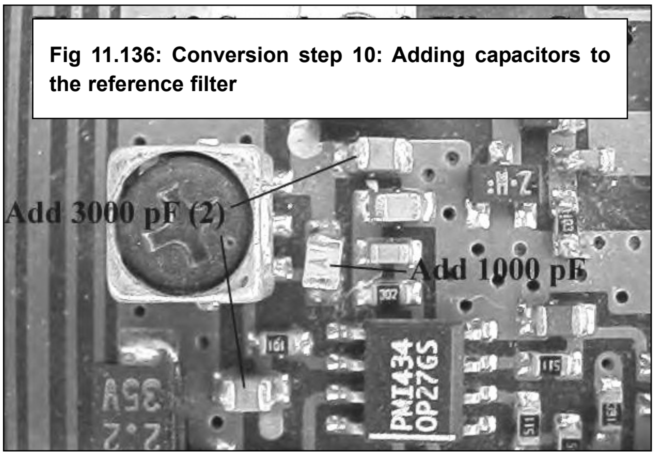


Fig 11.136: Conversion step 10: Adding capacitors to the reference filter

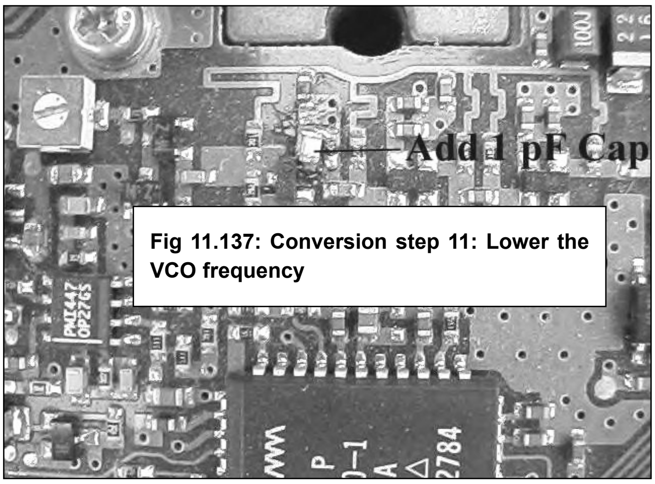


Fig 11.137: Conversion step 11: Lower the VCO frequency

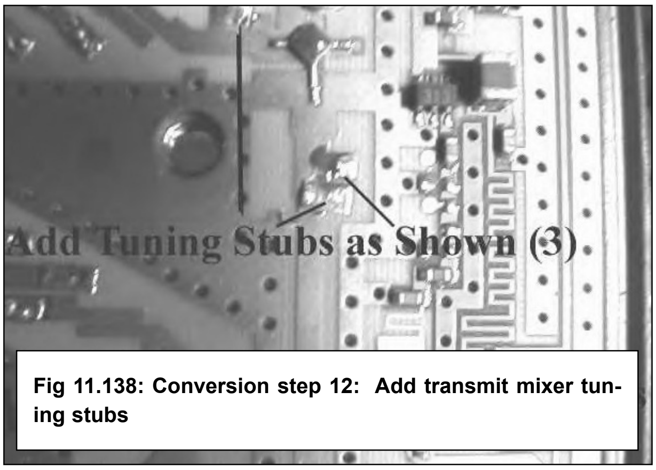


Fig 11.138: Conversion step 12: Add transmit mixer tuning stubs

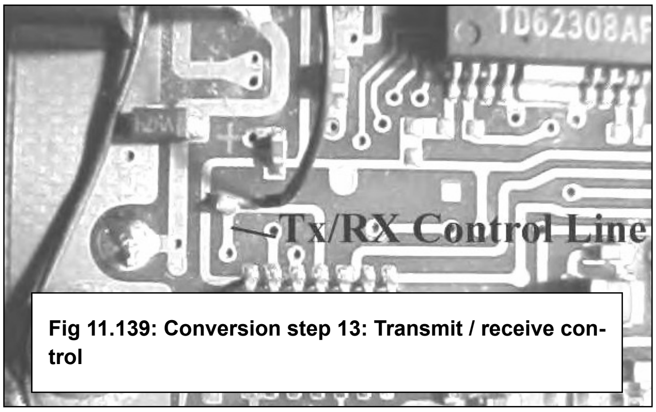


Fig 11.139: Conversion step 13: Transmit / receive control

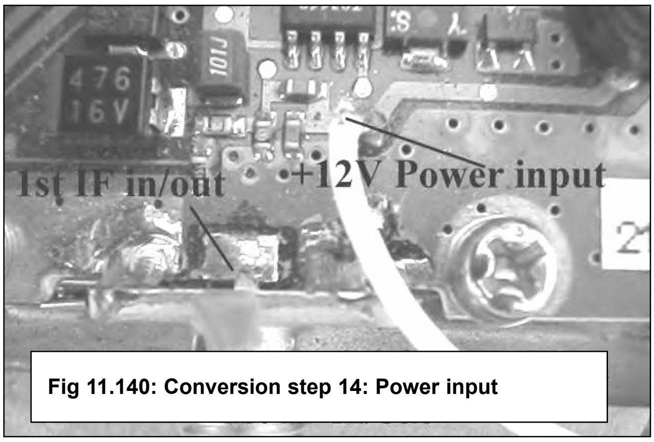


Fig 11.140: Conversion step 14: Power input

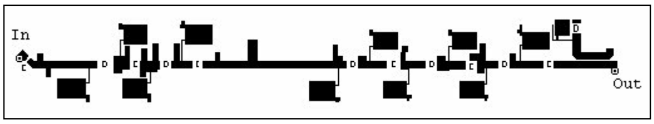


Fig 11.141: 1W PA board prior to tuning. -15dBm input gives +5 to +10dBm output with 10V at approximately 1A

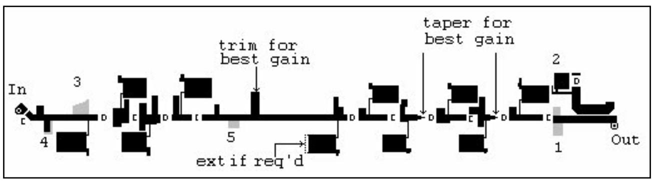


Fig 11.142: 1W PA board after tuning. The shaded tabs were added and tuned in the sequence shown. Results vary slightly from board to board. Key: [= coupling capacitor, D = devices, Input coupled with 2pF

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 receive 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 2m 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 11.141** 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.

To convert the 1W PA

These conversion notes were produced by Ken Schofield, W1RIL [61]: 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 11.142** 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

OPTICAL COMMUNICATION

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.

Latest Developments

The following contribution by Stuart Wisher, G8CYW, continues the development of optical communication introduced in the 8th and 9th editions of this Handbook by David Bowman, G0MRF (see also the CD attached to this edition).

The use of small lasers as the active element in the transmitter has now been superceded by the use of power LEDs which have been developed since the first article was published [62]. The author has been greatly aided by a small group of radio amateurs without whom much of this development would not have been possible. This work is dedicated to the memory of Tony, G8NPP who is sadly no longer with us: tragically he was killed in a traffic accident driving to a group activity in December 2008.

The AM Laser System

G0MRF's system (**Fig 11.143**) had separate head units for transmission (a laser) and reception (an expensive OPT301 opto IC). A more economic solution is to use a separate photodiode and low noise op-amp.

A SFH2030 photodiode and a low-noise NE5534 op-amp are used in the first receiver developed by G8CYW. The use of a precise 488Hz Morse tone for a modulated carrier wave (MCW) output and a very narrow filter in the receiver are continued, but with the inclusion of a 20kHz pulse width modulator (PWM) and a switch in the receiver filter circuit to broaden the bandwidth to permit voice communications.

The photodiode and amplifier are mounted in a 4 x 2 x 1 inch diecast box (**Figs 11.144 and 11.145**), acting as a receive

Fig 11.143: G0MRF's laser transmitter and receiver

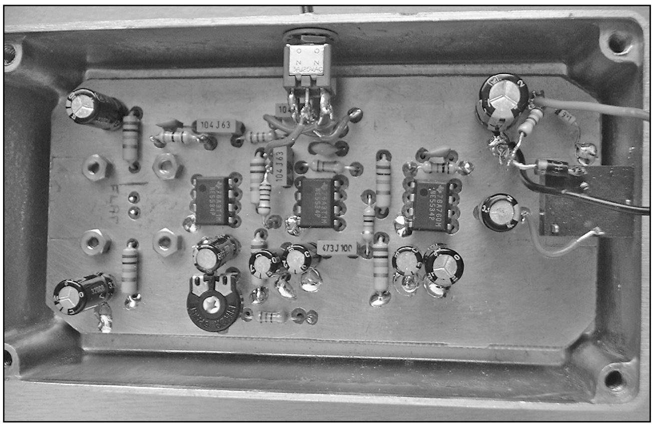


Fig 11.144: Original AM receive head. The switch adjusts bandwidth between CW and voice

Fig 11.145: The completed AM receive head, showing the photo diode and mounting tube

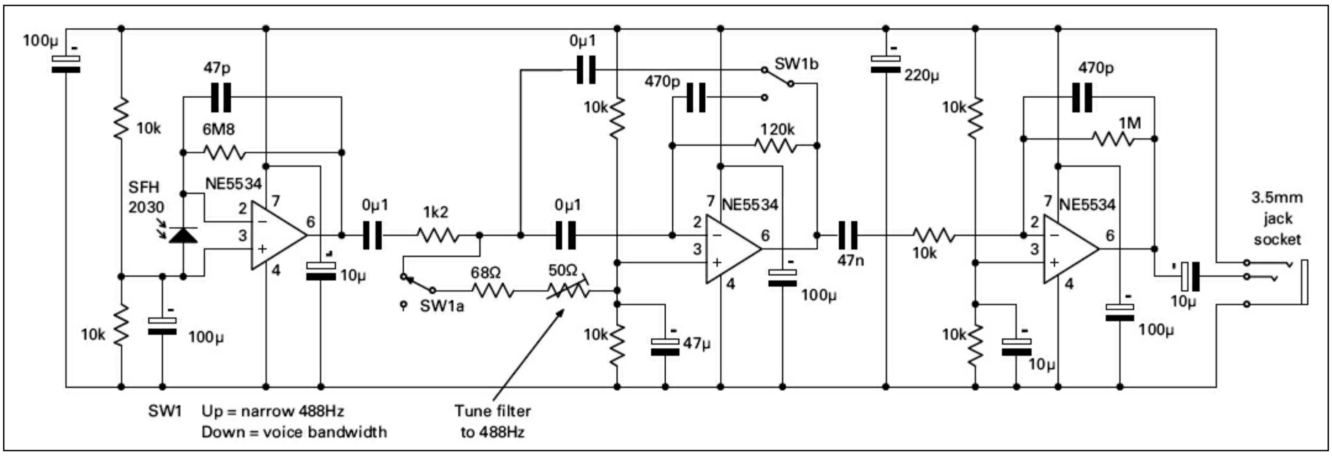


Fig 11.146: Circuit diagram of the AM receive head

head. The final stage of the receiver, a standard LM386 audio amplifier IC, volume control and loudspeaker, are contained in a separate box. A 3.5mm stereo jack plug, socket and lead are used to connect the boxes, carrying power, signal and ground. The transmitter circuit also has its own separate box, although there is no reason why it could not be accommodated in the receive audio amplifier and speaker box.

Fig 11.146 shows the circuit diagram of the receive head.

The Morse circuit is altered with the addition of an inverter to give maximum light output on key up (no modulation), which helps when aligning over a distance. The driver transistor produces an output of 3V at around 30mA, ideal for the average laser pen; laser pointer or laser spirit level. Also included is an output for a front panel LED so that it would not only indicate when switched on, but it would act as a low power transmitter beacon for short range testing (which is very useful). **Fig 11.147** shows the CW/PWM transmitter circuit diagram and **Fig 11.148** shows the completed unit.

The performance of this system is such that the MCW tone can be received at a distance of 6km with no lens on the receiver; at 15km, a one inch lens is required.

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

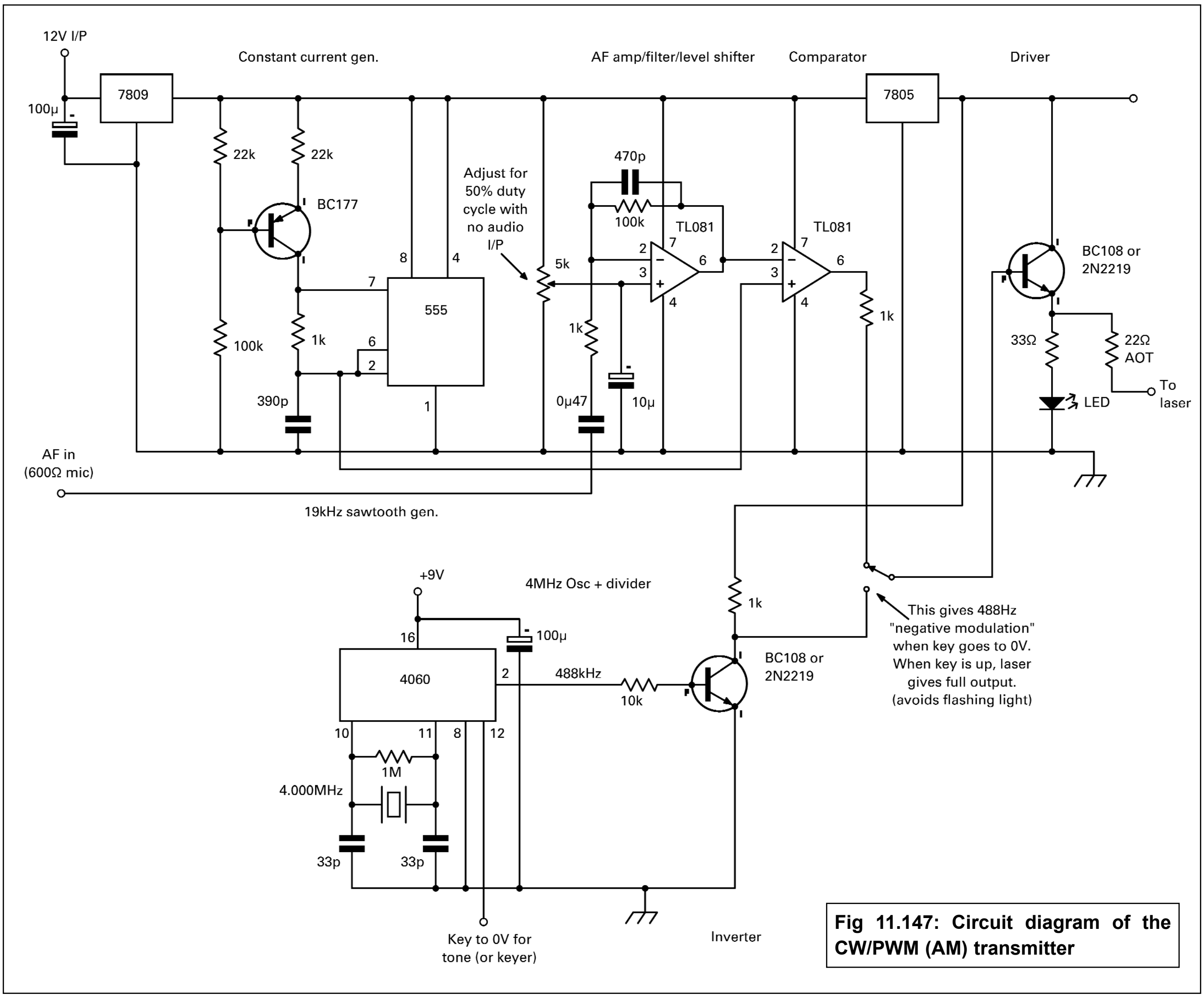


Fig 11.147: Circuit diagram of the CW/PWM (AM) transmitter

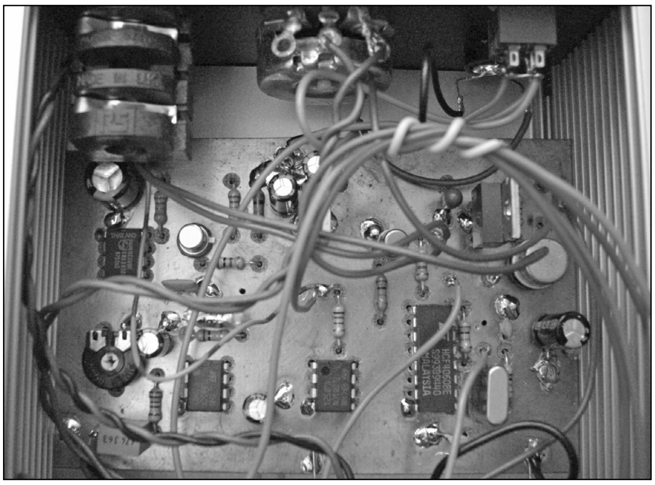


Fig 11.148: Completed CW/PWM transmitter

Fig 11.49: Parts used in constructing the receive optics

Optimum optics

All that is required of the optical part of the system is to collect as much of the available beam as possible and concentrate it onto the photodiode in much the same way as a microwave dish focuses radio waves on to the feed.

The optical system that evolved used a length of 110mm waste pipe, an end cap and a joining piece (all available from DiY warehouses). A Blue Spot 100mm magnifying glass was located at a pound shop, the handle was cut off and the protrusion where the handle met the rim was then filed smooth. This was found to be a tight push fit into the end of the tube. (Be careful at this point: the band around the lens is not parallel. Place it narrow end upwards on a sturdy table and push the tube over it. When the tube rim touches the table top, the lens is properly installed.) **Fig 11.49** shows the constituent parts.

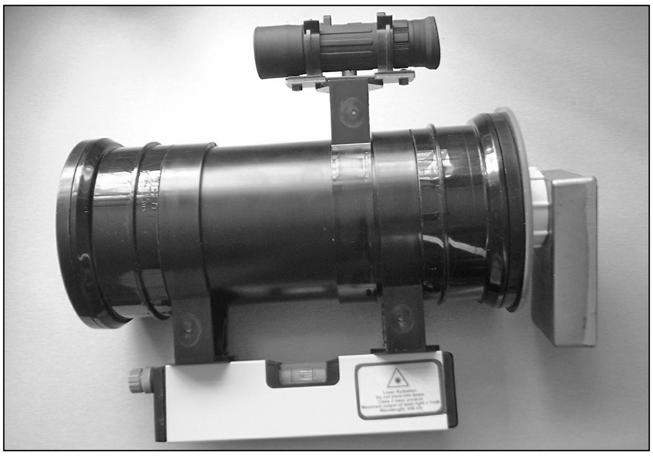


Fig 11.150: The complete receiver

The cylinder on the end cap is cut down to 20mm and the centre of the blanking disc drilled out to allow light to pass through. A 38mm diameter hole cut with a hole punch was found to be convenient. A 110mm pipe joining piece was cut exactly in half and one half is used to push the end cap, the cut end then is slid over the tube on the far end from the lens, making a loose fit that can be improved by using a turn or two of PVC insulating tape on the tube. The length of the tube depends on the lens. You need to be able to produce a focused image of a distant street lamp on the photo diode. The lenses were found to vary in focal length slightly; most required around 285mm from lens centre to the diode surface.

Lasers

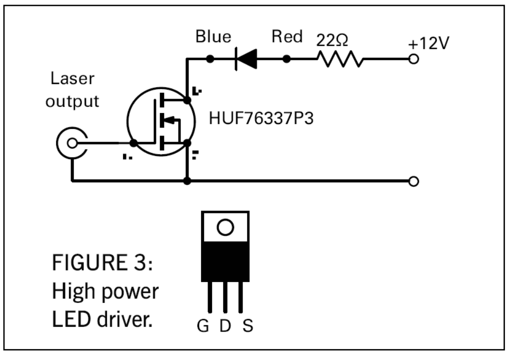
Originally, lasers were obtained from a cheap German supermarket, where a laser level kit was often available for £10. The kit includes a tripod, adjustable head and a spirit level with the laser in it. The spirit level can be modified to take two connections out to a 3.5mm socket to use it as the transmit head. A second kit provides everything needed to mount the receive tube. Simply bolt a section of the aluminum spirit level to two 110mm pipe wall clips (also useful for marking an accurate line around the tube before cutting) and slot the tube in place through the hoops. This means that a laser level is sacrificed to science, but this is the cheapest and easiest way. A further wall clip or two can be used to support a finder telescope or half a pair of cheap binoculars - an essential aid to lining up. This assembly then clamps in the adjustable head. A completed unit is shown in **Fig 11.150**, which also shows some 40mm waste pipe fittings used to hold the receive head. This arrangement also enables different heads to be slid in and out and for fine tuning the focus of the system, although once the focus is set it doesn't normally need adjusting.

The other half of the joining piece is used as a lens hood. This also seemed to finish the system off well. It has also proved useful as a holder for irises - cardboard discs with holes in - that act as signal attenuators. These can be used to assess how little signal is needed at a given distance, as an aid to calculating the potential range of the system. A 50mm diameter hole gives 6dB attenuation and a 25mm hole gives 12dB attenuation. Ignoring atmospheric effects, these equate to signal levels at twice and four times the range respectively.

Operation

AM works well at short range, but there are, however, some issues with the use of this system. There is a lot of QRM from street lights. The signal flutters at long distance due to atmospheric scintillation (twinkling), and aiming the laser accurately can be quite difficult. To address these issues, a high power LED (and lens arrangement similar to the receive lens) is now used which eventually got to the point where the signal could be detected 34km away.

Fig 11.151: High power LED driver



Power LED

The advantage of a LED transmitter is that it is much easier to aim because it has a broader beam than the laser. Also, it avoids the issues involved in aiming a laser over the countryside and, since the power density is much lower, it is safer. That said, a 1W LED using these optics still looks very bright, even when lined up over a distance of several kilometers.

Changing from a laser to a power LED is an easy move. Another tube and lens system is required - so there is a use for the second tripod, head and spirit level: to hold the second tube. At the rear of the end cap, a diecast box makes a good mount and heatsink for the power LED. The completed assembly looks rather like the detector in Fig 11.149.

The LED drive circuit (**Fig 11.151**) is an N-channel power MOSFET, which requires a small heatsink. The gate goes to the transmit electronics, the source to 0V and the drain to the LED cathode. The anode goes to +12V.

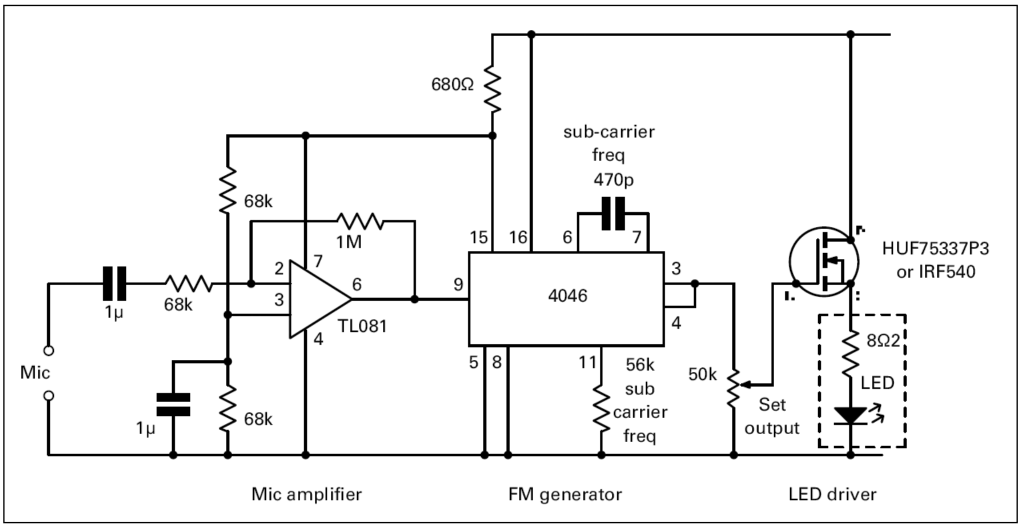
LED operation reduces the operating noise somewhat, because the wider (optical) beamwidth doesn't suffer from the 'speckle' and scintillation of signals that is common with lasers. But this remains an AM-based system, which has pronounced issues with fluttering signals plus QRM from streetlights and road traffic. These issues caused much thought to see if a solution could be found to get round these problems. A web search, looking for 'laser DX' and 'optical communication' revealed a wealth of material out there. A great source of information is the Optical Links site run by Tim Toast [63]. The progress made by various groups of optical communication enthusiasts can be found mainly in Australia, Czech Republic, Finland, Germany and USA, to name just some of the major contributions. Of special note are VK7MJ and group who have communicated by voice over 160km and KA7OEI and group who have exceeded even this. To date, most have now progressed into weak signal modes and the Australia/Tasmania groups have spanned the Bass Strait between Australia and Tasmania by cloudbounce, a distance of some 288km.

Since in the UK we do not have any huge mountains or dry flat deserts to provide long optical paths and our atmosphere is cloudy and misty most of the time, we cannot really compete on distance, so we re-defined our aims to involve immediate real-time microphone to loudspeaker communications.

FM System

Web searches produced a receive head design by VK7MJ that had a frequency response from audio to 50kHz and beyond. This could be used in an FM subcarrier system, centred on approximately

Fig 11.152: Basic FM transmitter operating at ~25kHz.



25kHz. At this frequency the QRM from street lights was significantly reduced (if not absent) and the effect of the limiter in an FM receiver helped to overcome the fluttering of signals.

The FM transmitter shown in **Fig 11.152** uses an op-amp based microphone amplifier/filter connected to a 4046 PLL IC VCO generator running at 25kHz.

The audio signal produces frequency modulation around the main carrier frequency. A MOSFET driver connected directly to the oscillator then provides adequate drive to the power LED head.

The receive head (**Fig 11.153**) starts with VK7MJ's design, followed by an amplifier tuned to the sub-carrier frequency feeding a NE566 PLL demodulator. All of this fits into a 4 by 2 by 1in diecast box and uses the same 3.5mm stereo jack system for power and signal connections that connect to the audio amplifier/ speaker box from the AM system. All the optics remain as they were for the AM system. A later modification is to make this head switchable between AM and FM by tapping directly into the output of the original VK7MJ circuit. The first dual-mode optical receiver was thus created. Just for good measure the transmitter box included a linear (rather than PWM), AM transmitter circuit to complete the dual-mode setup.

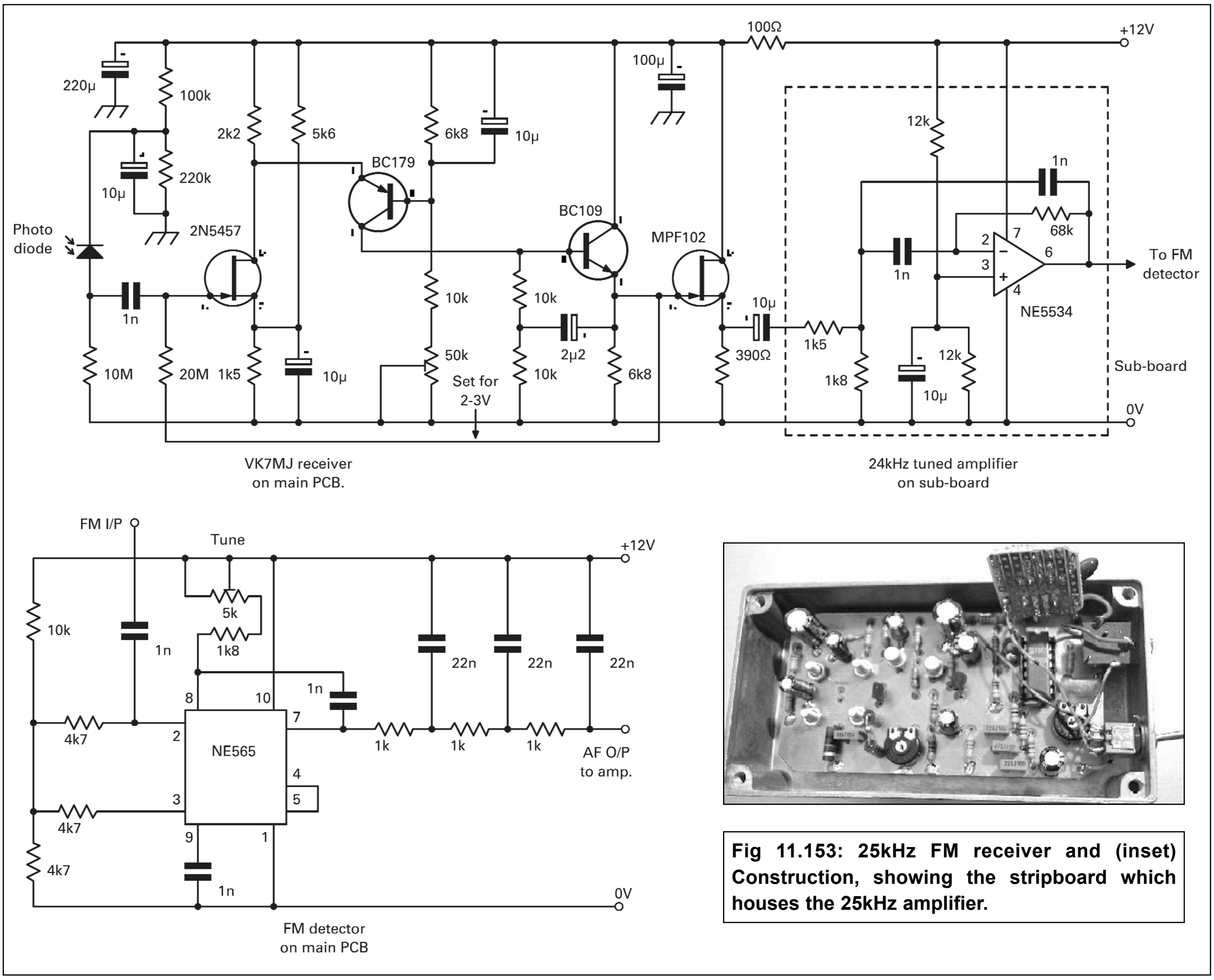


Fig 11.153: 25kHz FM receiver and (inset) Construction, showing the stripboard which houses the 25kHz amplifier.

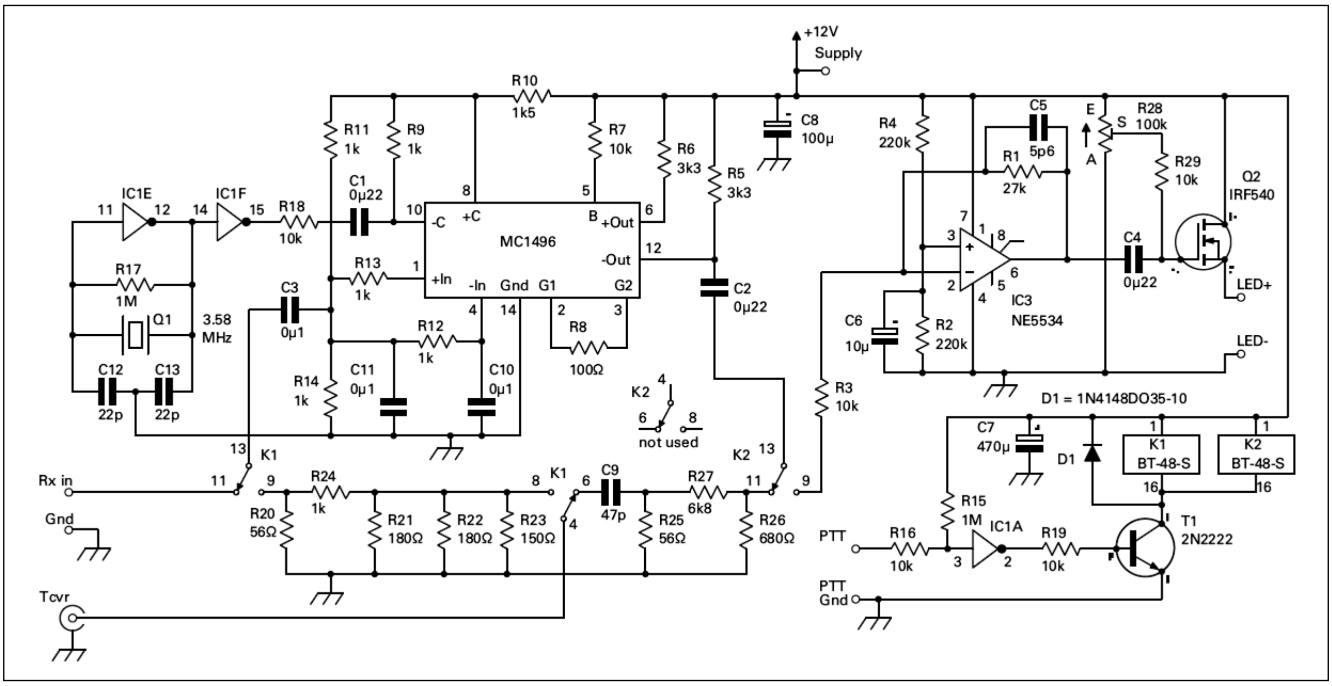


Fig 11.154: Circuit diagram of the transverter to convert 3.5MHz SSB to 20kHz for 'radio over light'

FM results

Short range tests at 6km showed FM to hold much promise. Very strong signals with no QRM or flutter were achieved. Going on to the 15km path gave similarly good results even in near proximity to powerful lights. Since then FM has been used over all paths tried up to 34km and strong signals and clear communication has always been possible.

Optical Transverter

After building and testing the FM system, G8CYW wondered what advantage, if any, might be gained by going to SSB. This mode is paramount for long distance voice communication over most, if not all, amateur bands where it is allowed, so why not on light? The thought of building stand-alone systems for receive and transmit for SSB was eclipsed by one of those *eureka!* moments: why not build a transverter? All of the gain and signal processing power of a small HF amateur radio transceiver (eg FT-817) could be utilised, converting the HF signals to light and vice versa.

Since the group were already operating on 20kHz PWM AM and 25kHz FM, thoughts turned to the possibility of sending and receiving single sideband (or any other mode for that matter) on around the same frequency. Crystals for 3.58MHz (actually 3.579545MHz) are readily available, enabling the 80m band to be used for the intermediate frequency. The chosen operating frequency is 3.6MHz RF, thus producing or receiving an optical signal around 20kHz. This is known as 'radio over light'. It even looks and feels like you are operating a real radio when making optical contacts.

The circuit

The transverter circuit is shown in **Fig 11.154**. A relay switches the HF transceiver between the receive and transmit paths in the transverter (via resistive attenuators). The transmit attenuator reduces the 0.5W output from the transmitter to a few millivolts. This signal is mixed with the 3.58MHz local oscillator signal in the MC1496 balanced mixer to provide the 20kHz signal. This is amplified by a NE3354 opamp which also filters out the unwanted 7.2MHz mixer product, and the signal is then fed to the gate of a power MOSFET. The MOSFET has adjustable bias via R28 to set the quiescent gate voltage (and hence LED idling current). The LED and resistor are contained in a separate box (the transmit head, described later), mounted at the focus of a lens in a similar manner to the previous system.

On receive, the amplified 20kHz signal from a receive head (of which more later) is connected to the input of the mixer, which up-converts the signal to the 80m band again thanks to the 3.58MHz local oscillator. The output of the mixer is switched to an attenuator to protect the HF transceiver's front end. The attenuator also protects the mixer IC against inadvertent transmission into the mixer output if the PTT fails. The local oscillator can be heard at a low level on the HF receiver if you tune down to the region of 3.58MHz. It is not strong enough to de-sense the receiver if you keep at least 10kHz away from it.

This linear transverter is useable on any mode (although FM and SSB are best) and on a range of frequencies. 3.600 MHz RF gives 20 kHz optical; 3.650 MHz gives 70 kHz and so on. Lower frequencies should be better because they are less demanding of the electro-optical system. It is best to keep to frequencies around 3.600 MHz (20 kHz optical).

The prototype transverter PCB and overlay are shown in **Figs 11.155 and 11.156** (in Appendix B). Note that the board should be double sided - the top is a ground plane that doesn't require etching. After you drill the board, identify all the ground connections and mark them (a marker pen pressed against the holes will usually be visible form the other side). Then turn the board over and, using a hand-held drill bit of about 3mm, clear the copper around the non-earth holes (eg the diode in the foreground of **Fig 11.157**). When you populate the board, solder all earth connections on both sides of the board (eg the four resistors near the middle). This will result in a good quality of screening.

Setting up

Before connecting power for the first time, turn R28 so the wiper is down to 0V (fully anticlockwise if you're using this PCB). Check the board current consumption - it should be around 20mA at 12V. Connect the LED head and, while monitoring current

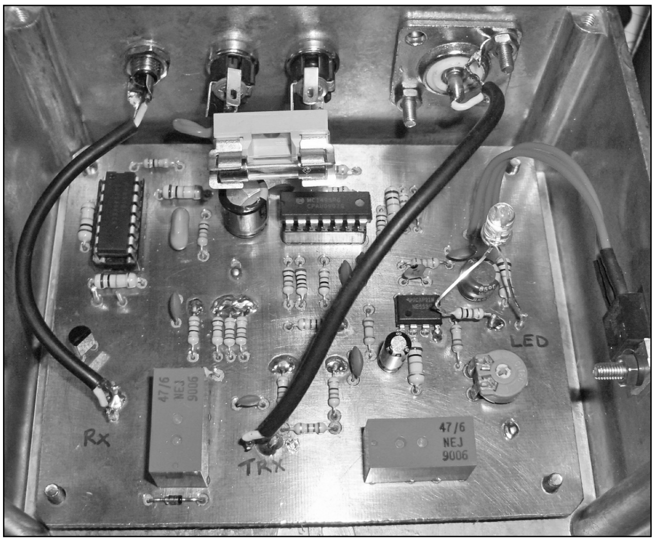


Fig 11.157: Linear transverter (built & photographed by M0DTS)

drawn, turn up R28 until the supply current is about 100mA more than you started, eg 120mA. This is akin to the bias setting on a conventional solid-state linear amplifier. It also gives adequate light for a distant receiver to line up on. Surprisingly, it is also left on when switched to receive, giving the distant station something to aim at.

Mods for optimum performance

If you intend using this on FM only, or over short distances, it performs perfectly well, but there have been some more recent modifications to the design above that squeeze a little more performance out of the system.

Several recent tests over long distances revealed an odd situation with receiver response. FM signals were end-stopping on the FT-817 S-meter, but when switched to SSB, the S-meter would not indicate more than S9 with equally strong signals. This was traced to the receive head (and transceiver, the same circuit on receive) putting out so much signal that it was causing the transverter to limit. Of course this is fine on FM, but it would be nice to have an equal response to SSB. The cure is to reduce the gain in the head by removing the second op-amp and connecting the output of the first op-amp via a 0.1μF capacitor to the output socket, this can easily be done on the existing two op-amp circuit board design. In the transverter the receive output attenuator was modified as follows: change C9 from 47pF to 1nF, remove R25 altogether, and swap R26 with R27. If the receiver S-meter is deflected by noise after this modification, simply fit a 220 ohm potentiometer as an input level control at the point where the Rx signal from the head connects to the transverter board. The signal on the pot wiper is fed to the Rx in port on the board. Adjust the potentiometer to the point where the S-meter shows no indication when the head is in total darkness. This modification slightly lowers the noise figure of the receiver and enables the S-meter to have full range on SSB signals.

Operating principle

It is easy to imagine what happens to an FM signal on transmit through the LED: the MOSFET is effectively in Class D and being driven hard, switching frequency modulated pulses to the LED at around 20 kHz.

It was problematic at first envisaging exactly what was happening with a single sideband signal, as only half of it will be conducted by the LED. This must be rather like putting a rectifier diode in series with your HF antenna! Only the positive half of the SSB signal gets converted into light (the LED is a diode after all). These half-signal pulses travel to the distant receiver. If you could pick this up directly, without a tuned circuit at signal frequency, it would sound like the most awful, overdriven distorted signal you have ever heard. But it does not sound distorted at all on an HF receiver because the tuned circuits in the receiver restore the waveform due to the flywheel action of a high Q tuned circuit.

Unlike many other transverters, there is nothing to tune up or adjust other than the LED bias potentiometer. Most who have built this have not even padded the crystal down to its design frequency (although there are spaces for capacitors on the board for this), just leaving it about 1 - 2kHz high is fine.

Do not exceed the LED current ratings. By monitoring the total transverter current, you can keep an eye on the average current through the LED. A 1A fuse is installed in the power supply line and it has blown several times on speech peaks.

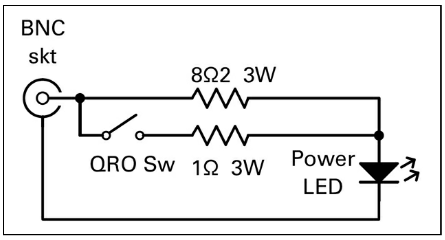
Transmit head

This consists of a power LED and switch mounted on a diecast box and current limiting resistors mounted inside the box. **Fig 11.158** shows the circuit diagram. In normal use the QRO switch is left open. When using SSB it is possible to close the switch, which lets a lot more current flow through the LED on speech peaks. Do it at your own risk - it gives about one extra S-point and several LEDs have been blown this way. Trying it on FM or CW is almost certain to blow the LED.

The LED is mounted on a 25mm square piece of 0.4mm fibreglass PCB on the base of the diecast box (see **Fig 11.159** and **Fig 11.160**). Power LEDs dissipate a few watts of heat so it is important to use heatsink paste between the LED and the board and again between the board and the box, which acts as a heatsink. Do not use normal-thickness PCB! The two power resistors are mounted in the box and a BNC socket is used for the drive connection. The box is mounted on a pipe-stop end with a central hole cut in it, then placed at the focus of a lens as described earlier.

An Osram Golden Dragon LR W5SM HYJY-1 LED, RS part number 665-6189 has been used here. It is quite important that you use this one because, although many others will work on transmit, It will be explained later how to make this particular LED

Fig 11.158: Circuit diagram of the transmit head.



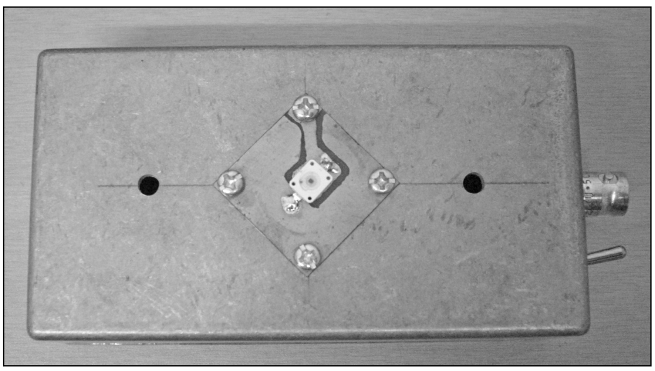


Fig 11.159: Power LED on transmit head.

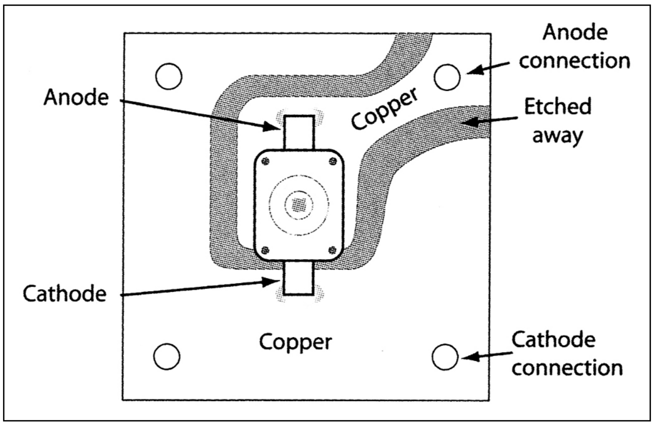


Fig 11.160: Power LED mounting

also work as a photodiode on receive. It is quite likely that the original LED will have been discontinued as the development of LEDs is proceeding at a rapid pace, so it will be a case of finding an equivalent and testing it.

Latest and most sensitive front end

Fig 11.161 shows the best front end used to date. It is closely based on a design by Clint Turner, KA7OEI, that is found on his website [64]. The first stage has been altered to be a reverse biased SFH2030 photodiode RC coupled to the gate of the FET. The relatively elaborate bias circuit is to make best use of the low-noise performance of the FET. Some of KA7OEI's low pass filtering has been removed to increase the bandwidth of the circuit which now responds up to about 100kHz. T3 forms a cascode amplifier with the FET while T4 controls the FET drain current. The low noise op-amp(s) provide further gain.

The following information on the 2N5457 FET front end bias applies to both the receive head and transceiver head to be described later. Variations in the characteristics of the 2N5457, bias generator MPSA18 and its diodes can cause an incorrect FET drain voltage. It is vital for maximum sensitivity that the drain voltage sits at around half the power supply voltage. If the drain voltage is higher than 6V, a quick fix is to connect a resistor from drain to 0V, which will greatly improve matters. Some constructors have used a 4.7kΩ fixed resistor whilst others have used a 10kΩ variable resistor and adjusted it until the drain voltage is at 6V. A more careful approach is to adjust the value of the resistor (currently 120Ω) connected to the drain of the FET. This has affected around half the circuits built so far.

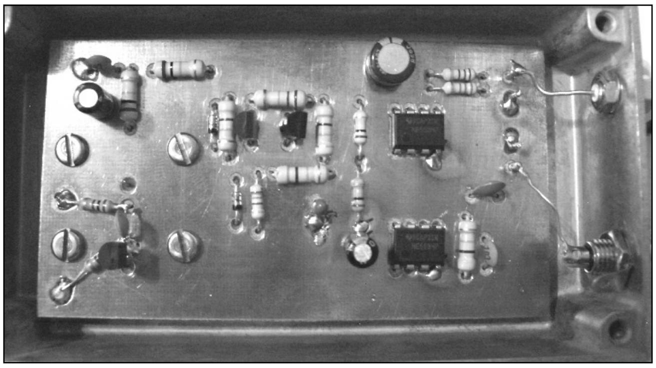


Fig 11.162: Receive head (two op-amp version)

This circuit when biased correctly is so exquisitely sensitive that there is no alternative to a diecast box for shielding: even the power is fed down a shielded cable from the transverter. Use a 3.5mm stereo jack plug, socket and screened cable to bring power in and take the signal back to the transverter on separately shielded cables. The SFH2030 photodiode just peeps out of a 6mm diameter hole in the box. This is normally placed at the focus of a lens on a pipe end cap, drilled as before. Any daylight reaching the photodiode will vastly increase its noise output so, while testing, keep the head, mounted in the pipe end cap, face down on the bench to exclude the light. Sliding it slowly to the edge of the bench instantly reveals when the tiniest amount of light enters because the noise level rises so much. You end up skulking around in the dark with this one! Without a lens, this head gives a noticeable increase in noise from moonlight and, with a lens, Jupiter and the brighter stars are easily detected.

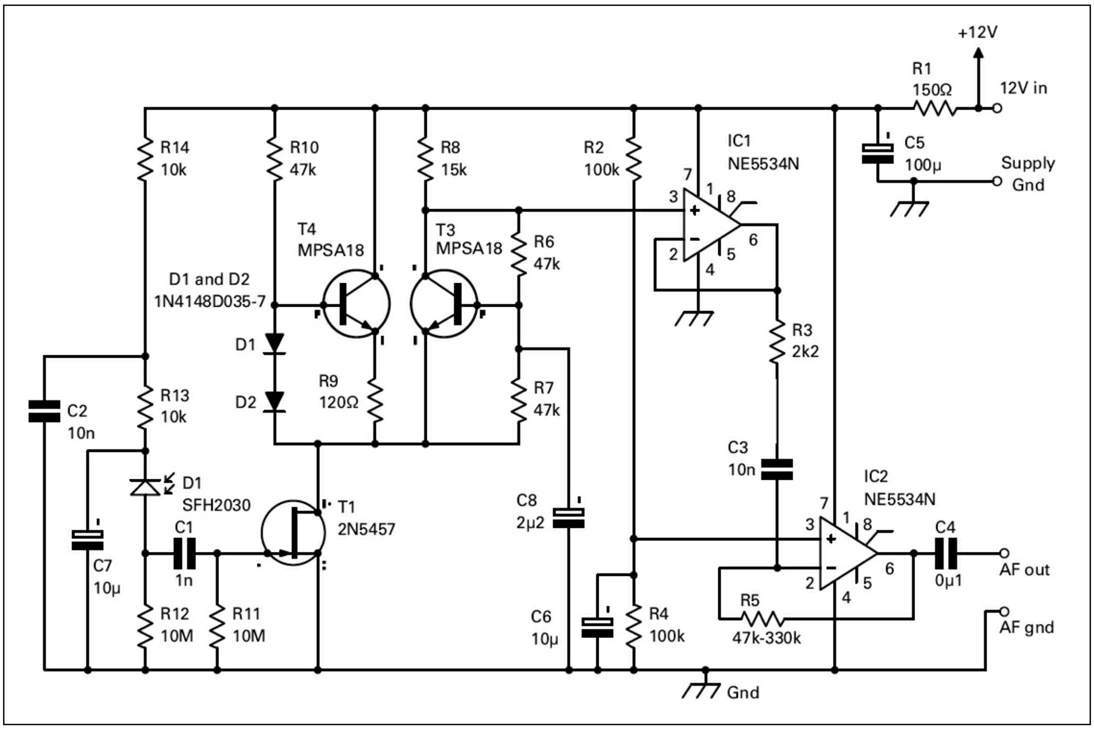


Fig 11. 161: Receive head circuit diagram (this is the two op-amp version)

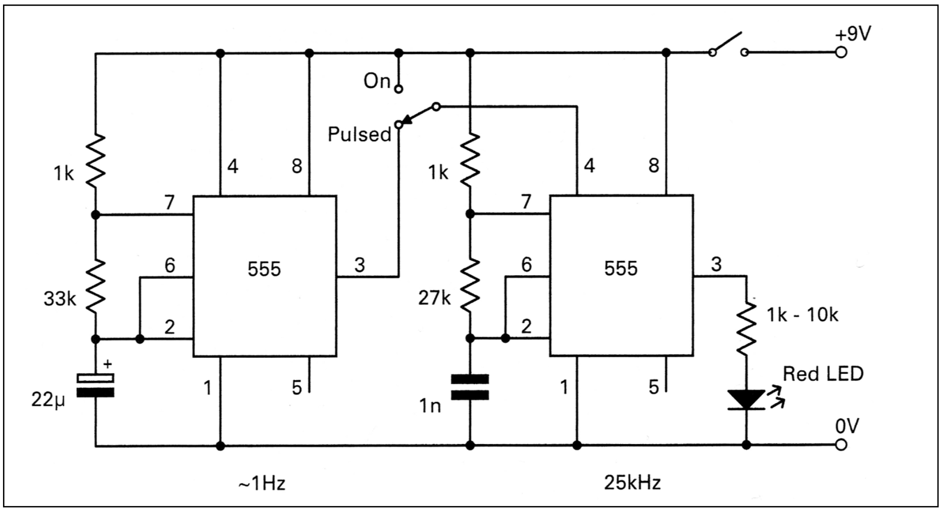


Figure 11.165: Circuit diagram of the test bench beacon

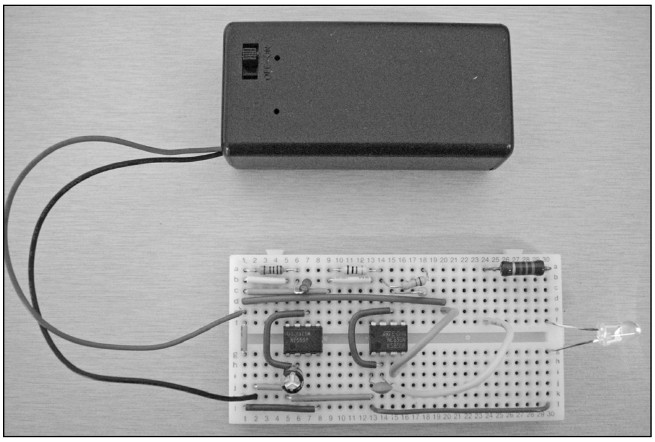


Fig 11.166: The test bench beacon, still on its prototyping board

Fig 11.162 shows the completed receive head, usable with one or two op-amps, with the PCB foil pattern and overlay in **Figs 11.163 and 11.164** (in Appendix B). As with the transverter, the PCB is double sided for screening; use the same drilling technique.

Test Beacons

It is useful to have a small beacon transmitter that can output an optical signal for bench testing. Nothing more complicated than a pair of 555 timer ICs is required, as shown in **Fig 11.165**. The right hand 555 oscillates at about 25kHz and drives a red LED via a series resistor. 1kΩ gives a bright light for long distance testing up to 500m or so; 10kΩ is quite dim and suitable for indoor use. This beacon tunes in as a carrier a little above 3.605MHz on the FT-817. The left hand 555 is optional; when switched in, it keys the 25kHz oscillator, making the signal more easily identified.

A plug-in breadboard version of the test beacon is shown in **Fig 11.166**.

Extreme beacon

One of the longer distance contacts, at 65km, was nearly a failure because the stations could not locate each other for some considerable time. White light from powerful torches looks just like car headlamps; red lights look like car tail lamps. Then a Xenon strobe was used, identification is much easier with a regularly flashing lamp but, understandably, it was looking a little dim at 65km.

Enter the extreme beacon, **Fig 11.167**. In principle, it is similar to the test bench beacon described above, but uses a power FET on the output to drive a 20W LED. This has 25 individual LED chips arranged in a 5 by 5 matrix on a substrate. It runs on about 12V at around 2A. The LED does require a substantial heatsink - reckon on dissipating about 15W as heat.

With this LED and its heatsink at the focus of a 100mm lens in the optic tube (**Fig 11.168**), it produces a 2° wide pattern of light flashing at either 2.5Hz (for visual identification) or 25 kHz for receiver alignment. Do not look directly into the beam: the intensity is enough to cause eye damage, even at some considerable distance.

The LED Transceiver

Tim Toast, who runs the Optical Links website [63], regularly scans the web for optical communication-related material. In October 2010 he provided a link to a paper, LED used as APD, written by a team at the University of Salerno, Italy. They had

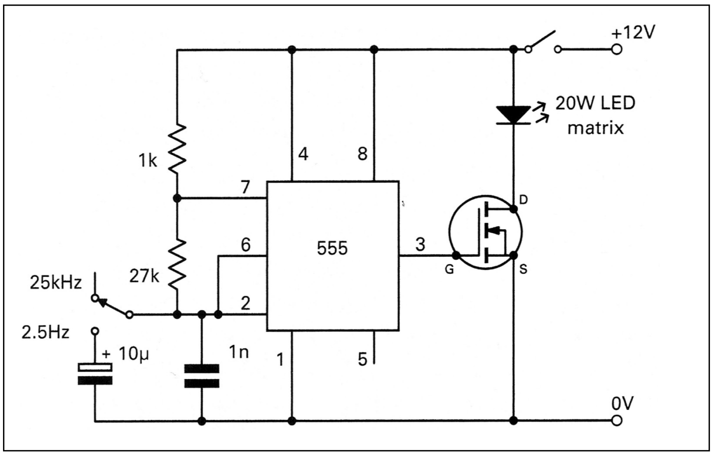


Fig 11.167 Circuit diagram of the extreme beacon

Fig 11.168: The extreme beacon LED mounted in one of the now-standard housings

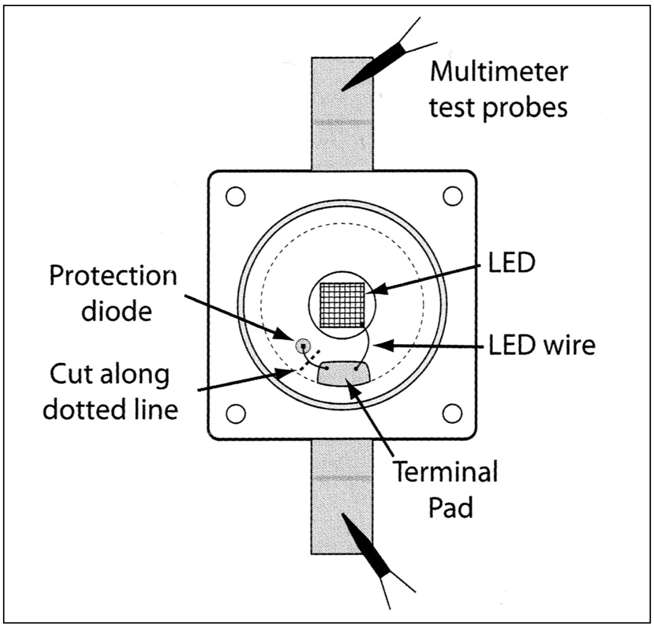


Fig 11.169: Where to cut the protection diode wire

discovered that some GaP and GaAsP/GaP power LEDs, when reverse biased to large voltages, acted as photo sensitive diodes and as avalanche photo diodes, the latter being very expensive.

After reading this paper, G8CYW set about to repeat their experiments. He used the test bench beacon, several power supply units, a few resistors and capacitors, a selection of red LEDs and an oscilloscope. It was found that a signal could be recovered from a particular high brightness 5mm LED without any reverse bias; then as the reverse bias was increased to over 30V, the recovered signal was enhanced. Another receive head was rapidly put together, with a relay to switch the LED from forward bias (so that It could be used as a transmit head), then to reverse bias it in an attempt to use it as a photodiode.

There is also one very obvious advantage using an LED in this way as a transceiver, you only need to aim only one set of optics: if you can hear the other station; you are ready to work the other station.

This low power LED transceiver was tested over a 6.5km path and found to receive and transmit as well as the separate units used before. This may well have been the first occasion where a single LED was used to receive a free space optical signal over a distance of several kilometres.

LED Transceiver Mk2

The concept of the LED transceiver was proved, but the light level on transmit was way down on that from the power LEDs. So, the Golden Dragon LED used in the separate transmitter was looked at again. It is an InGaAlP device, a type not covered in the original research paper. Also, it could not operate in reverse bias due to the presence of a protection diode in reverse-parallel with the LED chip. There is even a warning in its data sheet that the LED is not intended for reverse bias operation. It was noticed that both diodes were set in a silicone gel and the minute gold leads to both diodes were separately visible within the gel. Enter the concept of 'microsurgery' on the LED.

A sharp knife was used to cut through the gold wire to the protection diode by simply pressing the blade down vertically on the wire. Checking with a multimeter on the diode range indicated the operation had been successful: with the meter leads one way round the LED lit up and the meter indicated about 1.6V; the other way round was now open circuit, having previously indicated the 1.065V forward voltage drop of the protection diode. In short, the LED now has a dual use; one way round it is a LED, the other way round it is now a functioning photodiode. Several operations have since been performed and an LED has not been lost yet. **Fig 11.169** shows the LED, diode and cut point.

This LED encouragingly developed a photovoltaic output of over 1.1V when strongly illuminated. Using the test bench beacon as a signal source, it gave an enhanced output voltage when reverse biased to 43V. This was then the centre of a redesigned front end which incorporated the same modified KA7OEI circuit as before, plus relay switching to use the LED on transmit as well, as shown in **Fig 11.170**. The LED series resistors were also included on the board design to make the complete transceiver head.

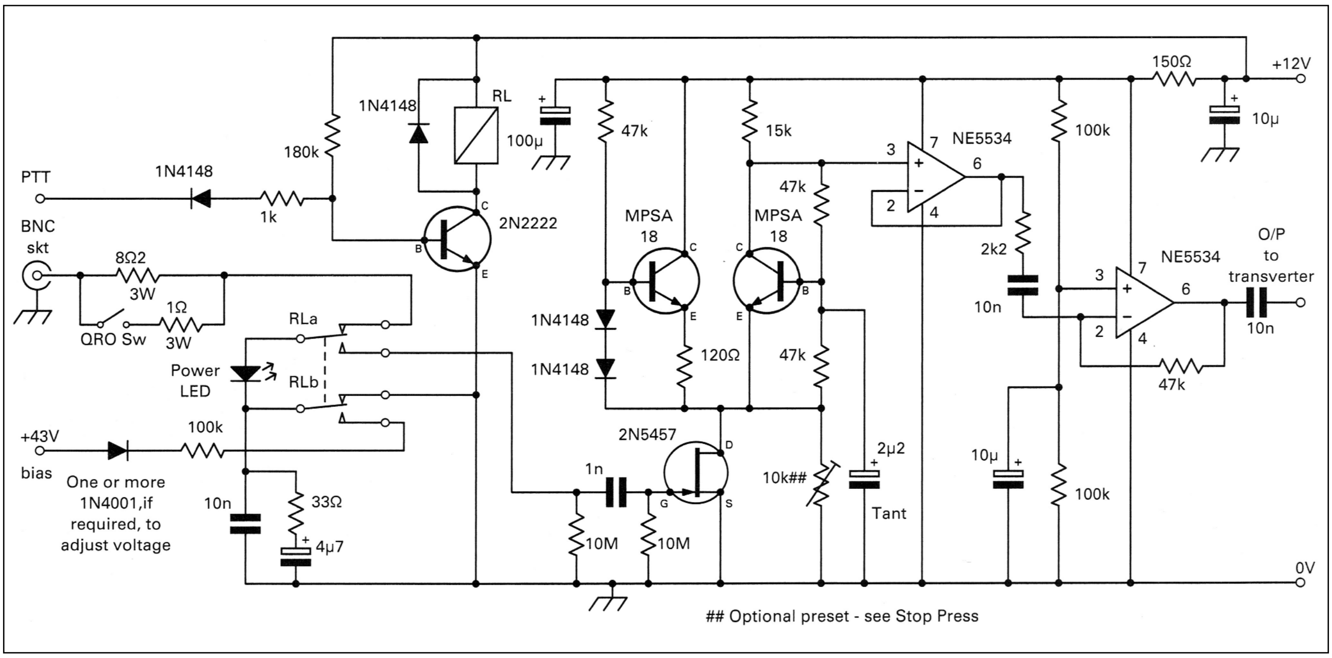


Fig 11.170: Circuit diagram of the LED transceiver head (two op-amp version).

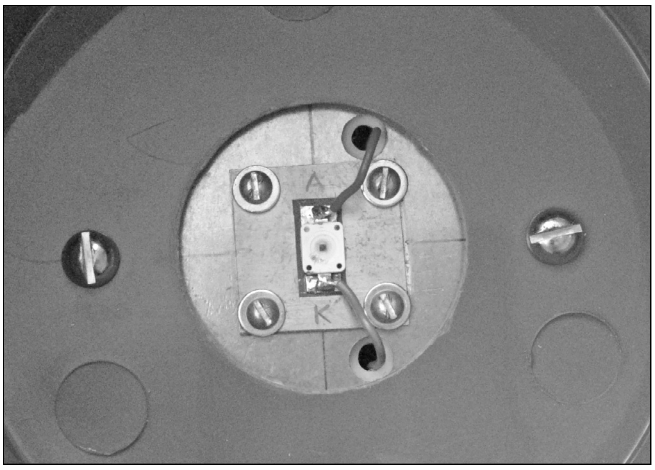


Fig 11.171: Transceiver LED mounted on the head assembly

The transceiver LED is mounted on a thin fibreglass PCB (**Fig 11.171**), this time in a design to minimise capacitance to ground on receive while retaining adequate heat conductivity on transmit. Use heatsink paste as before.

It is thought that the exact value of reverse voltage will be an 'adjust on test' item, as one version of the transceiver required 48V to bring it to optimum performance. The bias current is extremely low (about 100nA) and is switched off by the relay when not in use (the reason for the PTT arrangement, it defaults to transmit when not powered). A combination of 12V keyfob batteries and/or button cells is used to achieve the best voltage for the particular transceiver. You can adjust the actual voltage by placing 1N4001 diodes in series to drop the voltage if required: it seems that you need to hit the optimum voltage to an accuracy of about half a volt. Do not use zener diodes, as they would generate noise.

The batteries are contained within a piece of 15mm copper water pipe attached inside the box using Terry clips, as can be seen in **Fig 11.172**. Others have used N cell holders, which are a good match for the keyfob batteries. Don't try to use an inverter or voltage multiplier to supply the bias voltage; noise would drown out the wanted opto signal.

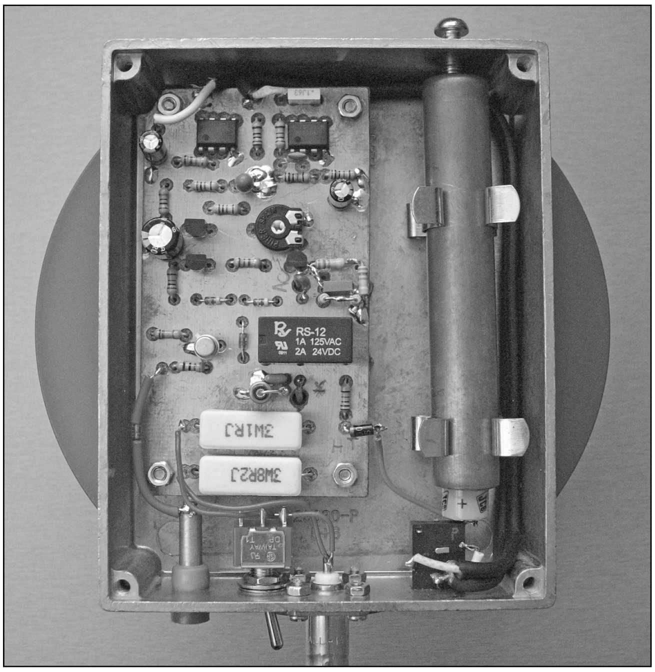


Fig 11.172: General arrangement of the transceiver (two op-amp version). The copper tube contains the 43V bias batteries

The PTT line can be connected to the FT817 (which grounds a pin on transmit) as well as the transverter PTT input and it should operate as required.

The Fresnel Lens Rig

This uses flat A4 page magnifiers made from acrylic sheet. They are available from stationery shops or, rather cheaper, from many pound shops. They are actually Fresnel lenses which are of quite good quality. They are not quite the same size as an A4 sheet of paper, being about 21cm by 28cm. Compared to the Blue Spot lenses that fit the plumbing pipes, they have over 8dB further gain when used on receive or transmit. This figure was arrived at by simple calculation of the increased area and confirmed using the FT-817 S-meter on receive and a light meter on transmit. The transmit beamwidth of the Fresnel lens system works out at about a fifth of a degree - which is sharp, but not impossible to aim. It is however too sharp to aim if you do not know exactly where the other station is, hence the extreme beacon (with its 2° spread) described earlier. Later versions of this also employed a secondary meniscus lens of 28mm diameter and 28mm focal length mounted close up to the LED (about 10mm from the LED). This extra lens captures more of the light from the LED and directs it toward the Fresnel lens. This extra lens broadens the beam slightly, to about 0.3 degrees, but the beam is still brighter as a result.

This Fresnel lens system was introduced before the LED transceiver was developed, so G8CYW's version has two lenses side by side (**Fig 11.173**). This makes it simpler to operate because the separate receiver and transmitter, aligned side by side, automatically look at the same point in the distance. It is simply necessary to line up the receiver on at a distant (beacon) signal and you're ready to transmit back. With the advent of the transceiver, it is only necessary to use a single lens, but the second lens has been successfully used to house the extreme beacon. This simplifies the alignment procedure when operating from a hilltop.

The Fresnel lenses tested have a focal length of 350mm, but it would be wise to test each lens individually by measuring the lens to image distance when producing a focused image of a distant streetlamp. The LED (or photodiode) is simply supported at this point. The ridged surface of the Fresnel lens must face towards the distant station. All these parts are mounted in a box, shown in Fig 11.173.

Fig 11.173: The Fresnel lens rig, suitable for a transmitter and receiver, or a transceiver and beacon

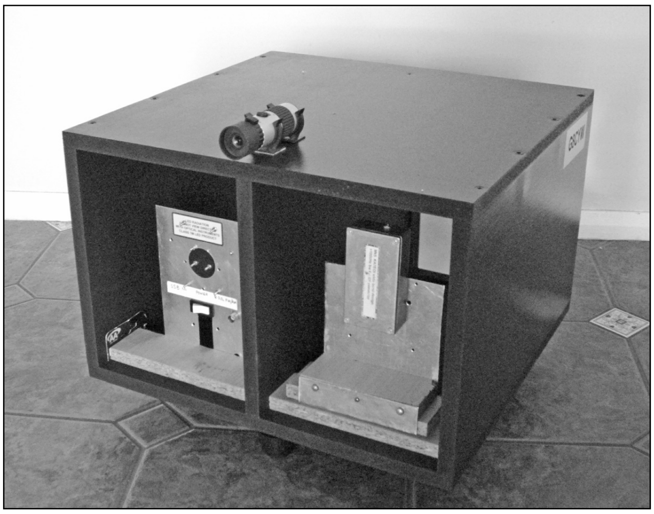


Fig 11.174: The Fresnel lens rig, rear view, showing sliding mounts for Rx, Tx, LED transceiver or extreme beacon

Two fixed cabinet feet are installed under the front of the black-painted lens box and one adjustable foot, on a screw thread, centrally at the rear. This gives adjustment in the vertical plane; for horizontal (azimuth) adjustment, it is simply necessary to move the whole box around on a square of MDF clamped on to a folding workbench or large tripod. This has proved adequate for contacts up to nearly 120km distance.

SSB has been proven ultimately win out over FM on weak signals as demonstrated when the UK distance record was increased to 117.6km, (M0DTS at Danby Beacon near Whitby and G8CYW near Alnwick, Northumberland), by operation over a difficult path over the North Sea. This path is technically non-line-of-sight, requiring help from atmospheric refraction. This gear seems to be a good addition to those who climb up hills for contests etc, to add 623nm (the wavelength of the red light produced by the LED), to the selection of wavelengths available for communication.

Infrared Communication

Analysis of several contacts has revealed that the maximum light level in which the system will operate giving the lowest noise level on receive is during nautical twilight when the sun is at least 12 degrees below the horizon. This places a constraint on the operating time in the summer months.

A recent development has been to use the near infrared band (IR), instead of visible red light. Several IR LEDS are readily available which produce radiation at wavelengths of 850nm or 940nm. Atmospheric data suggest that the shorter wavelength may be attenuated less over long paths, but both will be investigated. The LED transceiver is easily adapted for use with either type of IR LED, and the response as a photodiode (once the protection diode has been removed) to visible light seems to be reduced. The high bias voltage is not necessary and can be connected to the main LED transceiver supply of around 12V.

This is still at an early stage, although a first contact (believed to be a world first), between G8CYW and G8KPD at a distance of 6.3km has been made in full sunlight, with strong signals reported both ways. Both stations are currently constructing further optimised IR transceivers including extra optical filtering using readily available cheap long-pass IR filters intended for photography, which will result in a lower noise background on receive. The problems of operating in full daylight are akin to those on the higher microwave bands, accurate compass bearings and telescopic sights are only the starting point here.

FURTHER INFORMATION

A comprehensive list of further reading and useful component suppliers can be found on the CD attached to this edition of the *Radio Communication Handbook*.

REFERENCES

[1] CCIR, 1937.

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

[3] *International Microwave Handbook - 2nd edition*, editor - Andy Barter, G8ATD, RSGB 2008.

[4] *Microwave Projects, and Microwave Projects 2*, editor - Andy Barter, G8ATD, RSGB 2003.

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

[6] Avago (see box overleaf) Web: http://www.avagotech.com/

[7] Minicircuits. Web: www.minicircuits.com.

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

[9] http://www.kuhne-electronic.de/en/shop/147_Transverter/article:130_MKU_10_G2

[10] www.microwavers.org/

[11] http://myweb.tiscali.co.uk/g4nns/FeedHorn.html

[12] Maplin Electronics, www.maplin.co.uk

[13] Paul Wade, W1GHz HDL_Ant, - see www.w1ghz.org/antbook/contents.htm

[14] Farnell Electronics, http://uk.farnell.com/

[15] G3NYK: www.alan.melia.btinternet.co.uk

[16] ‘GHz Bands’ Column, *RadCom*, June 2008

[17] Sam Jewell, G4DDK web page, www.g4ddk.com

[18] RSGB RadCom Plus website at www.rsgb.org/membersonly/publications/radcomplus/index.php

[19] '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

[20] 'High precision frequency standard for 10MHz', Wolfgang Schneider, DJ8ES, *VHF Communications* 4/2000 pp177-90.

[21] Eisch-Kafka Electronics web: www.eisch-electronic.com

[22] Sam Jewell, G4DDK, sam@g4ddk.com

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

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

[25] RF Elettronica, www.rfmicrowave.it/

[26] ‘Matching an LDMOS power transistor at 1.3GHz with an output of 120W’; *Proceedings of Congress on VHF-UHF 2002; Munich Technical College, 9th. - 10th March, 2002*

[27] G H Engineering, The Forge, West End, Sherborne St. John, Hants, RG24 9LE, Tel: 01256 889295, Fax 01256 889294, email: sales@ghengineering.co.uk, web: www.ghengineering.co.uk/products/

[28] Terminal boxes from Farnell: www.farnell.com/datasheets/77332.pdf, I used the KL1508 400 x 300 x 120mm box.

[29] http://membres.lycos.fr/f1rhr/Txatv/PA1G2mod.pdf

[30] www.qsl.net/dl4mea/13ss/13ss.htm

[31] www.nr6ca.org/spectrian.html

[32] ON4IY, www.qslnet.de/member/on4iy/spectrian.html

[33] 'GaAsFET power amplifier stage up to 5W for 10GHz', Paeter Volg, DL1RQ, *VHF Communications* magazine, 1/1995, pp.52-63.

[34] '6cm transverters in modern stripline technology', Peter Vogl, DL1RQ, *VHF - UHF Munich* 1990, pp49-66.

[35] '6cm transverters in stripline technology, part 2', Peter Vogl, DL1RQ, *VHF Communication Magazine* 2/1991, pp69-73.

[36] 'Determination of parasitic inductances on capacitors', Roland Richer, *Special project at Dominicus vo Linprun Grammar School Viechtach*, 1994.

[37] '76GHz amplifier', Sigurd Werner, DL9MFV, *VHF Communications Magazine*, 3/2003, pp163-169.

[38] 'Medium Power Amplifier @ 77GHz', Infineon Technologies Data Sheet, T 602B_MPA_2; 05.05.99.

[39] Data Sheet IAF-MAP7710, Fraunhofer Institut for Applied Solid-State Physics, Freiburg.

[40] 'Frequency multiplier for 76GHz with an integrated amplifier', Sigurd Werner, DL9MFV, *VHF Communications Magazine*. 1/2002, pp35-41.

[41] 'Amplifier for 47GHz using chip technology', Sigurd Werner, DL9MFV, *VHF Communications Magazine*, 3/2002, pp160-164.

[42] 'A simple concept for a 76GHz transverter', Sigurd Werner, DL9MFV, *VHF Communications Magazine*, 2/2003, pp77-83.

[43] 'Combining power at 76GHz: Three possible solutions discussed', Sigurd Werner, DL9MFV, *VHF Communications Magazine*, 1/2004, pp13-19.

[44] Hittite Microwave Corporation, 20 Alpha Rd., Chelmsford, MA01824, www.hittite.com

[45] FM ATV standard for IARU region 1, Lillehammer conference report 1999

[46] Convenient procedure for filling VX-2 memories, *CQ-DL* 10/04, DARC-Verlag, Baunatal, 2004

[47] Frequency input module for 10GH ATV transmitter module, Alexander Meier, DG6RBP, *VHF Communications Magazine*, 4/2005, pp217 - 221

[48] Dubus www.dubus.org/

[49] Watkins Johnson. www.triquint.com/prodserv/more_info/default.aspx?prod_id=FP31QF

[50] www.qsl.net/ct1dmk/reflock.html

[51] '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.

[52] 'A 1W linear amplifier for 1152MHz', Sam Jewell, G4DDK, *RSGB Microwave Handbook Volume 2*, pp8.21-8.23.

[53] Richard L Campbell, KK7B, *Proceedings of the Microwave Update*, ARRL, 1988.

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

[55] Richard L Campbell, KK7B, *Proceedings of the Microwave Update*, ARRL, 1989.

** Devices from Agilent, Avantek or HP

Originally Avantek Semiconductors manufactured the devices. Hewlett-Packard bought finished units (for example LNAs and VCOs) from Avantek for its range of test equipment as well as discrete transistors. Many of these were custom items although still bearing the Avantek logo. As the volumes supplied grew Hewlett-Packard purchased Avantek and the devices then became branded as HP ones. A little later during the HP reshuffle of its markets and operating divisions, the decision was made that the HP name would only be used for the computing systems, and a new name Agilent was coined for the semiconductor division and other allied divisions. In 2006 the name of Agilent was being used both for the old HP test equipment range and the semiconductor devices. HP decided that Agilent would only be used for the test equipment divisions and so Avago Semiconductors was invented for the semiconductor divisions.

For information on devices shown in this chapter as originating from Agilent, Avantek or HP see http://www.avagotech.com.

[56] Down East Microwave, www.downeastmicrowave.com

[57] 'VHF and microwave applications of monolithic microwave integrated circuits', Al Ward, WB5LUA, *ARRL UHF/Microwave Experimenter's Manual*, ARRL, 1990. pp7.32-7.47.

[58] *Microwave Handbook, Vol 3,* edited by M W Dixon, G3PFR, RSGB, 1992, pp18.116-18.118.

[59] 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.

[60] 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.

[61] '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.

[62] 'A simple laser communications system', David Bowman, G0MRF. Web http://g0mrf.com/laser.htm

[63] Tim Toast: www.aladal.net/toast/comlinks.html

[64] KA7OEI Optical Comms: www.modulatedlight.org/optical_comms/optical_index.html

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