Chapter 14

Surplus

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Building Blocks

Compact Power Supply for the Qualcomm Omnitrak Amplifier

Dave Robinson WW2R

I thought it was time to redo my 10GHz portable unit to increase power without increasing the size, so that it could continue to be mounted at mast-head. It was decided to use the Qual-comm Omnitrak PA (ref 1) with the DB6NT original transverter design (ref 2), G4DDK004 oscillator and the G3WDG preamp (ref 3).

There was insufficient space in the box for the Omnitrak PSU board that came with the amp, especially as its TX/RX switching was already provided in the DB6NT module. I therefore needed to devise a small PCB, which could provide 10V at 1A max and –5V at around 50mA.

For the positive rail, the LT1086 1.5A low volt drop regulator would be used again (ref 4). However the usual 7660 chip was not capable. Looking through some regulators data sheets (ref 5) I came across the LT1054 chip which had many advantages over its predecessors: -

1 .It is capable of providing 100mA (7660 max is around 20mA)

2. It is rated for an input voltage of 15V maximum so can be run directly off 12V supply (7660 maximum is 10V)

3. Most importantly, it has an internal regulator that regulates the negative output voltage. (You don't need to worry about the 120 ohm internal resistance of 7660)

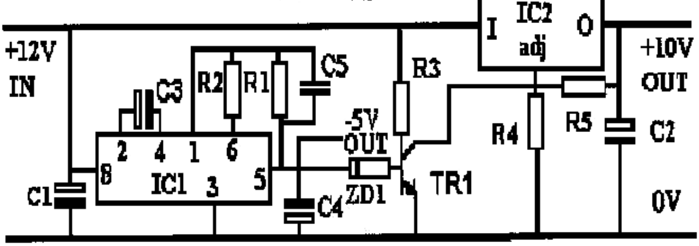
Accordingly, the circuit shown in **Fig 1** was devised. Failure of the negative supply reduces the positive supply to around 1.2V. The component listing is shown in **Table 1**. The PCB layout is shown in **Fig 2** and the overlay in **Fig 3**. Note that the components are mounted on the track side of the board and that pin 7 of IC3 is chopped off and not soldered to the track below it. Surface mount resistors and capacitors can be used on the board. The PCB was mounted on the back of the ampli-fier casting (the back of which had been milled flat) taking care to insulate the metal tab of IC2 from ground.
The output voltages of the 3 modules built so far were measured as 9.82 and –4.89.

Purists may be tempted to adjust the resistor values slightly to achieve exactly 10.0V and –5.0V. I tried it, but doing so made no difference at all to the Qualcom's 1.1W output power, probably due to its on board bias regu-lators.

References:

1. http://www.ntplx.net/~wz1v/w1ril.html
2. Dubus Technik III pp324-333
3. Dubus Technik IV pp276-339 (also Proceedings Microwave update 1991 and 1995)
4. "Using the California Microwave 11-026700 transmitter assembly" Micro-wave update proceedings 1995
5. http://www.linear.com/prodinfo/dslist.html

FIG 1.CIRCUIT



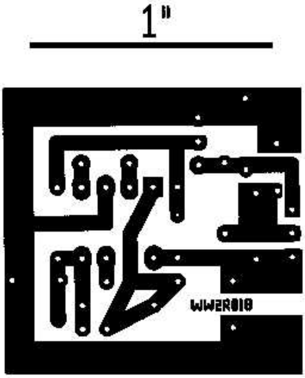


FIG 2. PCB LAYOUT

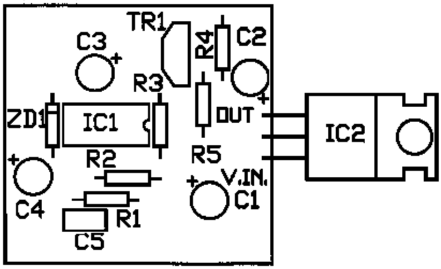


FIG 3. COMPONENT OVERLAY

Table 1: Component list

Resistors	Caps	Semiconductors
R1 100k	C1 10uF 25V	IC1 LT1054
R2 20k	C2 10uF 25V	IC2 LT1086
R3 10k	C3 10uF 25V	ZD1 4.7V 0.4W
R4 820R	C4 100uF 10V tantalum	TR1 2N2222a
R5 120R	C5 2200pF	

Sub 2dB Noise Figure for the Drake 2880 Downconverter

David Bowman, G0MRF

A satellite band Drake 2880, fitted with Toko filter, MGA86576 and 22pF caps in the I.F, 1.94dB noise figure and 38dB conversion gain…. interested ? - Then read on…

There have been a series of modifications detailed for the Drake 2880. The following improvement was tested at the AMSAT Colloquium held at the University of Surrey in July '99. This procedure is only suitable for Drakes operating in the Satellite sub band of 2400 or 2450 MHz. A second Drake modified was within 0.01dB ! This modification is consistent.

It is assumed that the following standard changes have been carried out:

1. Crystal changed and IF coils and chip caps removed - JN1GKZ
2. 2 x 22pF added to IF - G0MRF
3. Front end FET replaced with Hewlett Packard MGA86576.

After these modifications you should already have improved the performance to beyond that of the original manufacturer's specification. 2.4dB noise figure with a 28dB conversion gain was the average measured over a two day period.

Assessing the circuit for a next step, I identified the image filter as a source of excessive signal loss. The 86576 under optimum conditions should be able to produce a NF of 1.6dB. The Drake's image filter has around 4dB of insertion loss and this is undesirable when the front end is a single device. Additionally, we also saw some interaction between the stripline filter and the small cover which isolates the RF / IF parts of the circuit. In some cases we measured up to 2dB difference in noise figure between "lid on and lid off". So the stripline filter had to go!

Finding a replacement wasn't too difficult. A couple of years back I designed a 2.4 GHz to 432MHz converter which used a Toko dielectric filter. The filter is centred on 2450 MHz and is used in wireless local area networks (LANs) It has a pass band of +/- 50MHz but more importantly, the insertion loss is only 1dB.

To change the filter requires a little surgery with a modelling knife, the removal of the 3 striplines, fitting 2 ground pins and the repositioning of the mixer coupling capacitor.

Procedure

1. Desolder mixer coupling capacitor which is close to the output stripline.- Keep for later!
2. Cut across the input stripline with a sharp knife. This extends the 50 Ohm track which feeds the filter by about 3mm (The width of the stripline)
3. Cut the track between the output stripline and the mixer in the position shown in the photo (see next page).
4. Remove unwanted parts of the striplines by heating with a soldering iron and lifting the copper clear of the board.
5. Position the filter on the board and carefully mark the positions for the two ground pins.
6. Drill the PCB with a 1mm drill ensuring that the drill also marks the die cast case under the board.
7. Remove the board and fit / solder the pins from the top. Cut flush on the underside of the board.
8. Take a larger drill, 4 or 5 mm, and look for the marks in the casing. re-

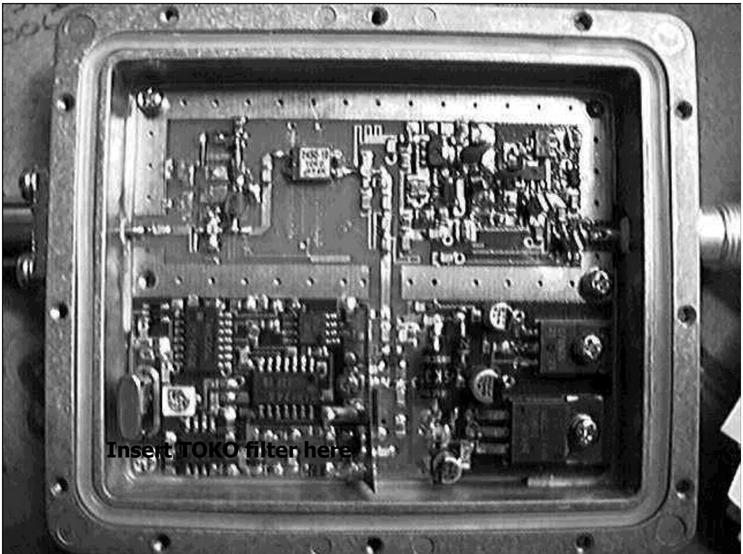
move a small amount of metal from the casing to allow any protrusion from the pins to be recessed
into the case. - This allows the board to lay flat when refitted.

9. Refit board into case. Solder filter to 1 track and the heads of the two pins.

10. Solder a coupling capacitor directly to the filter output terminal and to the track feeding the mixer. Replace all covers / screws.

You now have one of the hottest Drakes around!

THE DRAKE 2880 DOWNCONVERTER



DRAKE1.JPG

A Checklist for putting the Toshiba 20 and 40 watt 3.4GHz Linear Amplifiers on the Air

John Jaminet, W3HMS

Introduction
This paper was created from several messages received on the Internet Lists in September/ October 2001 by Steve, **N2CEI**, at DEMI, Rick, **K1DS** and an article in the MUD 2003 Proceedings, pages 66-70, by Dave, **WW2R**. Important new info was received from Owen, **K6LEW** in November 2001 on the importance of voltage regulation. It is incorporated in the text. In addition, our own **N3TWT-ATV** experiences with 24/7 operation in FM ATV (key down for months on end!) service are reflected in this update.

Basic Information
There are **two** versions of this amplifier:
a. 20 Watts output @ 3.4GHz with 1mW input.
b. 40 Watts output @ 3.4GHz with 1mW input.
Supply Voltage: + 12.6VDC (exactly)

Vendor Info on the 20 Watt Amplifier
This is a Toshiba 20W Linear Microwave Amplifier for use in the 3.44 to 3.68 GHz range. It is sold in the original manufacturers packaging.

Heat sinking: Required…it gets hot..
Size: 5" x 8" x 1 ".
Weight: About 2 pounds.
Input power: For full output, this is about 1 milliwatt or 0dBm.
Specifications: The specs on these amps are very "tight" and are typically as follows:
Linear Gain = 42.5dB; Gain: Ripple = .1dB
Linearity is -45dB; Return Loss (in and out)= 25dB
DC Power Supply current: Typically +12 VDC at 10.6 amps.
Power Jack: The DC power and control connector is a DB-15 male.
DB-15 Pinout: There are 3 rows of 5 pins each.

The ground is made from pins 3,7,10,11 connected together.
The +12VDC lead is made from pins 1,2,12,13 connected together All VCC and ground pins need to be connected to handle the 10.6 amps.
Pin 9 is the enable pin (TTL) which must be connected to ground in transmit to switch the internal power supply on.
When pin 9 is not grounded, the 12VDC supply draws about 15mA. and the amp is in stand-by mode.
This is a Class A amp and as a linear amp, it will draw about 10.6A with no signal input. Pins 4,5,8,15 are assorted alarm output pins low true….no more is known.

The 40 Watt Amplifier
This is a new Toshiba UM2683A 40W Linear Microwave Amplifier for use in the 3.4 to 3.6GHz range. It is sold in the original manufacturers packaging. This amplifier differs from the "2683B" "20W" version on other auc-

tions because there is a TMD0305-2 MMIC instead of the discrete circuitry in the front end. The TMD305-2 part is a 3.4-5.1GHz amp with 2 watt output and 22dB Power Gain. Turning the two pots at the far left on the lower board in the photo fully clockwise (shutting down the attenuator) and peaking the power with the third pot (2nd and 3rd stage bias) yielded 46dBm, 40 watts @ 3.456GHz using 12.0-12.6VDC (as measured at the connector) with about 0dBm input power.

Power supply requirement is 12.6VDC @ 15amps after readjusting gain and bias.

Size

The size of these amps is 5 x 8 x 1 inch and weight is almost 2 pounds. Input power required for full output is about 0dBm. DC power and control connector is a DB-15 male.

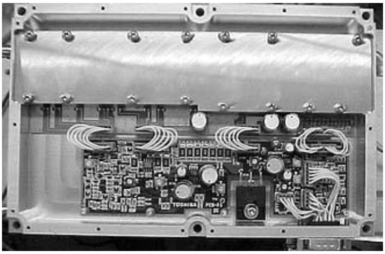
Pinout is as follows:

Ground are pins 3,7,10,11; +12VDC are pins 1,2,12,13;
All VCC and ground pins need to be connected to handle the 15 amps.
Pin 9 is the enable pin (TTL) which must be grounded to switch the internal supply on.

When not grounded, the 12VDC supply draws only about 15mA. and the amp is in a stand-by mode. Since this is a linear amp, it will draw about 14-15 amps without signal input. Pins 4,5,8,15 are assorted alarm output pins low true…no more is known. Only new amps are being shipped.

Heat Sink Requirements

These amplifiers get VERY HOT! Heat sinking is required, particularly for FM ATV service many hours per day. We have found that the heat sink should be about 2.5 or more times the size of the amp and be blown with two high cfm blowers of no less than about 85 cfm capacity. We use two 12VDC blowers for twice the air flow and for redundancy in case of the failure of one



blower.

In physics, we learned that heat rises and so it does indeed! Resist the temptation to mount the fins anyway but pointing UP as we have seen some disastrous results with the fins pointing down.

Even with fins up and 2 blowers we find the amp runs about 20 degrees higher than the temperature on the amp chassis, even in winter. We have mounted a small RS dual indoor and outdoor thermometer permanently on the chassis. We use the outdoor probe on the hottest part of the amp and heatsink. As I write this, an amp under test is showing 75F from the chassis probe and 55 F on the chassis. In summer, in a non-air conditioned space, the difference can be 30 degrees F. The results of poor heat removal have been small fires at the interior of the final amp shielding as revealed by the charred parts of two amps. Photos are available.

We recommend you use a very stiff power supply designed for 24 hour per day operation with at least twice the normal load of 15 amps. In our case this is 30 amps adjusted to give 12.6VDC on the amp terminals. Note that some commercial power supplies arrive giving 13.6VDC output. In some cases, the output is adjustable internally by a pot that is tough to find and adjust; a design indeed by Murphy, de Sade, or the village idiot.

Overall System View

When the amplifier is integrated into a

system such that relays, etc. are involved, the TX mode requires that the amplifier Pin 9, the PTT, be grounded for the amplifier to amplify, i.e. it turns on the power supply built into the amplifier. For RX, removing ground from Pin 9 places the amplifier into standby, drawing perhaps as much as 15mA, while the T/R relays all reposition for receive.

There is no feed-through via the amp. Changing from RX to TX involves just grounding Pin 9.

There are no relays internal to the amplifier for switching RF between RX and TX, hence, external relays for switching between RX and TX are required.

Critical Voltage Levels

Please refer to Step 9 (below) which is VERY critical. These amplifiers, especially the 40W version, will go into "fold back" and be hard to recognize in so doing. The 12.6 VDC +/- 0.2 is critical because the output FET regulator has an extremely narrow window due to the heavy current drain.

If you allow your primary voltage to get outside of this window, the FET bias voltage , as adjusted by R150 as measured on Pin 1 of the regulator interconnect to the amp section, you will note the need for drive in excess of 0dBm for 40dBm output (40W). (W3HMS used about 950 microwatts drive for 40 watts output). Excess drive is not good and indicates fold back which generates excess heat causing the "final" to generate more heat than it should.

These amps should draw 15 amps key down, not 14 amps, though some do vary. The set-up is simple enough but it has been necessary for several amps to be readjusted. This is because some operators did not see the importance of the relationship between the primary voltage at 12.6VDC

In summary

This is to say that R150 adjusts for 10.3VDC ONLY when the primary voltage is 12.6VDC with the amp under load (Pin 9 grounded) and the amp is drawing current.

If 12.6VDC in either direction is exceeded by about 0.3 - 0.4 VDC, adjusting R 150 will NOT regain 10.3VDC.

Thus, make absolutely certain you have sufficient heat sinking mounted to the amp for any tests or operation. The devices in this amp are extremely expensive and will not accept much heat. This amp gets VERY hot very quickly without a heat sink attached. One could also use a muffin fan blowing across the heat sink. One of the "alarm" functions available on the subminiature DB 15 appears to provide a "temperature" alarm. It is not known what the temperature should be when this function is energized but there is a voltage on one of the alarm lines that appears to relate to an increase in temperature. This amp does not self protect, i.e. shut down, so be very careful !

Perhaps the voltage could be used to control a relay which would remove PTT-ON from pin 9 but we do not know how much current it can handle. It may not be enough to control a relay, but it might control a NPNB transistor which could in turn control a relay, but this is not sure.

Output Power Monitoring

In an NTMS Feed point "article, also cited in MUD Proceedings 2003 on Page 66, Dave Robinson, WW2R, found a terminal in the amp which provides 5VDC equal to 40 watts output. This terminal is in a 6-pin connector at the right end of the amp, 3rd pin from left, when the shielded amplifier box is above the connector.

Another way to say it is the output connector will be on your right and the input connector on your left. This is the forward voltage, + voltage where 5VDC = 40 watts output. Doug connected a wire to an empty pin on the DB15 connector. I chose to drill a hole on the closest wall in a clear spot for a feed-through capacitor and then connected a 0-10VDC meter via a shielded cable.

Conversion Steps

♦ Obtain power plug DB-15 male at RS or other. RS # is 276-1502 for about $2

♦ Find and install heat sink size about 5" by 8" (size of amp) or larger and muffin fan if desired

♦ Calculate and obtain the attenuator value needed if you have more than 1mW output from your rig. FYI, the DB6NT units have about 200+ mW output and I will use 24dB of attenuation. ♦ Connect DB 15 power plug pins # 3,7,10,11 to ground.

♦ Connect DB 15 power plug pins 1,2,12,13 as the + 12.6VDC lead usable at 15 amps load.

♦ Connect a switch or relay to ground pin #9 for XMT ♦ Install antenna relay and necessary DC power

♦ Remove cover to see the 4 pots R217, R210, R136 on 20 watt model, R138 on the 40 watt model, and R150. These first 3 are the first 3 counting from the left and R150 is directly to the right of the voltage regulator

♦ Counting from left, turn pots R217 and R 210 fully clockwise

♦ Connect exactly + 12.6VDC power and power up

♦ Measure +12.6VDC at pin 4 on P6 to ground

♦ Connect input signal on 3.4561GHz (3.4001 in Europe) at the level of 1mW or 0dBm

♦ Monitor supply current and regulate to 15.0 amps with R138 on 40W model and R136 on 20W model

♦ Connect up the antenna or dummy load using good quality N or SMA fittings*

♦ If needed, adjust R150 for 10. VDC (CRITICAL) on right hand pins of P3 and the 10.3VDC from the regulator controlled by R150 and measured at Pin 1. This 12.6VDC should be measured inside the unit with the cover off while the amp is under load, i.e. key down or Pin 9 grounded on the sub miniature DB 15.

Given that this pin is hard to use when the amp is fully set up, we found that we had about 0.09 VDC drop under load between the terminal strip on our chassis and internal pin 1 of Plug 6. Knowing this, we now do set up by putting 12.66VDC under load on our terminal strip. With the amplifier off the air, thus no load, we see 13.05VDC on our terminal strip.

NOTE: You do NOT need any RF drive for this set up as the amp runs in Class A. If you have a power supply with remote sensing use it. If it does either voltage or current drain sensing, use voltage sensing. The voltage MUST stay steady at 12.6VDC or something will surely be damaged, most probably the regulator, to be followed shortly thereafter by the costly output FET.

It is not necessary to monitor the drive level if you have reliable and repeatable equipment for the "exciter". Once you establish 0 dBm at the input port to the amplifier it should stay adjusted. If somehow the equipment is not reliable as to its output, and there is time to adjust, then monitoring the input level would be necessary.

Most modern transmitting equipment of today is reliable such that once power levels are set, the input level to the amp should remain set.

In an NTMS Feed point "article, also cited in MUD Proceedings 2003 on Page 66, Dave Robinson, WW2R, discussed his experience with this amp. In one tune up case of a new amp, we could not find the voltage controlled by

R150 as addressed above so we followed Dave's advice to tune R150 and R138 for maximum output.

One could easily think about installing two VDC measurement jacks and make R150 an external pot so as to monitor and adjust both the 12.6 and 10.3VDC (at the VR) levels. No, the better choice is to monitor the output from a known and acceptable starting level. If you see the output begin to vary, then it is time to check the input level.

To clarify this point, R150 adjusts the output of the regulator BUT it does so in a very narrow window based on the primary input voltage of 12.6VDC and the amount of current being drawn and R138 will adjust this on the 40W version. ♦ Peak the power output with the third pot (This is R136 or R138) to yield 40 watts output

Testing

We have found that the output connectors can become quite HOT if the SWR is higher than desired, often too hot to touch! N fittings are desired but SMA fittings may simply have to be used per your individual needs. We recommend a careful "touchy/feely" exam even as little as 3 minutes after the on the air test commences.

Using the California Microwave 11-026700 transmitter assembly

Dave Robinson WW2R, G4FRE

Introduction
A considerable number of the California Microwave "Transmitter assembly 11-026700-08(h)" have been seen at rallies over the past 5 years. Many people have bought them, but it is hard to believe this, judging by the scarcity of "on-air" models. This paper details how to get them going on 3456MHz along with details of one method of incorporating it into a transverter and will hopefully encourage some to dust them off and help populate the band.

Equipment Description
In commercial use by AT&T for 3.7 - 4.2GHz microwave links, the amps were powered from positive ground 24Volts via a rack mounted California Microwave "Power supply 52-090095-0". The Power Amplifier unit consists of a 12.5 x 5.5 x 1.5 " aluminium can, containing the electronics, attached to a 13 x 6.75 x 1.5 " heatsink. The input connector is SMA, the output is via waveguide, but this is easily converted to SMA. A 6 core cable terminating in a 6 pin plastic plug supplies the operating voltages from the PSU.

The Power Supply unit consists of a 19 x 6.5 x2.5" box with one 2 pin connector, for applying -24V and a 6 pin connector for outputting the regulated voltages to the PA. It provides outputs of +10.25 and -5.3 Volts. It has an on/off switch and 4 test points.

The PSU is bigger than the PA, in addition it's -24 volt requirement makes using it portable a problem.

The first problem to be tackled therefore is to build a PSU to build the amp off 12V. In its original state the amplifier's performance is poor at 3456MHz and will need retuning.

Power Supply Unit
If the 6 screws securing the aluminium can of the PA are removed, along with the 6 smaller screws securing the aluminium screen below, it will be seen that the 6 core cable is connected as follows:

Orange and Purple	Drain Supply
Brown and Blue	Gate Bias
Grey	Ground
Red	rectified RF Sample

The orange and purple wires supply +10.25V at 1.7 and 1.5A respectively from isolated sources in the PSU. The 10.25V requirement precludes the use of the 78H series of regulators when running the PA off a lead acid battery as they require a minimum of a 2.5 Volt drop across them. A better idea is to use one of the LT108X series of low dropout voltage regulators which typically require 0.5 to 1V across them to maintain regulation. A single LT1084 regulator, which has a short circuit current of 5A is used.

The brown and blue wires supplying the -5.3V need around 10mA with the 10.25V supply on , 20mA with no 10.25V applied. This can be supplied by a LT1044 IC. The DC-DC converter chip is supplied from a 7808 regulator. The more usual ICL7660 is not used as it has been found more prone to expiring. The output of the LT1044 feeds an LM337T adjustable negative voltage regulator set to give -5.3V. The resistors around the LM337 have been increased, without adverse effect, from the usual recommended values, to reduce the current loading on the LT1044 IC.

It was decided to include extra circuitry to protect the amplifier from the loss of the negative rail, which has been known to destroy the amplifier. The circuit used COMPLETELY removes the positive 10.25V supply in the absence of the gate bias, rather than the more often published design such as (1) and (2), which only shuts down the regulator to 1.2V output.

The complete circuit diagram is shown in Fig 1. A PCB was designed for the circuit, excluding the components for the 10.25V regulators which are mounted directly onto the heatsinks, and the power relay. The PCB layout Fig 2. and the component overlay in Fig 3. The component listing is shown in Table 1.

Correct operation of the PSU should be ensured before proceeding. Normally the relay should operate and +10.25V and -5.3V should be measured at the appropriate points. If the -5.3V rail is removed (for example by disconnecting pin 8 of IC2) the relay should drop out and +10.5V rail should disappear.

Power Amplifier

Firstly the output connector has to be changed, to allow the usual RF connectors to be used on the amplifier. Carefully unsolder the brass pin from the output track of the amplifier. With a hex wrench, undo the 6 set screws holding the waveguide assembly in place and the four screws holding the tube to the edge of the PCB. It is possible, with a careful saw cut to leave part of the waveguide assembly with the 5 feed throughs and the earth lug and 4 mounting holes. For the faint hearted the safer option is to unsolder the 6 wires connected to the bracket and remove the whole assembly. This is replaced with a short length of 0.5" aluminium angle fitted with 4 bolt in feed throughs and a ground lug. (The brown and purple wires are connected to a common feed through). This method also shrinks the amplifiers area, allowing a changeover relay to be connected directly to the output connector.

The mounting plate left on the edge of the PCB has the correct fixing centres for a 4 hole SMA chassis mount female socket of the type with the long centre pin and the extended ptfe insulation. The centre pin and insulation are cut to length. The two holes for the lower screws are tapped and four long screws and 2 nuts hold the connector in place.

As a guide to the correct operation of the onboard PA regulator board measure the gate and Drain voltages of the devices:-

Device	Vgs	Vds
TR1	-1.1V	7.3V
TR2	-1.1V	8.1V
TR3	-1.2V	9.8V
TR4/5	-1.5V	9.8V

These voltages, are approximate and vary slightly between amplifiers, but give good indications of potential problems, and are measured with a 50 ohm load connected to the input and output of the amplifier to avoid instability corrupting the results.

Connect 10W power meter rated at least to 4GHz to the PA output. If a 3.9GHz source is available (such as a Midwest microwave "brick" awaiting conversion to 3456) lower the output to 1mW using attenuators as required (beware, some models produce over 250mW!) and apply to the PA input. The PA output should be at least 7.5W The next step is to apply a 1mW of 3456MHz to the amplifier. The output could be as low as 3W, so retuning is necessary. It will be observed that the board has extra printed stubs on the board, mostly on the inputs and outputs of the devices. These should be connected and disconnected with solder "blobs"

AFTER DISCONNECTING THE POWER SUPPLY BEFORE EACH ADJUSTMENT to maximize the output power working in sequence TR1-5, optimizing the gate matching first then moving onto the Drain. If necessary extra pieces of copper foil, available from hobby shops may be added to the board.

A recently discovered alternative to soldering the foil is to use adhesive backed copper foil which is available in various widths from stained glass craft shops. This was discovered to be the vital ingredient in my wife, Meg's N2NQI's newly mastered hobby and has meant many trips to supervise the correct widths being purchased. When properly tuned the amp should give at least 7W, it has been consistently found to give slightly less output at the lower frequency, despite repeated optimization efforts. Finally the inner cover should be replaced over the circuit boards, this has RF absorbing foam under it to maintain amplifier stability. The outer "can" was discarded.

The so far unused red wire drives the negative terminal of a 10mA meter via a series resistor to ground to indicate relative output.

Conclusions

Hopefully this article will encourage more activity on 3456 over longer distances. Remember the old adage of the ham bands "Use them or Lose them".

References

1. Power Amplifier for 13cm. E.Gobel, DUBUS 2/94 pp22-29

2.5GHz TV converters for amateur satellite reception

Sam Jewell, G4DDK

Figure 1.

Drake 2880 converter together with replacement crystal for the PLL and the ceramic 2.4GHz RF filter

Receive converters
Back in 1997 I published a construction article for a 2.3/2.4GHz receive converter for satellite and terrestrial use. The construction article was published in the Proceedings of the Weinheimer Tagung 42. This converter was similar to the design published in the RSGB Microwave Newsletter that same year by Dave Bowman, G0MRF, but I elected to use a 144MHz IF rather than 435MHz. I felt that a 144MHz IF was more appropriate for a mode S satellite receiver since it was my belief that 435MHz would be a more popular uplink band than 1269MHz. Consequently using 435MHz as the receiver IF would require the use of two 435MHz-capable transceivers (or transceiver and one multi-mode receiver/ scanner) in order to operate effectively. Time will tell if this assumption is right. Several months after I published my converter design I was at Microwave Update '98 in Sandusky, Ohio, when Toshi, JA1AAH and his friends appeared with several boxes of the Drake 2880 2.5GHz converters for sale at $25 each. Of course I had to have one.

When I got home I modified mine in accordance with the instructions on the JN1GKZ Website. The PLL crystal was changed to 8.8125MHz to give a local oscillator injection frequency of 2256MHz. The IF filter modifications of removing the two coils and capacitors was also done. I wasn't too impressed with the measured performance, but it was clear that with its compact weatherproof construction, it would be a very useful 2.4GHz converter when P3D finally got aloft. I put the Drake converter safely away, as the P3D launch was on hold yet again and I didn't have any immediate need for another 2.4GHz converter.

My own one worked just fine. Together with Lehane, G8KMH, I was providing the microwave measurements at the AMSAT-UK Satellite Colloquium at the University of Surrey in 1998. We were both surprised at the large number of Drake converters that appeared courtesy of Richard, G3RWL. Richard had already converted several units and Lehane and I, together with Dave, G0MRF, were able to measure their noise figure and insertion gain using a HP8970A noise figure meter and HP346A noise head. I think it would be true to say that the results were also a little disappointing. We measured converters both in unmodified form as well as those that had been modified according to the JN1GKZ instructions. The general spread of results showed that using the original published modifications, you could expect a noise figure of around 6.5dB and an insertion gain of about 18dB. These are highly averaged figures.

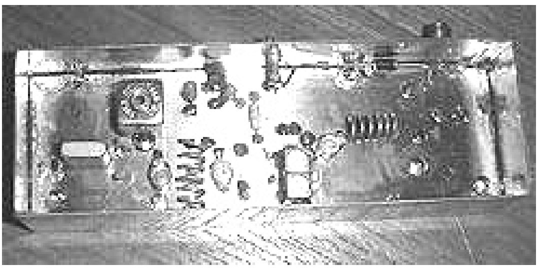
Unmodified Drake's measured at 2.5GHz RF and 200MHz IF showed that the units would comfortably meet their original specification if used within the intended band. Dave and I did many measurements of IF and RF response using sweepers, detectors and regular network analysers and concluded that the IF filter rolled off rapidly below 200MHz, and it also exhibited a marked dip around 145MHz! The RF filter also rolled off rapidly below 2.5GHz. These measurements led Dave to try some suggested modifications to the filtering which resulted in the noise figure of the test unit falling to around 3dB and the insertion gain climbing to a more comfortable 25dB. These modifications were published on G3PHO’s 'World above 1000MHz' web site:

www.g3pho.org.uk.

Essentially, these modifications involved lengthening the lines of the printed RF filter and adding one extra IF filter capacitor.

A further modification was to change the front end GaAs FET for an Agilent MGA 86576 GaAs FET MMIC. Although less straightforward than the filtering changes, the resulting improvement in noise figure is worth having. It should be noted, the length and quality of the grounding connection between the source leads of the MGA86576 and the ground plane on the reverse of the board must be short and clean (no excess solder) or the device will self oscillate - guaranteed! However, it was obvious that the noise figure could be further improved. At the 1999 Amsat-UK Colloquium Dave introduced the ceramic filter change. This involved replacing the existing, lossy, printed RF (image) filter with a small ceramic dielectric filter meant for 2.4 - 2.45GHz LAN products. When this filter was incorporated the noise figure of the converter fell to around 1.8dB, which is close to the claimed noise figure for the MGA86576 device. The converter also became more 'docile', possibly indicating less radiation loss from the printed filter, with it's always present potential
for instability.

With the modifications above the Drake 2880 is a very worthy converter for use with AO-40. However, there are still several pitfalls for the unwary. Dave Bowman, G0MRF has further



details of the Drake conversions on his web site:

www.g0mrf.freeserve.co.uk/

and may also have some of the conversion filters, MMICs and crystals for sale.

Beware the helix antenna

The MGA86576 is susceptible to input mismatch instability. One very popular UK produced multi-turn helical antenna has been measured with a very low input return loss of only a few dB. A Drake converter attached to one of these antennas will self oscillate. The use of several feet of lossy coax between the antenna and the converter will stop the oscillation, but is self defeating in so much as the overall system noise figure may then be unacceptable. The converter needs to be right at the feed point for optimum results.

The commonly accepted feed point impedance of the multi-turn helix of 140R is only obtained when the helix is centre fed1 . The more popular method of feeding on the periphery produces a much higher feed impedance requiring the use of a 1/4 turn matching section on the first 1/4 turn of the helix. This has to be very carefully tuned by bending the matching section towards or away from the reflector plate to obtain an acceptable match. A carefully matched section will start close spaced to the reflector and progressively further away at the other end.

A few further notes of caution must be mentioned when talking about helix antennas.

WA5VJB has made measurements on many of the popular G3RUH helix fed dishes in the USA. Very few of these have produced the expected results in terms of gain and circularity (and match?). Whilst G3RUH went to great trouble to get his design right for publication, it seems that many builders of these antennas don't follow the instructions. If you are using a helix fed dish and are not receiving the AO-40 S2 beacon very well when others are, it may be your dish feed! Also, the axial-mode helix is not always an axial mode! A long axial mode helix can behave like a normal mode helix (rubber duck) at some much lower frequency and may inadvertently receive high levels of lower frequency signals, especially since the axial mode helix is probably beaming into the sky and then looks and works for all the world like a vertical rubber duck antenna! A low loss, 2GHz high pass, filter between the helix and the converter may be a good idea in areas where there are lots of commercial VHF transmitters. This is especially true if you use a very wideband preamplifier in front of your 2.4GHz converter.

Converter supplies

It seems like the supply of Drake 2880 converters has, at least temporarily, dried up. I understand an alternative device has been found in the USA. However, in Eire, there is a similar 2.5GHz MMDS service to that in the USA. I have seen at least one design of MMDS converter in Eire that looks like it could also convert to 2.4GHz. I don't know if it has the same design PLL local oscillator to that in the Drake (many of these designs are similar) but maybe one of our Irish friends could investigate this possible source of converters?

Close

Receiving and analysing signals from the 2.4GHz satellites is both fascinating and easy, but it can be full of pitfalls for the unwary. If you don't get the results you expect, do check out what I have said above.

Programming a DMC 23GHz Brick Oscillator

Chris Bartram, GW4DGU

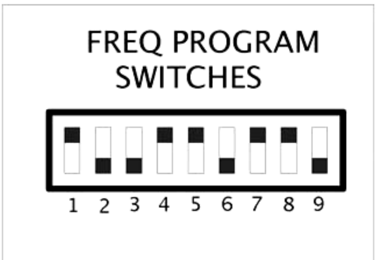
Despite my usual attitude to modifying surplus kit, I'm in the process of putting a 'quick converter' together using surplus bits in order to see if my 2.4m dish is accurate enough to use on 24GHz EME (and for some 22GHz radio astronomy).

A year or two ago I brought a DMC-110366 PLO brick, believing the seller that it might be useful as a 1.3GHz signal source. In fact the brick operates on around 23.5GHz and has good close-in noise performance and stability. After much playing with the DIP switches, I've managed to extract a very useful frequency from the unit: 23.616GHz, which is 24048-435MHz. As I've always used 70cm as an IF for 3cm, I'm rather pleased!!

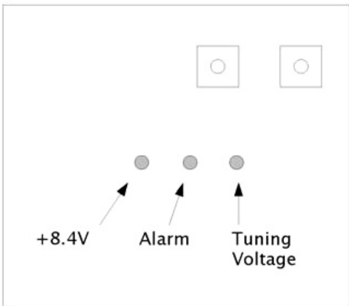
The brick requires a single +8.4VDC supply at about 1A max. There are alarm (essentially a lock detect) and tuning voltage outputs. The output level seems to be of the order of +8dBm at the 'XMTR LO' SMA output and around +4dBm from the 'receiver LO' output. I say 'seems' as I don't have a calibrated power meter and the spectrum analyser calibration is a bit suspect at 23GHz.

Alignment is a matter of setting the DIP switches and then adjusting the DRO centre frequency, using the only unmarked adjustment on the top panel of the unit.

Monitoring the tuning voltage output should show the effects of tuning the DRO. The best setting is when the tuning voltage is at the centre of its limits.



Location of DIP switches on DMC-110366 23GHz LO Module for 23.616GHz Output (435MHz IF = 24048MHz)



DMC 23GHz LO Power and Control Connections

Some DMC '23 Whitebox' information

Steve, G8JMJ

Editor's note: The following plea for information appeared on the UKuG Internet Reflector:
"Does anyone have any information on the various modules within the DMC "white boxes" for 21/23GHz ...
are they any use ?"

The previous pages describe GW4DGU's findings regarding the oscillator contained in the equipment but here's some more information, this time from Steve, G8JMJ. who posted this on the reflector in reply to the query above.

The DMC-110366 oscillator brick uses some form of a dual loop, but I've not had the time to analyse it properly yet.
A few points:
• The unit has two crystals, a 14.000MHz (TO5 can style) PCB mounted and what appears to be another one buried on the PCB below, contained in a small oven (will confirm and measure it's frequency)

- The sapphire trimmer on the top of the unit trims the 14.000MHz oscillator and can be used for fine frequency adjustment. The other oscillator can be adjusted by removing the large screw-in plug on the side of the module.

The bit switch positions have the following step sizes.

1 =3D 1.75MHz
2 =3D 3.50MHz
3 =3D 7.00MHz
4 =3D 14.00MHz
5 =3D 17.5MHz
6 =3D 35.00MHz
7 =3D 70.00MHz
8 =3D 140.00MHz
9 =3D 280.00MHz

The big toggle switch buried inside the module (accessible through a hole on the top) switches the DRO into free-run mode to allow for frequency setting with the loop open.
It should be noted that, although the 'synthesiser' circuitry can step these relatively large steps, the DRO cannot and so will require manual retuning before 'locking' occurs. Not all codes are valid. I have had the unit locked where the LSB (1.75MHz) step switch is in-operative.

On the other bits, the LNA runs from +8.4VDC as does most of the entire box. The gain at 24GHz was only about 17db. I've not measured NF yet.

The TX up converter works reasonably well at 24GHz with an IF of 200 to over 1300MHz (obviously lower conversion gain at the top end).

The TX image filter will retune so that the top end of it's response is at about 24.1GHz, insertion loss about 1.5db but with about 1db of ripple. Bandwidth, from memory, is about 300MHz.

Useful 10GHz Filter with 10MHz Bandwidth

David Wrigley, G6GXK

In the past have had some problems with my 10GHz transmitter. In order to ensure that I was peaking up the right signal I thought that it might be useful to feed the output through a narrow filter before measuring the output power. There are, of course, many such filter designs (ref. 1) in the Microwave Handbook and there is also the G3KNZ program (ref. 2) which permits you to design your own filter, as the following example shows:

Waveguide Band-pass Filter Design using WG16
Filter type B,
Ripple= 0 dB, Frequency= 10368.1 MHz, BW= 10 MHz
WG16 Internal Dimensions = 22.86 x 10.16 mm
Material=CU, Skin depth= 2.237868E-02 mm, Iris Thickness= 1 mm

Waveguide Filter Dimensions:
Cavity No: 1 = 18.2 mm (less 5% = 17.3mm actual to allow for tuning screws)
Iris Hole size: 0 = 6.4 mm
Iris Hole size: 1 = 6.4 mm
Average Unloaded Q of Cavities = 5500
Insertion Loss = 1.64 dB
Lg/8 Matching screw spacing= 4.7 mm

I was searching for some scrap copper waveguide in order to make just such a waveguide filter, when my eyes fell on a discarded **14GHz** filter which was originally part of a Marconi 14GHz up-converter, many of which were sold off at meetings last year. The filter had been discarded because it was originally thought that it couldn't be re-tuned to any useful frequency. You know the feeling - it looks so nicely made that one feels instinctively that it will come in useful one day!

A further examination revealed that it was indeed a very useful piece of microwave kit. The filter is a six stage unit built up of separate silver plated cavities interleaved with copper iris plates and the whole thing bolted together by means of four longitudinal bolts. The cavities are basically very rugged sections of WG18 and the cut - off frequency for that is about 9.5GHz. So, it should be OK at 10GHz. The thing to do is to remove some of the iris plates and calculate what the new cavity length is for 10368MHz, but first we can, out of interest or curiosity, check that the original design is what we thought it should be and get some experience using the Steve Davies, G3KNZ filter program running under GWBasic.

Fig 1: This is the self adhesive label from the original filter unit

Using components of old 14GHz unit Original design:
WG18 section 15.8 by 7.9mm
Cavity length 12.35mm
Iris Dia 5.7 on outer units, all the rest 5.1mm
Iris thickness 0.3mm
Design frequency marked on front panel, 14275MHz
Several runs were made, adjusting the bandwidth until the iris diameters came out right.

Running the program:
Filter type B,
Ripple= 0 dB, Frequency= 14275 MHz, BW= 50 MHz
WG18 Internal Dimensions = 15.8 x 7.900001 mm
Material=AG,Skin depth= 2.554082E-02 mm, Iris Thickness= .3 mm

Waveguide Filter Dimensions:
Cavity No: 1 = 13.4 mm (Actual cavity = 12.35 , therefore using a tuning allowance of 8% on calculated)
Iris Hole size: 0 = 5 mm (Actual holes mainly 5.1 mm in 6 cavity filter)
Iris Hole size: 1 = 5 mm
Average Unloaded Q of Cavities = 5033
Insertion Loss = .49 dB
Lg/8 Matching screw spacing= 3.5 mm

Revised cavities for 10368.1MHz Design Filter type B,
Ripple= 0 dB, Frequency= 10368.1 MHz, BW= 10 MHz
WG18 Internal Dimensions = 15.8 x 7.900001 mm (cut-off freq 9.5GHz)
Material=AG, Skin depth= 2.176689E-02 mm, Iris Thickness= .3 mm
Waveguide Filter Dimensions:
Cavity No: 1 = 34.3 mm (less 8% allowance for tuning = 31.5mm)
Iris Hole size: 0 = 6.6 mm (Used 5.7mm)
Iris Hole size: 1 = 6.6 mm
Average Unloaded Q of Cavities = 4646
Insertion Loss = 1.94 dB
Lg/8 Matching screw spacing= 9 mm

Conclusion

For a 10368MHz filter, the desired cavity is 31.5mm long. Divide this by the current length of a cavity block (12.35) and we get 2.55 blocks. So just over two and a half blocks are required for the new cavity. One was cut in half fairly easily using a junior hacksaw with a sharp new blade and the cavity block held in a vice. The sawing marks were removed from each piece using a broad smooth file. The cavity blocks were then rearranged to form the 10368GHz filter. Tuning up a single cavity is fairly straightforward provided one has a signal source and a detector. I used a Gunn Diode oscillator and diode detector. The Gunn Diode was first set up to 10368MHz using a frequency counter, then the filter was peaked up using a diode detector. I later tried to check the pass band of the filter unit by means of manually sweeping the Gunn Diode oscillator with a frequency counter on the output, but this was difficult due to the very sharp tuning which caused the frequency counter to lose lock either side. What was clear was that the filter had a bandwidth of a few MHz either side of 10368MHz. I found that both tuning screws in the cavity had an effect and had to be adjusted for maximum output.

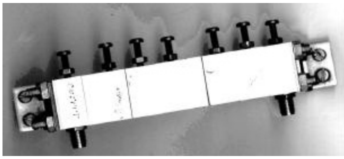


Fig 2: The finished filter

you can figure out the posit ions of the half blocks by means of the gaps in the tuning screws
Clearly many filters from 10GHz to 18GHz could be constructed from these bits – a sort of universal filter kit and it will no doubt be useful in the future .... not bad for an initially rejected item!

References

Ref 1 Examples of practical Waveguide filters 1995 reprint Microwave Handbook Vol. 3 page 18.18. (G3JVL)
Ref 2 Waveguide filter design program 1993 reprint Microwave Handbook Vol 2 page 12.26 (G4KNZ)

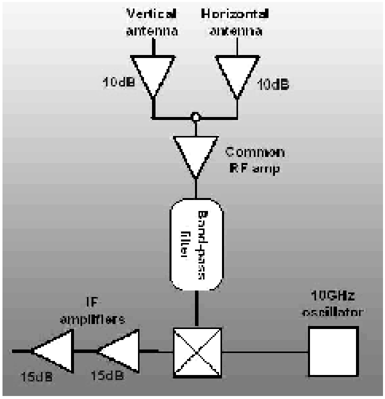
A (re) introduction to 10GHz via modified Satellite LNBs

Bernie Wright, G4HJW

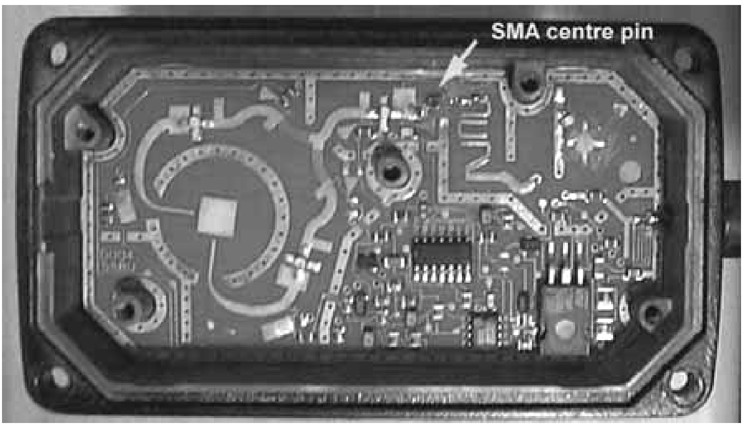
Dropping in late to the Lincoln Hamfest a few years ago, I found one of the traders selling off a box of about 30 satellite LNBs for £10.00, and just couldn't say no. Most LNBs are as shown here to the right - the older ones may have more stages and an active, rather than diode, mixer but the block diagram shown should still be quite representative of what is currently around.

The MMIC IF stages were an obvious candidate for re-use, and it was straightforward enough to convert the stages to provide amplifiers with a gain of 40dB or so from HF to a couple of GHz. With these, built onto the back of a dipole/reflector plate combination, noise free signals from the local 23cm TV repeater could be piped down through long lengths of pretty "ropey" coax without having to worry about signal loss.

It was a year or so later that another use came to mind. Interest in SETI had rekindled a general interest in radio astronomy for several of us and the notion of producing an 11GHz interferometer, using two 60cm dishes with their LNBs, began to appeal. We knew that the RF stages of most LNBs



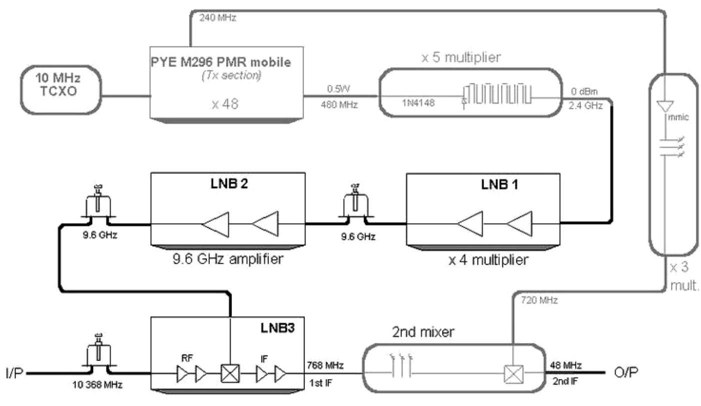
of the time gave about 10dB gain, and that there were always at least 2 stages, so early on we thought about getting in after the second stage on each dish mounted LNB, bringing the feeds out to a combiner (OK, just a



BNC T piece!) and feeding the result in to a third LNB, which would only be modified by fitting an sma socket in place of integral antenna probe. The loss in even "bog standard" quarter inch coax over a few metres would not approach the 20dB gain figure of the RF stages. It was soon clear that, with many of the LNB types, you could simply drill through both the pcb and the aluminium housing it was fitted to, after the second RF amplifier, You would then be able to open up the casting hole and sweat an sma socket onto the ground side of the pcb. It turned out to be very easy to do. On LNB types where part of the casting did get in the way, a hacksaw and file soon resolved the problem.

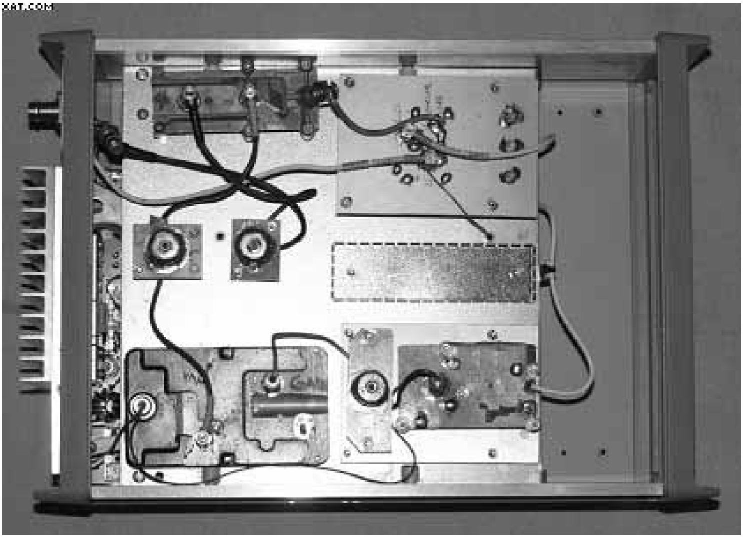
The 1-2 GHz IF from the third LNB was fed via a couple of further MMIC stages into a diode detector. After much adjustment (and fun), good interference patterns were obtained from sun transits with this arrangement. During the following year, the idea of replacing the ceramic resonator based LO with something more stable grew. It had been 20 years since my last experience with 10GHz (and those free running 723A/B klystrons) but I at least knew that narrowband operation had now made klystrons and Gunn diodes (and Barrett diodes, if anyone remembers those) all but obsolete.

About half of the surplus LNBs had a discrete diode pair mixer, with MMIC IF gain stages and a single GaAs FET oscillator. Again, it was easy enough with most of these to drill a hole through both the pcb and casting at the point where the FET oscillator fed the diode mixer and fit another sma connector so that an external crystal multiplier LO could be applied. It took a while to figure out a suitable multiplier, but, in the end, a Philips FR5000 PMR 1W VHF driver module was used to drive a pair of 1N4148 diodes. A 12 stage(!) pcb interdigital filter followed it to give a very clean x6 multiplication up to just over a GHz. Some 'Heath Robinson' MMIC amplifiers and pipe cap filters then took the LO up to 10GHz. I can't remember now what the IF was but it was single conversion to something around 150MHz. With this converter, fed from a second LNB just used as an antenna-with-integral-preamp (one of the interferometer units), a trip down the M11/A130 soon had GB3CMS coming in at good strength on the road side at Ford End. The real downfall with this converter was the poor stability of the crystal oscillator – but it was a start



and the bug had bitten. The frequency stability problem eased when it was realised that 10MHz, 12.8MHz and 14.4MHz TCXO's – all readily available, would multiply up to provide useful LO frequencies. Since 14.4 MHz also multiplies up to exactly 10 368MHz, this later became the basis for a CW/FM transmitter [likewise a 12.8MHz TCXO was/is being used as the basis for a low power beacon in the Cambridge area].

The current 10GHz converter here uses a 0.5ppm 10MHz TCXO in a double conversion configuration.



The first IF is 768MHz [9.6GHz LO] and this is brought down to 48MHz using a second LO of 720MHz, all driven from the 10MHz oscillator as below: A more practical approach to the LO multiplier at 480MHz, for anyone who has access to an ex Ionica head, would be to use the duplexer to select the 3.2GHz harmonic . These items have a surprisingly wide tuning range, and both RX and TX sides will cover 3.2GHz (9.6GHz LO) or 3.456GHz (10.368GHz). They can also be operated end-to-end from the TX to the RX port, giving more/sharper selectivity.

Liberal use of surplus satellite LNBs was [deliberately] made in this converter, though no attempt was made to use the integral image filter as part of the 9.6GHz LO multiplier. However, this did work well on the transmitter strip to multiply 3456MHz up to 10.368MHz, giving 40dB plus rejection of all other harmonics (without any modification to the etched filter). Instead, a couple of 15mm pipe cap filters were used.

These, together with the single image pipe cap filter, can be seen in this photograph of the converter: Yet to be tried is modifying the LNB oscillator so that the stage acts as both buffer and selectivity block using the integral ceramic puck. Has anyone already tried this approach?

The single image filter seems to have been quite adequate with such a high 1st IF, and the through loss not an issue when fed via a remote LNB antenna-with-integral-pre-amp arrangement.

Incidentally, RG223 seems to have a

loss of 4 to 5 dB/m at 10GHz, so up to 2 or 3 metres of interconnection lead between antenna and converter is OK. For longer lengths, such as chimney to shack use, it just needs another LNB configured as a 20dB amplifier added as a line amp.

Two LNBs are used in the LO path (laziness rather than necessity) configured as 20dB amplifiers. The first takes the 2.4GHz feed at about 0dBm, ensuring that this amplifier clips. The fourth harmonic is selected by the pipe cap filter and fed to the second LNB for amplification. About 10dBm is then available, so another pipe cap filter was put in for good measure prior to feeding the third LNB, taken from a mini-dish and fully used as the 10 368 to 768MHz converter stage. The early Cambridge mini-dish LNBs seem to work very well (see photograph on the next page) and the converter is fed from another one of these.

Laziness dictated the use of a separate mini-dish for transmit. With 300mW of CW, there has been no need to power down the receive LNB when on transmit. Since there is already a very noticeable increase in background noise as the dish is dropped down to the horizon, no attempt has been made to optimise the LNB probe feed at 10 368MHz.

GB3CCX at 150km is pretty much always copyable with this set up and thermal noise from nearby trees and buildings can be noticed as the antenna is rotated. So all in all, the LNBs have proved to be a good buy.

Every type of unit that was modified to act as antenna/ pre-amp units had a noise figure that were good enough to detect ground noise easily. Some of the mid 90's (analogue) Cambridge units had three cascaded stages following the antenna, but most have only two. A good rule of thumb is to expect 10dB of gain per stage. Using a hot air gun to remove devices has proved reliable.

Careless handling after that point has destroyed a few devices through static damage, but that is all The only comment I would make is that current limiting on the GaAs FET stages never seems to be provided, so that shorting the gate to ground always causes device destruction.

As for other LNB uses – I wonder if anyone has attempted using dual output units, which have the RF section duplicated, as the basis for a phasing transmitter/ transceiver?

HS400 - A useful and simple modification

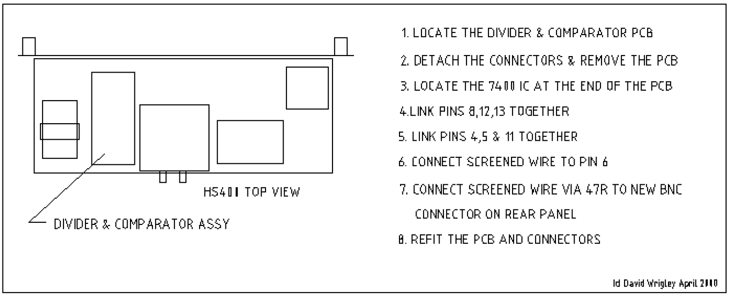
David Wrigley G6GXK

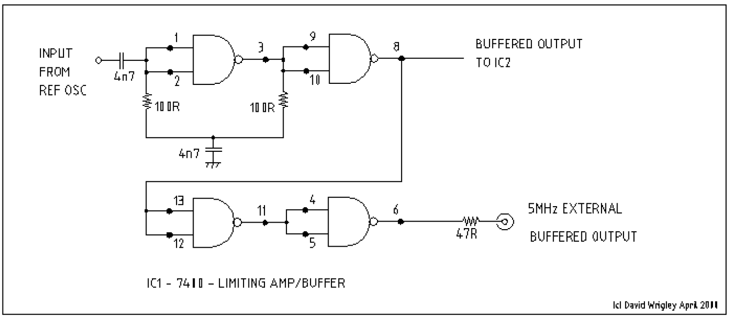
The HS400 contains a very useful and high quality reference oscillator. We all need a frequency source that we can rely on to set up the local oscillator in our transverters and any frequency counters that we may have. This 5MHz reference oscillator is just the job to use as a reliable local source. It was the author's intention to control this oscillator by synchronising it with MSF – hopefully more of that later. The first stage is to get at the voltage control connections and the output. These are all available on the connector at the end of the Divider and Comparator PCB which is located next to the reference oscillator.

A simple connection would be possible but unwise. These oscillators can be affected by load even when buffered by a single stage and it is vitally important to keep it low and constant.

It was decided that the connection would be made further away from the oscillator and the output of the Limiting Amplifier would seem appropriate. With amazing good fortune there are two spare gates in the 7400 (IC1), which have been used in this modification to further isolate the output from the internal circuitry. The only further consideration is the possibility of shorting the output. Well the man who doesn't make mistakes is a myth – we all do. So I have added a 47R resistor in series with the IC – this also cuts down the output voltage swing to a more reasonable level and limits the potential for damage to other equipment. Well that's it. The diagram gives a bit more detail.

The next stage is to build a PCB to





control this oscillator from MSF via a PIC chip controller feeding a Digital to Analogue Converter IC.
The purpose will be twofold.
1. The provide a long term lock to MSF accuracy
2. To minimise drift in the short term between MSF updates.
It should be possible to get an accuracy of 1 part in 10^10 using this system which is 1Hz in 10GHz. This might be bit over the top in terms of

HS400 – Electrical Frequency Control of the oscillator

David Wrigley G6GXK

The HS400 contains a very useful and high quality reference oscillator. We all need a frequency source that we can rely on to set up the local oscillator in our transverters and any frequency counters that we may have. This 5MHz reference oscillator is just the job to use as a reliable local source. It was the author's intention to control this oscillator by locking it to MSF or another reliable source and these modifications are intended to allow us to do that. The previous modification showed how easy it was to get out a buffered 5MHz output to a socket on the back panel. This modification allows us to electrically fine tune the crystal frequency and will ultimately allow us to set it to 5MHz manually (and later automatically) to within a few parts per 10^10.

Anyone who has seen the schematic diagram for the HS400 will have noticed that a provision for electrical frequency control is in place. How strange then that it doesn't work! The reason for this is that whist the external wiring is in place; Toyocom does not fit the internal components into the reference oscillator.

To fix this problem we will need to get inside the reference oscillator can and make connections to the internal stabilized supply and then fit a Varicap diode and series resistor to the crystal circuit.

One Varicap? – well, I believe that using only one diode can theoretically produce some phase noise and that two diodes back-to-back minimise this effect. However, this 5MHz reference oscillator using only one diode has been used as a source for the Manchester 10GHz beacon and no significant phase noise has been heard in the receiver. I am reluctant to put in extra components unless I've experienced the need for it.

Schematic correction

Before we get deeply involved, it is a good idea to check the cable between the reference oscillator and the adjacent PCB. In the two units modified so far this cable did not agree with the circuit diagram. The circuit diagram should be:-

The Orange wire from SKT C/pin2 connected to SktA/pin 4 (not 2 as shown)

The Blue wire from SKT C/pin7 connected to SktA/pin 2 (not 4 as shown)

Getting to the oscillator PCB

You will need to remove the oscillator can from the HS400. The procedure is

1. Turn the top panel screws a quarter turn (opposite ways!) to release the top panel.
2. Locate the Reference oscillator can at the opposite end to the mains transformer.
3. Unplug the connector and slacken off the two clamp fixing screws to release the can.
4. The can will then slide out towards the front of the HS400 unit. Move the can over to a clean area of bench and have a small tray handy to hold the collection of screws to be removed.

To strip the unit,

1. Unfasten the four small cross-point screws (with washers) on two sides of the can.
2. The base and assembly will then be able to be pulled out by means of the projecting screwed rods.

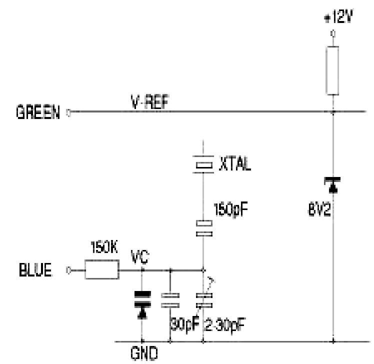
3. Unwrap the insulating fibre material and put this to one side.
4. The PCB in which we are interested is the small top one and can be worked on without any further dismantling at this stage.

The modification,

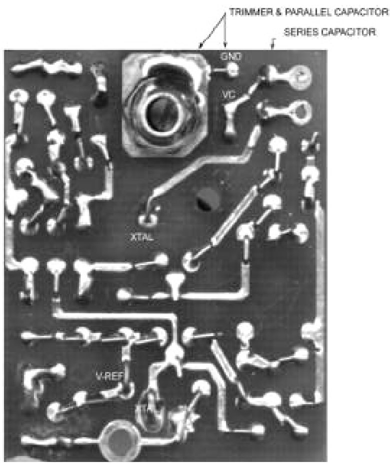
1. Fixing the Varicap diode and associated capacitor and resistor.

The Varicap diode used by the author was a small plastic (SOD323) 17pF (at 0V) type. The types likely to be suitable are the Farnell BB149A (316-2837) or alternatively (in SOT23) the Farnell, BBY31 (300-0310) or BBY39 (316-3076).

The Varicap diode is soldered across the points marked “GND” and “VC”. Also from “VC” is connected the (short) end of a wire ended 150K resistor. The other end of this resistor is connected to a wire and insulated with sleeving. The other end of this wire is soldered to the blue wire termination pad on the lower PCB. This wire provides an external connection for the frequency control voltage.



TOYOCOM 613 REF OSC - PARTIAL SCHEMATIC DIAGRAM



Useful 1GHz Oscillator Module

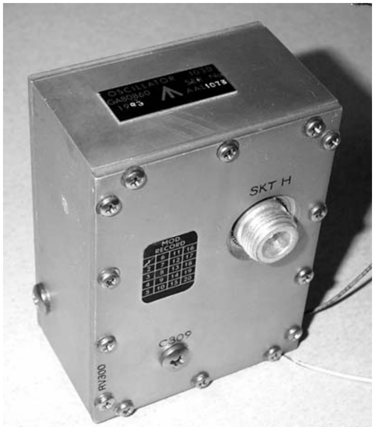
Gordon Fiander, G0EWN

Some of you may have noticed some extremely well constructed units for sale at various recent UK rallies. They are normally being touted by dealers as 'ovened oscillator units' with output at 1030MHz and are selling for around £2.50. I bought a couple at an Autumn rally at Wakefield.

On checking the units at home, the following became apparent. First the units are extremely well built; the milled brass enclosure alone being well worth the cost! (See photo). However, the units are not 'ovened'. Inside, a crystal oscillator (on 85.83333MHz) is followed by an amplifier stage, which in turn drives a diode multiplier feeding a tuned output line. The output line is tuned to pick off the 12th harmonic, in this case giving about 18dBm out on 1030MHz.

I used the bench supply set at +13.8V to test the units--both drew about 40mA. Having ascertained the units worked OK I fitted a 5th overtone crystal on 86.4MHz to one unit. The unit worked fine without need to retune, giving an output on 1036.8MHz; the 10th harmonic of this falling in the 10GHz band. The note was quite stable after a short period from turn on , even when listening on 10GHz, and the crystal could be easily adjusted to give a signal at the band edge or around the calling frequency on 10GHz. I intend to fit a DB6NT heater for even better stability and hope to be able to use as a marker to check RX (frequency?) when out portable.

The units need a stable supply as small voltage changes affect output frequency. I hope to change the second unit I have by fitting a crystal on



90.666 to give output at 1088MHz for a LO for another project; it might also work with 96MHz to provide an 1152MHz source and as such are a good buy.

Note that you will need a hot iron to desolder the existing crystal as silver solder appears to have been used throughout. Also if the supply drops below +11V the oscillator stops (it may have originally had +20v supply?). I have made a simple diode marker generator to go with the unit and can hear the signal throughout the shack. When the dish is pointed at the unit the signal is 59++

Surplus

Test Equipment

An External Mixer for the HP8555A Spectrum Analyser Plug in

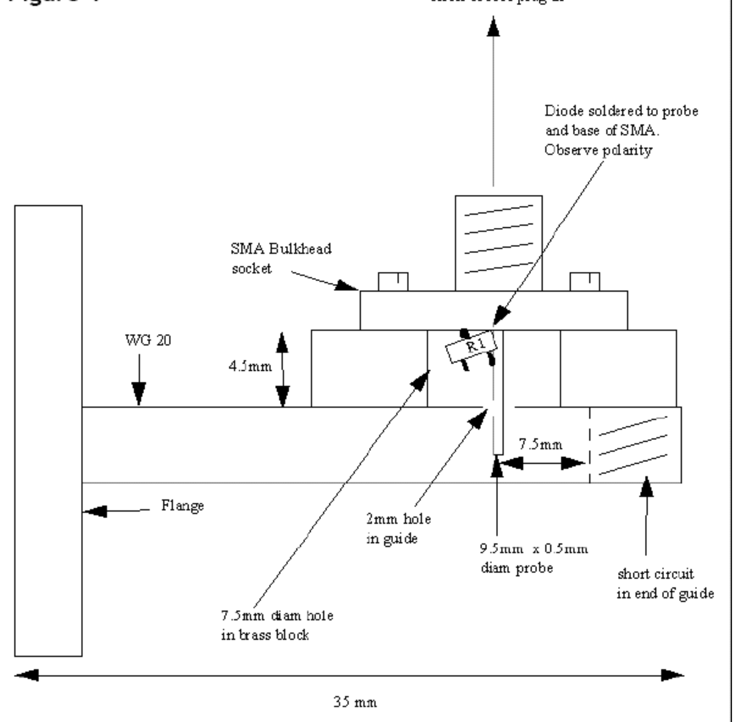
Chris Towns, G8BKE

This article describes an extension of this useful instrument to 24GHz and above, by the addition of a simple add on mixer.

In its standard form, the 8555A plug-in for the HP141T Spectrum Analyser will accept inputs in the range 10MHz to 18GHz via the input N type connector. There is a modification available to allow access to at least 24GHz via this connector but 24GHz via an N type is not really recommended. Indeed, I believe the modification precludes one from using the external mixer capability.

The 8555A will allow the use of an external HP mixer via a front panel BNC socket and this will extend the range of the instrument to 43GHz, just short of the amateur band! To obtain displays at this sort of frequency the 8555A LO is automatically multiplied in the external mixer and mixed with the test signal. In
conversation with Brian, GM8BJF it turned out that he was using a "home-

Figure 1



brew" external mixer to look at a 24GHz signal quite successfully and its was decided to try this approach to look at 47GHz. Although the amplitude calibration using this method is not really absolute, adjustments for maximum output, sideband selection and signal cleanliness can nevertheless be examined.

With no genuine surplus HP external mixers available, a small unit would have to be made as GM8BJF had done. However, without access to a 1N26E diode that Brian had used, experimentation with more readily available diodes was needed.

Since the mixer was going to be used at 24GHz and above, it was decided to make the unit in a very short length of WG20 which was to hand.

Figure 1 shows the outcome of using this approach.

The SMA socket was mounted on a small 4.5mm thick brass block soldered to the guide broad face, with a ~7.5mm diameter. hole in it to accommodate the mounting of the diode. A short circuiting brass block was soldered in to the end of the guide at the position shown. The probe from the SMA protrudes into the guide via a 2mm hole in the guide wall. The probe itself can be a wire extension of the SMA pin, or better still the SMA pin itself.

A number of diodes were tried including those out of LNBs. All worked well, but, in order to obtain a readily available source, the commonly available HP HSMS 8101 diodes which are available from Farnell (Stock No. 994-649) were used. Only really applicable for use in a surface mount applications their packaging is not ideal, but since maximum efficiency is not required, they seem quite satisfactory here. Care should be exercised in soldering in the diode as static can damage the device. Correct orientation of the diode should also be observed, since a positive bias voltage is applied to it from the 8555A. There are three leads on the diode package, but only two are used so it should be mounted with the lettering as shown, ensuring that it is correctly orientated. It is quite possible that other diodes such as the DDC4561 or even a 1N21, could also be used, but these have not been tried. The external mixer arrangement on the HP 8555A is clever in that the DC bias, IF and LO all share the same cable to the mixer. Thus if a short length of good SHF cable is used, suitable for use at 2-4GHz (e.g. SUCCOFLEX) no difficulty should be experienced in obtaining results. Adjustment of the "Ext. mixer bias" pot on the 8555A optimises the mixer /multiplication process for best signal into the analyser.

Since the frequency dial on the 8555A runs out at 43GHz, some other means has to be employed to determine the frequency of the wanted signal. Luckily HP also put the LO frequency on the top of the scales. Thus one knows what LO frequency the mixer is seeing! The IF centre frequency on the highest ranges of the 8555A is 2.05GHz. Knowing this, one can, by some arithmetic, work out what the LO frequency should be. As an example, to display a 47.088GHz signal, 47.088 minus the 2.05GHz IF will require an LO of 45.038GHz. This is obtained by a x12 multiplication of 3.75GHz LO in the mixer. Thus the frequency set pointer should be set as close as possible to the LO frequency of 3.75GHz on the top scale of the instrument.

Although I have not tried this, due to lack of a good 3GHz counter, it should be possible to connect a counter to the ext. mixer socket (watch the DC bias!) or the first LO output socket and with the 8555A set to "manual sweep" set the LO to precisely this frequency and then reset to "Int. sweep mode" again with the span set to say 1MHz. However, having said this, it appears that the "Signal Identifier" on the 8555A still operates in this non-standard mode thus one can check by the normal means if the signal being displayed is the correct one bythe "usual two divisions to the left" offset.

Note that for the external mixer to function and for the LO to be directed out of the instrument correctly the 8555A must be operated with the band selector set to one of the frequency ranges above 18GHz.

Also note that in this mode the "Input Attenuator" of the 8555A is not functional, although the "Log Ref. Level" control is. If the power of the signal has been measured previously, on a power meter, then a crude calibration of the vertical scale can be made. The signal being checked into the mixer should be kept to 2mW or less as driving it harder only produces more mixer products and tends to confuse the measurements. Keeping the input to this sort of level also ensures the longevity of the diode!
It is hoped than when I'm operational on 76GHz that the same method can be used to examine the signal at that frequency.

A similar diode mounting arrangement is being examined with a view to producing a useful waveguide noise source for noise figure measurements above 24GHz.

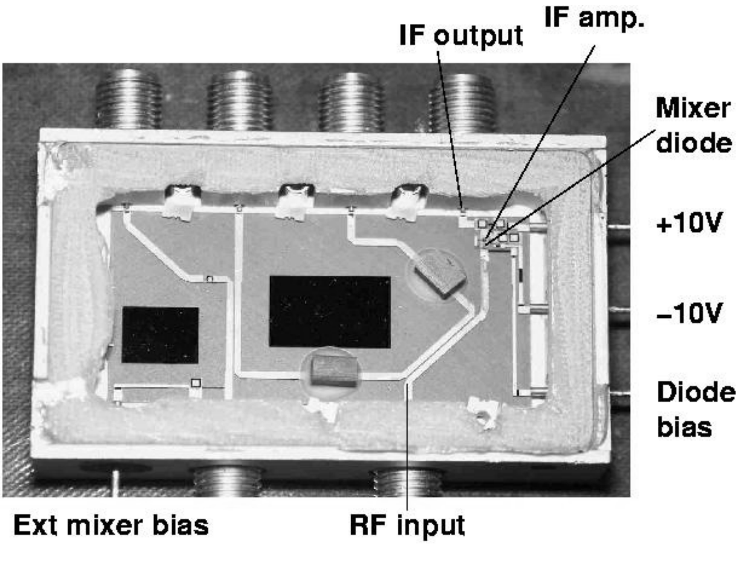
Repairing the Input Mixer of the HP8555A Spectrum Analyser Plug-in

Brian W, Flynn, GM8BJF

Recently I noticed that the sensitivity of my much prized HP 141T/8555A spectrum analyser had dropped rather drastically. I feared that the input diode had died for some reason and a d.c. check of the mixer diode, as described in the service manual [1], confirmed my worst fears. The diode was short circuit.

After a period of dejection I considered methods of repair. The first mixer is a thin film microcircuit which HP replace as a unit of it is damaged. There is also a small plug-in PCB assembly which contains select-on-test resistors, used to set the mixer bias and the 50MHz IF amplifier gain, to define the calibration of the unit. This PCB is changed along with the mixer. Deciding that there was little to lose, I removed the lid from the mixer package. The mixer substrate is housed in a gold plated metal case with a lid which is held in place with conducting epoxy adhesive. The lid can be removed by carefully prizing it off using a small screwdriver as a chisel while holding the mixer unit in a vice. This revealed the circuitry on the substrate shown in **Figure 1** below. The mixer diode is a chip device which is bonded to the substrate and the post mixer amplifier is a grounded base bipolar silicon transistor. The mixer diode output drives directly into the emitter of this transistor, presumably to give a good broad-band match to the mixer diode output impedance.

Initial attempts to replace only the mixer diode with a surface mount



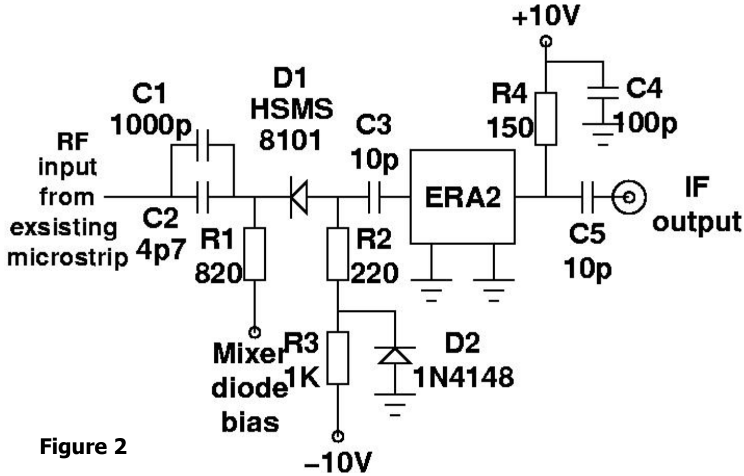


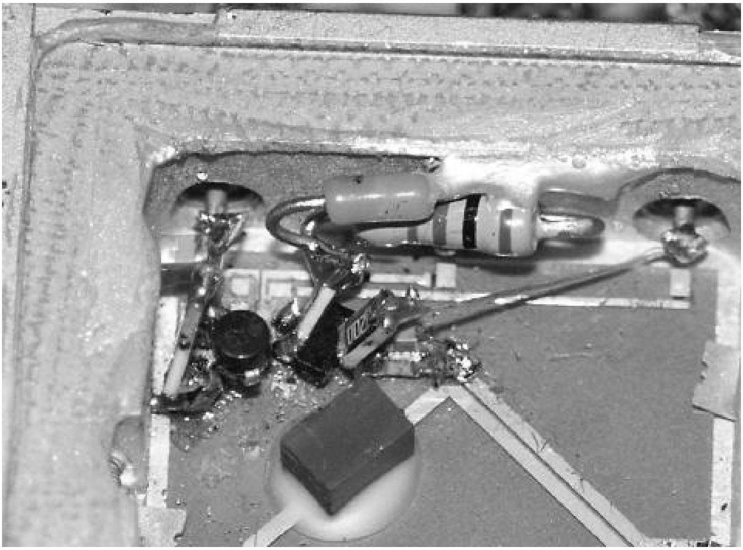
Figure 2

packaged device were unsuccessful. For this to succeed wire bonding facilities would be necessary. The next attempt was to replace all of the active circuitry as per the simplified circuit diagram shown in Figure 8-12 of the service manual. This was not very successful as the sensitivity dropped drastically.

The alternative approach adopted was to replace all of the active circuitry with the circuit shown in **Figure 2**, fabricated with surface mount components. The first step was to ground the two common leads to the ERA2 to the earth point on the substrate. This was achieved by careful soldering with a fine bit. Once the ERA2 was in place, the bond wires to the package connections were broken so that connection could be made to them. The ERA2 amplifier operates as a IF 2GHz pre-amplifier and provides a 50 ohm termination for the mixer diode. The HSMS-8101 is a Schottky barrier mixer diode which is available from Farnell. The capacitors C1 and C2 could be omitted if use can be made of the existing component mounted on the input microstrip. Attempts to solder to it were unsuccessful and resulted in its destruction, hence the use of C1 and C2. These were paralleled to give a broad band capacitor as the instrument operates from 10MHz to 18GHz. It would be preferable to retain the original part in circuit to maintain the response of the instrument at the lower frequency end of its range. The resulting modified mixer is shown in Figure 3. The result is not pretty, but it meant that the unit became usable again, (even if HP/Agilent would not approve of the accuracy!). The ground connections to C4 and D2 were made directly to the metal case with conducting epoxy. They were first soldered in place, in contact with the case, using the solder connections to locate them. The ground connections were then formed with sparing "blobs" of conducting epoxy.

Mixer Bias

The mixer is operated as a harmonic mixer to get the wide tuning range. It operates up to the fourth harmonic of the 2 – 4GHz YIG LO. To optimise the conversion loss the dc bias applied to the diode is set for each harmonic. This is controlled by fixed resistors



mounted on the daughter board assembly A10, which is mounted on plug-in board A6. Experimentation with the resistor values can improve the sensitivity. This was achieved by temporarily replacing the appropriate resistor with a 4k7 variable and varying it for best sensitivity on a suitable signal and then re-fitting a suitable fixed resistor.

Performance
The repaired mixer gave reasonable performance over the full range of the instrument. The amplitude calibration is not as good as it once was, but is within 2-3 dB on all ranges. To complete the calibration the resistors controlling the IF amp gain on the A16 assembly would repay attention. When time permits these will receive attention. After testing, the lid was replaced using more conducting epoxy.

Conclusions
This procedure was able to bring an otherwise useless piece of equipment back to life and allow it to be useful for amateur work. Manuals for the 141T mainframe and most of the plug-ins can be downloaded from the website:
http://www.logsa.army.mil/etms/find_etm.cfm
and can also be found at:
http://recordist.com/pdf
Unfortunately, the full manual for the 8555A is not publicly available on these sites but a nonetheless useful manual giving drawings and component locations and values can be had. The full operation and service manuals for the other 141T plug-ins, preselectors and tracking generators are available. There are also manuals for other items of test equipment available if you search the site.

Be warned they are large files!

References
1. Spectrum Analyser RF Section 8555A, Operating and Service Manual. Hewlett-Packard Company, Santa Rosa, USA, 1971.

Acknowledgments
The author wishes to thank Peter, GM4DTH for taking the photographs.

Parts listing

R1 820 0805
R2 1k 1/8W axial
R3 220 0805
R4 150 1W surface mount Farnell 507-878
C1 1000p chip
C2 4p7 ATC100A
C3 10p chip
C4 100p chip
C5 10p chip
D1 HSMS8101 Farnell 994-649
D2 1N4148
IC1 ERA2 Mini-Circuits

A New Pointer Drive for the HP8555A Plug-in

Chris Towns, G8BKE

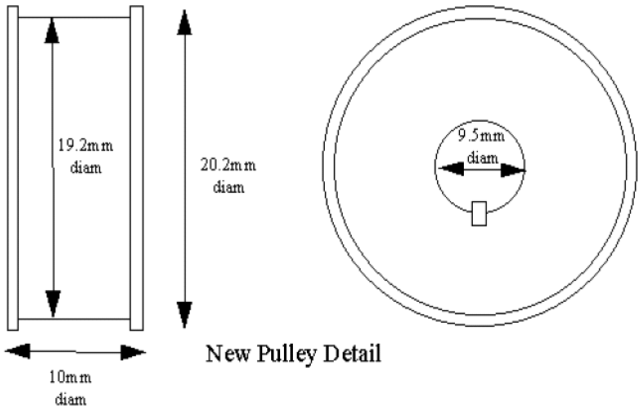
There are a large number of Hewlett Packard HP8555A 0.01 to 18GHz plug-ins for the HP 141T spectrum analysers around in both amateur circles and on the surplus market. One drawback of these units seems to be that the pointer drive mechanism, comprising a thin fibre belt, can wear with age and eventually snap. This renders the instrument pretty useless and to replace the belt is quite costly.

The following pages offer a cheaper solution which seems to be just as satisfactory as the original drive.
It involves replacing the cogged belt drive with a chord and pulley drive. Access to a small lathe is required to turn the new pulley in brass or aluminium.

To get at the drive mechanism one has to proceed with the following steps after the 8555A has been withdrawn from the 141T mainframe:

1. Remove the top cover (6 screws).
2. Turn the frequency controls to lowest frequency (fully anticlockwise) and ensure that the control is in the "normal" tune position (i.e. pushed fully in).
3. Remove the frequency control knobs via the small hex screws.
4. Next, with plug-in front panel facing you, unscrew the two countersunk crosshead screws on the left hand side of the unit and nearest the front panel.
5. Invert the unit and unscrew the two countersunk cross-head screws on the bottom of the unit nearest the front panel.
6. Looking again on the top of the unit, locate and carefully prise off the D type connector just behind the display. (This has a large loom of wire going to it). The whole front display including front panel should now lift upwards and away form the rest of the unit.
7. Remove the front panel escutcheon by using an Allen key to undo the "Ext mixer bias" and "Amplitude cal" pots. and also remove the circlip around the frequency control spindle.
8. This exposes three crosshead screws underneath the "BAND" escutcheon which should be removed.
9. At the rear of the front panel in amongst the front panel switches are two large nuts. One only has to loosen them off one turn, since the drum assembly is held in position with slotted brackets (clever HP!). The whole tuning drum assembly will then lift clear of the front panel.
10.Locate the bracket holding the three blue multi-turn pots and very carefully remove it by taking out the two cross-head screws holding the bracket to the display. Do not disturb the rotational position of the pots. The long tuning spindle will also come with this assembly together with the "normal / rapid" tuning clutch assembly. This contains 4 sprung ball bearings so be very careful with this!
11.This will leave on the display section the brass bracket carrying the display drive sprocket and a gear.
12.Remove this brass bracket by taking out the two long cross-head screws and then remove the circlip which holds the sprocket on the hollow shaft. You should now have complete access to the sprocket.

A new pulley can now be turned up in brass or aluminium, with a diameter of 19.2mm, a lip diameter of 20.3 mm and thickness of 10mm. as per the diagram. A hole with diameter



New Pulley Detail

of 9.5mm with a small key in it is also required to key it to the keyway in the existing shaft.

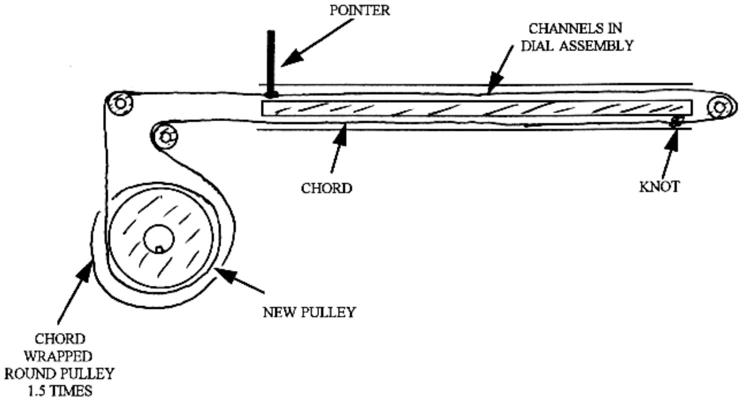
This can be made by filing a small slot in the pulley and soldering a small piece of brass in the slot.

The old sprocket can be used as a pattern. The new pulley must be a good push fit on the shaft otherwise backlash and wobble will occur.
For those unable to get access to a lathe, the original sprocket could be modified by carefully filing off the sprocket teeth and sticking two, slightly oversized, "cheeks" in thin brass or aluminium to the sprocket, thus effectively changing it to a pulley.
To re-assemble replace the hollow shaft including the new pulley and fasten with the circlip into the brass bracket and screw the bracket back on to the dial assembly.

Since the chord used for dial drives seem to be obsolete one now has to improvise! Fishing line of 2-3mm diameter, or something very similar, can be used as long as it will not stretch with time. This is threaded over the path shown in the accompanying diagram and wound round the new pulley 1.5 times. Care should be taken not to loose the 3 small pulleys over which the chord runs. The tricky part is getting the line knotted and in good tension at the same time.

Three hands can be useful!
Note the knot should run in the lower slot and must not be so large that it prevents the chord running freely. The purist might like to add a small spring at this point to maintain tension but, in practise, the loading on the chord is so light that this has not been found necessary. Ensure that the knot is located at the right hand end of the slot as shown and reverse steps 8 to 13 above, taking care again not to move the pot. shafts.

Rescue the aluminium pointer from the old belt by carefully cutting it off, but leaving enough material to fix it to the new chord with a blob of Evo-Stick. A small white mark exists on the tuning dial just above the words above "L.O." and the pointer is located opposite this mark. Ensure again that the tuning shafts are in the fully anticlockwise position i.e. lowest frequency during this process. Before full re-assembly check the free running of the chord from the low to high frequency end of the scale. Reverse steps 1 to 7.



The unit can then be placed in the main frame and checked for frequency calibration against known signals. Providing the tuning pots have not been moved during the process and the pointer has been fixed in the position shown, calibration should be as per the original.

The HP432A Microwave Power Meter
A legend in its own lifetime

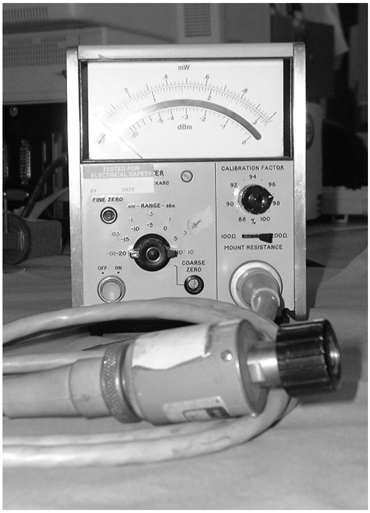
John A Share

Introduction

There can be few items of Instrumentation that have been in production for over thirty years. The HP432A Power Meter was designed in the late 1960s, entered production in the early 1970s and will cease to be available after September 2003. Entirely discrete analogue design, devoid of a digital display and built onto antique style circuit boards, these units continue to be used in numerous situations and applications throughout the world. They often appear on the second user market at a fraction of their original cost (currently c£10,000) and are a tempting purchase.

Model options relate to the location of the thermistor input connector and the provision of the battery pack. Option 001 has the rechargeable battery pack, option 002 has a thermistor input on the rear panel wired in parallel with the connector on the front panel whilst option 003 has the thermistor input only on the real panel.

Thermistor cables are available in 10ft, 20ft, 50ft,100ft and 200ft lengths and are suitable for use with any type of thermistor be it 100 ohm or 200ohm. The thermistors themselves are available in Coaxial or Waveguide Mounts and have a frequency range of 0.01 to 40 GHz. The most common heads are the Coaxial "N" type. They resemble a pepper pot, and whilst HP list only two models other manufacturers have produced alternatives. Most second user HP432As seem to be equipped with the 0.01 to 10 GHz head (HP478) and these are relatively inexpensive. The 0.01 to 18 GHz head (HP8478) is rather scarce and usually priced at a premium. Alternative 8 GHz heads are plentiful and are often priced at a few pounds each.



A calibration table should be found on the side of the head. With an HP unit this takes the form of an aluminium label with calibration points marked with a centre punch. For an accurate power reading at a specific frequency the HP432A calibration switch is set to the value marked on the label. Unless this requirement is critical such precision can be ignored and a median calibration setting used for the entire frequency range.

Coaxial heads have an internal resistance of 200 ohms. More exotic heads designed for direct connection to wave guide have an internal resistance of 100 ohms hence the two range options on the HP432 front panel.

The Package

For the Power Meter to be regarded as a usable system it needs to consist of three items, the meter itself, a head cable and a thermistor head. The battery is not essential, it is optional, and an Option 001 meter will function without this item. Head Cables are quite specific and, whilst it would pose few problems to make a suitable four core cable, locating a source of the connectors poses a serious difficulty. Fortunately these cables are available on the surplus market for a few pounds each. Low frequency thermistor heads (e.g. < 8GHz) are also readily available but intending purchasers are warned against buying heads that have been "repaired" or show obvious signs of tampering. They are notoriously difficult to repair. When purchased from a professional second user source, the package should include the workshop manual. These same sources are sometimes willing to make photocopies or supply originals but at a significant price. Without the manual these units are impossible to repair.

Circuitry

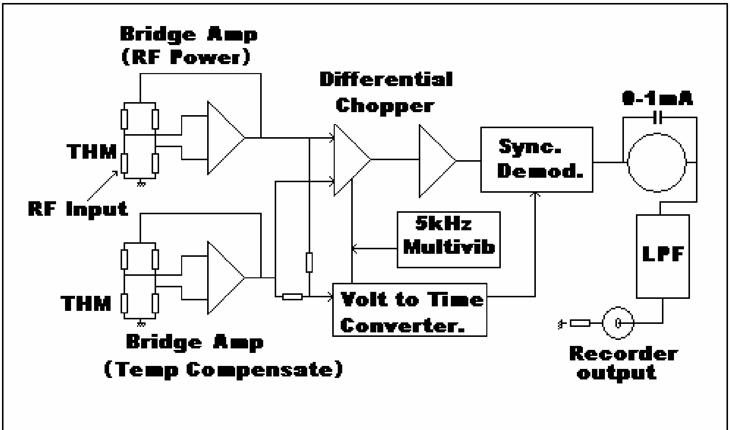
Input power warms the RF Power Thermistor creating a voltage (V_{rf}) proportional to the temperature change of the Thermistor, room temperature is sensed in an identical manner to create a compensation voltage (V_{comp}). The change due to the RF input is therefore the difference between these two voltages ($V_{rf} - V_{comp}$). Amplifying small dc signals poses serious base line stability problems and the classical solution was to convert the dc signal to ac, amplify, and then rectify back to dc. The term for this technique was a "Chopper Amplifier". Today there are Integrated Circuits that perform this role, e.g. 7650, and certain Operational Amplifiers are available to compete with their specification. In the HP432A this is performed using discrete components. Unfortunately ($V_{rf} – V_{comp}$) does not provide a "Power" measurement, it is necessary to include an "R" term as in the basic equation $P=V^2/R$, it is also necessary to remove the V_{comp} term. To achieve this the two values are summed ($V_{rf} + V_{comp}$) and the resultant value modifies the width of the 5 kHz signal to the synchronous demodulator. The demodulator output is therefore a pulse train of height proportional to the difference, and width proportional to the sum of V_{rf} and V_{comp} . The capacitor across the meter is an essential part of the integrating circuit that causes the meter to display $(V_{comp})^2 – (V_{rf})^2$. The Low Pass Filter is an integral part of an Auto Zero Circuit and the user output "Recorder Output" is a dc level devoid of the pulse train into the meter circuit.

Maintenance

Experience with a number of HP432A over the past fifteen years has clearly demonstrated that they are very reliable. There has never been a catastrophic failure and they are extremely easy to repair.

Whilst HP insist that of the sixty semiconductors used in the design over half are selected devices, it has been found that they can be generally replaced with off the shelf items of similar specification and some quite simple device testing to form matched pairs. One or two differential pairs are exotic devices and might prove difficult to locate they can, with a little ingenuity, be substituted.

The majority of failures have been due to ageing of switch contacts and judicious application of switch cleaner has affected a ready cure. A single case of a "sticky meter" was traced to the meter scale flaking and depositing flecks of white paint into the move-



Simplified Outline of HP432A Power Meter.

ment. Despite numerous attempts to halt the decay the problem continued. Eventually the meter scale was photocopied, the original paint removed entirely and the photocopy glued in place ... an invisible repair! Damaged heads must be regarded as irrepairable.

Applications

The HP432A has seven ranges, the most sensitive being 10 microwatts. The open end of a Coaxial Thermistor mount is an N type male connector. Numerous attachments can be fashioned onto N type sockets so that the HP432A functions as a very sensitive signal detector. In this manner wave guide modes have been determined by fashioning a dipole and rotating it to detect TE or TM modes within a guide that was excited from a signal generator. A closed loop will function as a general “sniffer” for detecting signals in a multiplicity of circuits. Low power circuits, for example amplifiers, can be connected directly to the thermistor head; it then functions in a dual role as load and detector.

For measuring real power, the least sensitive range of 10 milliwatts might appear somewhat limiting until it is realised that Coaxial Attenuators are plentiful and very inexpensive. Typically -10db, -20db or -40db with frequency ranges up to 18 GHz, when used in conjunction with directional couplers it is possible to measure immense amounts of power. In one application we use a -30 db bidirectional coupler / -30db attenuators / HP478 heads to measure the forward and reflected power in a system using a Klystron Amplifier that delivers 750 watts at 8.2 GHz.

Integrating an HP432A into a computerised system is made difficult because of the absence of remote range selection. In reality, this has not been found to be a real obstacle. However, the inherent drift that requires repeated “Zero Resets” can be automated. The reset circuit is simply a biased toggle switch and a pair of relay contacts wired directly in parallel with the switch and driven from PC interface card is a simple solution. The switch is readily accessible and there is adequate space to mount a small relay on a custom PCB within the enclosure.

On the rear panel of the HP432A is

a "Recorder" output. This has a maximum amplitude of 1.000 volts for full scale deflection of the panel meter, irrespective of sensitivity range. Connecting this to the input of an ADC within a PC through a long lead is prone to problems, mostly related to earth loop induced 50 Hz. The Recorder shield is not grounded at the panel; it is bushed. Ideally the connection to this should be differential but the connector is not two pin! Similarly digitising a remote single ended 1 volt signal with a 10 bit ADC invites problems of noise and hence loss of resolution. A working solution has been to pass the HP432A output through a passive 50 Hz notch filter, amplify the signal using a differential chopper amplifier to 10 volts and connect this signal to the single ended ADC input through the long lead. The difference is impressive!

There is a delay between the application of the input signal to the head and the moment when the output is a true representation. This is not fixed but depends on the step amplitude of the input signal. For the Analogue display this is of little importance. However, when using a PC, it is all too easy to forget that there is a significant delay (in PC processing terms). In general a full range step input change of 10mW requires 1 second of delay but this can be reduced to 100 ms with virtually no loss of accuracy. Where the input is fluctuating it is possible to reduce this delay even further but experience has shown that 20ms is probably the lower limit and the thermal lag of the head then becomes noticeable.

Conclusion

Many of my HP432As are well over twenty years old. They have proven to be incredibly reliable and have been used in many, many roles and applications. They are simple to configure, easy to use and give unambiguous readings. Down time figures would indicate that they have been by far the most reliable items of equipment I have ever purchased.

The new price of an HP432A is daunting; their availability on the second user market at one hundredth of that price is certainly attractive and, when seen at surplus sales at one thousandth of that original price, they represent the steal of all time.

Repairing HP Power meter heads

Tom Williams, WA1MBA

Several people have asked me if it is possible to repair an HP 478A-style thermistor mount (Power Head for 431 / 432 power meters). Surprisingly, in SOME cases, the answer is YES!

The first step is to do triage - figure out if the head is completely fried. Some of these heads will withstand 1 watt for 10ms and appear broken but can be brought back to life.

The only way to tell if a thermistor resistance is within specification is to measure the resistance under simulated operating conditions. This is complicated and I'll outline how to do it at the end of these notes. In the case of a power head that pins the power meter (one way or the other) you don't care about resistance being within spec - IT ISN'T. What you do care about is whether it is totally fried or can be brought back.

To determine if it is repairable, measure the resistance between pins 1 and 2, and also between 3 and 4. The resistance should measure between 1000 and 5000 ohms. Use a good digital ohmmeter on the 20K scale so that it draws very little current. Don't leave the meter connected for more that a second or two. The absolute maximum current that a thermistor can take is 13 mA. If you read an open or a short, you have blown thermistors, and might as well trash the unit. If they are both about 1K to 5K, proceed to repair/adjust.

Don't forget to check the cable... if there is an open, or a faulty connection, this can cause the same symptoms as a bad head.

The procedure is as follows

1. Take the can off the N connector by removing three set screws. Slide the cover can off.

2. Plug the unit into your 431 / 432.

3. Set the resistor mount ohms to 200 (assuming you are repairing a std 478A - some of the microwave heads are 100 ohms, so set the control accordingly).

4. Turn the power on.

5. If the meter is pegged down scale

a. Set Range to 10mW

b. Set ZERO and VERNIER to mid-range (on the 432, carefully count the turns of the screw-driver coarse- zero pot and set it to mid range)

c. On the back of the thermistor mount there are two small brass screws. Take your time. Turn one of the screws 1/8 (yes 1/8) turn clockwise. Then turn the other screw 1/8 t clockwise.

If there is a sudden jump in meter indication when advancing either screw, back it off 1/8 turn, and do NOT advance that screw any further. If either screw bottoms, do not apply force - it is likely that if a screw bottoms, the thermistor is fried.

d. The best result is when at some point in the alternating "turning of the screws" the meter rises.

Once it starts to rise, trim it to zero by turning each screw a little.

6. If the meter is pegged upscale,

a. Set meter to Zero (as in 5 above step b)

b. Set RANGE to highest power position which will not peg the meter.

c. Turn one of the little brass screws counter clockwise (leftwards) to obtain a meter reading of half the deflection noted in step b.

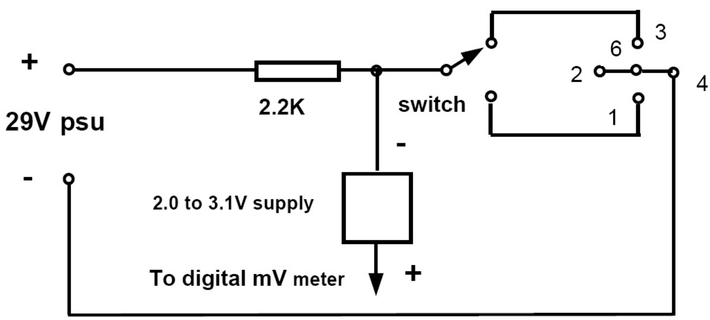
d. Turn the other screw counter clockwise to zero the meter. If it is impossible to zero the meter, at least one of

the thermistors is fried.
7. Replace the cover.
Congratulations! The head works, is no longer calibrated, but is probably within one dB of its original calibration. Check it with a known good source or another meter to get a calibration factor.

Note. Some waveguide type units have only one adjusting screw. Follow approximately the same procedure, adjusting only the one screw.

If you really want to check a mount for thermistor match you need a 29V DC power supply, a second high resolution supply with floating terminals adjustable 2 to 3.1 V, a switch, a 2200 Ohm resistor and a millivolt resolution digital volt meter. You set up the circuit below and adjust the adjustable supply as a bucking voltage to get a very small reading on the volt meter and then switch between the two thermistors to make sure that the readings do not vary by more than .030 volts.

Non-operating units with readings as high as .150 difference can usually be repaired per instructions. I don't recommend that you do this, first of all, its not usually necessary, and second of all, if you get it wrong you can fry the thermistors!



Curing Cyclic Instability and other problems in the Adret 5104 synthesizer

John Hazell, G8ACE

This Adret Synthesiser, new to me, had fully working frequency switches and the reference oscillator was set correctly.

However, the red alarm light had a cyclic flash and the output would jump just a few Hertz.

The Adret is not the easiest unit to work on without extender boards. With the initial investigation limited to checking signals on the linking cables, a blank was drawn as the fault appeared to be everywhere. By substitution, the fault was narrowed down to 6883, the 1MHz Spectrum Module. It was found that the PLL controlling the 50MHz oscillator was hard at one end, the varicap volts being at around 11.5v. The 50MHz is locked to the 10MHz reference so this was duly cranked and the PLL came off its stop with the reference 150Hz low. So the loop was capable of working and the instability disappeared. A 'Quick' fix is always desirable to prevent the unit joining those on the 'too difficult' shelf! A 15pF capacitor was inserted in series with the crystal to raise its frequency. The crystal, it is assumed, must have aged somewhat.

The value of the C is quite critical. 15pF centred the loop at 6.8v. The 10MHz reference could then be varied -100Hz/ +50Hz with the loop remaining in lock.

This was considered an adequate fix.

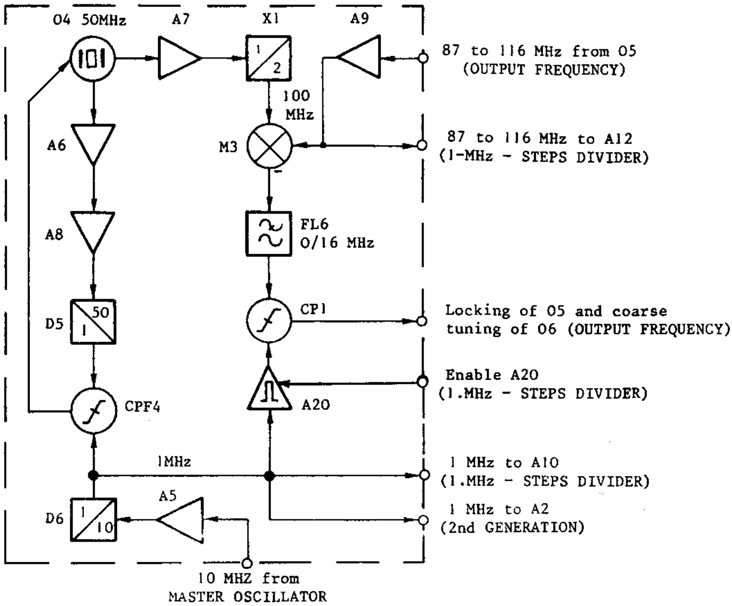
A quick check for the 50MHz loop, therefore, is to vary the reference oscillator over its whole range whilst viewing the Hz digits of the output on a counter. Do this only if you can reset the reference correctly of course. The block diagram of the 6883 module is shown overleaf, along with the circuit for the 50MHz oscillator taken from the Adret manual **which is essential if one is to service this unit ... editor**.

Another stock fault is the prescaler chip, a Plessey device. I replaced it with two 10131 chips wired as divide by 10 to cure the following fault. The front switches fail to work, the output being stuck up around 130MHz. Chris, G8BKE has had and fixed this fault recently. I have had it and so has another local, so it's fairly common.

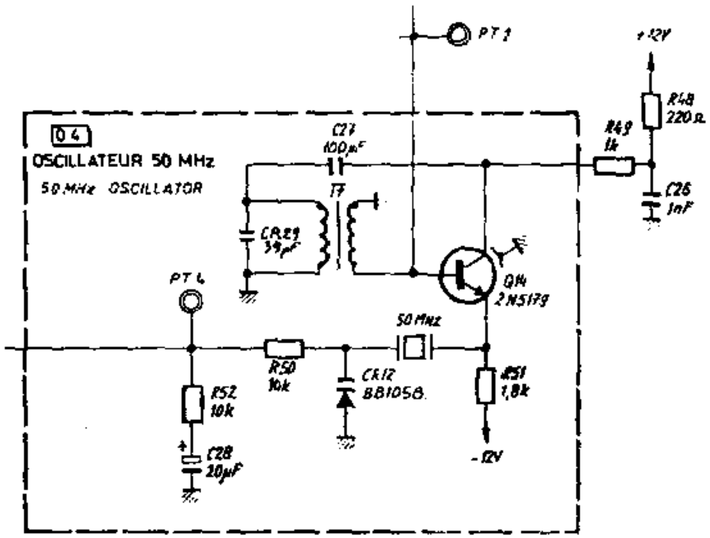
The other little snippet is the genlock input.

It's always active waiting for a 5MHz signal. If the internal oscillator is, in my case, within 20Hz, the oscillator simply snaps in. Another unit, newer, needs to be within 4Hz.

There's no need to touch the master frequency knob, unless of course you want it to be right without the external signal.



04 50MHz
A7
X1
A9
87 to 116 MHz from 05
(OUTPUT FREQUENCY)
100
MHz
A6
M3
87 to 116 MHz to A12
(1-MHz - STEPS DIVIDER)
A8
FL6
0/16 MHz
D5
CP1
Locking of 05 and coarse
tuning of 06 (OUTPUT FREQUENCY)
Enable A20
(1.MHz - STEPS DIVIDER)
CPF4
A20
1MHz
1 MHz to A10
(1.MHz - STEPS DIVIDER)
A5
D6
1 MHz to A2
(2nd GENERATION)
10 MHZ from
MASTER OSCILLATOR



PT 1
+12V
R48
220Ω
04
OSCILLATEUR 50 MHz
50 MHz OSCILLATOR
C27
100pF
R49
1k
T7
C26
1nF
CR21
39pF
PT 4
Q14
2N5179
50 MHz
R50
10k
CR12
BB105B
R51
1,8k
R52
10k
C28
20µF
-12V

More Notes on the Adret 5104 Synthesiser

Mike Scott, G3LYP

A while back, I bought an Adret 5104 which, on testing, I found to be faulty, giving output on about 125MHz which was not controlled by the switches on the front panel. I then obtained a second unit which was working, and by swapping plug in modules between the two units, I found that the faulty module in the first unit was the output module designated 6887.

At the time I had a quick look inside the module and noticed that some of the PCB was a bit brown, indicating that it had been quite hot. As the second unit was working I decided to keep the first one as spares.

I wanted to make a signal source for testing AO-40 antennas and receivers using the Adret and a modified DK004 LO source. Unfortunately when I switched on the Adret, I found that it was no longer working, giving output on about 74MHz which, like the first unit, was not affected by the switches. As I had previously established that all the modules in the first unit, with the exception of the output, were working, I thought that there was a good chance that the fault in the second unit would be in a different module. Not so! The output module was again the faulty one.

The Adret is a very complex device and after going through the manual several times, I still don't fully understand it. However, there was no choice but to have a go!

One of the problems in servicing the units is that, because of the modular design, most of the circuitry is in metal boxes plugged into a motherboard and it is virtually impossible to work on the modules while they are in situ.

The manual refers to extender cards which are used to make the modules accessible while remaining connected to the motherboard.

The 6887 output module has only an eleven pin connector and not all of the pins are used (some of the modules have 35 pin connectors) so in my case it was possible to solder wires to the mother board and to the PCB and so get access to the module on the bench while still connected to the main unit. There are also a number of SMB connectors on the top of most of the modules and by using a commercial "T" adapter or fabricating one from bits of coax and SMB connectors, it is possible to monitor the signals at many points in the unit. The manual recommends the use of a spectrum analyser and 100MHz oscilloscope, neither of which were available to me. Nevertheless, with a frequency counter, a multimeter, and a 20MHz 'scope, it was possible to do quite a lot.

The Manual contains several fault trees, but working through these confirmed what I already knew, that the fault(s) were in the output modules. These modules contain three basic circuits.

The first is an oscillator which covers 90 – 120MHz, and which provides the output of the Adret after buffering and amplification.

The second circuit is a virtually identical oscillator covering 87 – 117MHz which is fed back to the 1MHz Spectrum module (6883).

Both oscillators also provide output, via separate buffers to a mixer (an MCL device similar in appearance and pinout to the SBL- 1).

The 3MHz IF from the mixer is

passed through a multistage filter to the 11 pin connector and via the motherboard to the Phase/Frequency module (6886). There are also a number of other bits of circuitry associated with the control of the two oscillators. On the first of the faulty modules, a check of the signals at the various SMB connectors indicated that there was no output from the 87 – 117MHz oscillator sub-unit. A check with a multimeter revealed that the 2N5179 base-emitter junction in the buffer between the oscillator and the output socket was defective. This transistor was replaced with a BFY90 which is listed as an equivalent. After replacing the module in the motherboard, the unit started to work normally.

Unfortunately, tracing the fault in the second unit took considerably longer. I established that both oscillators were working, although the output of the 90 – 120MHz oscillator was at about 75MHz. The 87 – 117MHz oscillator appeared to be working normally and was controlled by the switches on the front panel. On checking carefully, I discovered that it worked up to 114MHz but, on switching the 1MHz switch to the next step (115MHz), the frequency fell back to 107MHz. On checking the DC level on the "Approach" line to this oscillator I found that the voltage at Test point 1 (All the five test points on this module were connected to colour coded wires and brought to the top of the module to enable monitoring while the module was in situ) rose steadily to a similar level as on the now working module (approx. 8v) but on switching to the next step, the voltage fell back to about 6.8v. On the good module the level rose to 8.8v. The "Approach" line is fed via a 4.7k resistor (R9 to a group of four transistors (Q8 – Q11) which compare the "Approach" signals from both oscillators and feed a signal back to the Phase/Frequency Comparator. On the working module the voltage at Test point 3 (the bases of Q8/Q9) was only marginally lower than at TP1 and ranged from 3v to 8.8v depending on the switch settings. On the faulty unit the voltage at TP3 was very low (less than 1v) on all switch settings. The immediate conclusion was that one of the transistors in this group was faulty. Unfortunately, checking with a multimeter suggested that all were in good condition, but as the voltage from the meter was only about 3v, I decided to replace all four. Luckily, all were common types and easily replaced. This was a good decision, because on re-installing the module, all was well and the output ranged from 90 – 120MHz with all the switches functioning as expected.

Having a working unit for comparison was a great help, and it would be useful if the Microwave Committee could keep a record of Adret owners who might be willing to help owners of faulty units.

As a final note, not having the correct mains lead, I replaced the mains connector on the back of both my units with IEC connectors fitted with integral filters.

The MACOM PE334 Synthesiser

Andy Talbot, G4JNT

At rallies and Microwave Roundtables there have been for sale a range of microwave synthesiser bricks.
The data sheet supplied with these had several errors and this note has been provided to help users get them working.

Power supply requirements:
+20V at approximately 400mA for the analogue circuitry on the pin marked
+5V at 150mA for the logic

Frequency control is via a 26 way IDC type connector, with TTL level signals on 12 parallel lines. Logic level is positive true, and the lines float high to a logic '1', they need to be pulled to ground for a logic '0'.

The pin configuration is detailed below, shown looking at the synthesiser module.

There are four types of synthesiser module covering different frequency bands and step size:

SAC12-01
4.7 - 5.1GHz Step size 0.625MHz

Pin 1 is missing.
25 23 21 19 17 15 13 11 9 7 5 3 (1)
26 24 22 20 18 16 14 12 10 8 6 4 2

SAC32-00
6.35 - 6.76 GHz Step size 0.25 MHz

Connections :

1 to 11	No connection	19	B5
12	B11 (MSB)	20	B4
13	B10	21	B3
14	B9	22	B2
15	B8	23	B1
16	B7	24	B0(LSB)
17	B6	25	Ground
18	No Connection	26	Ground

SAC32-01
6.65 - 7.16 GHz Step size 0.25 MHz

SAC42-01
10.95 -11.45 GHz Step size 3.125 MHz

Programming
The three lowest frequency units are programmed by applying a straightforward binary code to the 12 parallel programming lines.

The frequency of each of these can be set by determining the value of the programming number, N, then applying this value to the lines.

The 11GHz module requires two separate six bit programming numbers to be calculated.

Got to the next page ….

SAC12-01	Freq (GHz)	= 2560 + 0.625.N	or N = 1.6 * (F - 2560)
SAC32-00	Freq	= 6160 + 0.25.N	or N = 4 * (F - 6160)
SAC32-01	Freq	= 6144 + 0.25.N	or N = 4 * (F - 6144)
SAC42-01	Freq	= 8000 + 125.L + 3.125.M	

For example, to set the SAC32-00 to 6400.25MHz,
N = (6400.25 - 6160) * 4 = 961
Convert to Hex / Binary = 0x3C1 = 001111 000001

Only frequencies that are within (or close to) the allowed frequency bands can be programmed, thus the full range of programming numbers possible are not all valid, any attempt to set these invalid codes results in the synthesiser VCO free running. It is possible that by adjusting the VCO centre frequency by tuning the cavity, the lock range can be moved. This has not yet been tried.
The X band module, SAC42-01 is more complex in its frequency setting than the other types. The control lines do not all follow a simple binary sequence but are weighted as follows :

B0 - B5 Six bit binary value M giving 64 * 3.125 MHz steps up to 200MHz
B6 - B11 Six bit binary value L giving the 125 MHz steps.

Note that in some cases a specific frequency may be obtainable with more than one programming code, codes within the ranges shown in the examples below have been tested :

10850 MHz	=	0x5A0 = 010110 100000	Below lower freq spec.
11000 MHz	=	0x600 = 011000 000000	
11125 MHz	=	0x628 = 011000 101000	Two codes for this freq
11125 MHz	=	0x640 = 011001 000000	" " "
11275 MHz	=	0x670 = 011001 110000	
11400 MHz	=	0x6B0 = 011010 110000	
11571.875 MHz	=	0x6FF = 011011 111111	Above frequency spec.

Ailtech Noise Sources

Charles S. Osborne, K4CSO

At the risk of losing some choice hamfest finds, a posting of an Ailtech noise source on EBay reminded me of something I've discovered and should pass along about those noise sources:

If you have an Ailtech 7616 noise source or variant that won't calibrate, i.e. seems dead, don't give up on it. The noise source may actually be a separate module inside, made by Noise Com, MSC, or unlabeled others. Usually what goes is the attenuator which is used to calibrate the unit. It's a thick film hybrid. Over tightening the N-connector can twist the connection, cracking the attenuator's ceramic substrate, open circuiting it ... Or, at very least, it will be very unpredictably intermittent.

Now the good news ... most of the units I've disassembled had SMA connectors on the noise source internally and a normal SMA inline attenuator following. Disassemble the unit down to the noise source and use it with a new SMA inline attenuator and you are often times back in business.
I've bought a number that looked like they had been drug up and down the flight line, if not outright submerged. Most were in the $5—$10 range since the labels were gone and they were clearly a long shot. I'm 4 for 5 working once I tossed the cracked or fried 16dB input attenuator.

The noise source is typically 250,000 K or 28-30dB ENR without the 16 dB attenuator. The attenuator value is factory selected to get the unit into the 15.4 dB ENR = 10,000 K noise customary range. Changing to a good 18GHz 26dB attenuator (even if its a 16+10dB arrangement) gives you a 5dB ENR unit which is better for measuring extremely low noise preamps (if your noise figure meter will calibrate at 5dB ENR).

If you go to a conference where a noise figure meter is available, you can measure the same preamp with a known head and compare your "rebuilt" one and get even closer on the ENR re-calibration at that spot frequency.

I guess my secret's out ... no more $5 noise sources once this group gets turned loose on them at the hamfests! Now, if I could just revive my 20 year old HP8970A to use and calibrate the ones I have.

Agilent says they obsoleted the HP8970A in 1987 and its successor the B version in 2001. Story of our lives, right?

The interesting thing is they are offering $10K to get them out of our hands in a buy back deal on one of the new 8974A $40K noise figure analyzers. I've got about as much chance of getting one of those as a 100ft yacht.

Below: The Ailtech 7514 **P**recision **A**utomatic **N**oise **F**igure Indicator (PANFI)
This item is, from time to time, found on the UK surplus market and is well worth acquiring if you can also find or make a noise head.

Care and feeding of Minicircuits Lab ZHL42 amplifiers

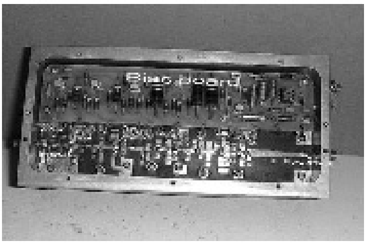
Sam Jewell, G4DDK

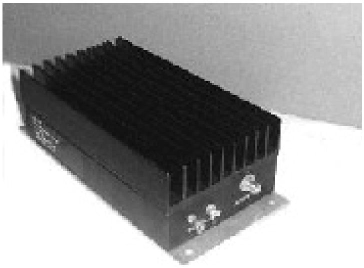
Minicircuits Lab introduced these amplifiers in the late 1980s as high power, linear gain blocks for commercial and industrial lab use. They were quite expensive and used then new GaAs technology.

A wide range of variants have since appeared including 10MHz to 4.2GHz types (ZHL-42W), 40dB gain types (ZHL-4240) and lower gain 100mW output amplifiers (ZHL- 1042J). However, the GaAs FET input stage was very prone to being destroyed by too much signal. A very common failure situation was simply turning on signal generators that didn't remember the last set level and that would power up with maximum output of perhaps 50mW.

This was almost certain to cause the amplifier to fail. When they first appeared they were rated at +10dBm maximum input, this was later revised down to +5dBm and my own advice is to ensure the input doesn't exceed 0dBm.

The purpose of this article is to show that these amplifiers can easily be made to work again and are (arguably) more useful to the amateur community after this simple repair than in their 'as new' condition. Why? How? Read on!

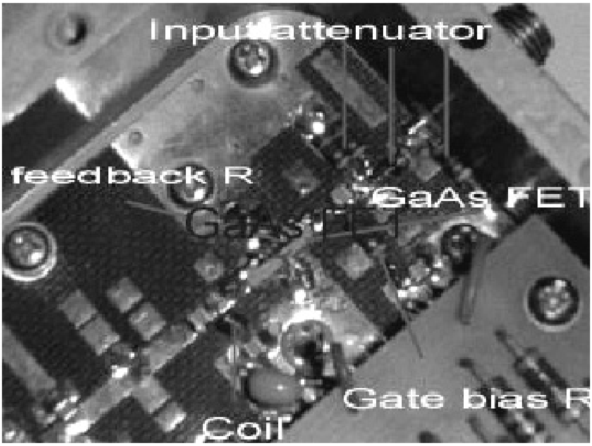




700MHz to 4.2GHz 1Watt output, 30dB gain, 1dB ripple broadband

After trying unsuccessfully to replace the blown input FET I tried the more practical approach of merely strapping out the input stage. First, check this is where the fault is. (You can be better than 90% certain it will be). It is easy to prove this is the stage at fault as the bias condition around the device will be found to be inconsistent with, usually, the negative gate bias voltage on the first stage non existent. Carefully remove the 820R gate bias feed resistor, 680R feedback resistor together with series chip capacitor, close-wound drain feed inductor and then the GaAs FET device. Cut a 1mm wide, 12mm long strip of thin copper tape to bridge between the gate and drain connection points. That's it!

With this 'repair' the gain of the amplifier will be found to have decreased by about 10dB. However, it is now possible to increase the input signal level significantly without fear of damaging the amplifier. I have applied up to 100mW without damage after making these changes. The great thing about this mod is that the achievable



output power from the amplifier is now significantly increased.
The following table shows the results from one such amplifier 'repaired' in this way. You will note that the frequency response is now noticeably less flat than 1dB. However, for amateur purposes this probably doesn't matter too much. Some additional gain can be obtained by removing the input 3dB attenuator and replacing it with another copper strap.
These amplifiers are an ideal way to make a multi-band transverter or to boost the output of your PLL Brick oscillator to produce a useful output power for a 23 - 9cm, personal beacon.

Frequency	**432MHz**	**1.3GHz**	**2.32GHz**	**3.4GHz**
Output power (sat)	1.4W	2.25W	1.7W	2.0W
For input power	Not measured	10mW	10mW	25mW
700mA at 15V	Rising to 875mA at saturated output power			

Finding a Use for Defective Marconi Power Meter Heads

Jan-Martin, LA8AK

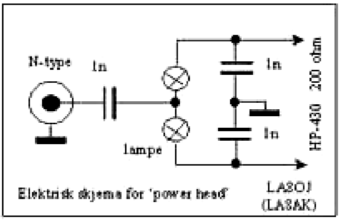
The following information was gleaned from the website of the late Jan-Martin Noeding, LA8AK.

It's presented here as a starting point for what could be the "answer to a maiden's prayer" as far as those of us who have blown Marconi power meter heads in the bench drawer.

Although the diagrams are not very sharp, as taken from Jan-Martin's webpage, they show the general idea of using miniature watch illumination lamps to make a workable sensor.

The editor would be very interested to hear from anyone trying this out and also from someone with a source of the miniature lamps. A check at the Conrad website, the company mentioned in the diagram below, failed to find anything suitable.

Using defective MI 6444 probe to repair old HP 431B and Narda power meters



Repairing power head using miniature watch illumination lamps from Conrad in Germany
As shown for Marconi power head