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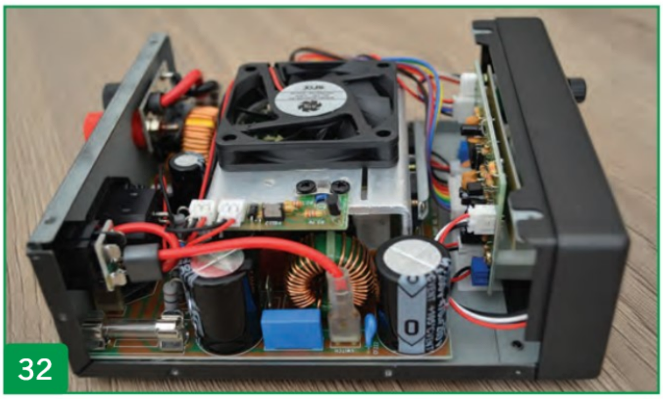

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Cover image: QRM finding and elimination with Ian, G0CNN.

RadCom THE RADIO SOCIETY OF GREAT BRITAIN'S MEMBERS' MAGAZINE

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New Products

Elecraft K4

Waters and Stanton have received the first sample of the EU certified Elecraft K4 SDR direct sampling transceiver. Weighing less than 5kg, this new 100W radio has a large colour touch screen, covers 160m to 6m and includes general coverage receive from 100kHz to 54MHz. The basic K4 has full dual receiver capability, with separate receive antenna provision. The high-definition panoramic display offers split screen option to enable displays for both receivers. Solid state switching provides silent, full break-in operation (QSK). For CW, PSK and RTTY there is on-screen decode and provision for transmitting PSK and RTTY with a paddle key. USB input includes the provision for a PC keyboard or mouse.

Extensive interfacing for datamodes is offered and a copy of the user manual can be pulled up onto the screen. For those interested in remote control, there are a wide number of options available within the K4. There is also provision for the inclusion of internal 144MHz and 432MHz transverters that are planned for next year. Like all Elecraft radios, the K4 can be easily updated using the included USB data cable.

For those wanting even more capability, there is the K4D that includes the KRX4 receiver board. This adds an extra set of bandpass filters and second receiver direct sampling to enable full diversity reception. Options for both models include AT4 Auto ATU and SP4 Speaker.

Currently, all K4 series radios are being built to order at the Elecraft facility in the USA. The anticipated delivery time is currently around two months from date of order, but customers can call Waters and Stanton for the latest update on 01702 206835. All orders placed in the UK at Waters and Stanton will benefit from the extended UK warrant supported at their Portsmouth Service Facility. This avoids Customs complications and delays that would occur if for any reason, the radio had to be returned to the USA. www.hamradiostore.co.uk

Radio carry case

LAM Communications has launched a new carry case for those that enjoy portable or DXpedition operating. It can be seen via their website and is available to view in the show room.

The Vine Antennas RST-RigCase is a portable transceiver carry case providing a robust and heavy duty solution for safe transceiver mobility. It comes with high quality pre-cut foam to modify to your preferred size and shape depending on the equipment you wish to store. The case is sized to fit many of the popular transceivers and receivers available, and is secured with a two-step locking system for extra security. If the equipment is being taken portable, such as a Field Day, then you'll be pleased to know it is waterproof to IP66.

Measuring 420 x 350 x 205mm and weighing 3.14kg, the case retails at £59.95 from LAM Communications. www.hamradio-shop.co.uk.

Extension speaker

The Vine Antennas RST-SPK-L super-size extension speaker is now available from LAM Communications. It has ABS enclosures and a sturdy metal grille and is suitable for free standing or wall mounting via an adjustable fixing bracket (supplied) as well as alternative M6 threaded mounting posts. Measuring 280mm in height and 220mm across the front, the depth including speaker bracket is 210mm. The super-size speaker retails for £69.95. www.hamradio-shop.co.uk.

70cm 6-element Yagi

M0CVO Antennas has released a new antenna for the 70cm band. The 70cm 6-element Short Fixed Array Yagi. A short fixed array provides the gain of a longer antenna in a more compact design. This antenna offers 11.02dBi (8.87dBd) forward gain and is ideal for portable or fixed station use. The feedpoint type is optional (BNC, N-Type or UHF) and can be chosen at the point of order. The boom length is 1m and it can be used for horizontal or vertical polarisation. The antenna retails for £64.95 and you can read more at https://m0cvoantennas.com.

Antennas

A 15 element Yagi beam array for 70cm

One characteristic associated with a high gain Yagi beam array is its narrow beamwidth. This can become a useful feature when working a weak station located directly ahead of the antenna. If other stronger stations are in the same general direction as the weak station, then a narrow beamwidth can be helpful by reducing interference when the working frequencies are close.

High gain Yagi beam arrays working on VHF and UHF bands become more practical possibilities to construct than for HF because of the shorter wavelengths involved. The theme for this month's Antennas is a 15 element Yagi beam array for the 70cm band. The concept of the antenna is shown in **Figure 1**, with **Photo 1** showing the antenna in use.

PHOTO 1: The 70cm band 15 element Yagi beam array in use.

The antenna's design model

The antenna was designed and modelled using the MMANA-GAL antenna analysis application **[1]** based on a central operating frequency of 432.25MHz with the 'default' ground option selected. The model predicted the 15 element Yagi beam array as having a free-space gain of about 14.5dBi (12.4dBd), a front-back ratio (F/B) of around 19.4dB and a beamwidth of 32°.

Figure 2 shows the antenna's predicted radiation patterns with the antenna modelled horizontally polarised at 7m above ground level (AGL). At this height above the (default) ground, the antenna's predicted vertical radiation pattern shows many forward lobes have formed ahead of the antenna, with the five lowest lobes having radiation angles of 1°, 4°, 7°, 10° and 13°. These five lowest lobes are at shallow angles to the horizon and should allow both local and longer distance stations to be worked when tropospheric conditions are favourable and the path towards the distant station is not locally obstructed. The higher angled lobes can be an advantage if there are obstructions nearby because they may allow these obstructions to be cleared, enabling RF signals to be directed towards the horizon if the trosopheric conditions allow. At 7m AGL, the best predicted gain for the antenna was 20.3dBi, corresponding with the second lowest forward lobe at 4° angle of radiation (shown with a small red arrow in **Figure 2**).

The antenna's predicted horizontal radiation pattern shows significant RF signal concentrated ahead of the directors, with a 3dB beamwidth of around 32° (ie between 344° and 16°). The model has predicted side lobes developing at angles of around 43°, 65° and 81° either side of the main forward lobe. The rear lobe is comparatively broad with rear side lobes either side of it at 107° and 128°. However, the rear lobe is small and the predicted front to back ratio (F/B) is around 19.4dB.

Construction

Using the MMANA-GAL predicted dimensions as a starting point, the 15 element Yagi beam antenna for the 70cm band was constructed using 8mm diameter aluminium tubing for the elements. To support the elements, the antenna's boom was made from a 2.5m length of 20mm diameter aluminium tube. The actual dimensions for the elements and their spacing on the boom are shown in **Figure 1** and **Table 1** for reference. To hold the antenna's reflector and thirteen director elements in position, fourteen clamps were needed, which were salvaged from a scrap antenna.

The Yagi beam array's dipole centre was constructed using an ABS box and its dipole used 8mm diameter aluminium tubing, as shown in **Photo 2**. Each of the dipole's leg tubes were passed through "push-fit" holes drilled equally either side of the ABS box. To construct the dipole, a tight fitting plastic rod was inserted between each tube, leaving a 2mm gap. Solder tabs were attached to the tube ends using self-tapping screws passed through holes drilled in each tube's end, which also held the dipole together. These tabs allowed the coaxial feeder cable to be soldered to the centre of the dipole. The ABS box was mounted onto the antenna's boom using self-tapping screws so that the dipole was correctly positioned in place on the boom. To weatherproof where the dipole's legs

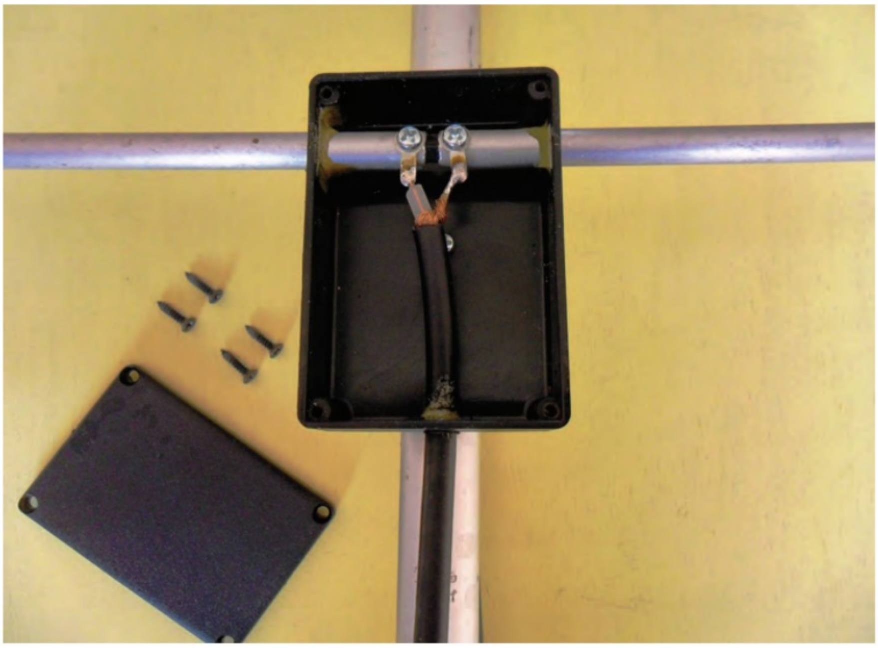

PHOTO 2: Homemade dipole centre used for the 70cm band Yagi beam array made using an ABS box, aluminium tubes and a plastic rod.

and coaxial cable entered into the ABS box, sealant was applied around the tubes and cable. Once completed, the ABS box lid was attached to its base and a smear of sealant passed around the outside to protect it.

A commercially made boom mount was used to support the antenna; it was salvaged from a scrap television aerial. To hold the boom-support, containing the antenna, to the top of a mast a commercially available clamp was used as shown in Photo 1.

There is an alternative solution for the element clamps. If it is necessary to make up element clamps, then the elements could be held in place using clamps made from modified plastic pipe clips, as shown in **Photo 3**. This approach has been previously used for other home constructed 70cm Yagi beam arrays and found to be a satisfactory arrangement. These pipe clips can be obtained from hardware stores and builders' merchants. The element clamps were intended to support 8mm diameter elements on a 20mm diameter boom. Therefore, the plastic pipe clips obtained were the type intended to hold 20mm diameter tubing. Each pipe clip was shortened to allow it to sit on top of the boom. A groove was filed, using a small round file, along the pipe clip's top to accommodate the 8mm diameter element that had a 4mm diameter hole drilled at its centre. Then a long M4 bolt, wing nut and washer were used to secure everything in place.

Matching

The 15 element Yagi beam array was fed using 10m of low-loss 50Ω coaxial cable that included a current choke 1:1 balun formed by winding the coaxial cable into an inductor. The 1:1 balun comprised four turns of the cable with an internal diameter of 20mm and was held together using cable ties. The 1:1 balun was situated as close as practical to the dipole connection **[2] [3]**.

Testing, adjustment and tuning

This design for a 70cm Yagi beam array has been constructed several times and the lengths of the reflector, driven element and first director (D1) may need to be slightly varied to obtain a good match to the 50Ω coaxial feeder cable. In addition, the space between the driven element and D1 may need to be varied to obtain the best match. The differences between these antennas were possibly caused by slight variances in the diameter of the aluminium tubes used for these elements. Best practice is to cut these elements about 10mm longer than suggested, then adjust the space between the driven element and D1 to obtain a reasonable match. Once the space between the driven element and D1 has been optimised, the length of each element can then be gradually reduced in turn to obtain the best match. Another approach used to tune the antenna was to add flush fitting aluminium tubes of 50mm in length to the ends of the reflector, dipole and D1 to enable the length of these elements to be adjusted to obtain the best performance. Once the lengths of these elements were found, the length of each adjustable element was measured to enable it to be replaced with an element of the correct length. Alternatively, once the adjustable elements had been optimised, their adjustable sleeves could be held in position using small self-tapping screws

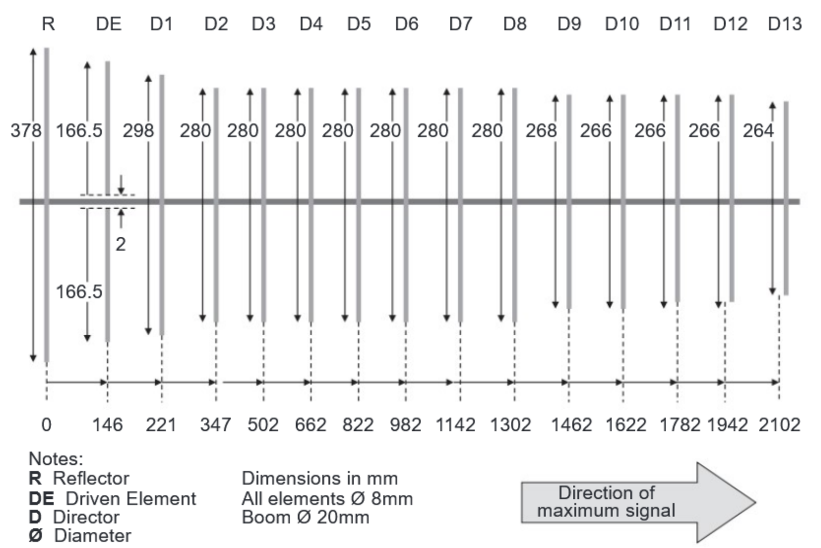

FIGURE 1: Concept of the 70cm band 15 element Yagi beam array including the actual dimensions used (centre-to-centre spacing). See also Table 1.

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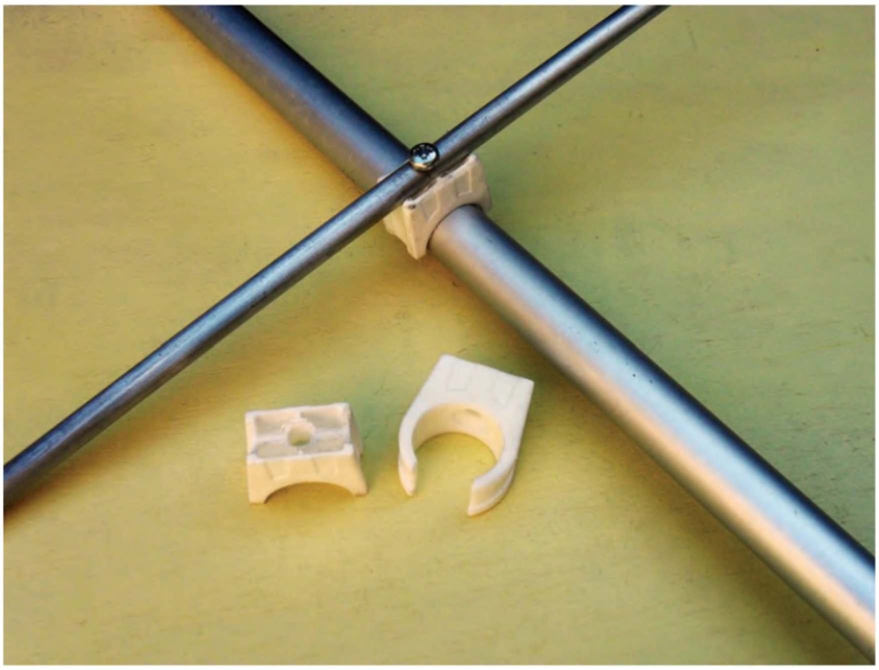

PHOTO 3: Example of an alternative homemade element clamp using a modified 20mm plastic pipe clip.

passed through suitable holes drilled through each sleeve and the element (as used for the dipole in Photo 1).

The antenna was tested on 432.25MHz using an MFJ 269c antenna analyser. An SWR of 1:1 was obtained after adjustment. When the antenna's adjustment and tuning had been completed and with the Yagi beam array at 7m AGL, the antenna was driven with an initial power level of 2W that was increased to 20 watts. The SWR continued to remain low at under 1.3:1 when measured between 430.05MHz and 434.5MHz, with an SWR of 1:1 measured from 431MHz to 433MHz. The feeder used for the tests was 50Ω coaxial cable, whose measured loss was about 3dB. The antenna's directivity was monitored by tuning the transceiver to the GB3UHF beacon, turning the antenna and observing the level change on the S-meter. This indicated a beamwidth of about ±15°.

Details of the dimensions of the aluminium tubes and boom used to construct the 15 element Yagi beam array are summarised in Table 1 for the antenna shown in Photo 1.

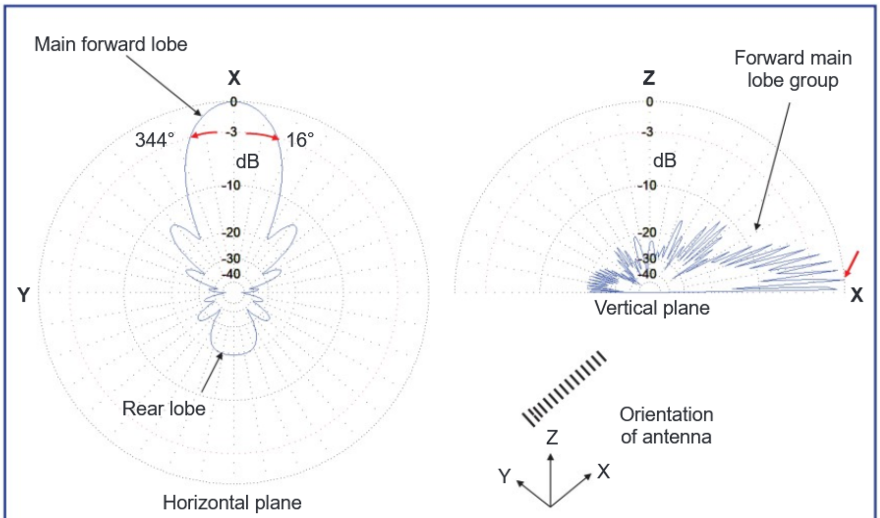

FIGURE 2: Horizontal and vertical plane radiation patterns predicted by MMANA-GAL for the 70cm band 15 element Yagi beam array at 7m AGL.

On air performance

The 70cm band 15 element Yagi beam array was found to be very directional, with stations that were apparently poor weak signals becoming an extremely good 'copy' when the antenna was pointed directly at them. Taking the antenna's gain as 20dBi gives a power gain of approximately 100 times over an isotropic radiator. Therefore, when running an RF input power of 10 watts at the antenna, the effective isotropic radiated power (EIRP) is about 1kW. This EIRP has proved to be very effective when working weaker more distant stations. The antenna has enabled many contacts to be made during UK activity contests and occasionally into northern Europe during good tropospheric conditions.

Websearch

[1] MMANA-GAL basic V3.0.0.31, freeware antenna analysing application. Original code by Makoto Mori JE3HHT. MMANA-GAL Basic and MMANA-GAL Pro by Alex Schewelew, DL1PBD and Igor Gontcharenko, DL2KQ. 1999 onwards

[2] RSGB *Radio Communication Handbook*, 14th edition edited by Mike Browne, G3DIH. Section 14, Transmission Lines, pages 14.13 to 14.15

[3] RSGB *Transmission Lines Explained* by Mike Parkin, G0JMI: Section 6, Transmission Lines, Antennas and their Connection, pages 64 to 77

TABLE 1: Actual dimensions and element centre-to-centre spacing used for the 15 element 70cm Yagi beam array. All elements are 8mm diameter and the boom is 2500mm of 20mm diameter.

Element	Length	Distance
Reflector	378	n/a
Driven element	335 †	146
Director 1 ‡	298	221
Director 2	280	347
Director 3	280	502
Director 4	280	662
Director 5	280	822
Director 6	280	982
Director 7	280	1142
Director 8	280	1302
Director 9	268	1462
Director 10	266	1622
Director 11	266	1782
Director 12	266	1942
Director 13	264	2102

† Driven element consists of 2 x 166.5mm legs with 2mm gap (see text)

‡ Length and position of this element may be adjusted for best match

Review

Astatic SWR Power Meter PDC2

The Astatic SWR Power Meter model *PDC 2* is described as a three-function test instrument that claims to be able to measure standing wave ratio (SWR), RF power and also can act as a field strength meter. It appears to be designed for the Citizens' Band market and has an impedance of 52Ω.

Out of the box

On opening the transparent packaging, the meter appeared to be well-made, despite being light. Both front panel meters were clearly marked and easy to read.

Curiously, the instructions within the first paragraph refer to a "Model PDC-1 test instrument" and also despite many helpful tips for employing the meter, such as providing a table correlating SWR with power loss, they make no mention of how to use it for its third intended function, a field strength meter. Nevertheless, it comes with a small wire antenna that screws into a socket on the left of the meter to fulfil this third function. Obviously, you will require a patch lead with at least one PL259 plug to connect it to your rig, the other plug being dependent on the output socket on your transceiver.

In use

I set it up and used it with my Yaesu FTDX101MP transceiver, which has a built-in SWR and power meter, hence I was able to compare the readings on the PDC2 with those on the Yaesu.

There are two power range settings: 10W and 100W, which are selected by a sliding switch on the front panel. I tried it initially on 28MHz as it is supposed to be a CB users' product and the readings were in accord with those on my transceiver, both in terms of SWR and power output. I also tried it on all the other HF bands and 50MHz and, once again, the readings were in line with those on the Yaesu. The instruction leaflet, which is written in English and French suggests that it should not be used at frequencies above 148MHz, so by inference, it should work fine on both the 4m and 2m bands.

Conclusion

This is a an inexpensive SWR and power meter and I found it as accurate as the inbuilt meter in my FTDX101MP. If I could only have one piece of test equipment, I would choose a SWR meter – I believe every amateur operator should have one, particularly if experimenting with different aerials. Amateurs returning to the hobby after years of being off the air will also find this a useful bit of kit if they are 'restocking' the shack.

It costs a very reasonable £39.95 and is a worthy addition to any shack. Being so light and compact, is ideally suited to portable operation too. All in all, I found it very straightforward to use. I would like to thank Moonraker (UK) Ltd for loan of the meter, https://moonrakeronline.com.

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The Astatic PDC2 SWR and Power Meter with Field Strength Antenna attached.

An SDR transceiver for the amateur bands Part 1: Receiver

A talk by Heather Lomond, M0HMO at the RSGB Convention in 2018 finally inspired me to experiment with software defined radio (SDR). I had already tried seeing if I could get back on 3cm using a RTL USB SDR dongle with a laptop and a satellite LNB. I realised that it was a huge leap up from trying to operate 'blind' with an analogue transceiver and converter, which I had been using previously. Julian, M0UGA then persuaded me to invest in a HackRF SDR module and start experimenting. **Photo 1** shows the major components I ended up employing; it was an interesting journey.

My first steps, with a single HackRF operating with SDRAngel software for Windows, were not a success. Trying to transmit and receive on a single HackRF usually resulted in several seconds delay in changeover – and mostly crashed the software. Hence, after the Convention I decided it might be possible to create my own software, using the GNU-Radio environment, to produce a Python coded program to run the HackRF.

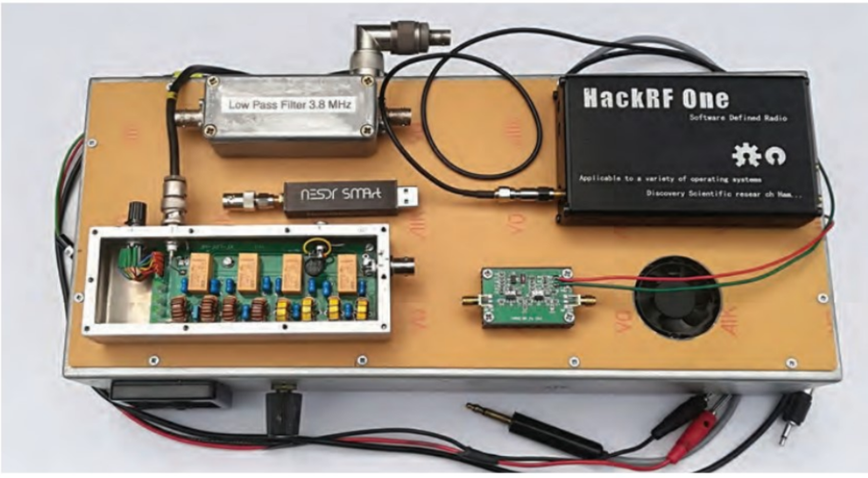

PHOTO 1: **General view of the hardware that eventually made up my SDR transceiver.**

In order to do this, the first step was to create a receiver that is capable of checking the performance of any home-made transmitter. This is where our old friend the SDR dongle comes into its own. Available for next to nothing, the basic RTL USB plug-in dongle (designed for TV reception) will do, but for a bit extra, it is worth purchasing one with a TXCO high stability oscillator and a metal case with SMA connector like the lower one in **Photo 2**. G3WCB found that the Noo-Electric NESDR-SMart was also a good device and I have used these ever since. The basic model covers about 26MHz to 1.65GHz and is very sensitive and accurate. Although it can be used with any Windows PC and downloaded software, I decided to start as I meant to go on and try to use it with Linux and later with a Raspberry Pi.

As with all software projects, it is essential to have a constant version of any underlying operating system. Hence the reason I chose the Raspberry Pi. Once I had produced something that worked, I could keep the entire project together on a single memory card and dedicate an inexpensive Raspi to it if necessary. Fortunately, the latest version of Raspi software for the Pi3, 4 and 400, called 'Buster' 'Jan-May 2021', comes with GQRx. This is a great program that produces a good general coverage amateur radio receiver 'out of the box'. Not only is it capable of receiving CW/SSB/FM etc but it has good dynamic range and displays both a waterfall and spectrum that can be expanded to extend across an entire HD screen attached to the Pi. Tuning from 26MHz to 1.6GHz continuously, spectrum widths from a few kHz to 1.8MHz can be viewed and displayed just like a real spectrum analyser (although not as accurate). The display shows over 90dB of dynamic range, but the device overloads above -20dB, so the dynamic range is just over 70dB. However, for the purpose of testing and setting up home built (or commercial) transceivers it is certainly satisfactory (see **Figure 1**). This shows the reception of the active portion of 70MHz during a UKAC contest with my GNUradio transmit software also running on a Raspi400.

I have also used the RTL dongle receiver on the receive side of my antenna coaxial relay, using an extra relay so that it could be used as a panoramic adapter for my older analogue transceivers.

Now that I had a stable, wideband VHF-UHF receiver working on the Raspi, its coverage could be extended downwards into the HF bands by creating a downconverter from a double balanced mixer and a local oscillator. Some newer versions of the RTL module do have the extended coverage

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FIGURE 1: **Screndump of GQRx receiving 70MHz during a UKAC contest (left) plus my GNUradio transmit software (right).**

PHOTO 2: 'RTL' SDR dongles.

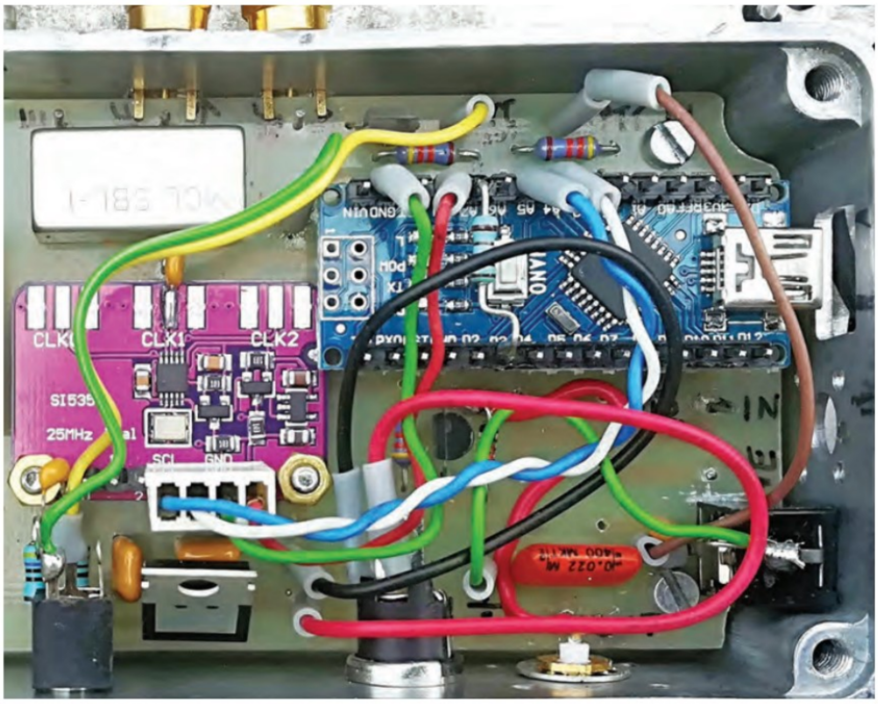


PHOTO 3: Assembled Si5351 synthesiser/mixer section.

down to MF – but not the NESDR-Smart, though it has such good frequency stability. M0ICR suggested trying out the Si5351 synthesised clock generator module as an alternative to a crystal oscillator. The module promised an inexpensive way to produce any frequency from audio to VHF and it can be controlled by an Arduino Nano, with which I was familiar. The output frequency of the Si5351 is held in its memory after the initial boot, so an extension to the Nano synthesiser software can be used to create a transmit sequencer for the transmitter stages of the project. The resulting assembly is shown in **Photo 3** and the interconnections in **Figure 2**; it will be described in the second part of this article.

The si5351 can be obtained from Adafruit and they have a useful library for the Arduino to simplify a lot of the complex programming of registers in the Si5351 chip **[1]**, so that any frequency can be set with only 15 lines of code. I used the Arduino IDE programming environment, which is available for the Raspberry Pi (under Prefrences > Add/Remove Software> Search for Arduino, then install AVR_development IDE).

When programming the Si5351, it is important to try to choose integer division ratios from its internal VCOs so that the minimum of spurious is produced on its output. I chose x28 multiplication from its 25MHz crystal to lock the VCO to 700MHz and a division ratio of 25 to produce 28MHz as the module's output. Any frequency will do, but I found that 28MHz produces a received signal from 80m as 31.5-31.8MHz with virtually no unwanted mixer products visible. The subtraction mixer products are below the frequency cut-off for the RTL-SDR and the result was a pretty clean, interference free display on the GQrx receiver spectrum. I found there was no need for any preamplification on the mixer as the RTL-SDR is designed to be low noise at VHF – indeed I usually put a 6-10dB attenuator on the RTL input to reduce the HF sky-noise (and VDSL interference) from the antenna. Later, I added low pass filters for each band in the antenna feed, which are needed to remove harmonics of the transmitter amplifier.

It must be noted that the Si5351 is designed as a multiple clock generator and not a synthesised pure sinewave oscillator. It produces a lot of mixer products and although it

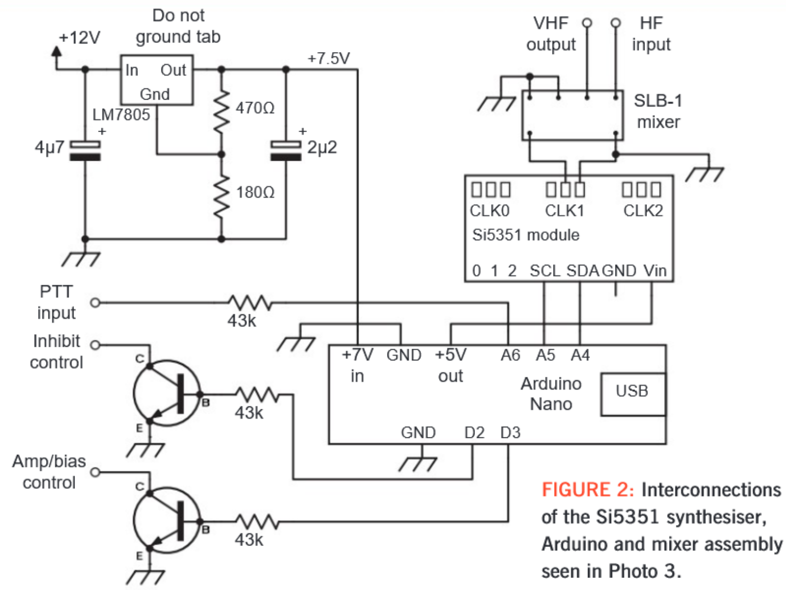


FIGURE 2: Interconnections of the Si5351 synthesiser, Arduino and mixer assembly seen in Photo 3.

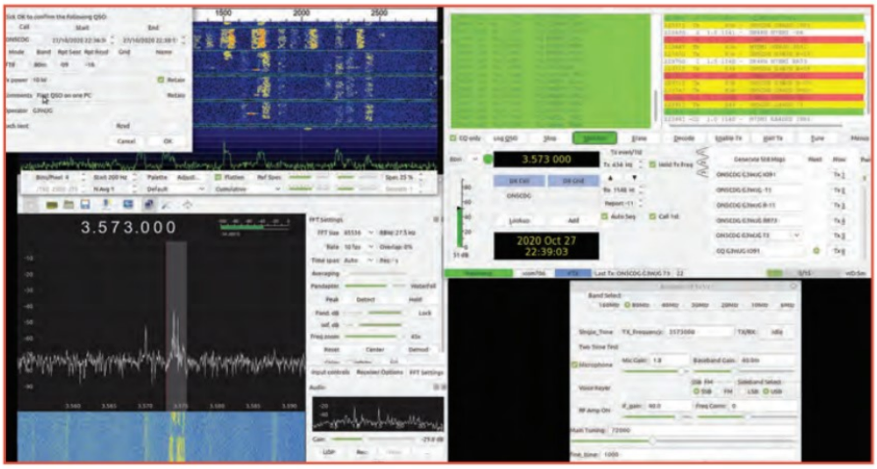


FIGURE 3: The somewhat busy screen of my SDR transceiver operating on a PC-based Linux-Mint platform.

works OK as a receive LO over a selected narrow band it is not suitable to produce a wideband transmit up-converter. I found It needed to be kept well away from low level preamplifiers in the transmit chain (see later).

The mixer I used was a Mini Circuits SBL-1 which is readily available new or used. This device is fairly wideband and its local oscillator (LO) input can be driven by the few milliwatts directly from the Si5351 module. The SBL-1 has good isolation between the ports (~60dB) so that the 28MHz local oscillator is reduced to a level that does not overload the RTL dongle or radiate from the antenna port.

Using this upconverter, the GQrx receiver will now cover the shortwave bands by tuning from 31.5-56.45MHz, but the frequency on display can be altered by subtracting the Si5351 local oscillator frequency (enter minus 28.000000MHz in the box marked 'LNB LO' on the input control page of the control panel), which will make the large display of GQrx read the correct shortwave frequency being received.

Any frequency error due to the Si5351 module can also be added or subtracted from the 28.000000 as well, so that the frequency on display is corrected to within a few Hz. It is best to correct any error in the RTL-SDR master oscillator by adjusting the value (in parts per million) in the box marked 'Freq correction'. This can easily be done on UHF by tuning a local 70cm beacon and adjusting the ppm for zero beat on its true CW frequency (or by observing the audio beat on the audio spectrum monitor and adjusting accordingly).

The Si5351 datasheet shows that you can change the 25MHz XCO on the Si5351 module for an external 3V clock of 10MHz, eg derived from a GPS locked standard. On a spare module I did this and created a locked frequency standard at VHF, the harmonics of which could be received well up into UHF.

At this point I had a reasonable receiver with up to 1.8MHz of panoramic display, capable of tuning from 3.5MHz to beyond 1296MHz. But in addition, the display has 'point and tune' capability such that any signal appearing on the selected spectrum can be instantly tuned to by a mouse click or wheel scroll.

The GQrx program from the 'Buster' operating software will work on a Raspi-3B for a quick test, but the processor runs hot (>80°C and the red thermometer symbol is visible in the top right of the screen) if you leave it on, eg to listen to Radio 2(!) However, with the Raspi400, the temperature rarely goes above 47°C, even with my extra transmitter program and a logger running .

For the 'Full Monty' all-modes transceiver, on a single screen with WSJT-X and map coverage, as per **Figure 3**, I had to resort to using a full size PC operating Linux-Mint 19.3 (but only because I haven't yet figured out how to internally cross connect the audio using 'Jack' or 'Port Audio' software on the Raspi400).

An alternative would be to hard-wire cross-connected audio to a separate Raspi3 to run the MGM decoder software.

In order to make a complete transceiver I use a HackRF One as the transmit generator. These have been out for a while, are one of the least expensive options and they can be purchased in a solid metal case. The HackRF has a capability of generating RF signals from 3.5MHz to 6GHz with an output of up to 10mW, though the output above 3.5GHz is very low (but useful enough to generate a test signal on the 6cm band). Although the HackRF One can also be used as a receiver I found it not as good as the RTL dongle. Also, in order to change from transmit to receive it requires a complete switch of software and drivers, which cause too much delay between overs to be used on the air in a normal QSO. In addition, when operating via AMSAT-Oscar-100, I found it essential (at least initially) to have the capability of full duplex to be able to monitor one's own transmissions.

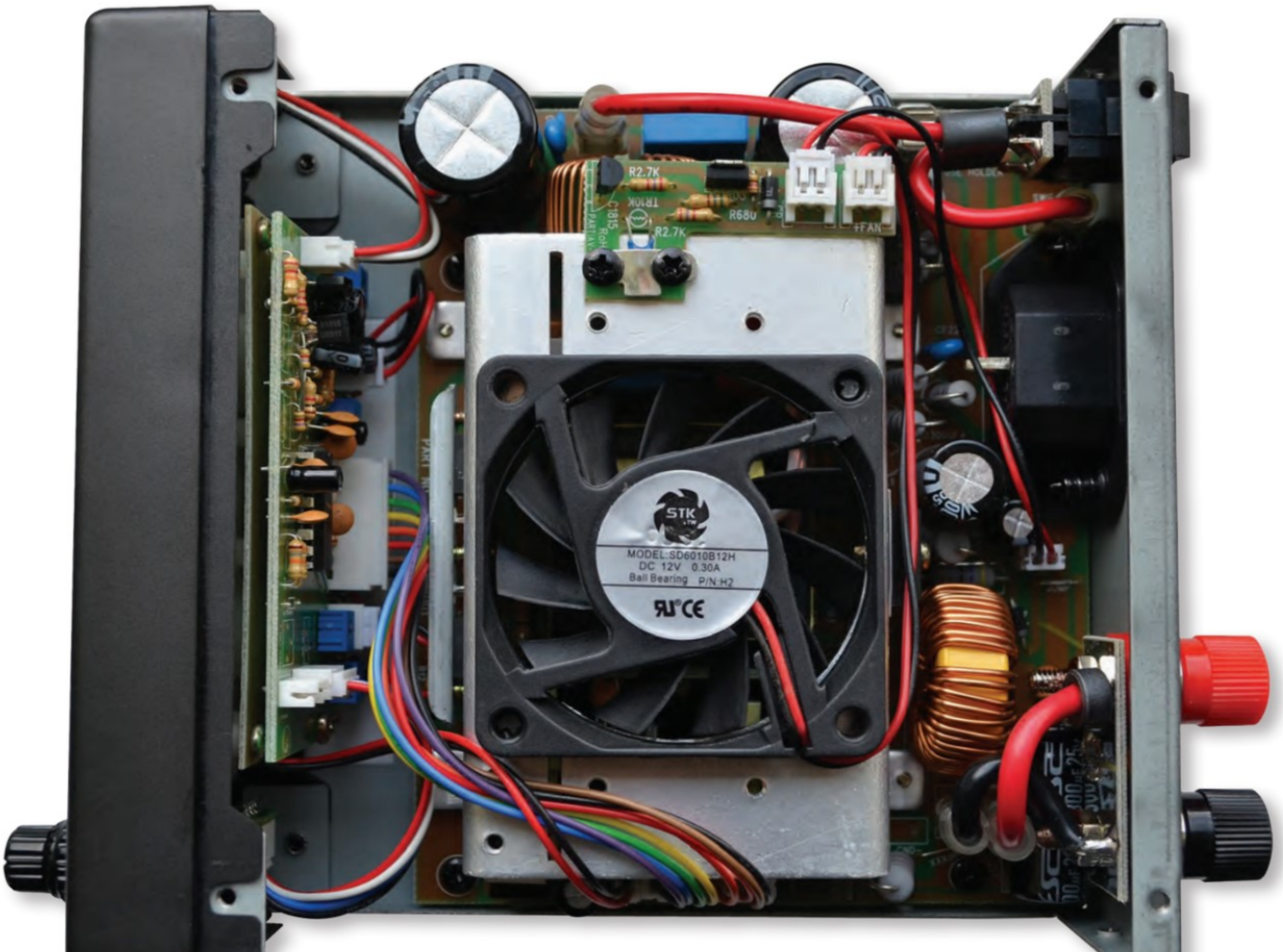
Next month

In the concluding part I will describe the use of Python code that is created from the GNUradio Companion environment, which can also be loaded onto a Raspi 3, 4 or 400 running 'Buster' to drive the HackRF transmitter, plus the sequential switching, synchronising and additional amplifier(s) needed to raise the signal from the HackRF One to form a low power transceiver on the HF and VHF bands.

Websearch

[1] https://learn.adafruit.com/adafruit-si5351-clock-generator-breakout

Watson Power-Mite-NF power supply



Taking the top off the unit, revealed a well-populated circuit board underneath a chunky aluminium heatsink with a 60mm diameter fan on top.

The Watson Power-Mite-NF is a compact 13.8V 25A switch mode power supply, weighing just 1.2kg and small enough to fit in a briefcase or suitcase.

Power supplies have come a long way since the days of older, linear models, with switch mode technology meaning they can be made smaller and lighter. As an example, the the 1980s Kenwood PS-50 13.8V 20A PSU was 173mm (w) x 107mm (h) x 296mm (d) and weighed 7.2kg.

The Power-Mite-NF comes in at 165mm (depth) x 147mm (width) x 55mm (height). To put that in perspective, this means the Power-Mite weighs only 16.6% of the venerable PS-50 and has a total volume of less than a quarter. And there are other advantages too.

Out of the box

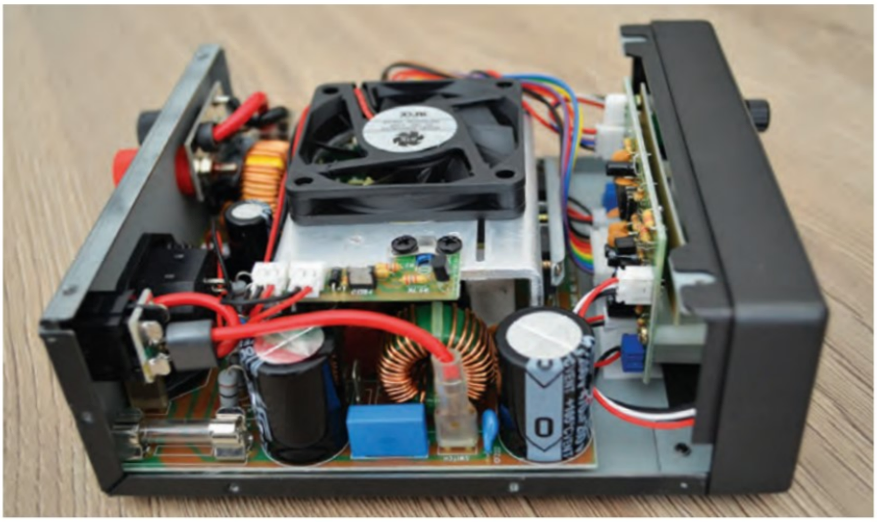
The Power-Mite-NF's output is continuously adjustable from 4.0-16V via a knob on the fascia. It also has a back-lit voltmeter and an ammeter on the front.

The voltage knob has a detent or click position for 13.8V, so it is relatively easy to set it for an average amateur radio.

Also on the front panel is a noise offset knob, which adjusts the switching frequency of the power supply from 19kHz to 31kHz. If you do find noise breaking through on a particular HF frequency it is, therefore, an easy job to just tweak the noise offset until it goes away.

On the back are an On/Off switch and two large threaded connectors for the DC output and I was pleased to see that you can completely unscrew these. I generally use ring connectors

The Watson Power-Mite-NF is a compact 13.8V 25A switching-mode power supply.



There is a user-replaceable fuse on the board that was revelaed when the lid was removed.

on all my radios and it is frustrating to find some power supplies do not allow you to completely remove the positive and negative connectors to allow you to put them on – so well done Watson!

Also on the back is an AC input socket that uses a standard DIN connector (the unit comes with a suitable power lead with a 13A plug fitted). This measured 175cm long, which should be enough for most applications. Incidentally, the plug was found to be fitted with a 13A fuse. This could easily be reduced to 5A with no effect on the power supply's performance I would have thought. In fact, on the side was a sticker that said it was fused at 220V/4A, which was odd.

The unit is said to be user-switchable to 90-125V AC on the inside so it could be suitable for holiday use in the USA or similar.

Returning to the front of the unit, on the left are two small LEDs. The top one is green and indicates that the unit is operating. The bottom LED is red and acts as a warning that something isn't quite right, such as a short circuit. The Power-Mite-NF reportedly has short circuit protection. It also has overload, over-voltage, and over-temperature protection. According to the supplied graphs, the unit can supply up to 25A continuously until the heatsink reaches 80°C. Power will then be reduced. Another graph suggests that the heat sink could take 8.5 minutes to reach 80°C with a load of 23.2A. To be honest, if you think you are likely to need in excess of 25A continuously I would recommend buying a PSU capable of, say, 45A.

A look inside

Taking the top off the unit, which involves removing eight cross-head screws, revealed a well-populated circuit board underneath a chunky aluminium heatsink with a 60mm diameter fan on top.

There is also a user-replaceable fuse on the board, but no mention of it in the instructions – worth bearing in mind if you own one of these PSUs.

In use

I connected the PSU up to my Yaesu FT-991A and noted that the fan was on as I powered it up. With a current draw of about 2A on receive I wasn't surprised. The fan volume isn't too obtrusive – about the same as a personal computer's fan.

After an hour of operation on receive the case temperature at its hottest point was measured as 29.8°C. So, warm to the touch, but nothing too serious.

I did some noise tests and I could detect noise on Top Band (1.8MHz), where I found an S9 noise signal. A quick turn of the front panel knob and it was gone. It was the same story on 80m (3.5MHz).

I could detect no noise at all on the higher bands, that is 40m (7MHz) through to 10m (28MHz), despite turning the noise offset knob through its entire range.

On VHF and UHF no noise could be detected from the Power-Mite either.

Turning to power tests, I found that when operating CW at 100W, the fan speed did not increase at all on a typical QSO and the case temperature remained at less than 30°C.

On SSB, a five-minute 100W QSO with Pia at school station DL0SGH near Frankfurt resulted in no further temperature increase at all.

Switching to FM and using the 10m band with the FT-991A set at 50W output, the rig was drawing about 11A and the case of the Watson Power-Mite-NF remained under 30°C after a two-minute over.

As the PSU had worked so well I thought it was time for some FT8 activity, so setting the power output of the radio to 50W I stayed on 10 metres and worked a few stations, with no change in the PSU's fan speed. At the end of the session, my infrared thermometer still read around 31°C. In fact, my FT-991A's fan was more likely to increase in speed than the Power-Mite's.

A DecibelMeter app for my iPhone, which was laid next to the PSU's fan, clocked the Power-Mite at 85dB, which was the same as my FT-991A's fan in normal use.

Conclusion

I used the Power-Mite-NF for four days and came away quite impressed. It is really a fit-and-forget addition to the shack and it just got on with its job.The Watson Power-Mite-NF costs £84.95 and is available from Waters and Stanton in Portsmouth, https://hamradiostore.co.uk/.

My thanks to Peter Waters, G3OJV for the loan of the unit for review.

Steve Nichols, G0KYA
infotechcomms@googlemail.com

A high performance loop amplifier

Introduction

This project has its roots in an article that I wrote (as G4UAZ) for *Cheltenham Amateur Radio Association News* in December 1984. It eventually got picked up in *RadCom* Technical Topics by Pat Hawker, G3VA and, latterly, can be found referenced on the internet **[1]**.

I had not critiqued my design thoroughly at the time; I just used it. I was aware that it is not necessarily optimum. So, when I retired over two years ago, modernising, fully developing and optimising the design became a nice follow-on project.

Technology and components have changed a lot in the intervening 35-odd years. Surface mount components give a great choice of available devices, modern PCB manufacturers make small runs of prototype PCBs very affordable (see later); open source circuit simulation and PCB design make professional-grade design and evaluation techniques available from home.

I discovered that manual soldering of surface mount components is entirely possible with a high magnification binocular viewer, fine tweezers and a miniature soldering iron bit: it was not the barrier I had imagined.

Surface mount parts are readily available from the major suppliers online and they will deal with individual private orders. The cost of building the finished device was low, in the order of £20 per unit for the few I made (although this cost needs to be shared over a few devices as some components are not available in one-offs, but many suppliers will provide small quantities).

The other major thing to enable usage of these techniques is computer simulation and computer aided design for PCBs. I used QUCS SPICE to experiment with the designs by simulation to check out their gain and bandwidth before committing to designing a PCB and having samples manufactured.

This sort of loop antenna with a small 1 metre diameter wire (or preferably tube) loop proves very effective when placed down the garden away from the house, provided the amplifier is adequate. Getting it away from mains-borne, ADSL internet and machine noise (including computers, switch mode PSUs, etc) leads to good MF and HF reception. Improvements on the original design in sensitivity and signal handling were aided by modern components. Modern transistors tend to have better gain-bandwidth products than their predecessors, but some shortfalls in the original design's simplicity have also been sorted out. For such a simple thing the new loop antenna does a decent job.

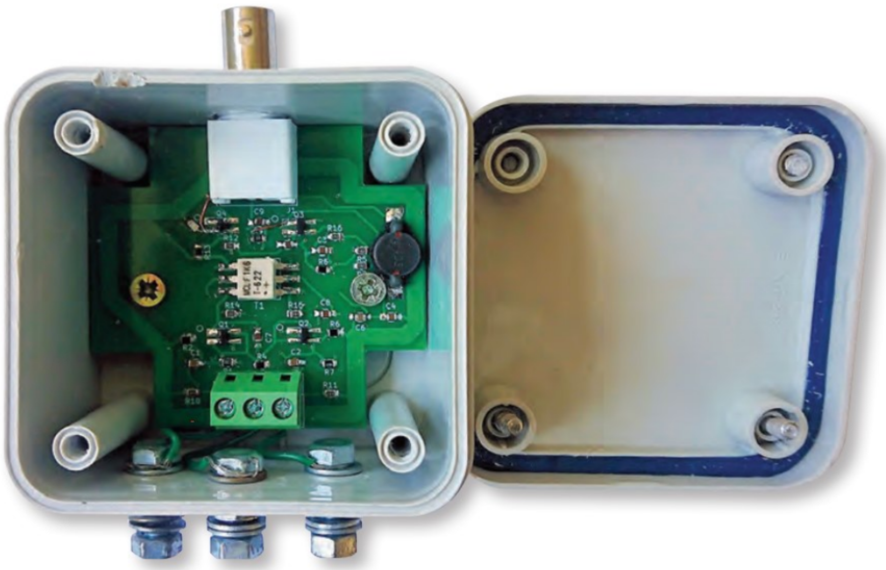


PHOTO 1: Completed loop amplifier, housed in a Schneider type NSYTBS775 weather-resistant case.

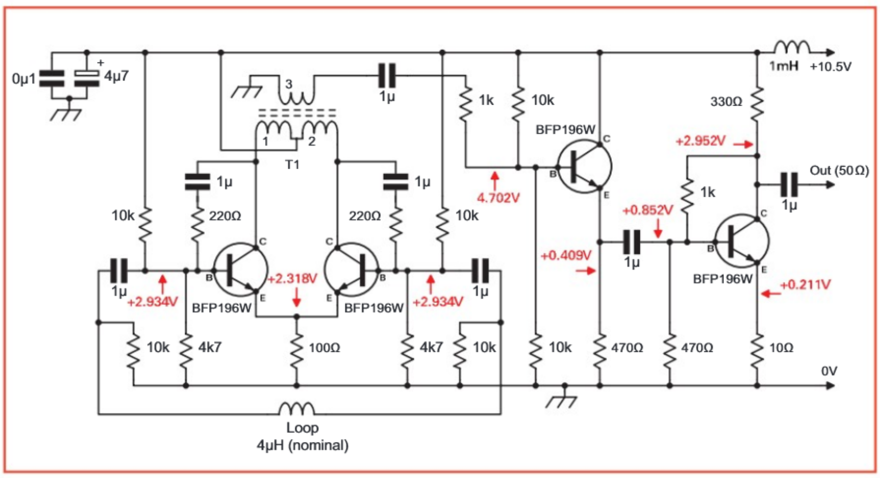


FIGURE 1: Circuit of the loop amplifier, with key voltages marked (from simulation, with 10.5V supply). Transformer T1 is Mini-Circuits T-622-KK81 (1:1:1, 50Ω, 1-200MHz).

In the beginning

The original design was a two transistor long tailed pair with a trifilar-wound output transformer. There was one immediate design flaw with the original output transformer that, if remedied, could make it much better. It was reflecting too low a load onto the long-tailed pair. Ideally it is improved by making it a step down transformer. The output is not much lower as the load is improved, but the transformer becomes trickier to design and make.

With low output the issue of pickup of extraneous signals on a long feeder can be a bit of a difficulty. It leads people to install all sorts of chokes and earthing to try and suppress it. So I decided to elaborate the design with a few extra surface mount components to allow me to use an off-the-shelf 1:1:1 transformer (that has excellent bandwidth and coupling coefficient) and to provide a bit more signal out. It is hardly a significant cost, so I went back to some fundamentals and simulation to consider how best to optimise the design.

Goals

What is wanted is a general coverage MF, HF and low VHF receive aerial that will work with SDR over wide bandwidth and good sensitivity, enough to pull in practically anything on any frequency that is there to be heard, within the reasonable constraints of a general coverage (ie not highly directional) antenna. And, of course, be very small and discrete so that it can be used in pretty much any location. In some cases, convenience of installation should allow it to be placed and aligned away from interferers so that man-made noise pickup is reduced – an increasingly useful feature. Ideally it will be sensitive enough at a very quiet location to hear the background atmospheric noise. As we do not intend to transmit on the loop, it gives us the option of using a very low noise amplifier to extract the extremely small signal that results from an untuned, physically small pickup element.

Small loops are a very good option. “Small” here means a fraction of a wavelength in circumference at the highest operating frequency. When used with the plane vertical they are vertically polarised and can be used close to or even *on* the ground. Again, this is very handy for installation. There is practically no advantage to be had raising it up (and yes, I have checked this). Maximum pickup is off the ends of the loop and there is a null perpendicular. This is the opposite of large loops (as used in resonant quads or delta loops). Small loops will receive vertically polarised high or low angle sky-wave signals, unlike a vertical monopole (which is low angle only in most cases), so they are good

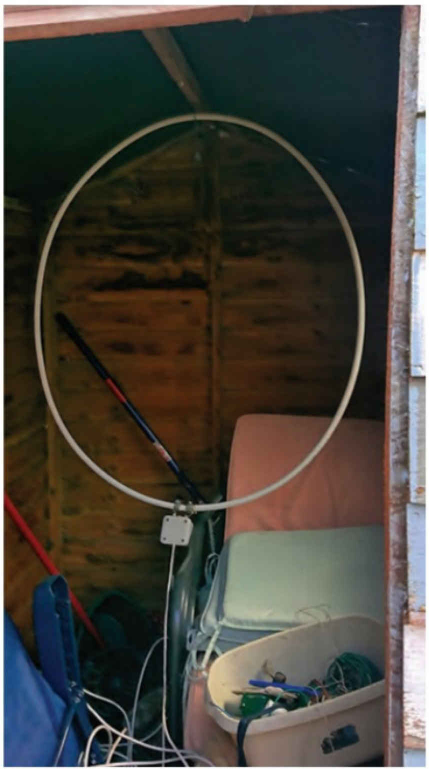


PHOTO 2: A 1m diameter loop can be put in all sorts of unobtrusive places.

for daytime LF short skip reception or other NVIS propagation.

So, what do we require from a loop amplifier to be useful?

- Low noise factor and enough gain such that its output is dominated by atmospheric noise pick up from the loop.
- Balanced input rejecting common mode, which ensures that signals or man-made noise picked up common mode (or as you may say, ‘on the feeder’ or elsewhere) don’t get through to the receiver, and the loop null is preserved. It is best if the output is matched to the transmission line so any forward stray pickup (due to poor balance of the feeder or installation) is not reflected back to the receiver.
- Adequate strong signal handling so the downstream receiver linearity is the limiting factor to spurious signals.
- Input designed to extract maximum energy from the loop in the best way to give wideband coverage MF to low VHF for example with high sensitivity.

This is all a pretty tall order. I also wanted to ensure all components were current (and widely) available surface mount devices, including the odd RF transformer in the design; that power consumption was reasonable; and the design is highly reliable (long ‘mean time to failure’). Designing for low cost and automatic PCB assembly is now consistent with prototyping at home too.

Design basics

First, let’s look at the theoretical characteristics of a small loop. If in free space we have a small vertical closed loop with an electromagnetic field travelling in its plane, there will be a voltage and current induced proportional to the magnetic flux density threading through it (which is related to the electric field strength by the impedance of free space). If we break into this loop and connect our amplifier, it will see the equivalent of an inductor (the loop inductance) in series with a resistor (radiation resistance in series with loop resistance) and a voltage source. The inductance for a circular loop is approximately

$$L = \mu_0 r\left(\ln\left(\frac{16r}{d} - 2\right)\right)$$

where $\mu_0 = 4\pi \times 10^{-7}$ newtons per amp squared, r is the loop radius in metres and d is conductor diameter. That’s about 4μH for a 1m diameter loop with a 15mm conductor.

This inductance presents a reactance that rises with frequency and dominates the impedance looking into the loop from the amplifier, so limits how much current you can suck out. It goes up more or less in proportion to the radius. The radiation resistance Rr is approximately given by

$$Rr \approx 177\left(\frac{A}{\lambda^2}\right)^2$$

where A is the area of the loop in square metres. For a small loop this will vary from micro-ohms to milli-ohms from MF to HF so the physical resistance of the loop conductor will dominate over radiation resistance – and, if ignorably low (as we would want it to be), the main contribution to reactance is the inductance. In a tuned loop you tune that out and can directly couple into the radiation/resistive resistance with a transformer. That requires a tuning capacitor – and retuning for even small changes in frequency. For convenient wideband, in terms of amplifier design, we want the input resistance to be as low as possible such that at frequencies where the inductive reactance becomes low in comparison (such as at MF), the input resistance is at least comparable to that reactance so we get as much signal current out of the loop as possible. At higher frequencies (where the inductive reactance dominates) this is less important, not least because the loop area is larger compared to the wavelength so more signal is available. This amplifier input resistance needs to be a lot lower than normal nominal 50Ω input amplifiers.

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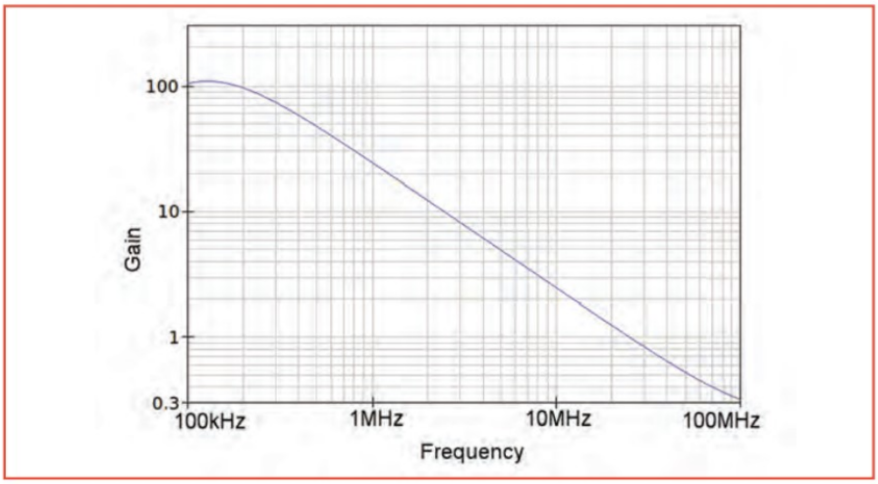

FIGURE 2: Frequency response, as simulated.

The voltage induced by an electromagnetic field in the loop is simply proportional to flux density and area, V= ωBA (noting that flux density is related to the magnetic field by B=μ_0H with H in units of amps per metre). The thing to note is that as the frequency (f=ω/2π) goes up, the voltage goes up proportionately too because the loop diameter is a bigger fraction compared to wavelength. You get a bigger voltage too for a bigger loop at fixed frequency, but this is reduced by the fact the inductance goes up with the radius. In principle we should get bigger signals from bigger loops but in practice I have found it to be marginal – with one exception. If you operate at a frequency below that where the inductive reactance is smaller than the input resistance then the signal falls off sharply with frequency. Above, we find the increasing reactance of the loop inductance is cancelled out by the larger signal voltage so sensitivity is pretty much the same over MF and HF if the design is any good. This frequency below which performance falls off is something of a performance yardstick, so performance at MF is good for critical comparisons of different active loop systems. The increasing inductive impedance with frequency and the fact the induced voltage rises with frequency for a given field strength can equalise each other to give a system response that is approximately flat with frequency within the operating range, a very useful feature.

So, just to summarise what the maths tells us, a small 1m diameter loop presents itself to the amplifier as a small inductor with a very small voltage source in series. If we have a very low input impedance to the amplifier we can use a smaller loop with good sensitivity. In order for the current to be limited only by the size of that inductance at the lowest frequencies (say the 630m band) the input resistance of the amplifier must be a handful of ohms (or less). As the frequency goes up though the inductive reactance increases, tending to reduce the current, but counteracting this the voltage goes up so sensitivity is not impacted. With my design this reactance also plays a part in the shunt voltage feedback of the first stage so the overall gain of this stage goes down with frequency, which is a factor for ensuring flat sensitivity across the bands.

Original design and re-design

My original amplifier design in *RadCom*'s Technical Topics was based on a long-tail pair of transistors giving a balanced differential input and an output transformer converting the balanced output to a single-ended unbalanced signal. Feedback was used to linearise the response and set the bias point. This has two effects. First it reduces intermodulation effects. Also, as the gain is set by the ratio of feedback resistance to the input reactance, it means that the amplifier closed loop gain reduces with increasing frequency due to the loop inductance and this equalises the voltage increase as the loop diameter becomes a larger fraction of the wavelength. When, with reducing frequency, the loop reactance becomes small relative to the amplifier impedance, the system sensitivity falls off relatively quickly. These characteristics are worth maintaining. So the redesign set about optimising the gain, input impedance, feedback and operating conditions for the transistors, also making sure common mode rejection was excellent.

Simulation

I used QUCS SPICE simulations, which allows a great deal of optimisation to be done quickly without picking up a soldering iron. The final design has input resistance of about 5Ω, which I have confirmed by measurement with my NanoVNA. This is a critical factor of the design. A low noise balanced amplifier with this level of input resistance across the band is not easy to achieve, but it is necessary to get a good level of signal out of a small 1m loop across the MF/HF band.

Circuit description

The new circuit as used in the QUCS simulation is shown in **Figure 1**. The circuit was simulated first at DC so the bias conditions can be checked. These are quite close to the measured bias voltages on the built circuit so you can check these with a multimeter as an initial check after assembly. The circuit will operate happily from 12V to 14V with overall current draw of 60-70mA. It will work down to 5V with reduced gain and signal handling.

The circuit figures the same long tail pair input configuration as in the original. The emitter resistor is not bypassed so it presents series negative feedback to common mode signals but differential mode signals flow directly from one transistor to the other and are not reduced by series feedback. There is shunt voltage feedback though, which in the new design is separated from the bias resistors so feedback and bias settings are independent. This stabilises the gain and helps linearise the characteristics which aids strong signal handling and makes a small but useful reduction in the input resistance. The transistors are CATV distribution type devices so are designed to handle strong signals with high gain and low noise up to very high frequencies – yet are very cheap. The differential to single ended conversion is done in T1, the Mini-Circuits transformer in the collectors. As this is 1:1:1 it is necessary that it should only be lightly loaded so there is an emitter follower buffer and a final gain stage to provide a reasonable gain. The feedback in both these stages again stabilises gain and linearises the characteristics. The result is still quite a simple circuit, but which receives signals as well as some of the best designs out there – some of which are quite complex.

This is the simulated gain-frequency characteristic with the loop's inductance included is shown in **Figure 2**. It shows the fall off with frequency that will cancel the anticipated rise in source voltage as the loop area becomes a bigger fraction of the wavelength as frequency goes up. Sensitivity across the HF spectrum should be about constant with a 1m diameter loop. I find LF signals (like amateurs on 630m, or the non-directional beacons just below the Medium Wave broadcast band) are covered quite well.

Strong signal handling of the improved design is fine in my location. The third order intercept point I estimate as +70dBm, which is OK. The 2nd order intercept point is a little less good at +40dBm or so, but there is very little problem in practice. If there was a problem then filtering can ameliorate 2nd order interference (although an input filter would have to be designed to match 5Ω…). As an experiment I made a high pass filter like this to reject MW signals but despite strong MW signals at my location I do not bother using it.

The component ratings are modestly specified to enhance reliability in the style of MIL217C standard that, although old, was aimed at designs like this with discrete transistors and RC components. The standard recommends that

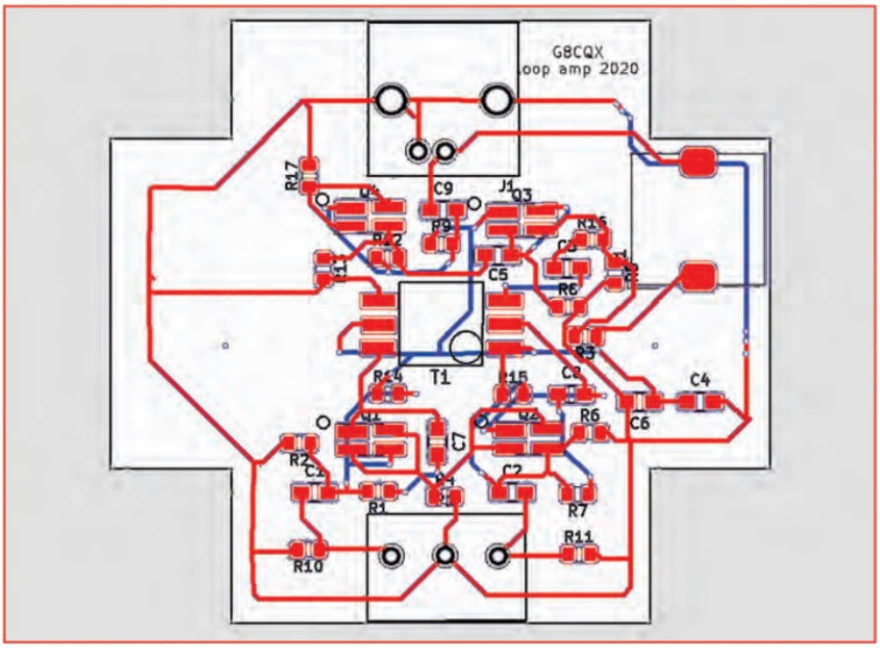

FIGURE 3: **Amplifier PCB design (not to scale). Gerbers can be downloaded from [2].**

stress on active and passive devices is kept well below maximum ratings to provide reasonable mean time between failures (MTBF). I suspect there are some designs that run smoking hot to get impressive intermodulation numbers, but then burn themselves out prematurely.

Practical implementation

I used Kicad for the schematic capture and board layout to generate manufacturing Gerber files for the PCB manufacture. You can upload the file set onto a manufacturer's web site and order prototypes at the drop of a hat. The longest delay is mail from the far east, if that is where you source your boards from. *[PCB manufacturers offering ten boards for US$5 (plus postage) have been mentioned recently in* RadCom *– Ed].*

In order to make hand-assembly more feasible I chose resistors and capacitors in the largest commonly-available footprint, 0805 size. The PCB design shown in **Figure 3** and **Photo 1** includes a few components not shown in Figure 1, such as components to extract power from the feed, an input terminal block and coax output connector. I should stress though that it's not essential to use this PCB layout and it's perfectly possible to construct the circuit using conventional through-hole components, given a suitable substitute for the BFP196W transistors. Only the transformer is really critical.

I identified a very cost effective environmentally sealed ABS box, Schneider type NSYTBS775, which the PCB was designed to fit. There will need to be a 5mm hole in one face for the BNC and 3.5mm holes in the opposite face for screw terminals. The BNC hole might need a bit of drifting in order to get the board in at an angle. You'll have to fit the external connection bolts have to be fitted after the board is put in. Of course many other sealed housings would do.

Power

The loop amplifier is powered over the coax feed (the components for this are on the PCB but not shown in Figure 1). A supply of 12-14V is best. It will work down to 5V but signal handling will not be as good. Power is injected at the receiver end using a 'bias tee', which can be homebrew or commercial. Mine were mainly 'ugly' construction but I have designed a PCB and you'll find the details of this, plus all the Gerber files and a Bill of Materials for this and the amplifier at **[2]**. My tee features a fuse to protect against short circuits on the feeder, a quite important consideration.

The loop itself

I did a lot of experimentation with different sized loops and have concluded the best compromise is a 1m diameter 15mm tube loop. Multi-turn or much bigger loops don't help as the inductance scales up too. One metre diameter is a good size to fit in many places, as suggested by **Photo 2**.

My loop is made from about 3.1m of flexible 15mm Pex-Al-Pex underfloor heating pipe, which has an aluminium core. You need to strip back the outer plastic to reveal the aluminium in the core; I made contacts with braid and hose clamps, with the braid clamped in stainless nuts and bolts. Terminal wires then connected to the PCB terminal block. Pex-Al-Pex is usually available only in long lengths so it might be worth asking a friendly plumber if they have offcuts. Other suitable possibilities include copper pipe, although bending 15mm diameter water pipe would be rather tricky. Smaller diameter, such as 'microbore' (6mm or 8mm diameter) central heating copper pipe is more manageable.

Deployment

I had two prototype antennas on test in my garden and back yard for over a year, along with one in the shed (see Photo 2). It does not need any extra chokes, earthing or balun. My back garden loop is fed with 50m or so of ordinary TV coax. The loss doesn't matter too much as there is enough gain in the amplifier and pick-up on the coax is below the minimum signal level from the loop due to the rejection characteristics. Although the Pex-Al-Pex tube normally only comes in white (and copper in, er, copper-colour), it is straightforward to paint the loop (eg leaf colour) to match its surroundings. It thus becomes practically invisible. (If you're using a copper loop then paint can also seal out oxygen and moisture, preventing oxidisation. Whether this will actually make a difference to performance in the long term by reducing skin effect losses at the surface is a matter for debate).

Results

Results are very promising. HF Signals are perhaps typically 10dB down compared to my 60m long wire but the signal to noise ratio is often 10dB or more better. The great advantage of a small loop is that it can usually be placed in the part of the premises with the lowest QRM – for example, at the bottom of the garden, furthest from noise sources. The receive performance is directly related to the quietness of the location: background noise dominates over the amplifier noise.

Although it is vertically polarised the loop will pick up signals at high angles from the sky, unlike a vertical antenna, so is good for short skip. The null perpendicular to the plane of the loop can help with interference rejection, too.

I use mine a lot for general listening on HF and to monitor WSPR signals on a RSP1A and SDRUno. I have had quite a few transatlantic spots on 630m and 160m, plus worldwide reception on 80m and 40m overnight and on 20 and 10m in the day – right through the sunspot minimum.

Although I reserve all design rights, I am content for people to use these to build their own copies or supply to others at cost in the hope others may enjoy the performance this design provides for such a modest investment.

Websearch

[1] https://www.qsl.net/m0ayf/G8CQX-Loop-antenna.html

[2] https://github.com/john-g8cqx/G8CQX_LoopAmplifier2020

PSU failure protection

Introduction

Most radio amateurs own at least one shack power supply for use with a main station transceiver. This article serves to document the associated over-voltage risks and presents a modern solution to the problem of protecting your radio. Along the way it explains some basic power supply design concepts.

The over-voltage protection board described was produced as a club project for the RSGB-affiliated Kingdom Amateur Radio Society (KARS), who affectionately labelled it 'Bernard's Over-voltage Board'. The name BOB stuck!

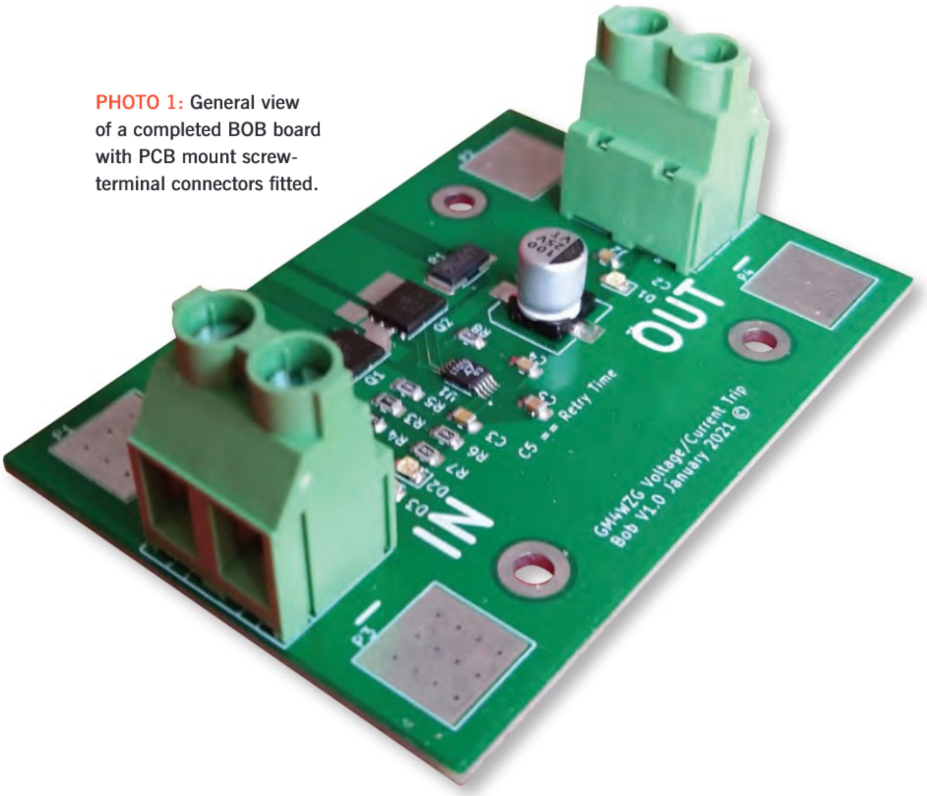

PHOTO 1: General view of a completed BOB board with PCB mount screw-terminal connectors fitted.

Background

Last year my elderly Yaesu FP-700 power supply unit (PSU) failed. This was an expensive failure: it put 24V across a 100W HF radio that was designed for 13.8V. To avoid any such similar circumstance from ruining another day, a crowbar **[1]** circuit was duly designed and implemented. This comprises a thyristor triggered by a Zener diode that puts a dead short across the power supply if the voltage exceeds a given figure.

But to apply a crowbar short circuit to a PSU is a fairly brutal act and it relies on the integrity of an old-fashioned fuse for its success. Traditional fuses are notorious for their slow and unpredictable action, taking anything from a few seconds to several minutes to blow when subjected to twice their rated current. Even at four times rated current, a delay of half a second is generally regarded as pretty fast.

Bearing in mind semiconductors can be destroyed in a few milliseconds, a better way was sought. I came across a modern, purpose-designed power supply protection integrated circuit, the LT4368. It protects against overcurrent, overvoltage, undervoltage and even reverse polarity. More on that later, after a look at the various types of power supply you're likely to encounter in and around the shack.

Safety first

All power supplies can be dangerous to work on and unless you know precisely what you are doing, I strongly recommend you do not ever open one up. Switching supplies are particularly hazardous because most of the circuitry is not isolated from the mains. Even when disconnected, capacitors may be charged to well above mains voltage and contain sufficient stored energy to kill or seriously injure. The protection circuit described later in this article does not require any modifications to a PSU so there should be no need to meddle.

Power supply types

Linear PSUs are preferred by many radio amateurs over switching PSUs because of their negligible RF noise output. **Figure 1** shows a the basic elements of a typical 13.8V linear PSU. AC mains is stepped down to about 18V AC by the transformer T1, rectified by the bridge rectifier BR1 then smoothed by C1, at which point the

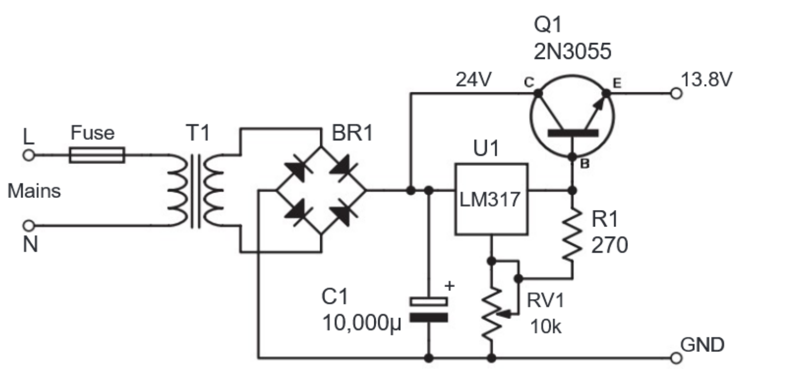

FIGURE 1: Simplified 13.8V linear PSU. If the transistor fails short circuit your radio's in trouble!

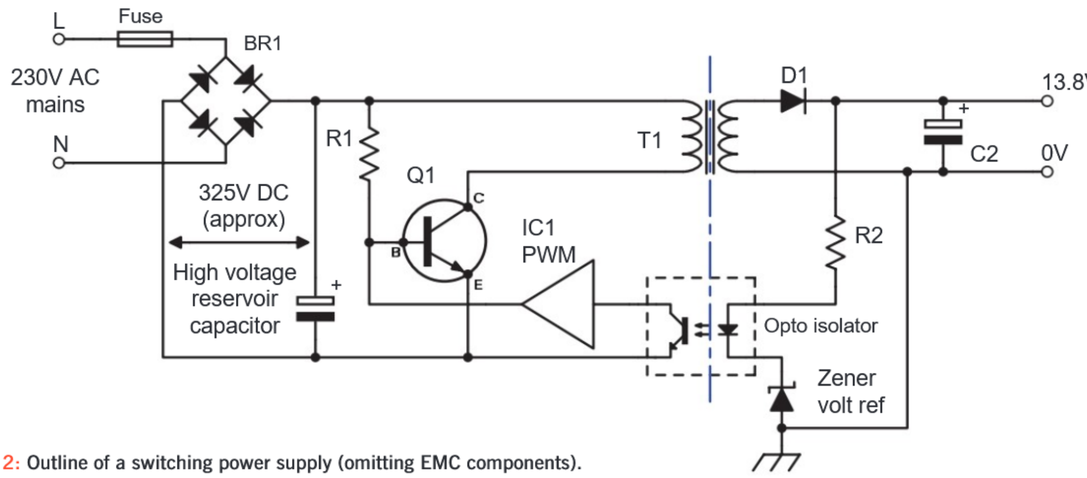

FIGURE 2: Outline of a switching power supply (omitting EMC components).

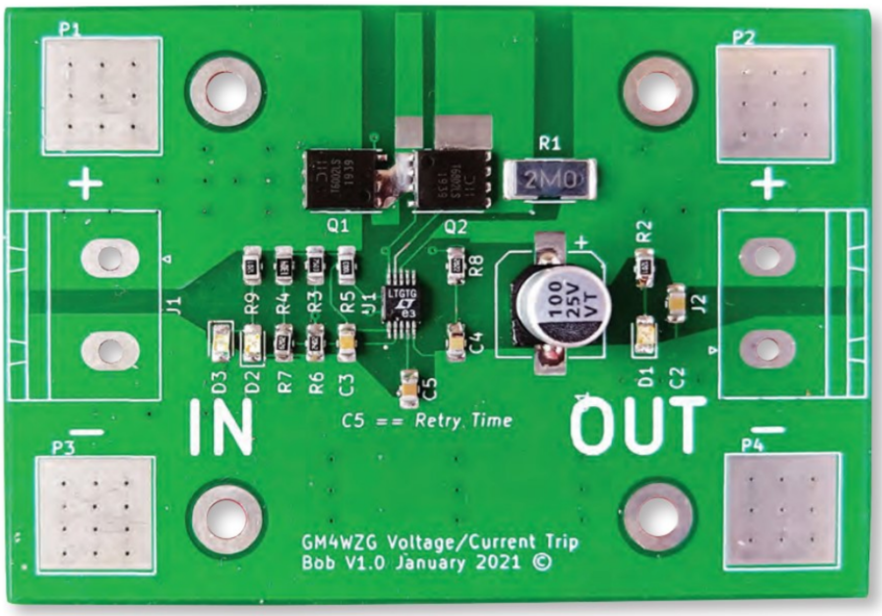

PHOTO 2: The BOB board has clear screen printing to identify component locations.

voltage is 24V DC. This is regulated down to the output voltage of 13.8V by passing through NPN transistor Q1, configured as an emitter follower whose base is maintained at a fixed level by a LM317 regulator IC. But there is a potential problem. When output transistors fail, they tend to fail as a short circuit. Failure of Q1 therefore would result in the application of 24V to the load – with potentially catastrophic consequences.

The other major type of PSUs you're likely to encounter, switching (switch mode) are efficient, smaller and cheaper to manufacture than linear supplies. **Figure 2** shows some of the basic elements present in a typical switching PSU. AC mains is directly rectified by bridge rectifier BR1 and smoothed by C1. IC1 is a pulse width modulator (PWM) IC that provides pulses at, typically, 100kHz to Q1, which turns on and off according to these pulses, thus providing high energy pulses to the primary of T1. (In practice Q1 is more usually a MOSFET but bipolar transistors can also be used; the distinction is not important for this description).

The pulses give rise to an AC output voltage across the secondary winding of T1 that is subsequently rectified by D1 and smoothed by C2. Owing to the high frequency of the pulse generator, the transformer and the smoothing capacitors can be a lot smaller than their cousins in the linear power supply.

The output voltage is fed back to IC1 via an opto isolator. If the DC output voltage is too high then IC1 reduces the pulse width, thus reducing the output and if the output voltage is too low it increases the pulse width thus increasing the output. In other words it functions as a fully regulated supply. The opto isolator and the windings of the transformer together provide the necessary safety isolation (dashed blue line) between the mains and the low voltage DC side.

If Q1 fails short circuit then constant high voltage DC is applied to the primary of T1. But, because there is no pulse action in T1, there can be no voltage on the secondary side and there is no risk of damage to equipment. Hopefully the fuse will blow, protecting BR1 and T1 from consequential damage.

But beware! If the feedback from the opto isolator fails to represent the true output properly, or IC1 fails in some way (perhaps as a result of RFI from a local high power transmitter), the DC voltage at the output could increase to unwanted levels.

Note that this circuit and description omits some essential components that perform functions such as interference suppression – a switch mode PSU built with just the components shown in Figure 2 would be *very* unwelcome in the shack.

Whatever the kind of power supply we have, it is a good idea to implement safe and independent over-voltage protection. This is especially pertinent to the lower cost units on the market.

We need to examine closely any manufacturer claim concerning the incorporation of over-voltage protection in their products. Many switching supplies rely on a very simple circuit comprising a small metal oxide variable resistor (MOV) and a fuse, then claim they are "fully overvoltage protected". This is not enough to give peace of mind to the owner of a £5,000 amateur

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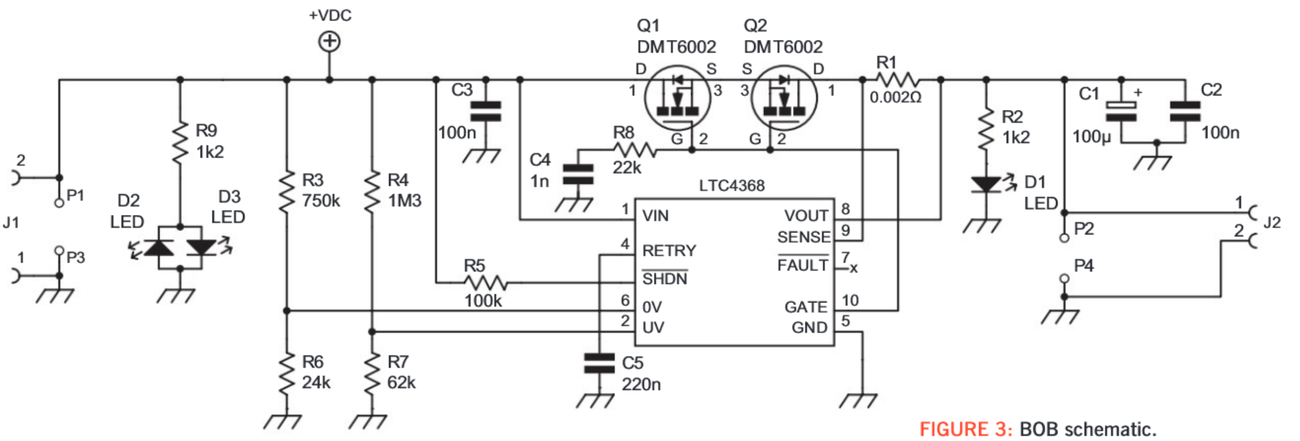

FIGURE 3: BOB schematic.

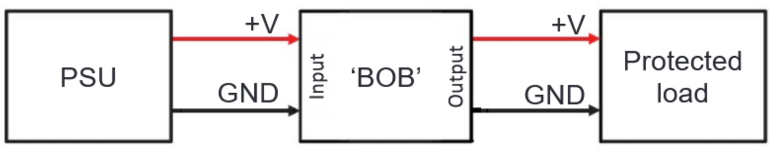

FIGURE 4: Using BOB is very simple.

radio transceiver, which might well be damaged in far less time than it takes for a fuse to blow.

Modern methods

Nowadays there seems to be a chip for everything, particularly in the area of power supply design. The Analog Devices LT43681 is described as a "100V Under-voltage/Over-voltage and Reverse Protection Controller with Bidirectional Circuit Breaker". The LT4368 will always trip in less than 8μs from when its set-point is exceeded, even if by only the smallest of amounts. This closedown is fast enough to prevent damage to almost any load.

The manufacturer's data sheet **[2]** of course provides full details and example schematics. For the design of BOB I more or less followed manufacturer's recommendations, with the exception of a small change to make the design calculations easier – and the addition of more RF decoupling. The full circuit diagram is shown in **Figure 3**.

In the circuit, external power at 13.8V is applied at J1. D2 and D3 are LEDs connected back to back. When power is connected properly, green LED D3 illuminates. If the unit is accidentally connected to power rails in reverse, red LED D2 will illuminate, indicating the fault condition. The LEDs are very clearly visible at 13.8V and are not damaged by input voltages up to 40V.

Power is switched to the DC output terminal J2 by N-channel Power MOSFETs Q1 and Q2. There is always an internal diode within a MOSFET and even if switched off, the MOSFET will conduct in one direction through this diode. Using two MOSFETs as shown puts these diodes back to back so there is never a conducting path unless the MOSFETs are deliberately switched on.

MOSFETs have a rating called Rds(on). This is the resistance across drain and source and, for power devices, is pretty low. At full saturation this typically causes a voltage drop across the two MOSFETs of less than 100mV – more or less the same as that of a typical fuse. BOB should therefore have no practical impact on the 13.8V supplied to the rig, even with 25A current draw. Also, the low voltage drop means that very little power is dissipated in the MOSFETs – essentially removing any requirement for heatsinking.

The output current passes through R1, a very low value resistor used to sense the current. Its value is chosen such that at maximum permissible current draw (25A), 50mV will be developed across it. This triggers the LTC4268 to switch off the drive to the MOSFETs and hence remove power from the load. As noted earlier, the LTC4268 switches in less than 8μs.

Two potential dividers are placed across the input. The junction of R3 and R6 feeds the over voltage comparator of the LTC4368 and the junction of R4 and R7 feeds the under voltage comparator. If the junction of R3 and R6 exceeds 0.5V the LTC4368 will trigger an over voltage condition. If the junction of R4 and R7 drops to below 0.5V the LTC4368 will trigger an under voltage condition. With the resistor values given BOB will switch off at below 11.5V or above 16V.

C2 and C3 are decoupling capacitors. The unit will be used in high RF environment so this pair of low ESR MLCC capacitors is deemed prudent.

PCB and construction

A simple PCB was designed, which can be seen in the photos. No kits are available but I do have a small supply of PCBs that I'll be happy to supply at cost **[3]**. The small PCB uses SMD components, none of which are considered to be 'difficult' to solder. At least one club member successfully constructed a board as his first foray into SMD work, so it does seem suitable for the less-experienced.

Testing and use

The BOB should be connected as shown in **Figure 4**.

To date, many BOBs have been produced and each one was tested by being subjected to overvoltage and to a dead short on its output. Interestingly, the BOB is so fast that if you short a 25A-capable output with a 100mA fuse, the power is shut off before the fuse has a chance to blow. A similar test has been carried out by using a low power 2N3904 transistor to short the output. The transistor always survives. An impressive demonstration of performance indeed.

Websearch

[1] https://tinyurl.com/sk8r9kw

[2] https://www.analog.com/en/products/ltc4368.html

[3] See https://gm4wzg.co.uk/bob

QRM finding & elimination

Area power outage

How often do you have a power cut? Hopefully, for most of us, not very often. How often do you have a pre-planned area power outage? These things are rare, but I had the good fortune to experience one recently. A power pole near to where I live had broken and needed replacing (**Photo 1**) and my local area was to experience a 9-5 power cut on a Saturday. Not wishing to miss such an opportunity, I set up my Yaesu FT-991A to run off a 12V leisure battery and also ensured my laptop and iPhone were fully charged. I wanted to measure and record the before-and-after noise levels I was receiving.

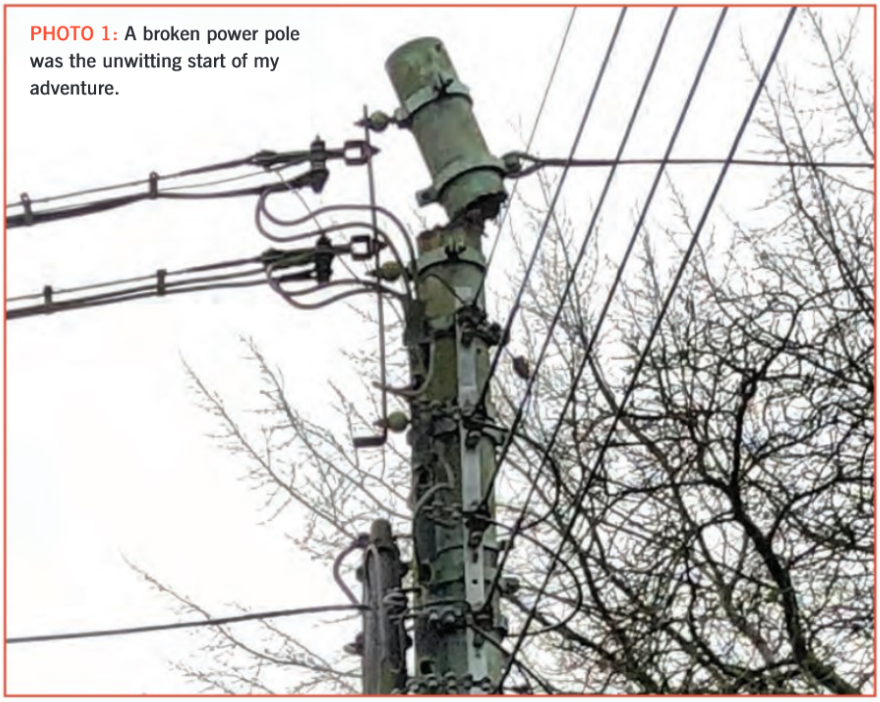

PHOTO 1: A broken power pole was the unwitting start of my adventure.

Before & after noise levels

I made video recordings and measurements of noise across the HF bands early on the morning of the planned cut. There was typically S5 to S7 on the lower bands and S3 on the higher ones. In particular, there was a distinctive interference pattern across most bands in varying strength. It was around 10kHz wide and repeated every 70kHz, present 24 hours a day and seemed to reset itself every 60 seconds. Making these 'before' recordings and measurements for each band in turn only served to remind me how noisy some bands were.

Then, bang on 9am the power went off… cue protests from all the house alarms in the neighbourhood!

The 'after' measurements when the power went off were truly amazing. My noise floor dropped away to effectively nothing and I even heard New Zealand for the first time from home! What also caught my attention was that the QRM I was receiving disappeared as well. As the power cut was only within my immediate area, say 100 houses, I now knew this QRM was local to me. I'd somehow got used to its presence and worked around it where I could. But knowing it was local spurred me on to do something about it. Could I actually track this thing down and eliminate it for good?

Gaining local support

First impressions count, so my first step was to write and print off a light-hearted note to my immediate neighbours. This was to save me door knocking and repeating the same story over and over.

We'd all just experienced the same power cut, so the timing was right. My note briefly explained my hobby, increasing noise levels and what I had experienced. I went on to explain the potential sources and it may be as simple as a faulty phone charger. I said clearly that I'd be happy to replace whatever was causing the interference, when it was identified.

I asked for their help in that I may want them to power down their houses to help me trace the source. I even included links to two videos I'd posted onto YouTube, one with the noise and one with the New Zealand station.

I keep my amateur activities discreet, so some neighbours won't have known about my hobby. My introductory note seemed to do the trick. Next time I bumped into my neighbours (socially distanced!) the topic came up: had I found it yet? When do I want them to turn the power off? Didn't it sound horrible on the video…

All was looking promising, but I explained I wanted to try and find it myself first. This would be an interesting challenge for me, and it would save them resetting all the timers in the house again (which, to be fair, can be a pain).

Simple DF loop design

The obligatory internet research on direction finding drew my attention to a design for a simple non-resonant loop **[1]**. This looked easy to make and well within my competence levels – assuming it worked! It is based on a 900mm length of RG213, as shown in **Figure 1**. The braid is removed for about 20mm from the centre of the cable. One end is terminated in a PL259 plug in the normal way; on the other end you only connect the braid to the plug, making sure the centre conductor can't short to anything. These are then plugged into a SO239 T-adapter; the third socket provides the output to your patch lead to the receiver. **Photo 2** shows my prototype, which I attached to a length of 20mm plastic conduit by cable ties. A little insulation tape over the gap in the braid will keep rain out, at least for the length of time you'll want to use it while standing *in* the rain.

Direction finding in practice

Not having a handy short wave radio or a mobile transceiver meant I had a side-project of mobilising my FT-991A into a bag, complete with a 12V battery large enough to give me about

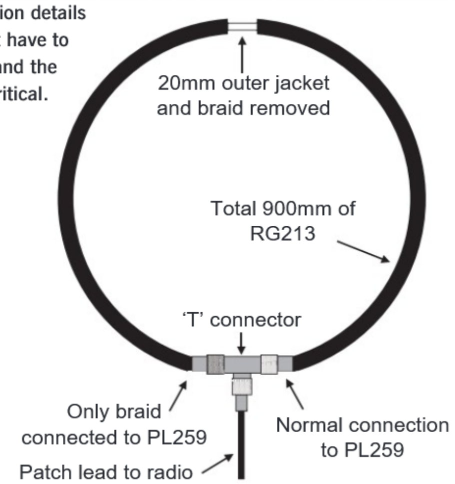

FIGURE 1: Construction details of the loop. It doesn't have to be perfectly circular and the dimensions are not critical.

an hour of receive time. With the radio strapped to my front and the loop in hand (**Photo 3**), I wasn't sure if I was more Ghostbuster or Dad's Army's Private Godfrey.

With the loop connected and the radio powered up, it was immediately obvious that the loop worked. There was a strong null face-on to the loop, with the interference clearly audible when the loop was side-on. I then proceeded to walk around the local area, making a mental note of whether the QRM was stronger or weaker. I found it easier using the sound of the QRM to track rather than using the silent null. It is only when you move towards or away from the source of the signal that you know which side of the loop is receiving strongest – it could be in front or behind you. In this way you can then establish the direction of the source.

My wandering around the locality, loop in hand, caused much intrigue to passing locals. No, I wasn't metal detecting or looking for bats, I explained, and there was genuine interest and wishes of Good Luck. One chap I spoke to even knew of RAYNET!

The results of my triangulation can be seen in **Photo 4** (the red line indicates my dipole). It suggested a property outside of my immediate estate, so this was someone I hadn't delivered my leaflet to.

At this point my battery died. Without my live demonstration, I approached the property owner and explained my predicament – again, in a friendly way, playing down the potential cause as it may be one of many things, but potentially something small and easy to fix. I explained if they didn't mind, I would replace whatever it was free of charge – if we could track it down. This first encounter went well and we agreed to arrange a future date to try and trace the source.

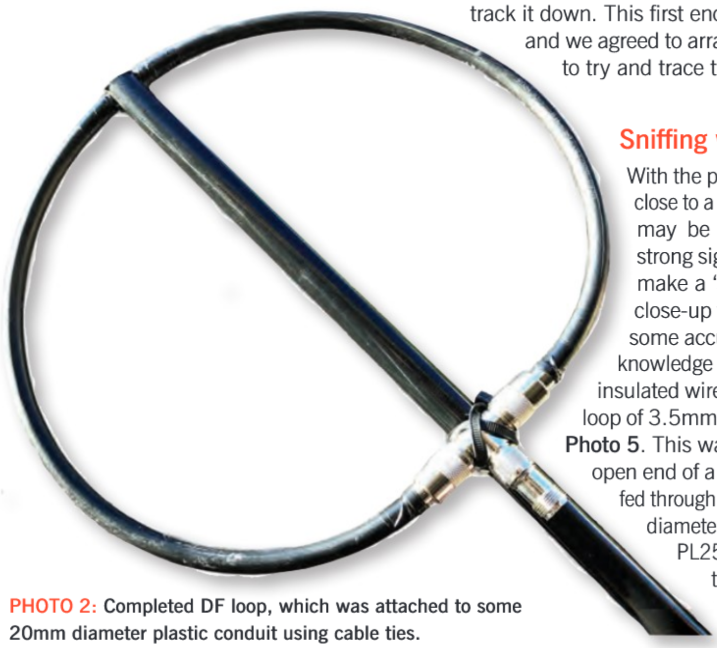

PHOTO 2: Completed DF loop, which was attached to some 20mm diameter plastic conduit using cable ties.

Sniffing wand

With the potential that when close to a source the receiver may be overwhelmed by strong signals, I decided to make a 'sniffing wand' for close-up work. Armed with some accumulated internet knowledge I used some 1mm insulated wire to make a 3-turn loop of 3.5mm diameter, seen in **Photo 5**. This was attached to the open end of a piece of coax and fed through some more 20mm diameter plastic conduit. A PL259 was soldered on the end of the cable so it could plug into the radio.

The idea of the wand is to hold it up close to a suspect source to see if it is emitting radio noise. Due to social distancing restrictions I wouldn't be able to enter the property, but at least the owner could do this detective work on my behalf and the wand may allow tests to take place without powering down the whole house (or by switching off breakers one at a time).

Closing in

I met the homeowner at the appointed time and explained our options on how to approach the exercise. It was a large house with various outbuildings. Consequently there were numerous consumer units and associated breakers. We agreed to power off one consumer unit and go from there. Luck was on our side and as soon as the power went off, the QRM stopped! Working through the breakers on that consumer unit narrowed it down to the one labelled 'outside & garage'. As I was outside I used the loop to see if I could detect a strong signal source. Strangely, as I walked in front of a set of garages the signal wasn't particularly strong. The owner then mentioned that there was another single garage on the other side of the house. Turning the loop back in that direction did show a signal and walking around to the other side of the house confirmed this. The QRM appeared to be emanating from this other garage.

PHOTO 3: Let the hunt commence.

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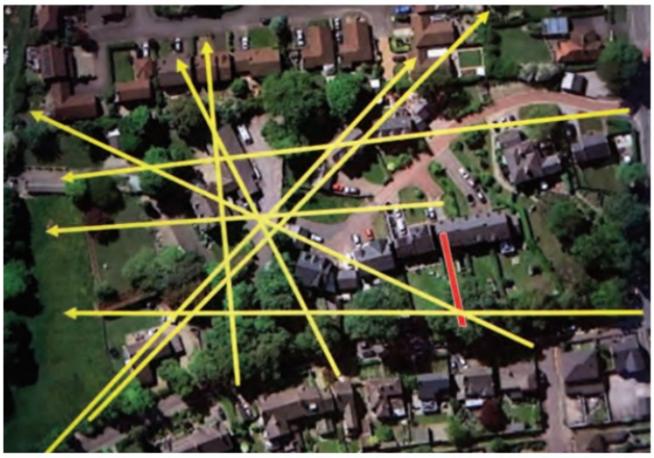

PHOTO 4: **My DF bearings suggested an area of interest.**

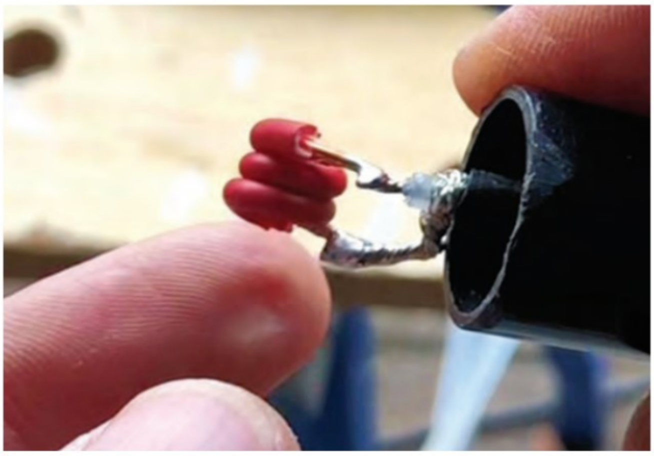

PHOTO 5: **The business end of the 'sniffing wand'.**

PHOTO 6: **This battery charger turned out to be the source of the QRM.**

Source of my QRM

Garage door open, it became apparent that there was only one thing plugged in – a charger trickle charging a lawnmower battery. The noise on the radio was horrific. Not wishing to miss an opportunity, I attached the sniffing wand for its inaugural outing. It worked! Not only was the charger producing radio interference, so was the extension lead it was plugged into, the double wall socket and the dedicated garage consumer unit – and all the connecting wiring. This may have even gone right back the house, who knows.

Turning the charger off was like night and day: all noise gone in an instant. To double-check, I returned home to the shack and we did a co-ordinated off-on test while I listened on the main station radio. Sure enough, the noise disappeared across all bands. I even think the noise floor in between the 70kHz spikes may have also reduced. The charger identified as the noise source is a (CE marked) 'FOXSUR FBC1205D 12V 4Ah-100Ah Pulse Repair Charger' (**Photo 6**). It is readily available online. It is manufactured in China and has Cyrillic as well as English labelling.

The offending charger was about 130m from my home. It is reasonable to assume that *any* radio user within about 150m of the charger would suffer comparable interference – and a quick piece of maths suggests this means about 55 acres being affected to a similar level of interference that I'd suffered. A sobering thought.

Happy ending

The homeowner was fully cooperative throughout and I agreed to buy him a new charger (and there was also a surprise bottle of nice wine for him too). He let me take the offending item away with me, which I've since passed on to the RSGB EMC Committee to help with their work on interference. I've also raised the issue with Ofcom.

As a matter of courtesy, I have sent another note around my immediate neighbours explaining the successful outcome and thanking them for offers of help. As was only fair, the note did not identify the charger-owning homeowner.

Perhaps I was lucky, but getting neighbours onside early I think helped a lot and also did a little bit to promote our hobby. It also helped raise a wider awareness to radio noise. Not laying blame or questioning their electronic purchases was key. Explaining that streetlights, or the power lines themselves could be a source of interference widened the net and again avoided blame or finger pointing.

Even the general public took and interest while I was direction finding. Some followed me down the main road to hear the signal getting stronger.

I also should have done this much sooner. I'd forgotten how long I had worked around this QRM. It turns out to have been about six months. If I'd only realised sooner that I had the ability to knock up some very simple detecting equipment and track it down I would have done it months ago. I just assumed it was out there somewhere far away that I couldn't influence and was something I would have to live with.

My noise floor will never be the near-zero I experienced during the power outage, but it's now largely QRM free, not messy and perhaps a little less noisy overall. Spikes on the band display are now actual stations to tune in on, not just simply more QRM. If you have a QRM problem it may be worth taking a leaf out of my book and doing something similar.

Websearch

[1] https://qrm.guru/how-to-locate-the-noise-source-building-df-loop/

Videos related to this article:

Area power off – https://www.youtube.com/watch?v=R52V7Nf8uBM

Loop build & QRM locating – https://www.youtube.com/watch?v=56IUtdBK-7U

Wand build & QRM removal – https://www.youtube.com/watch?v=Pwbk8yP6SIk

Design Notes

Transmitter output impedance

What's the output impedance of your transmitter? 50Ω? Well, that's what it says in the manual and on that label. It's a matched system – and the maximum power transfer theorem says that maximum power is transferred when the load impedance equals the source impedance. Doesn't it?

NO (and a qualified NO). We'll come back to Z_{OUT} shortly, but the maximum power transfer statement is often misunderstood. Look at **Figure 1**, which shows a voltage source (like a battery) in series with an internal resistance, R_{SOURCE} feeding into R_{LOAD}. If we define R_{SOURCE} to be fixed then, sure enough, we get maximum power transfer when R_{LOAD} equals this, with exactly half the source voltage appearing across the load. Half the power is delivered; the other half lost in the internal source resistance. As a source of power, the box is 50% efficient.

But consider now if instead R_{LOAD} is fixed and we make the *source* resistance variable. This is usually the case with a transmitter – our load is 50Ω by definition. Maximum power transfer now occurs when R is zero and we have all the voltage from the source across the load; any increase in R_{SOURCE} just loses volts and power. The transmitter is 100% efficient when R_{SOURCE} = zero and the maximum power transfer theorem only applies when it is the load impedance that is allowed to vary. For decent efficiency we definitely don't want to match our transmitter to the source impedance. What the actual output impedance of a real transmitter may be is a bit of a mystery, so we'll now look at what it might be and see how a transmitter's actual output impedance can be almost anything, can change when you drive it harder, add a filter, or a different matching network, or an ATU.

Take the simplest linear output stage possible, a single transistor operated in Class A and matched to a 50Ω load using a transformer, as shown in **Figure 2**. We'll make the transformer a 2:1 turns ratio so it transforms the 50Ω load impedance to 200Ω seen at the collector. Vary the drive to the base of the transistor and consider what happens. Provided it doesn't saturate or cut off, the collector current varies linearly with drive. Let's say we get 100mA RMS current variation, which is about 280mA peak to peak. This current (the varying AC part) passes through the 200Ω load resistor and develops an RMS voltage across it of 20V, delivering a power of I^2R = 2 watts. Now change the load, make that 50Ω load become 25Ω and see what happens. The transformed load resistance at the collector is now 100Ω, but the RMS current through it is still 100mA, so the PA stage now delivers only 1 watt. What has happened? We've reduced the load resistance and the power output has fallen. This can be explained by the constant current nature of a transistor operated in its linear region. The *current* delivered is 'forced' by the drive, while the *voltage* on the collector is allowed to be whatever it wants to be, depending on the load. This fixed current with arbitrary collector voltage means the device is functioning as a current source; when current doesn't vary much as the voltage changes, the impedance must be high. So, the output impedance of this simple transmitter is high, not quite infinite due to imperfections in the device, but way-way more than the load resistance presented to it.

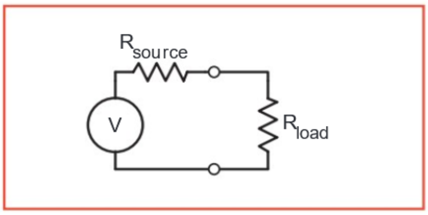

FIGURE 1: Source and load resistance illustrating the maximum power transfer theorem. If R_{LOAD} is considered to be the variable one, then maximum power transfer occurs when $R_{LOAD} = R_{SOURCE}$. If R_{SOURCE} is the one that can change, the same rule does not apply.

Note that in this example the supply voltage doesn't enter into the calculation, it is just the varying collector *current* that generates the output power as it passes through the transformed load resistance. However, the current has to come from somewhere, and that is the power supply. The voltage that is dropped across the resistor must peak at a voltage below that of the supply. If it tries to go beyond this, the transistor tries to drive more current than the supply and load resistor can support, the output limits, or clips, or saturates – or, in the other direction, cuts off. The RMS voltage of the fundamental RF carrier allowed by this peak value clipped to the supply is therefore $V_s / \sqrt{2}$.

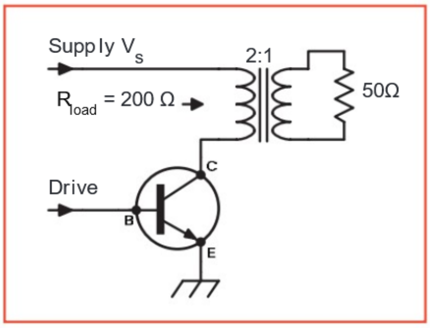

FIGURE 2: Basic Class A PA stage that can be operated either linearly or clipped / saturated. The resulting output impedance changes dramatically at the transition from linear to clipped operation.

Maximum power that can be delivered to the load is therefore $(V_s / \sqrt{2})^2 / R_{LOAD}$ or $V_s^2 / (2 \times R_{LOAD})$. This is the classic equation seen when designing the output matching for all high power amplifiers.

Class C/D

Now let's keep the same circuitry, but formally change the way the PA stage operates. We'll drive it hard so it firmly switches on and off, delivering power to the load from a supply voltage that is fixed at V_S = 30V. We have a Class D switching amplifier. When the transistor is switched hard on, 30V appears across the 200Ω load and, for that half cycle, delivers $30^2 / 200$ = 4.5 watts.

When the device is switched off, we will assume the transformer behaves as an inductor, tries to keep the current flowing so the voltage on the open-circuited collector flies up to 60V, imposing 30V across R_{LOAD} in the opposite direction and maintaining the 4.5 watts. This is 4.5 watts of square wave so there are 3rd, 5th, 7th etc harmonics that all contain some of that 4.5 watts (and that we need to filter out later), but let's not worry about that. What is the transmitter output impedance now? We no longer have the case of a varying current being forced through the load, but we have a fixed voltage of ±30V being switched across it. If we replace our output by that 25Ω resistor, transformed to R_{LOAD} = 100Ω, we get a power out of

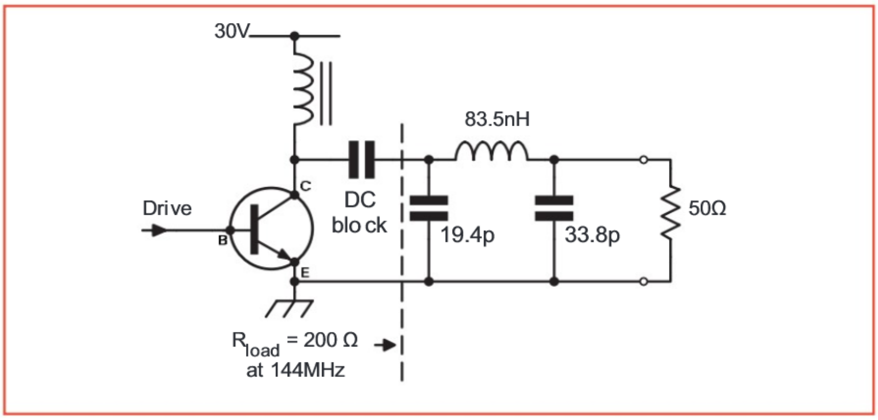

FIGURE 3: 144MHz PA stage using a Pi matching network. Values shown transform a 50Ω load to 200Ω seen by the transistor at this frequency. When the load isn't 50Ω, the impedance seen by the device changes in a surprising way, as discussed in the text.

$30^2/100\Omega = 9$ watts. A switched voltage source has a low (theoretically zero) source impedance.

So look what has happened. The same transmitter output stage, when operated in its linear region, acts as a variable current source and exhibits a high output impedance. As it runs into clipping, the output impedance becomes low. In a practical output device, where the transition from linear to saturation has a region of change and a bit of real output loss resistance, what on earth is the actual output impedance? And how can it even be defined when it is linearity and drive dependent? There lies the problem. The output impedance of a simple transmitter can be over the place depending on how hard you drive it.

Valves, especially triodes, do not even behave as true current sources; there is a definite V_A versus I_A characteristic, called the anode impedance. To a lesser extent this is also seen in MOSFETs and bipolars. For high efficiency PA stages, this anode (or drain or collector) resistance needs to be a lot higher than the transformed R_{LOAD}.

The simple PA design used here could just as well be a push-pull output stage operated in Class AB. There is still a linear relationship between drive and collector or drain current except that it is shared between two devices, alternately, depending on which half-cycle of the RF is underway. So the same rules about its high output impedance in linear operation and low impedance when saturated still applies.

Feedback

In wideband PA stages for HF and VHF, feedback is often applied across the output devices. This can be a simple as a resistor with DC blocking capacitor from drain to gate of a MOSFET, or from an extra feedback winding on the output transformer in a bipolar design. Now, during the linear operating region any attempt to reduce the voltage on the drain by reducing the load resistance will cause the negative feedback to try to raise it by increasing drive. We now have a causal relationship between load current and voltage introduced by the addition of feedback. This lowers the output impedance from 'high' to some other, probably just 'high-ish' value.

If we were to add a *lot* of negative feedback we might, with a great deal of design effort, get the internal impedance down to a value that transformed via the matching to 50Ω and we would indeed have a transmitter with 50Ω output. But it wouldn't be very efficient and is usually not what we want.

As a final thought to this section, consider an emitter follower amplifier. These always offer a low output impedance. Why? Because the common collector / emitter follower configuration has 100% negative feedback giving near unity voltage gain, hence the low impedance.

Matching network and output filters

The fun really starts when we don't employ wideband transformer matching but use something like a Pi network, as shown in **Figure 3**. The values there are designed to match the same impedances as the example earlier, 50Ω transformed to a 200Ω load seen at the device collector, with C and L values chosen for operation at 144MHz.

Using the NETCALC software **[1]** we can (rather tediously) work backwards, substituting various values other than 50Ω at the output load and see what the resulting R_{LOAD} is. After doing this with four values of load resistor I got bored and wrote a spreadsheet to do this over a range of output resistances. (SPICE or some other circuit simulation package would work just as well). The results are illuminating. Changing the output load in either direction above or below 50Ω generally decreases the resistive part of the load seen by the device and also introduces a reactive term. **Figure 4** shows the complex load impedance seen at the collector resulting from various values of load resistance. The blue line is the real part of the impedance, the red one the reactive part and the orange line is the modulus of the transformed impedance (which doesn't depart too far from 200Ω). This is for a Pi network. Other matching networks such as a Tee will behave differently.

Most HF and VHF transmitters nowadays don't use Pi or similar matching networks, but what about that output filter? It's a multi element Pi network and when a non-matched load appears, this will be transformed to a totally unknown and partly reactive load impedance at the transmitter devices – and one that almost certainly varies wildly with frequency, especially as we get near to the cut-off frequency. So although we might think that operating our transmitter into an open circuit 'ought to be 'safer than into a short circuit, if there are low pass filters, or L/C or transmission line matching circuitry in place – think again.

Mismatch protection

This uncertainty of what load appears when our complete transmitter system with all its matching and filtering runs into a mismatch makes protection difficult. We want to stop the devices from being asked to deliver too high a current – which happens if the transformed load is too low. The possibility of excessive voltage appearing across them is a risk but less likely to cause damage with clipping to the supply rails, but if the load appears reactive, resonances could just possibly occur that give unwanted high voltages.

In most transceivers up to UHF manufacturers usually incorporate an output VSWR bridge. Measurements from this can be presented to the user to show how good is the output match and can be used as part of an auto ATU tuning algorithm (see last month). It is also a good automatic warning. By measuring the output power and return loss/VSWR, a decision can be made as to whether at a particular threshold this might result in damaging currents or voltages on the output devices. In which case the drive can be reduced to keep within safe limits. But note that word 'might'. A high return loss

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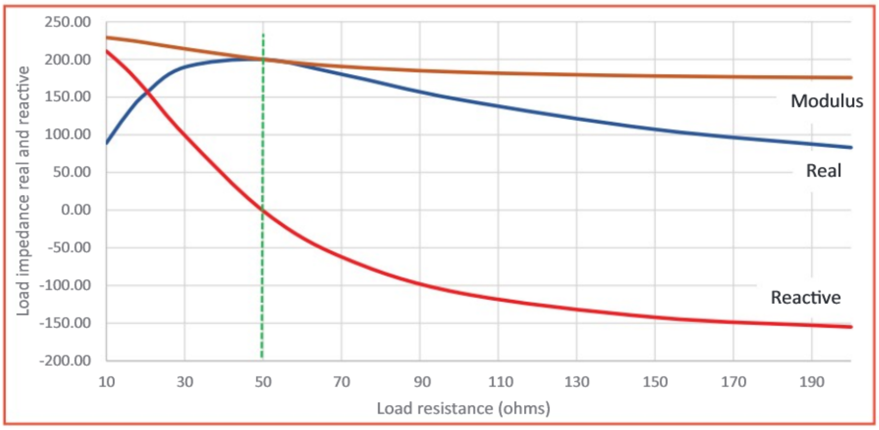

FIGURE 4: Complex load impedance seen by the PA device in Figure 3 as the output load is changed. The real part is shown in blue and the reactive part of the resulting load impedance in red with the modulus of the two in orange. The dotted line shows the condition for 50Ω output load.

may just as easily end up presenting a low PA current situation and the drive 'could' have been safely increased instead. Using output VSWR is not an ideal way to provide PA mismatch protection; it's just *simple* to do. Ideally, protection circuitry would monitor the PA current and the RF voltage appearing on the device terminals. It would then know the nature of the mismatch.

Coplanar waveguide

After reading the woes of my first 100dB attenuator described in the July Design Notes, David Bolton, GW8UQC wrote in to say, "I just wonder if what you saw with your attenuator might be as follows. What you have is not microstrip, it's co-planar waveguide (CPW). I used TXLINE – it's part of *Microwave Office* – but is available free online. The gap needs to be in the of order 0.275mm to give 50Ω with a line width of 1.5mm – I got that as being 'a big bigger than the BNC pin'.

"The impedance of the CPW behaves roughly as shown in **Table 1**.

"It's important to note that the fields from the centre line are not TEM, they are sort-of confined at high frequencies, but at low frequencies they are only poorly confined. What I think happens is that although the case is at a nominal ground at both ends (and the PCB is at close to the same RF potential), the sides of the box and the PCB ground plane are *not* at the same RF potential. Then what you see is a change in the coupling between grounds with frequency, which effectively 'shorts' the whole attenuator structure; exactly how it changes depends on the nature of the metallic box and its filling.

"One might expect this to reduce the attenuation monotonically, but at some point the fields from the line will become better confined by the CPW structure, coupling between box and ground plane become better defined and the attenuator becomes 'well behaved'.

"Then your wire between the box and GP are forcing the two grounds to the same potential, assuming that the wire impedance is small at these frequencies, the coupling is reduced. Had you extended the PCB to the box walls and soldered the PCB to the box you might not have observed the effect.

"Similarly, if through-hole wire links are placed at the edge of the GP next to the through line (about every mm should do it!) each side of the through line, the result would have been, well, different: I think you'd have still had coupling, which would just have been at a different frequency. The answer might well be found by EM simulation; *Sonnet Lite* might be good enough for the job, but that's a task for long cold winter nights. These are just thoughts that arrived during the small watches of the night.

"I can tell you that, of all the problems I hate, getting things into the box, over a wide bandwidth, is up there with going to the dentist. The problems usually start with a young engineer saying 'so I'll just put in a box, it'll be fine'!

"CPW has a big place in broad band systems. It maintains a better control over impedance over a rather wider frequency range than microstrip, should be lower loss and is slightly less prone to other things round it. The downside is that I'm not sure how to fabricate it on the home lab bench; gaps tend to be small.

"As a late thought, with the resistors stood on edge, the capacitance shunting the resistor is reduced. And if capacitors are stood on edge, they behave rather better. The company ATC **[2]** (who make splendid – and frighteningly expensive – microwave capacitors) has a good note on this topic."

TABLE 1: Impedance of co-planar waveguide (CPW) as a function of gap (see text).

Gap	Impedance (Z_0)
2.5mm	85Ω
2.0mm	83Ω
1.5mm	80Ω
1.0mm	74Ω
0.5mm	61Ω
0.25mm	47Ω

Websearch

[1] NETCALC RF network calculator: the original DOS version by GM3SEK and a Windows version by G8AUC can be found at www.sss-mag.com/netcalc2.html

[2] See https://atceramics.com/technical-notes and, more specifically, https://atceramics.com/userFiles/uploads/pdfs/vertical_orient.pdf

Addendum: Hybrid Loop Antennas

The Hybrid Loop Antennas article by David Hayes, M0GDX, published in the July 2021 *RadCom*, was of interest to many people. One thing that could potentially put off some from reproducing the loop was that no diameter was given for the three loading coils, each made from six turns of RG-213 wound on a plastic former, used to expand its coverage to 160m. The diameter is not critical, however M0GDX has confirmed that the former he used was 280mm diameter.

Reference was made to the cardioid loop design by Cross Country Wireless. An error was made in the re-drawing of the design so that the ground connection should go to one side of the loop and not the transformer output. The corrected design can also be found on the Cross Country Wireless website **[1]**, also listed in the original article.

We apologise for any inconvenience caused.

[1] http://www.crosscountrywireless.net/cardioid_loop_antenna.htm

Mystery component detector

The TC1 'Multi-function Tester' is a clever little component tester. It can identify and measure resistors, inductors, capacitors, diodes, Zener diodes, FETS and transistors.

According to the manual (which you can download from [1]), it can also measure IGBTs, thyristors, TRIACs, infrared (IR) remote control signals and even batteries up to 4.5 volts.

These little boxes are not particularly new; the earlier versions usually had monochrome displays. I just stumbled across a YouTube video about them and decided to buy one. They are fairly cheap and I thought that it would be handy in the shack for verifying I am using the correct value of capacitors, identifying transistors etc.

There are dozens of models available as bare circuit boards or already cased in a nice plastic box, often (like this one) with a convenient, integral Li-ion rechargeable battery.

The unboxed versions sell online for around US$7 to $8 and the cased version that I bought was $11.15. I bought it via Amazon, so they automatically charged me freight and GST (the New Zealand equivalent of VAT). All up it cost me NZD 45.36 (£23).

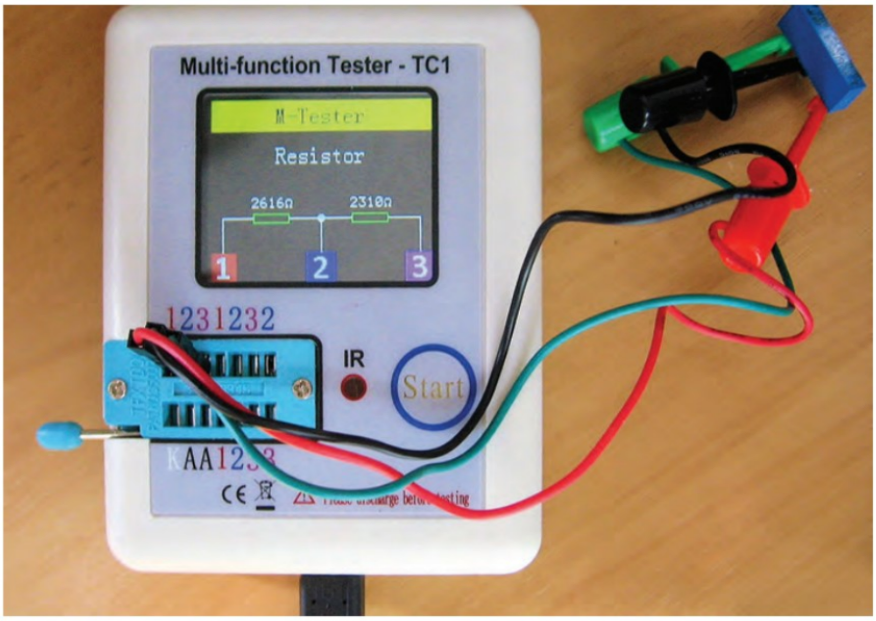

PHOTO 1: The TC1 measuring – and correctly identifying – a 5kΩ 'trimpot' potentiometer.

Accompanying the tester itself were a very short micro USB charging cable, three test leads, some break-off pins and even some sample components – a LED and a small electrolytic capacitor – to test.

The unit has a 160x128 pixel TFT colour display and a ZIF (zero insertion force) socket for connecting the device under test (DUT).

One thing I like is that components can be inserted into the test socket any way around and the unit can still measure them. With my digital multimeter I have to know which pin is the base, emitter, and collector of a transistor, but with this unit, you just connect it to pins 1, 2, and 3 in any order and it will tell you the pin-out of the device. You can even connect back-to-back diodes, two diodes in series, or a potentiometer (as seen in **Photo 1**). In this case the unit will display the resistance of each half of the circuit.

When measuring transistors it identified the connections, NPN or PNP, measured the current gain (h_{FE}), the base to emitter voltage (U_{BE} or V_{BE}) and the collector current (I_C). It will also display if the transistor has a protection diode between the collector

PHOTO 2: Measuring a MPSA63 PNP Darlington pair transistor. The gain (h_{FE}) of 127,000 is real.

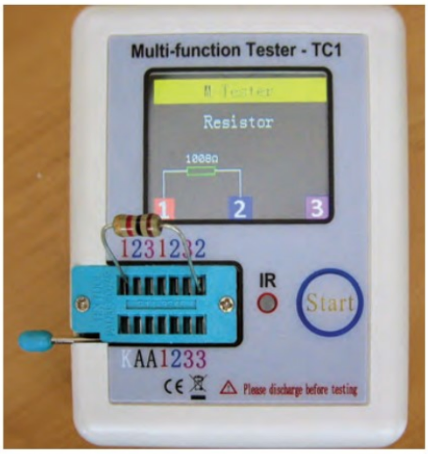

PHOTO 3: A 1kΩ resistor may not be a challenge but confirms you've read the colour code correctly.

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PHOTO 4: How an electrolytic capacitor shapes up, including measuring its ESR.

PHOTO 5: Less detail is displayed for smaller values of capacitor.

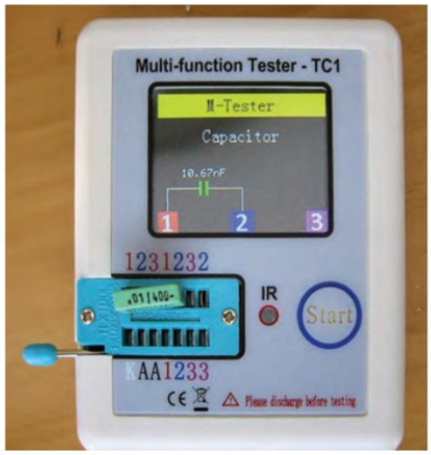

PHOTO 6: The forward voltage and polarity of a LED are identified correctly.

and the emitter. Among others, I tested an MPSA63 PNP Darlington pair and the gain came out as an impressive 127,000 – the transistor's data sheet promises a minimum of 10,000 at 10mA. A BD138 medium power NPN gave an indicated h_{FE} of 170 at 6.2mA, which compares favourably with the data sheet's minimum guaranteed gain of around 85 at that current.

Capacitor measurements include equivalent series resistance (ESR) and the voltage loss (V_{loss}) percentage, which can indicate if a capacitor is drying out or 'leaky'.

I tried out the infrared (IR) test by pointing my TV remote control at the tester and holding down the '1' button. The TC1 displayed the received IR waveform, the 'user code' and the 'data code'. It is not a particularly useful function for me but could be useful for some.

When measuring diodes, it measures the reverse bias capacitance (C), reverse current (I_r), and the forward voltage drop (U_f or V_f).

I did not have much success measuring 1.2V Ni-Cd or 1.5V dry cell batteries. They all read around 745mV when my digital multimeter was reading around 1.3V across the battery at the same time. Maybe it is a calibration issue; I haven't bothered to investigate further. Do not attempt to measure above the 4.5 volt limit. This unit is not able to cope with voltages above 5V.

I did not expect the unit to be particularly accurate. I assumed that it would be useful to make quick checks that the components I am soldering into a circuit are the right value, for example, 1kΩ not 10kΩ, or 100nF not 100pF. But I compared seven capacitance and resistance measurements against my old Expert 97 3½ digit multimeter and my new UNI-T UT890C 4½ digit meter. I used the UNI-T meter as the reference as it should be the most accurate of the three devices. The TC1 was well within 1% agreement with the UNI-T over all the measurements and it had roughly twice the accuracy of the Escort meter. That is remarkably good going for such an inexpensive, versatile piece of test kit.

Websearch

[1] https://www.circuitspecialists.com/content/430516/csi-tc1.pdf

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[1] www.rsgb.org/radcompix
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