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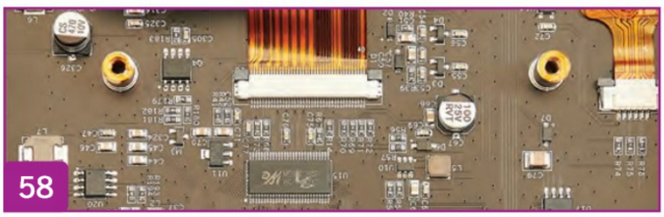

Features

Reviews

Improved Antenna Rotator Controller. Design by Kevin, M6CYB

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

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New Products

G63 UHF Mini Handheld FM Transceiver

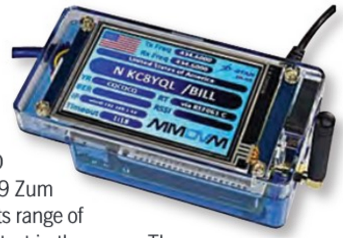
New from MyDEL this month is the G63 sub-miniature handheld UHF radio. If you are looking for a UHF handheld portable radio that you can pack in any bag and take with you anywhere then look no further. The G63 comes pre-programmed and ready to go. If you would like to program it yourself then feel free to do so as the G63 comes with a USB programming cable as standard. This sub-miniature FM handheld radio has all the same great features as the larger units, and with 3W of RF and 300mW of RX audio power, rugged casing, and sturdy controls, you will have no trouble in loud or outdoor environments. The G63 is available for £49.99 from Martin Lynch and Sons via: https://www.hamradio.co.uk/g63

General
Automatic Wireless Copy Frequency
Type-C Charging
High / Low power
Battery Power Saving
Low Battery Warning
Receiver
Sensitivity $<0.16\mu V$(12dB SINAD)
Squelch Sensitivity$<0.2\mu V$
Intermodulation 50dB
Audio Power$\geq$300mW
Audio Distortion:$<5\%$
Receiving current$\leq$300mA
Standby current $\leq$20mA
Transmit
Output power 3W/1W
Modulation mode: FM(F3E)
The maximum deviation $\leq\pm$5KHz
Residual radiation$<$60dB
Current$\leq$1000mA

TINY SA ULTRA

Have you ever thought about testing your own antenna with a spectrum analyser but never made any measurements because a spectrum analyser is such an expensive item? Well, during the festive season you can get testing because now high-specification RF measurements are available to everyone with the TINY SA Ultra. However, this one's in no way tiny with its 4-inch touch screen and impressive specification. The compact SA will measure frequencies between 50kHz and 6GHz. But wait there's more as it's not just a spectrum Analyser, it's also a signal generator up to 800MHz with a level accuracy of +/-2dB and a reference generator with fundamentals at -35.6dBm! Outputs are at: 1MHz, 2MHz, 4MHz, 10MHz, 15MHz or 30MHz. The TINY SA Ultra is available soon from Mirfield Electronics priced at £144.95 with SMA cables and USB charging lead included. More information at: www.mirfield-electronics.co.uk

ZUM Radio Elite 3.5 LCD ZUMspot Kit



Here we have some advanced technology in a small package with the ZUM Radio Elite 3.5 LCD from Zum Radio. Since 2009 Zum Radio has been advancing its range of HotSpot units. This is the latest in the range. The Elite 3.5 LCD ZUMspot is a small and efficient Multi-Mode Digital Hotspot including the following key features:

- High performance 32-bit ARM processor
- Fully assembled and tested in exclusive custom injection moulded case
- Includes a Raspberry Pi 4
- Supports DMR, P-25, D-Star, YSF, NXDN, POCSAG and M17
- Onboard LEDs to show status (Tx, Rx, Status and Mode)
- Up to 10mW RF transmit power
- SMA antenna connector, UHF antenna included
- Includes pre-programmed and pre-configured SD card with Pi-Star
- The open-source firmware (MMDVM) is pre-loaded
- Built-in 3.5" LCD screen
- Includes 3A USB-C power supply
- 1 year warranty

The ZUM Radio Elite 3.5 is available for £279.99 from Martin Lynch and Sons: https://www.hamradio.co.uk

RADIO HISTORY – Roscoe, VE1BC

In this book, author Spurgeon G. Roscoe takes a look at the world of amateur radio in North America and the special role the Halifax Amateur Radio Club (HARC), of which he is a member, played in it from the time of the World Wars all the way to the present day. Roscoe presents an account that is intimate and wide ranging, covering topics from the role of amateur radio on the Canadian ships of the 1920s and 1930s and during wartime, the involvement of women in the craft, the technical details of how some of the equipment has evolved over time and a dedicated look at the American Radio Relay Legue. Both expansive and detailed at once, this book is essential reading for anyone involved or interested in the hobby. The Book is available from Amazon.co.uk

Antennas

Square and Delta Loop Antennas

This month's theme is a comparison between the square and delta forms of the loop antenna. Essentially, this antenna comprises a loop of wire connected across the feeder cable's ends. Both the advantages and disadvantages of each form of the antenna are examined.

Loop antenna overview

A loop antenna is formed using a wire that is one electrical wavelength (λ) long, making it a resonant antenna. The loop's area determines its effectiveness and a circular loop gives the best performance because it has the largest area. However, a circular loop is difficult to construct and the next best practical alternatives are square and triangular shaped loops.

The square loop antenna

Each side of a square loop antenna [1] is electrically λ/4 long and this gives the antenna the advantage that it requires only half the space to install it compared to a λ/2 dipole. Another benefit is a 1λ loop antenna has approximately a 1.35dB gain advantage over a λ/2 dipole. However, the antenna's lowest horizontal side needs to be high enough above the ground to ensure people, pets, plants or structures can not make physical contact with it. As a result, the antenna's upper horizontal side will need to be suspended vertically more than a λ/4 high above the ground and this may be a drawback. The antenna will also need to be installed so that it complies with ICNIRP requirements as appropriate. The concept of a square loop antenna installed at a domestic is shown in **Figure 1**. The feed point for a square loop antenna is in the middle of one of its sides as shown in **Figures 2a** and **2b.** As shown in Figure 2a, if the feed point is situated in the middle of a side that is horizontal to the ground, then the antenna is horizontally polarised and either the lower or upper side could be used. Usually, the feeder cable is connected to the lowest horizontal side as the most practical option. Referring to Figure 2b, feeding the antenna using the midpoint of one of the vertical sides enables the antenna to be vertically polarised. Figure 2a shows the RF currents flowing in a horizontally polarised loop. The RF currents flowing in the horizontal sides are additive, while the RF currents flowing in the vertical sides cancel. This results in the antenna radiating a horizontally polarised RF signal because virtually no RF signal is radiated from the vertical sides.

Figure 2b shows a vertically polarised loop where the RF currents are additive in the vertical sides and cancel in the horizontal sides. This results in the antenna radiating a vertically polarised RF signal. The antenna's feed point impedance is balanced and is about 120Ω. However, the actual feed point impedance is influenced by the antenna's height above the ground. Unlike a dipole, a loop antenna does not tend to be influenced by 'end-effect' and this results in a resonant

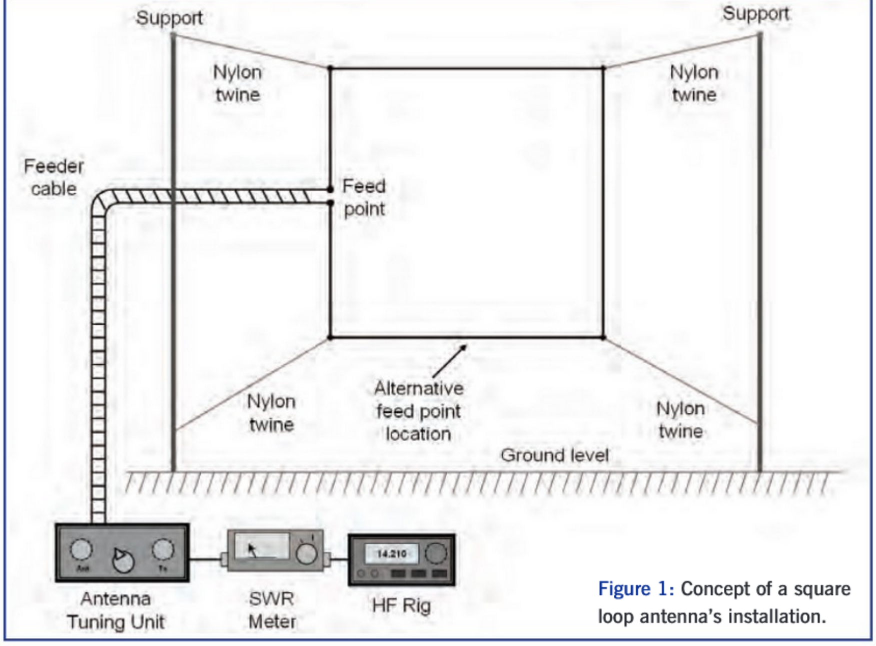


Figure 1: Concept of a square loop antenna's installation.

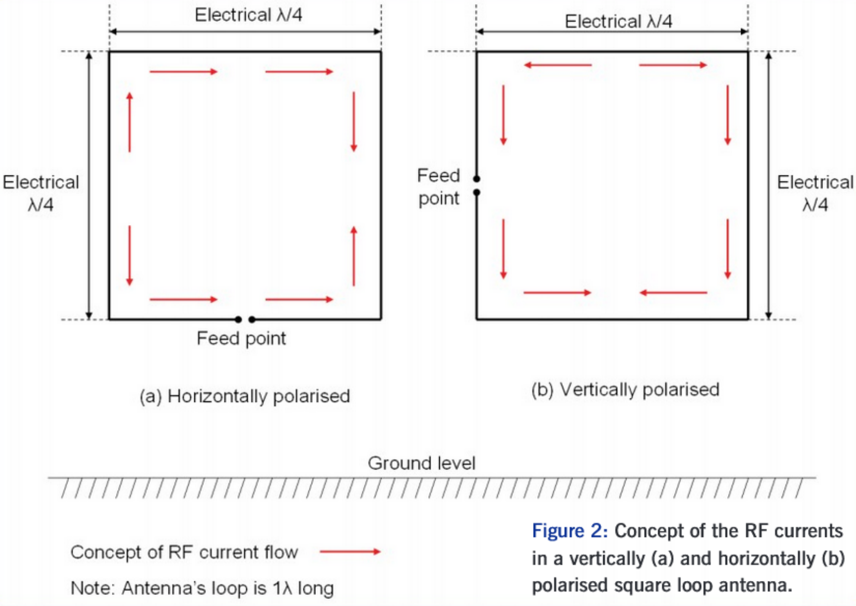


Figure 2: Concept of the RF currents in a vertically (a) and horizontally (b) polarised square loop antenna.

loop antenna being slightly longer than one wavelength by a factor around 1.028 to 1.06 [2].

The delta loop antenna

A delta loop antenna is made from a wire of one electrical wavelength (λ) in length that is formed into an equilateral triangle shaped loop, making a resonant antenna [3]. The triangular shape of the antenna resembles the Greek letter delta (Δ) from where it derives its name. The area of a delta loop will be less than a square loop, however its performance is still similar to a square loop when their wire lengths are 1λ long and both antennas use the same polarisation. However, compared to the square loop, the delta loop has the advantage of requiring only one high support. This feature simplifies the antenna's installation, a benefit for both portable as well as fixed station operation.

Figure 3 shows the concept of a delta loop antenna installed at a domestic location. The feed point for a delta loop antenna is usually located at an RF current maximum along one of the loop's sides as shown in **Figures 4a** and 4b. As shown in Figure 4a, if the feed point is situated at the midpoint of the horizontal side that is parallel to the ground, or at the loop's apex, this enables the antenna to be horizontally polarised. Referring to Figure 4b, when the antenna's feed point is situated at an electrical quarter wavelength (λ/4) from the apex down an upright side, then the antenna is vertically polarised. Either of the upright sides could be used as is convenient. Figure 4a shows the RF current flow when the delta loop is horizontally polarised. The RF currents in the upright vertical sides cancel and result in virtually no RF radiation from the vertical sides. However, the RF currents flowing in the horizontal side are additive and result in the antenna radiating a horizontally polarised RF signal. The feed point could also be placed at the loop's apex or the centre of the horizontal wire. However, the usual practice is to place the feed point centrally in the horizontal side as the most practical option. **Figure 4b** shows the general RF current flow in a vertically polarised delta loop. The RF currents in the upright sides are additive and result in the antenna radiating a vertically polarised RF signal. However, the RF currents in the horizontal side are subtractive and result in virtually no RF radiation from the horizontal side. If the ground under the horizontal wire has poor RF conductivity, this can tend to impair the antenna's performance. This can be improved by using a wire ground screen under the antenna which should not be connected to the antenna. For both horizontally and vertically polarised delta loop antennas, when the antenna is near-vertical to the ground, the feed point impedance is balanced and is about 100Ω. However, the actual feed point impedance is influenced by the antenna's height above the ground. In a similar manner to the square loop antenna, a delta loop antenna tends not to be influenced by 'end-effect' and this results in a resonant loop antenna being slightly longer than one wavelength by a factor around 1.028 to 1.06 [2]. Another advantage of using a delta loop antenna is that it requires less than three quarters of the space required to install a λ/2 dipole. However, the antenna's apex should be more than 0.3λ above the ground to allow the antenna to be suspended vertically. This could be a drawback depending on the wavelength used because the lower horizontal side may be too low to avoid people, pets, or anything else making physical contact with the antenna and also meet with ICNIRP requirements.

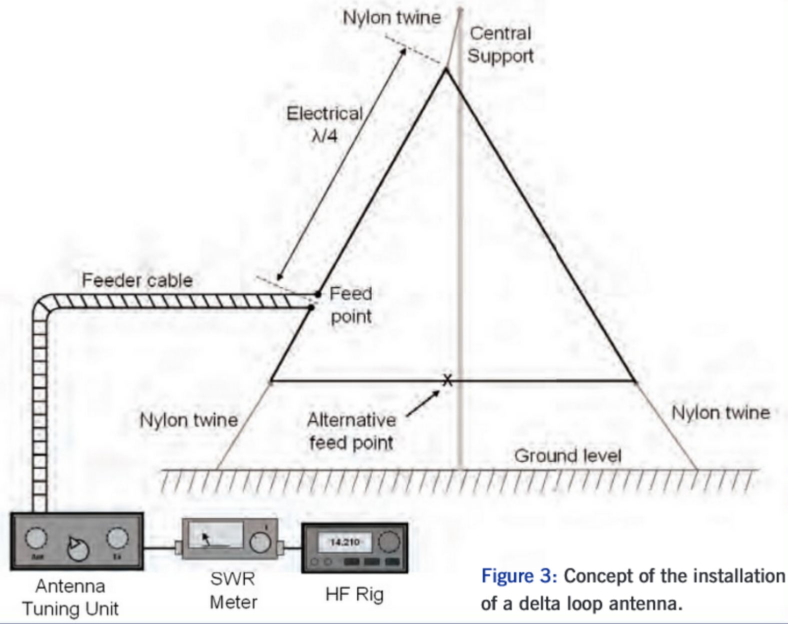


Figure 3: **Concept of the installation of a delta loop antenna.**

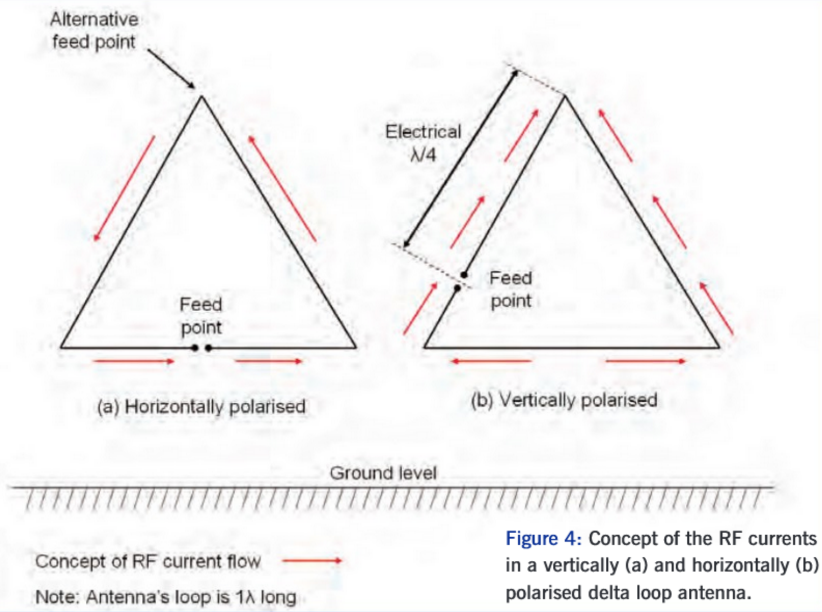


Figure 4: **Concept of the RF currents in a vertically (a) and horizontally (b) polarised delta loop antenna.**

Mike Parkin, G0JMI
email2mikeparkin@gmail.com

Radiation pattern comparison

A square loop and a delta loop antenna each having an electrical 1λ perimeter were modelled using MMANA-GAL with the lowest side at λ/8 AGL (above ground level), with the application's default ground setting selected [4]. This enabled radiation patterns for both types of the antenna to be predicted when using either horizontal or vertical polarisation. **Note:** Using as an example a loop antenna for the 40m band, if its lowest side is at λ/8 AGL then this side will be about 5m above the ground. However, if the loop antenna is for the 20m band, its lowest side will now be about 2.5m above the ground. Therefore, to comply with ICNIRP requirements, a higher frequency band loop antenna may need to be installed at a greater height above the ground.

Horizontally polarised antennas: Figure 5 shows the predicted radiation patterns for both antennas when the feed point was situated in the lower horizontal side of each antenna, making the antennas horizontally polarised. The radiation patterns for the square loop antenna are the blue curves, while for the delta loop antenna these are shown as the red curves. The horizontal plane curves show a slight variance between the antennas, with the delta loop giving better all round coverage along the Y-axis compared to the square loop by up to about 4dB. However, this difference would probably not be that obvious when working stations. The vertical plane curves show significant radiation at more acute angles above each antenna. This indicates both antennas should favour working medium distance and more local stations, although it should still be possible to occasionally work more distant stations when the propagation conditions allow.

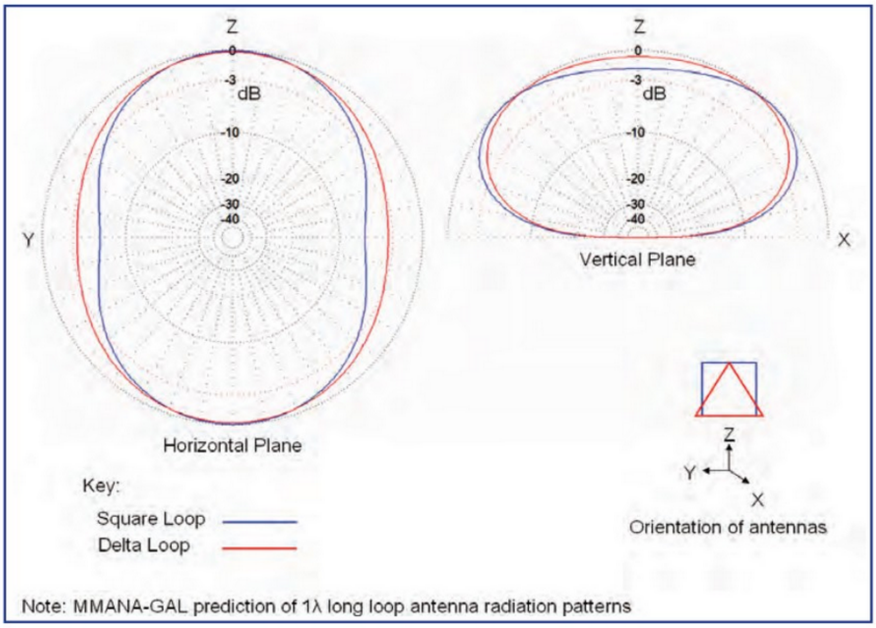


Figure 5: Predicted radiation patterns for horizontally polarised square loop and delta loop antennas.

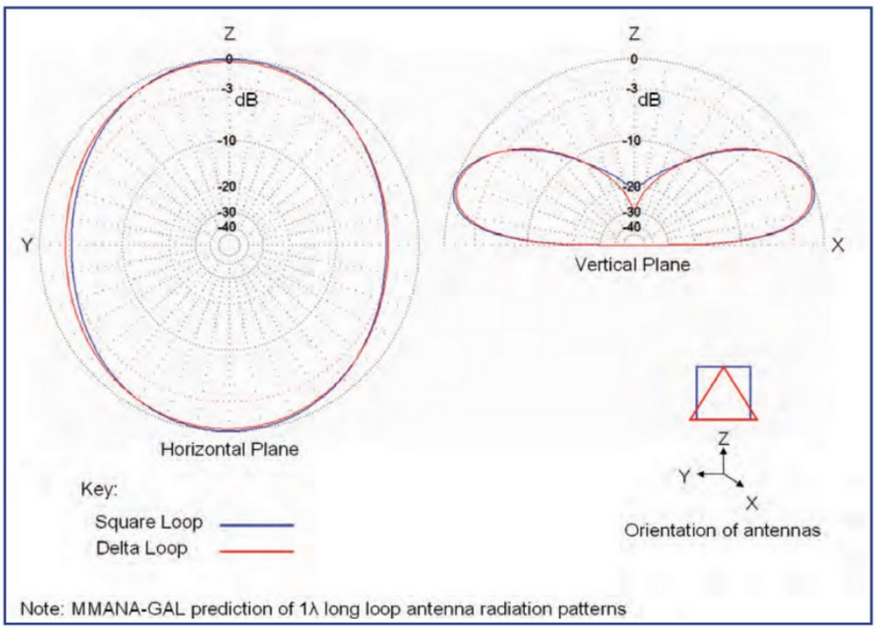


Figure 6: Predicted radiation patterns for vertically polarised square loop and delta loop antennas.

Vertically polarised antennas: **Figure 6** shows the predicted radiation patterns for both antennas when the feed point was situated in one of the upright sides of each antenna, making the antennas vertically polarised. Again, the radiation patterns for the square loop antenna are the blue curves, while for the delta loop antenna these are the red curves. The horizontal plane curves indicate the antennas give a reasonable all round coverage. In the vertical plane the prediction indicates both the antennas radiate RF energy at a low angle towards the horizon and should favour more distant stations to be worked depending upon the propagation conditions. Compared to Figure 5 in the vertical plane, the prediction for the vertically polarised antennas indicates significantly different radiation patterns compared to the horizontally polarised antennas. Both **Figures 5** and 6 indicate similar performances for either a square loop or a delta loop antenna when their perimeters are 1λ long, have the same operating frequency and are at λ/8 AGL. Based on the angle of the radiated RF power level referred to the horizon, this indicates which polarisation type is the most advantageous to use where:

Horizontal polarisation should favour working more local and medium distance stations for either antenna.

Vertical polarisation should favour working medium and more distant stations for either antenna.

Reference

[1] ARRL Antenna Book 23rd edition, edited by H.W. Silver, N0AX. Chapter bm5 Loop Antennas, pages 5.1 – 5.9.

[2] All about Cubical Quad Antennas 2nd edition by W.I. Orr, W6SAI. Published by Radio Incorporations Inc, 1970. Chapter 3, page 39.

[3] RSGB Radio Communication Handbook, 14th edition edited by Mike Browne, G3DIH: Section 15, Practical HF Antennas, Pages 15.26.

[4]] MMANA-GAL basic V3.0.0.31, freeware antenna analyzing application, by Makoto Mori JE3HHT. MMANA-GAL basic and Pro by Alex Schewelew DL1PBD and Igor Gontcharenko DL2KQ. 1999 onwards. Default ground: Dielectric 13, 5 mS/m, others 0.

An improved antenna rotator controller

Photo 1: Front panel of the rotator controller, 4 preset buttons, CW & CCW and Azimuth control.

Introduction

The Yaesu G800DXA antenna rotator at G4WIM was ten years old and overdue for maintenance. As I had just installed a new MM0CUG telescopic mast it seemed like an ideal opportunity to perform the service. A strip down revealed the ball races were corroded and needed to be replaced – easily done. A closer look at the shack end controller revealed nothing untoward but it was noticed while bench testing that its repeatability and accuracy were marginal for my 23cm 44-element Yagi. For 24cm DATV every fraction of a dB is valuable and thus having a reliable accurate rotator is essential. This led to the idea of creating an improved controller with better accuracy/repeatability and ease of use especially when employing high gain narrow beam width UHF antennas. As the controller design was nearing completion, I thought it would be interesting to include a RDF Doppler facility as the extra hardware required would be fairly minimal. Therefore, when not being used as a rotator controller it can be switched into Doppler RDF mode. Part two of this design describes how the RDF function was incorporated, how it works and how to access it. Referring to the original Yaesu G800 controller, while functional it can be difficult to calibrate, and only has limited functionality: clockwise and counter clockwise rotation, speed and a basic preset control. During development of this design, several prototypes were built ultimately resulting in a design employing a STM32 series Nucleo board. The selected STM32F401re Nucleo board has an M4 Cortex 32 bit processor running at 84MHz controlling a 320x480 TFT display with touch using a SPI interface. The Nucleo also has a 12 bit ADC resulting in 0.2° azimuth resolution, along with a 12 bit PWM output for generating analogue voltages – useful in this application for controlling the rotator speed.

This solution offers significant enhancements of the basic Yaesu controller as follows:

- Set ability to within 1° using a rotary control

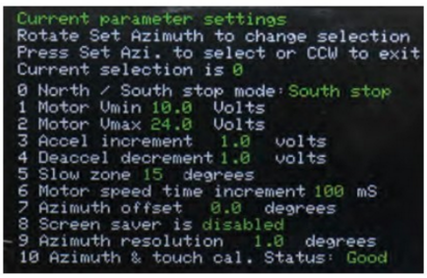


Photo 2: Parameter settings menu.

- Four non-volatile presets for ease of use
- Configurable controlled acceleration/de-acceleration between points to avoid stress on the mechanics and overshoot.
- Slow speed manual control for fine tuning the heading
- Automatic calibration to ensure repeatability to within 1° (absolute accuracy is determined by how well the antenna system is aligned to true North and the linearity of the rotator

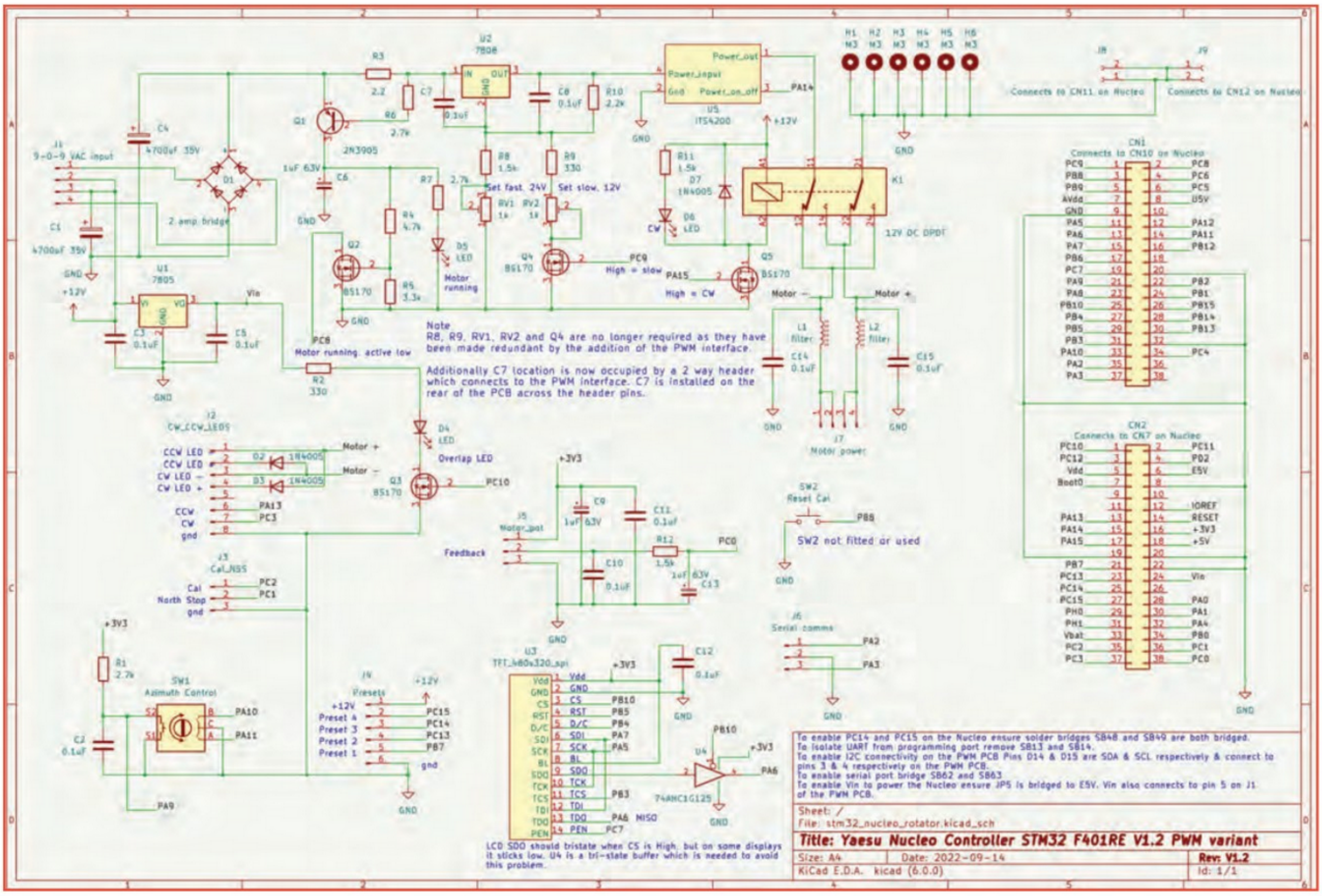


Figure 1: STM32 Controller PWM Variant schematic.

potentiometer amongst other factors).

- North/South stop operating modes
- Configurable 450° rotation with over/under lap control and indication
- Low cost 3.5" SPI LCD display with configurable screen saver
- Touch screen and adjustable azimuth resolution
- USB port for easy firmware upgrade
- Serial control of azimuth, screen saver and other parameters
- On screen configuration menus for most parameters and functions
- Solid state load switch for the rotator motor to protect against overload/short circuits etc.
- Optional RDF mode to be described in part 2.

Although the design is primarily intended for Yaesu rotators it could be adapted for other models which use low voltage DC motors and a potentiometer for feedback. *I'm currently working on an interface board to control an AC reversible motor.* It is not intended to be a fully featured K3NG type controller with a Nextion graphics display, rather just a mild upgrade from a standard Yaesu type using a standard low cost TFT 3.5" display.

Main Circuit description overview

The controller can be considered in two parts. One part contains the microcontroller (STM32) plus firmware and the other part is for power/motor control. The STM32F401RE Nucleo board has plenty of spare IO pins, RAM and flash (524,288 bytes available) which is just as well because the compiled code is 147KB or thereabouts. The STM32 has four outputs to the power/motor control side:

- Motor on/off
- Motor speed control via PWM and an external interface PCB
- Motor direction, clockwise or counter clockwise (DPCO relay)
- Over/under lap LED

Motor speed control voltage is configurable from 8V to 24V. Similar to what the minimum and maximum voltages are from the Yaesu G800 controller. The controller ramps up the voltage at the start of the operation and ramps it down as it approaches the target heading. The start and end voltages along with the increment size and timing are all fully configurable by means of a menu. Having a controlled gradual start and stop reduces stress on the rotator gears and limits overshoot when stopping. If any Azimuth movement is aborted (by pressing CW, CCW or Set Azimuth) the rotor does a controlled emergency stop – ie ramps the voltage down quickly. *Originally the design only had two speeds (fast and slow) set by a pair of preset resistors. The schematic and PCB still shows this functionality but as indicated there, parts are no longer required. Instead, a small PWM interface PCB is used which also supports the EEPROM functionality, as described later.* The LCD updates the heading and reports the status of the power signals to the controller ie:

- Motor is running, active low
- Voltage from sensor potentiometer to display actual beam heading

The STM32 uses SPI running at 50MHz to update the screen and interact with the touch screen.

Note: when building the code you have to edit a file to configure the display connections and SPI speed etc – this information is detailed in the firmware under a tab called "important notes".

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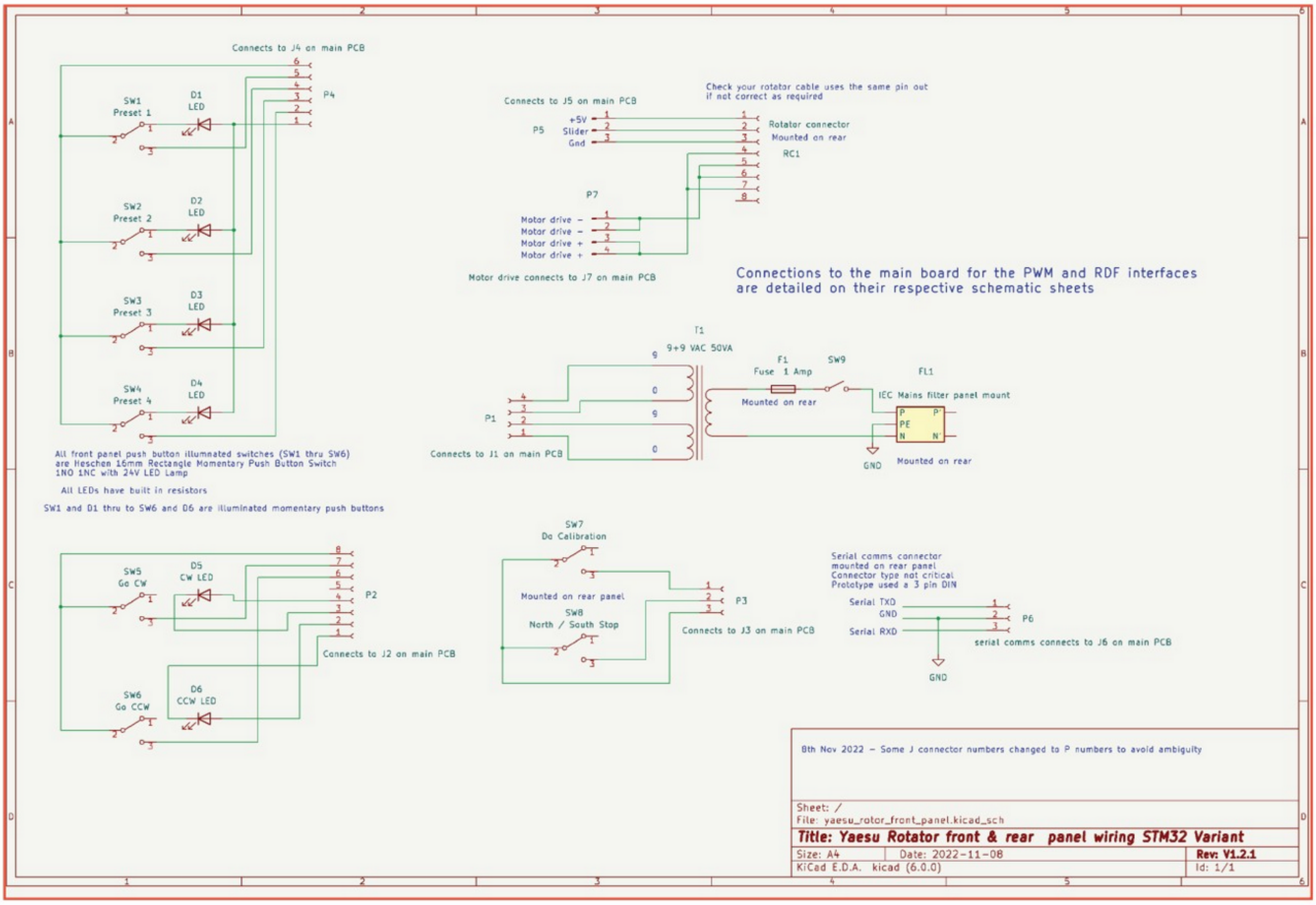


Figure 2: Front and back panel input and output schematic.

When the controller is moving the antenna, the LCD updates the heading approximately 25 times per second.

Main PCB detailed circuit description

Referring to **Figure 1** STM32 schematic, the electrical design is relatively simple but does have some safety features to prevent damage to the controller or rotator. The power supply uses a conventional 9–0–9 V, 50VA toroidal transformer and full wave bridge rectifier to generate just under 30V DC (no load) for the motor voltage regulator which is set to generate between 8 and 24V by adjusting its ground pin voltage by means of the PWM voltage interface pcb. It also produces 15V DC from the centre tap which is regulated down to 5V for the STM32 and the PWM variable voltage controller driven by the PWM output from the STM32 on PC9. For the motor power control, I opted to use a 7808 fixed voltage regulator and make it adjustable for several reasons:

- 7808 devices are very common in junk boxes
- They have built-in protection circuits for current limiting
- By controlling the ground reference pin from the PWM variable voltage converter, any voltage from 8 to 24V can be generated
- Absolute regulation and accuracy are not important due to motor DC cable losses
- 7808 regulators and less expensive than the adjustable LM317 types

The choice of secondary voltage is determined by the maximum DC voltage desired for the motor in use. I opted for a 9–0–9 to enable 24V for the motor maximum and to minimise heat dissipation when running at 10V slow motor speed. There is also a motor current sensing circuit comprising Q1, Q2 and R3 which is used during calibration to determine when the motor has stopped due to it hitting the end stops. It also operates should the end stops be hit during normal manual operation and shows an error message on the LCD. Power to the motor is turned on and off by means of a 'smart' load switch. This device provides short circuit, surge and 'load dump' protection. The rotator direction is reversed by means of a DPDT relay K1 operated by Q4. The relay is rated at 2 amps and must handle, not switch the current. The Yaesu motor draws circa. 600mA, so there is adequate margin – indeed this relay is larger than the relay used in the Yaesu controller. If the design is adapted for a more powerful motor then the PSU and relay may have to be uprated. The LEDs in the manual control buttons (CW and CCW) reflect what the rotator motor is doing as they are back-lit and powered from the motor supply thus their brightness changes with the voltage applied to the motor. They can be observed to be getting brighter during the acceleration phase and dimmer during deceleration. To avoid any potential RFI problems to or from the controller the power/signal pins are all RF filtered, as is the mains inlet. The unit is built into a metal case for the same reason. Figure 2 shows the front and rear panel wiring. The Bill Of Materials is at: https://tinyurl.com/yer5srrm

PWM Interface Circuit description

This is a necessary add-on PCB which serves two functions.

1. EEPROM support for storing various parameters and calibration data.

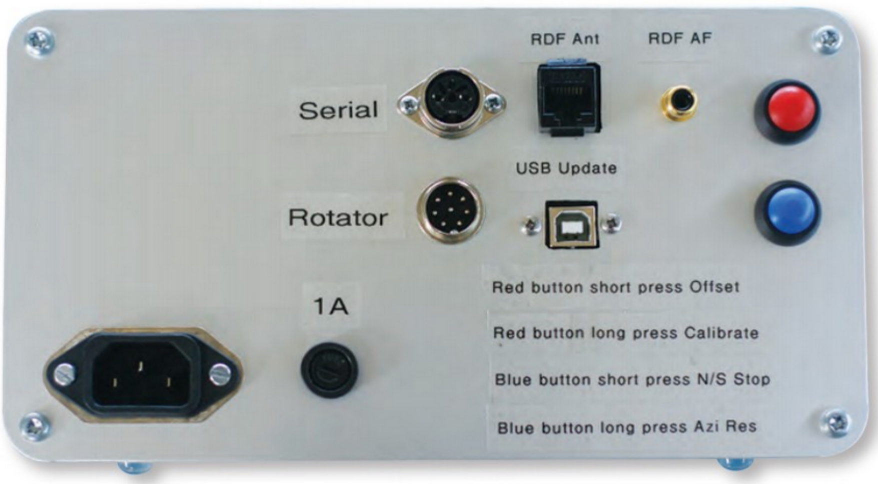


Photo 3: The Back Panel of the unit with all interfaces shown.

2. Converting the PWM output into a DC voltage ranging from 0 to 16V.

Referring to schematic Figure 1, U1 is the EEPROM which connects to the main PCB via a two-wire I2C interface. It is powered from the 3.3V supply taken from the Nucleo PCB. U2 is a simple power supply inverter. It takes 5V from the Vin pin on the Nucleo and inverts it to -5V to ensure the output of U3 can go down to 0V. The PWM output from PC9 on the Nucleo PCB is filtered by R6 and C4 into a DC voltage ranging from 0 to 3.3V. U3 with a gain of 4.9 increases the maximum voltage to 16.17V. As the output of U3 is connected to the ground pin of the 8V regulator U2 on the main PCB, this means a final output voltage range of 8 to 24.17V. However, the firmware limits the maximum to 24V. In operation, the power supply can be configured from the main menu to set the Vmin, Vmax, voltage step size and timing such that practically any voltage slope can be created for a given rotator.

Firmware

The code is written using the popular and free Arduino environment. The 'sketch' is extensively documented over several tabs to aid maintenance and modification to other platforms. The Arduino environment uses C++ making it very easy to read and understand its operation. Additionally, by using a high-level language (as opposed to assembly or machine code) it makes it fairly trivial to develop and add features to the code or to port it to other platforms. I hope the sketch and its embedded comments are adequate to enable further development/improvement by others. Note, when the code is first built it is likely that some library items will be missing and thus cause a build failure. Looking at the error output will show what's missing and needs to be installed, usually in the form of a zip library. The STM32 environment needs the following libraries and the IDE needs to have the STM32 board manager installed. A link is provided further down on how to accomplish this.

```
#include <Coordinates.h>// convert cartesian to polar requested azimuth
#include <SPI.h>
#include <TFT_eSPI.h>
#include <Adafruit_GFX.h>
#include "Free_Fonts.h"
#include <FlashStorage_STM32.h>
```

A simple web seach of the above names should take you to the latest version which can then be downloaded and installed as a zip library. *As noted above – there is a tab called "important_notes". Within this are instructions on how to edit the TFT spi driver set up. Failure to follow these changes may result in the display not operating as intended.* Significant time has been spent writing and optimising this code, but could doubtless be improved. Initially it was thought the code would be simple but over time it became apparent that this was not to be the case. Considering what it has to do in more detail it becomes quite involved – more so when it has to accommodate both North and South stop operation and management of the Yaesu 450° overlap when in full automatic mode, touch screens, menus etc. Plus, it has to display a moving compass smoothly and without significant delay. All told, the code has taken about 10 months to develop and test to a stage where it is relatively 'bug free'. The development of the code was aided by the use of a pseudo rotator ie a geared DC motor driving a 10-turn potentiometer. Using this test jig greatly accelerated code writing and debugging as the jig could 'rotate' much faster than an actual rotator. In essence, the code first does a set-up to configure the inputs/outputs and parameter loading before entering the MAIN loop which cycles around every 100uS checking to see if anything needs to be done. There are also three interrupts, two from the rotary encoder to set required azimuth and other parameters and a third from the encoder push button to select/deselect/abort. As the code is largely non-blocking this means that while the rotator is moving automatically to the desired heading, the desired heading itself can be changed 'on the fly' by means of the rotary encoder or touch screen or aborted by pressing CW, CCW or set azimuth. I'm not going to try and explain the code any further as there's lots of comments in the sketch explaining what's going on. The STM32 F401RE code can be accessed online. Note how in the comments at the start of the code it explains how to set up the programming tool options. Make sure you select the correct Nucleo board. This link https://github.com/stm32duino/Arduino_Core_STM32 shows how to install the STM32 board manager for the Arduino IDE. The firmware can be found at: https://tinyurl.com/yun29u23 I've created a shared folder for BOMs and other supplementary information.

Construction notes

As this unit is mains powered, take care with the 240V AC wiring to ensure it is not exposed.

Mechanical mounting

I made some simple aluminum brackets to hold the main PCB in place behind the front panel to avoid any mounting screws on the front panel. One of the brackets supports the PWM interface PCB and acts as a heatsink for the regulators. The other bracket is used to support the optional RDF analogue interface. All the PCBs use through-hole components for ease of assembly. The only surface mount parts are the load switch U3, and optional tri-state buffer U4, which is easily installed. When building the main PCB take care to ensure correct polarity of the components, especially the bridge rectifier D1 and smoothing capacitors C1 and C4. Note: also the two regulators U1 and U2 are not mounted on the PCB – instead they are fastened to the side mounting plate for cooling. **MOST IMPORTANTLY** *ensure that U2 is installed with an isolation kit to prevent its metal tab connecting to ground.* The entire unit is built into a 3U high case from Lincoln Binns, UnioBox3, width 224mm depth 160mm and height 120mm SKU U3S-224X160 – similar proportions to the original Yaesu controller. The toroidal power transformer is mounted on the underside of the chassis plate. The front panel mounting bracket also acts as a heatsink for the two voltage regulators. The front panel has the 3.5" display, four illuminated preset buttons,

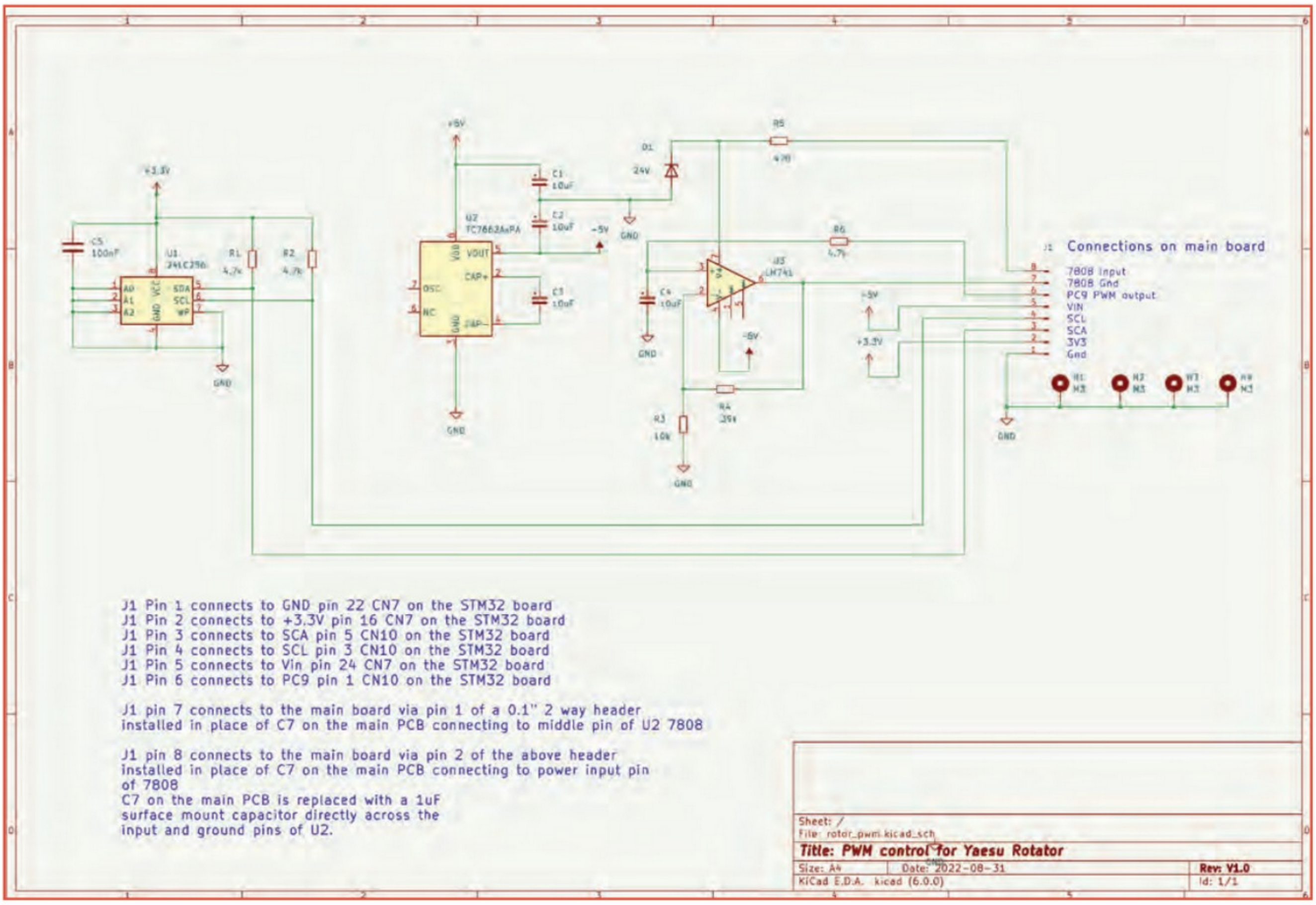


Figure 2: PWM and EEPROM Schematic.

clockwise and counter clockwise illuminated buttons, an overlap LED, a rotary/push button control to set and select the heading, power on/off switch. See photo 1. The rear panel has a filtered mains IEC inlet socket, mains fuse, auto calibration button, North/South stop selection button and connectors for the rotator, USB and serial communication. There are also connectors for the RDF function, audio input and antenna rotation. See **photo 3**. The pcb is designed to accommodate the STM32 nucleo along with the rotary encoder, overlap LED and LCD display. I would recommend staying with a metal case for RFI and screening reasons. Photo 1 shows the front panel.

Calibration

I found accurate calibration of the original Yaesu controller to be difficult, and at best not very accurate. Referring to photo 2. This design only needs the press of a button on the rear panel (or selection of sub menu item 2) to perform the calibration and results in a +/- 1° repeatability limited only by mechanical back lash. Having accurate calibration is the key to success. Calibration is performed in North stop mode but is equally applicable to South stop mode. Before attempting to perform calibration it's vital that the rotator runs clockwise or counter clockwise when the respective manual control button is pressed. If it runs incorrectly, simply reverse the motor connections. It's also necessary to make sure the polarity of the feedback potentiometer is correct. Going clockwise should result in an increasing voltage. In the case of the G800, when it is fully CW the potentiometer voltage should read maximum value ie about 3V assuming a 3.3V supply. Finally, it's a good idea to make sure the feedback potentiometer is correctly centered electrically before attempting a calibration. Initially set the rotator to its mid position, then activate the calibration function. This causes the rotator to go fully CW until it hits the limit switch which stops the motor, this is sensed by the controller and the feedback potentiometer voltage measured. The rotator then goes fully CCW for 450 degrees. Again, once it hits the end stop the voltage of the potentiometer is measured. The controller knows approximately where the end stops are and slows down as they are approached. Doing this allows the controller to calculate the ADC counts per degree and the offset from zero volts. Other calculations drive the Azimuth display. The calibration process can be aborted at any time by pressing the rotary encoder button.

Assuming the calibration process completes successfully the resulting data is displayed on the LCD and can be stored in non-volatile memory ready for re-use or ignored. The input to the ADC is filtered and has a high impedance, thus if the rotator potentiometer is worn and needs to be replaced, any relatively low value potentiometer will work. The Yaesu G800 uses a 500Ω part. For the G800 with a 3.3V reference for both the potentiometer and ADC, this results in very nearly 8 ADC counts per degree. Thus for 450 degrees (360 + 90 overlap) the total count range would be circa 3600 ADC counts. With a correctly centered potentiometer in the rotor this means the ADC count range should be from around 0390 to 3500. The above underlines the importance of making sure the feedback potentiometer is correctly centered within the rotator.

Operation

Once the unit has been installed and calibrated the display will show 'ready' along with default preset settings and current beam heading. The calibration process assumes the antenna is pointing North and has been accurately aligned with a compass. *If it turns out that true North is not quite where it should be on the display, then there is a menu option to adjust Azimuth offset such that indicated headings correspond to actual beam headings.* If the unit has not been calibrated then a warning message is displayed, and system defaults are used. A menu setting (sub menu item 4) allows cal/uncal status to be manually overwritten. It does not delete the stored data. This is useful for test purposes. Alternatively, an "f" command can be sent via the serial port to force the loading of default cal data. If no buttons are pressed, then after 2 minutes (configurable main menu item 8) the screen saver activates and displays a moving Yaesu (or other message/callsign) on the LCD. Operating any control on the front panel will bring the controller out of screen saver mode. Referring to **Photo 1**, in operation, the antenna can be adjusted as follows:

1. Briefly pressing one of the four presets will cause the rotator to automatically move to the desired preset heading. A full 360° movement takes just over 1 minute, depending on the maximum voltage that has been set.

2. Set the desired heading (green arrow and text on the display) with the rotary control, then select that heading by pressing the rotary control in. You can also simply touch the screen to set the heading and then press the rotary control knob in.

3. Using the manual CW or CCW buttons. This method is slower and takes 4 minutes for a 360° rotation. Manual control is purposely slow to avoid crashing into end stops and abrupt stops generally. To traverse quickly around the compass, it is best to use the set azimuth rotary control or touch screen function to get to the approximate location and then use manual control. Pressing the rotary control button or CW or CCW buttons while an operation is in progress will cause an abort (this includes calibration).

The four preset positions can be memorised by first using the rotary control and/or touch screen to set the desired heading, then press and hold the desired preset button to store that position. The LCD will indicate the heading has been stored. A momentary press of a preset will cause the rotator to drive toward that heading (if the screen saver is in operation, two presses will be required, the first to cancel the screen saver and the second to activate). The relevant CW or CCW push button will illuminate while the rotator is operating, and its brightness will reflect the speed, brighter being faster. When in auto mode the rotator will speed up then slow down as it approaches the desired heading. Automatic control will avoid entering the overlap zone, but if the rotor is driven into the overlap zone manually, it will then treat the overlap zone as if it were part of the normal area. The North/South stop functions allow the end stop to be placed in one of two places as follows. North stop: this allows CW rotation from due North, East, South, West, North then into overlap for East. CCW is the reverse of the above. South stop: this allows CW rotation from due South, West, North, East, South then into overlap for West. CCW is the reverse of the above. A brief press of the rear-mounted North/South stop button (or main menu item 0) will toggle between North and South stop modes. The current 'stop' state is shown on the LCD. Being able to select a North or South stop is useful depending on where most of the activity is likely to be. For example, if you were to live in a location with nearby high ground to the North likely blocking signals from that direction, then it would be sensible to use a North stop. A long press of the rear-mounted North/South stop button will enable the azimuth resolution to be changed to 0.1, 0.5-, 1-, 2- or 5-degree resolution (or main menu item 9). Instructions appear on the TFT screen. When in operation the rotary control will update the smaller desired azimuth arrow on the LCD in a smooth way with no hesitation. It is possible to set parameters using the serial port. The serial port is 115200 baud with 3.3V signal levels, thus an external interface adapter is required, typically a FTDI USB to serial converter. The available serial commands can be displayed using a serial connection and sending h? Below are a few examples:

Sending the charater 'a' will result in the current azimuth being returned sending 'c5' sets the interval of the screen saver text to 5 seconds. Available range 1 to 60 seconds, default 5 seconds. 'd45'sets the required azimuth to 45 degrees, range available 0 to 359. 'f' forces the default cal data to be loaded over the current data set and set as valid. Sending the letter 'e' toggles the screen saver function between on and off. 'f' forces dummy cal data to be good, this removes the 'No Cal' message. 'r45' results in the requested azimuth being set to 45 degrees and activating the auto mode. Available range 0 to 359. 's60' sets the screen saver timer to 60 seconds. Available range 10 to 9999 seconds. Setting it to 's0' disables the screen saver. Default 30 seconds. 't' returns the current screen saver settings If a PC with serial comms is not available then built-in menus can be used. The rotator menus can be accessed by pressing CW and holding then pressing and holding CCW then releasing both buttons. You will then be presented with a top level menu containing most key items. Menu item 10 accesses a sub menu for azimuth and touch screen calibration.

Further notes on the STM32

Using STM32 Nucleo board (as opposed to an Arduino Mega2560 or similar) enables high speed SPI communications to the TFT, and implementation of a touch screen. Additionally, as the Nucleo has a 12 bit ADC the azimuth resolution is better than 0.5 degree and the 32 bit M4 Cortex processor running at 84MHz allows a much faster/smoother screen update. However, the STM32 Nucleo does not have any true EEPROM (Electrically Erasable Programmable Read Only Memory), instead it simulates EEPROM by using its flash memory. This does lead to some constraints such as speed and longevity of the flash memory which is limited to 10,000 writes. To overcome this limitation the PWM variable voltage interface PCB also contains a true EEPROM device via an I2C data bus. Accessing the EEPROM memory space takes appreciable time (at power up it takes 10 seconds to load the various parameters, the code has been written to limit unnecessary writes).

Important – as noted on the schematic

To use the Nucleo board it is necessary to configure six solder bridges and change the position of a jumper to enable power from Vin rather than the programming USB port. Two solder bridges (SB13 and 14) are removed relating to UART functionality and the other two solder bridges (SB48 and 49) are made for IO connectivity (PC14 and 15), the details are included in the STM32 schematic. In addition, the Arduino IDE needs to be configured to work with the STM32 series of chips. SB62 and SB63 also need to be bridged to enable the serial port connection. As mentioned above, certain libraries need to be installed before the firmware will compile correctly and upload into the target. This link https://github.com/stm32duino/Arduino_Core_STM32 shows how to install the STM32 board manager for the Arduino IDE.

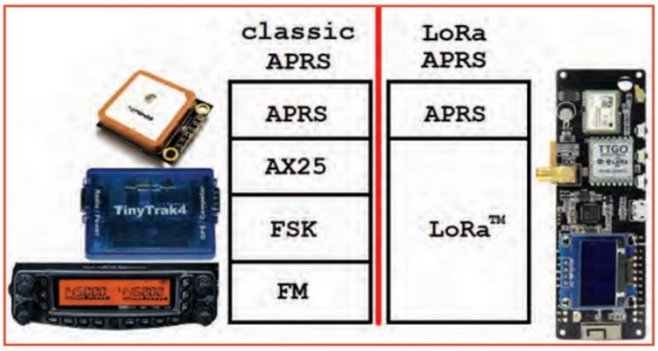
Conclusion and future enhancements

This has been a fun project to develop and results in a useful addition to the shack. Earlier versions used an Arduino Mega 2560 but it quickly became apparent that a STM32 Nucleo was going to be required to support extra features such as:

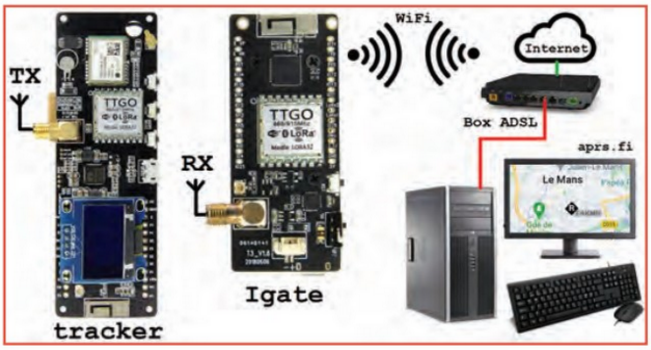
Configurable moving text screen saver, Azimuth offset, Azimuth resolution, ramp up/down timings and voltage outputs. All of the above have now been implemented. I hope that others find this design useful and have similar success. I currently have a small number of PCB's board sets for this latest version and, for those who are not comfortable with the programming side of things, I can provide pre-programmed STM Nucleo boards.

Keep a look out for Part 2 where I add RDF functionality to the project.

APRS with TTGO LoRa module Introduction



Hardware comparison between a classic FM beacon and LoRa.



APRS transmission chain.

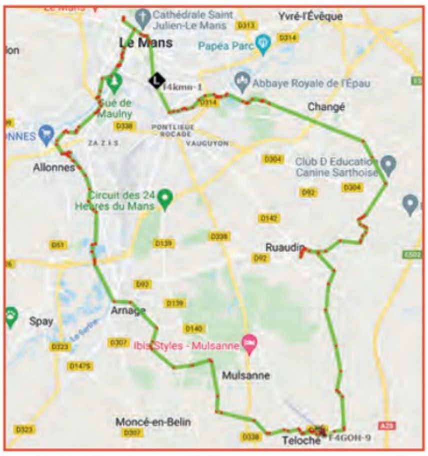
APRS (Automatic Packet Reporting System) developed by Bob Bruninga, WB4APR, is a real-time local communication system for the rapid exchange of digital data (geolocation, weather beacons, telemetry, messages, etc.). A classic APRS beacon requires an FM transmitter (144.800MHz) and a TNC (Terminal Node Controller) using FSK (frequency-shift keying) modulation. In recent years there has been a proliferation of modules based on the ESP32 microcontroller. Among these modules, there is the TTGO T-Beam, which combines all the following elements to create a compact APRS beacon, ESP32, LoRa SX1278 modem delivering 20dBm, GPS, Power manager, optional OLED I2C display and 3.7V battery holder. This results in a significant reduction in hardware volume compared to a conventional FM device. The user only needs to implement the software in the ESP32 microcontroller and build the box.

Peter, OE5BPA, the initiator of the project, had the idea of using the technology dedicated to IOT (connected objects) to transmit APRS frames and thus be able to use existing TTGO modules [1].

Reminder of APRS transmission chain

The beacon or mobile station (TTGO T-Beam) periodically sends its geographical position (longitude and latitude). The fixed station (iGATE with a TTGO module that does not have a GPS receiver) receives the position and, via its Wi-Fi link, reports it to the APRS servers on the Internet.

The user connects to aprs.fi with a browser and then obtains the position of the mobile station. An external user may not be aware of the technology used (FM or LoRa). However, it is necessary to have enough LoRa iGATEs to relay the information in the best conditions. I assessed the beacon around my area while mobile. The reception was good within a radius of 10-15km. Of course, the



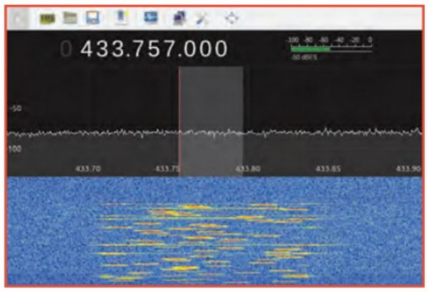
Test around Le Mans with an iGATE at the Touchard high school - Washington F4KMN.

antenna was not the original one, but a 70cm Jpole antenna, with 20dBm.

LoRa modulation

LoRa stands for Long Range. It is a technology that allows connected objects to exchange tiny amounts of data at low speeds. The LoRa radio is based on spread spectrum transmission.

To learn more about Lora technology, I recommend reading the excellent book written by Sylvain MONTAGNY at the University of Savoie Mont Blanc [2]. For developers, you will find a description of the LoRa library used in the project by Philippe Slimier [6]. Most LoRa modems cover the industrial, scientific and medical (ISM) radio frequency band on 868MHz, but there are SX1278 modules using the 430/440MHz UHF amateur band. You will therefore need to be careful when purchasing the TTGO to ensure



LoRa modulation in the GQRX waterfall.

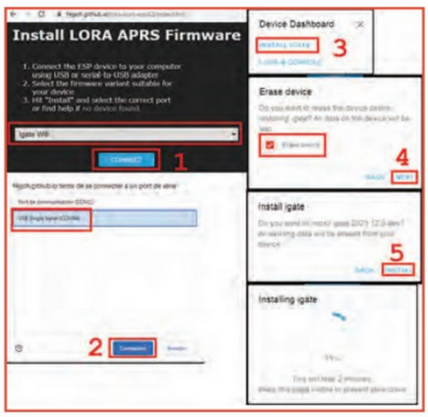


Figure 1: The different steps to program the iGATE software.

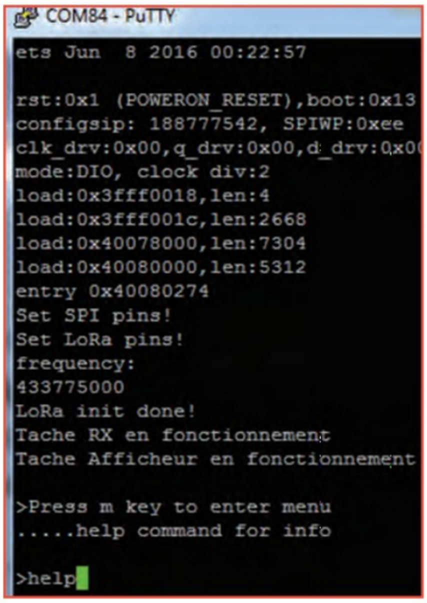
that you have the correct frequency range for the transmitter and receiver.

iGATE Software programming

I advise you to start by programming the TTGO iGATE first, and then check the Internet connection is correct via Wi-Fi. There are two possibilities for this:

- Use Peter's original program on his Github

Touchard High School Radio Club home screen Washington F4KMN.



To access the configuration menu, press the “m” key when starting the ESP32.

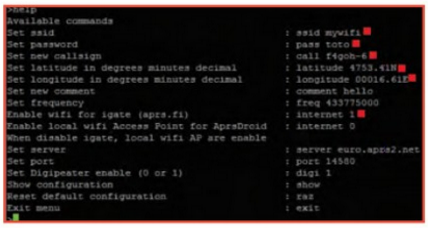
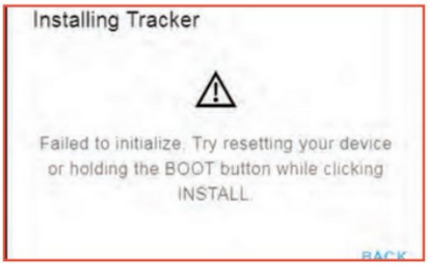
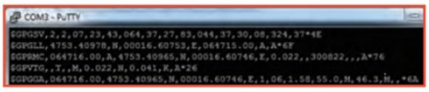


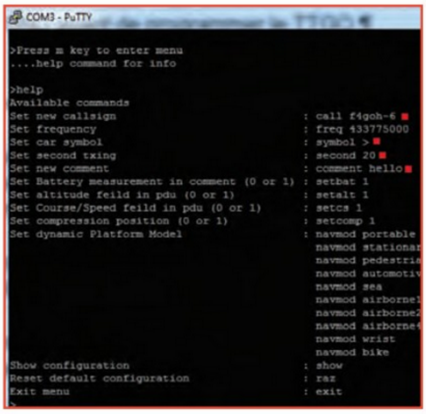
Figure 2: At least six parameters are needed to configure the iGATE (the ones in red)



In case of an error, GPIO0 must be connected to GND before programming the TTGO.



The Reset GPS program displays NMEA frames on the serial port at 115200 baud.

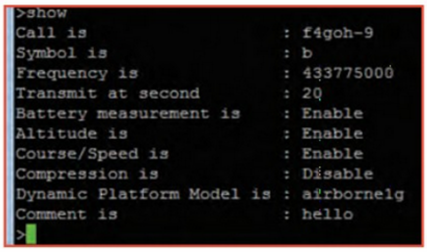


At least four parameters are needed to configure the tag (in red)

Dynamic Platform Model Details

Platform	Max Altitude [m]	MAX Horizontal Velocity [m/s]	MAX Vertical Velocity [m/s]	Sanity check type	Max Position Deviation
Portable	12000	310	50	Altitude and Velocity	Medium
Stationary	9000	10	6	Altitude and Velocity	Small
Pedestrian	9000	30	20	Altitude and Velocity	Small
Automotive	6000	100	15	Altitude and Velocity	Medium
At sea	500	25	5	Altitude and Velocity	Medium
Airborne <1g	50000	100	100	Altitude	Large
Airborne <2g	50000	250	100	Altitude	Large
Airborne <4g	50000	500	100	Altitude	Large
Wrist	9000	30	20	Altitude and Velocity	Medium
Bike	6000	100	15	Altitude and Velocity	Medium

GPS configuration models, by default it is in “Portable”, incompatible for a weather balloon.



An example of a weather balloon configuration.

Display of a frame

site. But first you will need to install visual studio code, recompile the program and modify a JSON configuration file. This is not an easy thing for anyone without experience in this field.

- Opt for a ready-to-use online programming tool. Indeed, from Peter's source code, I created my own 'version' to facilitate software programming [3].

After connecting the module with a USB cable to the PC, check that the USB serial COM drivers are installed. Windows 10 and Linux users will have no problem. Note the COM port number or newly-installed COM port. Using the Chrome browser, enter the URL [4], select “iGATE Wi-Fi” from the drop-down menu. Click on “connect”, choose the correct COM port (as found previously) and follow the programming procedure as shown in Figure 1.

Once programmed, the OLED screen displays the F4KMN school logo. The iGATE must now be configured, using a serial terminal such as Putty at a rate of 115200 Baud.

The iGATE is now programmed. The command line text remains in the serial terminal window. In Figure 2 there is an example of the configuration which is displayed. You will have to configure: the access point SSID Wi-Fi with the password, the callsign, the geographical position of the iGATE (longitude, latitude in degrees minutes decimal), activate the internet connection, display the results of the modifications (show command). The parameters are saved when exiting the menu by pressing 'exit'. Once configured, the DOTT should connect to the APRS server. An icon indicating the location of the iGATE should appear on the aprs.fi website.

This module has a bright future among makers and DIY fans. The amateur radio community is obviously not forgotten by using the existing APRS protocol. Building a beacon very quickly and at a low cost is now easy. I changed the software primarily for the launch of the weather balloon at Touchard High School, Washington in June 2022 [5]. But the application can be used in many different fields.

73 and good traffic in LoRa APRS.

Web links:

[1] https://github.com/lora-aprs
[2]https://www.univ-smb.fr/lorawan/wp-content/uploads/2022/01/Livre-LoRa-LoRaWAN-et-lInternet-des-Objets.pdf
[3] https://github.com/f4goh/lora-aprs-esp32
[4] https://f4goh.github.io/lora-aprs-esp32/index.html
[5] https://www.touchard-washington.fr/2022/05/31/le-3-juin-prochain-le-lycee-touchard-washington-prend-de-la-hauteur/
[6]https://github.com/PhilippeSimier/Esp32/tree/master/24_Lora

Anthony Le Cren, F4GOH, KF-4GOHf4goh@orange.fr

Design Notes

Your Input Please

Design Notes this month has been generated solely from input that has been sent in, all of which turned out to be very opportune since I had nothing to write about; JNT labs has not been doing much technical stuff in recent months – or at least nothing to report on here. Which does highlight one big problem: if this column is to continue in the same technical vein that it has done for the last sixteen years. I can't continue to write about just my own experiments, work and findings – there are too many other things in life worth pursuing as age progresses. I need input from you, the readers. What are you doing, developing, making, playing with, discovering? Let us know! Please send anything you have to the Email address below.

Aliases and Images

Oh dear! In October's column we looked at an aspect of Direct Digital Synthesizers. This prompted Paul, G8DVO, to write in pointing out an error in nomenclature I appear to have made many times in the past. He writes: "I get frustrated when you use the term aliasing to describe the image frequencies produced by a D to A Converter or a DDS. Aliases are a sampling thing (as in an ADC) while images are a reconstruction thing (as in a DAC). This is a common misuse of the term but that doesn't make it right. I want to make a clear distinction between aliases and images". "Aliasing occurs in the output of an ADC when the Nyquist rule is broken ie input signal frequency > output Nyquist frequency, or 0.5 times the sample rate. In this case there will be less than 2 samples per wave cycle of the input signal but those samples now represent another wave cycle with a different period (the alias) which is contained within the output Nyquist band. Hence the name alias meaning different. The signal and aliases from the ADC will appear as positive and negative frequencies because the source is real. Aliasing can also occur when a digital signal is down-sampled for decimation but here the output can appear as positive and/or negative frequency depending whether the source is real or complex. Normally an anti-alias LPF is used on the input with a pass band < Nyquist, and sufficient stop-band attenuation to block any input signals > Nyquist. In some applications aliases can be useful in which case a Band Pass Filter is used to pick out the required alias term.

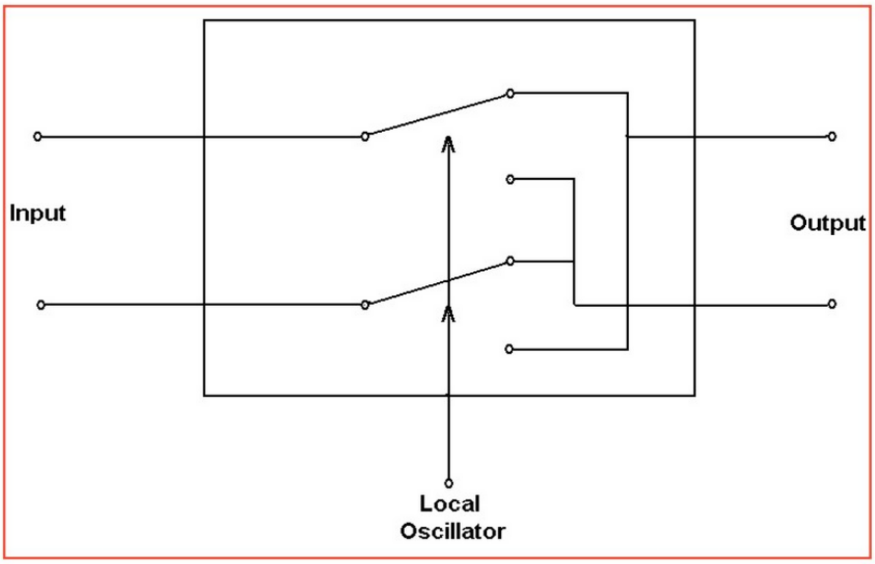


Figure 1: The ideal switching mixer.

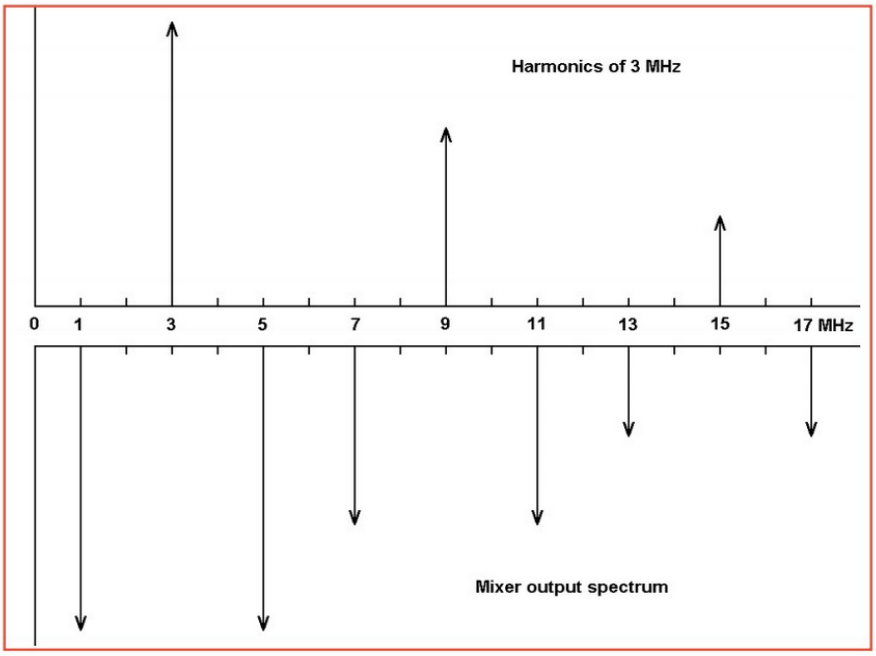


Figure 2a/b : Harmonics of a 3MHz square wave / Mixer outputs and their harmonics.

"Images occur in the output of a DAC due to the pulse modulated form of the output signal and has nothing to do with breaking the Nyquist rule. The DAC output spectrum contains the actual signal frequency (not a different one) plus an upper and a lower image frequency at each integer multiple of the sample rate. Images also occur when a digital signal is up-sampled for interpolation but here the images are on the upper and lower sides of the sample rate multiples if the source is real and on upper or lower side if the

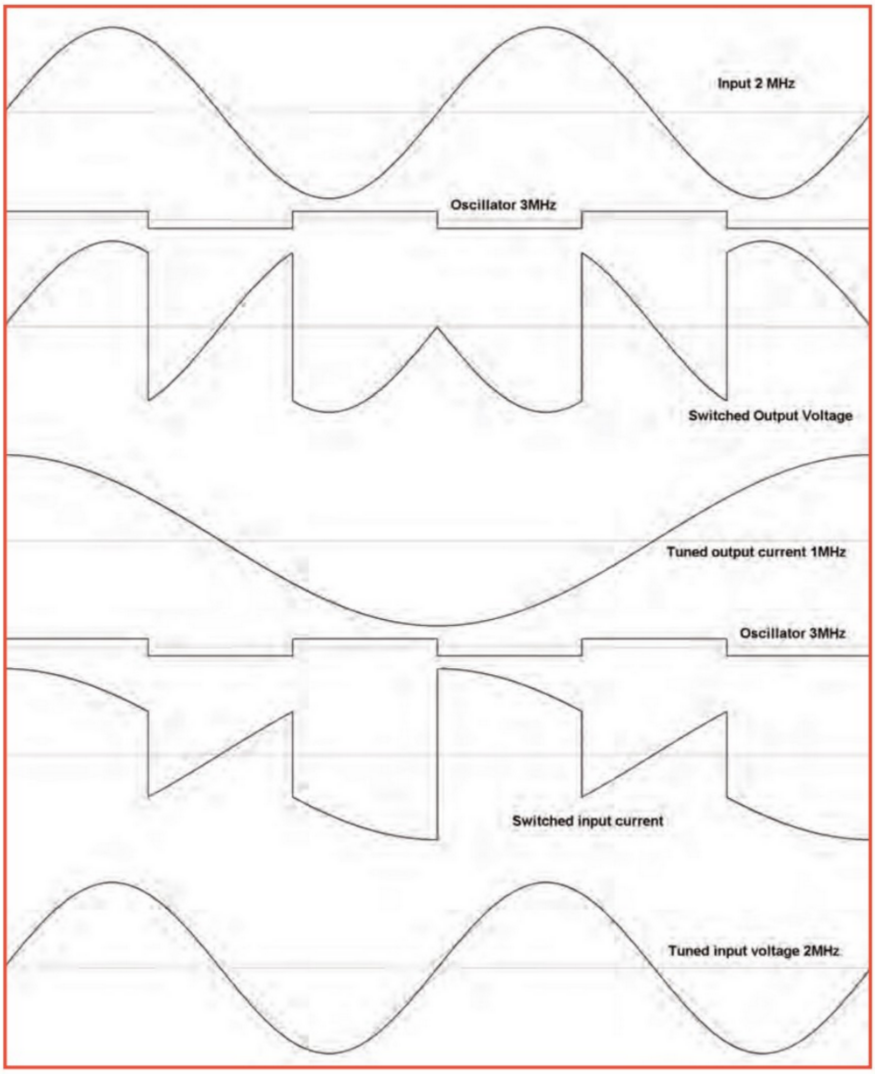


Figure 3: **Voltage and current waveforms.**

source is complex. Normally an anti-image LPF (aka reconstruction filter) is used on the output with a pass band < output Nyquist Frequency to pass the wanted signals, and sufficient stop attenuation > output Nyquist Frequency to block the image frequencies. In some applications images can be useful in which case a BPF is used to pick out the required image. "The confusion possibly arises because some try to explain aliases using Nyquist zones centred on sample rate multiples. However, unlike the image sample rate multiples, these Nyquist zones don't actually exist. They are just a visual model to determine what a particular alias frequency might be in the output Nyquist band for a given input frequency and sample rate." In defence, all I can say is that my only excuse is that as similar arithmetic is used to calculate images and aliases, it's all too easy to interchange the terms. So remember – aliases are input phenomena in an ADC; responses in a receiver, for want of a better description. Images are generated in a DAC. Think transmitter or a digital frequency source. Is there some simple mnemonic way to remember this?

Switching Mixers – Puzzles and Surprising Findings

After switching mixers were highlighted in a discussion on one of the Technical Groups, Peter Martinez, G3PLX, sent in the following that he had originally written some years ago, intended for publication. But that didn't happen, so here is the article in its entirety. Over the years the author has played with diode mixers such as the Mini-circuits SBL series, similar devices using semiconductor switches like the 4066, and related modulators and demodulators such as the Tayloe detector (also known as a Quadrature Switching Detector or QSD). Sometimes they worked well, sometimes the conversion efficiency was poor and distortion was evident. Sometimes the performance improved inexplicably when minor changes were made to the input or output circuitry. These puzzles have led the author to make a detailed study of the basic circuit theory behind these mixers. This study shows that, although a broadband resistively-terminated switching mixer is easy to understand and reproduce, resonating the input and output of such mixers can cause malfunction if not properly done. This article highlights this aspect of the design of such mixers and lists a set of rules which can be applied to the design of circuits using switching mixers so that these problems do not arise. Think of a perfect switching mixer as a black box containing a two-pole two-way switch and two pairs of terminals. With the switch in one state, one pair of terminals is connected straight to the other pair and in the other state the two pairs are connected with the terminals transposed. It's a polarity-reversing switch as shown in **Figure 1**. We tend to think of one pair of terminals as the input and the other as the output, but it works just as well either way round. The switch is toggled by the oscillator input which can therefore be thought of as a two-state (logic) input rather than an analogue one. Diode rings are one type of switching mixer, as are mixers using semiconductor switches. Active mixers using transistors or integrated circuits, in which the input and output ports cannot be interchanged, are not of this kind and are not considered in this article. Connect a 1 volt DC source to the input of such a mixer and a 50Ω resistor to the output. With the oscillator input held in one state, the output voltage will be 1 volt DC and the input current will be 20mA DC. With the oscillator toggling at 3MHz, the output will be a square wave of amplitude 1 volt peak. The power dissipated in the load will be 20mW, the same as in the DC case. We know from Fourier analysis that this square wave can be regarded as the sum of a series of sinusoidal waveforms with frequencies of 3MHz, 9MHz, 15MHz and so on, each one having an amplitude inversely proportional to the frequency. **Figure 2a** shows the resulting spectrum. The sum of the powers at all these frequencies is clearly 20mW so we can calculate the power at each frequency. In the case of the fundamental frequency of 3MHz, this comes to 16.2mW, with the remaining 3.8mW in the harmonics. Now change the input signal to a 1 volt RMS sinewave at a frequency of 2MHz. The input current will be 20mA RMS. The output will toggle, at the 3MHz rate, between a 2MHz sinewave of one polarity and a 2MHz sinewave of the opposite polarity. Mixer theory tells us that the spectrum of this waveform contains two main components, at 1MHz and 5MHz, that's 2MHz each side of 3MHz. In

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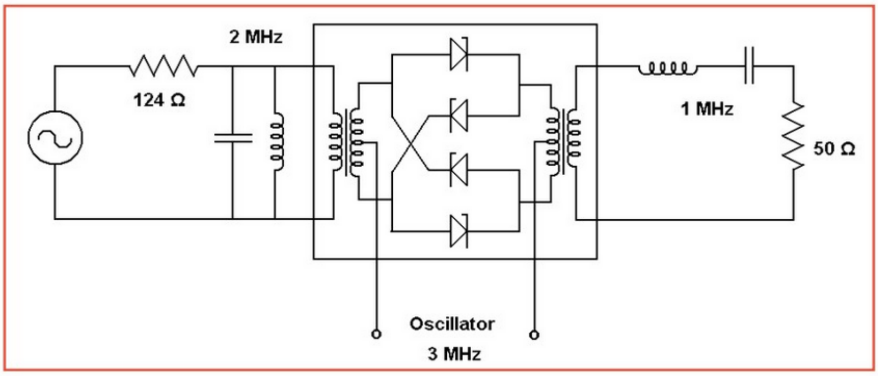

Figure 4: **The finished 'ideal' switching mixer, showing typical use of a Schottky diode ring**

addition, there will be smaller pairs of components 2MHz each side of 9MHz, 15MHz, and so on, each pair corresponding to one of the harmonics in the 3MHz square wave above. **Figure 2b**, shown upside down for clarity, shows the resulting spectrum. The combined power in the main pair is 16.2mW so the power output on each frequency is 8.1mW. If, as is usually the case, we are only interested in one of the two main output frequencies, this gives us a figure for the conversion loss of the mixer, namely a power ratio of 8.1/20, or 3.92dB. So far we have only considered what happens with a resistive load on the output and *all* the output components dissipated in this load. In practice we would want to select only one of the output frequencies and reject all the others. There are several techniques we could use to do this. Some give nice results. Others do not! Suppose we replace our single 50Ω resistive load with several 'loads' in parallel, each one consisting of a 50Ω resistor in series with a series-tuned circuit, each tuned to one of the output frequencies. The combined result will be equivalent to the original single 50Ω resistor but now each output frequency has its own resistor. The power in the 1MHz load will be the same 8.1mW that we calculated above. The insertion loss is still 3.92dB but we can see that there's a total of 11.9mW dissipated in the other resistors and it is now clear how the conversion loss arises – the input power on 2MHz has just been split between all the output frequencies. The mixer itself is lossless. Suppose we open-circuit all the output branches except the 1MHz branch. We are no longer dissipating 11.9mW in the 'other' load resistors. We started with 20mW but we now see only the 8.1mW output in the 1MHz resistor. Where has the other 11.9mW gone? The answer lies in the statement 'The mixer itself is lossless'. If the output power is 8.1mW, the power taken from the source *must* have dropped from 20mW to 8.1mW. The input voltage is still the same and removing the unwanted outputs has lowered the output current so the input current has also dropped. Let's take a closer look at the voltage and current waveforms in various parts of the circuit as we have it now. The voltage waveform on the 50Ω load, at the output end of the 1MHz series-tuned circuit, is a sinewave at 1MHz. The voltage on the input end of this series-tuned circuit is still the same as it was when we had a single resistive load, a 2MHz sinewave switched at 3MHz. The current in the load is a 1MHz sinewave so the current coming out of the mixer will be the same. This current flows through the mixer switches, so the current at the mixer input *must* have a switched waveform similar in shape to the voltage waveform at the mixer output but this time it's a 1MHz sinewave switched at 3MHz. Recall that when the load was resistive the input current was a nice clean 2MHz sinewave. Tuning the load to clean up the output voltage has left us with an unclean input current! This switched current waveform contains two main components, at 2MHz and 4MHz, with pairs of harmonics as before. To clean it up we need to remove the 4MHz and harmonic components, leaving only the desired 2MHz current to flow through the source. We can achieve this by adding a 2MHz parallel-tuned circuit across the input – no current flows through it at 2MHz but it shorts-out the other frequencies. Both input voltage and current are now sinewaves and we know that the input power is 8.1mW so we can calculate the input impedance from Ohms law ($R=V^2/P$) to be 124Ω (that's actually 50 x $\pi^2/4$). The finished mixer circuit now has a parallel-tuned input which matches a 124Ω source and a series-tuned output which matches a 50Ω load. **Figure 3** shows all the waveforms and **Figure 4** the final circuit. Remember that we can interchange the input and output terminals, which makes a mixer with a series-tuned input and a parallel-tuned output. The impedance now steps up by the factor $\pi^2/4$ from input to output. It is most important to recognise that we *must* have a parallel-tuned circuit at one end and a series-tuned circuit at the other. A parallel-tuned circuit forces the voltage across it to be a sinewave and a series-tuned circuit forces the current through it to be a sinewave. The mixer switches transform a sinewave voltage on one end to become a switched voltage at the other and the same transform applies *independently* to the current through the switches. Parallel-tuned or series-tuned circuits at *both* ends would force sinewave voltages or currents at both ends which the mixer would transform to become switched waveforms at both of the other ends! Such a mixer reduces to being a short circuit or an open circuit. This example with both the input and output at high frequencies can be extended to cover the case where the input or output is DC or low-frequency as in a modulator or demodulator. For example, if we have a capacitor across the output of a switching demodulator, we must *not* have a parallel-tuned circuit across its R.F. input. It does seem that this aspect of switching mixer design is little-known. In many textbooks no mention is made of this important interaction between the input and output. Some circuits seen by the author are wrong and in others a network has been added, usually at the mixer output, to ensure that the load impedance is resistive at all frequencies, not just at the selected output frequency. Such a network always contains some resistive elements. This makes the mixer behaviour immune to the circuit impedance at its other end but results in a conversion loss. There would seem to be a case for changing this network to one (without resistors) which meets the rule described here. For example if the input to an upconversion mixer in a low-frequency receiver is a pi-section lowpass filter, there should be a series-tuned circuit at the mixer output. This may result in an improvement of up to 3.92dB compared to the use of a resistive network. The Tayloe detector or QSD, used in direct-conversion receivers, is another kind of switching circuit which needs some care in its design. In the Tayloe, the antenna goes to a four-way switch, each output having a capacitor across it. The switch is 'rotated' at the frequency of the wanted signal. The low-frequency signals on the four capacitors can be processed to make an image-cancelling demodulator. There is normally a need for a filter between the antenna and the switch input. The output impedance of this filter must be high at harmonics of the switch frequency in order to avoid shorting-out the stepped waveform which is present at the switch input. There *must* be a series inductor here, not a shunt capacitor. To summarise the points highlighted in this article: Some aspects of switching mixers design are not well known. Some switching mixer circuits may not work properly. Some may be acceptable but with a greater conversion loss than necessary. A switching mixer with broadband resistive terminations at both ends is easy to design but it will have an insertion loss of 3.92dB. A switching mixer with a broadband resistive termination at one end can be either series- or parallel-tuned at the other end. A switching mixer which is tuned at both ends must be parallel-tuned at one end and series-tuned at the other. Such a mixer is lossless in both directions. The impedance presented by the parallel-tuned end of the mixer in (4) will be $\pi^2/4$ times that at the series-tuned end. The secret of switching mixers, therefore, lies in the careful use of tuned circuits at each end.

Best of 2022's Books

Three of the best from 2022

By the time you read this you might have completed your seasonal gift list. Just in case you haven't, here are three of the best sellers from 2022.

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NanoVNA Explained for Radio Amateurs

A practical guide to Nano Vector Network Analyers

Mike Richards, G4WNC

The introduction of the NanoVNA series of vector network analysers (VNA) has made this sophisticated RF measurement technique accessible to many radio amateurs. However, there is a steep learning curve to negotiate before achieving the level of understanding required to produce reliable results. This new book guides you through that learning curve and provides practical guidance for a wide range of common measurement scenarios.

The first section of the book provides background explanations, whilst the second concentrates on practical measurement guidance. VNA basic principles are well covered and are followed with details of the NanoVNA design compromises and their impact on measurements. Those looking to purchase or upgrade a NanoVNA will find the section covering the project evolution helpful. This takes a close look at the various models of NanoVNA and offers some purchasing advice. The NanoVNA has enjoyed plenty of firmware development to expand its capabilities, and this area is well covered and includes links to official firmware sources. The remainder of Part 1 provides introductions to the Smith Chart, S parameters and the all-important calibration process. This section concludes with a close look at how to expand the capabilities of the NanoVNA using your computer and the popular NanoVNA Saver and SimSmith software. In Part 2, the author concentrates on practical applications for the NanoVNA and covers an extensive range of measurement situations. The description for each measurement includes detailed menu navigation steps, along with clear illustrations of the connections and items required, such as test leads, terminating plugs etc. The test methods are also well described and include tips to help avoid common pitfalls. The first few examples cover everyday antenna-related measurements such as impedance matching, feeders, ATUs and resonant stubs. A perfect present for the holiday season.

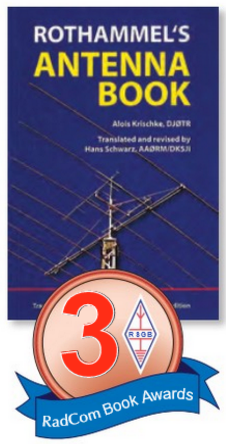
Amateur Radio Insights -Work the World with DMR

Digital Mobile Radio Explained

by Andrew Barron, ZL3DW

I have sung the praises of Andrew Barron many times in the pages of *RadCom* – and deservedly so, as he has a real knack for clear, concise explanations of even the most technically demanding subjects. And there's the rub: most people would agree that digital radio brings a whole new world to us amateurs but many have found the learning curve just too steep, discovering to their cost that digital radios generally don't work right out of the box in the same way we're accustomed to with their analogue brethren. It is not as simple as entering a couple of frequencies and setting a CTCSS tone the way you would for an FM radio. There are a lot of new terms to discover including dashboards, zones, receive groups, colour codes, code plugs, hotspots, Parrot, talk groups, and time slots. Also, acronyms like MMDVM, CPS, IPSC2, DMRMARC, TGIF, and DMR+. Use your DMR radio to its fullest potential this festive season.

Rothammel's Antenna Book



Known in Germany for years as the "Antenna Bible", this is the first English translation of the 1500+ page reference for antenna design covering LF to the microwave bands. The book begins with comprehensive sections covering the general principles of antennas and electromagnetic waves. This sets a solid foundation for the terms used in the rest of the book. Next is an examination of a broad range of transmission lines and matching systems. The book's main section provides a detailed look at almost every type of antenna. The antennas are narrowly grouped by type. For example, there are separate chapters for horizontal HF monoband, horizontal HF multiband, long-wire arrays, broadside and end-fed arrays, loop antennas, compact antennas, horizontal directional HF antennas, etc. Each chapter begins with an introduction describing the key features of each antenna type and references to the articles that either introduced or refined the designs. If your favourate amateur doesnt own a copy, now's the time to get them one.

Review

2m Band Pass Filter from Antennas-Amplifiers

The Serbian-based Antennas-Amplifiers have released a heavy-duty 2m bandpass filter, rated up to 1.5kW, which is now available from Waters & Stanton [1].

In this review we'll look at some of the reasons you may need a bandpass filter (BPF) in the first place and the features of this specific model.

The purpose of a filter in the output of any transmitter is to prevent spurious emissions such as harmonics and unwanted mixer products from reaching the antenna. Bandpass filters specifically pass the frequencies between two cut-off frequencies and heavily attenuate those both below and above these frequencies.

It's good practice to prevent unwanted signals (harmonics) leaving your shack in the first place, which a bandpass filter can help to minimise. This is especially pertinent when you want to run an amplifier at the full legal power of 400W. You also need your signal to be as clean as possible, as a courtesy to other users on your intended band and other bands. Of course, the usual linear amplifier rules of minimising the drive and operating well inside the maximum output of the amplifier still apply, which will minimise non-linearity and harmonics in the first place. See, there is a reason we can buy amplifiers rated well above our maximum legal power here in the UK. An amplifier rated at 1kW will work away all day at 400W, without being driven anywhere near its limit, resulting in a cleaner transmitted signal. This is where we initially minimise the production of harmonics at our station, then using a filter to further ensure that any which remain don't get to the antenna.

A bandpass filter can also help to attenuate unwanted, out-of-band signals which are strong enough to overload your receiver, affecting its ability to receive wanted, in-band, signals. They effectively allow only signals in between the lower and upper frequencies of the bandpass filter and attenuate those which are below and above these.

This review covers the 1.5kW version of the Antennas-Amplifiers 2m Bandpass Filter, which is very much heavy-duty in anybody's book. They do offer lower-rated versions for 2m and other bands, as well as a 4kW-rated 2m device. The 1.5kW model would be most suited for a serious contest or EME station that intends to run full legal power (400W) using a linear amplifier. The filter is rated according to the SWR, allowing for some tolerance on mismatch at lower power levels. At 1:1 it can handle 1.5kW, at 1:2 it's down to 1kW and at 1:3 it's only 800W.

The front of the bandpass filter, displaying the N-Type connectors for the input and output.

The left-hand side has a PC-style 12V exhaust fan to help with heat dissipation at high power levels.

The bandpass filter covers 144MHz through to 148MHz, more than adequate for our UK 2m allocation and includes the temporary 146-147MHz allocation available via a Notice of Variation [2] for Full licence holders. While we'd all want zero loss when inserting a filter inline, in reality there's always some loss. On this unit, the manufacturer's claimed attenuation through the pass band is as follows 144MHz: -0.09dB, 14MHz: -0.086dB, 146MHz: -0.09dB, 148MHz: -0.089dB. A very respectable loss of -0.09dB or less.

The unit design could be described as utilitarian. It's a simple, large grey metal box. This makes sense given where the unit is likely to end up, out of sight! Usefully, the unit comes with wall fixing points, so you could position it where your coax comes into the shack. The filter uses sensible N connector types to connect the input and output.

The unit comes with an exhaust fan to keep everything nice and cool. Additionally, the metal case is large with intake vents on the other side, to assist heat dissipation. To use the fan you'll need to connect a 12V supply to it.

Antennas-Amplifiers say the intention for this device is for it to be used as a transmitting filter primarily, although the design also allows it to be used as a receiving filter, for instance, to protect a wideband LNA or other sensitive devices from strong out-of-band signals.

It would effectively protect the transceiver from nearby HF transmitters, for example in DX expeditions or in multi-multi contest environments. This BPF will suppress any signals from HF (1.8-28MHz) stations, 50-54MHz and 430-440MHz. I remember the need for a bandpass filter like this a couple of years ago when we were setting up 6m, 2m and 70cm stations to all operate at the same time, all using full legal power.

Performance

The claimed attenuation results of the filter are: 1.8MHz: -100dB, 3.8MHz: -91dB, 7.1MHz: -81dB, 14MHz: -67dB, 21MHz: -58dB, 28MHz: -51dB, 50MHz: -34dB, 54MHz: -31dB, 144MHz: -0.09dB, 145MHz: -0.086dB, 146MHz: -0.09dB, 148MHz: -0.089dB.

The filter will supress harmonics produced by your station, which are especially noticeable and problematic for others when you're using a linear amplifier as well. The manufacturer supplied the following harmonic suppression and interfering signal suppression ratings [3]:

288MHz: -40dB, 296MHz: -37dB, 430MHz: -41dB, 432MHz: -42dB, 440MHz: -44dB, 576MHz: -36dB, 720MHz: -34dB, 880MHz: -33dB (GSM – Mobile phone), 930MHz: -31dB (GSM – Mobile phone), 960MHz: -29 dB (GSM – Mobile phone), 1296MHz: -17dB.

Conclusion

Not every VHF station will have the need for such a heavy-duty bandpass filter for the 2m band. But if you're running full legal power, then this could be a useful edition to your station. It would be especially useful to both multi-band contest and DXPedition stations.

Thanks to Roger, G3MEH for review input and Waters & Stanton for supplying the filter. It's currently on sale for £269.95.

Web search

[1] Waters & Stanton product page: https://www.hamradiostore.co.uk/product/dual-bpf2

[2] 146 and 147MHz NoV: http://rsgb.org/nov

[3] Manufacturers product page: https://antennas-amplifiers.com/product/band-pass-filter/2meter-band-pass-filter/2m-144mhz-bandpass-filter-1500w

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Review

SV4401A Handheld

Vector Network Analyser

Since the introduction of the NanoVNA, vector network analysers (VNA) have become commonplace in the amateur shack. The original device spawned many variants with increasing levels of capability. The SV4401A reviewed here is an example of an advanced handheld VNA providing continuous coverage from 50kHz through to 4.4GHz with N-type connectors, a large 7” display and a rugged steel case, **Fig.1**.

Fig.1: The SV4401A VNA.

In the Box

The SV4401A is supplied in a smart, rigid carry case offering good protection for the VNA and includes a practical zipped pocket in the lid to hold your test leads and calibration sets. Also included are a couple of N-Type to SMA adapters and an SMA calibration set. This set included the following parts: open, short, load and a male-male coupler. There were also two 530mm long SMA test leads, three SMA female-female couplers and a USB C cable for charging and computer control. Including a calibration kit and test leads was very welcome as I could start using the VNA immediately. There was no manual in the box, but there is an excellent 50-page PDF manual available for download from the manufacturer’s support website at: https://chelegance.github.io/WIKI/METER/SV4401A/SV4401A This website also carries the latest firmware, VNA software, and other instructions, a very helpful resource. The SV4401A was significantly heavier than any of the NanoVNA range and weighed just under 1.2kg.

Inside story

As usual, I couldn’t resist looking inside the case, and I’ve shown a photo of the control board in **Fig.2**. Here you can see the two 18650 size rechargeable batteries. These are standard removable batteries that are easy to replace. **Fig.3** shows the separate RF board that is mounted above the main PCB. The sensitive RF circuitry was carefully screened, and the two ports were well-spaced to minimise crosstalk. The eagle-eyed may also notice that the RF board uses SMA connectors, and the front panel N-Type sockets are, in fact, SMA to N-Type adapters. Whilst this is not ideal, any oddities from that connection will be accounted for during the calibration process. However, it’s worth making an occasional check that the internal SMA connectors have not worked loose. They were very firmly tightened in the review model. The provision of N-Types will be particularly helpful when using the VNA with VHF and UHF antenna systems. The additional weight of the SV4401A also helps to add stability when using N-Types.

I’ve redrawn the manufacturer’s block diagram in **Fig.4**. Those familiar with the NanoVNA range will spot that the architecture is very similar to the NanoVNA V2. One of the most important differences between the original NanoVNAs and the SV4401A is its use of a fundamental clock source instead of the 3rd and 5th harmonics used by the early models. This produces a better quality test

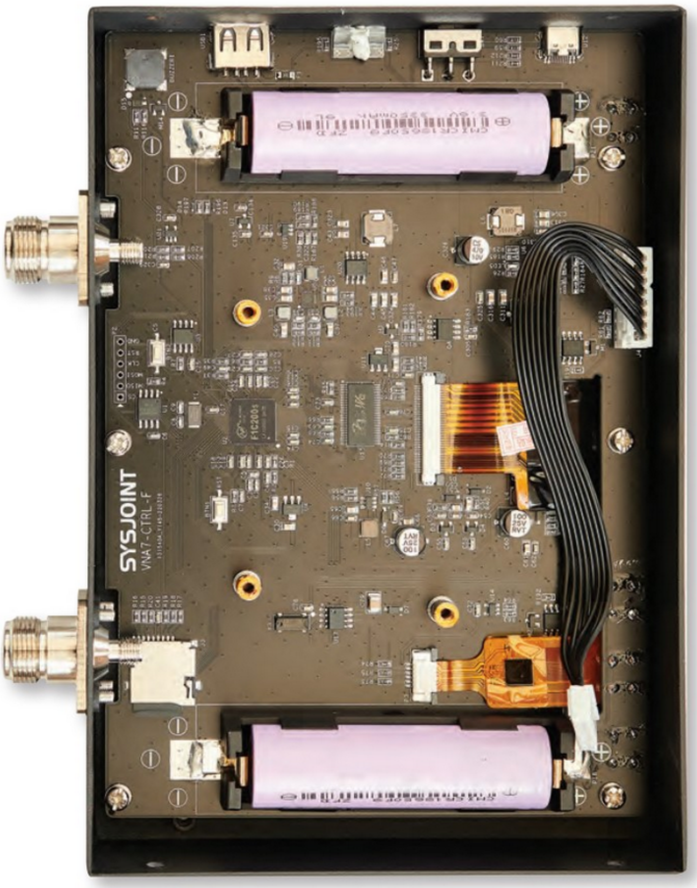

Fig.2: The main control board.

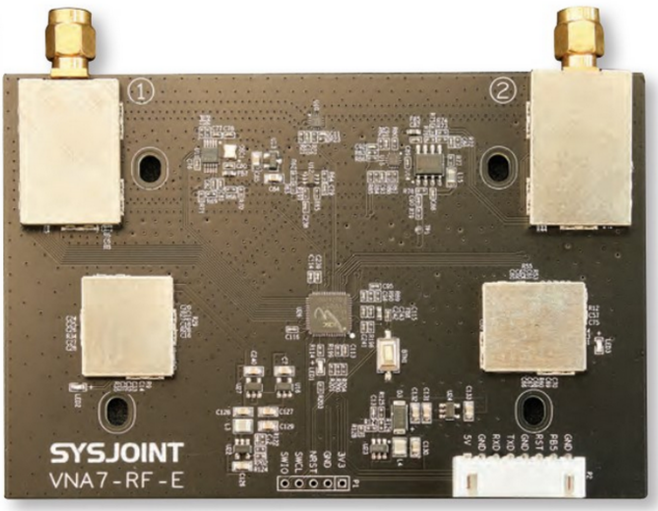

Fig.3: The RF daughter board with SMA connectors.

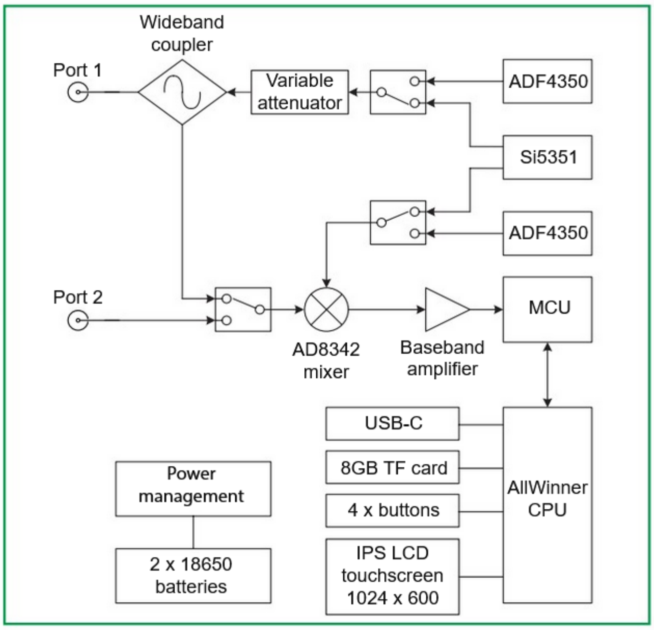

Fig.4: Simplified block diagram.

signal (stimulus) at a consistent power level throughout the 50kHz to 4.4GHz range. The higher output level also improves the dynamic range.

I'll quickly run through the block diagram here. A Si5351 clock generator supplies the stimulus (test signal) for Port 1 for LF through the VHF frequencies, whilst the first ADF4350 switches in to cover the higher frequencies. The coupler is not described in the manual, but is almost certainly a wideband resistive RF bridge that feeds the AD8342 active mixer chip. A second output from the Si5351 provides the LF to VHF local oscillator feed for the receive mixer. This switches to a second ADF4350 to provide the local oscillator at higher frequencies. The output of the mixer is applied to a broadband amplifier and then to the ADC (Analogue to Digital Converter) inside the microcontroller chip. The AllWinner F1C200s microcontroller provides all the control and display functions. One significant difference with the SV4401A is the very large (180mm diagonal) IPS display. This provides a clear and wide-angle view, which is very helpful for fieldwork. The RF design is similar to the NanoVNA V2 and one of the characteristics of this design is the use of a pulsed stimulus carrier, even when set to a single frequency in CW mode. It's the ECAL calibration process that causes the pulsing, but this limits the VNA's usefulness as a signal generator and also compromises some crystal measurements. This latter issue occurs because crystals are electromechanical devices that require a settling time after applying the energising signal. The pulsed carrier of the VNA doesn't leave sufficient settle time to make accurate measurements. However, the main benefit of this architecture is the much wider frequency range combined with the use of a fundamental clock for the stimulus signal right through to 4.4GHz. One of the many practical refinements in this model is the 8GB TF card that's used for storing s-parameters, calibration data and screenshots. In addition, I was able to save a complete VNA configuration file and give it a sensible name. This makes quick work of returning to a commonly used setup. The screenshot saves were also very useful and a good way to keep a visual record of important measurements. The SV4401A also provided a simple way to access the card from a PC. With the PC connected using the supplied USB-C cable, you power up the SV4401A with the Ctrl button pressed. This gives you the option to present the SV4401A as a virtual disk drive. You can then access the drive on the PC and move files around.

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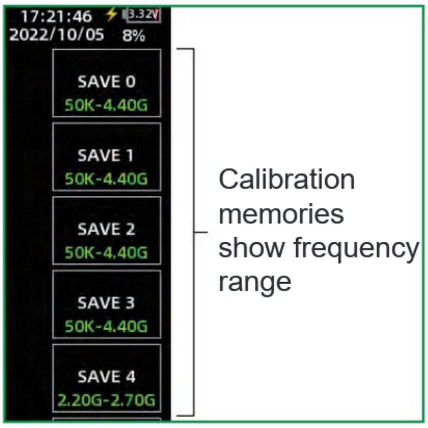

Fig.5: Screenshot of a memory location showing the stored frequency ranges

Specification summary

Frequency range	5-kHz – 4.4GHz
RF output power	-42dMb to -12dBm
Maximum Input	0dBm DC< 5V
Screen	7” IPS LCD, resolution 1024 x 600 pixels
Frequency accuracy	<±1ppm <140MHz, ±5kHz >140MHz
S21 dynamic range	75dB <3GHz
S11 dynamic range	50dB <3GHz, 40dB >3GHz
Sample points	1001 (101-1001)
IF Bandwidth	12.5kHz (6.25kHz, 3.12kHz, 1kHz, 300Hz,100Hz and 30Hz
Dimensions	190mm x 130mm x 30mm
Weight	1.15kg (1.65kg with accessories and case)

Measuring VNA Performance

Like most amateurs, I am not equipped with the latest professional spectrum analysers and VNAs to make detailed measurements of the SV4401A. However, with some ingenuity, I will show that we can get a good idea of the SV4401A's performance. I will be comparing measurements made with the SV4401A, with the manufacturer's published data. For example, I have a new 2.4GHz low-pass filter made by Crystek, and they publish the full frequency response and input/output return loss in the datasheet. If we use the SV4402A to measure the performance and get similar results, there's a fair chance that the SV4401A is doing a good job. My first step was to do a calibration run with the supplied leads and calibration kit. This was a swift process as the sweep time was just over 2 seconds for a full spectrum sweep. Once complete, I saved the calibration result into one of the 12 available memory slots. A nice touch here is that the memory display shows the frequency range stored in each location, **Fig.5**.

The first test was the s21 insertion loss, which produced the result shown in **Fig.6**. As you can see, there was quite a lot of ripple on the trace, although it followed the expected curve. This ripple is caused by impedance mismatch on Port 1 and 2 of the VNA and is common to most of the low-cost VNAs. Providing a clean resistive load up to 4.4GHz is no easy task and is one of the many reasons why commercial VNAs are so expensive.

For testing purposes, the simplest solution is to add 10dB matching attenuators on either side of the Device Under Test (DUT) as shown in **Fig.7**.

This provides a good quality 50Ω termination for the filter and masks the port impedance variations of the VNA. This comes at a price because the effective dynamic range is reduced by 20dB. You can see the much cleaner curve in **Fig.8**. This plot also shows that the right-hand side of the curve is close to the noise floor of the SV4401A. The next test was to look at the return loss of the input and output ports of the Crystek filter. I could dispense with the matching attenuators for this test and connect the filter directly to the test lead on Port 1. With the far side of the filter terminated in a 50R load, the SV4401A produced the curve shown in **Fig.9**. As you can see, this is a very close match to the Crystek figures.

The next step was to measure a 2.4GHz filtered amplifier by Analog Devices. This is an evaluation board (CN0417) designed to be used with their Pluto SDR. The amplifier includes a 2.4GHz SAW band-pass filter and delivers approximately 20dB gain across the 2.4GHz band. When measuring amplifiers, you must be careful not to overload the VNA's delicate input circuitry. This is especially true with low-cost VNAs as they don't usually have much input protection. The safest solution is to insert an attenuator in the amplifier output that matches its gain. In this case, a 20dB attenuator would do the job. However, an alternative approach is to adjust the output level of the VNA to produce a safe level at the amplifier output. The SV4401A has an adjustable output level that's ideal for this situation. I could adjust the output in 1dB steps from -12dBm and -42dBm. For this amplifier test, I set the level to -40dBm, to give an amplifier output of around -20dBm. This is well below the 0dBm maximum input for the SV4401A. The SV4401A firmware takes account of the output level setting, so the resulting plot displayed the amplifier gain correctly. The s21 gain curve closely matched the manufacturer's graph in this case. Whilst it's very convenient to run the VNA barefoot like this, there is a risk that an amplifier could deliver a fatal transient into the VNA. Therefore, I prefer to use an attenuator between the amplifier output and the VNA. Next up for testing was a directional coupler from Mini-Circuits (ZX30-20-4). This offers 20dB directional coupling between 5MHz and 1GHz. I connected the coupler with Port 1 connected to the input and port 2 on the Coupler output. The forward output was terminated in a 50R load. The measured results were very closely aligned with the Mini-Circuits specification. I even measured the forward loss of the coupler in fractions of a dB, **Fig.10**. In this supercritical measurement, I had to use 10dB attenuators on either side of the coupler to dampen the VNA mismatch ripples. In addition to basic VNA measurements, the SV4401A has some predefined analysis tools available. The first is Time Domain Reflectometry (TDR). This is shown on the display side panel and provides an accurate distance measurement to any discontinuities

Fig.6: 2.4GHz filter response with mismatch ripples

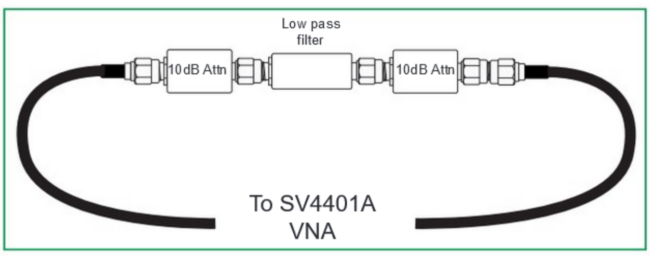

Fig.7: Connecting masking attenuators to a 2.4GHz filter

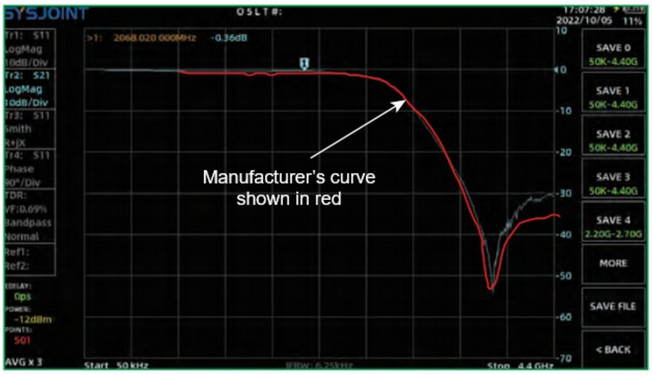

Fig.8: 2.4GHz response with masking attenuators fitted.

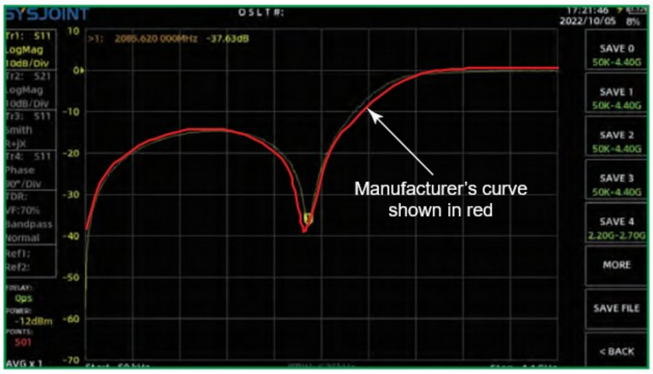

Fig.9: 2.4GHz filter return loss measured and manufacturer's result.

in a cable. This is useful for fault location or to aid with cutting precise cable lengths. Next is the LC matching calculator. This clever addition automatically calculates the LC values necessary to match the load connected to Port 1. This avoids transferring the VNA data to SimSmith to complete the calculation. The markers on the SV4401A are well thought through and can be set similarly to other VNAs. One neat addition is the Sweep analysis panel. This is activated from the marker menu and provides automated analysis of low-pass, high-pass and band-pass filters. It automatically calculates the -6dB and -60dB points, cutoff frequency and the roll-off rate. There is also an SWR sweep analyser for those working with antennas. This will list all the frequency ranges where the SWR falls below a preset value. This proved to be a quick way to check a multiband antenna. I think these tests have shown the SV4401A to be a competent unit that can handle measurements up to 4.4GHz. Great care is needed when working towards the top end of the range to avoid misleading results. Whilst it's good to have test leads and calibration kits supplied, the items are unmarked, so the quality is unknown. If you want to use the SV4401A for measurements above 1GHz, I suggest you buy known quality replacement test leads and calibration kits. My favourite source for these is SDR-Kits.net When measuring the performance of reactive devices such as LC filters, you may also need to use good quality 10dB attenuators to mask the impedance irregularities of the VNA.

Splendid Display

The 180mm display of the SV4401A is worthy of special mention as it stands this budget VNA apart from the competition. The large colour display is bright, clear and uses a capacitive touchscreen rather than the resistive units found on many budget VNAs. The touchscreen feels much more like a modern phone or tablet and is very responsive. The provision of such a good screen has enabled the designers to include many more features in the display. For a start, just about everything on the display can be touched and adjusted. For example, a single touch of the averaging display brings up a keypad entry where you can change the averaging. The same is true of the output power and most other displayed items. This means no more hunting around the menus to find the attenuator or other common features. The display traces also have another level of control, and you can toggle them on/off with a single press. However, a long press opens up a menu where you can change the format, scale and reference position. This makes the interface a delight to use. When using my NanoVNAs, I usually connect up the PC to avoid the frustration of using the screen and its menus. With the SV4401A, I found I was happy to work directly with the unit and not bother with the PC. The excellent screen with its conveniently-angled stand makes the SV4401A ideal for use on the workbench whilst creating your latest project.

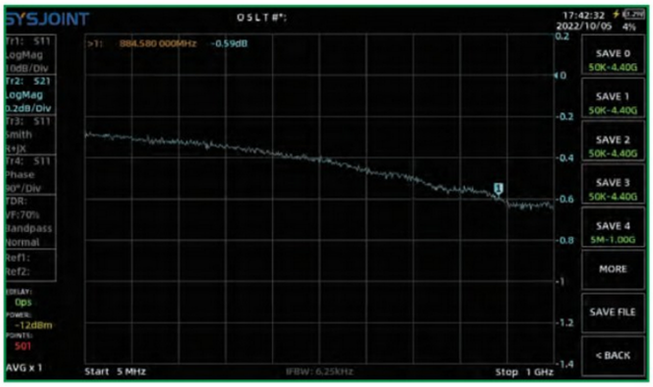

Fig.10: Forward loss plot of a Mini-Circuits directional coupler.

Computer control

The SV4401A supports computer control using a modified version of the popular NanoVNA-Saver software. In addition to providing a larger display area, PC control enables more detailed analysis thanks to the sweep segmentation process in NanoVNA Saver. This lets you cascade multiple sweeps to create plots with much higher resolution. For example, with the SV4401A set to 501 points and NanoVNA Saver set to 5 segments, the software will cascade 10 sweeps at 501 points each between the start and stop frequency. This will deliver an overall resolution of 5010 points. Using NanoVNA Saver also lets you save an infinite number of calibration runs and traces. This latter feature is ideal for keeping a record of your antenna characteristics so you can check for signs of ageing. It's not appropriate to cover all the features of NanoVNA Saver here, but it is a compelling addition to any VNA. To use NanoVNA Saver, I had to first upgrade the SV4401A firmware to version 0.2.1 and then download the adapted version of NanoVNA Saver (v0.3.10) that's available via the SV4401A WiKi. The upgrade process was straightforward and well-documented. Don't be tempted to upgrade NanoVNA Saver to a later version, as it won't recognise the SV4401A!

Summary

You've probably gathered by now that I liked the SV4401A VNA. I believe it is the best incarnation of a budget VNA that I've come across, and my wallet is being sorely tested! You need to budget for some decent test leads and a calibration kit, but that still makes SV4401A good value for money. Those interested in using the SV4401A for VHF/UHF antenna work will be pleased to see N-Type connectors, even if they are only SMA to N-Type adapters. The refined display and built-in features make this an excellent unit for the shack. The SV4401A costs £389 inclusive of VAT and is available from ML&S Ltd. My thanks to ML&S for the loan of the review unit.

EMF exposure measurement,
an experimental approach.

The problem

It should not have passed anybody by that there are new UK licence conditions which aim to limit the strength of the electromagnetic fields (EMFs) that we generate in areas open to the public. As it stands, it is almost impossible to specify the harm that is being banned and so, an international expert group has agreed a set of field strength limits as a baseline to minimise risk, we are instructed to avoid exceeding them. For EMF exposure assessment the conventional wisdom is that absolute measurement of fields by amateurs is daunting so they may otherwise be estimated indirectly from the "effective isotopic radiated power" and installation characteristics. This should work very well on those bands (VHF, UHF, microwave) where you can more or less assume the radiator is a point source not exceeding its directivity gain specification and the public will be in the "far field". The calculators available ought to give results which are likely to be ok. But HF situations like my backyard with temporary wires strung up, my attic shack end-fed long-wire or ad-hoc/portable operation leave more doubt as to whether they are anything like stock configurations. They are highly unlikely to be dangerous, especially if the maximum power limit is not used, but it might be difficult to prove exact compliance with a standard set up or the spreadsheet approach. To use the well-worn engineering phrase though, these indirect assessments are near enough for all practical purposes. I hanker after something a bit more positive though. If confronted by a neighbour worried by an imagined invisible threat from an antenna, I would far rather show them a measurement than a spreadsheet.

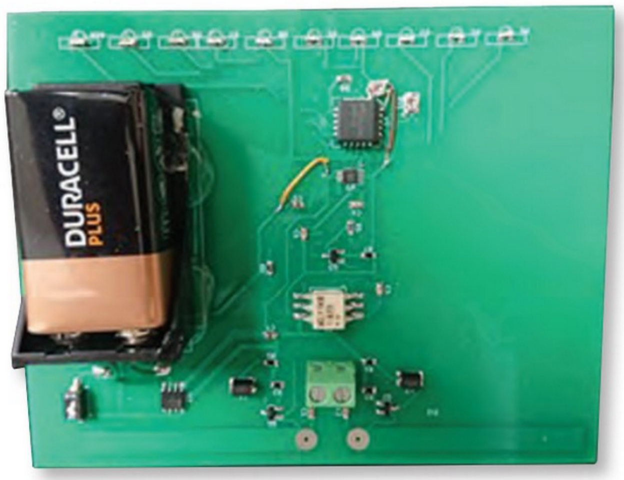

PHOTO 1: The assembled EMF meter.

Field Strength measurement

What about field strength measurement? Relative field strength measurement has been around for as long as radio, and the amateur has always been able to build or buy simple devices which show relative field strength. I guess

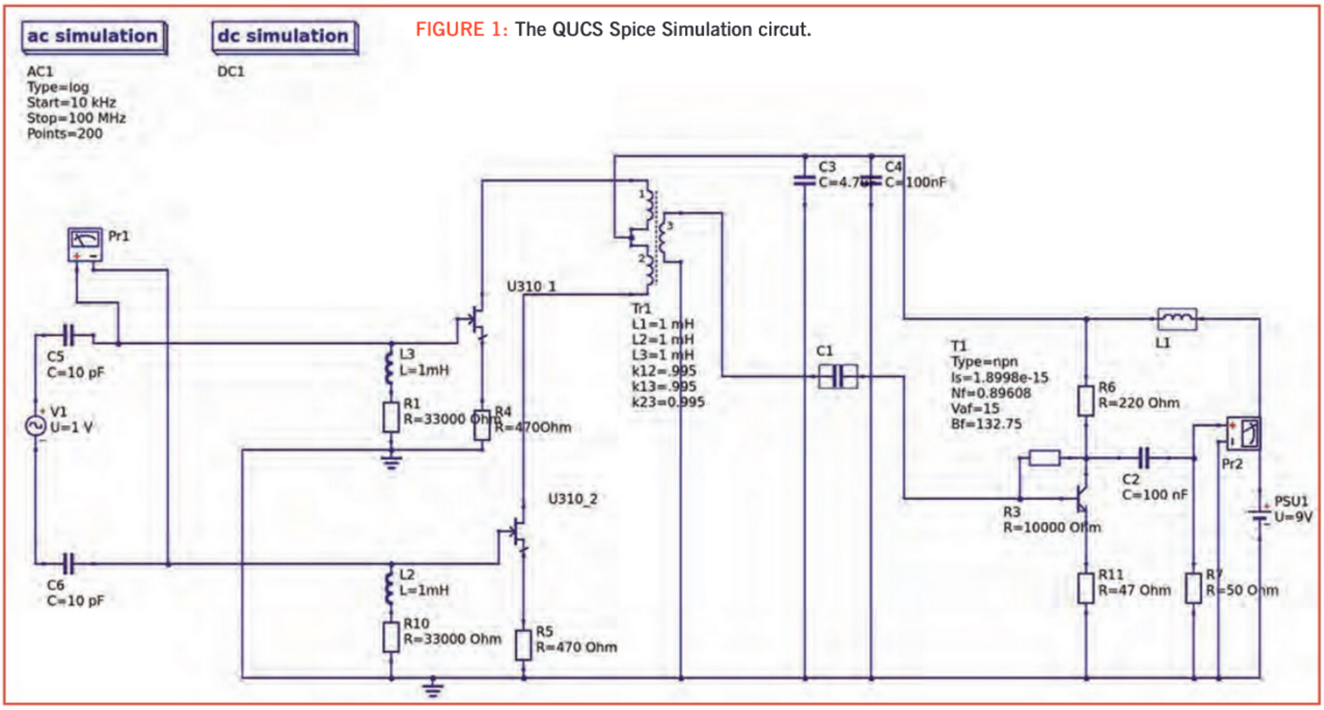

FIGURE 1: The QUCS Spice Simulation circut.

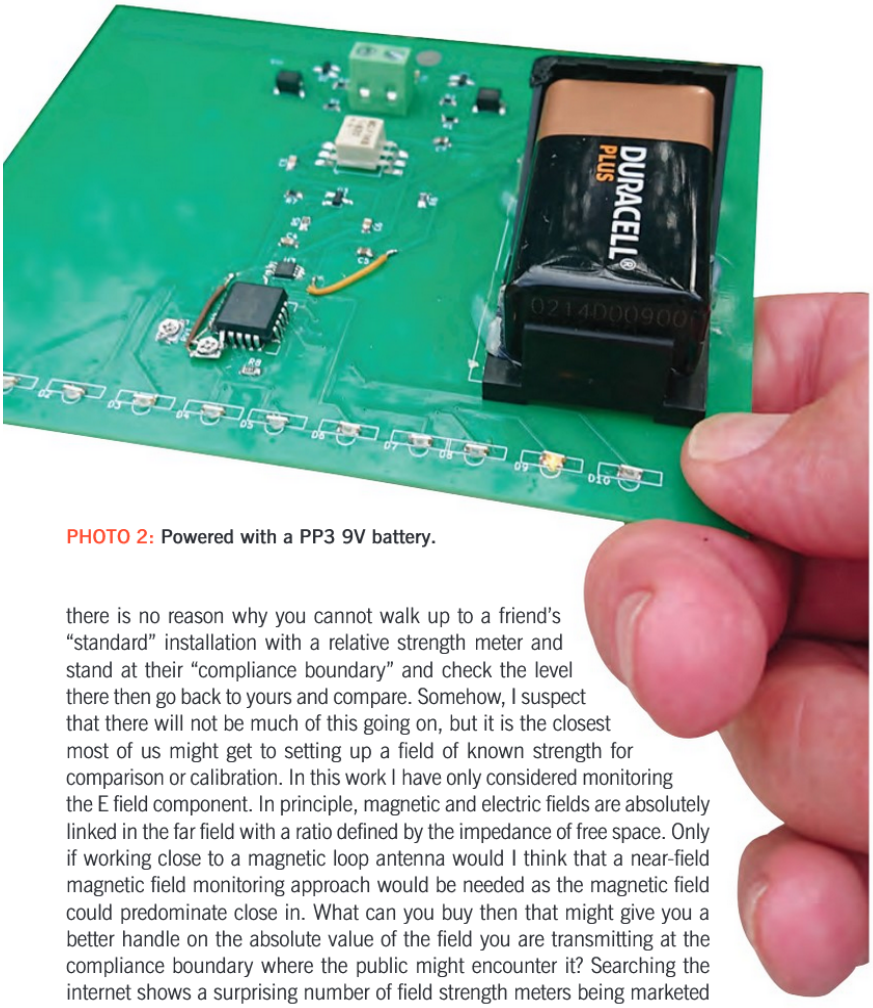

PHOTO 2: Powered with a PP3 9V battery.

there is no reason why you cannot walk up to a friend's "standard" installation with a relative strength meter and stand at their "compliance boundary" and check the level there then go back to yours and compare. Somehow, I suspect that there will not be much of this going on, but it is the closest most of us might get to setting up a field of known strength for comparison or calibration. In this work I have only considered monitoring the E field component. In principle, magnetic and electric fields are absolutely linked in the far field with a ratio defined by the impedance of free space. Only if working close to a magnetic loop antenna would I think that a near-field magnetic field monitoring approach would be needed as the magnetic field could predominate close in. What can you buy then that might give you a better handle on the absolute value of the field you are transmitting at the compliance boundary where the public might encounter it? Searching the internet shows a surprising number of field strength meters being marketed for purposes ranging from serious professional installation to hunting for fairies and ghosts, with a similar spread of cost. Reading the specifications of the devices that one assumes are at the serious end of the scale gives no indication or specification of accuracy, although display intervals are cited in field strength values. Another clue is inclusion of "calibration factor" controls. That makes me suspect that if you need an accurate field measurement with one of these you first have to set the calibration factor using a transmission with known field strength. They are probably near enough for monitoring hazardous leaks from a faulty microwave oven but uncertain for other uses. It would be possible to make a "cell" to provide a calibrated field; two large conducting plates (but small compared to wavelength) spaced apart with a known RF potential difference applied across them could generate such a calibrated uniform test field whose value is the driving RF PD divided by the distance between plates. Sounds quite hard to do properly.

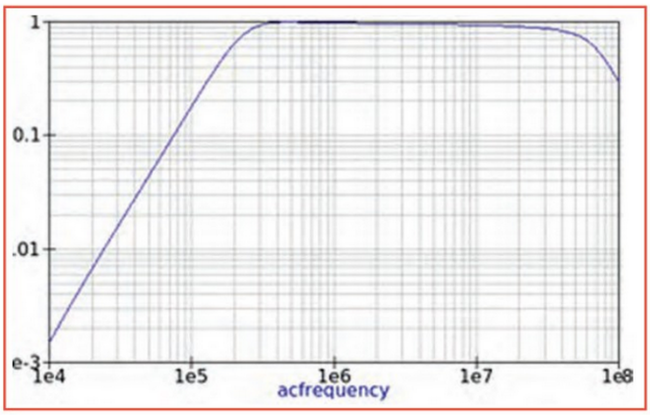

FIGURE 2: Simulated response of a high impedance buffer amplifier.

Pick-up theoretical sensitivity

What would the calibration factor of a pick-up element be in theory? The simplest thing to consider would be a short electrical wire for a pick-up. You expect this to produce a voltage in proportion to the electric field. But what do you measure it against? Ground would be an option in theory but in practice this would be somewhat arbitrary unless we have set up a special test situation which would not help measurement in useful real locations. Then the best thing here is to use another identical conductor to make an elemental dipole. You have a small open gap in the middle across which we are interested in the induced voltage from an electric field. We assume it is in a free space environment. You would expect this voltage to be proportional to the length of the dipole elements and the field strength. I found a rather nice derivation of the formula for this voltage (**reference 1**) which starts with Maxwell's equations and the boundary conditions at the dipole. The math at Reference 1 is turbid but essentially, they say that when the dipole is put in the field a response field is set up due to induced currents. Looking at boundary conditions allows the formulae describing them to be simplified to fit in; one is that the electrical potential along each element must be zero (we assume they are perfect conductors) as intuitively it shorts it out. The field model for this is that the induced current sets up a field in exact opposition to the incident field at the boundary. Another boundary condition is that conducted current must be zero at the ends and across the gap. These observations with appropriate limiting approximations (like the small size of the elements compared to wavelength) are used to collapse the complicated formulae to a very simple one giving the voltage across the gap as $V_{oc} = E_0 h$ where h is the "half height" of the dipole (the length of *one* element of the dipole pair) and E_0, the field strength per unit length. One might well have guessed that the open circuit voltage would be the dipole length times the field strength, so it was worth looking this up, as the "half height" scaling is not so intuitive. This is the theoretical scaling factor we wanted so we can go ahead with this as a ground truth and build a calibrated sensor if it uses a short dipole of known length much smaller than the wavelength being measured.

A practical calibrated sensor

A lot of field strength meters use a simple small element and a diode detector. The problem there is that if we use our short dipole this breaks the boundary condition that no current flows across the gap as the diode allows some through. So we cannot rely on the simple formula to self calibrate the pick-up. We need a buffer with very high input impedance, so it responds to the voltage without drawing current. It just so happens that whilst experimenting with compact active receiving antennas, I developed a design for just such an amplifier. This uses a pair of JFETs on the differential

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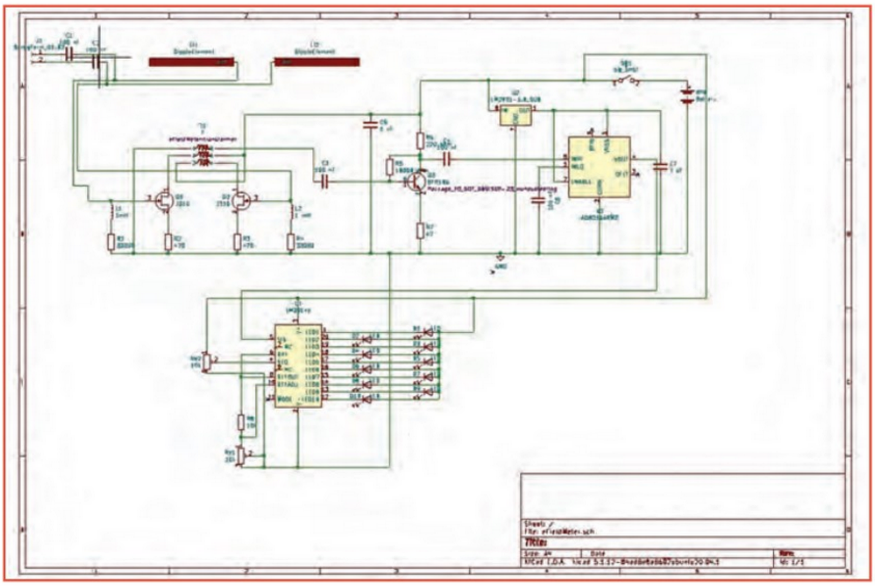


FIGURE 3: EMF exposure meter schematic (view online via the website at Reference 3).

input feeding a wideband transformer to produce the unbalanced output via an output buffer. The simulation of the circuit given in figure 1 shows in figure 2 that this has unity voltage gain with a flat enough response from 1 to 30MHz. In Figure 1 & 2 I have used a circuit similar to this with a bit more output buffering and with 1-2m long dipole elements as an experimental active antenna and it works well. The snag with a high input impedance is that the resistors in the gates will generate Boltzman thermal noise. The choke is there in an attempt to keep it out of the signal path. I do not know how well this worked and at no time was this E-field active antenna any more sensitive than my best active loop antenna. But it is ideal to use for the E-field strength meter as it will preserve the calibration factor so we can read the field strength in volts per meter from the scaled output voltage Vo with Eo =Vo/h. In practice the voltage measurement chain will need calibration with a known signal level inserted instead of the pick-up signal, which is relatively easy to do.

Experimental EMF meter

I thought the best thing would be to assemble an experimental EMF exposure meter on a single self-powered PCB. It would need a detector and indicator. For the former I decided to use a wide range surface mount logarithmic detector device: the AD8310. This is overkill in terms of dynamic range, but it is easy to use with low component count. The output of this is fed to a LED bar graph voltmeter based on a LM3914VX/NOPB PLCC Display Driver. Figure 3 shows the overall circuit schematic from which the Kicad designed surface mount 2-layer PCB Gerber files have been produced for online ordering of manufactured samples. It was designed to be able to be hand soldered which is how I assembled mine. Power is from a 9V PP3 battery. The probe pick up is 2x5cm dipole elements on the top copper layer, which can be easily seen in the photo of the assembled device, see Photo 1.

Calibration

Although we know the calibration factor of the pick-up element here, we still need to calibrate the detector and meter to a known RF voltage. I connect a known RF input level directly to the amplifier input via the 2-way connector block. If you have a calibrated signal generator, that is useful. In my case, I used my FT-817 on its lowest power setting (about 1/2W) with a 5dB 2W attenuator to give about 3V rms measured from a calibrated 50 ohms diode detector. I use a Crystek detector but have found a homemade one with a zero bias Schottky diode gives the same readings. I then insert about 30dB further attenuation (whatever is needed to give exactly 0.1Vrms from the measured voltage using the dB ratio given by $20\log_{10} V_o/V_{in}$) from a stepped attenuator to deliver 100mV rms into 50 ohms. In my case the precise actual figures were 2.8V from the detector requiring 29dB on the stepped attenuator. Note that when you connect the stepped attenuator output into the connector block, the high impedance load presented by the amplifier should mean that the full Potential Difference from the Thevenin equivalent source is applied to the amplifier (0.2Vrms) which is the desired level. We want to adjust RV1 so the top LED is just on at this level and adjust RV2 with no signal, so the bottom light is just off. It may be that there is some noise "fuzz" in the no signal case on the lower LEDs and a bit of iteration is needed for best results. What matters is the top LED indicating the reference RF input of 200mV PD. Lower LEDs then show reducing field at -2dB or so per LED. This will then equate to a top LED field strength of 1V/meter because the half height of the dipole is 5cm. This is not a bad level and is nice and easy to remember and juggle in your head. It is more than 20dB less than the maximum exposure limits at HF. Testing at low power and seeing how much headroom you have available, then after to turn the power up is quite a good modus operandi.

Field test

I undertook a live test in my East Devon alternative address backyard. The centre of the transmitting horizontal dipole was 7m up and this was the point of maximum field underneath. 10W power was used on the 10m band. At -2dB on 1V/m I had plenty of headroom for turning up the power without worrying about public being present. Running this configuration through the online RSGB calculator comes up with a compliance distance of 1.7m for 10w and 3.8m for 100W. All nicely less than the antenna height. At 400W it is marginal at 7.2m which again is in reasonable agreement with my measurement which is a bit more permissive if one is to be pedantic.

And finally

This has not been a rigorous design (for example, I haven't tried reducing the post detection bandwidth) and I have only done limited testing. But it all seems to work well enough on the three models I have built. I have uploaded the zip archive of the Gerber files to manufacture the PCB as well as the schematic and Bill of Materials files to the Git Hub repository at Reference 2. I am more than happy if anyone wishes to try out the design by building it for their own use or at cost for others. It is quite straightforward to assemble (take care with LED polarity. I used a multi-meter measuring ohms to light them up to ensure I got them the right way round!). Note that if you do intend to rely on a device like this, the licence still places a duty on licensees to keep records of assessments. It is no good just having the device, you also need to carry out tests and document them to remain within the EMF exposure terms. Using the device to confirm calculations is the most robust position to be in.

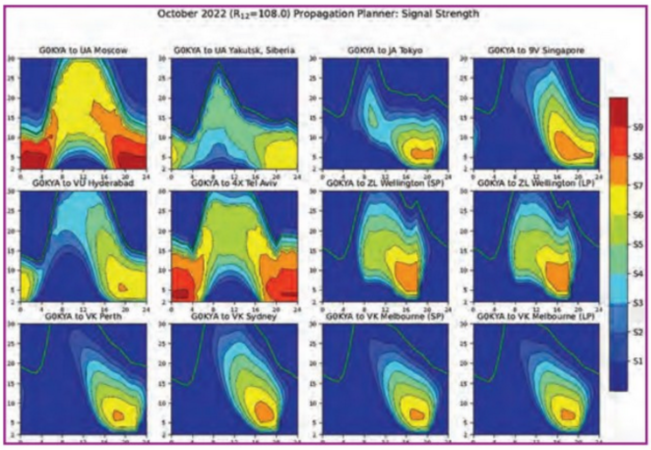
For more information about EMF visit: rsgb.org/emf

Reference 1: Voltage Across the Terminals of a Receiving Antenna by Kirk T. McDonald, Joseph Henry Laboratories, Princeton University, Princeton, NJ 08544, June 25, 2007; updated February 26, 2018

Reference 2: https://github.com/john-g8cqx/EFieldMeter

Reference 3: https://photos.app.goo.gl/2fUqcVFsw1KV25q29

Producing HF Propagation Charts using Proppy



Example propagation charts from Proppy.

The HF propagation predictions from Gwyn Williams, G4FKH, published in *RadCom* each month (page 96 this month), give you a quick look at the paths that may be open to 28 locations around the world.

These are based on 100W being fed to a dipole and were augmented last year to include FT8 using letters as well as numbers.

Now, we are well aware that some readers may be able to do far better than the predictions, with their high, directional Yagi antennas and running 400W. Or, perhaps you are a Foundation licence holder running just 10W, in which case they may appear over-optimistic. To this end, we introduced two online features that allow you to specify your antenna, height, power and mode, which enable you to tailor the predictions to your station.

Using the VOACAP and ITURHFPROP propagation engines will give more accurate, tailored results. VOACAP is the older of the two prediction engines while the newer ITURHFPROP application is in accordance with ITU-R P.533-14. Either is more than capable of giving you the predictions you need. The tools can be found at:

www.rsgb.org/voacap and www.rsgb.org/proppy.

Creating propagation charts

Recently, the Propagation Studies Committee has been looking at providing graphical propagation charts. The thought is that some users might prefer a more colourful approach to their HF predictions. The trouble is that publishing all of these charts would take up a lot of space in *RadCom*. Plus, these would again be a compromise in terms of power/antenna gain/mode.

The solution, thanks to Proppy's programmer James Watson, M0DNS, has been to give you an online tool to produce your own customised charts. Just go to https://soundbytes.asia/proppy/planner

Proppy input form.

The first thing to do is move and drop the TX pin on the map to your location. This will automatically generate your latitude and longitude in the "TX Site" at the bottom of the page. Note: stations west of the Greenwich Meridian will be marked as negative. Then go down to the next line marked "load presets" and select whether you want "Worldwide", "Europe" or "RadCom 1, 2, or 3".

RadCom 1 is Asia, Australasia and the Middle East. RadCom 2 is Honolulu (short and long path), South America and Africa, and RadCom 3 is North America. You can, if you wish, add to these or put in your own locations. Press load and you will see the target locations come up in the form.

Next, move down to the plot selection box where you can choose between "Basic Circuit Reliability", "Signal to Noise Ratio" or "Signal Strength". Move to the next box where you can select the colour scheme. Then select the output format, choosing between PDF, PNG or SVG – the latter two are graphical file formats. We suggest selecting PDF.

Now, move down to the "System" box where you select the "Month", mode (in "Traffic"), noise level, and Smoothed Sunspot Number "SSN" source. If in doubt, select "Standard Curves". Finally, go to "TX Site", put your callsign in, check your latitude and longitude are correct and input your antenna gain. Then, go to the top of the form and press "Generate Plan".

Once the calculations are complete, you will be given a URL where your predictions can be found. Either save the file or print it out as you wish.

The charts show frequency on the left (Y axis) and time of day along the bottom (X axis), in four-hourly increments. The colours on the chart refer to the legend on the right and whether you selected "Basic Circuit Reliability", "Signal to Noise Ratio" or "Signal Strength". Obviously, you need to do three separate runs if you want to cover all of the locations listed on the RadCom charts. Let us know how you get on with the tool. Our thanks go to James Watson, M0DNS for all his work on this project.

Steve Nichols, G0KYA
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Feature

Ramblings of an 1950s Radio Engineer...

Amateur Radio has altered so much since the 1950s that it is interesting to look back and record some of the changes for posterity, before all of us oldies have gone SK. In the 1950s there was very little commercial equipment available for radio amateurs, mostly of USA origin, and too expensive and difficult to obtain. Everyone in the local club, the West Kent ARS in Tunbridge Wells, had transmitters built by themselves.

Picture 1: Place in the coal fire for 15 minutes before soldering.

Picture 2: A Home Brew Transmitter.

Picture 3: Typical Transmitter Wiring.

Many used CW on the HF bands, with plenty of contacts all over the world. At that time there were a large number of people who were very proficient in Morse code, which was the main method of radio transmission in the three Armed Forces. In many ways the requirement to pass the Morse test for an Amateur licence was good, as it introduced new blood to its capabilities and we found how enjoyable it was. The necessity to be able to receive Morse code was deemed important by the powers that were in case the amateur was causing interference with essential services. Something which greatly influenced those times was the availability of a great deal of ex-WD (War Department) equipment from WW2, mostly from Lisle Street in Soho, London, and we almost all used a receiver from there. I operated over the years with an R1224, an R1475, an R107 and the last one, best of all, an AR88. We also bought cheap units for stripping down to re-use many of the components – resistors, capacitors, valves, and transformers, for example. In the shack the receivers and transmitters were always separate, and when going from one to the other the antenna had to be switched over. I used an open knife-switch, as many of us did. Antennas were always homebrew – there were very few commercial items available. G3JRD came into existence in 1953. Fine – although we had no mains electricity, living as we did way out in the country on the Kent/Sussex borders. I later met someone else who, when first on the air, was in the same predicament in Kent. A little ex-WD 98 cc petrol generator had been obtained, and a 12V car battery, which had to be replaced in a car, was still good enough to stabilize the generator voltage. The receiver for the first year was a pre-war ex-WD R1224, needing a 2V 'accumulator' (a rechargeable portable lead-acid battery) for the heaters, a 9V bias battery and 120V DC high voltage supply. It was a remarkably good little set, with five valves, a tuned RF stage, mixer, a 465kHz IF, audio amp and PA. The positive feedback reaction control was used for CW, turned up to just below the oscillation point, which also peaked up the selectivity, and provided a very big increase in sensitivity. With practice and patience, it was as good as many of the modern 'all buttons and knobs' transceivers. When it was finally taken to the Amateur Radio Museum at Duxford Airfield many years later, no one had even heard of the type. My homebrew transmitter was for CW only, as we had to do a year of CW for the first year at a maximum of 10W before the Full licence was obtained. This was just a few months after I had worked for a while at the Rugby SW Transmitting Hall, putting HF transmitters of up to 70KW on the air! At that time the output power was specified as the PA DC input power (the voltage x current on the anode) – simpler than PEP. There were no Foundation or Intermediate licences. The petrol generator, sited in what had been a stable attached to the house, had to be running whenever I went on the air. Working someone in Wales who bemoaned the fact that he only had DC mains was rather subdued when he learned that I did not have any at all. One very big advantage of living in the country without mains was that there were no TV sets in the area as TVI was a big problem for many amateurs at that time. Perhaps typically British, some TV receivers even had an IF of around 14MHz, such a wonderful choice! The result was break-through which was the fault of the TV set, not the amateur, but imagine trying to convince an irate neighbour! Having no mains meant using hand tools only – a metal chassis had to be made, and holes cut in it for such things as valve holders. All the necessary holes were made labouriously with a hand drill. Soldering was with an old-fashioned blob of copper on the end of an iron rod (Picture 1), heated in a fire in the winter, or a paraffin Primus stove, which had to be lit first.

It was not very easy to get the iron to the right temperature. SWR meters were non-existent as far as I know, and tuning was an interesting exercise. First the antenna was disconnected and the tank circuit of the PA tuned for resonance, indicated by a large reduction in current. The 807 was quite capable of this treatment – a semiconductor PA would have blown up. Then the antenna was reconnected and various adjustments made to resonate at exactly the same frequency, but now with a very small dip at resonance. It was assumed that the RF was now going 'up the spout' successfully. Picture 2 is a typical home-brew valve transmitter, CW only, crystal controlled, using the very robust 807 valve in the PA. There was a lot of metal-bashing before starting the wiring. The meter measured the anode current to the PA. Valves needed a high voltage to operate, and the PA stage of a transmitter frequently used an 807-beam tetrode which might have up to 600V DC on its anode, though the specification was 400V. The anode would glow, but an 807 from the USA could take it.

A lot of care was necessary with such high voltage. If the crystal was a sufficiently active one, it was possible to change the frequency slightly with a small shunt variable capacitor, enough to separate the frequency from another transmission close by. We also used homebrew VFOs, but good stability was difficult to achieve unless care was taken to make all the

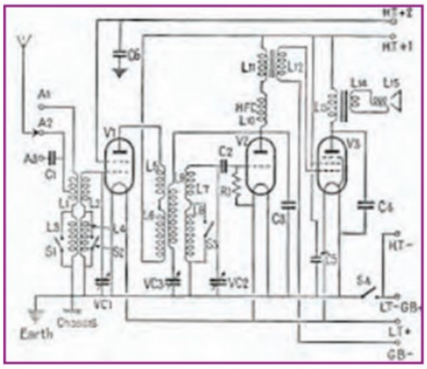

Fig 1: A typical circuit diagram.

construction very rigid, and the heat from the valves caused drift. Picture 3 shows typical transmitter wiring.

The PA valve holder in this case is an 'up market' ceramic one on the right-hand side. After all the work involved, it was exciting when it came to testing, and very satisfying if it worked, and one had a good QSO with a fellow amateur who had gone through the same process to get on the air. Reading circuit diagrams was a completely normal activity, and it only took a quick glance at one before it was known what sort of equipment it was. Without that ability it would have been impossible to assemble a rig and get on the air. (Fig 1)

A quick inspection shows that this is a tuned RF receiver (TRF), with three antenna inputs, an RF stage, detector, and PA, covering two bands. There were so many valve types, most with different configurations, that reference books listing hundreds of base connections were a must in the shack. This (Fig 2) is a small sample of one of the many pages.

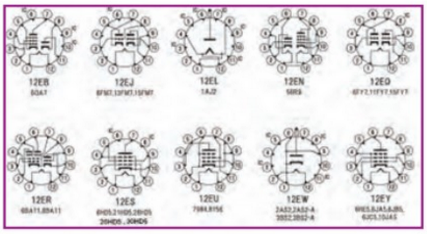

Fig 2: One page from the book of valve bases.

On a few occasions when there was no resistor of the correct value, a carbon resistor from a stripped-down ex-WD unit was found that had a lower resistance than the one that was required, and its side filed down until it was the right value. It was then varnished and used in the gear being built. If an ex-WD crystal was too low in frequency, it could be taken apart, the crystal ground on a piece of glass with car valve-grinding compound or jewellers' rouge, to bring its frequency up, cleaned with spirit, and hopefully it still worked, but at the required frequency. This might all seem a lot of hassle, but imagine anyone getting on the air in the really early days, in the 1920s. It was probably quite difficult to find a source of components, and there might have been a fair amount of travelling to find some of them. In order to get high DC voltages for the valves in the equipment from a 12V battery, the most efficient way was to use a mechanical vibrator (Picture 4) which were quite common in electronics in the 1940s and 50s.

Picture 4: The Mechanical Vibrator.

It chopped the DC up, to be fed into a step-up transformer and the output rectified and smoothed. The vibrators were not very long-lived, so were fitted with pins fitting into a holder similar to a valve base, so that they could be changed easily. Ex-service rotary converters were available, but their efficiency was only about 50%, so I did not use them much, taking too much of the limited power.

Working portable

Not many worked mobile (not many had cars!) but we did sometimes go portable, including on the highly popular Field Days organised by the RSGB. Picture 5, taken in an old car about 1955, using an R1475 receiver, ex-WD of course, and a rig built into two ex-WD chassis. Not a pretty sight! As it was not exactly "mobile", /P seemed appropriate; the card was left over when the West Kent ARS used it on a previous Field Day.

Picture 5: Working mobile from the 1955 car.

On returning to the UK after a couple of years working abroad, it was a new QTH in rural East Sussex, with mains supply (yippee!) and three quarters of an acre of garden for antennae. Heaven! SW bands were worked much as before, using about 40W and CW. As an encouragement to newcomers to our wonderful hobby, there are below two extracts of my logbook in 1981. These contacts were at 40W CW when using a homebrew valve transmitter and a homebrew dipole on the 21MHz band, when it was open. The receiver was an AR88.

It is interesting to note the length of the contacts – 8,14,19,10,16 and Australia for 37 minutes. We had good long QSOs on CW. When both of you were using straight Morse keys it felt surprisingly personal, almost like a verbal conversation, as everyone had slight differences in handling the key, possibly noticed sub-consciously.

Pictures 6 and 7: two extracts from my log books of the time.

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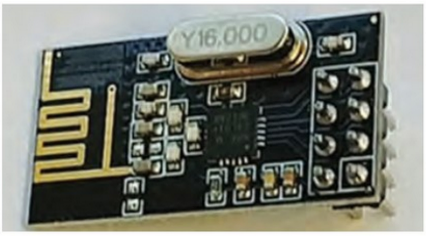
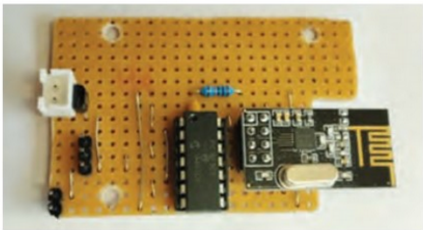
Wireless Morse Keys

Picture 1: Two Keys with transmitters fitted.

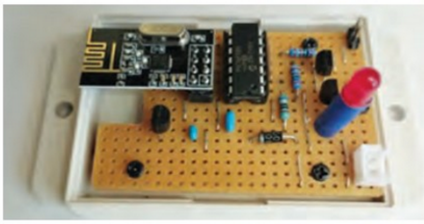


Picture 2: The nRF24L01.



Picture 3: nRF24L01 on the stripboard.

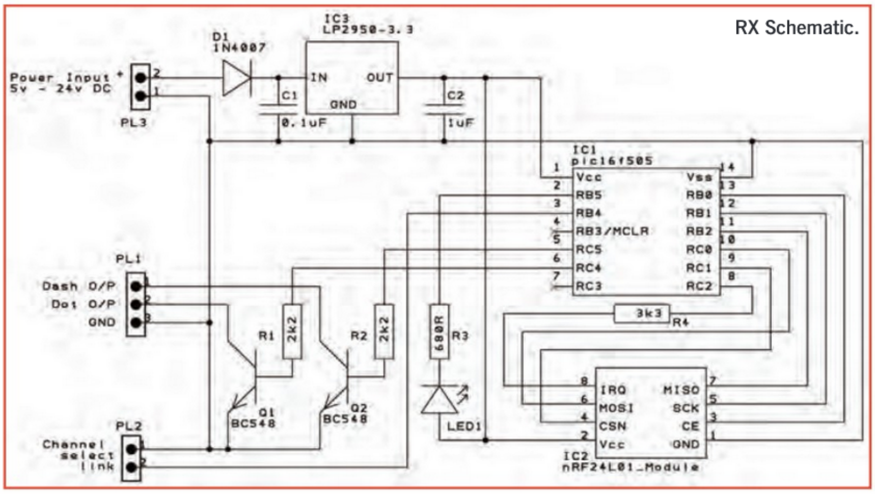
Picture 4: Iambic Key with Transmitter.



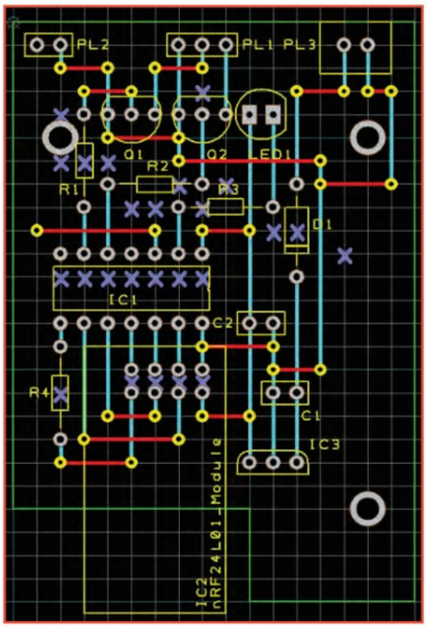
Picture 5: The Finished RX board.

I've for long been frustrated by having to swap connecting leads between Morse keys and transmitters. I thought it would be a good idea to have wireless connected keys, similar to a computer wireless mouse. At first, I thought of modifying a computer mouse but the complexity of the USB interface was rather daunting and there was no published information that I could find on exactly how they work. Then I stumbled across the nRF24L01 module (**Picture 2**) which is much loved by Arduino users and can be used to establish short range 2.4GHz data links.

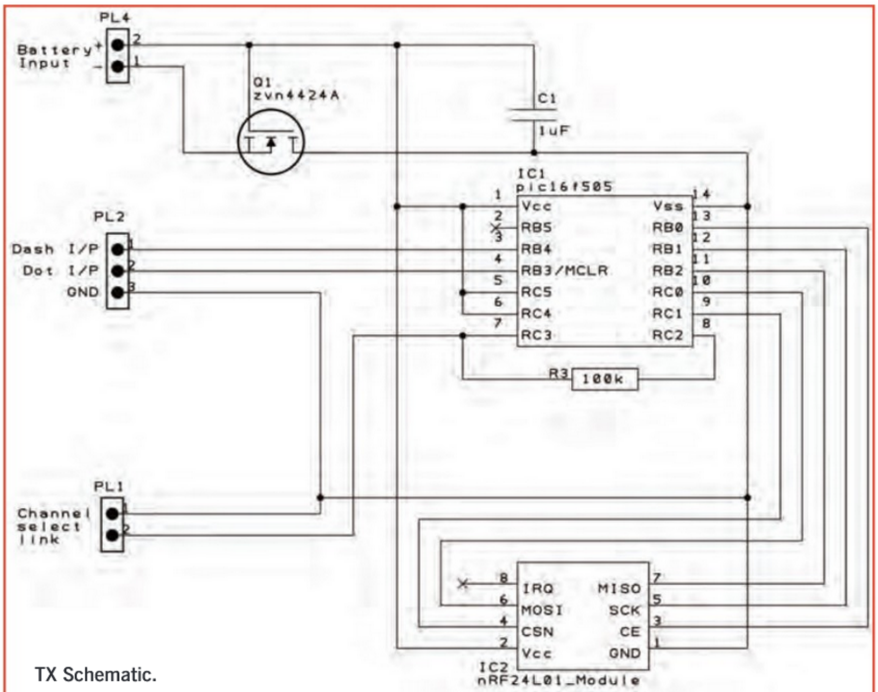
The modules are extremely cheap. To cut a long story short, I designed them into a module using a fairly modern PIC processor and a custom PCB which works very well. Rather than using direct cable connection, the keys are wirelessly connected to the station transmitters. (**Picture 1**). Each key has a battery-powered sender unit operating in the 2.4GHz licence-exempt band. Each station transmitter has a 2.4GHz receiver module connected to its key input. All units use the same frequency and protocol so that any key can operate any radio equipped with a receiver unit. Units are provided with a link option to operate on one of two frequencies in the 2.4GHz band (currently set to 2.479GHz and 2.482GHz). The units use modules based on the nRF24L01 transceiver chip made by Nordic Semiconductors. The modules are currently available at quite a low cost. Transmission is at 1Mb/s, using the nRF24L01 basic "Shock-Burst" message format. Each message simply transmits the state of the key input(s). If a dual paddle key is connected, the state of both the dot and dash paddles is transmitted. Messages are 48bits long and comprise preamble, address, payload and CRC fields. Whilst a key input is active, messages are continuously transmitted at a rate of two every 1.5ms. The transmitter output power is set at -6dBm (250μw), which gives an operating range of around 10m. Each key is fitted with a sender unit built on a small piece of stripboard. Control of the nRF24L01 module is provided by a



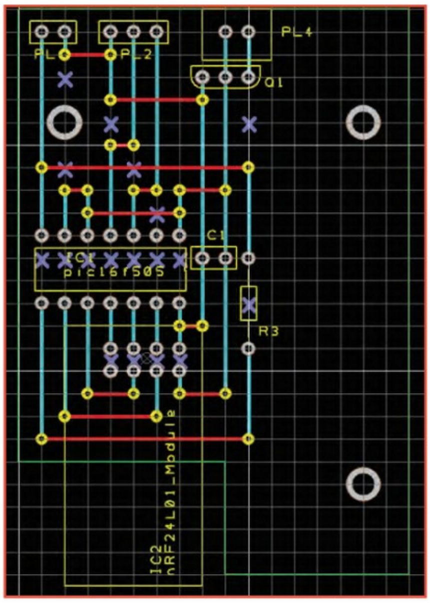
RX Schematic.



RX Layout.



TX Schematic.

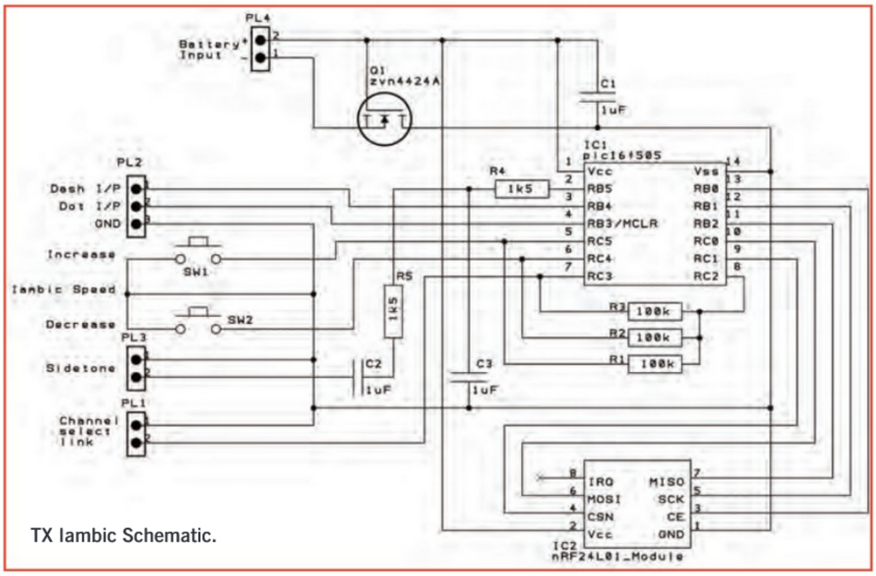


TX Layout.

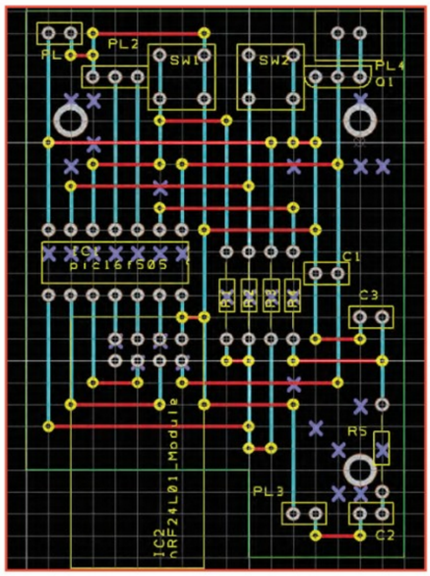
Microchip 16F505 microcontroller. The firmware for this chip has been written in "C" using the Microchip XC8 compiler. The unit is powered from an unregulated nominal 3V power supply (2 x AAA batteries). Average key down current is about 2.5mA. When there is no active key input, the unit goes into sleep mode, drawing about 0.7µA current, obviating the need for an on/off switch.

The Iambic keyer version is particularly useful with transmitters that don't have a built-in keyer function. A few extra components enable the iambic keyer function whilst using the same firmware in the PIC microcontroller as the basic unit. Two push buttons allow the keying speed to be stepped up and down through a range of 10wpm to 35wpm. Pressing both buttons simultaneously switches the unit between transparent mode (as the basic unit) and iambic keyer mode. A keyed tone output is also available for use with radios that don't have a sidetone output and for training purposes. The units are housed in standard flanged boxes from RS components.

Picture 4 shows the iambic keyer box mounted on a twin paddle key with the control buttons protruding through the top and the headphone monitor jack at the rear. The battery holder is fixed to the inside of the lid. The receiver unit (**Picture 5**) decodes and verifies the incoming messages. It has two open collector outputs to drive the radio transmitter's dot and dash key inputs. When used with a straight key, or the iambic keyer function, only the dot output is driven. The unit is powered via a 3V linear regulator which will accept a DC input voltage between 5V and 24V, allowing it to be powered from a DC supply or from the associated transmitter. There is also a monitor LED which illuminates when the unit is receiving a valid signal.



TX Iambic Schematic.



TX Iambic Layout.

Ray Scrivens, G3LNM