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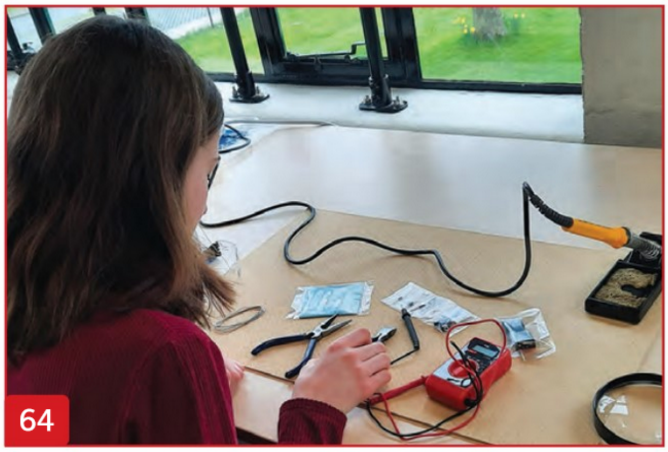

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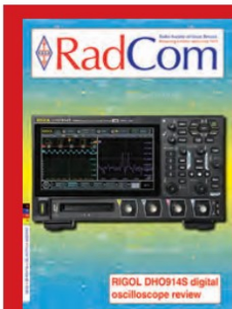

RIGOL DHO914S digital oscilloscope review
Cover design by Kevin Williams, M6CYB

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

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New Products

SDRplay introduces the RSPdx-R2

UK radio manufacturer SDRplay Ltd is introducing the RSPdx-R2 which is an enhanced version of its highly-popular multi-antenna port SDR, the RSPdx.

Jon Hudson, Sales and Marketing Director of SDRplay said "Global supply chain support issues have prompted some redesign of existing products to ensure continued supply for our Peterborough manufacturing partner. With each new member of the RSP, SDRplay tries to include performance enhancements. This has given us the opportunity to offer performance enhancements at the same time as assuring supply".

The RSPdx-R2 provides up to 10MHz spectrum visibility anywhere from 1kHZ to 2GHz with no gaps. It features:

- Upgrade to RSPdx: improved noise performance below 1MHz; improved dynamic range below 2MHz both in tuner mode and HDR mode
- Three software selectable inputs, including a BNC input for up to 200MHz
- 500kHz LPF for LF/VLF
- HDR mode for enhanced performance under 2MHz
- Notch filters on all inputs
- Rugged steel case

The suggested retail price for the RSP1dx-R2 is £225.60 including UK VAT. More details on SDRplay.com/RSP1dxR2

SDRplay recently launched its free multiplatform SDRconnect software which, as well as running on Windows, will also run on MacOS and Linux/Raspberry Pi. As with their SDRuno Windows software, the emphasis is on 'plug and play' making the SDRplay receivers a low-cost way to discover or rediscover the radio hobby for anyone who already uses a computer. There are even "band buttons" for all the amateur bands!

The UK manufactured RSP family of SDR receivers range in price from around £110 to £260 and are available directly from SDRplay Ltd (www.sdrplay.com), or from Martin Lynch & Sons, Moonraker, Nevada, Radioworld, SDR-Kits and Waters & Stanton.

Moonraker C8000 300kHz-50MHz Six Digit Frequency Counter

The Moonraker C8000 frequency counter is designed for measuring frequencies between 0.3-50MHz. The unit has 2 x SO239 connectors on the rear panel and also a very high-sensitivity RCA socket for reading very weak signal frequencies.

Specification

- Power supply required: 12-14V
- Power Rated: 1-50W
- Absorption: 150mA
- Frequency range: 0.3-50MHz
- Sensitivity better than 50mV
- Size: 125 x 170 x 35mm
- Weight: 0.86kg

Available at moonrakeronline.com for £99.95

New low-loss coaxial cables

The DX Shop Limited has launched two new high-performance coaxial cables with sensible pricing.

The new DX-L400 is a 10.3mm coaxial cable with low losses and high durability. The DX-L600 is a 15mm version of the same.

Both will stand repeated bending around rotators and with pump-up masts without loss of performance or mechanical failure so are suitable for fixed or portable installations.

Standard connectors are used so there is no need to invest in expensive, cable-specific terminations.

Supplied in 100m reels they are available from stock at The DX Shop Limited

Visit thedxshop.com for more information.

Antennas

If you live in a flat, or your shack is upstairs, and you choose to suspend an end-fed wire antenna between the building and a support in the garden, you are likely to experience EMC problems. This is because the braid of your coaxial cable is connected to both the antenna and the electric mains wiring via the radio.

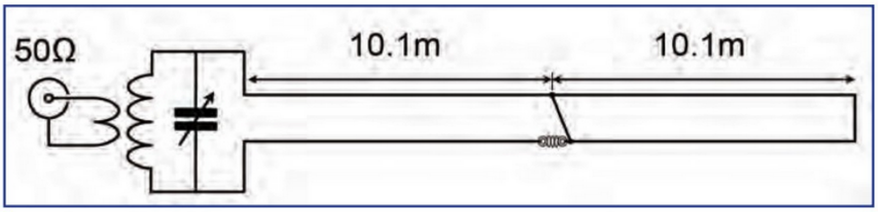

FIGURE 1: A diagram showing the end-fed dipole antenna with its essential isolating tuner.

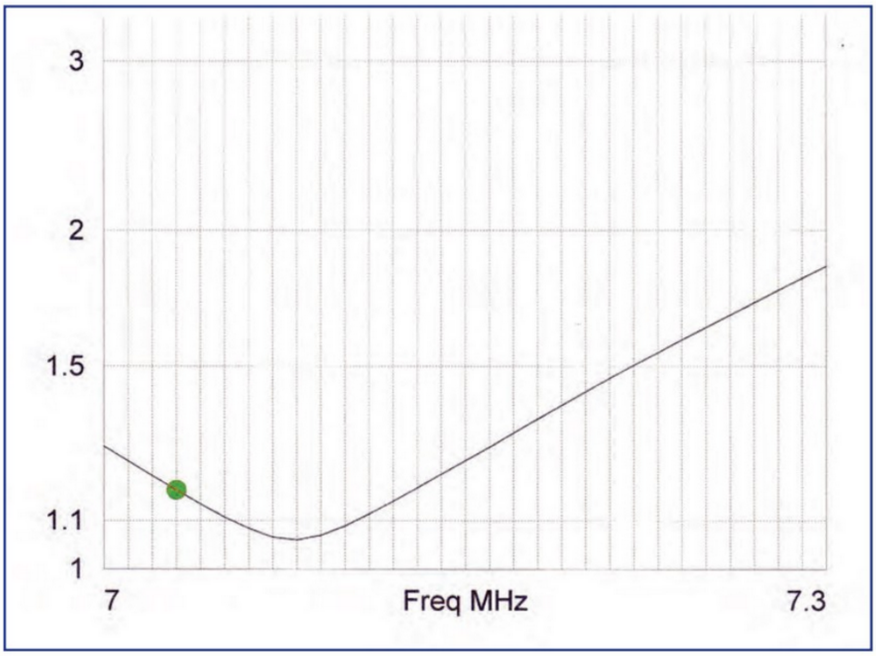

FIGURE 2: The computed VSWR (vertical scale) plotted against frequency over the 7MHz band, referred to an impedance of 1.5kΩ.

The problem

Although the braid of your coaxial cable may carry a low current at the matching transformer when the wire is a half-wave multiple, the current is likely to increase a quarter of a wavelength along the co-ax braid, just as it does in the antenna. Domestic wiring, and plumbing that is coupled or bonded to the mains earth wire will, therefore, support an RF current. This current, as well as wasting power, is likely to be effective in the rooms, above, below and beside you, and may expose your neighbours or your family and their electronic equipment to radio-frequency fields. Conversely, electrical interference from sources within the house may be introduced to the antenna via the braid of the coaxial cable. This reciprocity should always be bourn in mind.

An alternative

Here is an alternative end-feed technique that allows a *centre-fed* horizontal doublet to be driven from one end without a ground connection, and with its coaxial cable isolated from the antenna. It is a derivative of the Zepp antenna where, in this case, the traditional vertical section becomes horizontal. The potential advantages are that your radio will be less susceptible to noise carried into your receiver on the braid of the coaxial cable, you will reduce the RF exposure in and around the house, and you will have reduced sensitivity to local vertically-polarised interference. **Figure 1** shows the typical arrangement in which the left-hand half of a 7MHz dipole antenna is formed from a section of high-impedance twin-wire transmission line (450Ω ribbon cable). The other half of the antenna is a quarter-wavelength of line that has both legs joined at each end. This half could, alternatively, just be a quarter-wavelength of wire, but the twin wires have less tendency to twist when tensioned, and their lower impedance helps to increase the VSWR bandwidth, as shown in **Figure 2**. It also reduces the potential at the ends of the dipole. Providing that it can be tuned to resonance at the isolation transformer, the antenna need not be self-resonant. Dimensions may, therefore, be adjusted to fit the site.

How it works

When the transmission line half of the antenna is driven via the tuned transformer, the currents in each leg of the line are completely unbalanced because one leg is not connected to anything! Consequently, a large common mode current flows from the line to the assumed 70Ω resistance of the dipole. You will see from **Figure 3** that there are two currents in the line. The greater is the line's common-mode current that flows via the 70Ω feed resistance to the other half of the antenna. This current results in radiation. The other current is the balanced-mode non-radiating standing wave caused by the line's miss-termination. In this half-wave example, the impedance presented by the antenna at the transformer is a resistance of roughly twice the line's characteristic impedance at fundamental and third harmonic frequencies. This is 1.5kΩ when the line is formed from 15cm-spaced 1mm wires. To match this to 50Ω coaxial cable, the transformer requires a turns ratio of about 5. When the antenna is constructed from 450Ω ribbon cable, the impedance is about 900Ω and requires a 4-to-1 ratio. The advantage of ribbon cable is that it does not matter if it twists, whereas the open line will be short-circuited unless well tensioned to prevent twisting.

The tuned transformer

This type of tuner shown in **Figure 4** was once popular for matching the impedance of multi-band doublets like G5RVs to coaxial cable, when they were driven via high-impedance balanced lines. Either the high-impedance side with shunt capacitance, or the low-impedance side with a higher-value series capacitance, may be tuned to achieve sufficient magnetic coupling between windings, whilst isolating the coaxial cable's braid from the antenna. The antenna may be driven on other higher-frequency bands to which the transformer can be tuned. Taps on the inductor are then required to accommodate the various complex feed impedances on the 30m, 20m, 18m, 12m and 10m bands. The tuner design will depend on how you make the line section of the antenna, and may be developed experimentally. Suitable designs may be found in radio handbooks, and suitable tuners used to be available commercially, such as that from KW Electronics, which included provision for driving balanced lines. For 40m, I recommend starting with a 10-turn solenoid of bare copper wire of no less than 2mm diameter. Make it self-supporting using a 50mm diameter former, and with turns spaced 2mm. This will enable tap locations for minimum VSWR to be found experimentally before making fixed connections to relays or a band-switch. Turns of well-insulated wire are required at the centre of the solenoid for the 50Ω winding, because there are likely to be several hundreds of volts between the windings. A variable tuning capacitor, with wide plate spacing, is likely to be required, as determined by the power level, but start with low power and any available value that will tune through minimum VSWR.

Alternatively, use your VNA to find appropriate solenoid tap locations that result in a feed resistance of 50Ω. This can be a very satisfying exercise. When you know what is required, start searching for a suitable high-voltage capacitor. For example, with 400W and 1.5kΩ, the potential difference across the capacitor will be about 800V. By using relays for tap connections, and a variable capacitor driven by a stepper motor or a servo motor, it becomes practical to locate the tuner and antenna away from the house, and hence further reduce potential EMC problems. When operation just on 7MHz and 21MHz is sufficient, a simple remote tuner may have two capacitors that are switched by a single relay. A wideband 4:1 or 5:1 turns-ratio transformer, between the coaxial cable and the antenna, may be used in conjunction with an unbalanced tuner, but bear in mind that the end of a dipole has a high potential that will require appropriate insulation and minimal capacitance between windings.

A technique that permits looser physical coupling, by increasing the separation between the windings and hence incurring less inter-winding capacitance, is to use a series tuner on the 50Ω side, and a parallel tuner on the antenna side. The best location for the tuner, when the antenna is supported from a house, is immediately at the near end of the antenna, or just inside the window where the antenna is attached. Otherwise, some of the antenna must inevitably run through the house to the tuner. This will not be a very healthy arrangement because of the high EMF at the ends of a dipole. The near end may be located clear of the house when the associated tuner is remotely controlled, because the length of a coaxial cable with low VSWR is not critical.

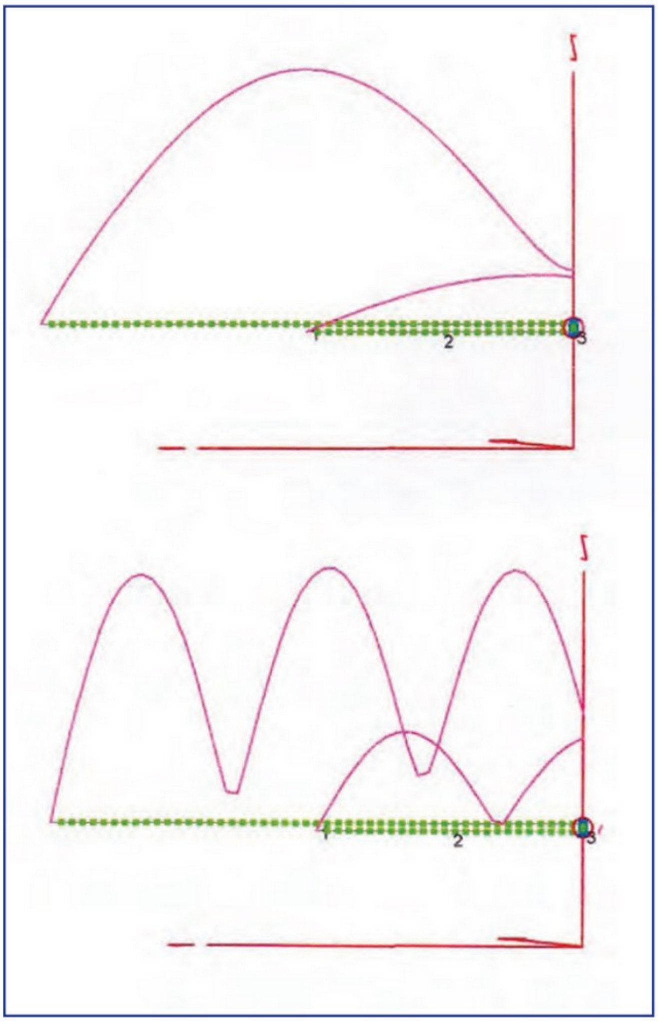

FIGURE 3: **Current distribution in the end-fed dipole: top at 7MHz, and bottom at 21MHz.**

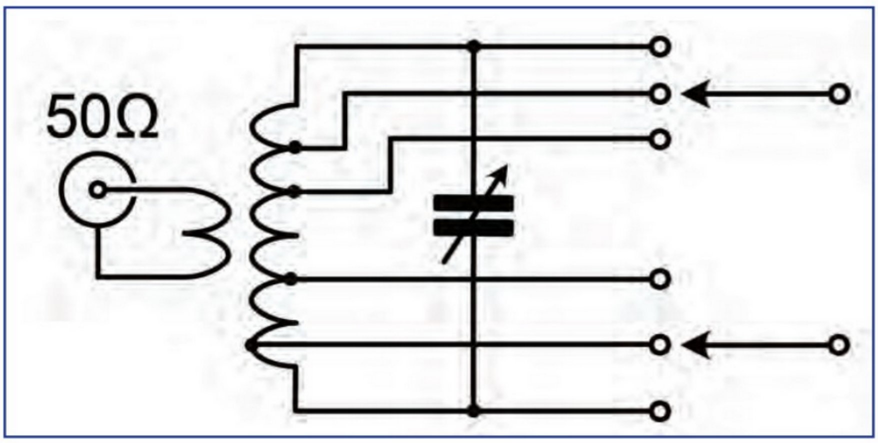

FIGURE 4: **The circuit diagram of a multi-band isolating tuner for the antenna.**

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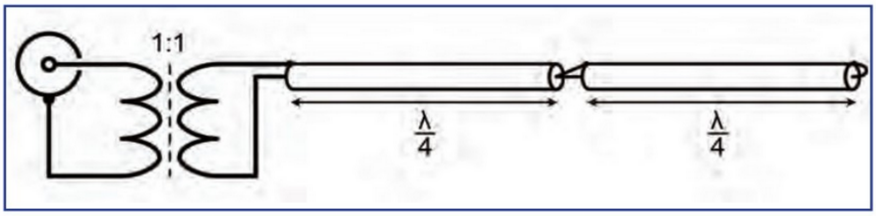

FIGURE 5: A suggested co-ax version of the end-fed dipole.

The radiation pattern

The radiation pattern in azimuth on any frequency is the same as that expected from a conventional centre-fed multi-band doublet. For example, there are two lobes at the second harmonic, six lobes at the third harmonic and four lobes at the fourth harmonic frequency. The pattern in elevation is that expected from a conventional horizontal antenna at the same height as the end-fed dipole.

Inverted-L

To fit a lower-frequency version within a restricted site, the end-fed dipole or doublet may be rigged as an inverted-L. Either all (as in the Zepp antenna), or a part of, the line section is rigged vertically, with the tuner at ground level. Because of its susceptibility to noise, it is preferable to have the vertical part remote from the house. A significant advantage over a conventional inverted-L antenna is the absence of both a radial ground and counterpoise system.

A co-ax variant

Although I have not tried this, it should be possible to replace the twin-wire line with coaxial cable as in **Figure 5**. The current on the surface of the braid replaces the common-mode current. The inner conductor of the line section is connected to both inner and outer conductors of the other section. No connection is made to the braid of the line section at the centre of the antenna because its outer (radiating) surface is fed from its inner surface. Measure the coaxial cable over the braid without consideration of the inner velocity factor. An isolation transformer with turns near 1:1 is then required for a half-wave antenna. However, a current balun is unlikely to provide sufficient isolation when inserted at the end of a dipole. On non-resonant bands, attenuation by the co-ax will detract from radiation efficiency. You may not be able to apply much tension to a coaxial cable unless it has a steel core. This configuration may be appropriate for higher bands such as 10m and 6m where a relatively-short length is required.

Feedback

I was pleased to receive feedback about last month's Vee beam article from Mike, M0FCD who fortunately has the space for a Vee beam. He discovered that, by replacing the termination resistors with inductive reactance, the backward radiation was increased. I tried this in my model to find that an inductance of 3.5uH is the optimum value for 14 to 28MHz for about 2dB extra reverse gain. Apologies for the W that crept in to replace Ω in Table 2.

Contest Calendar June 2024

Ian Pawson, G0FCT

RSGB HF Events

Date	Event	Times (UTC)	Mode(s)	Band(s)	Exchange
Sat 1-Sun 2 Jun	NFD	1500-1500	CW	1.8-28	RST + SN
Mon 10 Jun	80m CC DATA	1900-2030	RTTY, PSK63	3.5	RST + SN
Mon 17 Jun	RSGB FT4 Contest	1900-2030	FT4	3.5-28	Report
Wed 19 Jun	80m CC CW	1900-2030	CW	3.5	RST + SN
Thu 27 Jun	80m CC SSB	1900-2030	SSB	3.5	RS + SN

RSGB VHF Events

Date	Event	Times (UTC)	Mode(s)	Band(s)	Exchange
Tue 4 Jun	144MHz FMAC	1800-1855	FM	144	RS + SN + Locator
Tue 4 Jun	144MHz UKAC	1900-2130	All	144	RS(T) + SN + Locator
Wed 5 Jun	144MHz FT8 AC (4 hour)	1700-2100	FT8	144	Report + 4-character Locator
Wed 5 Jun	144MHz FT8 AC (2 hour)	1900-2100	FT8	144	Report + 4-character Locator
Sun 9 Jun	2nd 144MHz Backpackers	0900-1300	All	144	RS(T) + SN + Locator
Tue 11 Jun	432MHz FMAC	1800-1855	FM	432	RS + SN + Locator
Tue 11 Jun	432MHz UKAC	1900-2130	All	432	RS(T) + SN + Locator
Wed 12 Jun	432MHz FT8 AC (4 hour)	1700-2100	FT8	432	Report + 4-character Locator
Wed 12 Jun	432MHz FT8 AC (2 hour)	1900-2100	FT8	432	Report + 4-character Locator
Thu 13 Jun	50MHz UKAC	1900-2130	All	50	RS(T) + SN + Locator
Sat 15-Sun 16 Jun	50MHz Trophy Contest	1400-1400	All	50	RS(T) + SN + Locator
Tue 18 Jun	1.3GHz UKAC	1900-2130	All	1.3G	RS(T) + SN + Locator
Thu 20 Jun	70MHz UKAC	1900-2130	All	70	RS(T) + SN + Locator
Sun 23 Jun	50MHz Contest CW	0900-1200	All	50	RS(T) + SN + Locator
Tue 25 Jun	SHF UKAC	1830-2130	All	2.3-10G	RS(T) + SN + Locator

Best of the Rest Events

Date	Event	Times (UTC)	Mode(s)	Band(s)	Exchange (Info)
Sat 4 May - Sun 4 Aug	UKSMG Summer Marathon	All	All	50	4-character Locator
Sat 1-Sun 2 Jun	ARRL International Digital Contest	1800-2359	DIGI (no RTTY)	1.8-50	4-character Locator
Sat 1-Sun 2 Jun	UKSMG Summer Contest	1300-1300	All	50	RS(T) + SN + Locator + Member Number
Sun 2 Jun	UKuG Low Band	0900-1500	All	1.3, 2.3, 3.4G	RS(T) + SN + Locator
Sat 8 - Sun 9 Jun	IARU ATV	1200-1800	TV	432 & up	P# + SN + 4-digit code + Locator
Sun 9 Jun	Practical Wireless 2m QRP	0900-1600	AM, FM, SSB, CW	144	RS(T) + SN + Locator (5W max)
Sat 15- Sun 16 Jun	All Asian DX	0000-2359	CW	1.8-28	RST + age
Sun 16 Jun	WAB 6m Phone	0800-1400	AM, FM, SSB	50	RS + SN + WAB square
Sun 30 Jun	UKuG 5.7/10GHz	0600-1800	All	5.7 & 10G	RS(T) + SN + Locator

For all the latest RSGB contest information and results, visit www.rsgbcc.org

Design Notes

A multiplicity of signal sources

Nowadays, thanks to the online market places and far-eastern-made modules, we are spoilt for choice when it comes to generating waveforms for use as a local oscillator, transmitter source, or anything else for which we might want an RF signal. For over a decade in this column, we have looked at various synthesizer designs, but the descriptions have been a bit spread out and needed some searching. So here we will take a look at and summarise the range of popular phase-locked-loop (PLL) modules that can be found. An internet search with any of the keywords such as 'synthesizer', 'PLL' 'signal source' will find them, especially if coupled with the keyword 'buy'!

Figure 1 shows a range of PLL synthesizer modules, all with internal voltage-controlled oscillators (VCO) for simplicity of use. Starting at the bottom right, we have probably the first one of these to appear as a ready-to-go module: the Analog Devices ADF4351.

ADF4351 and fractional-N synthesis

Fractional-N synthesis is a technique that allows frequencies to be generated on a grid much smaller than the PLL's comparison frequency. This in turn means that spurs (or 'spurii') some distance from the wanted signal are minimised, and so cause less interference if used in transmitters or receivers. This was the problem with the older first-generation integer-N synthesizers where, if we wanted to generate a range of frequencies in steps of 25kHz, the PLL had to have a comparison frequency of 25kHz, which in turn could lead to generation of spurs either side of the carrier at multiples of this. Fractional-N synthesis 'jitters' the divide ratios in a pseudo-random way while keeping a high comparison frequency. The rate of jitter is made fast enough to be (mostly) filtered out in the PLL loop filter, and can allow arbitrarily-small increments in frequency to be selected. Phase noise, and close-in spurs are usually a bit worse than with a traditional integer-N arrangement, but troublesome spurs, especially those further out, are removed.

In general, a first generation fractional-N synthesizer gives an output frequency equal given by

$$F_{OUT} = F_{COMP} * (N + F / D)$$

where F_{COMP} is the comparison frequency which is typically 10MHz or higher, usually equal to the reference input. N, F, and D are positive integers and, by making suitable choices, a ratio can usually be found to give a frequency very close or exactly equal to that wanted. The ADF4351 limits the range of F and D to a maximum of 4095, so some ingenuity and compromise in selecting these values may be required if some obscure frequency is needed.

The internal VCO of the ADF4351 can tune over the range 2.2GHz to 4.4GHz and, for lower frequencies, an output divider can be selected which divides the VCO frequency by an integer power of 2 up to 64, allowing this chip to generate (almost) any frequency within the range 35MHz to 4.4GHz.

Bottom left of Figure 1 is another ADF4351 module made by SV1SFN for the more discerning users. It is better constructed, with more attention paid to decoupling and layout, and as a consequence produces slightly reduced spurious emissions and phase noise. There is no on-board 3.3V regulator on this module, so the user has to supply their own, which also allows the selection of an exotic low-noise one if desired to reduce phase noise even further.

FIGURE 1: **A collection of PLL synthesizer modules to be found from many online sellers. Bottom right is the ADF4351 device for 35MHz to 4.4GHz. Then going anti-clockwise, the MAX2870 is broadly similar but covers the frequency coverage from 23MHz to 6GHz. Top left is the 'deluxe' ADF5355 than can generate frequencies in the range 54MHz to 13.6GHz at tiny frequency steps thanks to its double fractional-N architecture. Finally, at bottom left is another ADF4351 module, supplied by SV1AFN, that has an improved board layout for reduced phase noise and spurs, but does need an external voltage regulator.**

MAX2870

This fractional-N synthesizer device, shown top right in Figure 1, appeared in ready-to-go modules a few years after the ADF4351. It is broadly similar in its fractional-N register set, but has a VCO that goes up to 6GHz, and an output divider up to 128, allowing frequencies to be generated in the range 23.5MHz to 6GHz.

Recently, here at 'JNT labs, I was looking for a source that, when multiplied up, could generate a signal in the 24048MHz band. Divided by four, this come out as 6012MHz, just above the specification for the MAX2870. Undaunted, I just programmed a set of register values that would generate 6012MHz (and a bit) and looked to see what it could do. The MAX2870 appeared quite happy there, but I didn't try pushing it any higher. **Figure 2** shows a narrow-band spectral plot of this device generating at 6012.5MHz. The plot width is 200kHz and the PLL phase noise is clearly apparent from the 'hump' in the baseline, as well as low levels of close-in spurs which are a usual characteristic of fractional-N synthesis. The discrete spurs separated by a few tens of kHz were a puzzle. They are not always present if I use a different reference input, so warrant further investigation.

The deluxe version

Shown above in Figure 1 (along with a small controller board, see later) is the more-exotic and expensive ADF5355. This device has a VCO tuning 3.4GHz to 6.8GHz, an output divider that goes to 64, allowing a minimum frequency of 54MHz, and an output doubler taking the maximum output frequency up to 13.6GHz.

Of greater significance is the extended double fractional-N engine. This

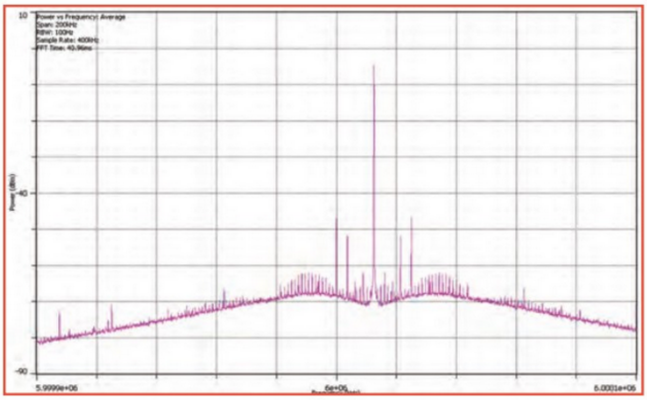

FIGURE 2: **The output spectrum of a signal generated by a MAX2870 device deliberately forced to go just above its specified frequency band, so that it can be used, after frequency multiplication, at 24.048GHz. The close-in phase noise, as shown here, will be increased by a factor of 20.log$_{10}$(4) or 12dB after the frequency multiplication, but should still be acceptable for a low-power personal beacon on that band.**

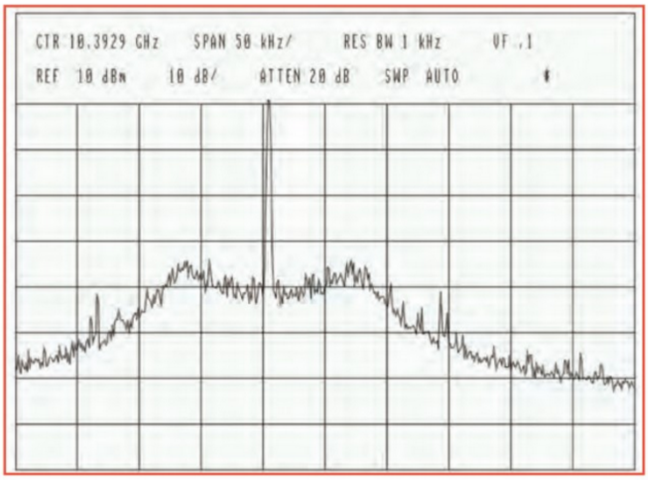

FIGURE 3: **The spectrum of the ADF4351 multiplied to 10GHz in a bread-board version aimed at a new-generation beacon source. Direct quadrature up-conversion followed by multiplication gives greater flexibility for generating MFSK data modes than offered by existing beacon sources.**

extends the old N+F/D ratio to now become $N + (F_1 + F_2/D_2) / 2^{24}$. F_1 can span the huge range from 0 to 2^{24}, while F_2 and D_2 are each limited to the range 0 to 16383 . This massive range of programmable values means that frequencies can be generated, at up to a few GHz or so, to a precision of milli- or even micro-Hertz. It comes at a price; this module typically sells for over £100, although price may have come down a bit recently, so is really only worth having if you really do need such frequency-setting precision. The output doubler leaks a large amount of half the wanted frequency. A source above 6.8GHz can probably be obtained better using an external frequency multiplier, rather than the built-in one. Programming the ADF5355 was covered in considerable detail in the December 2018 *Design Notes*.

10GHz source

When designing a new 10GHz beacon, I used an SV1AFN ADF4341 generating at 2592MHz as the frequency source. The output was passed through a base-band quadrature up-converter to generate the frequency modulation, then into a times-four multiplier to 10368GHz. The idea was to generate multiple frequency-shift keyed (MFSK) modulation, in this case the Q65-60D data mode, at a quarter of the wanted output frequency, and then multiply the whole lot. This beacon source is still in bread-board stage, and will be fully reported on later. **Figure 3** shows the 10GHz spectrum at the output of the times-four multiplier.

Synthesizer control

All the modules described here are supplied with no controller, and are programmed via a three-wire SPI interface. This appears on the header pins which carry serial data, clock, and load signals. Long ago, I adopted a universal generalised approach to programming SPI chips. A small PIC processor is programmed to accept serial RS232-type data sent as simple ASCII text; it then converts this to the three-wire SPI format. A complete register set, when finalised, can be blown into the PIC's non-volatile memory, so that the boot up is almost instantaneous next time. Slightly different PIC code is needed for different synthesizer chips, although the MAX2870 and ADF4351 share the same code. A typical register-setting command might be 'RDEAD1234' sent as nine ASCII characters followed by a carriage-return. The initial 'R' indicates that it is a register setting command, followed by 8 hex characters that represent the 32-bit data sent to the chip. All the chips here are organised so that the least-significant bits address a register number; the above command therefore sets register 4 to a hexadecimal value of 0xDEAD123. The use of simple ASCII text commands on a serial interface allows a terminal software package such as PUTTY to be used to programme registers individually in real time. Alternatively, custom register-setting and control software can be written to make control a lot easier. An example of this can be seen in the screenshot of **Figure 4**, controlling the MAX2870 chip. The register values shown in this screenshot, for generating the randomly-chosen output frequency of 28.415341MHz, were found by searching over all F and D numbers for the nearest solution, which here it managed to hit with an error of just 0.05Hz.

Full details of this approach to controlling synthesizer chips, and other SPI controlled devices, can be found at **[1]**. Many of these PLL synthesizers can now be found supplied with controllers already incorporated, often touch-screen ones. But whether these provide the full functionality of being able to set registers at will, I leave to you to investigate.

Measuring current at low voltage

Measuring the current consumption of something running from a low-voltage supply, especially if that current can vary over a wide range, can be quite difficult. This is especially so if the exact value of the supply voltage is important to operation. This issue arose with a piece of equipment designed to run from a single Li-ion cell. It monitored its own voltage supply to obtain the state of the battery charge and, depending on this, changed various operating parameters that affected current consumption. When voltage fell to 2.9V, it would go to sleep, just waking up periodically to measure the supply V_{DD}. During normal operation from 3.0V to 4.2V, the current consumption could vary over the range 50mA to 150mA, depending what was switched on, but in 'sleep mode' the current consumption was as low as 10uA. This would peak to around 1mA for a few tens of microseconds when it periodically woke up to measure the supply.

In trying to test and commission this, I adopted a circuit similar to that in **Figure 5**. Problems arise because a typical DVM on a current range drops 200mV across it at full scale. So, if a current of 50mA to 150mA were being measured, a DVM on its 200mA range would drop 0.05V to 0.15V. This is quite a significant portion of the total battery voltage, but more important is what happens when the various

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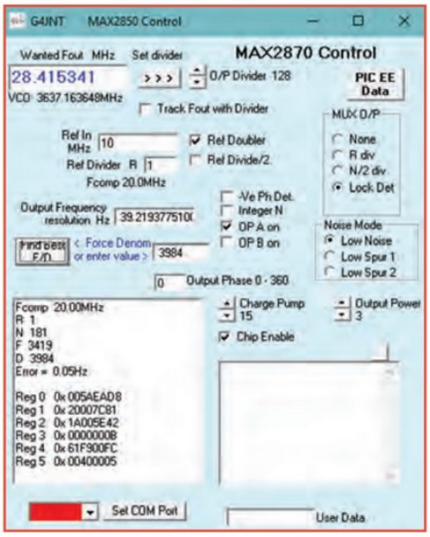


FIGURE 4: **Control software for generating register values for the MAX2870 device. This software communicates using RS232-type serial interfacing, sending ASCII text to the 'postage stamp sized' PIC controller on the RF module.**

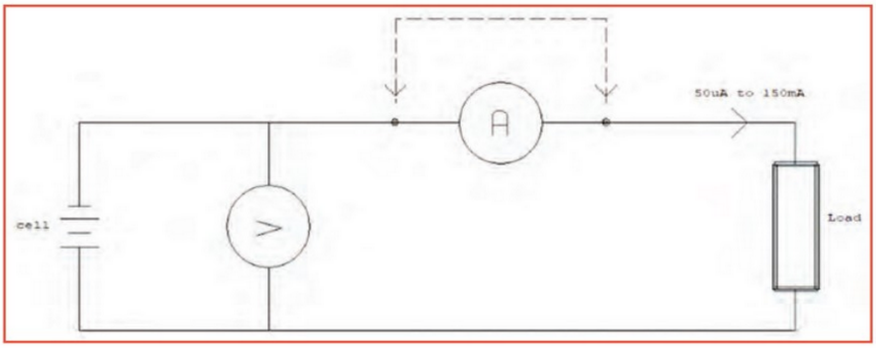


FIGURE 5: **Problems arise because of voltage dropped across the ammeter when trying to measure current consumed by a load that can vary over a wide range, especially when the supply voltage is low.**

switching actions are carried out. The varying load causes supply voltage jumps such that the whole system ties itself up in knots. This situation was just about ameliorated by using the DVM on its 2A range, where the resulting lower-resolution current measurement was acceptable and the voltage drop low enough not to matter. But it all went wrong when the current in 'sleep mode' needed measuring.

The current dropped to 10μA which couldn't even be detected on the 2A range, so the natural tendency was to switch to a lower range. This was a bad idea! The range switch in the DVM was break-before-make and briefly interrupted the supply, that then caused the processor to reset. The next solution was to select the 200μA range but use a jumper, shown dotted in Figure 5, to short-circuit the meter until sleep operation started. This worked fine, until it tried to measure its supply voltage. The short duration current increase to 1mA caused a brief additional voltage drop across the meter on the 200uA range that was enough to cause a PIC brown-out detection and close down. A large electrolytic capacitor across the processor supply solved this issue; the capacitor probably ought to have been there anyway. I eventually got the measurements needed and set things up properly, but the whole situation is all a bit of a mess. This problem must arise all over the place, especially in industry with testing modern electronic devices. Posting about this current measurement issue on the RSGB Technical Group resulted in a few responses.

Kevin, G7BCS wrote: "Generally, I use power supplies that have the right measuring capabilities to measure this sort of thing. In a previous life that was mobile phones, which can dart between μA on standby and a couple of A with the radio on and CPU going full chat. Capturing that sort of dynamic range is not easy. The best solution I found was the Keysight N6705 with the right plug-in module, but clearly this isn't available for an amateur budget! The principle is that it's measuring across a shunt that's inside the correction loop of the power supply; however, and if you've got a power supply with separate source and sense terminals, you might be able to put the multi-meter inside the loop?"

Malcolm, G8MCA, had another solution: "I have encountered this problem and similar and my solution was to build instruments to cope which also gave me another project. My usual problem is that a circuit will take 2μA to 3μA or so in sleep mode, but then 100mA when it wakes and sends data, and a typical ammeter cannot cope with that range. I have a log ammeter I built which is single range and measures 1μA to 100mA using the log relationship of a Schottky diode. It works well but drops 0.3V which you have to remember. I built another where the highest sensitivity is 40μA with 0.01μA resolution, and it uses a 1NA211 with 20Ω so it drops 0.2mV at 10μA. Both ammeters output data to a serial line as well which helps testing greatly. Both use the PIC16F1783 microcontroller."

And from Jim, M0BMU: "I expect the answer in an industrial situation would be to use a multi-meter with lots of digits; 5½ or 6½ digits are quite affordable in a professional context these days. Then you could set the meter to a high current range and still have reasonable resolution at the low current end. There are also many ICs available as high-side current-sense amplifiers, some of which probably have a low enough input offset to do this job.

"But assuming you want to use a basic 3½ digit meter, you would need to use a low-value shunt resistor at the high current end, say 100 milliohms to keep the voltage drop in the 10s of millivolt range. This would give you 1mA resolution with 200mV meter span, which I guess would be adequate for measuring a 100mA load. But that would be unhelpful for measuring the low current, for which you would need 100Ω shunt to get 1μA resolution. If an active/sleep signal is accessible in your circuit under test, you could have a dynamic arrangement with high-current and low-current shunts, and bypass the low-current shunt dynamically when the circuit is in the active state with a PMOS FET with a low $Rds_{(ON)}$, driven from the active/sleep signal; there are plenty of inexpensive ones with ON resistance in the <10mΩ range."

Measuring V_{DD} in a PIC

To measure the supply voltage using an analogue to digital converter (ADC) with its own accurate voltage reference, V_{REF}, the seemingly obvious way would be to use a pair of potential divider resistors to reduce the supply voltage to some value below this that the A/D can properly measure, then compensate the divider ratio in software. But modern PIC devices (and probably many other processors) have a facility built-in to make this easier. By reprogramming registers, the ADC configuration is moved around to use the power supply V_{DD} input as its reference. One channel of the ADC is connected to the fixed reference, V_{REF}, and measures this. Now a sort of upside-down operation is obtained, with the unknown V_{DD} being used to measure a known reference. The value of V_{DD} can be obtained from the equation $V_{DD} = V_{REF} \times 1024 / N$, on the assumption that we have a 10-bit ADC delivering a value N in the range 0-1023. When the ADC is needed subsequently to measure other parameters, the registers are reprogrammed to use V_{REF} as the ADC reference as normal. By this technique, two resistors and an I/O pin are saved, as well as the continuous current drain through the two potential divider resistors permanently across the supply.

Reference

[1] http://www.g4jnt.com/Synthesizers.htm

EMC

The item in the April 2024 *EMC Column* about natural and man-made radio noise levels has created some interest. What do these curves mean? How were they derived from the ITU-R curves? How can radio amateurs measure background noise in practice?

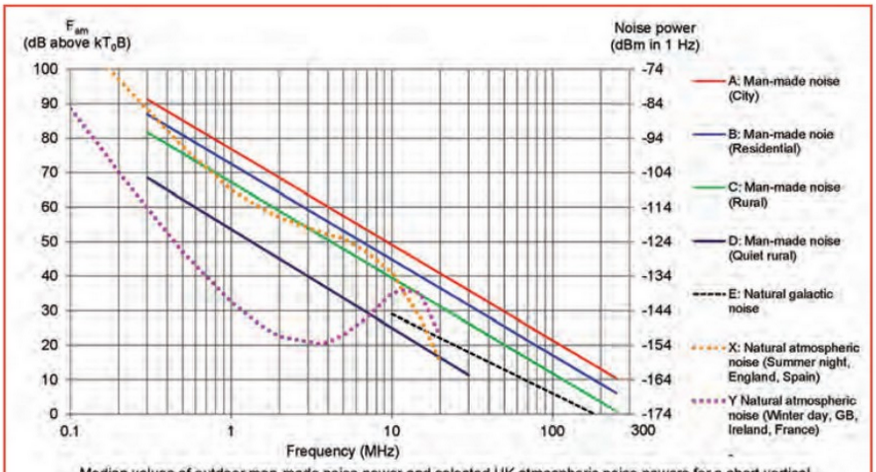
Here is a follow-up with some background information about background noise!

The starting point is International Telecommunications Union ITU-R Publication P.372-16 'Radio Noise'. This can be downloaded free of charge from ITU **[1]**. Compared to previous editions, version 16 has improved the presentation of figures, and many have been redrawn in colour. Radio noise levels are specified in terms of external noise figure Fa(dB), relative to thermal noise. But what does this mean in practice?

A resistor generates noise due to random motion of electrons unless it is cooled to absolute zero when all noise disappears. Noise power depends on absolute temperature in kelvin and the bandwidth of the measuring receiver. If the absolute temperature is 290K (17°C), and if the measuring bandwidth is 1Hz, then the noise power is -174dBm, that is -174dB relative to 1mW. This may also be written as -174dBm/Hz. This represents a fundamental limit to the maximum sensitivity that any radio receiving system can achieve.

If the temperature of a resistor is increased, then the noise power also increases. The resistor could be a physical resistor, or it could be the radiation resistance of an antenna. Radiation resistance is what a transmitter 'sees' when it transmits RF power into an antenna that is well matched and free of losses. The power from the transmitter appears to be going into a resistor but it isn't dissipated as heat, instead it is radiated as a radio wave into space. A half wavelength dipole has a radiation resistance of around 72Ω at the feed point. When you receive a signal from an antenna, it appears to come from the radiation resistance. The antenna picks up natural and man-made sources and this is sometimes called 'sky noise'. The Fa figures in P.372.16 tell us the sky noise power from the antenna in dB above -174dBm.

Figure 39 in P.372-16 is of particular interest to radio amateurs, as it shows expected man-made noise levels from 300kHz to 250MHz. These 'curves' are actually straight lines if the vertical scale is in dB and if the frequency scale is logarithmic. Table 1 in P.372-16 allows you to find the man-made noise levels at any frequency from 300kHz to 250MHz. I have put the figures from Table 1 into an Excel spreadsheet, which created the curves in **Figure 1** above. These are the same as the curves in Fig. 39 in P. 372-16 and I have added of a second axis in dBm and two natural atmospheric noise curves.



Median values of outdoor man-made noise power and selected UK atmospheric noise powers for a short vertical lossless grounded monopole antenna (Derived from Rec. ITU-R P.372-16 Table 1, Figures 39,16a,16b, 30a, 30b)

FIGURE 1: Man-made noise levels from ITU-R P.372-16

Peter, GM3XJE, has written more about this in his article 'Noise floor and all that' on pages 50 and 51 of this issue of *RadCom*.

Environmental categories

There are four environmental categories for outdoor man-made noise, A: City, B: Residential, C: Rural and D: Quiet Rural. You need to be well away from built-up areas to find the quiet rural level but in residential areas of the UK it is often possible to find noise levels down to the rural curve at certain times of the day and on certain frequencies. This is possible if all nearby electronic products comply with the relevant EMC standards and there is no VDSL noise.

It is of course also necessary to consider natural atmospheric noise levels which may be higher than the man-made noise curve for a particular environmental category. Figure 2 of P.372-16 shows the atmospheric noise level exceeded 0.5% of the time. It is important to note that this curve applies to the maximum level that is unlikely to be exceeded anywhere in the world including tropical latitudes where there are high radio noise levels due to thunderstorms. The median natural atmospheric noise levels found in UK and Europe are very much lower and to find these we need to look at maps of radio noise levels.

For example, the map in Fig 30a of P.372.16 shows that on summer nights in June, July and August from 8pm to 12am local time, Fa at 1MHz is around 60dB in Scotland and Ireland, 65dB in England and Spain, and 70-75dB in France and Germany. To find Fa at other frequencies we need to look at Fig 30b. For Fa = 65dB at 1MHz, we need to go half way between the 60dB and 70dB lines and this gives Fa = 40dB at 10MHz. The curve for England for frequencies up to 20MHz is shown as the orange curve 'X' in Figure 1. Atmospheric noise is slightly above the rural man-made noise curve below 700kHz and at 3-10MHz. Noise levels in Scotland and Ireland are about 1.7dB lower than England at 10MHz. They are about 1.7-3.3dB higher in France and Germany. Similarly, for winter day time, we can use P.372-16 Figs 16a and 16b for Dec, Jan and Feb at 12pm to 4pm local time. This gives the purple curve 'Y' for England, Scotland, Wales, Ireland and France. It is below the man-made rural curve at all frequencies.

Noise measurement

If we know the expected noise levels, how can we measure the actual noise levels at an amateur radio station in practice? In principle, you can measure the RMS noise power coming out of the antenna and convert it to the equivalent in 1Hz bandwidth, using the noise power bandwidth of the measuring receiver which is generally different from the -3dB or -6dB bandwidth. This gives a


PHOTO 1: **RF interference filters fitted to a solar PV installations.**


PHOTO 2: **Clip-on ferrite cores fitted to a solar PV installations.**


PHOTO 3: **Ferrite ring and clip-on ferrite cores fitted to a solar PV installations.**

result that can be compared to the ITU-R P.372-16 levels. In practice this works well at UHF and microwave frequencies but there are some practical difficulties for measuring man-made noise from 300kHz-250MHz.

First, the P.372-16 noise curves are measured with a short vertical lossless grounded monopole antenna. If you measure with a horizontally-polarised antenna, you will measure less noise at HF for various reasons (maybe 5dB less) so the measured result will not be comparable to the P.372-16 levels. Any practical HF antenna may have losses in the matching network (if any) and it also has ground losses so the feed point impedance contains both radiation resistance and loss resistance.

There is another way to measure background noise levels which is to convert the Fa curves in P.372-16 Fig 39 and Figure 1 to RMS noise field strength using Equation (7) In P.372-16, This is what I did in the April 2024 *EMC Column*. There are some important differences: the Fa curves slope down at -27.7dB per decade whereas the RMS noise field strength curves slope down at -7.7 dB per decade; the RMS noise field strength curves allow measured noise field strength to be compared to the radiated limits in EMC standards which are defined in terms of RF field strength; it is possible to use an active antenna with the RMS noise field strength curves, as long as we know the antenna factor. That is how the antenna output is related to field strength. This is not the same as antenna gain.

Solar PV VHF noise

RF interference from solar PV inverters continues to be an important EMC matter, particularly at HF, but there are also some VHF cases. Most cases relate to hybrid inverters which have a large storage battery, and most are made by two manufacturers who will remain anonymous for the time being. Ofcom has been involved in a number of cases and attempts have been made to apply mitigating measures to reduce the RFI from some installations. A Member has sent some photos, SDR plots and signal reports which make a very useful solar PV EMC case study. The object of the exercise was to reduce conducted RF interference that gets out of the inverter via the DC power ports, the wiring to the solar panels and the wiring to the storage battery.

Photo 1 shows two Schaffner FN2200-25-33 filters fitted in a box with the cover removed. These filters are rated at 25A and 1.2kV DC and they are suitable for filtering the DC output from solar PV strings. They are available from various electronic component distributors such as Farnell UK at £77.49+VAT each. **Photo 2** shows clip-on ferrite cores on the DC wires to the solar panels and **Photo 3** shows a ferrite ring and clip-on ferrite cores on the DC wires to the battery.

Here are the before and after results provided by our Member:

8MHz - Before S3 - After (trace);

10MHz - Before S9+ every 30kHz - After S7 every 60kHz (trace 30kHz intervals between) during daytime and trace at Night time;

14MHz - Before S9+ every 30kHz - After S8 every 30kHz (day time);

18MHz - Before S9 every 30kHz - After (trace);

21MHz - Before S3-4 every 30kHz - After S3-4 every 30kHz;

24MHz - Before S5-7 every 30kHz - After S1 every 30kHz;

28MHz - Before S6-7 every 30kHz - After S4 every 30kHz (night time) and trace every 30kHz (daytime)

It can be seen that there is some improvement at some frequencies, but this was not sufficient. Here are some comments about the installation:

- The filter box is mounted fairly close to the inverter which is good but any RFI filter is only effective if it has a good low-impedance earth connection back to the source of interference, in this case the inverter. There is a green and yellow earth wire but this goes away from the inverter, possibly to an earth terminal somewhere. This would be standard wiring practice for electrical installations, and it would be effective at 50Hz, but it is far too long to be effective at radio frequencies up to 30MHz where the wavelength is 10m.
- The filter earth wire also takes an indirect route back to the inverter, forming a loop that may radiate RFI. It should be replaced by a short, direct low-impedance earth connection back to the chassis of the inverter that follows the route of the DC power wires as closely as possible to minimise the area of the loop. A short direct wire, maybe 30cm long, to any available earth connection on the inverter should give some improvement but a wide metal strip or plate would be more effective.
- The filters are designed to be effective up to 30MHz and the clip-on ferrite cores on the black wires would provide some benefit at VHF, above 30MHz, by reducing RFI radiated directly from the wires between the inverter and the filter.
- The clip-on ferrite cores, and ferrite ring on the orange battery wires, are only likely to be effective at VHF and they are unlikely to have much effect below 30MHz because there are only two turns on the ferrite ring and only one 'turn' on each ferrite core.

The conclusion is that retrofitting high-performance RF interference filters, and adding multiple ferrite cores, gave insufficient improvement below 30MHz. The long filter earth wire is likely to be a factor, but the RF interference may also be getting out via other routes. In this particular installation, the inverter is being changed and it remains to be seen what improvement this makes.

References

[1] ITU-R Recommendation P.372-16: https://www.itu.int/rec/R-REC-P.372/en

[2] RSGB EMC Leaflets - EMC Leaflet 16: Background Noise on the HF Bands: https://rsgb.org/main/technical/emc/emc-publications-and-leaflets/

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Review

RIGOL DHO914S four-channel 125MHz digital oscilloscope

The DHO914S provides a great deal of functionality in a relatively-small enclosure, with low-power consumption, a multi-colour touch-sensitive display, and a clear user interface, for a retail price of £711.60 including VAT (February 2024) and a three-year warranty on the main frame.

FIGURE 1: The front panel.

Packed with functions

In addition to displaying and analysing analogue waveforms, this instrument also functions as a logic analyser, an arbitrary-function generator (AFG), a protocol analyser, and Bode plotter. There is also a fast Fourier transform (FFT) function to display the frequency spectrum of a waveform being viewed. See **Table 1**. The well-designed user interface enabled me to use many of the analogue functions without recourse to the user manual. A number of English pdf documents, and a software update, can be downloaded from the Telonic website **[1]**.

The front panel, shown in **Figure 1**, contains the touch-sensitive display, a number of clearly-labelled controls, the 50-way digital input connector, four coloured BNC analogue input sockets, the illuminated power on-off button, a USB-A connector, and the two probe-calibration terminals. The USB-A connector may be used with a FAT32 or NTFS memory stick on which to save screen displays, or a mouse to control the instrument via on-screen controls. The internal software also supports a USB expander so that a memory stick and a mouse may be concurrently connected, but it does not support an external keyboard.

The touch-sensitive display supports screen operations such as tap, drag and 'pinch & stretch', has multiple colours, and can display menus and sub-menus together with a virtual keyboard when required for data entry.

FIGURE 2: The rear panel.

The rear panel, shown in **Figure 2**, contains a small number of clearly-labelled connectors, a ground connection, a slot for a security retainer, ventilation holes for the internal forced-air cooling system, and two adjustable support legs, which allow the instrument to be oriented vertically or tilted back by approximately 22°. If required, the instrument may be attached via four tapped holes to a fixed mounting bracket, or a swinging arm above the work bench.

The DHO914S uses a Cortex-A72 6-core processor with 4GB of system memory, 8GB of non-volatile memory, and runs the Android operating system. It employs 12-bit 1.25GSa/s digital signal processing, with a frequency response of DC to 125MHz.

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FIGURE 3: PVP3150 passive probe.

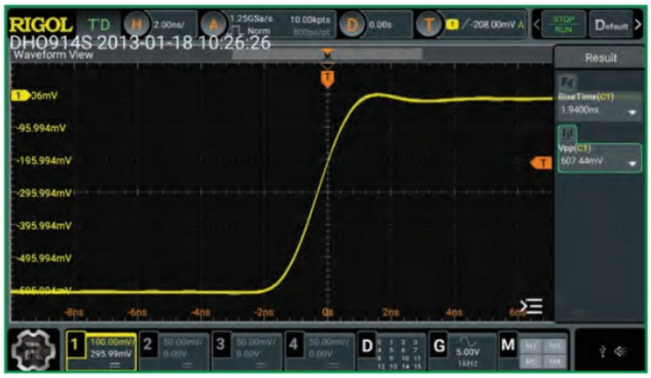

FIGURE 4: The DHO914 rise time is 1.94nS (provided by Telonic).

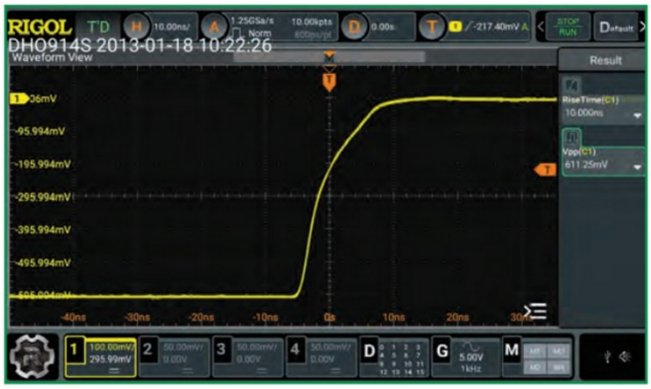

FIGURE 5: The probe rise time in the 1:1 position is 10ns (provided by Telonic).

FIGURE 6: The probe rise time in the 10:1 position is 1.85ns (provided by Telonic).

(Compare this with the 8-bit 1 GSa/s 100MHz processing used in the earlier DS1102 series.) An alternative instrument, the DHO924, provides a maximum input frequency of 250MHz, and is supplied with higher-frequency passive probes. A suffix 'S' indicates that the AFG is present as in the review model.

The horizontal time base has a range of 2ns/div to 500s/div with a resolution of 100ps and an accuracy of ±25ppm ±5ppm/year.

The instrument may be controlled remotely by connecting a PC to the USB-B or LAN interfaces, and using the standard commands for programmable instruments (SCPI) or using web control. Additional programming languages include Visual C++, Visual Basic, Python and Labview. RIGOL recommends the Ultra Sigma software which may be downloaded from **[1]**. This instrument also provides internal measuring and mathematical routines for the analogue inputs, and protocol analysis routines for the digital inputs.

The probes

Each of the four supplied passive probes have a pointer tip with a push-on cover, containing a spring-loaded metal hook, and a switch to select 1:1 or 10:1 attenuation (see **Figure 3**). A small access hole is present to adjust the frequency-compensation trimmer with the supplied trimming tool using the square wave signal from the calibration terminals on the front panel. The probe pouch also contains some spare indicator rings matching the four input channel colours. These are useful in identifying the channel to which the probe is connected. The frequency response of each probe is specified as DC-20MHz in the 1:1 position, with a rise time of 17.5nS. The frequency response in the 10:1 position is DC-150MHz, with a rise time of 2.3nS. I was unable to measure the rise time of the oscilloscope or probes myself as I did not have a sufficiently-fast pulse generator, so Telonic kindly provided the information in graphical form in **Figure 4**, **Figure 5**, and **Figure 6**.The user manual (section 4.3.7) explains how to adjust the 10:1 compensation trimmer. The adjustment of the probe compensation trimmer in the 10:1 position was a little coarse and there was a slight difficulty in getting an exact square wave.

In a similar manner to other (non Rigol) passive probes, the centre conductor of the probe screened lead has a significant resistance, in this case approximately 221Ω, between the probe tip and the BNC connector centre pin with 1:1 selected. This is to reduce the effects of resonances and reflections of very fast pulses **[2]**.

The PVP3150 passive probes are specified as having a standalone capacitance of 50pF +/-10pF in the 1:1 position and 10pF +/-5pF in the 10:1 position. Depending on the type of circuit being measured, it may be better to set the probe to the 10:1 position, thus reducing the capacitive loading on the circuit. Users must bear these values in mind when using the probes to make measurements in high-impedance or high-frequency circuits.

On the test bench

After attaching the power supply leads, a DC resistance of approximately 1Ω was measured between the mains-plug earth pin and the front-panel BNC connector bodies. The supplied earth lead may be employed between the rear panel connector and a local earth. Following connection to a mains supply, the on-off switch was illuminated in orange. On depressing the on-off switch, the illumination changed to green and the initialisation process then took a little over 50s, after which the default button illumination and operational screen were displayed, with a single straight yellow signal line present for channel one; no probes were connected at this time.

The internal cooling fan was fairly quiet and ran all the time that the instrument was switched on. Power was supplied by a mains power adaptor connected to the rear USB-C connector, and registered 12V at 4A, or 48W. You can substitute a suitable external battery plus connecting lead (not supplied) if desired.

FIGURE 7: The default start-up screen.

FIGURE 8: The horizontal measurement screen.

FIGURE 9: The vertical measurement screen.

FIGURE 10: Menu and sub-menu example, showing the AFG sub menu.

The user interface

The user interacts with the DHO914S via a series of push buttons and rotary controls on the right-hand side of the front panel, and a series of touch controls on the display. The functions of some of the rotary controls depend upon which mode of operation is currently selected, and indicator lights illuminate related controls for that mode. **Figure 7** shows the default display when initialisation has been completed.

Pressing the 'Measure' control button, for example, produces a sub-menu of horizontal control settings as shown in **Figure 8**, and pressing the 'Vertical' control on the lower left-hand side of the sub-menu displays the vertical settings as in Figure 9. The three buttons labelled 'Trigger', 'Acquisition' and 'Measure' in Figure 9 will take you back up one menu level. **Figure 10** shows the AFG sub-menu with a virtual keyboard for entry of numeric values and units. There are too many other menus to show in this review but they are explained in the user manual.

All screen displays in this review were made using the 'Quick' button to copy the screen to a .png file on the memory stick plugged into the USB-A socket on the front panel. The filenames take the form RigolDS0.png with the '0' auto-incrementing on each save to prevent filenames being duplicated.

Frequency response

Three frequency-response tests were carried out using a signal well above the minimum sensitivity of channel 1, and the results are shown in **Table 2**. In every case, millivolt measurements were recorded from the oscilloscope screen after pressing the 'Analyse' button and selecting 'DVM' and 'Counter' with a constant input-level signal. The dB measurements were calculated from the DVM display. All amplitude measurements were made using an un-modulated sine-wave signal.

In the first test, an HP8640B signal generator was connected directly to the oscilloscope via a 300mm length of 50Ω coaxial cable, with a BNC T-piece containing a 50Ω load immediately next to the channel 1 input socket to ensure that the coaxial cable was correctly terminated. At 1MHz, the DVM readings closely agreed with the 'pd' output setting of the signal generator. Most professional RF signal generators have their output amplitude calibration labelled either as 'emf' or 'pd', where 'emf' is the output amplitude into an open circuit load, and 'pd' is the output amplitude into a load impedance equal to the impedance of the signal generator, normally 50Ω. Therefore, the 'emf' output is twice the 'pd' (correctly terminated) output. This process is used because the input impedance of a unit under test, say a receiver, may not be exactly 50Ω making the 'pd' value uncertain, so test specifications can stipulate that the 'emf' value is used instead for consistency.

In the second test, a probe was set to 1:1 with the probe point connected to the inner pin of the signal generator output connector and the probe ground lead to the output connector body. The result at 20MHz includes the small drop in oscilloscope sensitivity noted in the first test, so the probe only contributes a loss of 1.42dB.

The third test was a repeat of the second test with the probe set to 10:1, having previously adjusted the compensation trimmer for the best square-wave display.

The second and third test results include any effects of the inductance of the probe ground lead.

The third test result at 125MHz includes the small drop in oscilloscope sensitivity noted in the first test, so the probe only contributes a loss of 1.5dB.

Some of the test frequencies in Table 2 may seem arbitrary, but these were based on starting at the specified maximum frequency for each item tested, and then stepping down through octave ranges

FIGURE 11: The waveforms of a divider chain displayed simultaneously.

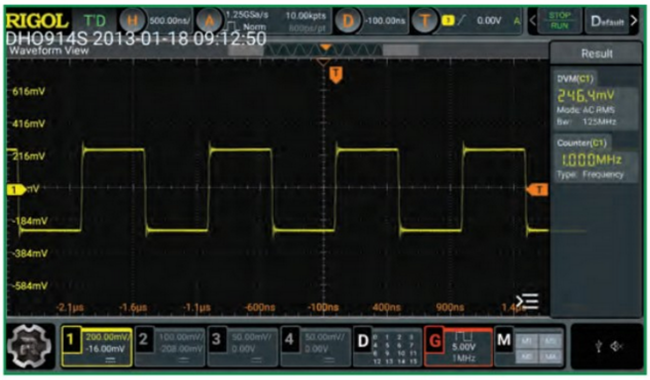

FIGURE 12: A square wave generated by the AFG.

FIGURE 13: A triangular wave generated by the AFG.

FIGURE 14: A 1MHz sine wave, amplitude modulated by a 1kHz triangular wave at 100%.

on the signal generator without making any other control changes. A small number of measurements were taken at frequencies outside the published specifications to see how the instrument behaved. No abrupt changes in performance were noted, just a gradual drop in sensitivity with increasing frequency. The counter display was stable up to 300MHz with a sufficiently-high input level.

Minimum sensitivity

Table 3 shows the minimum signal amplitude as a function of frequency that synchronises the display, and gives stable measurements of the signal amplitude and frequency on the most-sensitive range of 200µV per major graticule division. In each case, noise was clearly visible on the display. The signal generator was connected directly to the channel 1 input and terminated in 50Ω. The relative sensitivity was calculated from the DVM reading with a constant signal-generator output level. The frequency response at this very-low signal input level showed a steeper increase in loss with frequency compared to the frequency response shown in Table 2, but these tests were pushing at the measuring limits of the instrument.

Other points of note

At the maximum input setting, the vertical calibration is 10V per major graticule division, so the instrument can measure +/-30V or 60V peak-to-peak maximum without any external attenuation. In all the sine-wave measurements, the DVM reading agreed with the screen waveform peak-to-peak amplitude, converted to RMS, to within the tolerance of the visual measurement.

The user manual recommends that the instrument be switched on for 30 minutes before the self-calibration routine is run. After several hours of use, the instrument was barely warm and the power adaptor only slightly warmer. If left plugged in with the oscilloscope switched off, the power adaptor ran cold.

Figure 11 shows all four channels in use at the same time, displaying 1MHz, 500kHz, 100kHz and 50kHz signals respectively from a series of linked frequency dividers in a crystal calibrator. To obtain a stable display, the lowest-frequency input on channel 4 was used for synchronisation.

The instrument may be set to make and save a series of user-defined measurements via the 'Measurement' menu. These may include a series of mathematical functions, if required, and the period between each measurement may also be defined.

The instrument may be controlled remotely via the network connector and USB-B connector on the rear panel. Full details are included in the programming manual available from **[1]**.

The instrument weighed approximately 1.77kg, including the power supply and power leads, and measured approximately 266mm wide, 162mm high and 78mm deep including feet, control knobs and carrying handle.

Arbitrary function generator

The AFG, which is only fitted to the 'S' models, can be used as a single signal source with the following properties:

- variable AC output level from 2mV to 5V;
- selectable output waveform (sine, square, ramp, noise, DC, and 'arbitrary');
- variable frequency, from 2mHz to 25MHz, depending on the selected output waveform;
- variable DC offset;
- variable phase offset;

FIGURE 15: The Fast Fourier Transform display.

FIGURE 16: A 125MHz sine wave.

- selectable modulation type (AM/FM/PM);
- variable modulation frequency, from 2mHz to 1MHz; and
- selectable modulation waveform (sine, square, triangle, up-ramp, down-ramp), or none.

Examples of a few of the AFG output waveforms at 1MHz are shown in **Figure 12**, **Figure 13**, and **Figure 14**.

The AFG operations may be set by time, as well as amplitude, frequency, and wave shape. It should therefore be possible to conduct receiver AGC and transmitter ALC transient response measurements, as well as a more-conventional range of low- and high-frequency circuit and filter-response measurements, up to a maximum frequency of 25MHz.

Fast Fourier transform

An oscilloscope is mostly used to display waveforms in the time domain. The FFT option in the 'Math/Operator' sub-menu may be used to display a waveform in the frequency domain, ie its individual frequency components. **Figure 15** shows the 1MHz square wave from the AFG in the time domain on the left of the display, and some of its frequency components on the right.

In use

This instrument is good value given the broad range of functions provided in addition to the 'standard' oscilloscope functions. It may be overly complex and expensive for the radio amateur who has only basic requirements, and indeed there are less-complex and lower-cost instruments available from Telonic. However, if you are a more-advanced user, you will find that this instrument makes an important contribution to your shack.

I found it easy to synchronise the instrument display with the majority of incoming waveforms with amplitudes greater than about two minor graticule divisions. No loss of synchronisation occurred following a sudden reduction of signal level, as long as the displayed signal level was greater than two divisions; this was an inconvenience noted by the reviewer on the earlier DS1102E oscilloscope.

The lack of support for an external keyboard was not a problem, as the on-screen virtual keyboards were easy to use with my finger tips or a mouse. When I connected a 21in ACER colour display to the HDMI port, it almost immediately showed a replica of the oscilloscope display.

An amplitude-modulated waveform may be correctly synchronised to the modulating signal by applying the modulating signal to another input channel, and selecting that channel for synchronisation.

FIGURE 17: A 125MHz waveform with aliases.

Digital sampling

When using this, or any sampling oscilloscope, you should be aware of the constraints imposed by the digital sampling. The sampling speed defaults to being directly related to the horizontal scan speed set by the horizontal scale control, similar to the x time-base setting in an analogue oscilloscope. If the waveform being measured by sampling has a frequency component which is greater than half the sampling frequency, then the Nyquist, or sampling, theorem shows that alias products will be generated by the sampling process, and these will be visible on the screen. Viewing a few cycles of the input signal will be fine, as shown in **Figure 16**, but turning the scale

Table 1: The main instrument functions.

Function	Description
Digital oscilloscope	Four analogue input channels, with a maximum frequency of 125MHz, and capture rates up to one million waveforms per second, a 50MB memory depth, and 12-bit digital processing.
Logic analyser	16 digital channels, with 25Mpts memory depth and 625Msa/s.
Arbitrary function generator	Single output channel up to 25MHz, with selectable waveform type and modulation type. Model S only.
Protocol analyser	Supports decoding RS232/UART, I2C, SPI, CAN and LIN serial bus.
Bode plot	Model S only. Graphically display the frequency and phase response of an external system

Table 2: Oscilloscope and probe frequency responses (os signifies that the result is outside of the specification; # indicates that the probe compensation trimmer may need a minor adjustment).

Frequency (MHz)	Channel 1 direct	Via probe set to 1:1	Via probe set to 10:1	Comments
1	0dB (200.1mV)	0dB (302.5mV)	0dB 30.09mV)	Measurement reference level
3.9	-0.02dB	-0.02dB	+0.04dB	
15.6	-0.22dB	-0.99dB	+0.42dB #	
20	-0.18dB	-1.6dB	+0.74dB #	Probe 1:1 specification limit
31.25	-0.3dB	-3.76dB (os)	+1.37dB #	
62.5	-0.46dB	-8.98dB (os)	+2.34dB #	
125	-1.16dB	-17.5dB (os)	-2.66dB	Oscilloscope specification limit
150	-1.7dB (os)	-16.4dB (os)	-1.22dB	Probe 10:1 specification limit
200	-2.74dB (os)	-19dB (os)	-4.23dB (os)	Display synchronises, and the DVM reading is stable
250	-4.9dB (os)	-29.5dB (os)	-15.9dB (os)	Display synchronises, and the DVM reading is stable
300	-8.8dB (os)	-30.2dB (os)	-16.5dB (os)	Display synchronises, but the DVM reading is unstable

control anticlockwise automatically reduces the sampling rate, and you will see eventually a number of these alias products (**Figure 17**).

A more-complex example is shown in **Figure 18**. When viewing an amplitude-modulated waveform of several MHz on channel 1, and the modulating signal of several kHz on channel 2 with the display synchronised to channel 2, channel 1 shows significant alias products. The correct solution, to avoid seeing the alias products, is to override the default memory depth, and set it to its maximum value so that the sampling rate is also at its maximum (see **Figure 19**).

Other points

A small number of anomalous readings were noted at the very threshold of measurement of the vertical input, which were worst at the maximum frequency of 125MHz, becoming progressively better at lower frequencies and not detectable below 1MHz. With the vertical sensitivity set to maximum (fully clockwise), and a directly-connected synchronised input signal of less than 800µV at 125MHz from the signal generator, turning the sensitivity control one click anticlockwise caused the displayed waveform to increase in amplitude, whereas it should have decreased. One further click anti-clockwise caused the displayed waveform to decrease as expected. At these very-low signal levels, noise was clearly visible on the display. It is not clear if these unexpected results were a function of the analogue-to-digital converter or the processing software, but Telonic was able to confirm their existence. No other anomalous characteristics were detected during tests.

There is no warning indicator on the power supply adaptor to show that it is powered on.

The slim-line 13A mains plug, approximately 15mm deep, was easy to plug in, but more difficult to remove from a mains socket because of the small gripping area. The 5A fuse fitted to the UK mains plug seemed a little excessive in view of the low power consumption, and could probably be replaced with a fuse of lower rating.

The sixteen digital inputs, programming facilities, and remote-control options, were not tested during this review.

I recommend that you read the technical documentation available from Telonic and Rigol before purchasing this instrument to ensure that it meets your needs.

The instrument was supplied by Telonic Instruments Ltd (Telonic) **[1]**, the UK representatives of RIGOL Technologies Co. Ltd (Rigol) **[3]** and included a UK compatible fused power adaptor, four passive probes and an earth lead. My thanks to Telonic for the loan of this instrument.

Table 3: Minimum sensitivity as a function of frequency.

Frequency	Generator Output (pd)	DVM Reading (mV)	Relative Sensitivity
1MHz	100uV	100.5uV	0dB
16MHz	250uV	176uV	-3dB
64MHz	360uV	167uV	-6.7dB
125MHz	700uV	165uV	-12.5dB

References

[1] Telonic Instruments Ltd website: https://telonic.co.uk/

[2] Probe cable description: https://www.davmar.org/TE/TekConcepts/TekProbeCircuits.pdf

[3] RIGOL Europe website: https://www.RIGOL.eu/

FIGURE 18: An AM waveform with aliases.

FIGURE 19: The same AM waveform as in Figure 18, with no aliases.

Book Review

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The Radio Today guide to the Icom IC-905

By Andrew Barron, ZL3DW

Here we have the latest offering from Andrew Barron, taking us through the inner workings of the new and amazing Icom IC-905 SHF transceiver. This is the first transceiver to cover 2m, 70cm, 1.2GHz, 2.4GHz, 5.6GHz and 10GHz via an optional transverter.

The book provides a critique and hands-on view of the IC-905's use in the real world which I found refreshing. A good example of this is the chapter on 'Setting up the radio' which expands on the Icom manual which does not give a lot of detail on actual settings. A practical application of what to do to achieve a particular setup or function, whether that be a simple RTTY or a satellite configuration (such as QO-100), is explained here in simple steps. It also explains in clear language, the updating of firmware and additional software from Icom. The extra transverter that is needed for the 10GHz band is also covered in detail, including its setup and configuration.

One consideration that is prominent in the opening sections of the book is the additional equipment needed and practical constraints of using satellites and how these impact upon the IC-905.

The style of the book's composition is useful and the detail of the settings for the IC-905 are explained carefully. These are backed up by screen views to provide visual confirmation of configuration settings. This also gives a sense of comfort when inputting information into the transceiver as one always feels a sense of trepidation in case the settings are not appropriate and result in damage to the equipment.

I particularly like the advice Andrew gives on repeater and split operation. This can be a complex field, as seasoned satellite operators will know. He fully explains both the advantages and some shortcomings of the transceiver. One example of this is its inability to operate fully duplex. This is mainly due to the cross-band requirement of satellite access. In addition to this, the power output of the transceiver does not help under these circumstances and can require an external power amplifier.

Looking at the book as a whole, it has a clear structure as set out in its contents. However you are wishing to configure your transceiver, whether it be for general operation or for scanning a particular set of frequencies or modes, you will find it here. With a product like the Icom IC-905, there is so much that can be done, and so much detail in this book, that you may well disappear into your shack for many hours.

Size 174x240mm, 256 Pages, ISBN: 9781 9139 9553 9
Non Members: £15.99, RSGB Members: £13.59

The First Enigma Codebreaker

By Robert Gawlowski

The book charts the story of Marian Rejewski who is claimed to be the first codebreaker of the famed Enigma code. This story spawned nothing short of a revolution in the world of cryptography. The story shows the impact on the everyday lives of those involved.

The examination of Rejewski's life shows that an unassuming man, using his mathematical skills, changed the world of cryptography forever and in doing so compromised the seemingly unbreakable Enigma code. Born a Prussian, but of polish parents, he had a strenuous upbringing due to the harsh environment of the period. He studied mathematics at the University of Poznan as he had always had a liking for puzzles and problems. This was to set the scene for his future career.

Following university, he began work in his cryptographic career along with two other friends from the university who were working on Enigma. This was after joining an existing team which was having problems and considered Enigma to be unbreakable. The new team soon became an isolated and somewhat elite group working for the army. In those early days, the problem of tackling Enigma was to first build an Enigma machine in order to be able to replicate key sequences, although it would not of course decrypt a message. This was in itself a huge undertaking as the machine had 26 keys all linked to a stecker board with 26 cross-wiring points. In addition, the device contained three rotors plus a fourth rotor, which could be switched left or right to bias the settings. This was one of the most complex electromechanical machines ever devised. Once they had done that they could begin to analyse cryptonyms produced by it.

The book goes into detail regarding Marian's involvement with both Polish and French intelligence and the sequence of events leading up to the eventual breaking of the Enigma code which happened during the Christmas of 1932.

I found the book very interesting in that it mixes both the life story of Marian and the Enigma story. The account then follows the trail of Marian as he traverses across Europe to eventually arrive in England and take up a post with Ultra, ultimately handing over the baton to Alan Turing to continue the work which culminated in the Turing 'thinking machine' which could cryptanalyze German enigma traffic *en masse* and, more importantly, the same day!

I would recommend this book as it is a fascinating read. It is one of those books that you just cannot put down. Anyone who has any interest in history, or the enthralling story of Ultra, and the people who played such a vital role in it, will enjoy this book.

Size 162x226mm, 200 Pages, Hardback, ISBN: 9781 3990 6910 6
Non Members: £20.00, RSGB Members: £13.99 (30% OFF)

Review

REZ Ranger 80

portable antenna kit

FIGURE 1: The component parts of the antenna kit.

FIGURE 2: The kit comes with a back-pack.

It is always a great privilege to be given an opportunity to review a new piece of amateur radio equipment, and I was pleased to have been asked to test the REZ Ranger 80 portable antenna kit.

This is a well-made piece of kit that is easy to deploy, and not only would it be of interest to portable QRP operators who are all too aware that the 'best linear for QRP rigs is a good antenna', but also to those running standard-power rigs both for SSB and digital modes such as FT8.

Introduction

Some of you may know that that I am interested in portable operation, either on the North Yorkshire Moors, or at my other QTH on the Ionian Island of Kephalonia. Prior to trying the Ranger 80, I had been using the FMJ multiband HF directional/vertical antenna system from Alpha Antenna, which covers 80m to 6m. When portable, I tend to operate QRP using SSB, but having recently made the QRP Labs 'QDX' transceiver, I have also been able to operate on FT8 outdoors. Of course, I appreciate that others use base-station rigs, such as the Icom IC-7300, when working portable, and, as I have one of these rigs, I decided to use it as well as my QRP equipment to evaluate this antenna.

Opening the box

On opening the box or, should I say, the series of stout cardboard cylinders, I was immediately struck by the quality and solidity of the component parts. The various elements include a ground spike that is 17in (43.2cm) long, a tuning coil whose base has four banana sockets to which the four radials are attached (supplied), and a 'military' telescopic antenna which, when collapsed, is 17in (43.2cm) long. The telescopic whip of 7 sections extends to 9ft 3in (2.8m), and is constructed from brass and stainless-steel parts; it is described as having a 'stealth black coating'. **Figure 1** shows all the components of the antenna kit. The kit is not light, but I do not think that there is any great tension between weight and portability; moreover, the antenna comes with a large backpack (**Figure 2**) which can not only contain the constituent elements, but can also accommodate small rigs such as an IC-7300 and associated battery packs. The version provided to me could operate from 80m to 15m, but it is possible (for an extra cost) to obtain a kit to enable the antenna to be used on the 12m and 10m bands as well.

Setting it up

This is simplicity itself, as well as being intuitively obvious. Nevertheless, for those uncertain about deployment, there are several YouTube videos which are helpful. It took me just over a minute to assemble the antenna in one of my fields, and a similar time to dissemble it, clean the earth spike, and pack the whole system away. The manufacturers state that the earth spike is for use in 'soft or medium soils'. The whole deployment of the Ranger 80 was significantly quicker than setting up my Alpha antenna, which takes about five minutes. **Figure 3** shows the base of the antenna secured in the ground by the spike before deploying the radials. **Figure 4** shows the antenna set up in one of my fields for portable operation.

The manufacturers claim that the antenna will handle 200W on SSB and 100W with digital modes, such as FT8, on a 50% duty cycle. Power from the transmitter is delivered to the base of the tuning coil via an SO239 connector (**Figure 5**).

FIGURE 3: **The base of the antenna secured in the ground by the spike before deploying the radials.**

Tuning the coil

This can be broken down into two operations, namely band-tuning and fine-tuning. Tuning is carried out by sliding the collar up or down on the coil unit. **Figure 6** shows the coil assembly, and it can be seen that the collar is right at the top of the tuning coil; for coarse tuning, it is moved up or down over the turns of the coil as required. In general terms, the higher the collar, the higher the resonant frequency. It is essential to have an antenna analyser to tune the antenna. You need to connect your antenna analyser to the SO239 socket via a length of co-ax; it is important is that you make your measurements from at least 4m away so that body capacitance does not affect the reading. What I did was to estimate where on the coil the desired band-tuning point might be, take a reading, and move the coil up or down in an iterative process.

Now comes the fine-tuning operation. This is achieved by rotating the coil clockwise or anticlockwise and measuring the resonant frequency with the antenna analyser. Clockwise rotation decreases the frequency and anticlockwise rotation increases it. Make adjustments for minimum VSWR. I would recommend that, for each of the bands, you take a photograph of the height of the collar, and/or count or measure the collar's height. This may all sound involved and time consuming, but I found that I was able to carry out the band-tune operation, and then the fine-tune operation, within about two minutes the first time, and then much more quickly thereafter.

My measurements of the VSWR are shown in **Table 1**, together with the values reported by the manufacturer. Discrepancies may be ascribed to differences in ground properties.

So how well did it work?

My best QRP contact was on 60m with VO1NC

FIGURE 4: **The antenna set up in one of my fields for portable operation.**

in GN38 running 4W. Using 100W and FT8, I worked VK5MN in PF95. In addition to these, I had QRP FT8 QSOs with stations all around Europe, chiefly on 40m, and with multiple stations on the eastern seaboard of the USA and Canada on 20m, and a few JA stations on 15m. Using 10W SSB on 80m, I worked a variety of stations in all four countries of the UK, the Republic of Ireland, The Netherlands, Germany and France, all with reports no worse than 5 and 5.

Conclusions

I was unable to discern any significant rise in temperature of the coil after one hour of continuous operation on 15m using 100W. It is heavier, but much easier to deploy, than with my Alpha Antennas FMJ. On the negative side, it does not appear to come as standard with the capability to operate on the 12m and 10m bands, and you need to buy the additional kit to do so. The cost of the antenna is currently £599.95. Would I buy it? Possibly, if I did not already own my Alpha Antennas FMJ antenna which I can use from 80m to 6m. If you do buy one, it will last a lifetime and, should you not chose to employ the ground spike, the 3/8-24 mounting hardware will enable it to be set on a tripod.

Further information

There are several websites and YouTube videos that you may wish to study (**[1]**, **[2]**, **[3]** and **[4]**). I am grateful to Moonraker Group Ltd for the loan of this antenna. If you are interested in buying one, you can contact them on 01908 281705 or by email at **[5]**.

References

[1] https://www.rezantenna.com/

[2] https://youtu.be/2LXandLjEHQ?si=8d8aK70JQvZMPff6

[3] https://youtu.be/UhTj6eA1pb8?si=wYv-auqbLjlHStNX

[4] https://youtu.be/X5eAMEWfADE

[5] sales@moonraker-group.com

FIGURE 5: **Power from the transmitter is delivered via an SO239 connector.**

FIGURE 6: **The coil assembly showing the collar right at the top.**

Table 1: Factory and measured values of VSWR at various frequencies.

Frequency (MHz)	Factory VSWR	Measured VSWR
3.757	1.46	1.25
5.357	no data	1.1
7.159	1.04	1.2
10.136	no data	1.25
14.200	1.14	1.6
18.120	1.3	1.5
21.000	1.5	1.8

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Noise floor and all that

Nowadays, nearly all of us suffer from man-made interference in the HF bands in the form of wide-band non-directional background noise. This defines our 'noise floor', and limits our ability to make contacts. How bad is my station? Let's see how to find out.

The characteristics of the noise

QRM comes in various forms, such as interference from another station, or emissions from particular sources like noisy switching power supplies, or internet connections using VDSL etc. These types may be reduced by moving to an adjacent frequency, or by turning your antenna so as to place a 'null' of the antenna pattern in the direction of the interfering signal. The signals may be described as 'directional', in that they are generated by particular sources, and they arrive at your antenna from particular directions. However, try as you might, you are always left with a residual background noise, especially at HF. Some of this is generated in your receiver, but if the level of the noise increases significantly when you connect your antenna, the rest comes mostly from the accumulated emissions from electrical apparatus all around you. It seems to be the same no matter where you point your antenna. It extends from one end of the band to the other, and has the characteristics of random noise.

An example of this is shown in **Figure 1**. I use a 'SunSDR2 pro' transceiver in my shack and, being an SDR, it shows a spectrum plot and waterfall display over a wide range of frequencies. It is also accurately calibrated, and measures the power presented to the antenna terminals to within 0.5dBm. Note that, even in the regions where there are no signals being received from distant stations, such as around 7.005MHz, the floor of the spectrum is at around -110dBm. In 2.5kHz bandwidth, that is equivalent to about an S6-S7 signal as shown on the S meter. (The bandwidth of the receiver, 2.5kHz in this case, is shown by the lightly-shaded grey area on the spectrum plot.) When I disconnect the antenna, the noise floor decreases by more than 25dB, so I can confidently say that the noise is being picked up by my antenna. The user's guide for the radio states that the bandwidth used in the spectrum display is 9.53Hz, so we would expect the noise power in 2.5kHz to be 2500/9.53 times worse, or 24.2dB worse. The background noise within a 2.5kHz band is therefore -110 + 24.2 = -85.8dBm, and that is between S6 and S7.

David Lauder, G0SNO published a set of curves from the ITU in the April 2024 edition of *RadCom* suggesting that, at 7MHz, the noise floor E-field in a 'quiet rural' environment should be equivalent to about -50(dBμV/m) in 1Hz. Hmm! How do we relate this to my spectrum noise floor of -110dBm in 9.53Hz? Well, to do so we need to borrow an unfamiliar concept from radio astronomers, and invoke the second law of thermodynamics.

FIGURE 1: **The spectrum display shown on my receiver on 40m.**

Antenna temperature

Radio astronomers have been measuring noise signals since the 1940s. The signals emitted by celestial radio objects are mostly in the form of steady broad-band noise, and when they point their antenna at such an object, they measure the noise power received within the bandwidth of the receiver. They could express this power in a number of ways, such as dBm in 1Hz, but a favourite is to convert it into an 'antenna temperature'. The idea is based upon the fact that all resistors have minute fluctuating voltages across them which depend, amongst other things, on the temperature of the resistance (expressed in Kelvin, 273.16+°C). The radio telescope antenna (like every other antenna) has a radiation resistance in which the signals can be thought to be received, and the antenna temperature is then the equivalent temperature of the radiation resistance which would produce the same amount of noise. People familiar with EME operations at 144MHz and above will already be familiar with the use of antenna temperatures.

Let's take a concrete example. The maximum exchangeable noise power, P (W), generated by a resistance is given by P = kTB, where k is the Boltzmann constant (1.38 x 10^{-23}J/K), T is the temperature in K, and B is the bandwidth in Hz. Maximum exchange of power happens when the resistance is connected to a second resistance of the same value, so that the first resistance then dissipates kTB W in it. My spectrum noise floor at 7MHz is -110dBm, that is equivalent to a power which is 110dB less than 1mW. We take the base-ten exponent of the value in dB divided by 10 to get the ratio, ie 110dB is equivalent to a factor of $10^{(110/10)} = 10^{11}$. This is the number of times less power than 1mW. Expressed in W, -110dBm is 1mW/10^{11} = 1.0 x 10^{-14}W. The antenna temperature, T, in a bandwidth of 9.53Hz is given by T = P/kB = 1.0 x 10^{-14}/(1.38 x 10^{-23} x 9.53) = 7.60 x 10^{7} K – very hot! Of course, nothing actually is physically that hot. This number just means that the noise floor has as much power in it as would be delivered into the receiver's antenna terminals by a matched resistance at that temperature.

The antenna gain doesn't matter!

As we are dealing here with noise coming equally from all directions, it is as if we are in a hypothetical hot oven at a uniform temperature (in the example given at 7.60 x 10^{7} K). Our antenna's radiation resistance is 'at' this temperature. It does not matter what sort of antenna we have, whether just a half-wave dipole, or a stacked four by four over four etc, the radiation resistance of the antenna will be at that temperature. This may seem surprising. "Surely," (I hear you say) "the higher-gain antenna will pick up a greater signal, and therefore the antenna temperature will be higher." Not so. We have to take account also of the beam of the antenna. A higher-gain antenna has a narrower beam width than a lower-gain antenna. The higher-gain antenna will pick up a stronger signal, but from a smaller area of the inside surface of our 'oven'. The lower gain antenna picks up a weaker signal, but from a greater area, and these two effects exactly cancel. Not convinced? Let me show you a thought experiment which uses the second law of thermodynamics to prove the case.

Consider two perfect loss-less ovens, each at the same temperature. Since they are loss-less, neither gains or loses heat, and they will stay at the same temperature forever. Now let's join them together with a loss-less heat pipe (**Figure 2(a)**). Heat from each oven is able to pass along the pipe to the other, but because they are at the same temperature, the net flow, on average, will be zero, as the amount of power passing from top to bottom exactly equals that passing from bottom to top. This is an expression of the second law of thermodynamics, which perhaps is better explained in the inimitable words of the Flanders and Swann song **[1]**:

Heat won't pass from a cooler to a hotter;
you can try it if you'd like, but you'd far better
not-er;

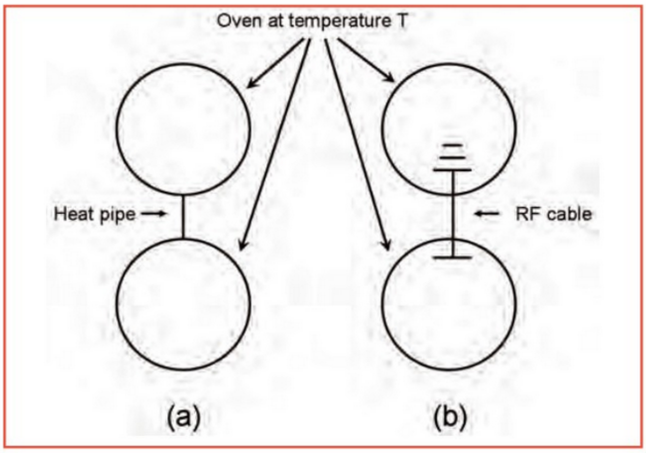

FIGURE 2: A thought-experiment which demonstrates that the antenna gain does not matter.

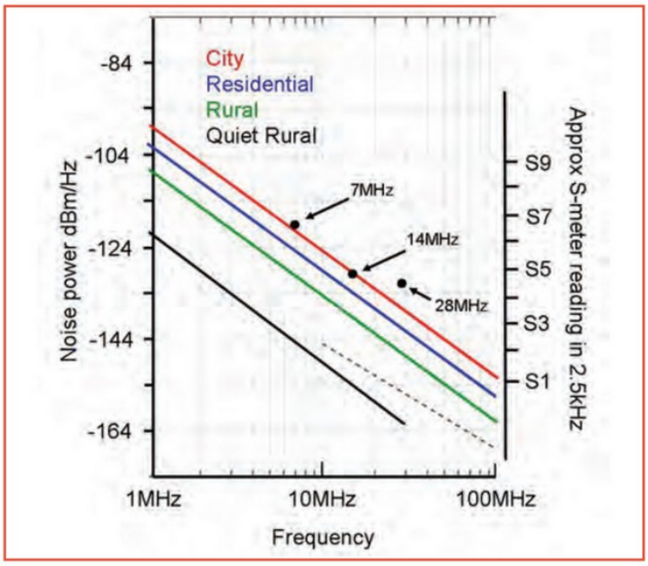

FIGURE 3: The ITU noise charts for 'city' (red), 'residential (blue), 'rural' (green), and 'quiet rural' (black) environments, adapted from [2] and re-expressed in dBm (LH scale). The dotted line shows the contribution made by noise from the Galaxy. My station's noise floor at three frequencies is shown by the black dots. The approximate equivalent S-meter reading in a bandwidth of 2.5kHz is shown on the RH scale.

'cause the cold in the cooler will get hotter as a rule-er;
because the hotter body's heat will pass to the cooler.

Now replace the heat pipe by connected antennas, one high-gain antenna in the top oven (**Figure 2(b)**), and one low-gain antenna in the bottom oven. Both have the same radiation resistance, say 50Ω, and power passes in both directions along the (loss-less) feeder connecting them together. The higher-gain antenna passes power to the lower-gain antenna which radiates it as radio waves, and *vice versa*. The second law of thermodynamics tells us, quite clearly, that the temperatures of the two ovens will not change, despite the antenna gains being unequal.

This result is useful to us in our analysis because it tells us that *all antennas are equal so far as this non-directional background noise is concerned*, and all will show the same noise floor on the spectrum plot.

The ITU plots

David's chart, reproduced from **[2]**, has a number of curves on it showing the background noise expected in 'quiet rural', 'rural', 'residential', and 'city environments'. I think that the noise E-field expressed in units of dB µV/m is a bit baffling and not helpful. To my way of thinking, dBs are defined to be ten times $\log_{10}$ of a ratio of powers. The unit µV/m is not a power, so you have to remember to use 20 rather than 10 and be very careful to keep this in mind. I have taken the ITU plot, and have re-cast it in terms of dBm. The lines show the ITU models for the four environments, which in Figure 39 of **[2]** are given as median values of the power per Hz relative to kT_0, where T_0 is set to 290K (about 17°C). Thus 0dB on the ITU plot corresponds to power $kT_0 = 1.38 \times 10^{-23} \times 290 = 4.0 \times 10^{-21}$ W. To convert to dBm, we use the expression

$$P_{dBm} = 10 \log_{10}(4.0 \times 10^{-21}/10^{-3}) = -174.$$

Thus, to re-cast the ITU plot (Figure 39 of **[2]**) to dBm/Hz, we just need to subtract 174 from the values on the vertical axis. The result of doing this is shown in **Figure 3** (left-hand scale). I have also included the approximate S-meter reading for noise in a 2.5kHz band on the right-hand scale. I say 'approximate' because the S-scale is defined for a pure sine wave, and the conversion to the equivalent noise power depends a bit on how it is measured.

The plot shows the power expressed in dBm in a bandwidth of 1Hz on the left-hand vertical axis, and the lines show the ITU levels plotted against the frequency (on the horizontal axis). My noise floor is -110dBm in 9.53Hz at 7MHz, so I need to subtract $10 \log_{10}(9.53/1) = 9.8$dB to find my noise in 1Hz. The answer is -119.8dBm. I have shown that as a black dot on the graph. I have also made measurements at 14MHz and 28MHz (also shown). We can now see from Figure 3 that, at 7MHz and 14MHz, my station's noise floor lies just above the 'city' curve, and quite a lot above it at 28MHz. Carnoustie, where I live, is hardly a city, but is certainly residential. This suggest either that the ITU models need to be revised upwards to take account of the much-increased background noise levels that we all experience, or that I have additional noise contributions in my measurements. I have gone to great lengths with filters to mitigate against noise going up the outside of my coaxial cable and into the antenna, but I can't be sure that there is not a wide-band noise source in a particular direction which is being picked up by my antenna.

You can check the noise floor at your station in a similar fashion. You must use a well-calibrated receiver, and take account of any amplifier or attenuator that you have. In my own SDR, this is taken into account automatically so that the spectrum plot always displays the power presented to the receiver. Look at your measured noise floor in dBm, correct for the bandwidth used to create your spectrum plot, then use Figure 3 to see how you are doing. I would be interested to know your results, so do drop me an email if you would like to do so, so that we can compare notes. In a recent *RadCom* article, it was stated that noise levels in the south east of the UK are 1000 times worse than in rural Australia. Mine is almost that bad at 7MHz and 14MHz, worse at 28MHz, and I don't live in a city in the SE of the UK, but on the east coast of Scotland. Perhaps, when there is a power cut, I will get my batteries out and see how much the noise level has gone down.

Dr David Lauder, G0SNO has written more about the ITU curves in his regular EMC column on pages 30 and 31 of this issue of *RadCom*.

References

[1] https://www.youtube.com/watch?v=VnbiVw_1FNs&ab_channel=notsyort

[2] https://www.itu.int/dms_pubrec/itu-r/rec/p/R-REC-P.372-16-202208-I%21%21PDF-E.pdf

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