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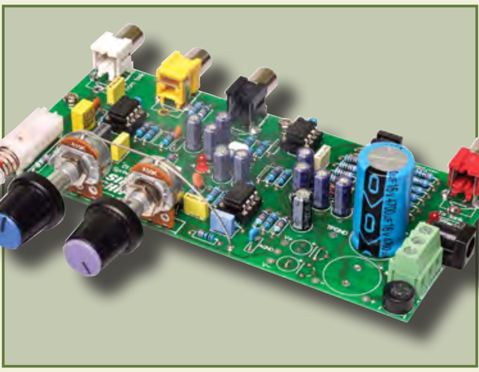

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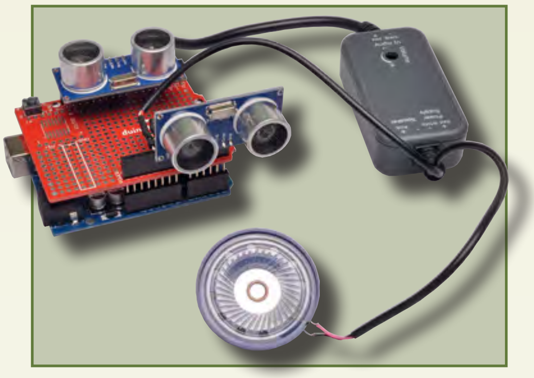

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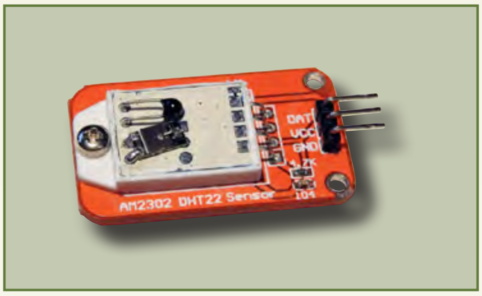

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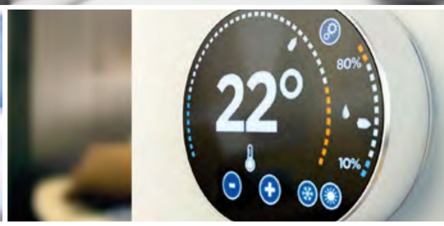

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EVERYDAY PRACTICAL ELECTRONICS

Off to a flying start

Finally, after a lot of anticipation (on my part) and a lot of hard work (on Mike Hibbett's part) I am delighted to announce a very important new project for our popular *PIC n' Mix* column. This month, Mike is looking at 'Practical DSP' (digital signal processing). I have wanted to cover this topic for a very long time, but never quite found the right combination of available author, space and cheap reliable silicon. But those crucial ingredients are now in alignment and I'm really looking forward to the upcoming months as Mike lays out a must-read guide to actually building a DSP system and implementing the all-important FFT (Fast Fourier Transform) algorithm in a practical useable circuit.

For those of you who are interested in negative feedback control, digital filtering or just advanced system experimentation, this series will be a fascinating introduction to a powerful technique – great things to look forward to!

SC200 amplifier parts sourcing

I hope you enjoyed the recent three-part series on the *SC200 amplifier*. We know quite a few of you are hoping to build it because we've had several emails about sourcing components. I've spoken to the project designer and between us we have a short list of items that we know are available.

First the transformer – the specified Altronics model is discontinued, as is the full amplifier kit. The easiest solution is to use two transformers: one 300VA 40-0-40 transformer and a 20 to 30VA 15-0-15 transformer (eg, Jaycar MT2086 or Altronics M4915B). The primaries of the two transformers should be wired in parallel. Arguably, this is superior to a single transformer since the 300VA transformer is then tasked with only supplying the power amplifier, so it will be able to supply slightly more peak current. It should also make chassis layout and wiring easier.

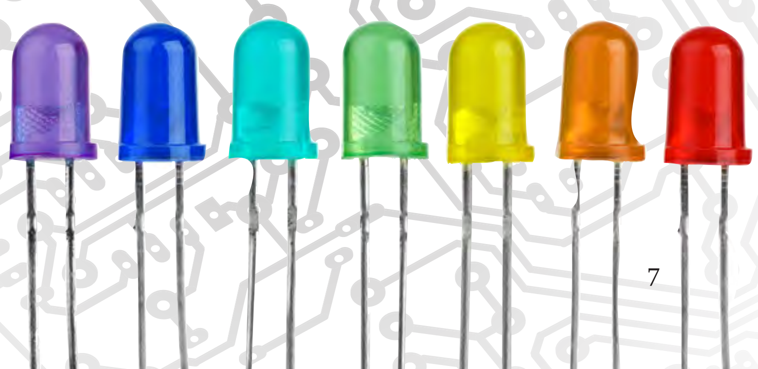
Second, the 'ferrite bead' choice isn't critical, but this one would work fine: http://bit.ly/2G8wEsL

Last, a couple of resistors are hard to find, but Farnell do have them:

Part 1435952: 0.1Ω, 1%, 3W 2512
Part 2476365: 6.8Ω, 5%, 3W 2512

It shouldn't matter too much if the 6.8Ω resistor is a 5% tolerance device, and not the specified 1%, since it operates with inductor L2, which will be poorly defined.

We're always here to answer your component sourcing problems, so don't let a hard-to-find part stop you building a project.

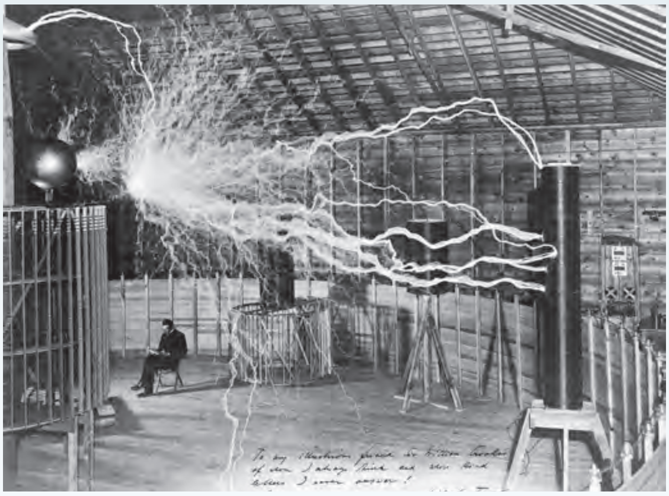


NEWS

A roundup of the latest Everyday News from the world of electronics

Wireless charging beyond the mat – report *by Barry Fox*

Remember Nikola Tesla? Around 1900 he was trying to transmit electric power wirelessly, over long distances, but failed because of the pesky inverse square law – received intensity diminishes according to the square of the distance from the source. So Tesla had to radiate huge amounts of power. And so today's wireless 'near field' chargers for phones need the device to sit on a charger mat.



Tesla's Colorado Springs laboratory had one of the largest Tesla coils ever built, which he called a 'magnifying transmitter' as it was intended to transmit power to a distant receiver.

So it caused quite a stir when, at the end of last year, Californian company Energous won certification from the US Federal Communications Commission for its WattUp Mid Field transmitter, which charges at distances of up to a metre. WattUp devices were then shown at the Consumer Electronics Show in Las Vegas in January. Energous claimed that WattUp 'delivers safe wire-free charging energy at distances of up to approximately 15 feet [4.5m]'. (**http://bit.ly/2Hbvmyz** and also: **http://bit.ly/2nXS4SH**)

Technical details of how Energous succeeds where Tesla failed were sparse, but the company has around fifty granted US patents and designs with nearly a hundred pending applications. The patents name Michael Leabman (Energous chief technical officer) as prime inventor and reveal a wealth of information on how WattUp works.

Radio bands

WattUp uses the ISM radio bands, which were originally reserved internationally for use with industrial, scientific and medical equipment, but are now also widely used for unlicensed communications such as Wi-Fi, Bluetooth and Zigbee. The first patent granted to Energous, in September 2015 (US 9,124,125) describes the basic principles of WattUp technology. The radiation is tightly focussed onto the target devices, to avoid hitting people or interfering with other equipment. Focussing is by shaping the energy into 3D 'pockets' of power, and this is done by simultaneously transmitting multiple radio waves with different phasing.

For instance, the patent explains, one wave may be at 5.7GHz and another at 5.8GHz. Targeting the power pockets can in theory be at distances ranging from 'a few centimetres to over hundreds of metres', the patent says. Devices to be charged are location-tagged by Bluetooth or even audio, and the power pockets focussed on the tags.

Significantly, when the US Patent Office looked for previous patents on similar technologies they turned up US patent 787,412 filed in May 1900 by Nikola Tesla of New York, for the 'Art of Transmitting Electrical Energy through the Natural Mediums'. But Tesla did not have the luxury of Bluetooth tagging and smart antennae, which could track and focus on tagged targets.

Several Energous patents describe powering a laptop or a TV from a wall transmitter, and then using it as a hub transmitter to power or charge other devices round the room. Other patents suggest using in-car antennae to charge and power devices inside the vehicle.

US 9,853,692 runs to 118 pages and explains more about the use of Bluetooth Low Energy chips for tagging and for device location and control.

Wi-Fi clash?

Observers have already warned that if WattUp works in the 5GHz ISM band it can interfere with Wi-Fi. (See: **http://bit.ly/2BrOJCT**). Now, US 2017/0373725 patent, filed by the Research and Business Foundation Sungkyunkwan University in Korea through 2016 and 2017, details the interference risks from wireless power technology to communication systems.

'In general' the Korean University patent says, 'in the wireless powered communication network, when the wireless power transfer apparatus uses the same frequency band as a wireless communication system such as Wi-Fi, the wireless power transfer apparatuses, unlike the wireless communication terminals, transfer a power signal without checking whether a channel is used. Therefore, there is a problem that the power signal of the wireless power transfer apparatus interferes

Wireless charging beyond the mat – *continued*

with the wireless communication system to interfere with wireless communication. Since each wireless communication terminal may perform communication only while the wireless medium is in an idle state, the transfer of the power signal of the wireless power transfer apparatus causes performance degradation of the wireless communication network.'

The Koreans believe the solution is to make the wireless power system behave more like a Wi-Fi communications network and continually exchange handshake signals that check whether the communications airspace is 'clear to send' and so avoid 'collisions'.

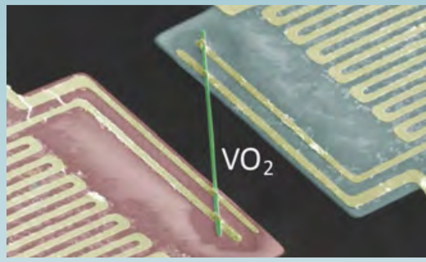
In response to questions on this, a spokeswoman for Energous has now explained that 'Energous received previous FCC certification for 5.8GHz Near Field transmitters. The recent FCC certification is for a Mid Field transmitter operating in the 900MHz band and the company has noted that the WattUp ecosystem will run in the 900MHz band for Near Field, Mid Field and Far Field solutions.'

So has Energous now changed its system from 5.8GHz operation to 915MHz operation? 'Yes. WattUp will launch and operate in the 900MHz band. However, Energous does have customer interest in 5.8GHz Near Field, closed-ecosystem solutions.'

Frequencies around 915MHz are already used for low power monitoring, and location and sensing systems, such as Zigbee light control (at 868MHz in Europe). So is the change by Energous to 915MHz operation to avoid interference to Wi-Fi? Says Energous: 'The 900MHz band has two main advantages over 5.8GHz for non-contact, charging at a distance: 1) The SAR (Specific Absorption Rate, a measure of the rate at which energy is absorbed by the human body) results are improved in the 900MHz band, and 2) Coexistence in the 900MHz band vs 5.8GHz is also improved.'

So how is interference to other systems like Zigbee avoided? Energous will say only: 'We have not publicly shared our coexistence technology/strategy.' So, clearly big question marks hang over claims that WattUp will 'work at 15 feet'.

Weird metallic conductor



A metal has been found that conducts electricity but produces little heat in the process. A team of researchers led by the Lawrence Berkeley National Laboratory have discovered this unusual property in vanadium dioxide (VO_2) at a temperature of 67°C.

Most metals are both good conductors of heat and electricity; classical physics explains that their electrons are responsible for both the movement of electrical current and the transfer of heat. VO_2 is different. When the researchers passed an electrical current through nanoscale rods, and thermal conductivity was measured, the heat produced by electron movement was ten times less than that predicted by calculations.

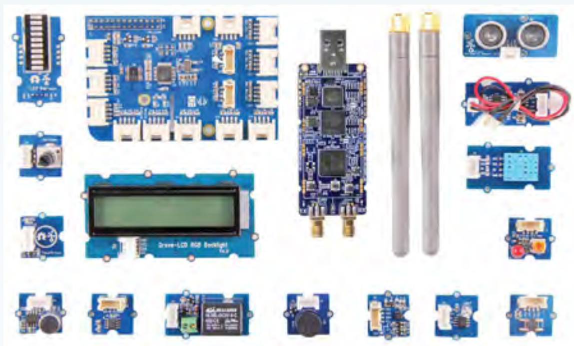
SDR starter kit developed for Raspberry Pi

Lime Microsystems has announced a Starter Kit for its LimeSDR platform based on the Grove Platform and for use with the Raspberry Pi. The £180 kit includes a LimeSDR Mini with antennas optimised for the 433/868/915MHz unlicensed bands, plus a GrovePi+ and an array of Grove sensors and outputs, many of which are supported by a Scratch extension, and other programming environments. The kit provides everything you need to get started learning SDR (software-defined radio) basics and developing IoT applications and is targeted at beginners and for educational use.

By combining the hardware components with Lime's ScratchRadio software extension, users will be able to quickly and intuitively create simple and fun applications that integrate software-defined radio (SDR) capabilities and peripheral I/O.

The LimeSDR is an award-winning software defined radio platform for cellular, IoT, DVB, GNSS and other RF applications. Together with LimeNET, it has a several-thousand-strong community of developers and backers, including EE and Vodafone.

Lime is now taking pre-orders for the Grove Starter Kit via its distribution partner: **www.crowdsupply.com/lime-micro/limesdr-mini/**



The Grove Starter Kit for LimeSDR Mini and Raspberry Pi

Wet string works!

TechnoTalk

Mark Nelson

Mention 'fibre broadband' and most minds turn immediately to optical fibre – but other options are available. Household string is made of fibres, either natural sisal or some plastic substitute, and as every electrical engineer (and radio amateur) knows, wet string does conduct electricity – up to a point. String's ability to carry broadband was actually investigated recently, and the results may surprise you (and no, this is not an early April Fools' spoof).

WET STRING WOULD NOT BE anybody's first choice for delivering broadband signals. But by the same token, neither would copper telephone wires that had been optimised for speech communication, in which the maximum frequency is around 3.4kHz. High-speed broadband operates at far higher frequencies (up to 17.66MHz for VDSL2), which considered in broadcast radio terms is the territory occupied by the 16m band that some shortwave radio stations use. Using copper telephone wire to deliver high-frequency broadband signals is not an ideal solution, although over short lengths it works tolerably well – for most users.

Deliberately degraded

So using telephone wire for broadband is a compromise, but not an entirely bad one. My computer at home is getting a solid 42Mbit/s over copper wires using BT Infinity (other providers are available) and I don't think I have any grounds for complaint, especially when some users in rural areas are hard pressed to receive a tenth of these speeds. But how far can you degrade the cabling and still pass a broadband signal though it? That's the question that exercised the mind of a technician at Andrews and Arnold Ltd, who have been in the Internet service provider (ISP) business since 1998.

Fully aware of the widely abused capabilities of wet string, he decided to see how far you could get squirting broadband signals down this distinctly 'alternative' type of fibre. You can see a report at: **www.revk.uk/2017/12/its-official-adsl-works-over-wet-string.html** but in a nutshell, a 'cable' of hairy string, two metres long and well doused in salty water, proved to be entirely capable of downloading from the Internet at speeds of up to 3.5Mbit/s. Uploading was not so rapid. Attaching empty tin cans to each end of the string added a voice capability too, although over a span of only two metres, this was perhaps a refinement too far.

Little things matter

Joking apart, after reading this, you might come away with the notion that as long as you use copper wire and not string, cabling has little or no influence on broadband signals. That's not the case, however. Poor installation practices can create all manner of signal loss problems that can drive you round the bend. The causes are seldom easy to trace and may exist either inside or outside your premises. Your ISP may well be to blame, but before you accuse them of providing data speeds far lower than you expected, do carry out a speed test with the computer plugged directly into the router at the point where the telecom wires enter your house, flat or office. The speed bottleneck that annoys you so much may lie somewhere along the wiring that you installed yourself; and you don't want to pay an eye-watering call-out charge!

Things to watch

If you reckon you are not getting the broadband speed that you should be receiving, check out where your router is installed. If it's not connected directly into the filtered NTE5 master socket (**www.increasebroadbandspeed.co.uk/tip4**), you may be throttling the signal. Factors that can affect broadband cable integrity include cable crushing, impedance bumps and selective suck-out. The minimum bend radius for data cable is four times the cable diameter, which is approximately one inch. If it is bent beyond this specified minimum bend radius, you upset the spacing between conductors, causing signal reflection and reducing transmission speeds.

Screw terminal blocks and jelly connectors don't affect voice signals but may cause nasty impedance bumps for data transmission. According to laboratory research by Krone (who designed the cable system used by BT), the major cause of data bit errors is signal reflections caused by impedance changes – or mismatches – at various points through the channel. When your equipment detects it has received a packet of garbage, it requests retransmission. This is fine for occasional errors, but as the incidence of garbling increases so do the requests for retransmission, becoming the major part of network traffic in worst-case situations.

Signal suck-out

Frequency-selective suck-out occurs when a spur cable off the main wiring route just happens to have a physical length that is related mathematically to the wavelength of the broadband carrier frequency. This so-called stub, which may have been left from an extension socket no longer in use, can suck out a significant fraction of the signal energy at and near (for instance) the quarter-wave resonance frequency of the stub, meaning that you will have to shorten or increase the length of the wiring spur – or disconnect it altogether. Signal suck-out can also occur in the cabling on BT's side of the NTE5, where it is notoriously difficult to track down and (obviously) you can do nothing to correct it. In extreme cases BT is obliged to tear out and replace all of the cable between your premises and the broadband cabinet in the street.

Optimum transmission depends on four key properties: characteristic impedance, high-frequency loss, delay and crosstalk. If any of these are badly impaired you will have some problems correcting them (and they are well beyond my proper comprehension of electrical transmission lines!). With luck, *you* won't have to. One last word: never use cheap cable for data wiring. Right now there is a lot of low-grade material on sale that uses conductors made of CCS (standing for copper-coated steel or copper-clad steel). It's a bit cheaper and certainly looks like normal cable. But don't be fooled: its resistance is higher than copper and it is also prone to fatigue at the point of termination, which can cause random faults especially with insulation displacement connectors (ICDs). Use this rubbish at your peril!

SPRING REVERBERATION UNIT

by Nicholas Vinen

Here's a blast from the past! Welcome to our *Spring Reverberation Unit* project for musicians. It uses an affordable, readily available spring 'tank' and a flexible power supply, so you can easily build it into your favourite amp – even if it's portable.

Despite the availability of digital reverb and effects units, many musicians, especially guitarists, still like the 'old school sound' of spring reverberation.

Put simply, a reverberation effects unit takes the dull sound of an instrument (including the human voice) being played in a 'dead' space and adds lots of little echoes.

These simulate what it sounds like to perform in an acoustically complex space such as an auditorium, which has lots of different hard surfaces for sound waves to reflect off, making for a much more 'live' sound.

Even if you're playing in a decent hall, adding extra reverb can make the hall sound bigger and grander. It's also a great way to help a beginner musician sound more professional.

To simulate all these acoustic reflections, rather than using digital processing, a spring reverb uses a spring 'tank' comprising two or more actual springs.

Sound waves are generated at one end of the springs using a voice coil, much like a tiny speaker, and just as sound waves travel through air, they will also happily travel down the metal springs. They are picked up at the other end by what is essentially a microphone.

Only, because of the (for lack of a better word) springiness of the springs, and the way they are suspended at either end, the audio signal doesn't just travel down the springs, it bounces around, generating echoes and since no physical process is 100% efficient, these decay, sound just like waves do when they bounce off walls, floors, ceilings, chairs and other objects.

It's a personal preference, but many prefer this effect to a digitally generated one.

The end result is something you really have to hear to appreciate, but it's surprising just how good a job the spring tank does of mimicking sounds bouncing around a hall.

Of course, the exact sound depends upon the exact tank used – some have two springs, some have three, some are longer or shorter and so on – but regardless of how natural it is, chances are you will find some configuration where it will add an extra dimension to your performance.

And being electronic, you can vary the reverb effect's intensity (or 'depth') and turn it on or off as necessary. But unlike a digital effects unit, you can't easily change other parameters such as the echo delay or frequency response.

Sourcing the spring tank

Fortunately, there are multiple suppliers of spring reverb tanks. You guessed it; most of them seem to be in China.

The one we're using is from a musical instrument component supplier called Gracebuy based in Guangdong, and at the time of writing this, you could purchase the tank for US$20 including free postage via the following 'shortlink': **http://bit.ly/2GvH9HG**

The same supplier sells this same unit on **ebay.co.uk**. Look for item 301900482233, or search for: 'Spring Reverb Tank Electric Guitar Amplifier 2 Spring Medium Decay TPSB2EB2C1B'.

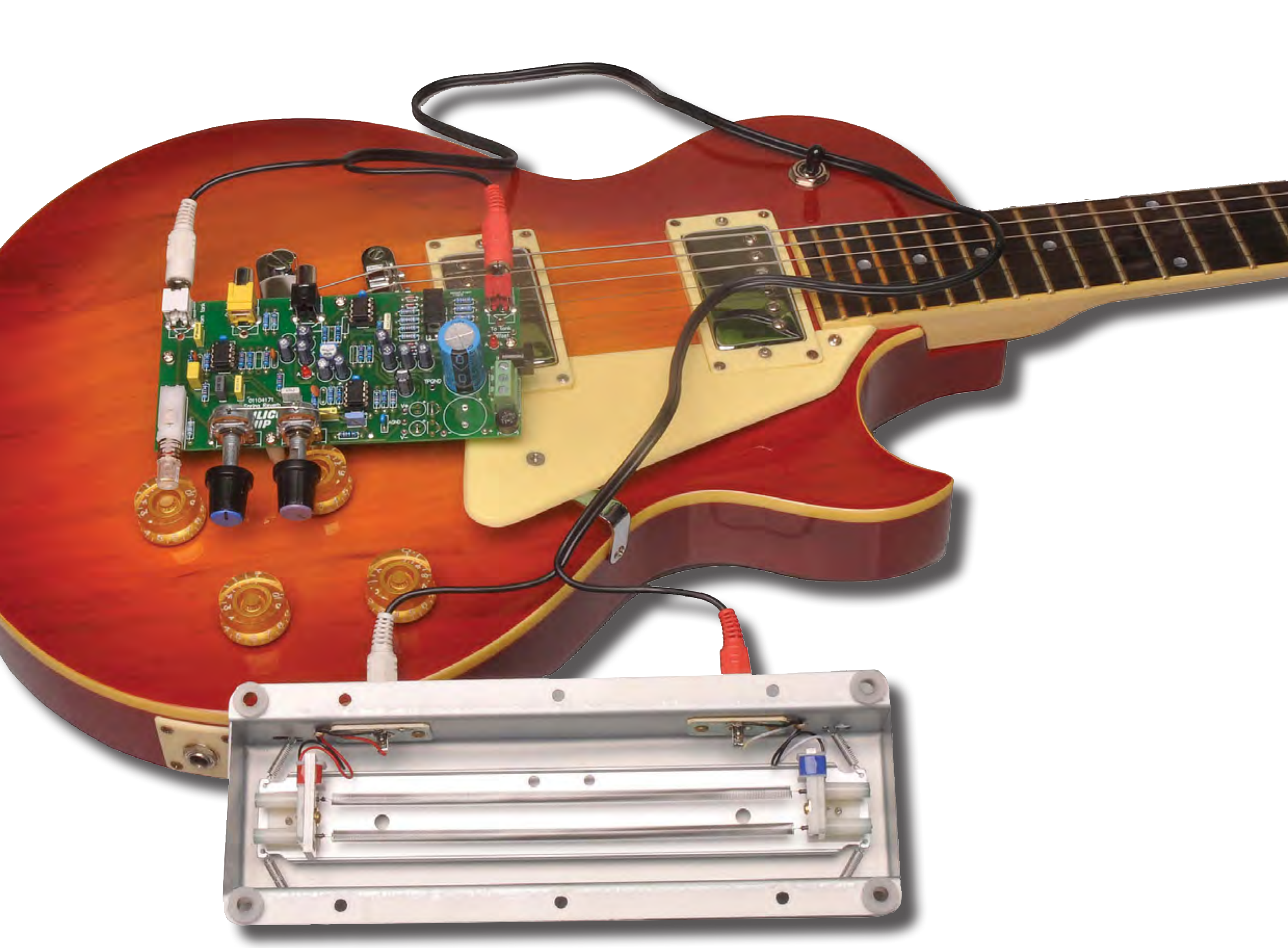
If you search eBay, you can also find other units, including some with three springs and/or longer springs. We haven't tried any of these but we would expect them to work with our circuit with little or no modification. So if you're feeling adventurous, here is a couple of eBay numbers for alternative spring tanks: 391937355232, 162782505026.

You can get an idea of the properties of the tank we're using by looking at the scope screen grabs in Figs.4-7. Three spring units will have triplets of echoes, rather than pairs, and longer units will have a larger gap between the stimulus and echo. Other tanks may also have a shorter or longer persistence time than the one we've used, depending on the properties of the springs themselves.

Note that most of the alternative tanks are larger than the one we've used (which is fairly compact; see the specifications panel) so make sure you have room for it in your amplifier's chassis (or wherever you plan to fit it) before ordering one.

Features and specifications

Reverb tank type: two spring
Anti-microphonic features: spring suspension, plastic mounting bushings
Spring tank dimensions: 235 × 87 × 34mm
Reverb delay times: 23ms, 29ms (see Figs.4 and 5)
Reverb decay time: around two seconds (see Fig.6)
Input sensitivity: ~25mV RMS
Frequency response (undelayed signal): 20Hz-19kHz (–3dB) (see Fig.2)
Frequency response (reverb signal): 200Hz-3.4kHz (–3dB) (see Fig.2)
Signal-to-noise ratio (undelayed signal): 62dB
Signal-to-noise ratio (typical reverb setting): 52dB
THD+N (undelayed signal): typically around 0.05% (100mV signal)
Controls: level, reverb depth, reverb on/off
Power supply: 9-15VAC, 18-30VAC centre tapped or 12-15V DC
Quiescent current: typically 30-40mA



You might think of it as 'olde world' but there's a surprising number of musos who say that a spring reverb always sounds better than a digital unit!

Improvements to the design

Compared to many other designs, an important aspect of this design is its ability to run off a DC supply. This was added so that buskers can add a spring reverb function to portable amplifiers, which may be powered by a 12V lead-acid battery or similar. In fact, the PCB is quite flexible and can be powered from 9-15VAC, 18-30VAC (centre-tapped) or 12-15V DC.

It's also possible to modify it to run off 15-30V DC, in which case you may need to increase the voltage ratings of the 1000µF and 220µF capacitors.

One small extra feature we've added, besides the new power supply options and related changes, is an indicator LED to show whether the reverb effect is active. It's built into the reverb on/off pushbutton switch, S1.

Basic concept

A block diagram of the *Spring Reverb Unit* is shown in Fig.1. The level of the incoming signal (from a guitar, keyboard, microphone, preamp...) is adjusted using potentiometer VR1 and is then fed both to a preamplifier for the spring tank and to a mixer, which we'll get to later. The preamplifier boosts high frequencies since the transducer which drives the springs is highly inductive and so needs more signal at higher frequencies to produce sufficient motion in the springs.

Between the preamp and the tank is the buffer stage, which has little gain but serves mainly to provide sufficient current to drive the transducer, which it does in bridge mode, for reasons explained below.

The output of the spring tank, which is delayed compared to the input and contains all the added reverberations, is fed to switch S1 which can shunt the signal to ground if reverb is not currently required. Assuming the signal is not shunted, it is fed to a recovery amplifier that boosts its level back up to a similar level to the input signal and then on to VR2, which is used to attenuate the reverberations in order to control the intensity or 'depth' of the effect.

The attenuated reverberations are then fed to the mixer, where they are mixed with the clean input signal to produce the final audio output, which can then be fed to an amplifier or mixer.

Circuit description

The complete circuit for the *Spring Reverb* module is shown in Fig.3. Note that two different ground symbols are used in the circuit. For the moment, you can consider them equivalent; we will explain the significance later, when we go over the power supply details.

The signal from the guitar/preamp/source is applied via RCA connector CON1 and then passes through a pair of electrolytic capacitors connected back-to-back (ie, in inverse series), which effectively form a bipolar electrolytic capacitor, to prevent any DC component of the signal from reaching the rest of the circuitry.

The signal then goes through a low-pass/RF filter comprising a 100Ω resistor, 4.7nF MKT capacitor and a ferrite bead. The –3dB point of the low-pass filter is around 340kHz, while the ferrite bead helps attenuate much higher

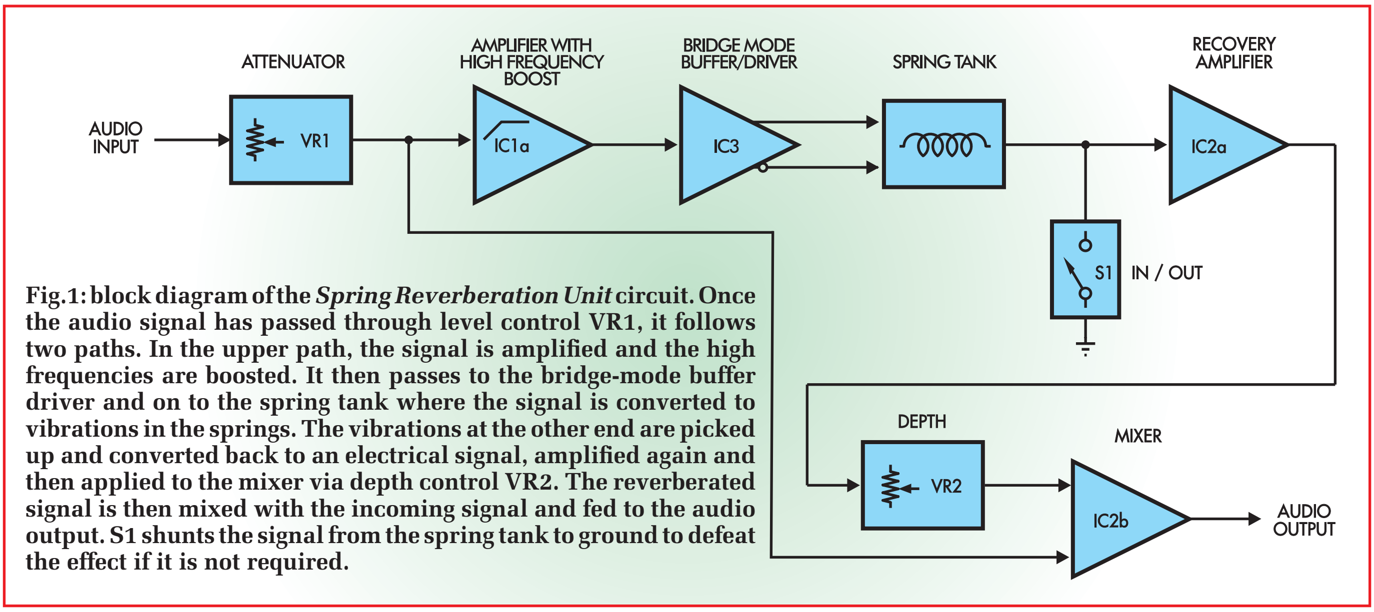


Fig.1: block diagram of the *Spring Reverberation Unit* circuit. Once the audio signal has passed through level control VR1, it follows two paths. In the upper path, the signal is amplified and the high frequencies are boosted. It then passes to the bridge-mode buffer driver and on to the spring tank where the signal is converted to vibrations in the springs. The vibrations at the other end are picked up and converted back to an electrical signal, amplified again and then applied to the mixer via depth control VR2. The reverberated signal is then mixed with the incoming signal and fed to the audio output. S1 shunts the signal from the spring tank to ground to defeat the effect if it is not required.

frequency signals (eg, AM and CB radio) which may be picked up by the signal lead. Both filters help prevent radio signal break-through. The audio signal then passes to 50kΩ logarithmic taper potentiometer VR1, which forms an input level control.

The level-adjusted signal from the wiper of VR1 goes to two different parts of the circuit, as shown in the block diagram (Fig.1); to the mixer, via a 47nF AC-coupling capacitor and to the tank drive circuit, via a 100nF AC-coupling capacitor. We'll look at the latter path first before coming back to the mixer later. The 100kΩ DC-bias resistor at input pin 3 of IC1a forms a high-pass filter in combination with the 100nF coupling capacitor, which has a –3dB point of 16Hz.

Note that in the original design, this part of the circuit used a 10nF capacitor which gave a –3dB point of 160Hz. The reason for having such a high roll-off was two-fold: first, the tank used previously had a very low input DC resistance and presenting it with a high-amplitude, low-frequency signal risked overloading the driving circuitry. And second, this helped attenuate 50/100Hz mains hum and buzz that may be from the guitar, cabling and so on.

Additionally, while it is possible to get good low-frequency performance, it's generally undesirable because it tends to muddy the sound.

We've shifted this –3dB point down because the transducer in the tank we're using this time has a much higher DC resistance and we've beefed up the driving circuitry, so overload is less of a problem, and this makes the reverb sound less 'tinny'.

However, you still have the option of reducing this capacitor value, possibly back to the original 10nF, if you find the unit has excessive hum pick-up. It really depends on your particular situation whether this is likely. Note that this solution to hum is a case of 'throwing the baby out with the bathwater'; at the same time as reducing the hum pick-up, you're also filtering out any genuine signals at similar frequencies.

Getting back to the signal path, IC1a operates as a non-inverting amplifier with a maximum gain of 101, as set by the ratio of the 100kΩ and 1kΩ resistors.

The 10nF capacitor in series with the 1kΩ resistor causes the resistance of the lower leg of the voltage divider to increase at lower frequencies, thus reducing the gain at lower frequencies. For example, a 10nF capacitor has an impedance of 16kΩ at 1kHz, thus the gain at 1kHz is reduced to 100kΩ ÷ (1kΩ + 16kΩ) + 1 = 6.9. The slope of the resulting filter is 6dB/octave and the –3dB point is 16kHz, which not coincidentally, happens to be the frequency at which a 10nF capacitor has an impedance of 1kΩ. In other words, the gain is reduced to about half its maximum (ie, 51) at 16kHz. You can see the effect of this filter stage in the frequency response diagram of Fig.2.

This 10nF capacitor also prevents the input offset voltage of IC1 from being amplified and creating a large DC offset at the output, while the 100pF capacitor across the 100kΩ resistor reduces the gain of this op amp stage at very high frequencies, preventing instability and also reducing the effect of RF/hum pick-up in the PCB tracks. The –3dB high-frequency roll-off point due to this capacitor is 16kHz.

Tank drive circuitry

Because the spring tank we're using has a fairly high input impedance of 600Ω at 1kHz, and because the springs themselves are quite lossy, the signal fed to the tank needs to have as large an amplitude as we can provide, given the supply rails available.

Note that the supplier lists the tank input DC resistance as 28Ω and its inductance as 23mH, but the actual measured figures are 75Ω and 83mH, giving an input impedance of just under 600Ω at 1kHz.

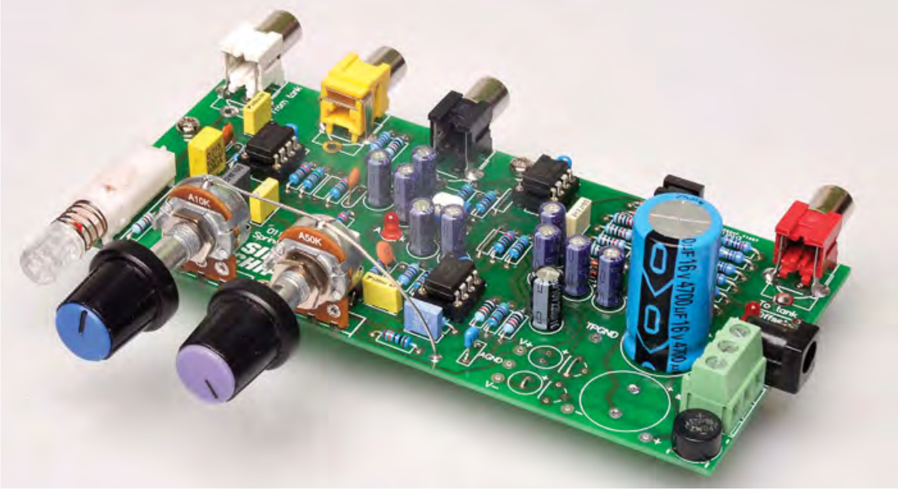
With ±15V supply rails, the LM833 and TL072 low-noise op amps we're using have a maximum output swing of around ±13.5V or 9.5V RMS. But since we've also designed this unit to be able to run off a 12V lead-acid battery (or equivalent) for busking purposes, and with a supply of only 12V, the output swing is much more limited at 9V peak-to-peak or just 3.2V RMS.

To improve this situation, we've redesigned the circuitry to drive the tank in bridge mode. This is possible since the driving transducer's negative input is not connected to its earthed chassis. That doubles the possible signal when running from a 12V DC supply, to nearly 6.5V RMS.

It works as follows. The output signal from gain/filter stage IC1a passes to both halves of dual op amp IC3. In the case of IC3a, it is fed directly to the non-inverting input at pin 3, while for IC3b, it goes to the inverting input at pin 6 via a 4.7kΩ resistor.

IC3a operates as a unity-gain power buffer. The output signal from pin 1 of IC3a goes to the tip connector of CON2 and hence the transducer in the spring tank via a 220Ω series resistor, but pin 1 also drives the bases of complementary emitter-follower pair Q1 and Q2 via two 22µF capacitors.

A DC bias voltage of around 0.7V is maintained across these capacitors due to the current flowing from the regulated V+ rail (typically +15V),



Here's the completed *Spring Reverberation Unit* (in this case to suit a DC power supply (see Fig.8[a]). Note the tinned copper wire link over the potentiometer bodies – it not only helps minimise hum but also keeps the pots themselves rigid.

through a 2.2kΩ resistor, small signal diodes D1 and D2, another 2.2kΩ resistor and to the V– rail (typically –15V). You can calculate the current through this chain at around (30V – 0.7V × 2) ÷ (2.2kΩ × 2) = 6.5mA, and this current sets the forward voltage across D1 and D2 and thus the average voltage across those two capacitors.

The voltage across these capacitors defines the quiescent base-emitter voltage of both Q1 and Q2, and thus their quiescent current, which is around 10mA. This is necessary to prevent significant crossover distortion when drive is being handed over between Q1 and Q2, as the output signal passes through 0V.

The two 10Ω emitter resistors help to stabilise this quiescent current by way of local negative feedback, since as the current through Q1 or Q2 increases, so does the voltage across these resistors, which reduces the effective base-emitter voltage.

The signal fed to the tank is also fed back to inverting input pin 2 of IC3a, setting the gain of this stage at unity. This closes the op amp feedback loop around Q1, Q2 and associated components.

The outer 'ring' terminal of CON2, which connects to the opposite end of the tank drive transducer, is driven by an almost identical circuit based on IC3b and transistors Q3 and Q4. However, so that the transducer is driven in bridge mode, the gain of this stage is –1, ie, it is an inverting unity-gain amplifier.

This is achieved by connecting its pin 5 non-inverting input to signal ground via a 2.2kΩ resistor and then using a 4.7kΩ feedback resistor and a 4.7kΩ resistor between the inverting input (pin 6) and the output of the previous stage, pin 1 of IC1a. The 2.2nF feedback capacitor rolls off the gain of this stage at high frequencies, giving a –3dB point of 16kHz and ensuring stability. The tank doesn't do much to preserve frequencies above 5kHz anyway.

By the way, we're using a TL072 op amp for IC3 instead of an LM833, as used for IC1 and IC2, because its lower bandwidth (and other aspects of the internals of this IC) makes it better suited for driving a complementary emitter-follower buffer. If you use an LM833 instead, the circuit will work but there is likely to be a spurious low-level ~1MHz signal injected which might upset the power amplifier.

This signal is due to the op amp having trouble coping with the extra phase shift introduced due to the transistors in its feedback path and it's hard to tame without adding some gain to the buffer stage, which we don't really need. Using a TL072 instead solves the problem and since all the gain is handled by the other two LM833 op amps (which have a lower noise figure), it doesn't degrade the performance at all.

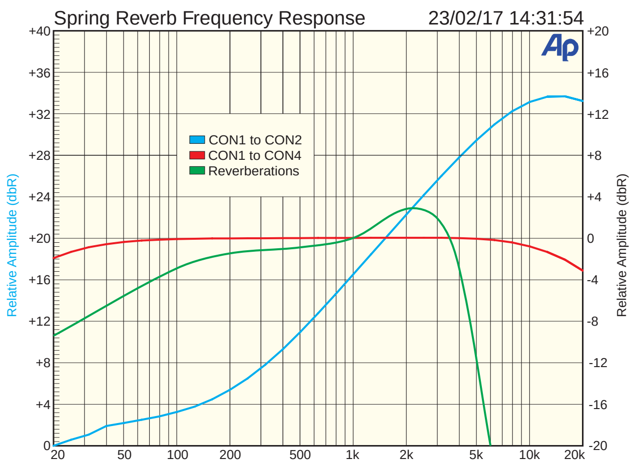


Fig.2: three frequency response plots for the *Spring Reverberation Unit.* The frequency response from input connector CON1 to spring tank driver connector CON2 is shown in blue and uses the left-hand Y-axis. The unit's overall frequency response, ignoring reverberations, is shown in red. The approximate frequency response for the reverberations is shown in green. This is difficult to measure since pulse testing must be used, otherwise standing waves cause constructive / destructive interference. Our curve is based on pulse testing at discrete frequencies and can be considered an approximation of the actual response.

Output offset adjustment

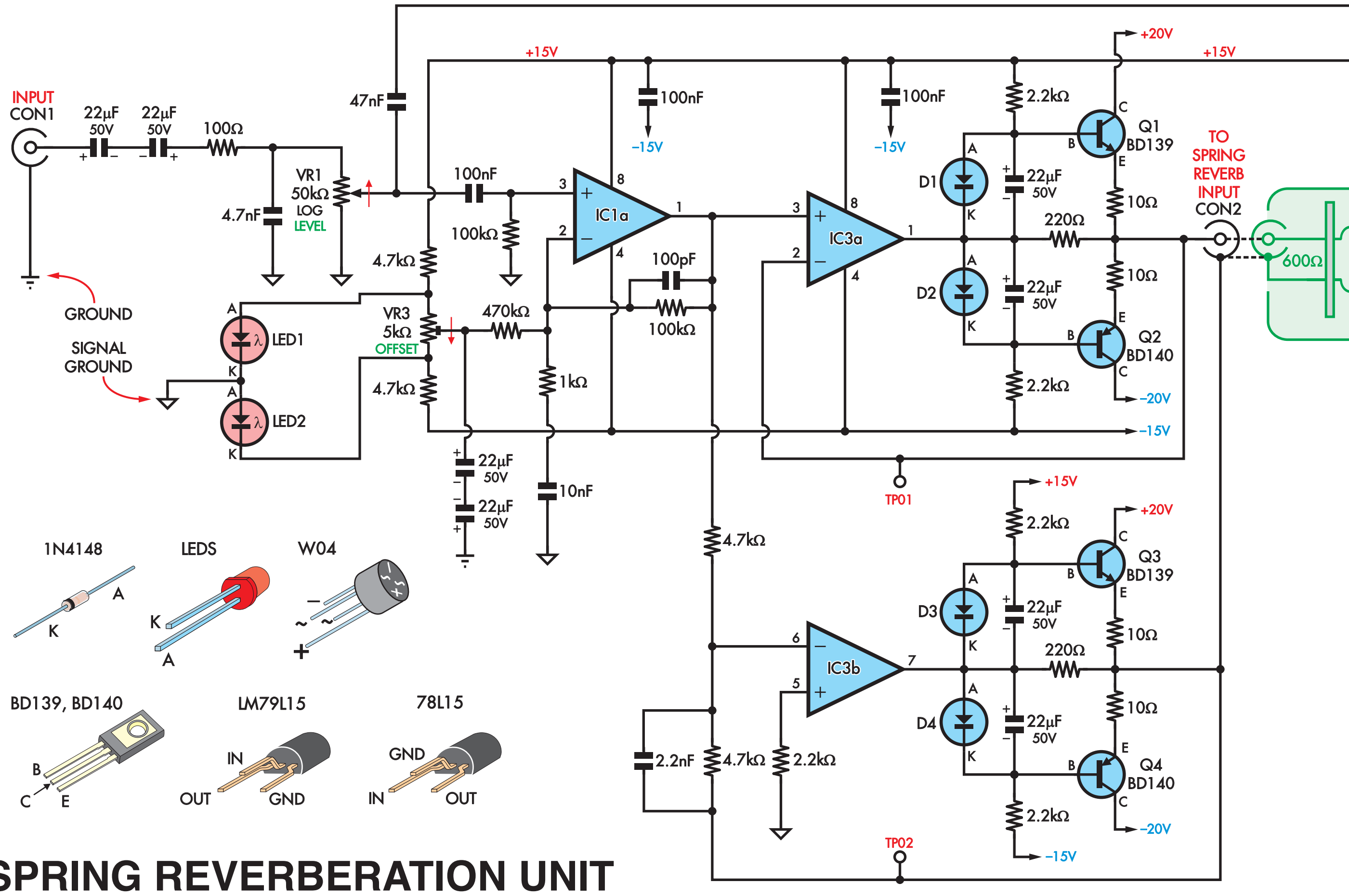
Since the transducer in the tank has a relatively low DC resistance, we'd like to avoid a high DC offset voltage across CON2 as this will waste power and heat up both the transducer and Q1-Q4 unnecessarily. It's not absolutely critical, but we've included DC offset adjustment circuitry because it's relatively simple and cheap.

But because this unit can run off an unregulated DC supply, we've designed it so that it doesn't rely on the regulated supply rails to provide a consistent offset adjustment.

Red LED1 and LED2 are connected across the supply rails with 4.7kΩ current-limiting resistors. The junction of LED1's cathode and LED2's anode is connected to signal ground. As a result, LED1's anode is consistently around 1.8V above signal ground while LED2's cathode is consistently about 1.8V below signal ground.

VR3 is connected between these two points and so the voltage at its wiper can be adjusted between these two voltages. Two back-to-back 22µF capacitors stabilise this voltage so that it does not jump around when power is first applied and the supply rails are rising. A 470kΩ resistor between VR3's wiper and pin 2 of IC1a allows VR3 to slightly increase or decrease the voltage at that pin, to cancel out any offset voltages in op amps IC1a, IC3a and IC3b.

Note that because IC3a has a gain of +1 and IC3b has a gain of –1, when you turn VR3 clockwise, the output voltage of IC3a will rise slightly while the output voltage of IC3b will drop slightly. Thus, there will be a position of VR3 such that the output voltages of these two op amps are identical when there is no input signal. This is the condition we're aiming for as it minimises DC current flow through the transducer connected to CON2.



SPRING REVERBERATION UNIT

Fig.3: complete circuit for the *Spring Reverberation Unit*, including the spring tank connected between CON2 and CON3 (shown in green). Only the output socket of the spring tank is connected to its case – this is to avoid earth (hum) loops. Note also that two different ground symbols are used; depending on the power supply arrangement, they may be connected together, or the signal ground may sit at half supply when powered from DC. Two different power supply arrangements are shown in the boxes at right and the PCB can be configured for one or the other. With an AC input, the circuit is powered from regulated, split rails of nominally ±15V while with a DC supply, the circuit runs off the possibly unregulated input supply.

Table 1 – expected voltages relative to TPGND

Supply	'+'	'–'	V+	V-	AGND
15VAC	+20V	-20V	+15V	-15V	0V
12VAC	+17V	-17V	+12V	-12V	0V
9VAC	+12V	-12V	+9V	-9V	0V
12V DC	+12V	0V	+12V	0V	+6V *(half V+)*

Signal recovery

The signal passes through the springs in the tank as longitudinal vibrations; and these are picked up at the opposite end by another transducer which is connected to the board via CON3. The signal from this second transducer is roughly –60dB down compared to the signal going in, so it is fed to another high-gain stage based around op amp IC2a, through another coupling/high-pass filter comprising a 100nF capacitor and 100kΩ resistor, with a –3dB point of around 16Hz.

Switch pole S1d is shown in the on position; in the off position, it shorts the signal from the tank to ground, so there is effectively no reverb.

IC2a is configured as a non-inverting amplifier with a maximum gain of 83 (820kΩ ÷ 10kΩ + 1). However, like IC1a, its gain is reduced at lower frequencies due to the 15nF capacitor in the lower leg of the divider, with a –3dB point of around 1kHz. As before, a capacitor across the feedback resistor ensures stability and reduces gain at very high frequencies; in this case, it is 10pF.

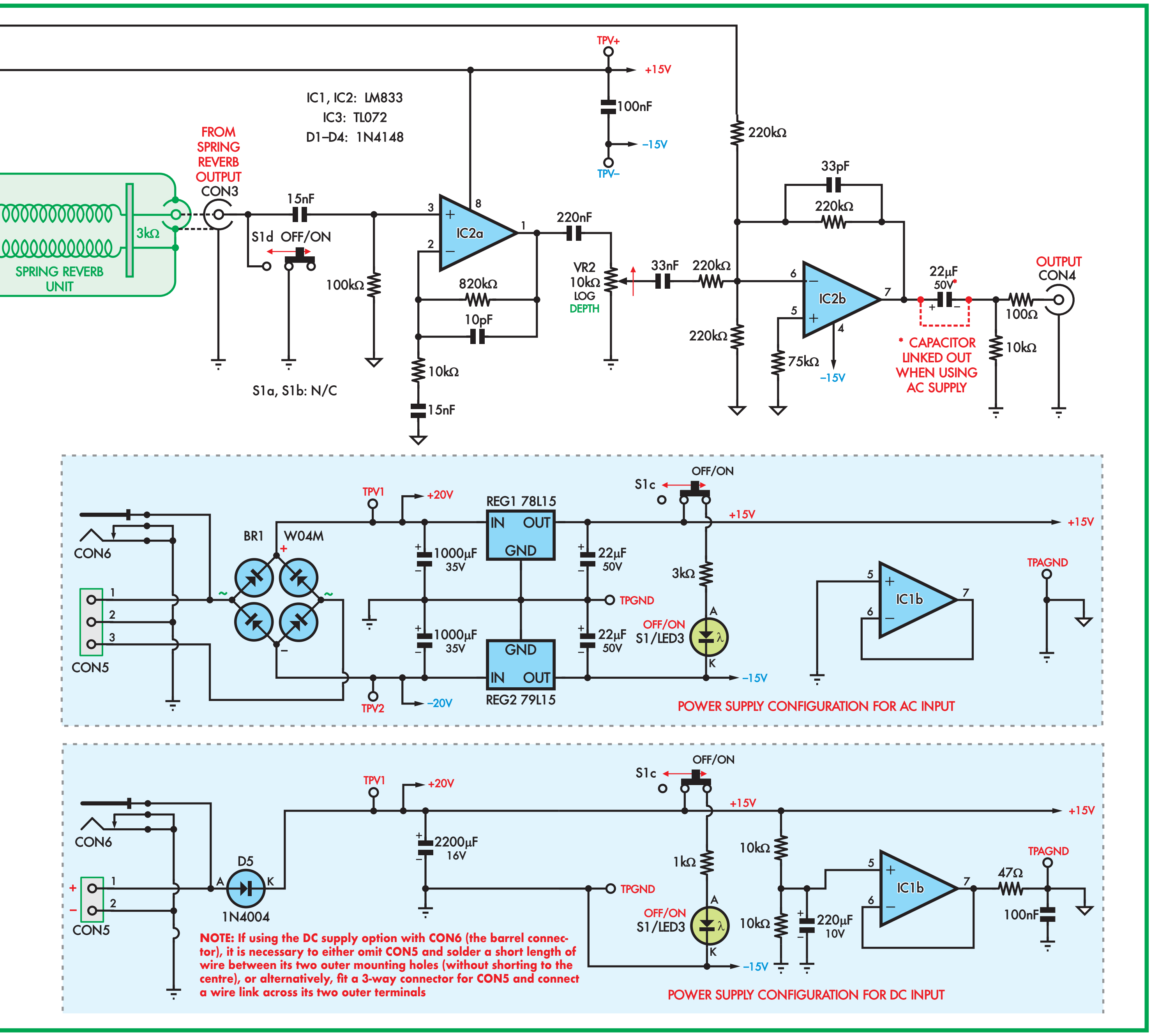
The recovered signal from the tank is then AC-coupled to 10kΩ log potentiometer VR2 via a 220nF capacitor. VR2 controls the level of the reverb signal which is fed to the mixer, and thus the 'depth' of the reverb effect. The resulting signal at its wiper is then coupled to inverting pin 6 of mixer op amp IC2b via a 33nF AC-coupling capacitor and 220kΩ series resistor.

The reason for using two coupling capacitors with VR2 is to prevent any DC current flow through it, which could cause crackling during rotation as the pot ages (note that we have done the same with VR1).

The mixer

You may remember that the signal from VR1 was fed both to the tank and to the mixer; after being coupled across the 47nF capacitor, it passes through a second 220kΩ series resistor to also reach pin 6 of IC2b. So this is the point at which the original and reverberated signals meet and you can see how VR2 is used to vary the effect depth, as the louder the reverb signal is compared to the input signal, the more reverberation will be evident.

A third 220kΩ resistor provides feedback from IC2b's output pin 7 back to its inverting input, while the non-inverting input (pin 5) is connected to signal ground via a 75kΩ resistor. This value was chosen to be close to the value of three 220kΩ resistors in parallel, so the source impedance of both inputs is similar. IC2b operates as a 'virtual earth' mixer, with both



its input pins 5 and 6 held at signal ground potential.

Remember that the action of an op amp is to drive its output positive if the positive input is higher than the negative input and negative if the situation is reversed. So the feedback from its output to its inverting input operates to keep both inputs at the same potential. Since the non-inverting input is connected to ground, the inverting input will be held at that same potential and the signals represented by the currents flowing through the three 220kΩ resistors are mixed and appear as an inverted voltage at the output.

The output of IC2b is fed to output RCA connector CON4 via a 22μF AC-coupling capacitor and 100Ω short circuit protection/stabilisation resistor. The capacitor removes the DC bias from the output when a DC power supply is used. If an AC supply is used, the output of IC2b will already swing around 0V so no DC-blocking capacitor is needed and it is linked out.

Note that the PCB has provision for two back-to-back electrolytics here (for use with an AC supply). However, IC2b's output offset should be low enough that most equipment that would follow the reverb unit (eg, an amplifier) should not be upset by it, hence we are not recommending that you fit them.

Power supply

Two different configurations for the power supply are shown in Fig.3 and you can choose one or the other depending on which components you fit. The one at top suits a transformer of 9-15VAC (or 18-30VAC centre tapped). AC plugpacks can be used. The power supply configuration at bottom is intended for use with 12V batteries or DC plugpacks and will run off 12-15V DC. However, it could easily be adapted to handle higher DC voltages of up to 30V if necessary.

Looking at the AC configuration at top, the transformer is normally wired to CON5. If it isn't centre tapped, the connection is between pin 2 and either pin 1 or pin 3. For tapped transformers, the output is full-wave rectified by bridge rectifier BR1, while for single windings, the output is half-wave rectified. The output from BR1 is then fed to two 1000μF filter capacitors and on to linear regulators REG1 and REG2, to produce the ±15V rails.

If your AC supply is much lower than 15V (or 30V centre tapped), you will need to substitute 78L12/79L12 regulators for REG1 and REG2 to prevent ripple from feeding through to the output. Similarly, for AC supplies below 12V (or 24V centre tapped), use 78L09/79L09 regulators.

Fig.4: the yellow trace shows the signal fed to the spring tank input, while the green trace at bottom shows the signal at the spring tank output. 23.6ms after a pulse is applied to the input, it appears at the output and then a second echo appears around 29ms after the initial pulse. You can see the next set of echoes due to the signal travelling up and down the springs again some 45ms later and note that each set of echoes has opposite polarity compared to the last.

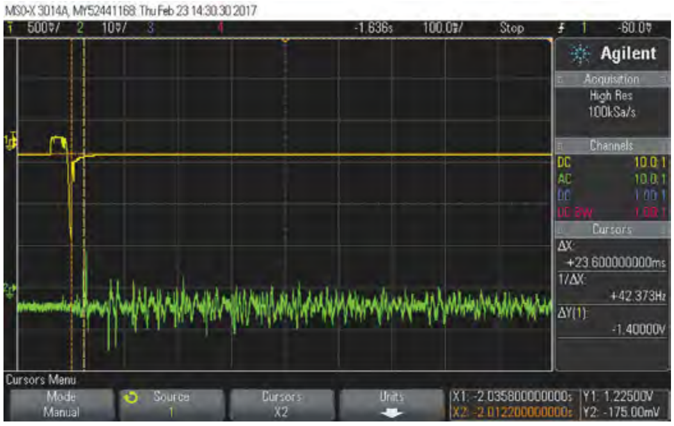


Fig.5: the same signal as shown in Fig.4 but this time at a slower timebase, so you can see how the reverberating echoes continue on for some time after the initial pulse, slowly decaying in amplitude.

Assuming the reverb effect is on, switch pole S1c will be in the position shown and so the LED within S1 will be lit, with around 9.3mA [(30V – 2V) ÷ 3kΩ] passing through it.

Op amp stage IC1b is not used with an AC supply and so its non-inverting input is connected to ground and its output to its inverting input, preventing it from oscillating or otherwise misbehaving. With an AC supply, the signal ground is connected directly to the main (power) ground via a link.

DC supply

For a DC supply, such as a 12V battery, the configuration at bottom is used.

If using the DC supply option with CON6 (barrel connector), it is necessary to either omit CON5 and solder a short length of wire between its two outer mounting holes (without shorting to the centre), or alternatively, fit a 3-way connector for CON5 and connect a wire link across its two outer terminals.

Diode D5 replaces the bridge rectifier and provides reverse polarity protection. The main filter capacitor is larger, at 2200µF, to minimise supply ripple.

Parts list – Spring Reverberation Unit

1 double-sided PCB, available from the *EPE PCB Service*, coded 01104171, 142 × 66mm
1 spring reverb tank (see text)
1 stereo RCA lead with separate shield wires
4 RCA sockets, switched horizontal or vertical (CON1-CON4)
1 3-way terminal block, 5.08mm pitch (CON5) OR
1 PCB-mount DC socket, 2.1mm or 2.5mm ID (CON6)
1 50kΩ logarithmic taper single-gang 16mm potentiometer (VR1)
1 10kΩ logarithmic taper single-gang 16mm potentiometer (VR2)
1 5kΩ mini horizontal trimpot (VR3)
2 knobs to suit VR1 and VR2
1 4PDT push-push latching switch with integral LED (S1)
8 PCB pins (optional)
1 100mm length 0.7mm diameter tinned copper wire
3 8-pin DIL sockets (IC1-3) (optional)

Semiconductors
2 LM833 low noise dual op amps (IC1,IC2)
1 TL072 low noise JFET-input dual op amp (IC3)
2 BD135/137/139 1.5A NPN transistors (Q1,Q3)
2 BD136/138/140 1.5A PNP transistors (Q2,Q4)
2 red 3mm LEDs (LED1,LED2)
4 1N4148 signal diodes (D1-D4)

Capacitors
10 22µF 50V electrolytic
1 220nF 63/100V MKT
2 100nF 63/100V MKT
3 100nF multi-layer ceramic
1 47nF 63/100V MKT
1 33nF 63/100V MKT
1 15nF 63/100V MKT
1 10nF 63/100V MKT
1 4.7nF 63/100V MKT
1 2.2nF 63/100V MKT
1 100pF ceramic
1 33pF ceramic
1 10pF ceramic

Resistors (all 0.25W, 1%)
1 820kΩ 1 470kΩ 3 220kΩ 3 100kΩ 1 75kΩ
2 10kΩ 4 4.7kΩ 6 2.2kΩ 1 1kΩ 2 220Ω
2 100Ω 4 10Ω

Additional parts for 9-15VAC powered version
1 78L09, 78L12 or 78L15 positive 100mA regulator (REG1) (see text)
1 78L09, 79L12 or 79L15 negative 100mA regulator (REG2) (see text)
1 W02/W04 1A bridge rectifier (BR1)
2 1000µF 35V/50V electrolytic capacitors, 16mm maximum diameter, 7.5mm lead spacing
1 22µF 50V electrolytic capacitor
1 3kΩ 0.25W 1% resistor

Additional parts for 12-15V DC powered version
1 1N4004 1A diode (D5)
1 2200µF 16V electrolytic capacitors, 16mm maximum diameter, 7.5mm lead spacing
1 220µF 10V electrolytic capacitor
1 100nF multi-layer ceramic capacitor
2 10kΩ 0.25W 1% resistors
1 1kΩ 0.25W 1% resistor
1 47Ω 0.25W 1% resistor

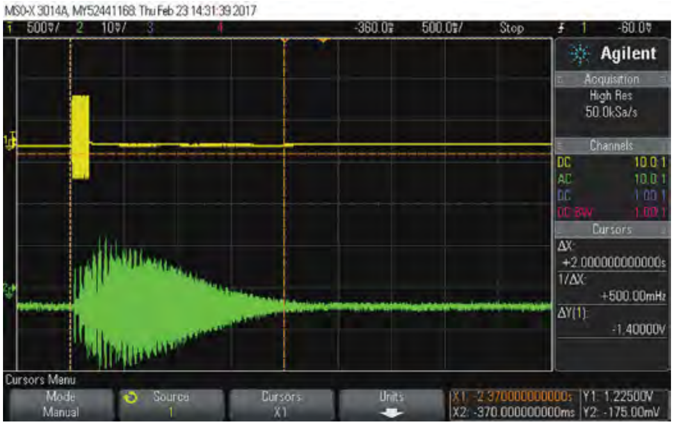


Fig.6: here we have a longer stimulus pulse, again shown in yellow, and the response shown in green on a much longer timebase. The reverberations continue for several seconds after the initial pulse but they have mostly died out after around two seconds (indicated with the vertical cursor).

Fig.7: this shows the output of the *Spring Reverberation Unit* with a short 1kHz burst applied to the input. You can see the original pulse at the left side of the screen and the reverberating pulses, which have been mixed into the same audio signal, repeated twice with decaying amplitude.

For DC supply voltages above 15V, substitute a similarly-sized capacitor with a higher voltage rating such as 2200µF/25V or 1000µF/50V.

The current-limiting resistor for LED3 has been reduced to 1kΩ so that it is still sufficiently bright with the reduced supply voltage, while IC1b is configured to generate a virtual earth at half supply. This is derived from the main supply via a 10kΩ/10kΩ resistive divider with a 220µF capacitor across the bottom leg to eliminate supply ripple from the signal ground.

Op amp IC1b is configured as a buffer, so that the signal ground has a low impedance and drives it via a 47Ω resistor, to ensure op amp stability.

A 100nF capacitor between signal ground and power ground keeps the high-frequency impedance of the signal ground low despite this resistor.

PCB construction

Assembly of the PCB is straightforward. It is available from the *EPE PCB Service*, coded 01104171 and measures 142 × 66mm with tracks on both sides, and plated through-holes. Two overlay diagrams are shown overleaf: Fig.8(a) shows the component layout for a DC supply, while Fig.8(b) shows the layout for an AC supply. Differences between the two will be noted in the following instructions.

Begin by fitting small signal diodes D1-D4, oriented as shown in Fig.8 and then use the lead off-cuts to form the wire links, shown in red. Both versions require five links to be fitted, but some of them are in different places so follow the appropriate overlay diagram.

Next, fit the resistors where shown. While their colour code values are shown in the table overleaf, it's a good idea to check the resistor values with a multimeter before fitting them and remember to slip a ferrite bead over the lead of the 100Ω resistor just above VR1.

The resistors fitted to both versions are almost identical; besides the variation in value of the resistor next to S1, the only other difference is that the three resistors to the right of IC1 are not fitted for the AC supply version.

For the DC supply version, you can now fit D5, oriented as shown.

If you are using IC sockets, solder them in place now, with the notched ends towards the top of the board. Otherwise, solder the three op amp ICs directly to the board with that same orientation. Note that IC3 is a TL072, while the other two ICs are LM833s, so don't get them swapped around.

For the AC supply version, solder BR1 in place with its longer (+) lead towards upper left, as shown in Fig.8.

Now proceed to install the two onboard red LEDs (LED1 and LED2) with the longer anode leads to the left (marked A on the PCB) and all the ceramic and MKT capacitors in the locations shown in the overlay diagram. Polarity is not important for any of these capacitors.

Note that LED1 and LED2 are lit as long as power is applied, so you could mount one of these off-board as a power-on indicator if necessary.

However, we think in most cases, constructors will be building the *Reverb* unit into an amplifier which already has a power-on indicator, so this should be unnecessary and LED1/LED2 can simply be mounted on the PCB as shown.

If you're building the AC-powered version, solder REG1 and REG2 in place now, oriented as shown. Don't get them mixed up. You will probably need to crank out their leads slightly using small pliers, to suit the PCB pads.

Now fit trimpot VR3, followed by illuminated switch S1. Make sure S1 is pushed all the way down onto the PCB before soldering two diagonally opposite pins and then check it's straight before soldering the remaining pins.

You can now install the small (22µF) electrolytic capacitors. These are polarised and the longer (+) lead must go towards the top of the board in each case, as shown using + symbols in Fig.8. If building the DC-powered version, there is also one 220µF capacitor that you can fit at the same time, but make sure it goes in the position indicated.

Next, mount CON5 and/or CON6, depending on how you plan to wire up the power supply. If fitting CON5, make sure its wire entry holes go towards the nearest edge of the board and if using a 2-way connector (for a DC supply), make sure it goes in the top two holes as shown in Fig.8(a).

Next, fit CON1-CON4. In each case, you have a choice of using either a horizontal switched RCA socket (as shown on our prototype) or a vertical RCA socket fitted either to the top or the bottom of the PCB.

Pads are provided for all three possibilities, and which is best depends on how you're planning on running the wiring in your particular amplifier.

As you will see later, we recommend using a stereo RCA-RCA lead to connect the main board to the tank, and the tank will normally be mounted in the bottom of the amplifier chassis while the *Reverb* board will normally be mounted on the front panel. So keep that in mind when deciding which RCA socket configuration to use.

If you want to fit PCB pins for the test points, do so now, however it isn't really necessary since the pads are quite easy to probe with standard DMM leads.

Transistors Q1-Q4 should be fitted next. Don't get the two types mixed up; the BD139s go towards the top of the board, while the two BD140s go below.

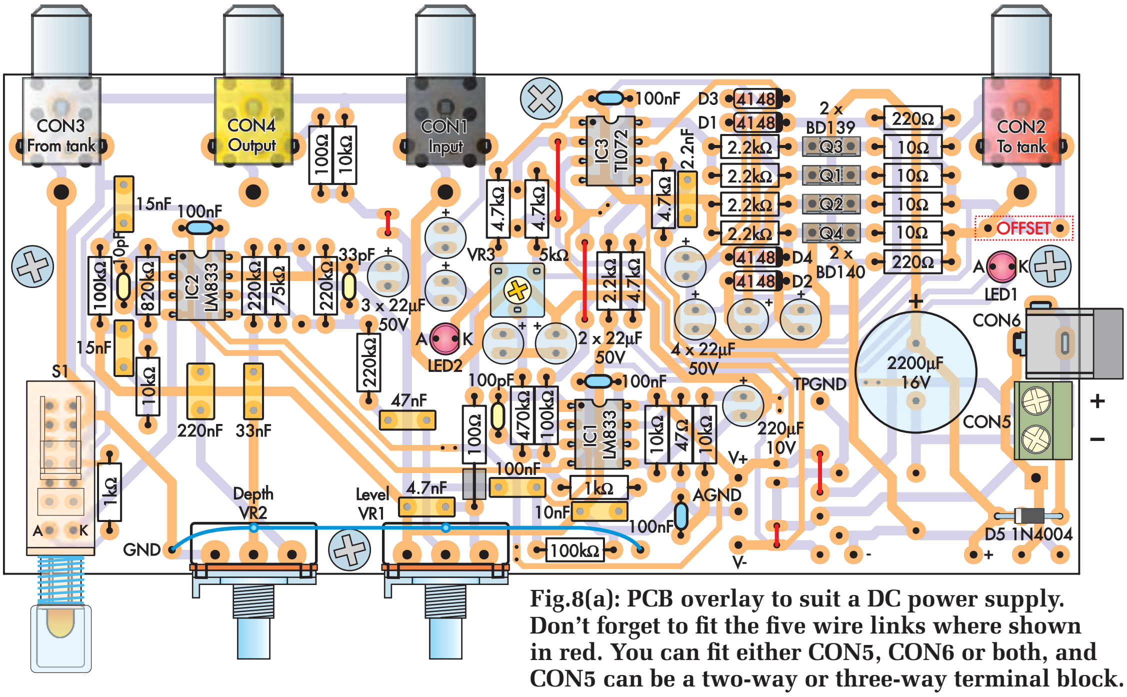


Fig.8(a): PCB overlay to suit a DC power supply. Don't forget to fit the five wire links where shown in red. You can fit either CON5, CON6 or both, and CON5 can be a two-way or three-way terminal block.

All four transistors are fitted with their metal tabs facing towards the bottom of the board as shown; if you're unsure, check the photo.

You can now solder the large electrolytic capacitor(s) in place; the DC supply version has one, located as shown in Fig.8(a), while the AC supply version has two. In all cases, the longer (+) lead goes towards the top of the board, as shown.

The last components to fit to the PCB are potentiometers VR1 and VR2.However, before installing them you must do two things. First, clamp each pot in a vice and file off a small area of passivation on the top of the body, allowing you to solder the ground wire later on.

Second, figure out how long you need the shafts to be to suit your amplifier and cut them to length. Make sure they're still long enough so that you can fit the knobs later!

Now solder the two pots to the board, ensuring that the 10kΩ pot (VR2) goes on the left side and then insert one end of a 100mm length of tinned copper wire in the pad marked 'GND', just to the left of VR2, and solder it in place. Next, bend the wire so it contacts the top of the two pot bodies and then solder it to the free pad to the right, as shown in Fig.8, and trim off the excess.

Now it's just a matter of soldering this ground wire to the areas where you scraped away the passivation from VR1 and VR2. Note that you will need to apply the soldering iron for a few seconds for the metal to get hot enough for solder to adhere.

Testing and set-up

The first step is to apply power and check the supply voltages. If you've fitted sockets, leave the ICs off the board for the time being. Having said that, if you have configured the board for a DC supply, plug in LM833 op amp IC1 (taking care with its orientation).

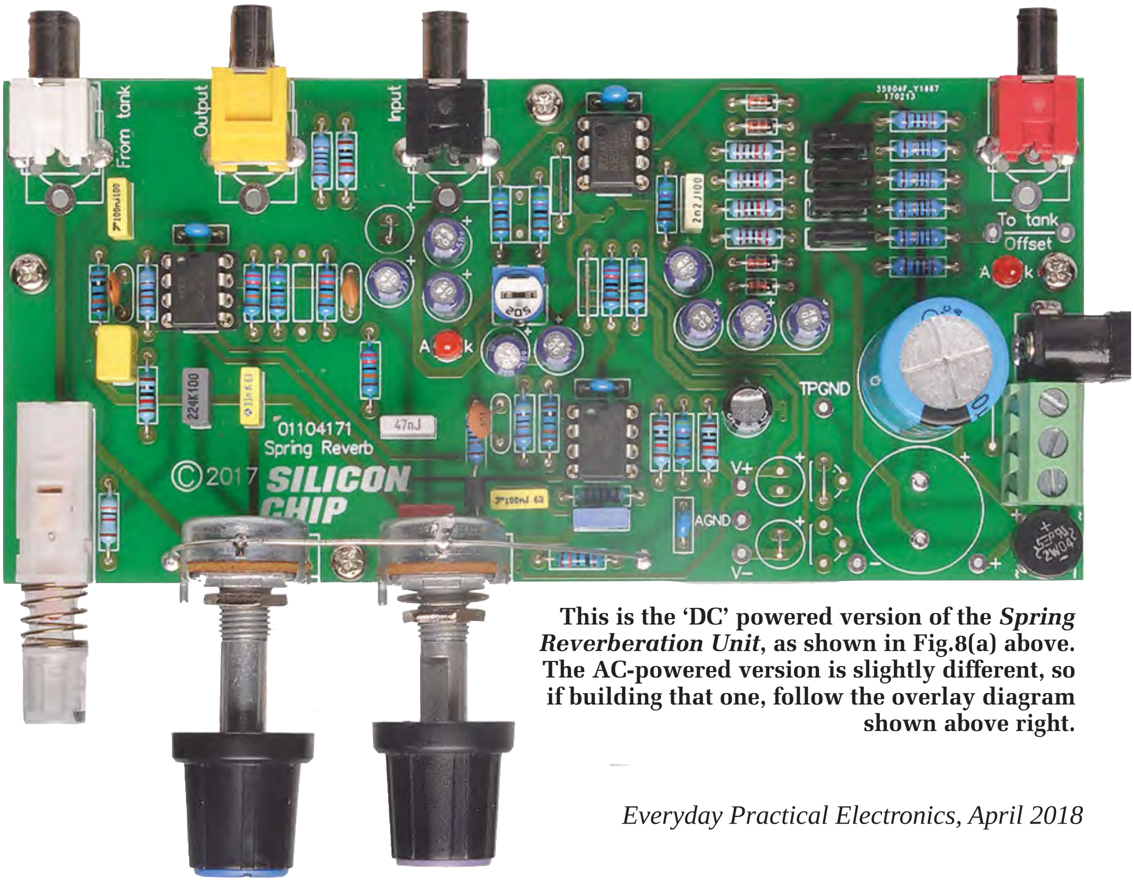
Apply power and check that the voltages at the five specified test points are close to the values given in Table 1 (on the circuit diagram).

Voltage variation on the '+' and '–' test points can be expected to be fairly large, possibly a couple of volts either side of those given. Voltages at V+ and V– should be within about 250mV of the optimal values, while for DC supplies, the voltage at AGND should be almost exactly half that at V+.

If you've fitted sockets, cut power and plug in the remaining ICs. Don't get IC3 (TL072) mixed up with the other two ICs, which are LM833s. In each case, the pin 1 dot must go towards the top edge of the PCB, as shown in Fig.8. Re-apply power for the remaining steps.

Measure the voltage between the two test points labelled 'OFFSET' in the upper-right corner of the PCB. You should get a reading below 100mV. If not, switch off and check for soldering problems or incorrect components around IC3a and IC3b. Assuming the reading is low, slowly rotate trimpot VR3 and check that you can adjust it near zero. It should be possible to get the reading well under 1mV.

If you have appropriate cables or adaptors, you can now do a live signal test. Use a stereo RCA/RCA or



This is the 'DC' powered version of the *Spring Reverberation Unit*, as shown in Fig.8(a) above. The AC-powered version is slightly different, so if building that one, follow the overlay diagram shown above right.

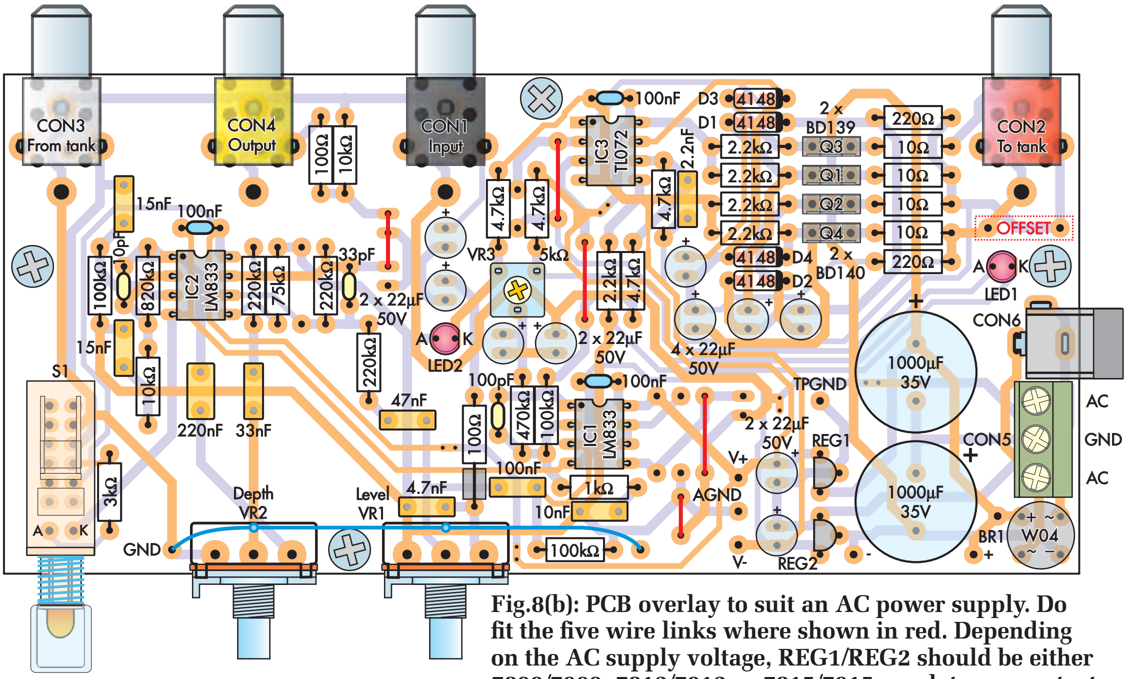


Fig.8(b): PCB overlay to suit an AC power supply. Do fit the five wire links where shown in red. Depending on the AC supply voltage, REG1/REG2 should be either 7809/7909, 7812/7912 or 7815/7915 regulators – see text.

RCA/3.5mm-plug cable to connect a mobile phone, MP3 player or other signal source to CON1. Turn VR1 and VR2 fully anti-clockwise. Use a cable with RCA plugs at one end and a 3.5mm stereo socket at the other end to connect a pair of headphones or earphones with a nominal impedance of at least 16Ω (ideally 32Ω or more) to CON4.

Power up the board, start the signal source and slowly advance VR1. You should hear the audio signal passing through the unit undistorted.

Now you can use a stereo RCA/RCA lead to connect the main board to the tank, via CON2 and CON3, matching up the labels on the board with those on the tank.

The tank should be placed on a level surface with the open part facing down. Continue listening to the signal source, then advance VR2. You should hear the reverb effect. If you're unsure, pause the audio source and you should continue to hear audio for several seconds until the reverb dies out.

That's it – the *Spring Reverberation Unit* is fully functional.

Installation

The tank should be installed with the open end down because the spring suspension is designed to work optimally in that position. Use the four corner holes to mount it since the tank is microphonic and these are designed to provide some isolation to prevent bumps from upsetting the springs too much. It would probably be a good idea to add extra rubber grommets under each spacer and avoid compressing them too much, for extra isolation.

As for mounting the PCB, you have three options. Option one is to mount it somewhere on the front panel of the amplifier so that switch S1 and potentiometers VR1 and VR2 are easily accessible. You then simply connect it to the tank using a stereo RCA/RCA lead.

If the panel it's mounted on is thin enough, it can be held in place using the two potentiometer nuts, although it would be a good idea to attach a small right-angle bracket to the mounting hole between the two pots, on the underside of the board via an insulating spacer, to provide a third anchor point on the panel.

The second possibility is to fashion a bracket from a sheet of aluminium with four holes drilled in it, matching the mounting holes in the board, with the side near the front of the board bent down and additional holes drilled in this flange for attachment to the front panel of the amp. You can then use self-tapping or machine screws to attach this bracket to the amp and then the board to the bracket. For bonus points, earth the aluminium bracket back to the GND pad on the PCB, to provide some shielding.

The third possibility is to leave S1, VR1 and VR2 off the board and mount it on top of the tank itself. We suggest using a long insulating spacer attached to one of the free holes on the tank's flange, supporting the PCB via the front or rear mounting hole, with a liberal application of thick double-sided foam tape on top of the tank to support the PCB.

You will need to trim the component leads carefully to make sure they can't poke through the foam tape and short on the top of the tank. In fact, it would be a good idea to silicone a sheet of plastic on top of the tank before applying the tape to provide extra insulation.

You would then mount S1, VR1 and VR2 wherever suitable and connect them back to the board using twin-core shielded cable for VR1 and VR2 (with the shield to the left-mount [ground] pin in each case). For the connections to S1, use regular shielded cable with the shield wired to the pin connected to ground and the central conductor for the audio pin, and a section of ribbon cable for the LED connections.

Using it

Using the *Spring Reverberation Unit* is straightforward. Push S1 in to enable reverb and push it again so it pops out to disable reverb. When reverb is enabled, S1 will light.

Adjust VR1 to give a near-maximum output level without clipping and then tweak VR2 until you get the desired reverberation effect.

With VR2 fully clockwise, the effect is overwhelming; you will probably find it most useful somewhere between 10 o'clock and 2 o'clock.

There are no controls on the spring reverb tank itself, just input (red) and output (white) RCA socket. All controls are on the PCB for this project.

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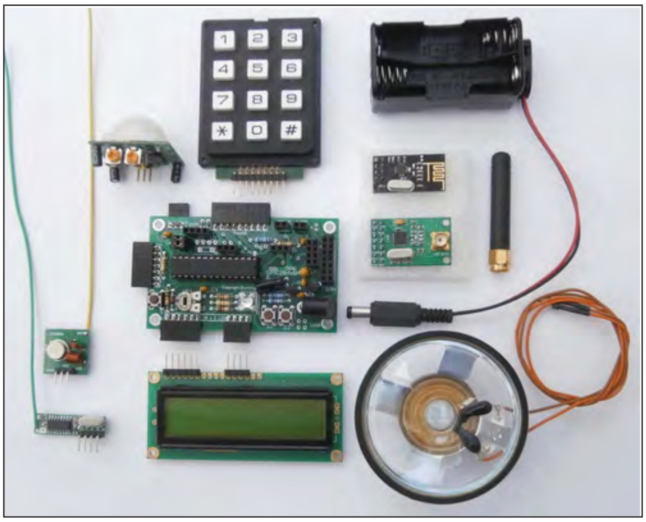
PIR and RF Data PIC Training Course 2nd Edition

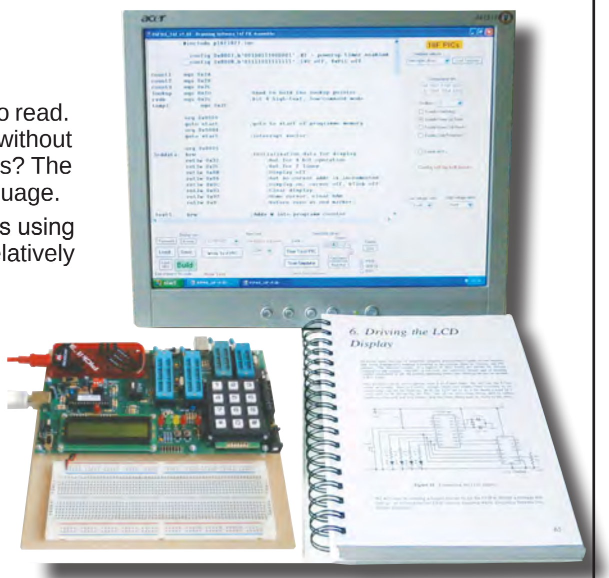
by Peter Brunning



If you want to expand your knowledge of radio communications this course will excite your interest. It centers around our new GPIC28rfv5H general purpose PIC circuit. This has sockets to fit all the attachments shown in the picture. Clockwise from 10 to 10 - PIR motion detector, keypad, nRF24L01+ 2.4Ghz transceiver, nRF905 433/868/915Mhz transceiver, DC power lead, loud speaker, LCD, simple 433Mhz receiver, simple 433Mz transmitter. The GPIC28rfv5H can be programmed using a Brunning Software programmer with BSPWA or using a PICkit3 with MPLAB-X.

We start with simple experiments to understand the GPIC28rfv5H. Run the RGB LED, write text to the LCD and use the keypad to input numbers. Then we start the PIR and radio frequency experiments using two GPIC28rfv5H. We use the simple Tx and Rx to understand the makeup of the data stream. Then we use the 433Mhz transceiver to create a movement detector system with a central receiver displaying the data. We repeat the experiments using the 2.4Ghz transceiver. Then we build a 2.4Ghz yagi antenna to measure the radiation around us. Microwave oven is top, mobile and wideband not far behind. We scan the nRF24L01+ across its band as a signal generator to get a reference. 238 page book 240 × 170mm wirobound to open flat, two sets of essential components, and PCB with microwave diode fitted for yagi, £145 including UK carriage. See website for details.

The Brunning Software P955H PIC Training Course



Imagine giving a Thomas Hardy book to a child who has just learnt to read. The child cannot understand the book so we create a new language without the complications. How absurd! So why do it with computer languages? The correct technique is to avoid the complications not to rewrite the language.

The first book teaches absolute beginners to write PIC programmes using assembler which is the natural language of the PIC. We use a relatively simple 8 bit PIC, simple programming techniques, and inspire the reader with immediate success. In the space of 24 experiments two projects and 56 exercises we work through from absolute beginner to experienced engineer level. The second book introduces the C programming language for 8 bit PICs in very simple terms. The third book teaches Visual C# programming for the PC so that we can create PC programmes to control PIC circuits.

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The complete P955H training course is £254 including P955H training circuit, 4 books (240 × 170mm 1200 pages total), 6 PIC microcontrollers, PIC assembler and programme text on CD, 2 USB to PC leads, pack of components, and carriage to a UK address. To programme 32 bit PICs you will need to plug on a PICkit3 which you need to buy from Microchip, Farnell or RS for £43.

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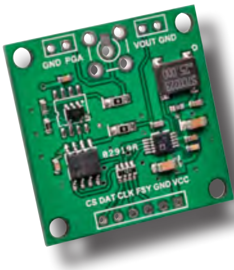
Touchscreen DDS Signal Generator



It can produce sine, triangle or square waveforms from 1Hz to 10MHz, with ±0.005% frequency accuracy; plus, it also has a sweep function. Its touchscreen LCD makes it very easy to drive and you can use it for audio or RF applications.

by Geoff Graham

This project combines a low-cost DDS function generator module with our touchscreen *Micromite LCD BackPack* module (first described in the May 2017 issue) to create a remarkably capable signal generator at a low price. It can generate sine, triangle and square wave signals from 1Hz to 10MHz, and you can specify that frequency with 1Hz resolution.

The Direct Digital Synthesiser (DDS) function generator module produces the actual waveforms, while the Micromite controls it and provides an easy-to-use graphical user interface (GUI).

As well as generating the basic waveforms, this unit can also act as a sweep generator, allowing you to test the frequency response of filters, speakers, IF (intermediate frequency) stages (in superheterodyne radios) and more.

Other features include an adjustable output level, selectable amplitude modulation for the sinewave output and a selectable log/linear function for the frequency sweep.

Many would consider a signal generator to be the next most useful tool to have on a workbench after the multimeter and oscilloscope. While this device will not compete with a £1000 synthesised signal generator, it does provide the basics at a tiny fraction of the cost.

The DDS function generator module is fully assembled and can be purchased for about £10 on eBay or AliExpress. Combined with the Micromite BackPack (which uses fewer than a dozen components), you can build the whole project in under an hour and without breaking the bank.

Analog Devices AD9833

The AD9833 waveform generator IC is the heart of the signal generator module used in this project. It uses DDS to generate its output.

Normally, it is difficult to digitally generate a relatively pure, variable frequency sinewave. Even the best Wein bridge (analogue) oscillators are notoriously difficult to stabilise and cannot be controlled over anywhere near the range of frequencies that this DDS unit can produce.

DDS involves a high-speed digital-to-analogue converter along with a ROM lookup table, a phase accumulator and possibly digital interpolation to produce a relatively pure, variable frequency waveform.

The waveform shape can be changed by using a different lookup table or using a reprogrammable lookup table. In other words, DDS is somewhat similar to digital audio playback from a computer or compact disc, but it normally operates at a much higher frequency.

For those who wnt to study the DDS chip in detail, we'll have a separate article on the AD9833 DDS IC and modules that are based on it in the June issue of *EPE*.

Frequency precision

Because the AD9833 module uses a crystal-controlled oscillator to produce the sample clock, the precision of the output frequency is determined by the precision of the crystal.

With the specified module, this is better than ±50ppm (our prototype achieved about ±10ppm). This also means that calibration will not be required and the frequency will not drift with time.

For example, if you set the output to 1MHz, you can expect it to typically be between about 999.999kHz and 1000.001kHz, or in the worst case, between 999.995kHz and 1000.005kHz.

Another benefit of DDS is that the phase of the output will not change when the frequency register is updated, and this in turn means that the output waveform will not have a glitch at the time of the change. This is vital for generating sweeps as it allows the frequency to be changed smoothly from one end of the sweep range to the other.

Because the waveform is digitally created with 1024 steps for each sinewave quadrant, the output is not perfectly smooth. The resulting harmonic distortion means that it is not quite good enough for noise or distortion measurements; its signal-to-noise ratio is about –60dB and its total harmonic distortion is typically 0.05%.

Having said that, it is more than adequate for general purpose tasks and

Features and specifications

General

Frequency accuracy: ±50ppm
Power supply: 4.5-5.5V DC at 350mA maximum
Output level: 10mV to 3V peak-to-peak (~3mV to ~1V RMS), 20Hz to 1MHz

Sinewave mode

Frequency: 1Hz to 10MHz with 1Hz resolution
Output level: as above up to 1MHz, reducing to 0.8V peak-to-peak at 10MHz
Amplitude modulation: on/off (1kHz square wave)

Triangle wave mode

Frequency: 1Hz to 1MHz with 1Hz resolution

Square wave mode

Frequency: 1Hz to 1MHz with 1Hz resolution

Sweep mode

Waveform: sinewave only
Frequency start/stop: 1Hz to 1MHz with 1Hz resolution
Sweep period: 50ms, 100ms, 500ms, 1s, 2s
Sweep law: linear or exponential
Trigger output: 250µs positive pulse at start of sweep

the ability to quickly and accurately set the output frequency makes it a pleasure to use.

DDS module with gain control

The output of the AD9833 IC is about 0.6V peak-to-peak, so the function generator module that we are using includes a high-bandwidth amplifier based on the AD8051 rail-to-rail op amp. This can drive low-impedance loads (eg, 50W) and provide higher output levels (up to 3V peak-to-peak).

To control the gain of the output amplifier, the module uses a Microchip MCP41010 8-bit digital potentiometer, which is under control of the Micromite (along with the AD9833). The bandwidth restrictions of the MCP41010 potentiometer result in a reduction in the output signal level above about 2MHz.

The output is still good for up to 10MHz, but the signal level for sinewaves will be reduced and the triangle and square waves will look more like sine waves, so we have specified both of these up to 1MHz maximum.

Micromite LCD BackPack

As with a number of our recent projects, this one is based on the *Micromite LCD BackPack* and relies on the touchscreen interface on the LCD panel to set the frequency and output levels – there are no switches or knobs.

The program is written in BASIC and because it is stored in plain text, you can see how it works and if you have the inclination, modify it to suit your personal preferences. For example, you can easily change the colours or add a special feature.

The *Micromite LCD BackPack* was described in the May 2017 issue of *EPE* and uses fewer than a dozen components. If you're reasonably experienced, you can build it in around half an hour. It includes a 3.3V regulator, the 28-pin Micromite PIC32 chip and touch-sensitive LCD screen. A complete kit is available from **micromite.org** – your one-stop shop for all things Micromite.

Note that if you want to try out the BASIC program for this project, you can do it on any Micromite with an ILI9341-based LCD panel connected; you do not need a DDS function generator module.

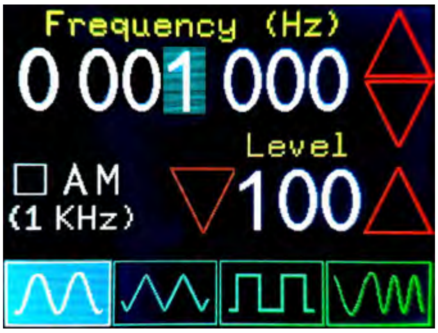
This is because the Micromite only sends commands to the AD9833 and MCP41010; it does not look for a response (and neither chip provides one anyway). So it won't know the difference; you simply won't get any signal output.

Driving it

In operation, the signal generator is quite intuitive, with everything controlled via the colourful touchscreen LCD panel. Probably the best way to appreciate this is by looking at the screen shots.

At the bottom of every screen are four touch-sensitive icons which are used to select the operating modes: sine, triangle, square wave and sweep. Touching one of these will immediately switch to that mode.

Starting with the sinewave mode (shown in Screenshot 1), the frequency



Screenshot 1: this is the screen displayed for a sinewave output. The frequency can be changed by selecting a digit to change and touching the red up/down buttons. The signal level (expressed as a percentage of full scale) can be similarly adjusted. The check box marked AM will enable a 1kHz square wave amplitude modulation.

is adjusted by touching the red up/down buttons on the right of the frequency display. The least-significant digit that you want to change can be specified by touching that digit and it will then be highlighted in blue.

A single touch on either the up or down buttons will increment or decrement the frequency, but if you hold the button down, the frequency will increment or decrement with increasing speed.

While you are adjusting the display in this way, the output frequency will follow in real time so it is easy to scan through a range of frequencies to find the one that you want.

If you want to simply jump to a specific frequency, you can touch and hold a digit on the display and an on-screen numeric keypad will pop up, allowing you to directly key in the frequency that you want (see Screenshot 2).

Touching the SAVE button on this keypad returns to the main screen with that frequency set, while the DEL button will delete the last digit entered.

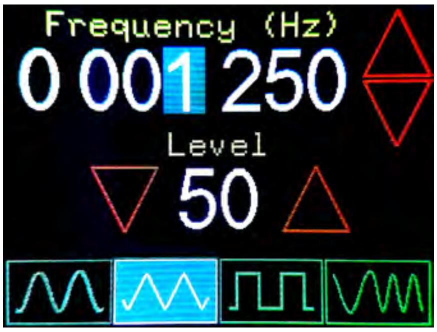
The process to adjust the signal level is similar, although you do not need to select a digit as the up/down buttons will always change the least significant digit.

Touching a digit in the level display will also take you to a numeric keypad where you can enter a specific level in the range from zero to 100% of full scale (about 3V peak-to-peak).

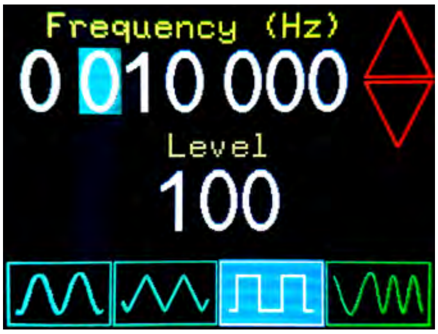
The sinewave screen has a check box for turning on or off amplitude modulation at 1kHz. This simply modulates the output with a 1kHz square wave and is useful for signal tracing in AM radios, both broadcast and shortwave, up to 10MHz.

The triangle waveform screen is similar to sine except that it does not provide an AM facility (see Screenshot

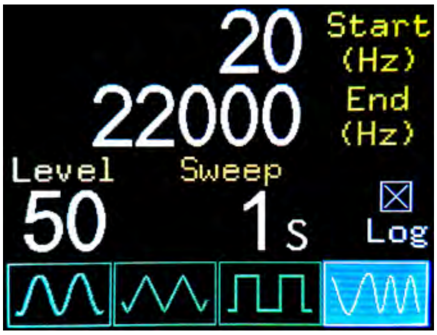
Screenshot 2: you can enter a precise frequency or signal level by touching and holding the frequency or level display. This keyboard will then appear so you can enter the value. The DEL key deletes the last number entered and the SAVE button saves the value and returns to the main screen.



Screenshot 3: the screen for generating the triangle waveform output is similar to that used for sinewaves. Along the bottom of the screen, the four touch sensitive icons are used to select the four operating modes – sine, triangle, square wave and sweep.



Screenshot 4: the DDS module does not allow you to change the level of the square wave output so this is fixed. Frequency selection is the same as the other modes – the frequency is changed by selecting the least-significant digit to change and touching the red up/down buttons.



Screenshot 5: the sweep output screen allows you to select the start frequency, end frequency, signal level, the sweep time and whether an exponential sweep is required. Touching entries like the start frequency makes a numeric keypad appear so you can key in the value that you want.

3). The square wave screen (shown in Screenshot 4) is also similar to the other two except that you cannot change the signal level (the MCP41010 digital potentiometer is not suitable for attenuating square waves).

All the changes that you make, including the waveform selection, are automatically saved in non-volatile memory and are recalled on power up. This means that when you turn on the signal generator, it will start up with exactly the same settings that you were using the last time.

Sinewave sweep

The sweep screen (Screenshot 5) uses a different screen layout. To select the start and end frequencies, you simply touch the frequency that you need to change and enter the specific frequency on the pop-up numeric keypad. You can select any frequency that you wish – you could even sweep all the way from 1Hz to 10MHz if you wanted to.

The output level is selected in a similar way, just touch the level display and a numeric keypad will pop up allowing you to enter that setting.

The sweep period works slightly differently; it will change every time you touch it, allowing you to step from a 50ms sweep time up to two seconds before wrapping around to 50ms again. Normally, the frequency sweep is performed in a linear manner with time, but you can select an exponential (ie, inverse log) sweep with the 'Log' check box.

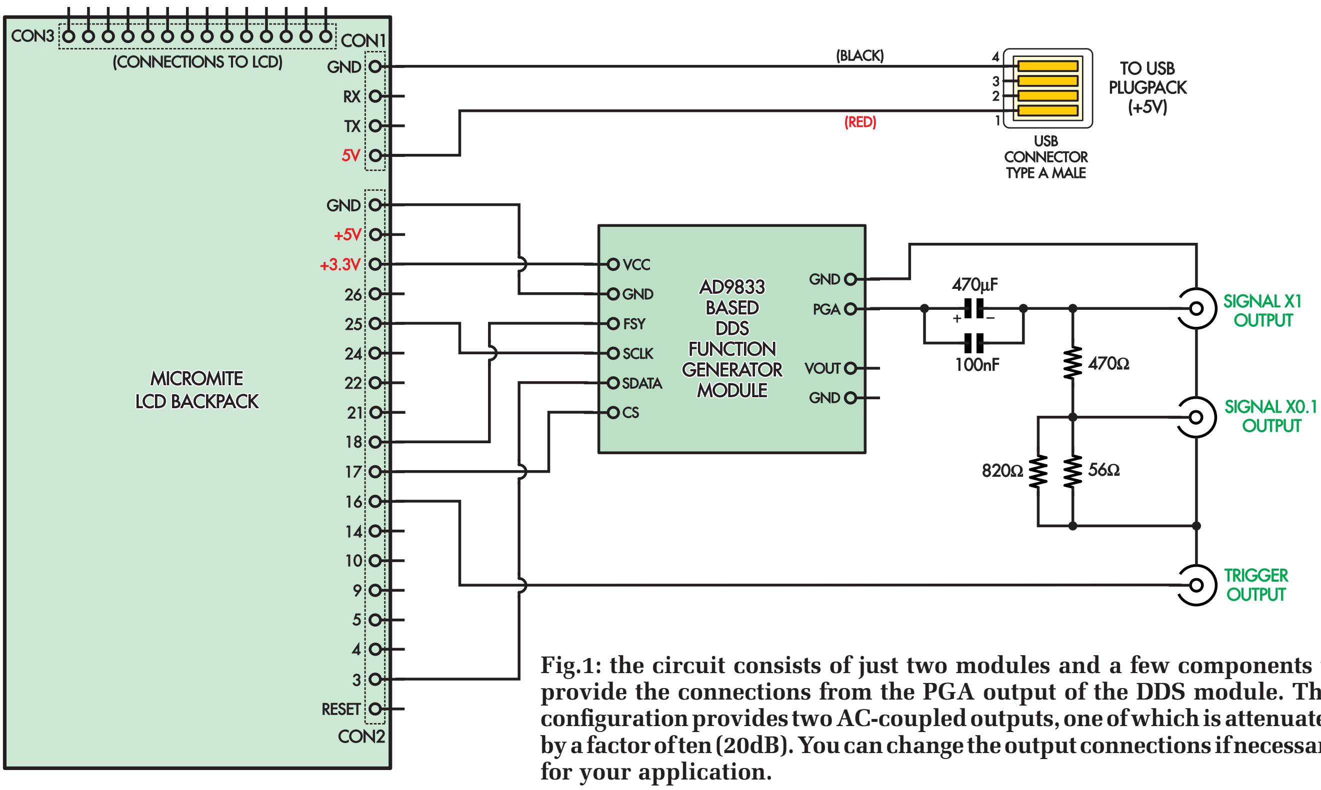


Fig.1: the circuit consists of just two modules and a few components to provide the connections from the PGA output of the DDS module. This configuration provides two AC-coupled outputs, one of which is attenuated by a factor of ten (20dB). You can change the output connections if necessary for your application.

With a linear sweep, it would take twice as long to go from 200Hz to 400Hz as it would from 100Hz to 200Hz. With an exponential sweep, it takes the same amount of time to go from 200Hz to 400Hz as it does from 100Hz to 200Hz, as both require a doubling in the output frequency.

This sounds more natural to human ears as doubling the frequency is equivalent to going up by one octave on a musical instrument.

The swept output is always a sinewave and at the start of the sweep, the Micromite generates a 250µs positive-going pulse on its pin 16 output, which is connected to the trigger output socket.

This signal can be used to trigger an oscilloscope so that it can lock onto the start of the sweep cycle for analysing the frequency response of a circuit or device.

Circuit details

The signal generator essentially consists of just two packaged modules connected together, hence the circuit is quite simple – see Fig.1.

There are six connections between the *Micromite LCD BackPack* and the DDS function generator module. These are for power (+3.3V and ground), the serial data lines to the DDS (DAT and CLK) and two additional signals: FSY, which when pulled low selects the AD9833 DDS chip as the recipient of serial data and CS, which similarly is pulled low when the MCP41010 digital potentiometer is being sent a command via the serial bus.

The DDS module can run from 5V, but we are using the regulated 3.3V supply rail from the *Micromite LCD BackPack* to avoid possible problems caused by potential noise from the output of a 5V USB charger. This noise can upset the AD9833 and MCP41010 ICs, which do need a clean power supply.

There are two outputs on the DDS module. One is labelled V_{OUT} and this is a fixed direct-coupled output from the AD9833 waveform generator (about 0.6V peak-to-peak). But we are using the PGA (programmable gain amplifier) output of the module and it is AC-coupled to two RCA sockets, one at the full output level and the second attenuated by a factor of 10.

Combined with the MCP41010 digital potentiometer in the DDS module, this gives an output range from 10mV to 3V peak-to-peak (equivalent to 3.5mV to 1.06V RMS).

The use of the 470µF coupling capacitor means that the output is usable to below 10Hz even into a 600W load. The parallel 100nF capacitor caters for higher frequencies, essentially bypassing any internal inductance of the larger capacitor.

The output from the module will swing from a little above ground to some maximum voltage determined by the MCP41010 digital potentiometer, below 3.3V.

If you will be primarily using the signal generator for testing digital circuits, you might prefer to dispense with AC-coupling and use DC coupling instead. You could even install a toggle switch to switch between these modes.

Similarly, you could use a switch to select different output attenuation levels if you wish. And you might consider using BNC sockets instead of the RCA sockets that we used.

The trigger output has simply been connected to output pin 16 of the *Micromite LCD BackPack*. You may wish to include a low-value series resistor (eg, 1kW or less) to protect the *Micromite LCD BackPack* from static discharge or accidental application of voltage to this terminal; it should not affect the trigger signal greatly.

Purchasing the correct module

If you search eBay or AliExpress for 'AD9833', you will find plenty of DDS modules (over 100 hits). However, you must be careful to purchase the correct module – there are a number of variations available and the firmware is written specifically to suit the module that we have pictured here.

It will probably not work with other modules, even if they also use the AD9833. So, check that the photograph matches perfectly and do not purchase anything different. Here is one which should be suitable: **http://bit.ly/2GpJHXD**

Many of the photos on eBay show the module with the I/O connector and SMA output socket already soldered to the board, but all the vendors that we purchased from supplied these two components separately. We did not find the SMA socket necessary in our application, but you could fit it if you want to.

Construction

Construction mostly involves assembly of the *Micromite LCD BackPack* and then mounting and connecting the DDS function generator module.

The *Micromite LCD BackPack* PCB is silk-screened with the component placement and values, so it is simply a case of populating the board and plugging it into an ILI9341-based LCD panel.

We suggest you use the 2.8-inch version of the *Micromite LCD BackPack*, and this is fully covered in the May 2017 issue of *EPE*.

If you have a PIC32 chip that is already programmed with MMBasic firmware then you will need to set up the LCD panel for display and touch, then load the BASIC code into the chip using a serial console.A detailed explanation of how to do this is provided in the *Micromite User Manual* and the May 2017 issue of *EPE*.

However, if your PIC32 chip is blank, you can load MMBasic and the code for this project simultaneously by programming it with the file **SigGenerator.hex**, which can be downloaded from the *EPE* website (along with the BASIC code). You will need a PIC32 programmer such as the PICkit 3 or the cheap DIY PIC32 programmer described in the Nov '16 issue of *EPE*.

If you do not have such a device, you can simply purchase a fully programmed microcontroller from **micromite.org**

Regardless, if your chip is programmed with **SigGenerator.hex**, all that you need do is plug the chip into its socket and connect the DDS module and you are ready to go.

The only point that you need to be aware of is that the touch calibration in the above firmware was done with a standard LCD panel. However, yours might require re-calibration if it is significantly different from the one that we used.

This can be done by connecting a USB-to-serial converter to the console, halting the program with CTRL-C and running the calibration routine by issuing the 'GUI CALIBRATE' command.

For further information, see the May 2017 *Micromite LCD BackPack* article or the *Micromite User Manual* (which can be downloaded from Geof Graham's website: **http://geoffg.net**

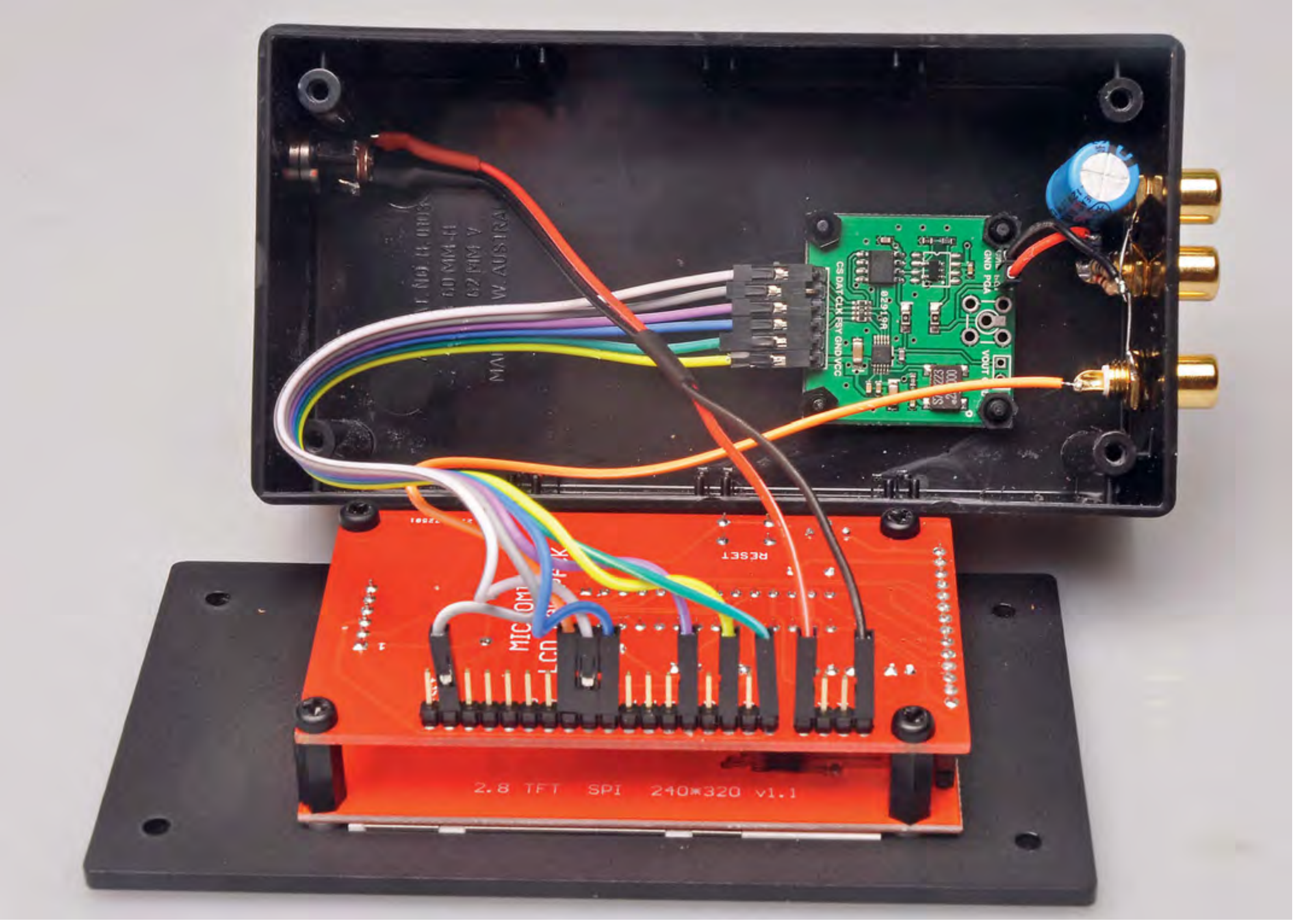
Putting it in a box

The *Micromite LCD BackPack* fits neatly into a standard UB3 plastic box, as we have done with similar projects based on the *BackPack*.

The easiest way is to use the laser-cut acrylic front panel which replaces the standard lid supplied with the box and is normally supplied with the kit. This provides a neat looking assembly with the display and *Micromite LCD BackPack* securely fastened.

You can purchase this panel from the *EPE PCB Service*, coded DDS Sig Gen Lid at **www.epemag.com** separately in either black, blue or clear colours.

Note that this panel is thicker than the lid supplied with the UB3 box, so the self-tapping screws supplied with the box may not be long enough. In that case, replace them with No.4 × 10mm self-tapping screws.



Interior view of the *Touchscreen DDS Signal Generator* showing the connections made from the *Micromite BackPack* to the module and internal connectors. You do not have to solder the extra through-hole components the way we did, as the UB3 jiffy box provides a fair bit of clearance.

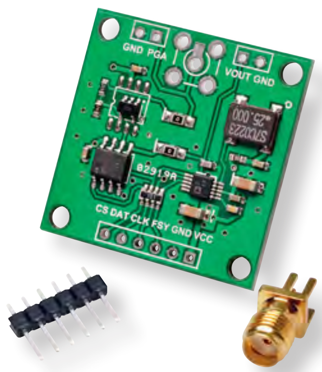
The first stage of assembly is to attach the LCD panel to the acrylic lid using an M3 × 10mm machine screw, a single M3 washer and an M3 × 12mm tapped spacer at each corner.

This arrangement ensures that the surface of the LCD sits flush with the acrylic lid. Then, the backpack should be plugged into the LCD and fastened by M3 × 6mm machine screws to each spacer. Details of the full assembly are shown in Fig.2.

The LCD and the *Micromite LCD BackPack* require a 5V power supply with a minimum capacity of 300mA. For this, you can use a 5V plugpack or a USB charger.

You can also find USB Type A to DC charging cables on eBay or AliExpress, which circumvents the need for cable rewiring. If you are using a plugpack, make sure that it is regulated and that its unloaded output does not rise above 5.5V, as this could cause damage.

For a USB charger, a suitable power cable can be made by cutting off one end of a standard USB cable (retaining the Type A connector on the other end) and soldering the free end to a



The *Touchscreen DDS Signal Generator* is based on this pre-assembled DDS function generator module which uses the Analog Devices AD9833 to generate the signal. It's amplified by an AD8051 high-speed op amp, and a Microchip MCP41010 digital potentiometer controls the gain.

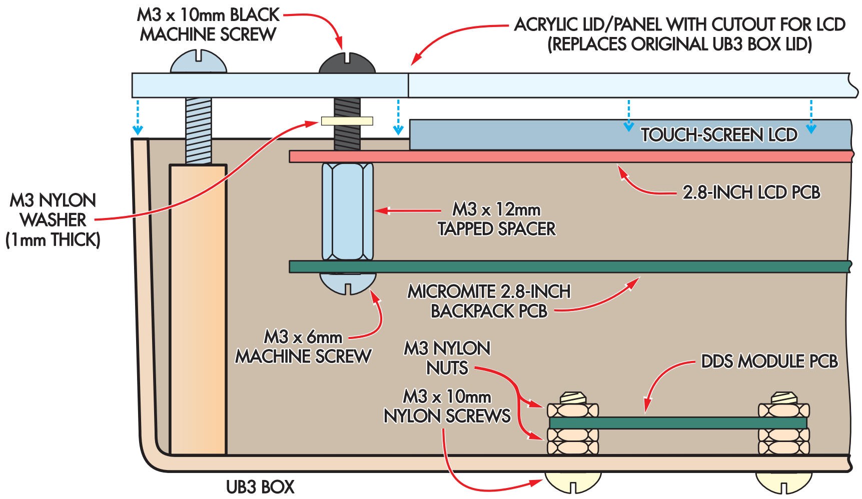


Fig.2: the DDS module is mounted in the bottom of the box using M3 machine screws, nuts and nylon nuts as spacers. By contrast, the *BackPack* is attached to the underside of the laser-cut lid. The wiring is not shown in this diagram.

suitable DC power plug. The red wire in the USB cable (+5V) should go to the centre pin of the plug and the black to the sleeve. The other two wires (the signal wires) can be cut short as they are not used.

A matching DC socket for incoming power can be mounted on the side of the UB3 box. Two flying leads from this socket should be fitted with female header sockets (also known as DuPont connectors) which fit over the BackPack's power header pins (CON1). Fig.3 illustrates the complete assembly.

The DDS function generator module can be mounted on the base of the UB3 box using four M3 machine screws and nuts. Use nylon M3 nuts as spacers between the base of the box and the module.

You need to select a spot for the module that will not foul the underside of the *Micromite LCD BackPack* PCB, particularly CON1 and CON2 which extend close to the bottom of the box.

Finally, connect flying leads from the module to the required pins on CON2 on the *Micromite LCD BackPack* and from the DDS outputs to the RCA (or BNC) connectors.

The most convenient method of mounting the output capacitor and resistors is to solder them directly onto the RCA/BNC connectors. Fig.4, overleaf, provides a convenient summary of all the connections to the DDS module.

We suggest that you wire up the connections from the module to the *Micromite LCD BackPack* using leads with female header sockets (DuPont connectors) at each end. These will simply plug onto the headers on both modules, which makes it easy to remove and/or replace the module if necessary.

Altronics have suitable pre-assembled leads (Cat P1017) as do Jaycar (WC6026) or search eBay or AliExpress for 'DuPont Jumper'.

Testing

Before connecting the DDS function generator module, confirm that the *Micromite LCD BackPack* is working correctly and has been programmed with the BASIC code. The testing procedure is described in the *Micromite User Manual* and also in the May 2017 issue.

Then it should simply be a matter of connecting the DDS module and checking its output. If it does not appear to be working, your first action should be to carefully re-check each connection. Then measure the voltage across the pins marked V_{CC} and GND on the module, which should give precisely 3.3V.

Remember that the module does not provide any feedback to the Micromite so the LCD might show the frequency, level etc and look like it is working but this does not mean that the module is actually alive and reacting to these commands (it is a one-way communication path).

If you have an oscilloscope or logic analyser, you can monitor the pins labelled FSY, CLK and DAT on the module.

Every time you change the frequency you should see a burst of data on these pins. Similarly, the pins labelled CS, CLK and DAT will show a burst of data when the signal level is changed. If these are not present, re-check the *Micromite LCD BackPack* and its connections.

A final test is to connect an LED with a suitable current-limiting resistor or an old fashioned moving-coil multimeter directly to the output of the module and set the signal generator to a 1Hz square wave.

You should see the LED or meter responding to the 1Hz output. If not, the simple option is to replace the DDS

Parts List

1 2.8-inch *Micromite LCD BackPack* module; see the May 2017 issue of *EPE* (kit available from micromite.org)
1 DDS function generator module with AD9833, AD8051 and MCP41010 ICs (see text and photos)
1 UB3 'jiffy' plastic box
1 pre-cut plastic lid to suit BackPack and UB3 box
1 USB charger plus USB cable with a male Type A connector on one end (alternatively, a USB Type A to DC connector charging cable)
OR
1 5V regulated plugpack
1 matching chassis-mount DC barrel socket
6 flying leads (120mm) with single pin female headers (DuPont connectors) on each end
5 flying leads (120mm) with single pin female headers (DuPont connectors) on one end and bare wire on the other
1 6-pin right-angle male header
4 No.4 × 10mm self-tapping screws
4 M3 × 10mm tapped nylon spacers
8 M3 × 10mm machine screws
4 M3 × 6mm machine screws
4 M3 nylon washers
12 M3 nylon nuts

Capacitors
1 470µF 16V electrolytic
1 100nF multi-layer ceramic

Resistors (all 0.25W, 5%)
1 820Ω 1 470Ω
1 56Ω

COMPETITION!

See Page 49 for a chance to win a Micromite BackPack!

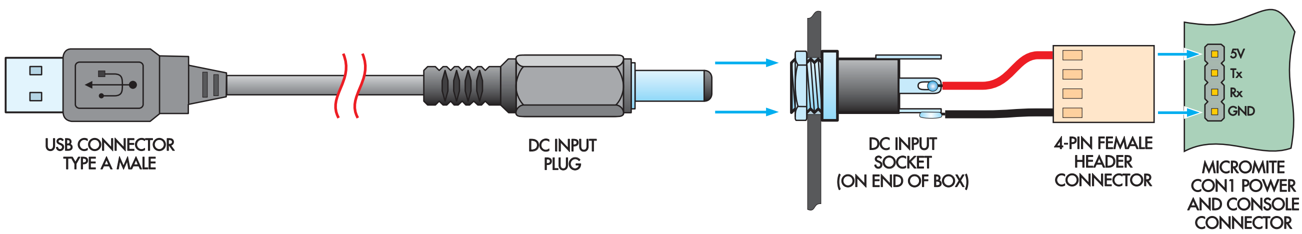


Fig.3: the *Touchscreen DDS Signal Generator* is powered from a standard USB plugpack charger. To make a suitable power cable, cut one end off a USB cable (maintaining the type A male connector at the other end) and solder the red wire to the centre terminal pin of a DC plug and the black wire to the outer barrel connection. The matching DC socket is mounted on the side of the UB3 box and is connected to CON1 on the *BackPack* PCB.

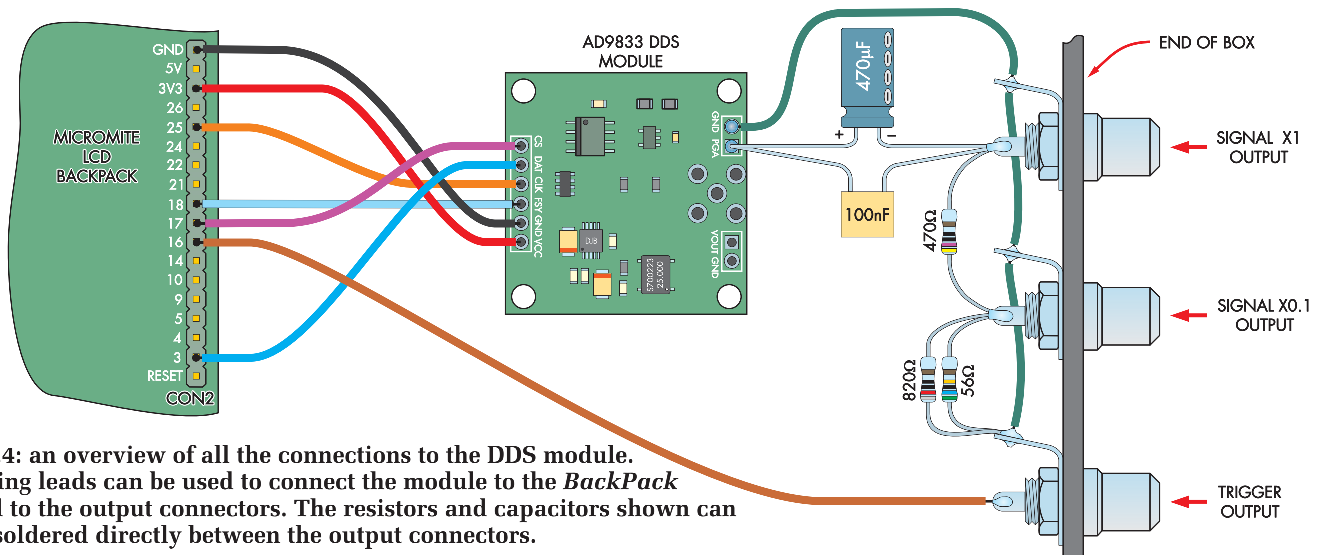


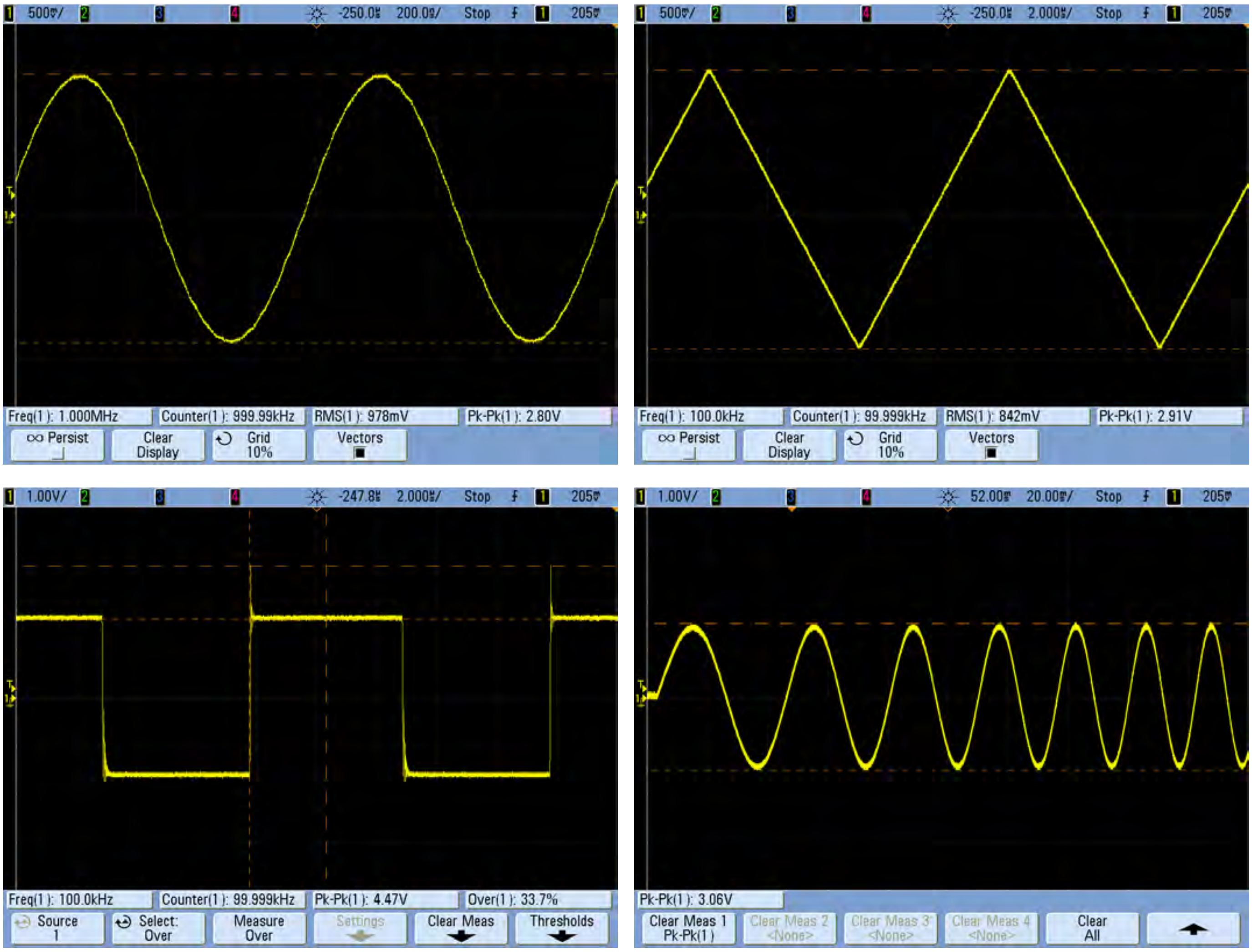
Fig.4: an overview of all the connections to the DDS module. Flying leads can be used to connect the module to the *BackPack* and to the output connectors. The resistors and capacitors shown can be soldered directly between the output connectors.

module. Scope 1-4 show waveforms that have been generated using the *DDS Signal Generator*.

Firmware updates for the Micromite and the BASIC software for the *DDS Signal Generator* will be provided on the author's website at: **http://geoffg.net/micromite.html**

Micromite parts

We strongly recommend you make **micromite.org** your first port of call when shopping for all Micromite project components. Phil Boyce, who runs **micromite.org**, can supply kits, programmed ICs, PCBs and many of the sensors and other devices mentioned in recent articles – in fact, just about anything you could want for your Micromite endeavours. Phil works closely with Geoff Graham and is knowledgeable about the whole series of Micromite microcontrollers.



Scope 1-4: These scope captures show typical output waveforms. The sinewave output is reasonably smooth despite being digitally created. There is some harmonic distortion, which means you cannot use this project for precise noise and distortion measurements, but it and the other outputs are quite suitable for general purpose tasks. The final scope capture shows a short sweep between 20Hz and 50Hz.

Improving your Arduino-based Theremin

By BAO SMITH

In the December 2017 issue we had a short article on building a no-frills Arduino-based digital Theremin. This month we show how to add a second sensor to the Theremin, which is used to control volume.

You can't really call something a 'Theremin' if all it does is alter pitch. So we decided to improve on the Theremin kit from Jaycar by adding a second ultrasonic sensor which is used to alter volume.

This extra HC-SR04 ultrasonic sensor is cheap and can also be bought from Jaycar (Cat. XC4442).

Adding the second sensor

The second sensor is aimed perpendicular relative to the first; and moving your hand closer to it increases the volume, decreasing it if you move away.

While the physical change to this kit is very simple, there is much more that needs to be altered on the software side to provide the volume-altering effect.

Because of the lack of space around the DIGITAL pins due to the pitch-controlling sensor being located there, we opted to plug the second sensor into the ANALOG pins.

Conveniently, the ANALOG pins on the Arduino Uno can be used as digital pins; however, when manipulating them, the pin number needs to be prefixed with 'A' – thus, A2 corresponds to ANALOG pin 2 on the board.

We have placed the additional sensor with V_{CC} on ANALOG pin 2, Trig on pin 3, Echo on pin 4 and GND on pin 5. We also slightly bent the 2-pin male header that the amplifier power supply connection was attached to so that the lead does not come into contact with the sensor.

As detailed in the December 2017 article, the pin locations of the new sensor can be altered (if necessary) by changing what is defined in the software. But it's easiest to use the same pins we have.

Then all that needs to be done is upload the new software to the board. The new software will still work with just one sensor, as shown last month, and can be downloaded for free from the *EPE* website.

Software

Once again, the software details are left to interested readers to explore. Instead, we will just go over some of the more important points. At the top of the **Ultrasonic_Theremin.ino** file there is a new macro called 'VOL_SENSOR', which is set to 1 by default.

When set to 1, the software will act as if both sensors are attached, and thus it attempts to request data from both sensors. If set to 0, the software functions as if only the pitch-controlling sensor is attached and thus it only polls one sensor.

The amplifier's audio signal level is determined by the value of the 8-bit OCR2B register, which can range between 0 to 255 inclusive.

Now that we have the additional sensor, a second distance measurement is computed (simultaneously with the first, to avoid slowing down the feedback loop). This distance measurement is then used to scale the sinewave value written to the OCR2B register, effectively attenuating the sound level depending on how far your hand is from the new sensor.

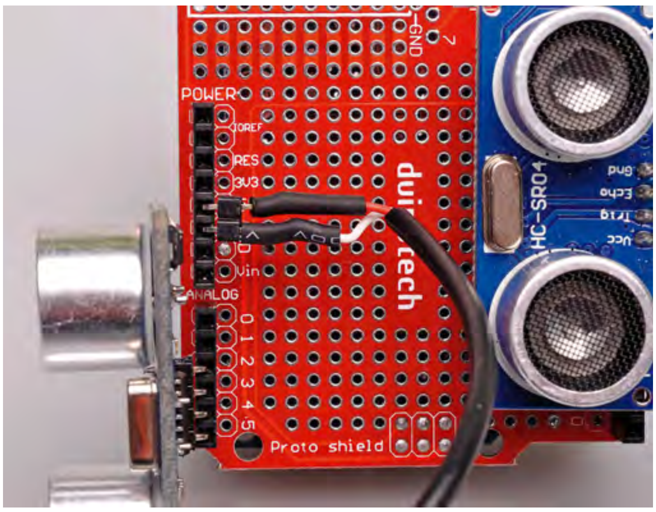
By default, the software uses the same MAX_DIST setting for both sensors to set their maximum detection range. If for some reason you wanted to use a different value for each sensor, you would need to modify the software.

The trickiest part of modifying the software to handle two sensors was the code to measure the distance for each simultaneously. This involves sending simultaneous trigger pulses, then waiting for both echo pulses to be received while separately timing the start and end of each echo, so that we can later subtract them and calculate the distance measured. We recommend that interested readers take a close look at this part of the source code to see how we did it.

Of course, one of the great things about Arduino is that you can download our software and easily experiment with making changes to see what effect they have.

More Arduino projects

If you're interested in building other Arduino projects, check out Jaycar's guides at: **www.jaycar.com.au/arduino**



The second ultrasonic sensor is fitted so that V_{CC} goes to ANALOG pin 2, while Trig goes to pin 3. Note that the amplifier power lead has been bent slightly so that there is better spacing between parts.

Measuring Temperature and Relative Humidity

Using Cheap Asian Electronic Modules – Part 4

The AM2302/DHT22 digital temperature and relative humidity (RH) sensing module provides about the simplest way to make a microcontroller project with temperature and RH-sensing capabilities.

by JIM ROWE

Low-cost modules capable of sensing and measuring both temperature and relative humidity (RH) have been available for a few years now.

Initially, these modules appeared as peripherals for Arduino and similar microcomputers, but they soon became an almost standard add-on for just about any micro-based project.

How humidity is measured

Relative humidity is the ratio of the amount of water vapour per volume of air at a particular temperature to the maximum amount of water which can be contained by that volume of air at that same temperature without condensation.

Another way to state this is that RH is approximately the ratio of the actual vapour pressure to the saturation vapour pressure. The saturation vapour pressure depends on the dew point temperature, which is the highest temperature for a given humidity level at which water vapour will condense and form dew.

This means that RH depends on three factors: the amount of water vapour in the air, air temperature and atmospheric pressure at the time of measurement.

Since the module described here measures both RH and temperature, if you assume a fixed barometric pressure (eg, at sea level it is typically close to 1 bar), you can compute the absolute humidity based on these two readings.

Just about all of these temperature/RH sensing modules are based on integrated digital sensors made by Chinese firm Aosong Electronics (based in Guangzhou), which also goes by the name MaxDetect Technology.

What's inside

Most modules currently available use their improved AM2302 sensor, which has alternative names: DHT22 or RHT03.

Aosong/MaxDetect say little about what's inside the AM2302/DHT22/RHT03, but mention that it contains a dedicated 8-bit microcontroller (see Fig.5), a temperature sensor and one RH sensor, the latter being based on a special polymer capacitor.

Curious to know more, I carefully cut away the slotted upper section of the plastic device body. All this achieved was to reveal the two sensors, fitted on the top of a very small PCB (18 × 14mm) which is potted inside the remaining part of the plastic body (see photo and Fig.6).

The polymer capacitor humidity sensor (Fig.1) works by measuring the relative change in the dielectric constant of the capacitor with varying humidity.

Since the change in value differs between capacitors, sensor calibration is required to provide accurate results.

A thermistor provides temperature sensing. The thermistor used is an NTC (negative temperature coefficient) type, made of a conductive material which decreases in resistance proportionally as the temperature rises.

The microcontroller measures the RH sensor capacitance and the thermistor resistance, and then converts the analogue readings to digital values.

This micro and a number of associated components are mounted on the underside of the PCB; we can't determine their exact configuration as it's impossible to remove the potting without destroying most of the circuit.

However, there is a YouTube video where someone has removed all the components from the device. Some of the pictures from that video are shown in this article, and the link to the video is at the end of the text.

Aosong/MaxDetect state that every AM2302 sensor is temperature compensated and calibrated in an accurate calibration chamber, during or after which the calibration coefficients are saved in the micro's one-time programmable memory.

Considering its low price, the claimed performance of the AM2302 is quite impressive. The RH measuring range is from 0 to 100%, with a resolution of 0.1% and an accuracy of ±2%, while the temperature measuring range is from –40 to +80°C with a resolution of 0.1°C and an accuracy of ±0.5°C. The long-term RH stability is rated as ±0.5% per year.

Fig.1: close-up of the humidity sensor, showing the two capacitor plates. Note the darker plate marked with red is much smaller than the gold one underneath.*

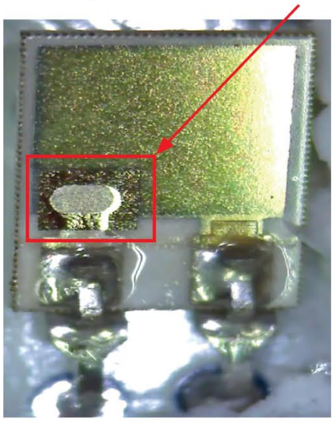
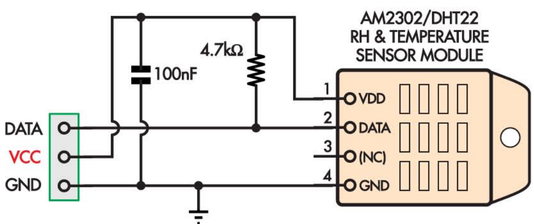
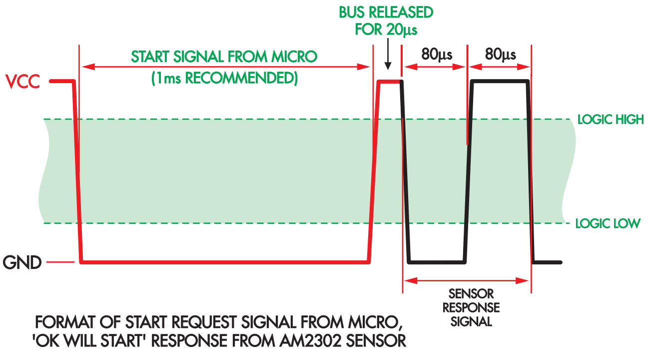


Fig.2 (below): complete connection diagram for the AM2302/DHT22 sensor module. The 4.7kΩ pull-up resistor allows for bidirectional communication with a single DATA pin.





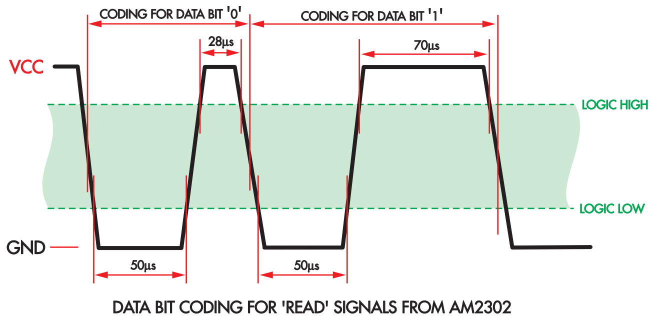


Fig.3: to wake the sensor from standby mode, the micro pulls the DATA line low for a minimum of 800µs and a maximum of 20ms. The DATA line then goes high for 20µs. This is regarded as a start request sent to the AM2302.

Fig.4: the micro differentiates between what type of bit it has received based on the pulse time; a data bit of value zero has a pulse time of 78µs while a one has a pulse time of 120µs.

The device is designed to run from 3.3-5.5V DC, with operation from 5V recommended. It has a nominal current drain of 1.5mA when measuring, or 50µA when in standby. It needs at least two seconds between measurements.

The AM2302/DHT22/RHT03 module measures only 25.1 × 15.1 × 7.7mm, while the PCB for the most common module using it measures 39 × 23mm – see our picture.

The sensor has four connection pins, although one is labelled 'NC' (no connection) in Aosong's data sheet.

As you can see from Fig.2, there's very little in a typical sensing module apart from the AM2302/DHT22/RHT03 device itself.

There are just two passive components on the board: a 100nF bypass capacitor from V_{CC} to ground and a 4.7kΩ pull-up resistor between the digital data bus line and V_{CC}.

The reason for that resistor leads us to discuss the way the device communicates with an external micro, over that single-wire bus.

How it handles data

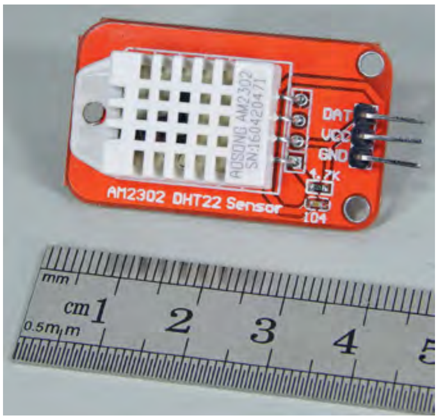
Although it's poorly explained in the AM2302 data sheet, here's the basic idea: when the DATA line is allowed to float at the logic high level (pulled high by the 4.7kΩ resistor), the sensor effectively sleeps in standby mode. To wake it up, the external micro must pull the DATA line down to logic low for at least 800µs, but no more than 20ms. In fact, they recommend that it be pulled down for 1ms.

Then the micro should release the DATA line, allowing it to float high again for about 20µs. This '1ms-low-followed-by-20µs-high' sequence is regarded as the micro sending a start request signal to the AM2302.

If the AM2302 responds to this wake-up call, it pulls the DATA line down to logic low for 80µs, and then allows it to float high again for another 80µs.

This is regarded as its 'OK, will start' response. This 'start request' and 'OK will start' sequence is shown in Fig.3.

Soon after this startup sequence, the AM2302 sends out its current measurement data as a sequence of 40 bits of data, grouped in five bytes – see Fig.7.



The module in question with the case still intact. The module has a fairly low profile, measuring only 7.7mm high.

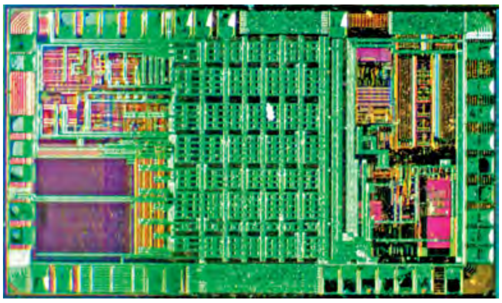


Fig.5 (above): the internal layout of the micro in the AM2302 sensor.*

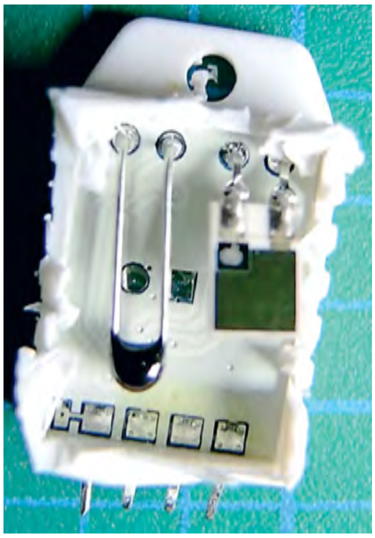


Fig.6 (right): the sensor module with the top of the case removed. The bead type sensor is an NTC themistor and to its right is the capacitive humidity sensor.*

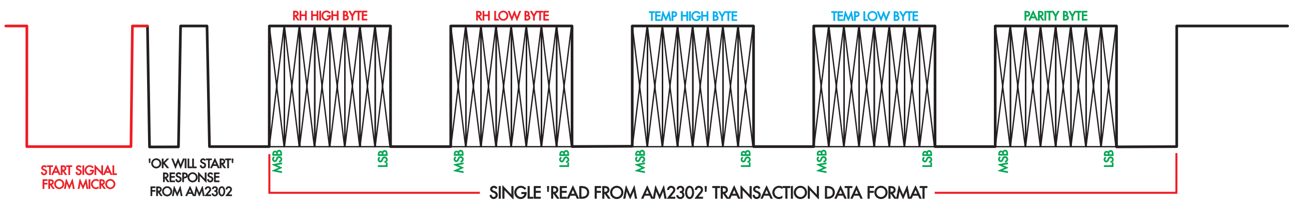


Fig.7: once there has been a start response from the sensor, the AM2302 sends out its measurement data in 40-bit sets. The first 16 bits is the relative humidity, the 16 bits after is the temperature and the final 8 bits are parity bits to pad the length of the data to 40 bits total.

The relative humidity reading is in the first two bytes (RH HIGH and RH LOW), followed by the temperature reading in the next two bytes (TEMP HIGH and TEMP LOW), and finally there's a checksum or parity byte to allow error checking.

All of these bytes are sent MSB (most-significant bit) first and LSB (least-significant bit) last.

It's also worth noting that both the RH and temperature readings have a resolution of 16 bits.

While this single-wire-bus transaction may look fairly straightforward, it isn't quite that simple – because of the special encoding that Aosong uses for the data bits themselves.

As shown in Fig.4, a binary zero is coded as a logic low of 50µs followed by a logic high of 28µs, whereas a binary one is coded as the same logic low of 50µs, but followed by a logic high of 70µs.

So both a zero and a one begin with a logic low lasting for 50µs, but a logic high that follows lasts for only 28µs in the case of a zero rather than 70µs in the case of a one.

As a consequence, data bits with a value of 0 last for a total of 78µs, while those with a value of 1 last for 120µs. So the time taken by each of those data bytes as shown in Fig.7 will not be fixed, but will vary, depending on the data bit values.

For example, a byte consisting of all zeroes (00000000) will last for only 624µs, while a byte of all ones (11111111) will last for 960µs. So in practice, the duration of each data byte will vary between 624 and 960µs.

The micro connected to the AM2302 needs to take this rather unusual coding system into account when it decodes RH and temperature data.

How it's used

You shouldn't have to worry about decoding the AM2302 measurement data yourself, because many people have already worked it out for most of the popular microcontroller.

For example, if you want to hook up an AM2302-based module to a Micromite, Geoff Graham has already solved this problem and provided a special command in his MMBasic programming language. It looks like this:

```
HUMID pin, tVar, hVar
```

Where HUMID is the command keyword and 'pin' is the micro's I/O pin to which the module's DATA line is connected.

'tVar' is the name of the floating-point variable you want to receive the returned temperature (in °C) and 'hVar' is the name of a second floating-point variable to receive the returned relative humidity (as a percentage). It's that easy!

If you're running the module from a 5V supply, you do have to make sure that you connect the module's DATA line to a Micromite pin that is 5V tolerant – ie, one of pins 14 to 18, 21 or 22 on the 28-pin Micromite.

So if you have connected the module's DATA line to pin 18 of the Micromite and have declared the temperature and RH variables as say 'temp!' and 'RH!' respectively, you'll be able to read the sensor's data with this one-line command:

```
HUMID 18, temp!, RH!
```

If you want to take a sequence of say 10 readings spaced apart by the recommended minimum of two seconds and print them to the console, here's the kind of simple program you'll need:

```
DIM nbr% = 10
DIM temp! = 0.0
DIM RH! = 0.0
PAUSE 1000
DO
  HUMID 18, temp!, RH!
  PRINT "Temperature = "temp!
    "C & humidity = " RH! "%"
  nbr% = nbr% - 1
  PAUSE 2000
LOOP UNTIL nbr% = 0
```

If you want to hook up an AM2302-based module to any of the Arduino versions, it's almost as easy. You have quite a choice when it comes to pre-written applications, some of which you'll find using these **github.com** links:

http://bit.ly/2DJHJ2O
http://bit.ly/2BBrAKU
http://bit.ly/2BBgju3

There are also sample programs on both of these websites:

www.aosong.com
www.humidity.com

So it's not at all difficult to use one of these low cost AM2302/DHT22/RHT03-based modules with a readily available microcontroller.

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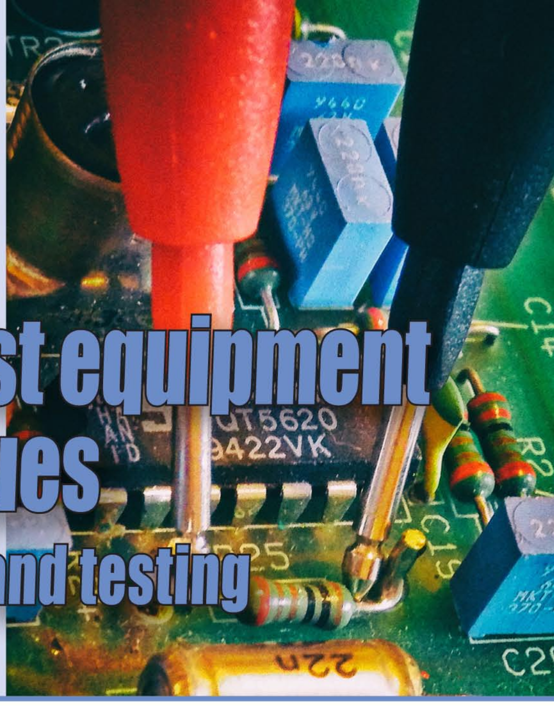
*** These pictures have been taken from the video at:**

http://youtu.be/C7uS1OJccKI

by www.youtube.com/user/electronupdate



Teach-In 2018

Get testing! – electronic test equipment and measurement techniques

Part 7: Radio frequency measurement and testing

by Mike Tooley

Welcome to *Teach-In 2018: Get testing! – electronic test equipment and measurement techniques*. This *Teach-In* series will provide you with a broad-based introduction to choosing and using a wide range of test gear, how to get the best out of each item and the pitfalls to avoid. We'll provide hints and tips on using, and – just as importantly – interpreting the results that you get. We will be dealing with familiar test gear as well as equipment designed for more specialised applications.

Our previous *Teach-In* series have dealt with specific aspects of electronics, such as PICs (*Teach-In 5*), Analogue Circuit Design (*Teach-In 6*) or popular low-cost microcontrollers (*Teach-In 7* and *8*). The current series is rather different because it has been designed to have the broadest possible appeal and is applicable to all branches of electronics. It crosses the boundaries of analogue and digital electronics with applications that span the full range of electronics – from a single-stage transistor amplifier to the most sophisticated microcontroller system. There really is something for everyone in this series!

Each part includes a simple but useful practical *Test gear project* that will build into a handy gadget that will either extend the features, ranges and usability of an existing item of test equipment or that will serve as a stand-alone instrument. We've kept the cost of these projects as low as possible and most of them can be built for less than £10 (including components, enclosure and circuit board).

This month

In this seventh part, *In theory* will be looking at the principles and techniques that underpin a wide range of RF measurements, including voltage, power, modulation depth, and standing wave ratio (SWR). *Gearing up* introduces a variety of common items of RF test equipment while *Get it right!* helps you avoid some of the pitfalls and provides useful hints and tips that will help you to improve the accuracy and relevance of your measurements. Finally, our seventh *Test Gear Project* is a wide-band RF 'sniffer' that can also act as a useful relative field strength indicator.

In theory: RF measurement principles and techniques

In last month's *Teach-In 2018* we discussed audio frequency (AF) measurements. This month, we turn our attention to measurements that are made at much higher frequencies, extending from 30kHz to 30GHz (and beyond). They include the range of frequencies used for radio and TV broadcasting, as well as the 'wireless' and Wi-Fi networks that we use in our homes.

Measuring RF voltage

Provided that the frequency of a signal is within the measurement range of an instrument, the easiest method of measuring RF signals is with the aid of an oscilloscope. In conjunction with a ×10 probe, most oscilloscopes are capable of making measurements at frequencies of up to 30MHz or more depending on the upper frequency limit of the instrument used. An example is shown in Fig.7.1 where the sinusoidal output voltage of a low-power 7MHz transmitter is displayed.

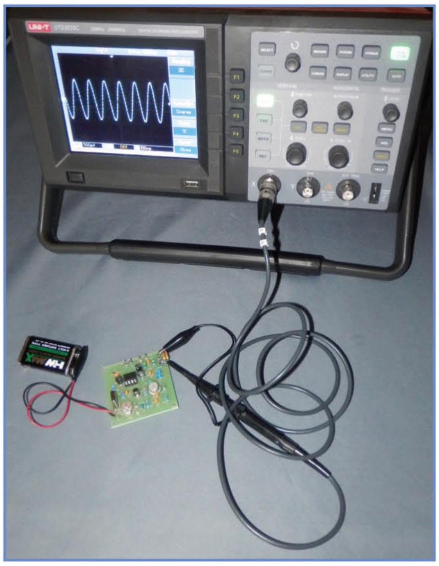


Fig.7.1. Using an oscilloscope and ×10 probe to measure the RF output of a low-power 7MHz transmitter

An alternative to using an oscilloscope for RF voltage measurements is a diode with a conventional analogue or digital meter, as shown in Fig.7.2(a). The probe will respond to the peak RF voltage but the meter can be calibrated in RMS volts. RF probes are often supplied with wideband voltmeters and they can extend the frequency range well beyond that which could be measured by an oscilloscope. To ensure the widest possible frequency range, the probe tip and ground connection need to be kept very short in order to minimise the stray reactance that would otherwise degrade the probe's performance.

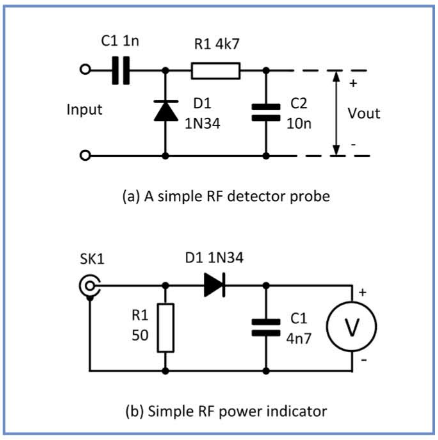


Fig.7.2. (a) A simple RF diode detector probe (b) A simple RF power indicator

A similar arrangement can be used to indicate RF power. In this case, the diode detector needs to be used in conjunction with a resistive load, as shown in Fig.7.2(b) and the meter needs to be scaled in mW or W. Since there is a square law relationship between power and RMS voltage, the instrument scale can be distinctly non-linear. Furthermore, the load resistor shown in Fig.7.2(b) must be appropriately rated in terms of power dissipation and it must be purely resistive (due to self-inductance, high-power wire-wound resistors must not be used). At frequencies above 30MHz, load resistors need to be specially constructed to minimize stray reactance and they should be fully screened in a metal enclosure and fitted with an appropriate coaxial input connector.

Modulation

Different types of modulation (see Fig.7.3) can be applied to RF signals acting as 'carriers' to convey signals such as speech, music and data. The depth of amplitude modulation can be easily measured using an oscilloscope and calculated from the relationship:

$$m = \frac{V_{max} - V_{min}}{V_{max} + V_{min}} \times 100\%$$

where V_{max} and V_{min} are the maximum/minimum envelope voltages, see Fig.7.4.

RF power measurement

As mentioned previously, at low RF frequencies it is possible to use an oscilloscope to measure RF voltage and

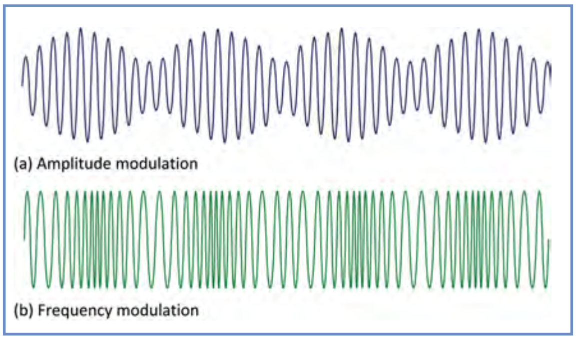


Fig.7.3. (a) Amplitude modulation (b) Frequency modulation

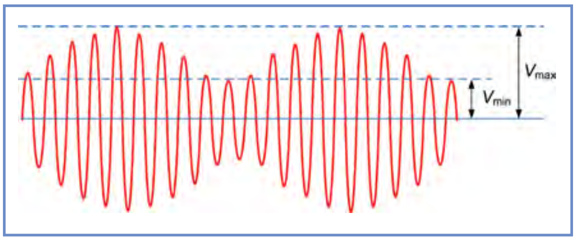


Fig.7.4. Measuring the depth of amplitude modulation

Fig.7.6. Common 75Ω and 50Ω RF connectors – (L-R) Belling Lee, PL-259, SMA, BNC, N-type

Table 7.1 Typical values of intercept (dBm) for the AD8318 logarithmic power detector

Frequency	900MHz	1.9GHz	2.2GHz	3.6GHz	5.8GHz	8.0GHz
Intercept	+22dBm	+20.4dBm	+19.6dBm	+19.8dBm	+25dBm	+37dBm

Fig.7.5. Measuring the output power from a hand-held digital transceiver (the power meter indicates an output of exactly 2W delivered to the 50Ω load attached to the power meter

thus, provided that the load resistance into which the power is delivered is accurately known, it is possible to determine power from the peak-to-peak value of a waveform observed on an oscilloscope. The following relationship is used:

$$P_{out} = \frac{V_{out(pk\text{-}pk)}^{\;2}}{8R_L}$$

Putting this into context, let's assume that the oscilloscope indicates a peak-peak voltage of 20V with a load having a resistance of 50Ω. The output power would be calculated as follows:

$$P_{out} = \frac{V_{out\,(pk\text{-}pk)}^{\;2}}{8R_L} = \frac{20^2}{8 \times 50} = \frac{400}{400} = 1\text{W}$$

An alternative to using an oscilloscope is that of using a dedicated RF power meter, in which case the instrument scale will already be calibrated in mW or W – see Fig.7.5.

Instrument connection

When carrying out RF measurements it is usually important to ensure that input and output impedances (as well as all connecting cables) are correctly matched. When applied to cables, the term 'characteristic impedance' is the impedance that would be seen looking into an infinite length of the cable at the working frequency. Typical values are 50Ω, 75Ω and 90Ω for unbalanced (coaxial) cables, as well as 300Ω and 600Ω for balanced (twin) feeders.

Characteristic impedance is a function of the primary constants of the cable, the two most significant being the inductance (L) and capacitance (C) per unit length of the cable (note that L is measured with the far end of the cable shorted while C is measured with the far end open circuit). The value of characteristic impedance (Z_0) is approximately given by:

$$Z_0 = \sqrt{\frac{L}{C}}$$

Connectors

Different types of connector (see Fig.7.6) are used with coaxial cables, depending on the frequency range, power handling requirements and characteristic impedance required. The popular BNC connector is available in both 50Ω and 75Ω versions (note that the diameter of the centre pin is smaller for the 75Ω version). Conventional TV connectors (originated by Belling-Lee) are used with 75Ω systems. N-type and SMA connectors have a constant 50Ω impedance over a very wide frequency range.

Measuring low RF power

When low levels of power (less than 100mW) are to be measured, simple diode detectors are generally unsuitable due to their non-linearity and poor sensitivity. Sensitive power meters with separate power sensing elements are available but, in recent years, an alternative low-cost solution can be based around the use of a specialised logarithmic power detector chip, such as the AD8318 from Analog Devices (see Fig.7.7).

Designed for relative received signal strength indication (RSSI) and automatic power regulation, the AD8318 is a demodulating logarithmic amplifier that accurately converts an RF input to a corresponding dB-scaled output voltage. The device uses progressive compression over a nine-stage cascaded amplifier, and each amplifier stage has its own detector (see Fig.7.8). Detector outputs are then summed to provide an output voltage that varies between –2.1V with no input and –1.4V (with 10mW input into 50Ω). The useful measurement range (without external attenuation) is from +5dB to –55dBm (see Fig.7.9) with an error of less than ±1dB.

The logarithmic slope of the AD8318's transfer characteristic (see Fig.7.9) is nominally –25mV/dB but can be adjusted by scaling the feedback voltage from VOUT to the VSET input (see Fig.7.8).

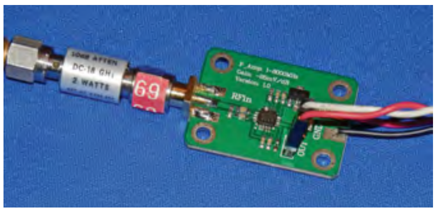


Fig.7.7. A logarithmic detector (the input on the left is fitted with a fixed 10dB attenuator rated at 2W from DC to 18GHz)

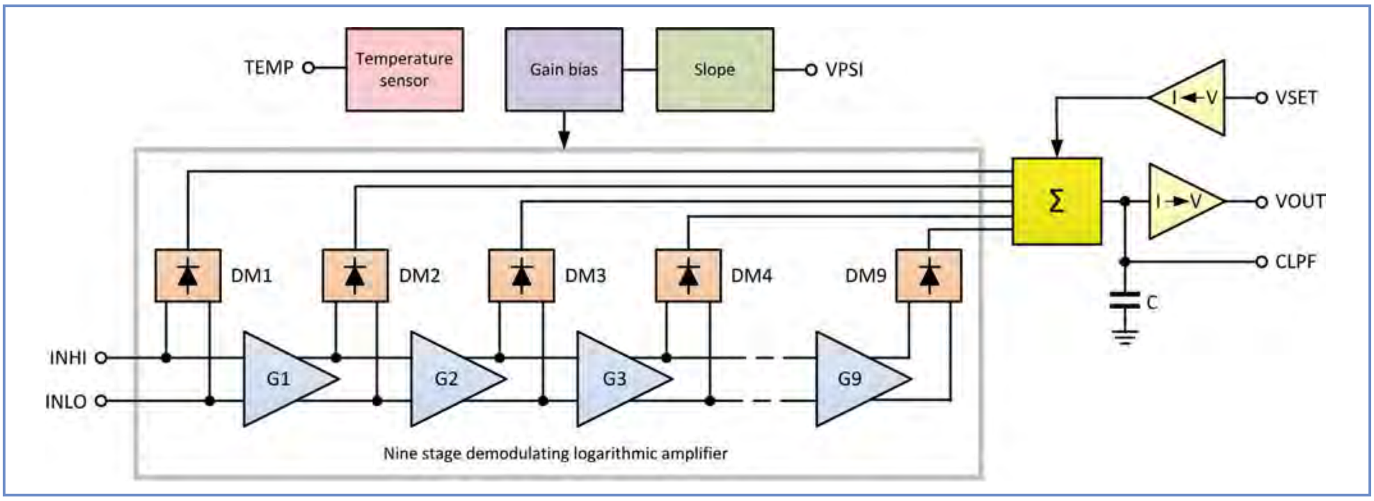


Fig.7.8. Simplified internal schematic of the AD8318 logarithmic detector

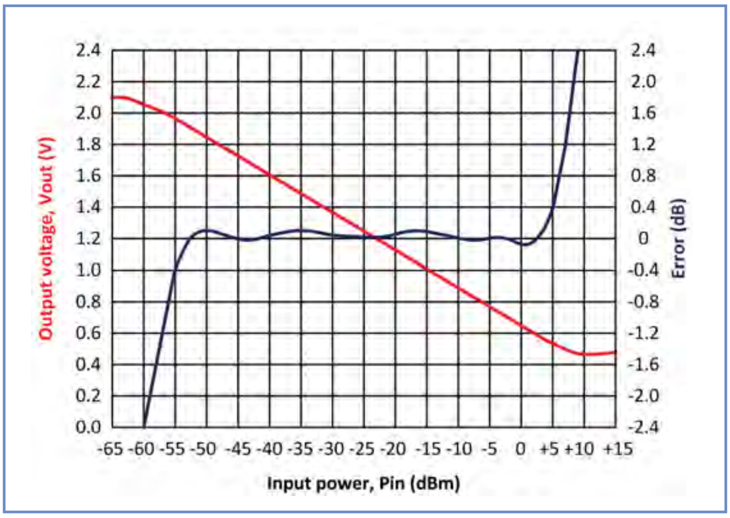


Fig.7.9. Output voltage (V) vs input power (dBm) at 2.2GHz

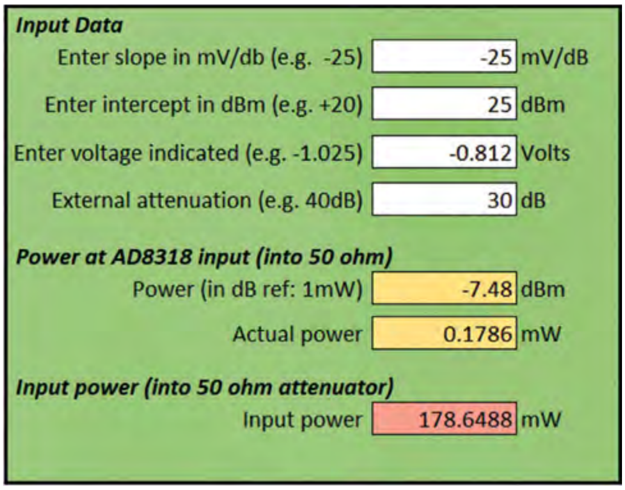


Fig.7.10. Using a spreadsheet to calculate the RF input power (see text)

The intercept (not shown in Fig.7.9) is nominally +20dBm (referenced to 50Ω but it varies somewhat with frequency, effectively shifting the transfer characteristic along the *x*-axis). Measured values of intercept at different frequencies are shown in Table 7.1 but to make life easier, we have made a spreadsheet calculator (see Fig.7.10) available for download from the *EPE* website. This allows you to calculate power levels at different frequencies with various amounts of external attenuation present.

Fig.7.11. Checking the output frequency of a dual-band FM transceiver. The digital frequency meter (shown on the right) indicates a frequency difference of 1kHz

Frequency measurement

Frequency is difficult to accurately measure using a conventional oscilloscope. Modern digital storage oscilloscopes (DSO) fare somewhat better in this respect as they can often display frequency values digitally. However, a digital frequency meter (DFM – see Fig.7.11) is a better solution and the desirable characteristics of such an instrument are high sensitivity, a high upper frequency limit as well as appropriate accuracy and resolution. A sensitivity of 100mV or less is suitable for most applications, and an upper frequency limit of 500MHz will be adequate for most measurements at HF and VHF.

Generating RF signals

An RF signal generator will be required for measuring the performance of radio receivers, amplifiers and filters. So that its output can be reduced to a suitably low level, such an instrument will need to be fitted with an accurate calibrated attenuator and the output level should be adjustable from a minimum of 1V RMS to less than 1μV. Precise frequency calibration is not essential if a DFM is available to monitor the output frequency (see Fig.7.12). The generator should have an output impedance of 50Ω and the output should be capable of being modulated in order to provide an AM or FM signal. Some instruments may also be supplied with internal calibration facilities, but this is no longer essential in the case of equipment that uses digital frequency synthesis where signals are derived from an accurate internal reference oscillator. Finally, to prevent leakage, adequate screening is required.

Fig.7.12. Bench test instruments: an RF signal generator used with a digital frequency meter (DFM)

Fig.7.13 shows the complete circuit of

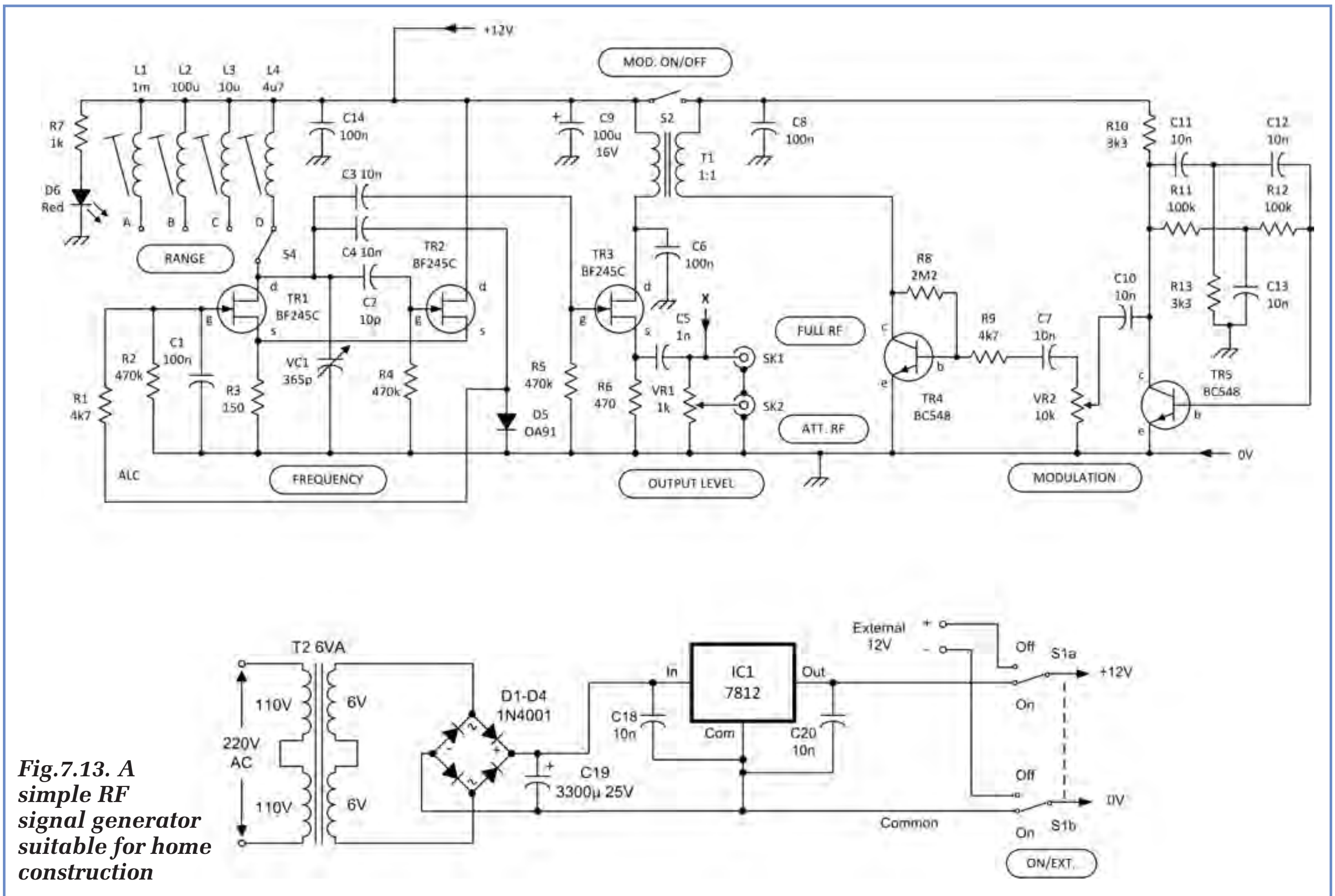


Fig.7.13. A simple RF signal generator suitable for home construction

a simple RF signal generator suitable for home construction. This instrument operates from a mains supply or 12V battery. It provides an amplitude modulated output from 250kHz to around 15MHz in four switched ranges.

Spectrum analysis

It is often useful to be able to display the frequency spectrum of an RF signal. This will allow you to identify unwanted harmonic components as well as those resulting from noise, spurious oscillation and unwanted mixing of signals. A dedicated RF spectrum analyser can be prohibitively expensive, but a low-cost solution can be based on a simple DVB-T tuner 'dongle', as shown in Fig.7.14. Such devices are widely available, and they are usually marketed as USB 'digital TV sticks'. Used in conjunction with suitable software running on a desktop PC, laptop or tablet, they will allow you to receive signals over a frequency range extending from around 25MHz to 1.5GHz. If desired, the frequency range can be extended down to around 150kHz by modifying the device, bypassing the front-end tuner/frequency changer and feeding the input signals direct to the I and Q inputs on the SDR decoder, as shown in Fig.7.15. Alternatively, a ready-made RF up-converter (such as the NooElec Ham it Up) can be fitted. A typical spectrum display is shown in Fig.7.16. This shows a 2MHz frequency range that extends from 104MHz to 106MHz in which signal levels range from –20dBm to –40dBm. The strong signal at 104.6MHz is a local FM broadcast station.

Fig.7.14. A low-cost DVB-T tuner 'dongle' can make the basis of a simple RF spectrum analyser

Noise sources

A wideband noise source can be used to carry out various RF measurements and is a low-cost alternative to using more sophisticated equipment such as tracking spectrum analysers. The inexpensive noise source shown in Fig.7.17 can be useful for testing amplifiers and filters.

The output of a noise source is normally specified in terms of Excess Noise Ratio (ENR). This is a normalised measure of how much noise is generated when the noise source is switched on compared with the noise that would be thermally generated with the noise source switched off.

$$\mathrm{ENR} = 10\log_{10}\left(\frac{T_h - 290}{290}\right) = 10\log_{10}\left(\frac{T_h}{T_c} - 1\right)$$

where $T_h > T_c$ (290K), T_h is the temperature 'hot', and T_c is the temperature 'cold'. (Note temperatures are measured in kelvin, not Celcius.)

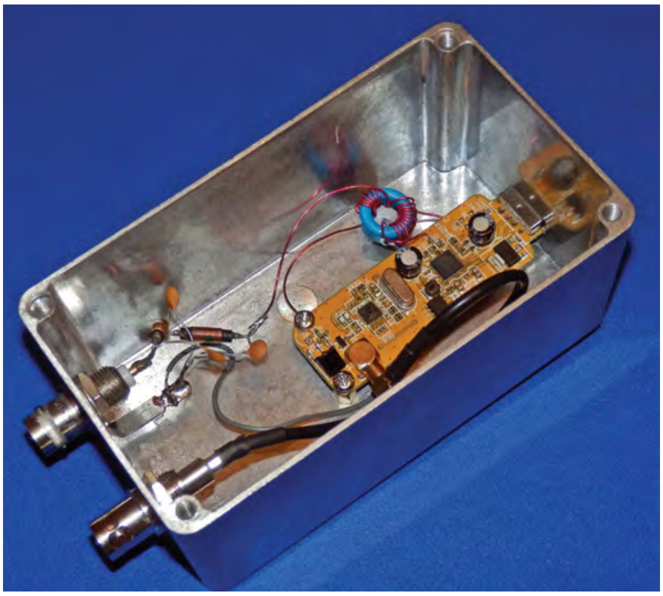


Fig.7.15. A DVB-T tuner 'dongle' modified to support MF/HF as well as VHF/UHF inputs. This permits operation over a frequency range extending from around 100kHz to well over 1GHz

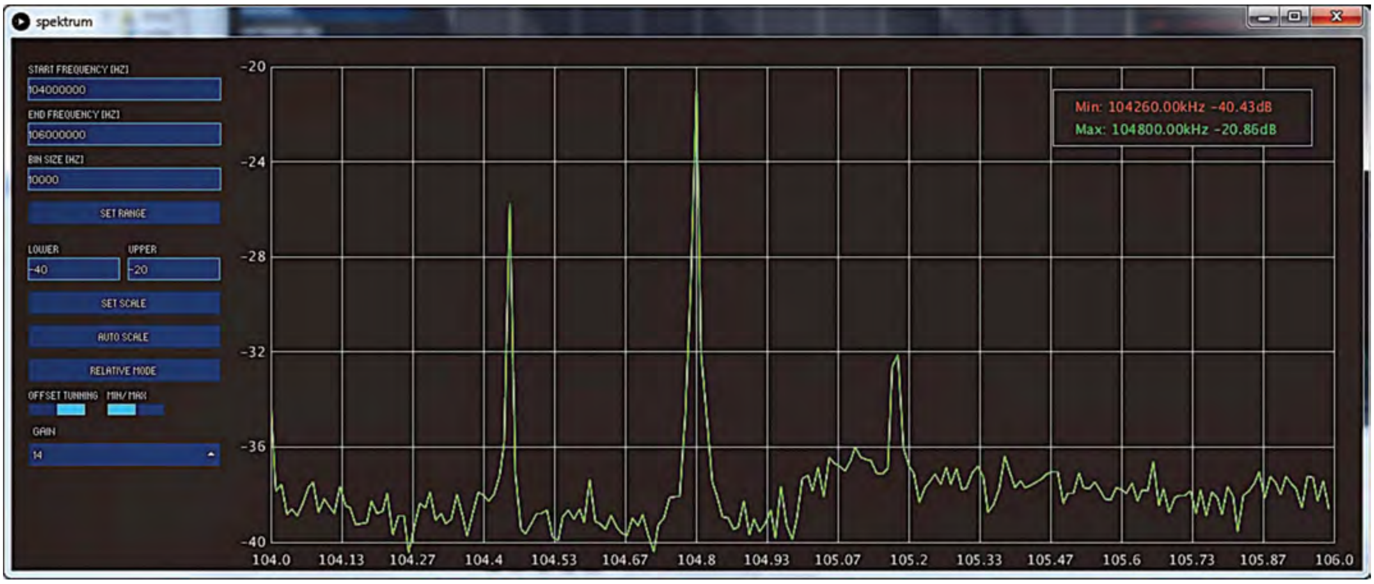


Fig.7.16. A typical spectrum display obtained from a DVB-T tuner 'dongle' and appropriate SDR software. The strong signal at 104.6MHz is a local FM broadcast station

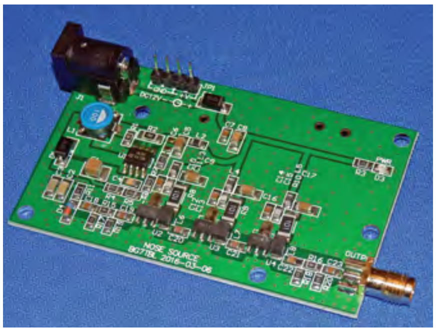


Fig.7.17. A low-cost wideband RF noise source can be useful for testing amplifiers and filters up to about 1GHz

When a spectrum analyser is used for testing a filter in conjunction with a noise source, the ENR of the noise source must be significantly greater than the spectrum analyser's noise floor. An ideal value would be 60dB but, for testing an RF amplifier the ENR should be very much smaller and 6dB to 10dB would usually be more appropriate. Thus, an RF noise source should normally be used in conjunction with an attenuator. Without an external attenuator, the measured broadband power output of the noise source shown in Fig.7.17 is around 10dBm (10mW) and so an external attenuator will invariably be required when using it with a spectrum analyser.

Measuring standing wave ratio (SWR)

Before describing instruments that can be used to measure SWR it is important to understand what we are trying to measure. In radio equipment an antenna is connected to a source of RF (eg, a transmitter or transceiver) by means of a feeder. This feeder (usually a length of coaxial cable) forms a transmission line, along which the RF energy is conveyed. The line is said to be correctly 'matched' when the impedance of the source, line, and load are all *identical*. The source is the sender or transmitter, while the load is the antenna.

If the system is perfectly matched and the line is loss-free, the voltage will be the same at all points along the line. If the system is perfectly matched and the line is 'lossy', there will be a linear reduction in voltage along the line. If, however, the system is not correctly matched, which would be the case if the antenna had a different impedance from the source and line, part of the energy will be reflected back from the load to the source. This results in standing waves of voltage and current along the line. The presence of standing waves is undesirable for various reasons, including additional power loss along the feeder. This can be very important when a long run of poor quality feeder is used. In summary:

- The SWR of a feeder or transmission line is an indication of the effectiveness of the impedance match between the transmission line and the antenna
- SWR is the ratio of the maximum to the minimum current or voltage along the length of the transmission line, or the ratio of the maximum to the minimum voltage
- When the line is correctly matched the SWR is unity. In other words, we have unity SWR when there is no variation in voltage or current along the transmission line
- The greater the number representing SWR, the larger is the mismatch and the greater is the loss due to an imperfect feeder.

SWR is easily measured using an instrument known variously as an SWR bridge, SWR meter, or combined power/SWR meter (see Fig.7.18). Despite the different appearance of these instruments they are all based on the same operating principle; sensing the forward and reflected power and displaying the difference between them using a meter scale or digital display. Fig.7.19 shows the internal construction of a low-cost SWR bridge for operation at up to 30MHz. This shows the main transmission line in the centre and the two coupled transmission lines either side of it feed

Fig.7.18. Typical low-cost combined RF power meters and SWR bridges

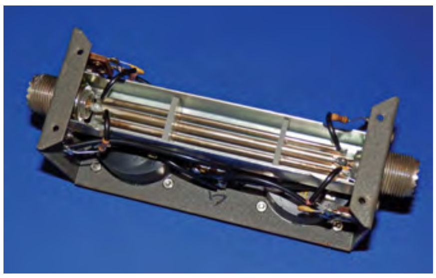


Fig.7.19. The interior of an HF SWR bridge showing the main transmission line and the two coupled transmission lines that feed the forward and reverse detectors

Fig.7.20. Checking output power and antenna SWR with a dual-band VHF/UHF base station transceiver

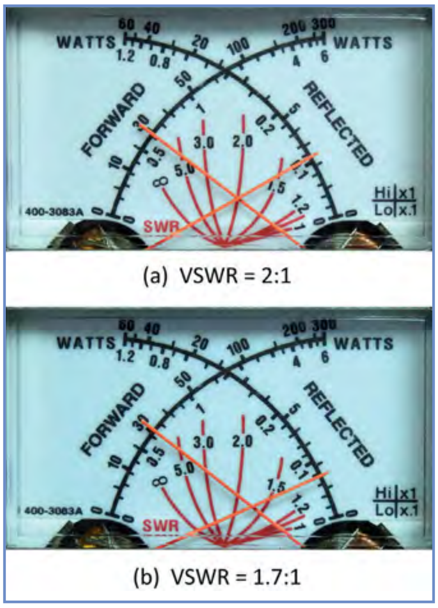


Fig.7.21. Reading a cross-needle SWR indicator

Fig.7.22. HF/VHF antenna analyser for operation between 1.8MHz and 170MHz. The instrument can display resistance, reactance, impedance, reflection coefficient and SWR

the forward and reverse diode detectors. Fig.7.20 shows a combined HF and VHF power/SWR meter being used to check the output from a dual-band VHF/UHF base station transceiver. The instrument has two separate sensing heads (one for HF and one for VHF) and measures power from 5W to 400W in four ranges at frequencies from 1.8MHz to 525MHz.

Instead of using a single meter with forward/reverse switching (or two separate meter movements) to indicate forward and reverse power, some SWR bridges make use of cross-needle indicators. These comprise two separate moving coil movements each with its own needle pointer. One needle indicates forward power while the other (simultaneously) indicates reflected power. The corresponding SWR is read along a line marked in the centre of the display, as shown in Fig.7.21.

Antenna and virtual network analysers (VNA)

If you are making regular measurements at RF (and particularly if you are working with antennas) you will find an antenna or virtual network analyser invaluable. A basic antenna analyser is shown in Fig.7.22. This popular instrument is capable of measuring SWR, impedance (resistance and reactance), and reflection coefficient over a frequency range extending from 1.8MHz to 170MHz. For automated measurements in conjunction with a desktop PC, laptop or tablet, virtual network analysers are available from several RF test equipment manufacturers.

Gearing up: RF test equipment

In addition to an oscilloscope for general RF measurements, you will need an RF signal generator and a digital frequency meter. Both are available at reasonable cost, new and second hand. For a basic RF signal generator, you can expect to pay around £100, but instruments are regularly available from on-line auction sites at bargain prices. Typical examples are shown in Fig.7.23 and 7.24. When choosing an RF signal generator, it is important to select an instrument with sufficiently wide frequency coverage and also with a calibrated attenuator. For basic radio and TV servicing such instruments are often designed to work with 75Ω systems, but for most professional applications a 50Ω output is more common. For AM measurements, depth of modulation can be measured using an oscilloscope, but dedicated modulation meters capable of measuring both AM and FM signals are available (see Fig.7.25).

SWR bridges are available at various prices and quality levels but it is advisable to avoid the cheaper instruments that invariably have cramped displays, limited frequency response and poor accuracy. A good quality cross-needle instrument from Avair is shown in Fig.7.26. This instrument covers the range 1.8MHz to 200MHz in two ranges; 30W and 300W full-scale and is available at around £50. Directional power meters from Bird (the 'Thruline' series of wattmeters) are highly recommended, but you may require a selection of different plug-in sensing elements in order to cover the required range of frequency and power. Following the purchase of a basic instrument these can represent a significant additional outlay.

Fig.7.23. AM/FM RF signal generator

Fig.7.24. A highly accurate RF signal generator that incorporates phase locked digital frequency entry

Fig.7.25. FM/AM modulation meter

Fig.7.26. A cross-needle SWR meter (like this instrument from Avair) can be a useful investment if you need to measure RF power or adjust antennas at HF and VHF

Other RF instruments and accessories

In addition to the instruments listed above, several other useful items of test gear can be acquired from second-hand and on-line sources. They include:

- RF voltmeters and AC meters with diode probes (see Fig.7.27)
- RF power meters with internal loads (see Fig.7.28)
- RF spectrum analysers (see Fig.7.29)
- Calibrated RF attenuators (see Fig.7.30)
- RF connectors, inter-series adapters, patch leads and probes (see Fig.7.31).

Fig.7.27. Part of the author's RF test bench showing a variety of RF voltage and power meters purchased from second-hand dealers and from on-line auction sites

Fig.7.28. A Marconi TF1152 power meter suitable for continuous power of up to 25W with an internal load rated for use up to 500MHz. Instruments like this can make an excellent second-hand purchase

Fig.7.29. A Farnell 352C spectrum analyser suitable for use from 300kHz to 1GHz

Fig.7.30. A Marconi switched attenuator suitable for use from DC to 1GHz with attenuation of up to 142dB in steps of 1dB

Fig.7.31. A variety of connectors, adapters and other accessories can be extremely useful when carrying out RF measurements

Get it right when carrying out RF measurements

- Always ensure that test leads and cables carrying RF signals are properly screened and fitted with appropriate connectors
- When using an RF probe always ensure that the ground connection is as short and direct as possible
- Avoid over-driving amplifiers, filters and attenuators and always keep input signals within the working range for the equipment on-test
- When carrying out measurements of RF power, use a matched and fully screened resistive load that is appropriately rated in terms of impedance and frequency range
- When in doubt about signal and power levels it is wise to use an attenuator ahead of the measuring instrument and then progressively reduce the attenuation as required
- When carrying out tests and adjustment on antennas, avoid physical contact with the antenna and maintain a safe working distance from it
- Don't rely on measurements of RF voltage, power and SWR when an instrument is working towards the end of its measuring range (accuracy will invariably be impaired as an instrument's limits are approached).

Test Gear Project: A handy RF 'sniffer'

Our handy RF 'sniffer' will provide you with a useful device for detecting RF energy. It operates over a wide range of frequency extending from around 3MHz to over 500MHz (with reduced sensitivity). The presence of RF is indicated using an LED with brightness depending on the level detected, but the instrument can be used with an external voltmeter as a sensitive field strength indicator.

The complete circuit of our *Test Gear Project* is shown in Fig.7.32. The circuit comprises a diode detector followed by TR1 and TR2 acting as a high-gain DC amplifier. To improve sensitivity, the diode detector is biased to the edge of conduction by means of a constant voltage source formed by D2 and adjustable via RV1. The input of the circuit is applied by means of a BNC connector that permits the connection of a measuring probe or short antenna.

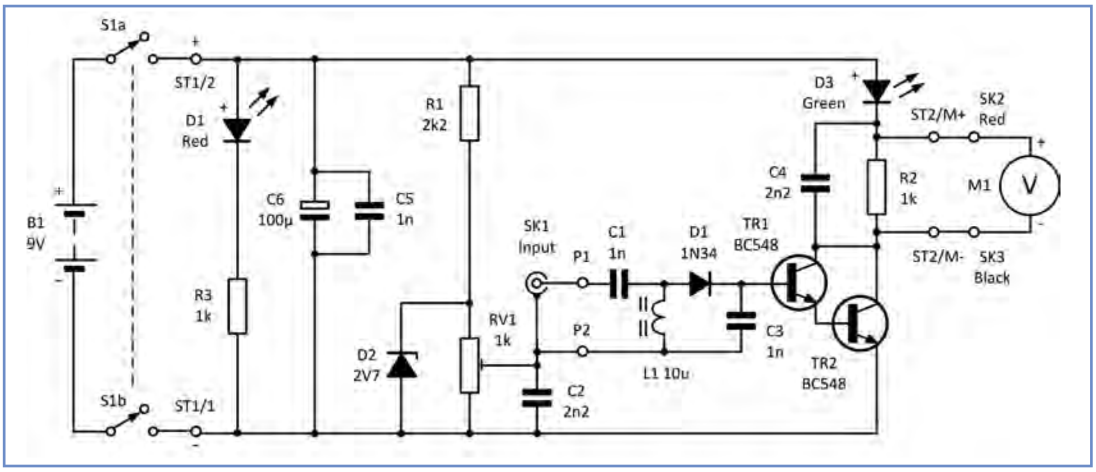


Fig.7.32. Complete circuit of the handy RF 'sniffer'

You will need

1 Perforated copper stripboard (9 strips, each with 25 holes)
2 2-way miniature terminal blocks (ST1 and ST2)
1 ABS case with integral battery compartment
1 9V PP3 battery clip
1 9V PP3 battery
1 Miniature DPDT toggle switch (S1)
1 chassis mounting female 50Ω BNC connector
1 red 2mm panel mounting socket (SK2)
1 black 2mm panel mounting socket (SK3)
2 BC548 transistors (TR1 and TR2)
1 5mm red LED (D1)
1 5mm green LED (D3)
1 2.2kΩ resistor (R1)
2 1kΩ resistor (R2 and R3)
1 miniature axial lead 10µH inductor (L1)
1 1kΩ miniature multi-turn pre-set resistor (RV1)
3 1nF disk ceramic capacitors (C1, C3 and C5)
2 2.2nF disk ceramic capacitors (C2 and C4)
1 100µF 16V radial electrolytic (C5)

Assembly is straightforward and should follow the component layout shown in Fig.7.33. Note that the '+' symbol shown on D1 indicates the more positive (anode) terminal of the LED. The pin connections for the LED and transistor are shown in Fig.7.34. The reverse side of the board (NOT an X-ray view) is also shown in Fig.7.34. Note that there's a total of 23 track breaks to be made. These can be made either with a purpose-designed spot-face cutter or using a small drill bit of appropriate size. There are also nine links that can be made with tinned copper wire of a suitable diameter or gauge (eg, 0.6mm/24SWG). When soldering has been completed it is very important to carry out a careful visual check of the board as well as an examination of the track side of the board looking for solder splashes and unwanted links between tracks. The internal and rear panel wiring of the test signal source is shown in Fig.7.35.

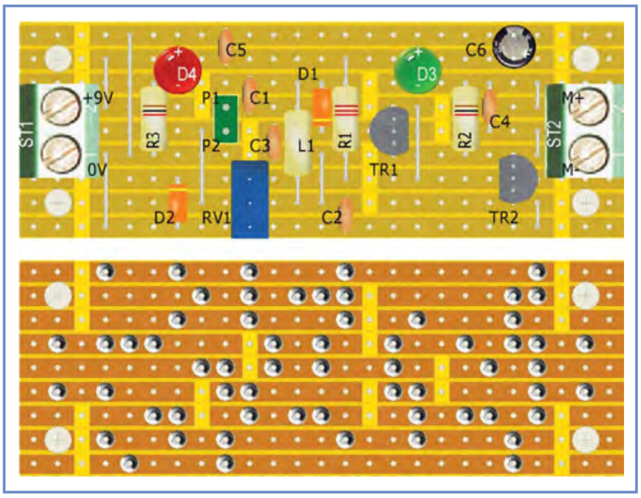


Fig.7.33. Stripboard layout of the handy RF 'sniffer'

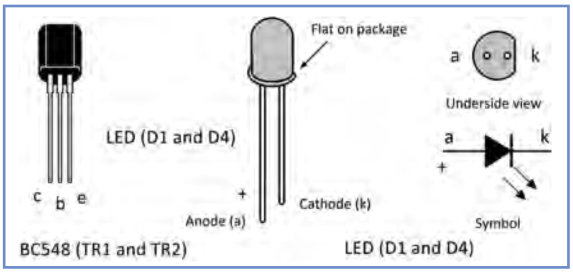


Fig.7.34. LED and transistor pin connections

Setting up

Setting up is reasonably straightforward. Switch 'on' and with no input connected to SK1, set RV1 to minimum position. Then slowly increase the bias voltage at P2 by advancing the setting of RV1 until the green LED (D3) just starts to become illuminated. Back off the setting of VR1 slightly until the LED turns 'off'. Next, connect a short antenna or pick-up loop to SK1 and, with the aid of an RF source, such as a wireless key-fob or other 430MHz remote control device, placed close to the antenna, check that the green LED becomes illuminated. If desired, connect a DC voltmeter on the 10V or 20V DC range to SK2 and SK3 and check that a reading is obtained. Note that a PMR handy talkie, or other transceiver with an output of between 1W and 5W, should typically produce an indication of between 250mV and 1.5V when positioned at distances of between 8m and 2m respectively.

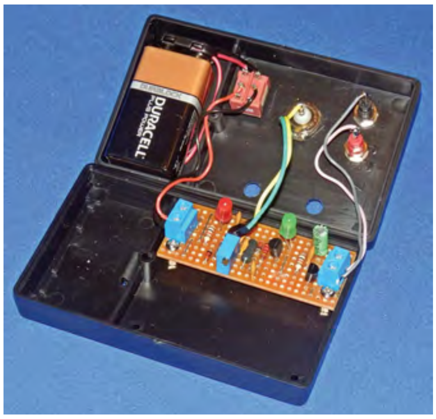


Fig.7.35. Internal wiring of the handy RF 'sniffer'

Fig.7.36. External appearance of the handy RF 'sniffer'

Fig.7.37. Using the handy RF 'sniffer' as a relative field strength indicator

Next month

In next month's *Teach-In 2018* we will be looking at digital measurements and associated test equipment. Our practical project will feature a simple logic probe.

NET WORK

by Alan Winstanley

Spectres from the past

THE NEW YEAR got off to a flying start with the breaking news that a flaw had been discovered in the microarchitecture of popular CPU chips manufactured for many years by Intel and others. While *Net Work* is not an IT feature, this month's column will help ensure readers are mindful of these potentially critical flaws. As things have turned out, similar defects affect processors found in most PCs, servers, tablets and smartphones. Risks were found in the way bytes of sensitive data, held in a processor's supposedly protected kernel memory, could in fact be accessed from outside. Something impossible had apparently happened, when proofs of concept demonstrated how data such as logins, cookie text or passwords could be extracted from a CPU's 'privileged' memory area. It appears the makings of this flaw had been mooted as far back as June 2017, referencing an Intel Xeon platform running Linux. (See Google's Project Zero Blog at: **https://goo.gl/ixkqex**). Interested stakeholders duly went public in January this year, causing pandemonium across the whole IT sector.

Hardware, not software

Secure data that had been cached could, in theory, be accessed by malware, infected web pages or scripts, due to the way that a CPU core tries to second-guess or 'speculate' what commands it might need to execute in the near future. This technique can leave remnants of sensitive data hanging 'in limbo' in the kernel where, as laboratory tests demonstrated, it could be intercepted (very slowly) by eavesdropping malware. These 'speculative side-channel execution' flaws are hard-wired into processor silicon and date back many years.

The processing bugs have been dubbed 'Meltdown' and 'Spectre'. Meltdown affects a wide range of legacy Intel processors and also a number made by ARM (as used in smartphones and tablets). Spectre tries to force programs to speculate the same way and reveal privileged data as a result; Spectre can reportedly affect AMD processors too. These are highly esoteric bugs that are said to be difficult to exploit, but the industry remains tight-lipped about whether the problem has already appeared in the wild. While it seems exceptionally unlikely, it is of course impossible to prove a negative, and the fall-out and legal liabilities created by these bugs will haunt CPU manufacturers for years to come.

Ghostbusters

As if personal computing devices aren't already vulnerable, cloud-based services such as Amazon Web Services were hit by the fallout too. The majority of web servers are Intel-based and by the time you read this, many if not most of them will have been updated, causing consternation in the IT industry because of a sometimes measurable performance hit to the system's operating speed. A very digestible resource, including demos of Meltdown and Spectre is hosted by Graz University of Technology at: **https://meltdownattack.com**

Where does this leave the everyday user? It's the typical game of cat and mouse that seasoned Internet users know too well, but there is a practical limit to what individuals can do to protect themselves. Microsoft states it has had no evidence that its customers have been attacked by Meltdown or Spectre. Early Windows patches issues by Microsoft 'bricked' some AMD-based computers, and due to rebooting problems Intel has since warned users to only install patches pushed onto them by their OS or motherboard vendor, and not manually try installing unproven CPU firmware patches at this time. It may never be possible to update a lot of legacy computer or mobile hardware, and users will rely on software-level protection to guard against a hardware flaw that may never come to anything anyway.

Keep up to date

It's beyond the scope of *Net Work* to delve deeply into individual computer systems, but we know that most readers use Windows-based or Linux machines and a number of Intel-based Macs are also caught in the net. The only thing users can do is to monitor updates and ensure that their personal computer's operating systems and software are up to date. In the case of Microsoft Windows, a Windows update will probably already have been installed (notably KB405689x), see information from Microsoft at: **https://goo.gl/cptrwQ**. Mac OS and Chromebook devices may already have been silently updated. Web browsers including Chrome and Firefox are also being updated to sidestep Meltdown and Spectre 'timing attacks' possibly originating on websites. Consider also using an ad blocker to prevent possibly hostile third-party adverts displayed on websites from attacking your computer. Still to come is the likely impact on Android smartphone or tablet owners and countless other smart devices containing embedded processors and software that may never be updatable.

Don't panic!

Endeavouring to take a balanced view, the chances of being on the receiving end of Meltdown or Spectre are probably tiny in the first place, helped by fact that an embargo was placed on information about the flaw which gave manufacturers a breathing space to patch it. Having updated web browsers, the OS and everything else in sight, the author will simply sit tight, the same as always, and take any more updates as they come. There has been no discernable difference in everyday PC system speed after the Windows updates were automatically installed.

Perhaps this is another nail in the coffin of home computing done the traditional way, and we will all eventually give up and opt for dumb 'client' computers (whether using faulty processors or not) networked to the cloud, where all the heavy lifting and anti-malware measures are taken care of for us. Microsoft, after all, touts Windows 10 simply as a constantly rolling 'service' rather than an operating system; and trying to keep a personal computer and its software up to date is proving increasingly onerous, unrewarding and time-consuming for its users. So hang on, there is more to come and it's going to be a bumpy ride!

Control system for car air suspension

by Julian Edgar

The world's first programmable, DIY air suspension controller for cars uses two PIC modules

Many luxury cars use air springs rather than steel springs. Air springs have multiple technical advantages over conventional steel springs, but the biggest is that the ride height of the car can be maintained at a constant level, irrespective of changing loads. Seeing these advantages, I decided to fit custom air suspension to one of my cars – a 2000 model Honda Insight.

Air springs are a lot more sophisticated than just squashed rubber balloons – but for the purposes of this article, it might be easiest to think about them in that way. Add more air to the spring, and ride height rises. Allow air to flow out, and ride height falls. The movement of air is controlled by electric solenoid valves, with the high-pressure air provided by an on-board 12V electric compressor and a storage tank. Ride height is monitored by potentiometer-based sensors. And the controller making these decisions? I decided to use a pair of PIC-based modules from Australian company eLabtronics, with the modules running custom-developed software.

The Honda Insight equipped with air suspension, set to its lowest ride height.

Controller demands

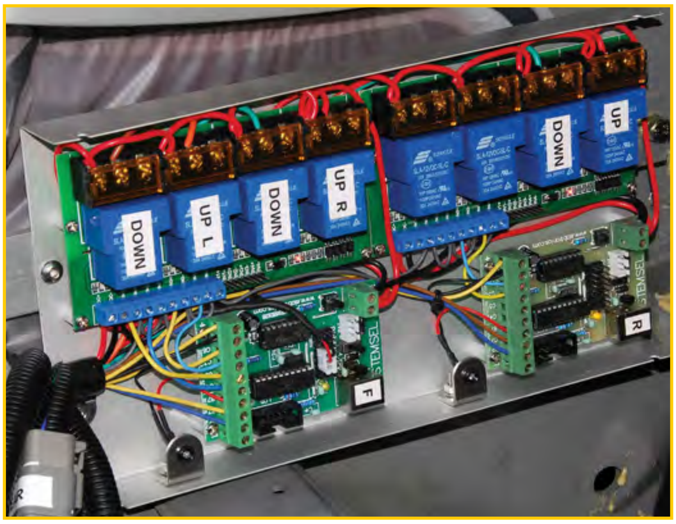
So what function does an air suspension electronic control system need to perform? At its simplest, it needs to monitor ride height and then trigger solenoid valves appropriately to maintain this height at a constant level. Hmm, sound easy? Well, let's add some complexities.

First, how do we measure ride height? As described above, typically, potentiometer-based sensors are used, mounted in parallel with the springs. As load increases, the springs are compressed and so the ride height decreases. But the springs also compress and extend with bumps – that's their purpose! So how can we differentiate the measured ride height variations being caused by bumps with the variation being caused by 'true' ride height changes? An averaging function is needed so that the fast movements of bumps is ignored, but the slower movements of ride height changes are measured.

But then it gets more complex. How long should this averaging period be? Movement of the springs with bumps are fast, so if we were to average over – say – 10 minutes, we'd get a good indication of the actual ride height. But what happens then if the car is stopped – and five people get into it? The suspension will compress, and we can't wait 10 minutes before correcting the ride height – the suspension will be on its bump stops!

Therefore, we really need two suspension height measuring modes – one that reacts quickly (people getting in and out) and one that reacts slowly (change of ride height caused by use of fuel, variation in load of mud or snow, or the temperature of the air within the springs changing). The 'fast reaction' averaging could be over (say) 10 seconds, and the 'slow reaction' averaging over (say) 10 minutes.

The Insight at maximum ride height. Neither of these settings is used on the road.



The control system uses two eLabtronics PIC-based STEMSEL controllers working with two quad relay boards. The relays control solenoids that allow air to flow in/out of the air springs.

A ride height sensor, shown before the wiring was completed. The car uses potentiometer-based sensors usually fitted to Range Rovers.

So why not just make the averaging time 10 seconds in all conditions? Imagine a long corner, where the car rolls and the outer suspension compresses. In that situation, a 10-second averaging technique will cause the system to try to inflate these outer springs! Thus, two different averaging modes really are needed.

But how will the system know which mode to select? In the case of the eLabtronics controller, we decided to select the correct averaging mode by detecting how often the ride height sensor was changing in output. This is because when the car is being driven on the road, the suspension height outputs change frequently per second, while when the car is stopped, the ride height outputs change much less frequently.

Hmm, now what about a driver control? A potentiometer input is also fitted for the driver, allowing the manual selection of ride height over a pre-fixed range. Finally, in the case of the Honda, I decided to control the rear air springs as a pair and the front springs separately.

So let's see how this is all coming together; the controller requirements are:

Analogue 0-5V inputs

- Rear suspension height
- Front-left suspension height
- Front-right suspension height
- Driver height control pot

Digital outputs

- Rear suspension up / down solenoids
- Front-left spring up / down solenoids
- Front-right spring up / down solenoids
- Mode LED (this shows whether the slow or fast averaging mode has been selected by the software)

Read all about it!

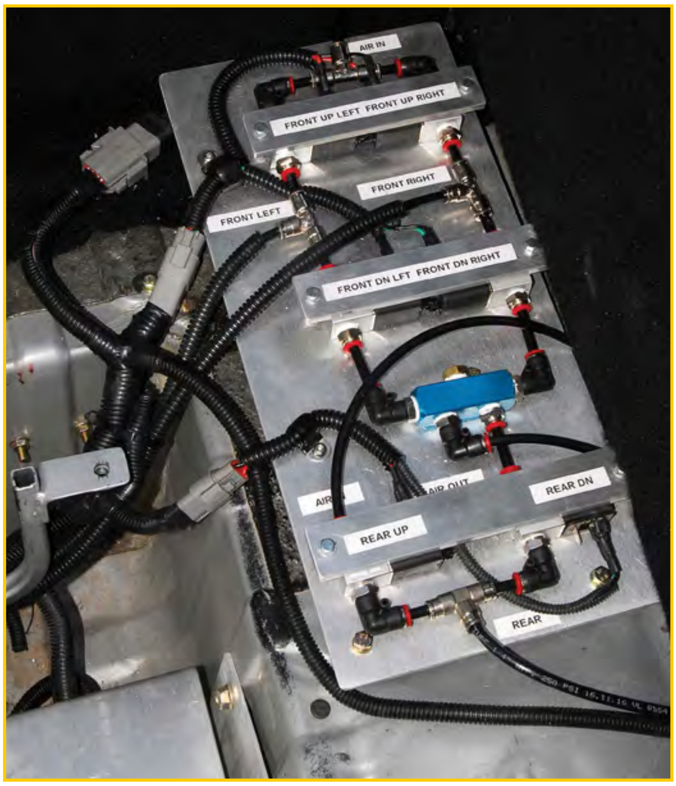
The development of the air suspension system in the Honda – including its electronics – is covered in detail in the book 'Custom Air Suspension', published worldwide by Veloce.

The controller

The selected eLabtronics product was the company's STEMSEL module. This is a prebuilt (but unboxed) module that uses a PIC18F14K50 microcontroller. The module has 12 digital or analogue input/output pins, four driver outputs that can drive relays (and so operate the solenoids), an on-board USB connector that allows reprogramming from a PC or laptop, and it can be powered from the car's 12V supply. In addition, it has a regulated 5V output that was used to power the suspension height sensors and the dashboard height selection pot.

The presence of only four driver outputs means that two STEMSEL modules were required – one for the front suspension and one for the rear. These work with two commercially available quad relay boards, with six of the eight relays used.

Now what about programming the module? While I devised all the logic of the control system, writing a program to achieve this is beyond my capabilities. However, eLabtronics had University of Adelaide intern engineer Daniel Calandro working with them, and Daniel eagerly took up the challenge. Using the proprietary eLabtronics visual CoreChart coding approach, Daniel developed the air suspension controller software.



The solenoids that control airflow. The rear springs are controlled as a pair and the front springs are controlled separately.

Part of my software design request was that key aspects of the software could be user-tuned to suit the application. These adjustable parameters were:

- Hysteresis (the required difference between the driver height request and the actual suspension height before the system reacted)
- Modes 1 and 2 averaging periods
- Sensitivity of the mode selection subroutine

The CoreChart program can easily be brought up on a laptop screen, the parameters in specific boxes altered, and then the revised program directly loaded into the module. This gives the required 'tunability'.

In use

So how well did the world's first user-tunable, DIY automotive air suspension controller work? If I were to say to you that everything was just perfect out of the box, I'd be lying, but with relatively few software re-writes, the system was quickly controlling the suspension height.

The user-adjustable tuning changes were very much a necessity – for example, I have ended up using different averaging settings for the front and back suspensions. I have also chosen to run quite a small hysteresis, so the suspension is maintained within a 10mm height window.

In on-going use, a few hardware changes have been needed. At one stage, the power feed to one of the modules came adrift and inputted large spikes into the 5V regulator, causing it to fail. We've since added some filtering protection on the power supply. The small 5V regulator is also a little marginal when powering multiple parallel pots, as is the case when powering the suspension height sensors and the dashboard pot. It will be easier to power these pots from a separate 5V supply.

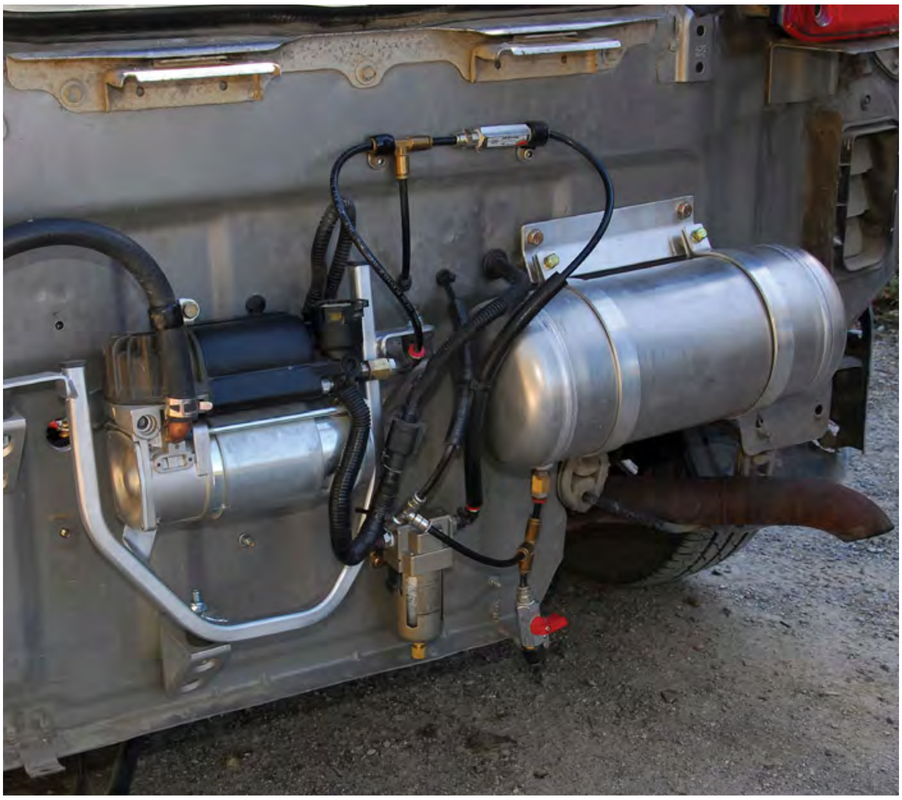
And did I mention the system is fun, too! The best 'party trick' is to sit on the open rear of the car, feel the suspension compress, and then almost immediately feel it rise underneath you to the original level. Hop off the back, and there's a 'shhhhhhhh!' as the air is released and the rear then drops back to its correct level.

On a more serious note, the use of air suspension in a small and lightweight car like the Honda Insight allows it to provide much improved ride quality (especially when carrying large loads) and also allows the driver to lower the car slightly on good road surfaces to improve aerodynamic efficiency and so reduce fuel consumption.

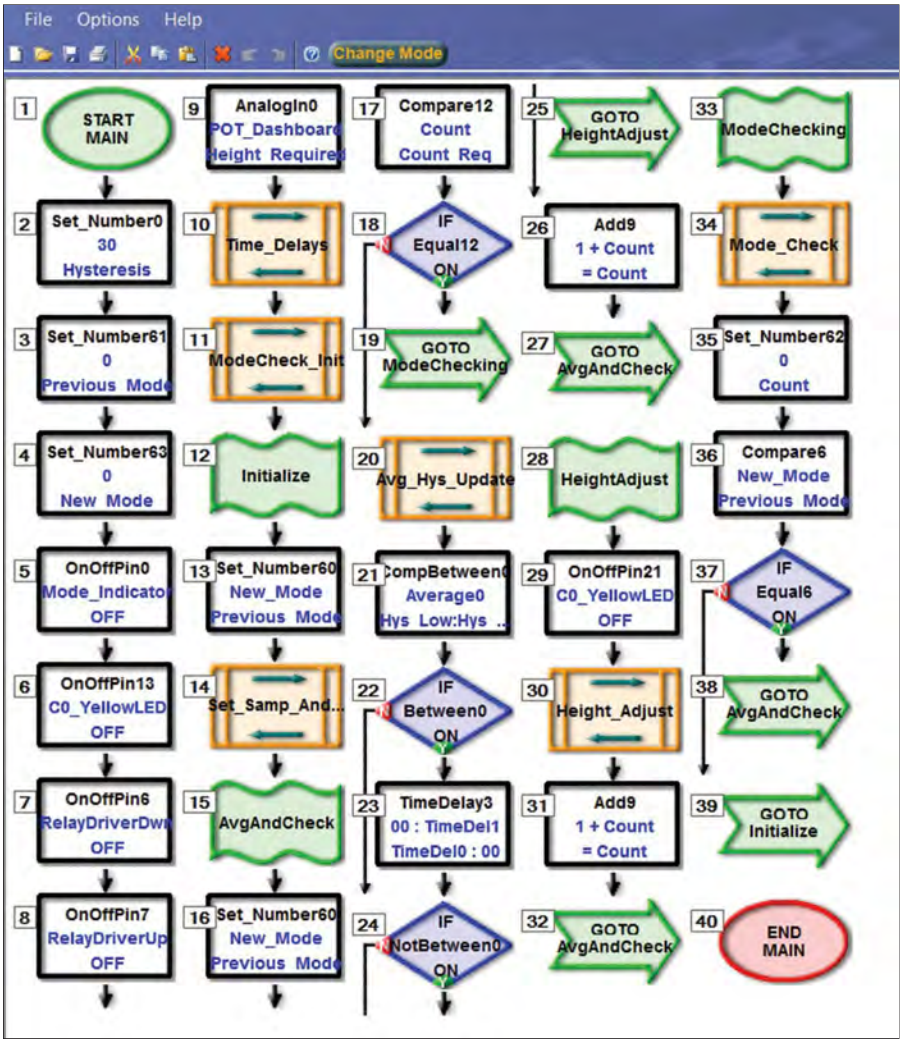
And all possible through the use of a few cheap microcontroller modules!

Disclaimer

I don't benefit financially from the sales of the modules. I am happy to freely work with the company to develop applications that benefit my car modification, and I am very happy for others to also gain from that.



The compressor and air tank – when re-fitted, the bumper completely covers these components. The compressor is from a BMW X5 and the tank from a Porsche Cayenne.



The eLabtronics CoreChart software for the air suspension controller. Changes are easily made to vital tuning parameters to suit different car applications.

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This month, ***EPE*** and **micromite.org** are giving you another chance to win a fully assembled 28-pin MicroMite BackPack Module complete with 2.8” Touch Screen. For details about this powerful, yet compact module, turn to page 12 in the May 2017 edition of ***EPE***.

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4. The winner will need to confirm a valid address for their prize to be shipped
5. A UK winner will have their prize sent via Royal Mail’s Special Delivery service
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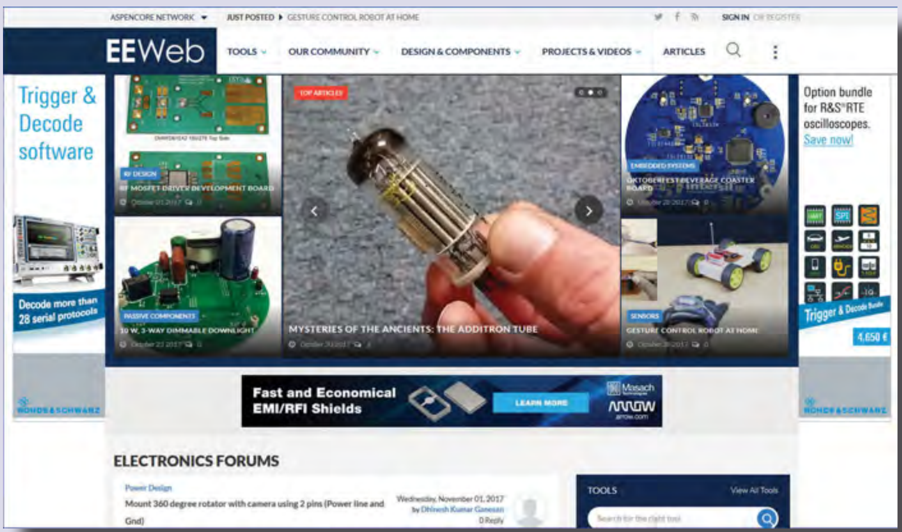


PIC n' Mix

Mike Hibbett

Our periodic column for PIC programming enlightenment

Practical DSP – Part 1

HELLO again! While Mike O'Keeffe takes a break from writing to dive into the joys of parenthood, I am back to fill his boots for a short time. This provides an opportunity for a clean break from Mike's current subjects, so we thought it would be a great time to look into a topic that we have not yet covered in *PIC n' Mix* – namely, 'digital signal processing' – or 'DSP' for short. It's a fascinating area with some fairly complex mathematics behind it, but we will focus on practical applications, and have some fun along the way. First though, here's a bit of background.

What is digital signal processing?

DSP is the process of taking a signal, converting it into a digital form, and performing signal processing or analysis on that digital representation of the original signal. The types of input signals that are suitable for use with DSP algorithms various enormously – audio signals from a microphone, video from a camera, vibration signals from a sensor on a rotating machine shaft are just some examples. You can even apply DSP to slowly changing signals such as daylight levels, or changes in heart rate (very relevant in neonatal care.)

Real-world applications run from the tuning of church bells during manufacture to the detection of planets orbiting distant stars. The latter example has led to the detection of planets outside of our solar system.

There are a number of signal processing techniques that digital signal processing encompasses, but the technique which has fascinated the author since university days is the transformation of a signal from the 'time domain' to the 'frequency domain'.

So, what do these terms actually mean? A signal whose value (amplitude for example) is recorded over a period, a *time* interval, is being represented in the *time domain*. A signal whose value is recorded over a range of *frequencies* is being represented in the *frequency domain*. You can see an example of these two representations of the same signal in Fig.1.

The example in Fig.1 is very simple, but it clearly demonstrates how to visualise the translation of a signal in the time domain into the frequency domain. Consider, however, a situation where your input signal is a mix of many different frequencies. Being able to transform such a signal, in real time, into its component frequencies enables some very interesting processing capabilities. Examples include looking for signals at particular frequencies (akin to bandpass filtering) or even simply visualising the frequency components of a signal – such as for an audio spectrum display on a Hi-Fi amplifier.

The process of transforming a signal from the time domain to the frequency domain is called 'Fourier transformation'. The 18th century French scientist and mathematician Joseph Fourier invented this mathematical system, long before its practical applications were realised.

The computational method for performing the transformation is called the 'discrete Fourier transform' (DFT), which deals with translating a finite set of input signal levels (ie, a short sample period) into a corresponding set of frequency components. The actual algorithm that is used within digital systems is called the 'fast Fourier transform' (FFT.) This is an algorithm that yields the same results as a DFT, but is highly optimised. Surprisingly, the DFT algorithm was also discovered long before any practical applications were available, and certainly well before the advent of computers.

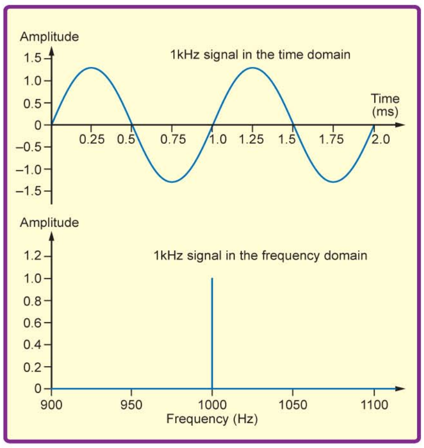


Fig.1. Two different ways of displaying the same signal – a 1kHz sinewave graphed in the time and the frequency domains

In this series of articles we will be exploring the application of the DFT to analogue signals. Many of the applications of digital signal processing require expensive digital hardware and sensing systems, but there are many practical applications that can fall within a hobbyist's price range, while still being useful, educational and fun.

The consequence of limiting the cost of the system we create will be resolution, signal frequency and accuracy. There are however many applications where those specific limitations are perfectly acceptable.

Fast!

The mathematics behind FFT is complex, and beyond the scope of these articles. However, for our purposes all we need to know is some of the background and principles of application. Fortunately, the engineers at Microchip have created highly efficient DSP software libraries that we can access and build into our projects, and Microchip provide these libraries for free. If this series of articles whets your appetite to dig deeper into the theory there are a number of very good university tutorial videos available for free, on-line. An excellent introductory overview is given here: **https://youtu.be/spUNpyF58BY**. (Incidentally, the same YouTube author also offers some excellent introductions to calculus, neural networks, machine learning blockchain technology and other important and topical mathematical subjects – thoroughly recommended!)

The principle of operation is straightforward enough: take a series of samples of your data at regular time intervals, and store them into a buffer. Adjust those samples to match the format required by your FFT function and then call the FFT function to transform the data from the time domain to the frequency domain. Last, examine the resulting data and take an action on it. This workflow is shown in Fig.2, highlighting what operations occur within the hardware circuit, and what occurs on-chip.

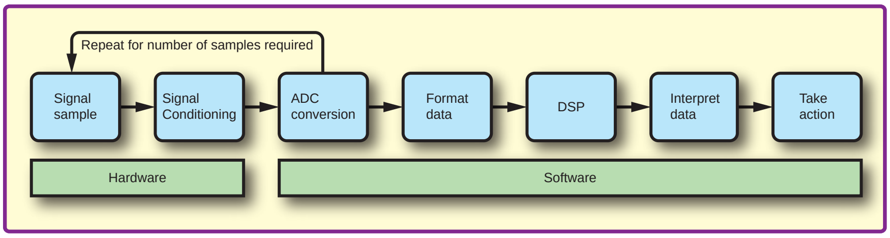


Fig.2. Overview of a DSP workflow (hardware and software)

The signal conditioning stage is a combination of scaling your signal to maximise the full range of the ADC input voltage swing, and also some low-pass filtering to minimise any high frequency signals or noise, above our sampling frequency, which many cause unexpected 'ghost' signals to appear in your data. This process is called 'anti-alias filtering'. We will explore the effects of aliasing next month.

Processor selection

Let's now move away from the theory and look at how we can start experimenting with a practical solution. Microchip offers a number of processors with DSP-assisting instructions built into the chip, but it basically comes down to a choice between three processor families: the PIC32MZ, dsPIC30 or dsPIC33. The PIC32MZ has the highest clock speed and largest memory availability, but we have selected the dsPIC33 family as this range offers devices with dual-in-line (DIL) packaging – ideal for simple breadboard tests. Specifically, we have selected the dsPIC33EP512GP502-I/SP, a 28-pin DIL device, which is cheap, readily available from distributors such as Farnell (part code 2406557) and has plenty of on-chip resources. It is also compatible with the PICKIT-3 debugger/programmer, our favourite low-cost interface. Here is a summary of what you get from this processor, all for the price of a pint of beer:

- 512KB Flash memory
- 48KB RAM
- 500,000 samples per second 12-bit ADC
- CAN bus interface
- 21 GPIO pins
- DSP-instructions
- On chip high-speed oscillator
- 70MHz clock speed
- 3.0V to 3.6V operation

Plus, all the usual timers, PWM, SPI, I²C interfaces that you would expect from Microchip. The pin-out for this device is shown in Fig.3.

With the on-chip high speed oscillator, calibrated to within 1%, there is no need to add an external crystal oscillator. This is ideal, as it allows us to build the test circuit on a breadboard, with just a few decoupling capacitors and pull-up resistor required to complete the minimal hardware.

Software

Many microcontroller vendors provide DSP algorithms within their free software libraries, even for processors that do not have special DSP instructions in hardware. Microchip provide two sources of DSP libraries: one within the Harmony software framework, designed specifically for the PIC32MZ family of processors, and second, a separate dsPIC DSP library, provided as part of the XC16 compiler, for the dsPIC family of processors. Since we have selected the dsPIC33 processor we will make use of the latter. Do be careful not to confuse the two should you search on-line.

We will build our software within the MPLAB-X IDE, using the free version of the XC16 compiler. With the addition of the PICKIT-3 debugger/programmer unit, this represents a very low cost set-up. We will cover the XC16 compiler installation next month.

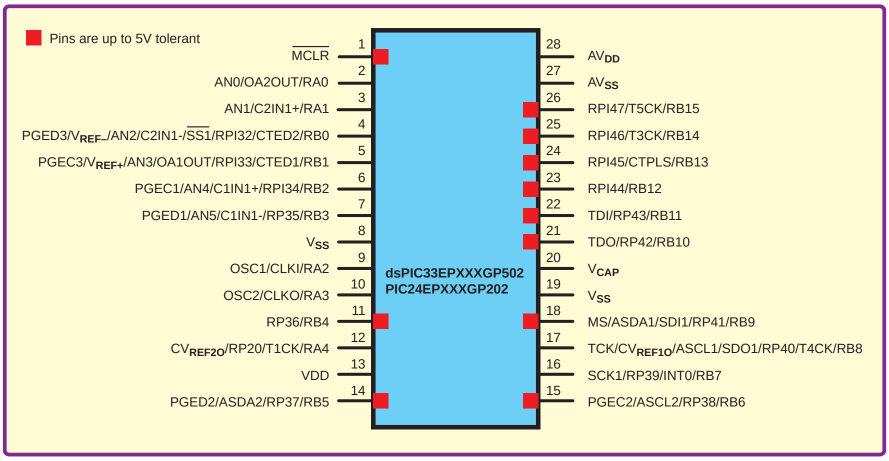


Fig.3. Pin out for our digital signal processing PIC of choice – the inexpensive 28-pin dsPIC33EP512GP502-I/SP, a DIL device.

Hardware design

Fig.4 shows the minimal components required to start testing the DSP capabilities of the dsPIC33 processor. Three 100nF capacitors, a 10µF tantalum capacitor and a 4.7kΩ resistor complete the basic requirements. Two LEDs have been added here to aid debugging; half a dozen hook-up wires need to be added to make this breadboard good to go and ready to connect to the PICKIT-3 debugger/programmer interface.

If you decide to follow along with us on this journey and are using a similar breadboard setup, make sure your breadboard is a good one, and not something you have had in your shed for the last ten years. While all the high frequency signals are on-chip, your ability to program the device reliably may be impacted by old, tarnished connections. Fault finding these issues can be frustrating!

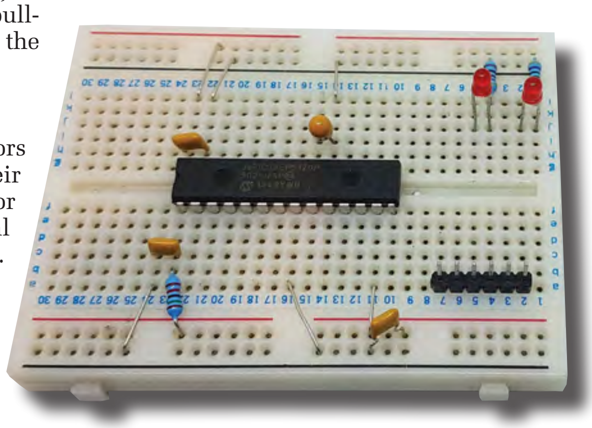


Fig.4. Our basic DSP hardware – details next month

The PICKIT-3 also serves as a power supply for the processor, freeing us from the need to provide an additional regulator circuit to the board. At a push, however, the circuit will run directly from a pair of AA cell batteries because the processor can operate over the 3.0V to 3.6V range.

As you can see from this board layout, initially, we are not connecting sensors. Our first software tests will be made with pre-computed data samples, allowing us to know exactly what data is going into the FFT function. This minimises the number of things that can go wrong. We will add sensor input via the ADC once the basics are working.

Our aim in this series of articles is to leave you with a reference hardware design and source code that will enable you to explore this fascinating branch of electronics to your own ends. We hope to hear from you on your progress in DSP!

Next month

In Part 2 we will discuss building the initial hardware and look at how to incorporate the DSP libraries provided by Microchip.

Circuit Surgery

Regular Clinic

by Ian Bell

High-frequency signals and low-pass filters

A LITTLE while ago user ***741*** posted a question about low-pass filters on the *EPE Chat Zone*. As regular readers will know, this forum has since been retired and *EPE* 'chat' is now hosted on EEWeb (**www.eeweb.com/forum**). The old forum is still available in read-only-mode, but new discussions can be started on EEWeb using the tag 'EPE Magazine' to mark threads as being *EPE*-related.

Returning to ***741's*** question, he asks: 'Op amps have GBW (gain-bandwidth product), say 10MHz. If this was an amplifier circuit (not a filter), then (ignoring phase gremlins), the relation $G = 1/\beta$ gets less accurate as A (open-loop gain) falls with frequency. At 100kHz, A is 100, so the relation holds well, but at 1MHz, we need to use $G = A/(1+A\beta)$... I think.

How does the performance of say an MFB (multiple-feedback) low-pass filter suffer with rising input frequency? Suppose we have a 1kHz low-pass filter. You feed in 10kHz, and it maybe works 'well'. You apply 1MHz, and it starts to act up.'

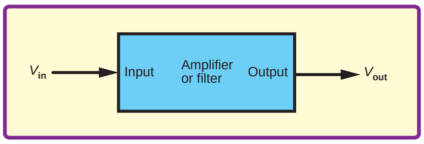


Fig.1. Simple electronic system; the gain is v_{out}/v_{in}.

Back to basics

Starting with circuit fundamentals, a filter or amplifier is an electronic system (see Fig.1) which reproduces the signal on its input (v_{in}) on its output (v_{out}). Typically, the output has different amplitude and power delivery capabilities. The ratio of output to input signal level v_{out}/v_{in} is the voltage gain (G) and very often (as in ***741's*** question) we are interested in how the gain of a circuit varies with frequency, known as the 'frequency response' of the circuit. Often, a frequency response is shown by plotting a graph of gain against frequency (see Fig.2, which we will discuss in more detail shortly). Such a plot can obtained by measurement in a lab, by simulation or by plotting a mathematical function derived from analysis of the circuit. The frequency response graph can be used to help decide if a circuit is appropriate for our intended purpose. Frequency response requirements may be used as part of the specification for a circuit design.

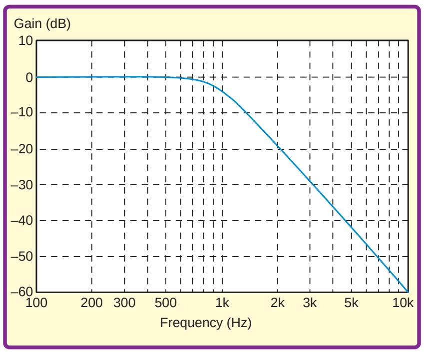


Fig.2. Example frequency response graph of a 1.0kHz low-pass filter.

The question also prompts some more specific points – how we interpret the well-known negative feedback gain equation $G = A/(1+A\beta)$ in the context of op amp filters and why the performance of specific low-pass filter circuits may degrade at relatively high frequencies. We will look at this after introducing some basics of frequency response plots and filter characteristics.

Filters are circuits that pass signals (from input to output) for a certain range of frequencies (called the pass band) while rejecting signals at other frequencies (in the stop band). For signals within the pass band the filter may have high gain, low gain or even attenuation – the key thing is not the specific gain, but the fact that the gain will diminish significantly with respect to the pass band for frequencies outside this pass band. As its name implies, a low-pass filter should pass signals only if they are below a given frequency, called the 'cut-off frequency'.

Decibels

On frequency responses graphs gain is often plotted using a scale in decibels (dB) versus the logarithm of frequency. The definition of a decibel is also logarithmic and is based on the power ratio of two signals P_1 and P_2 – specifically, the power ratio in decibels is given by: $10 \times \log_{10}(P_2/P_1)$ dB, where P_2/P_1 is the power gain (eg, P_{out}/P_{in}), or P_1 is an agreed reference level and P_2 is a measured value. The term 'decibel' means one tenth (deci, hence d) of a bel (symbol B). One bel is $\log_{10}(P_2/P_1)$, but as we use $10 \times \log_{10}(P_2/P_1)$ we are counting in tenths of a bel. The bel is named after Alexander Graham Bell.

The definition of the decibel is based on power, but we are often interested in voltage levels and voltage gains. Power is related to the square of voltage, specifically $P = V^2/R$ for a voltage (V) across a resistor (R). If we square something inside a logarithm it is equivalent to multiplying the log by two (without the square); that is, $\log(y^2) = 2 \times \log(y)$. So for voltage gains, we assume a reference resistor (R) that cancels out when we find $P_{out}/P_{in} = (V_{out}{}^2/R)/(V_{in}{}^2/R) = V_{out}{}^2/V_{in}{}^2$, we get a voltage gain in decibels of $2 \times 10 \times \log_{10}(V_{out}/V_{in}) = 20 \times \log_{10}(V_{out}/V_{in})$.

On the decibel scale, a gain of 1 is 0 dB, gains greater than 1 (amplification) are positive and gains less than one (attenuation) are negative. The graph in Fig.2 shows a filter with unity-gain at low frequencies and increasing attenuation at high frequencies. The scale points of –10, –20, –20, –40… correspond to attenuations of 1/3.2, 1/10, 1/32 and 1/100 respectively. To covert from a voltage gain (G) expressed in dB to a simple numerical gain find $10^{(G/20)}$. For a power gain use $10^{(G/10)}$.

Being able to interpret decibel values, and plots using decibels and log frequency is very useful when working with filters. The log-log scale allows details of responses to be seen over a very wide range of frequency and gain, which would be lost with a linear plot. However, if they are not read correctly the importance of small features may seem exaggerated – you need to be able to interpret the dB values in terms of their actual relevance to the circuit you are working with. Having a feel for what the numbers mean (eg, a –40dB output is 100 times smaller than the input voltage) helps.

Brick wall

An ideal low-pass filter would pass all signals below the cut-off and completely reject all signals above the cut-off. The frequency response graph would look like Fig.3. Unfortunately, this perfectly sharp cut-off, which is referred to as a 'brick wall' filter, cannot be physically implemented.

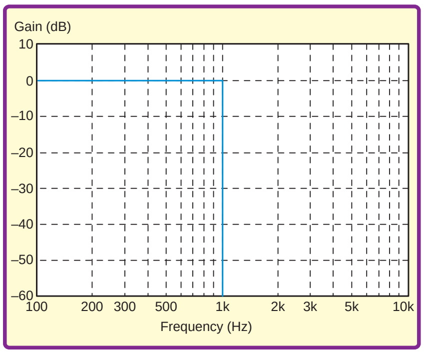


Fig.3. Frequency response graph of a 1.0kHz 'Brick Wall' low-pass filter. This perfect filter cannot be built as a real circuit.

Instead, real filters have a more gradual decrease in gain in the transition from pass band to stop band; for example, the frequency response in Fig.2. The question then arises as to where the cut-off is taken to be along the smooth curve. The usual answer is that the cut-off is defined to be the frequency at which the output power from the filter falls to half that in the pass band, this is $10 \times \log_{10}(0.5) = -3$ dB. For example, for the response shown in Fig.2, the cut-off frequency is 1kHz (the same as the brick wall filter in Fig.3).

All filters deviate from the ideal brick wall filter, so the question a circuit designer faces is what is good enough, or most suitable for the application? The purpose of a filter is to remove unwanted frequency content from a signal because it will cause problems in the following circuitry or output signal. The attenuation of the unwanted signal has to be sufficient to prevent its presence from being problematical. The filter specification becomes more demanding if the wanted and unwanted signals are close in frequency, because this requires a filter to be closer to the ideal, but impossible-to-achieve brick wall response. In many situations where a filter is required the unwanted frequencies will be close, or moderately close, to the wanted signal, and there may be very little unwanted signal strength at much higher frequencies. In such cases the performance of the filter at these higher frequencies may not be particularly important. However, this is not always the case and some applications demand filters with good performance over a very wide range of stop band frequencies.

It is therefore important to understand the limitations of some types of low-pass filter circuit at high frequencies. This is exactly the issue raised by ***741***.

Feedback equation

741's question makes reference to the feedback gain equation $G = A/(1+A\beta)$, in which G is the gain of the whole circuit with feedback, A is the gain of the amplifier to which feedback is applied and β is the proportion of the output signal applied as negative feedback. A is called the 'open-loop gain' of the amplifier, this does not change when the external feedback is applied – it is an internal property of the amplifier. However, the gain of the whole circuit is strongly dependent on the feedback, as well as being influenced by A. We have discussed feedback theory in *Circuit Surgery* (November 2016) and showed how this equation simplifies to $G = 1/\beta$ if A is very large (eg, hundreds of thousands), which it typically is for op amps at low frequencies. This is an important result because it means the gain of the circuit only depends on the external feedback components (eg, the resistors in an op amp amplifier circuit) and not on the value of the open-loop gain of the amplifier itself.

The $G = A/(1+A\beta)$ equation is obtained by drawing a system diagram to represent the relationships between the signals in the circuit and finding the gain (output/input) of the system. This diagram is an abstract representation of the behaviour of the circuit and is used because it provides a useful basis for mathematical analysis. System diagrams like this are frequently used in control systems design – the technique is applicable well beyond analysis of circuits such as amplifiers.

System diagrams

Some standard op amp amplifier circuits are shown in Fig.4. The most straightforward case for developing a system diagram is the non-inverting amplifier. Here it is clear to see that the output is fed back to the inverting input via a potential divider. The fraction of the output voltage that appears at the junction of the two resistors is the feedback fraction β. The circuit input is applied directly to the non-inverting input of the op amp. Thus the input is added and the feedback subtracted at the amplifier input. This leads to the system diagram in Fig.5, which can be used to derive the equation ***741*** quotes. This was done in the November 2016 article, so we will not repeat it here.

An important point about the system diagram in Fig.5, and the equation obtained from it, is that it applies specifically to the non-inverting amplifier. It may, but does not necessarily, apply to other amplifier circuits. For example, as was also pointed out in the November 2016 article, the system diagram for the inverting amplifier is not exactly the same as the one in Fig.5. This means that the equations $G = A/(1+A\beta)$ and $G = 1/\beta$ do not apply to the inverting amplifier; for example the simplified equation for the inverting amplifier is $G = (1–1/\beta)$. This is not always made clear when op amp feedback theory is discussed.

The non-inverting amplifier is sufficient for explaining the general principles of feedback theory, so the variants are not presented, and it is easy to assume the non-inverting equations apply to all circuits. This is directly relevant to ***741's*** question – in a later post he commented 'It's hard to work out what β is for many filters'. Filter circuits may have more complex feedback structures than basic amplifiers, so the assumption that the non-inverting amplifier feedback equation can be straightforwardly applied may be wrong.

To obtain equations for the input-output relationships of single op amp filters, like the MFB filter ***741*** mentions, we typically apply basic circuit theory (eg, Ohm's and Kirchhoff's laws) together with idealised op amp characteristics (zero input current, and zero voltage across the inputs due to the very high gain in combination with negative feedback). On the other hand, the basic theory that gives us $G = A/(1+A\beta)$ and $G = 1/\beta$ does help explain, in general terms, what happens to op amp circuits as frequency increases and this is related to high-frequency performance issues with filters such as the MFB.

Compensation

Fig.6 shows the frequency response of a typical op amp's open-loop gain and the closed-loop gain of an amplifier (eg, inverting and non-inverting amplifiers) built using the op amp. In

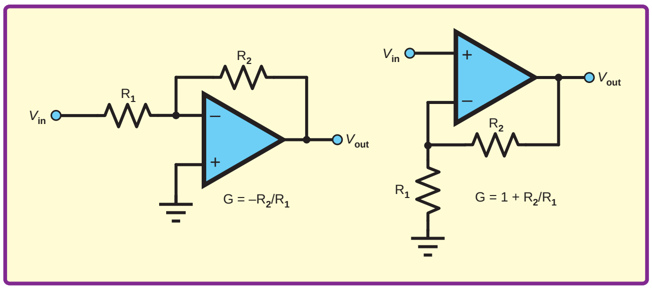


Fig.4. Op amp amplifiers

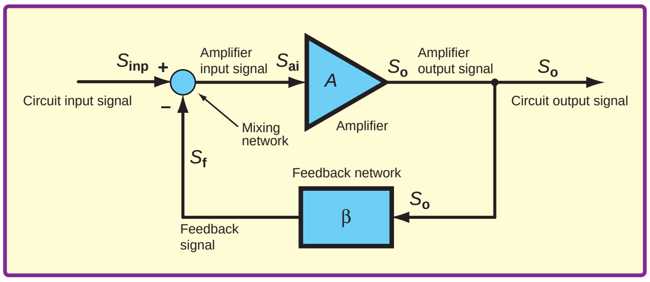


Fig.5. System diagram for the non-inverting op amp amplifier

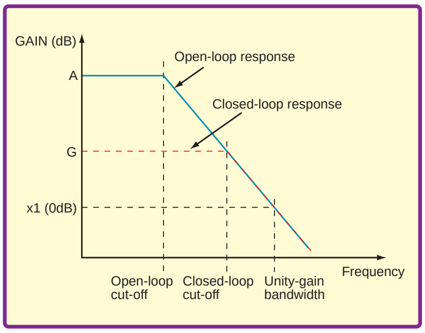


Fig.6. Typical form of the open- and closed-loop frequency responses of an op amp amplifier.

both cases this is a low-pass response. The direct-coupled, DC-amplifying, nature of the op amp circuit means that the gain does not also roll off at low frequencies, as it would for a capacitively coupled amplifier. The open-loop op amp has full gain at DC, but this starts dropping at a very low frequency compared to those at which many op amp applications operate. The cut-off frequency is typically 1 to 10Hz, with the falloff in gain above that frequency being 20dB per decade (tenfold increase in frequency). This low cut-off frequency for the open-loop gain is deliberately designed into the op amp to prevent instability when feedback is applied and is known as 'compensation'.

The closed-loop gain (G) of an op amp amplifier is set by the feedback resistors, as mentioned above, and is usually much lower than the open-loop gain. As frequency increases, the closed-loop gain remains constant at $1/\beta$ until the approximation $A = 1/\beta$ no longer holds due to the open-loop gain (A) at that frequency (and above) being too low. At these higher frequencies the closed-loop gain decreases along with the open-loop gain (see Fig.6). The cut-off frequency for the closed-loop response is typically at a much higher frequency than the cut-off frequency point of the open-loop gain.

As the op amp's gain extends from DC, the frequency at which the closed-loop gain starts dropping off is equal to the circuit's bandwidth (range of frequency it can amplify) at that closed-loop gain. The lower the closed-loop gain, the higher the frequency at which the closed-loop frequency response intersects the open-loop response and starts decreasing. The straight-line nature of the open-loop response in Fig.6 means that if you multiply the closed-loop gain and closed-loop bandwidth together for any closed-loop gain you get the same value. This is called the 'gain-bandwidth product' (GBW) and is an indication of how 'high frequency' an op amp is. Fixed GBW is common for op amps, but not universal, as it depends on the form of compensation used.

From the preceding discussion we see that application of negative feedback to a very high gain amplifier (op amp) makes the circuit gain insensitive to the gain of the amplifier itself. This effect breaks down at frequencies where the amplifier's gain is not significantly larger than the circuit gain. Negative feedback has other significant effects on circuit performance, for example it reduces the effective output impedance. Like amplifier-independent gain-setting, the feedback magic is lost for these effects too at high frequencies; for example, effective output impedance may increase at high frequencies. The equations $G = A/(1{+}A\beta)$ and $G = 1/\beta$ quoted by ***741*** may only relate to gain and may only apply to specific circuits but, together with the op amp frequency response curve, they demonstrate the general principle of feedback providing improved circuit performance, which is lost at high frequencies due to diminished open-loop gain. The problem with the MFB's (and similar low-pass filters') gain increasing at high frequencies is due to an increase in output impedance.

Analysis

Fig.7 shows a Sallen-Key filter, which is a single op amp filter circuit, like the MFB filter mentioned by ***741***. At very high frequencies the capacitors in the circuit behave like short circuits leading to the simplified version of the circuit shown in Fig.8. The op amp has been replaced by a unity-gain buffer (this is the role it performs in this circuit). Given that the buffer's input is shorted to ground, the buffer's internal voltage source is producing 0V. This is the same as connecting the internal side of the output impedance to ground, allowing us to further simplify the equivalent circuit to that shown in Fig.9.

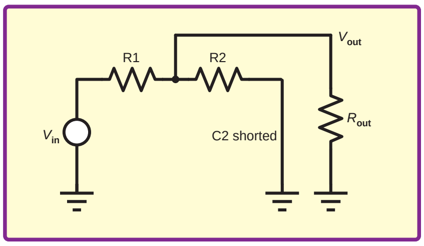


Fig.9. Further simplification of the equivalent circuit in Fig.8.

Looking at Fig.8 we see that R_2 and R_O are in parallel, but typically R_2 is at least 100-times larger than R_O, so we can ignore R_2. Thus v_{out} is related to v_{in} by the potential divider formed by R_1 and R_O, and so:.

$$V_{out} = \frac{R_O v_{in}}{(R_O + R_1)}$$

However, again, R_1 is typically about 100-times larger than R_O, so in this case we can ignore R_O in the sum of the resistances, giving an approximate value of v_{out} as:

$$V_{out} = \frac{R_O}{R_1} v_{in}$$

The output of the real filter circuit is a combination of the unwanted signal given by the above equation and the ideal filter response. At low frequencies the ideal response dominates, but as frequency increases, the op amp's gain decreases, so the effective value of R_O increases, increasing the contribution of the unwanted signal to the total output. At high frequencies this unwanted signal dominates the circuit's output. The MFB suffers in a similar way, although the equivalent circuit model is different.

Simulation

The behaviour of Sallen-Key and MFB filters at high frequencies can be observed using simulations. Fig.10 shows a schematic used for a simulation of MFB and Sallen-Key filters implemented using Linear Technology's model of their LT1002A op amp, which is supplied with LTspice. The Sallen-Key filter is also implemented with an ideal op amp

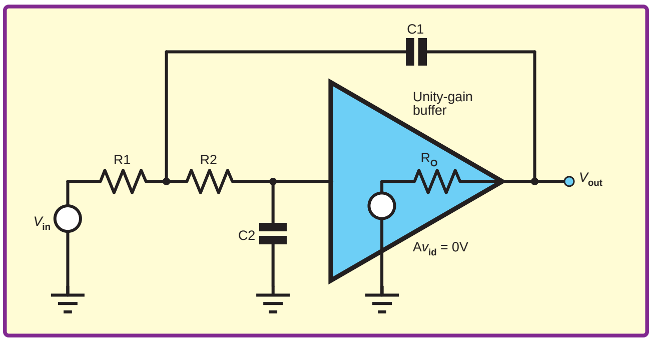


Fig.7. Sallen-Key filter.

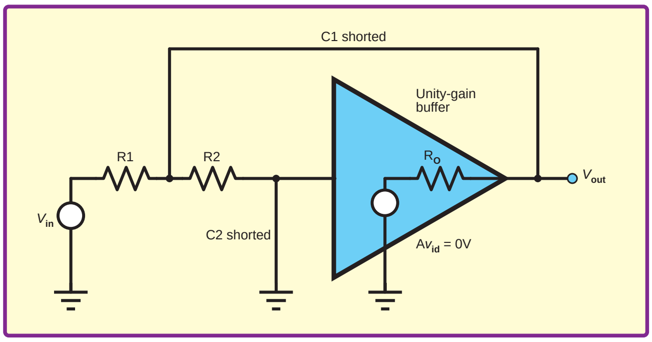


Fig.8. Simplified equivalent circuit for the filter in Fig.7 at very high frequencies.

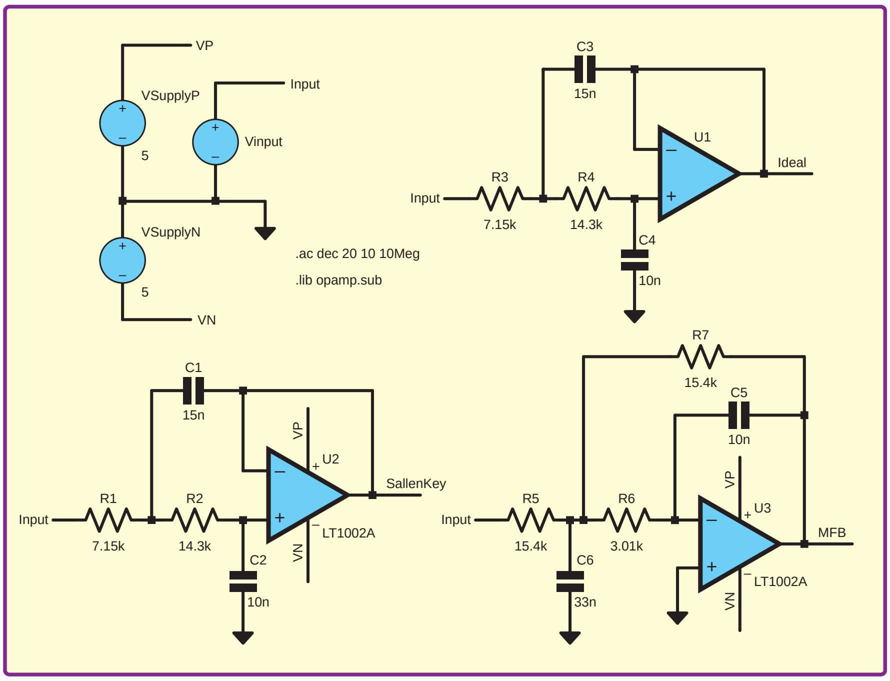


Fig.10. Schematic for LTspice simulation comparing ideal, Sallen-Key and MFB low-pass filters.

model, which will have zero output impedance at all frequencies and hence not suffer from the problem described above.

The results of the simulation are shown in Fig.11 and demonstrate the issue mentioned by ***741***. It can be seen that both the Sallen-Key and MFB filters suffer from the problem of increased gain at high frequencies in comparison with the ideal circuit, but the MFB circuit exhibits better performance. The exact performance of the circuits will depend on the op amp used and relevant external component values. The simulation only shows the effect of using the realistic op amp model. The influence of other components' non-ideal behaviour and physical layout are not included, but may need to be considered in real circuits at high frequencies. In this example, the Sallen-Key circuit attenuates signals above 1MHz by 40dB (one hundredth), whether or not this is acceptable would depend on the application. The MFB circuit is much better: –100dB at 1MHz is an attenuation of one hundred-thousandth.

Further reading

Readers interested in more details on this topic are directed towards Jim Karki's application reports from Texas Instruments.

Karki, J., Active Low-Pass Filter Design, Rev. B, 2002, *Texas Instruments Application Report SLOA049B*.
www.ti.com/litv/pdf/sloa049b

Karki, J., Analysis of the Sallen-Key Architecture, Rev. B, 2002, *Texas Instruments Application Report SLOA024B*.
www.ti.com/litv/pdf/sloa024b

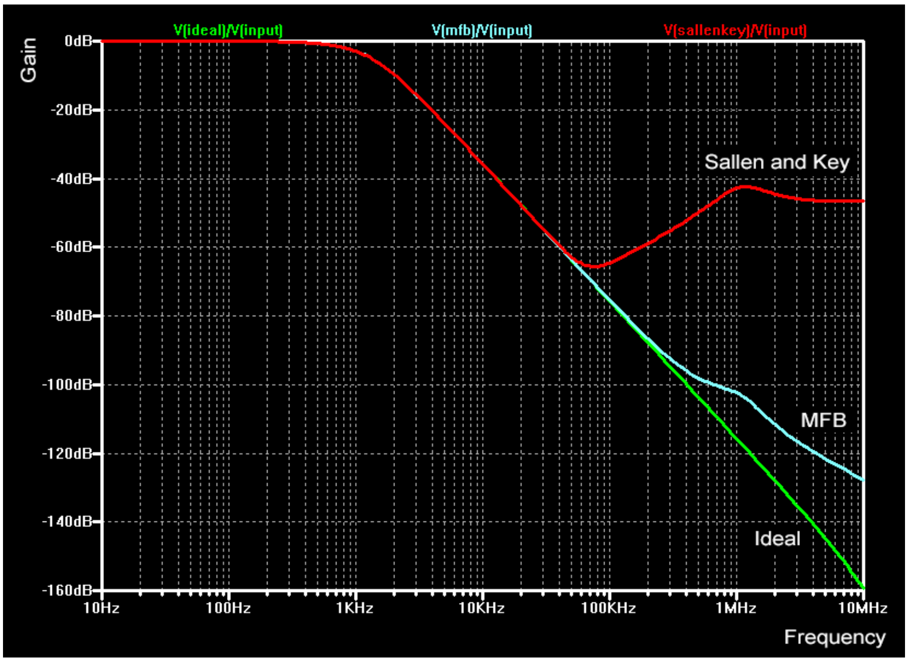


Fig.11. LTspice simulation results from the circuit in Fig.10.

AUDIO OUT

By Jake Rothman

Variable audio filters – Part 1

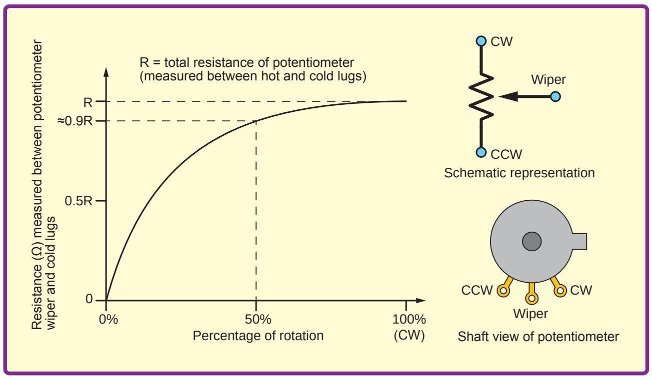


Fig.1. Inverse logarithmic (anti-log) potentiometers are needed for filter controls, since frequency goes up as resistance is reduced – the frequency increasing as the knob is turned clockwise (CW).

After amplifiers, filters are the next most important circuit elements in audio because the basic building blocks of sound are amplitude and frequency. Musical instruments, the human vocal tract and loudspeakers all embody multi-pole filters. In *Audio Out*, Jan 2016 I gave design details for a fixed single-frequency Sallen-Key filter block using *fixed* resistors and capacitors (see also *Teach-In 7*). However, a *variable* corner frequency filter is often needed for audio/music applications, since it is essential to be able to tune a circuit for the best musical effect. In most audio equipment, this is usually done manually with switches and potentiometers. In musical instruments, such as synthesisers, and dynamic equalisers, much faster control is needed, and to achieve this the filter's (corner) frequency needs to be electronically controlled and we will look at some ways of doing this next month.

Fig.2. Not all log/anti-log potentiometer curves have the same inflection points. Be careful when calibrating frequency controls – change the potentiometer and you may have to reprint the front panel. This Behringer fourth-order active crossover frequency control was remarkably inaccurate.

Going potty

Just as with volume, human frequency perception exhibits a logarithmic ('log' for short) response. We respond to octaves, which are frequency doublings. We hear the pitch difference between 200Hz and 400Hz, for example, as the same as between 400Hz and 800Hz. In conventional RC filter circuits the corner frequency is inversely proportional to the change in capacitance or resistance. Therefore, a potentiometer's corner-frequency-setting resistance has to decrease with clockwise rotation to increase the corner frequency. Also, for a variable frequency panel control to sound correct as it is rotated, the increase in frequency must be logarithmic. This means that inverse log (anti-log) potentiometers have to be used, as illustrated in Fig.1. Anti-log potentiometers are generally denoted by the letter 'C' after their value. These potentiometers are the opposite (inverse) to the normal log pots used for volume controls and can be hard to source because they are only used in audio, but they are available from Tayda (**www.taydaelectronics.com**).

Remember that all potentiometer audio-taper laws are approximations and the curves differ between different makes; for example, Alps potentiometers have their half-way point at 83.5% of track resistance, while Bourns are usually at 90% with respect to the full CW rotation track value. This does not matter much with normal volume controls, where the ear readily accommodates such differences, but where exact calibration of controls is expected, such as the exact frequency values on an active crossover, it can be a problem. (I measured the frequency setting on a Behringer active crossover to be 780Hz when set to 1.2kHz, an error of 58% – but at least it was correct at both extremes of the potentiometer rotation! – see Fig.2.) Some expensive equipment such as API EQ units avoid the variability of potentiometer audio-taper

Fig.3. Special dual-track potentiometer for filters with only three pins (six are normal).

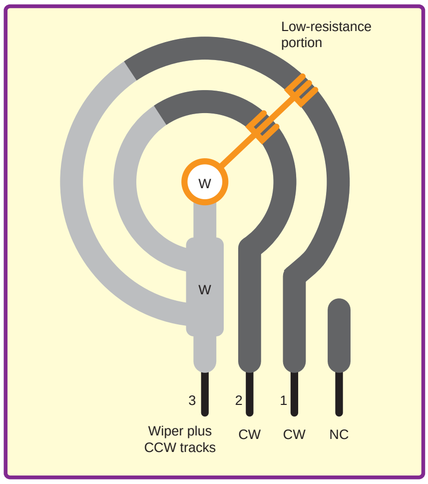
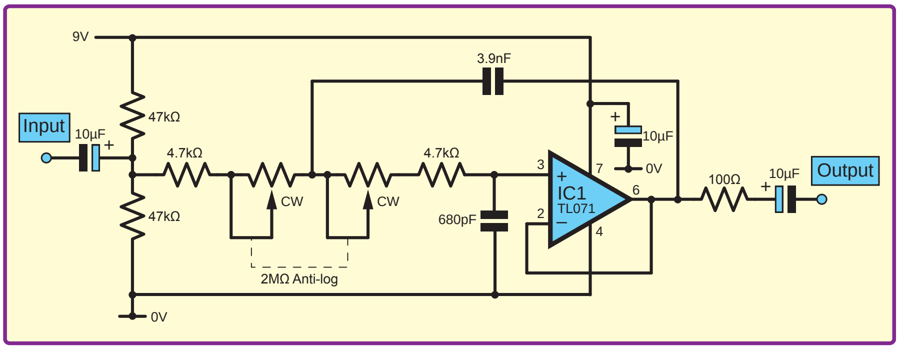


Fig.4. Internal track construction of a special filter potentiometer.



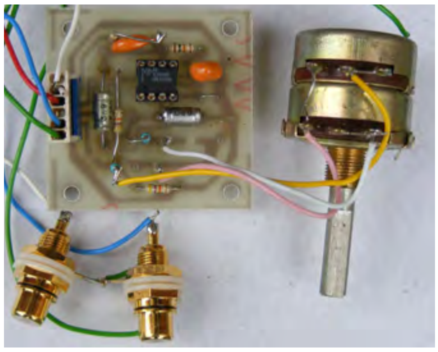


Fig.5. (left) Sallen-Key filter for a sound effects generator, (right) filter PCB with dual-gang potentiometer – note only three wires to pot needed

laws by using gold-plated switches and lots of well-defined 1% resistors for repeatability – essentially in the style of R&D test equipment. However, in musical applications, potentiometers are preferred because users need to be able to set a frequency between two discrete steps and hence have the ability to *smoothly* tune by ear.

Occasionally, some strange frequency controls turn up. When replacing some dual-gang equaliser (EQ) frequency potentiometers in an Allen and Heath GL3300 mixer, I was surprised to find they had three connections rather than the expected six, as shown in Fig.3 and 4. This was possible because both sections were wired as two-terminal rheostats/variable resistors and not potentiometers. It eliminates one wiping contact, a major source of noise and wear.

Variable Sallen-Key filter

Let's look at a real example. I recently modified the Sallen-Key filter design described in *Audio Out*, Jan 2016 to create a variable-corner-frequency filter by replacing the fixed resistors with a dual-gang potentiometer. This was used in a synthesiser based around a ring-modulator made for Eat Static (**www.eatstatic.co.uk**). Most musical instrument filters are second-order low-pass types, because this mimics the effect of sound boards and wooden bodies, and this is what is described here. An equal-component-value version of the Sallen-Key could have been used, but that would have introduced unwanted gain. The version I opted for is shown in Fig.5. It uses two different-value capacitors, so only a unity-gain voltage follower is needed. I used

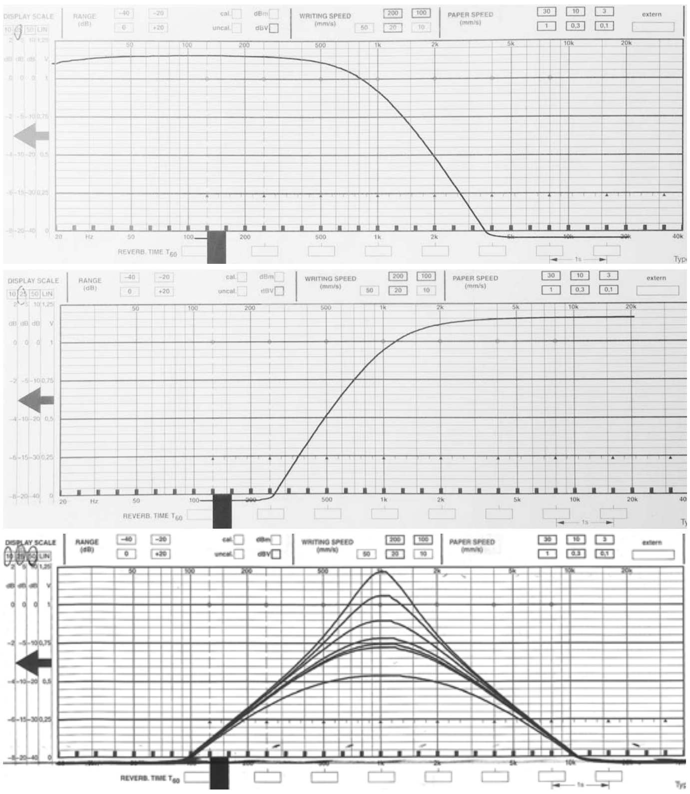


Fig.6. Response curves of state-variable filter: a) (top) Low-pass, b) (middle) high-pass c) (bottom) band-pass showing different Q settings. Note the scale is 1dB per a division. The whole vertical scale is 25dB.

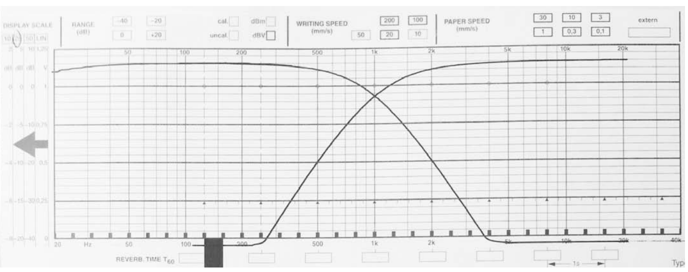


Fig.7. State-variable response curve showing crossover function by overlaying high-pass and low-pass response curves.

a 2MΩ anti-logarithmic dual-gang Plessey moulded-track potentiometer to get a large frequency range and long life. Because of the potentiometer's high resistance I had to use a TL071 FET input op amp. It was consequently noisy, but that didn't matter because there was a VCA after it, muting the hiss during gaps between notes. The circuit was constructed on the standard Sallen-Key PCB shown, also described in Jan 2016.

State-variable filter

Another fixed-component filter topology that can be adapted to use variable components is the state-variable filer (SVF). This design is more complex than Sallen-Key, but that added complexity simultaneously provides three filters in one: low-pass, band-pass and high-pass outputs (see Fig.6a, b, and c) which opens up many possibilities for more complex signal processing. It does this by using a pair of integrators to form a low-pass filter then subtracting this low-pass output from the input signal, generating an exact mirror-image high-pass output. Its dual high-pass and low-pass outputs (Fig.7) make it ideal for use in active crossovers because it can split the audio signal into bass and treble components to feed separate amplifiers for the woofer and tweeter. Active crossovers for PA systems usually use an SVF because only half as many resistors are needed to set the crossover frequency compared to using Sallen-Key filters. This halves the number of gangs required on the frequency potentiometer – an important engineering consideration.

One problem with the state-variable approach is that the high-pass and low-pass cut off frequencies are exactly the same for Butterworth response. If a large offset/overlap with a Butterworth characteristic or different value of Q around the crossover is needed, as is often the case with real loudspeaker drive units, then the designer must resort to two separate filters. The degree of overlap can be controlled to some degree however with the SVF by varying the Q (Fig.8).

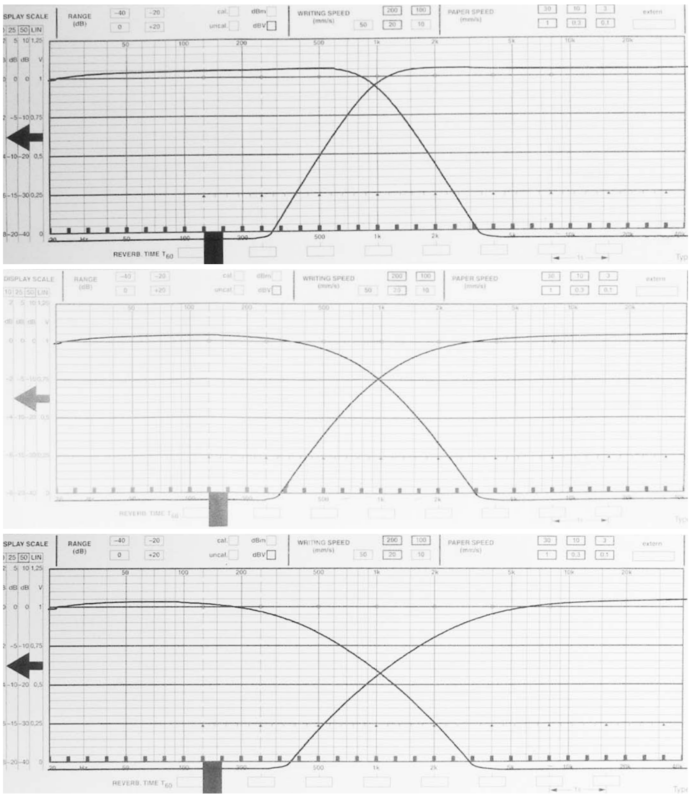


Fig.8. Varying the overlap: a) (top) minimal dip, moderate Q, b) (middle) maximum dip, minimum Q, c) (bottom) no positive feedback.

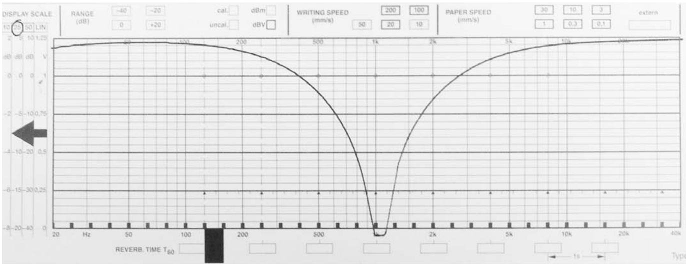


Fig.9. Notch function generated by adding high and low-pass outputs in a state-variable filter.

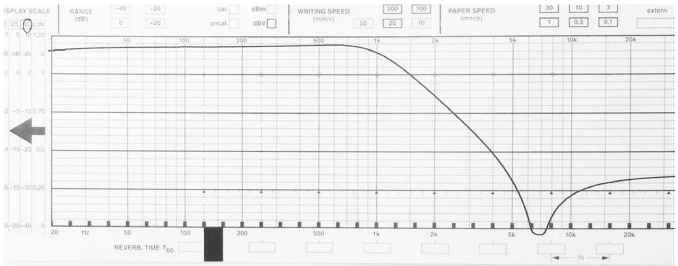


Fig.10. Elliptical filter function from a state-variable filter

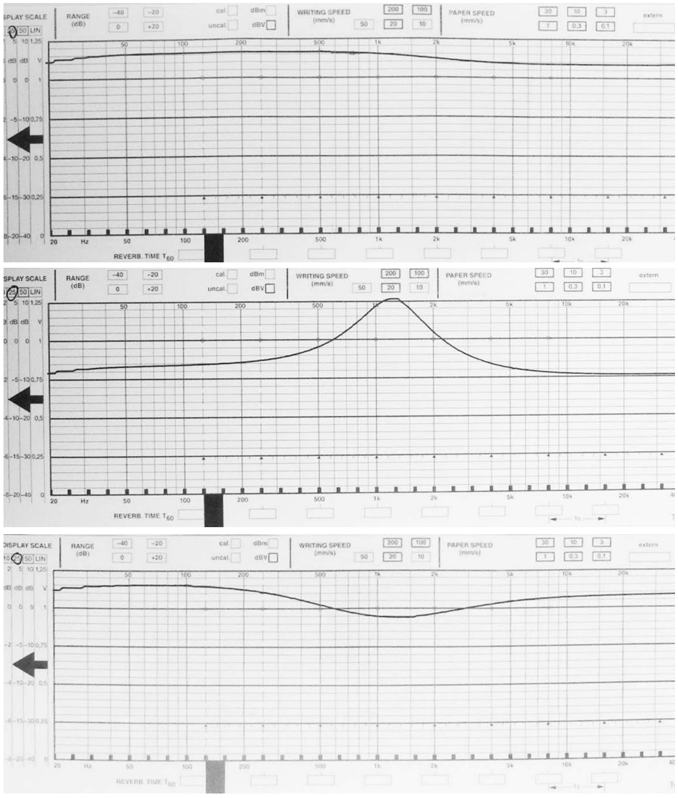


Fig.11. (top) Summation of high and low-pass outputs of a SVF. For a second-order design this only occurs at the –3dB crossover dip. Also shown are (middle) the maximum damping/lowest Q and (bottom) +3dB peaking response points.

The Q (a measure of the amount of resonance) of the SVF is more stable and controllable compared to the Sallen-Key. This makes it an ideal basis for parametric equalisers, an expensive tone control used in studios which allows smooth control of signal parameters: boost/cut frequency, Q and level. This allows one to exactly equalise any frequency response anomalies. Once you have used a parametric you will never want to go back to the fixed-frequency bands and Q of a graphic equaliser. Douglas Self's S*mall Signal Circuit Design* gives an excellent SVF-based parametric equaliser circuit.

Another advantage of the SVF approach, is that low-pass and high-pass outputs can be summed together electronically to give useful effects. Equal mixing of both outputs together will give a notch response (Fig.9). Unequal mixing, say with a ratio of 90% low-pass with 10% high-pass can give a roll-off with a notch, called an elliptical function, as illustrated in Fig.10. This is useful for active crossovers, where the notch can be used to suppress a speaker resonance. If one output is inverted before mixing, a flat response, albeit with a phase shift or all-pass response, is obtained – see Fig.11.This is very useful because the high-pass and low-pass outputs can be processed separately before mixing back together. Effectively, this means the bass and treble components of an audio signal can be given individual treatments before being recombined. This is the basis of multi-band processing and is useful in a 'de-esser', a unit where the high frequency sibilant sounds of speech are limited separately from the rest of the audio. Another application is exciters, such as the BBE Sonic Maximiser, which modulates the higher frequencies and boosts the bass. One more esoteric device is the Optimod compressor for radio stations. This splits the signal into five bands, compresses them then mixes them back together to give that characteristically 'impressively loud' American radio sound.

Fig.12 shows a second-order SVF, which is the simplest possible version. A PCB design is shown in Fig.13. The track layout and overlay are given in Fig.14. Capacitors are used to block DC through the potentiometers, preventing scratching if bipolar op amps, such as the NE5532, are used. Note the 330kΩ resistors (R17, R18) to maintain a DC path around the circuit. The spare op amp (IC2b) is available to use as an inverter or summing stage for mixing the outputs.

Electrical control

The potentiometer sections can be replaced by control elements, such as operational transconductance amplifiers (OTAs), light-dependant resistors (LDRs) or JFETs to provide electrical control. Blackmer gain cells, such as the 2180 voltage controlled amplifier (VCA) chips (which have much lower

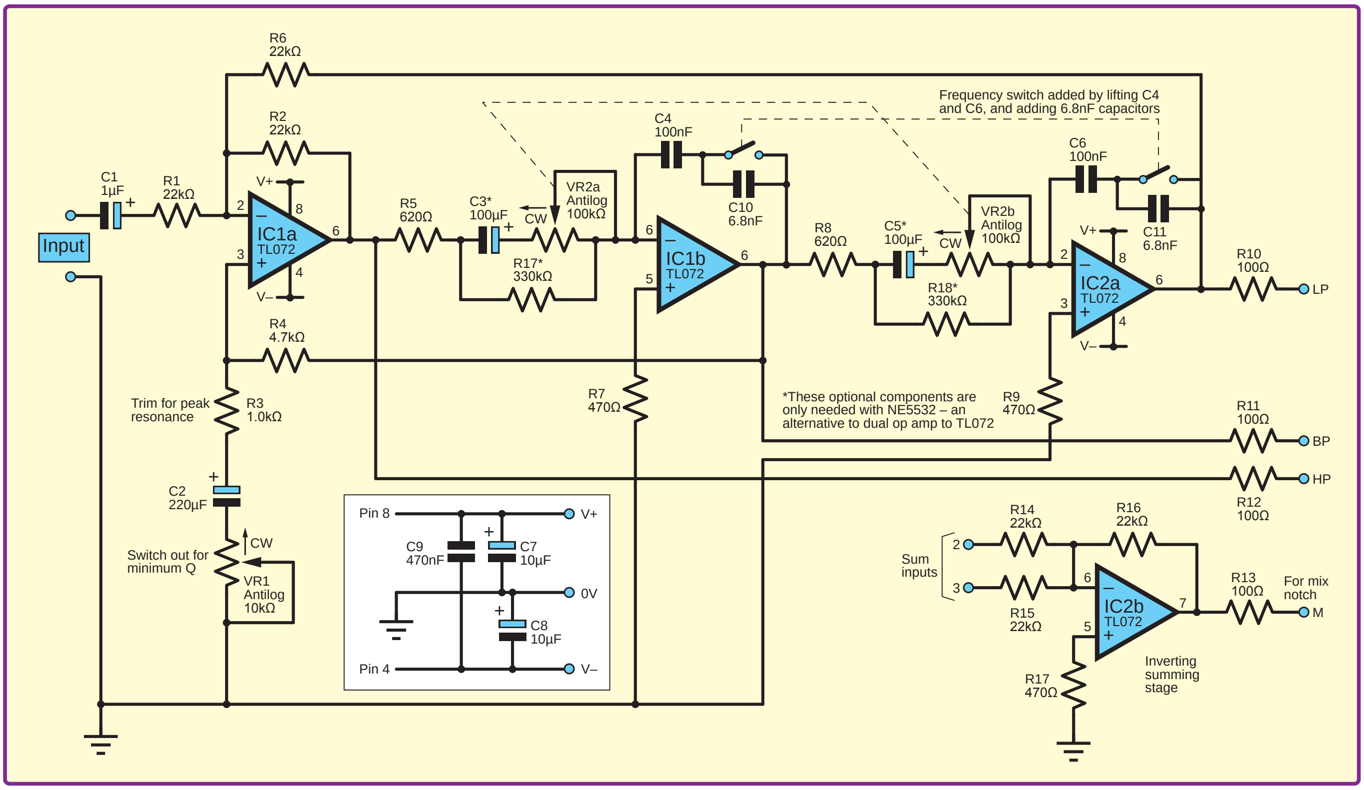


Fig.12. Second-order state-variable filter (SVF) circuit.

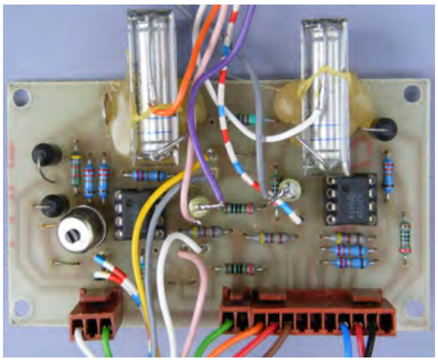


Fig.13. SVF built for design development use. Note the big polystyrene capacitors switched in to divide the frequency by a factor of ten.

distortion than OTAs and FETs) can also be used, resulting in a precision low-distortion (0.05%) Hi-Fi filter. Another advantage of the Blackmer gain-cell is its logarithmic change in frequency with control voltage, unlike OTAs, which are linear. Potentiometers, FETs and LDRs often have bad tracking between the sections, which can cause changes in the shape of the curve as the frequency response is varied. Using VCAs and OTAs gives consistency between sections and the respective control pins can be paralleled together and controlled from a single variable DC voltage. This approach is more reliable, accurate and cheaper to do than using a multi-gang potentiometer. On the other hand, potentiometers do provide the lowest distortion (typically less than 0.001%) and for this reason some designers avoid VCAs.

Fig.15 shows a third-order version of the SVF, which was used in some Fender bass guitar combos, Pulse and Ashly active crossovers (see Fig.16).

Fig.17 shows the fourth-order SVF configuration, such as that used by Behringer and Brooke Siren Systems active crossovers. It uses a four-gang

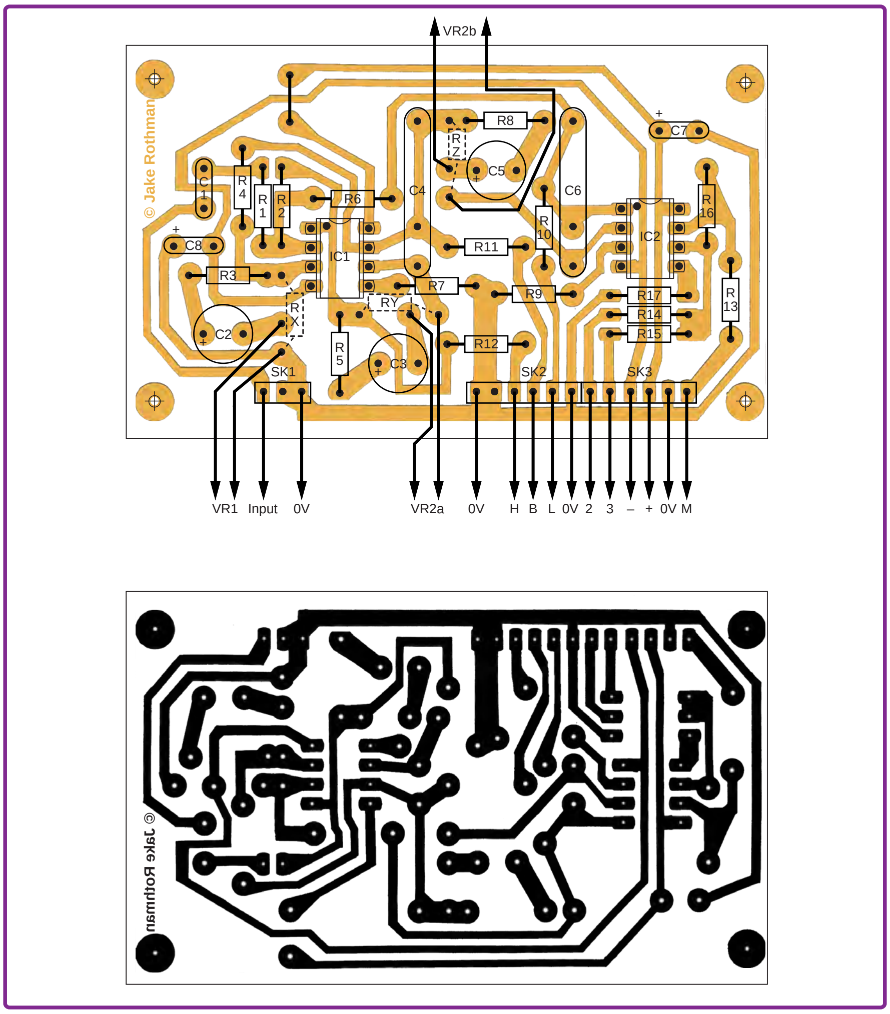


Fig.14. State-variable PCB layout/overlay.

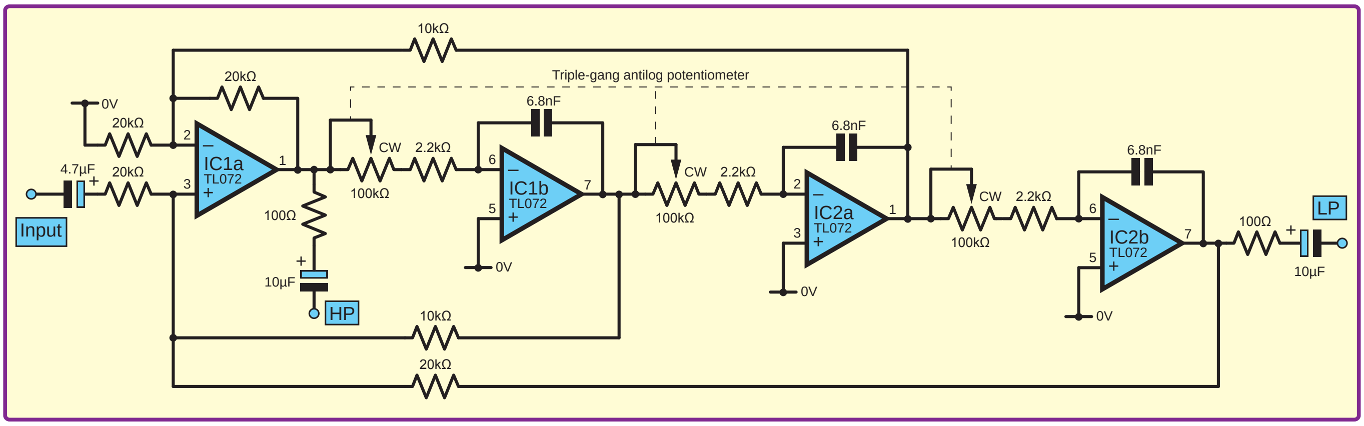


Fig.15. Third-order (18dB/octave) state-variable filter (useful in active loudspeaker designs).

anti-log potentiometer in its simplest form. In *Part 2* (next month) we'll insert some 2180 VCA chips to achieve voltage control.

Sloping off

A popular configuration of an active crossover filter is the 'Butterworth-squared' design, mathematical terminology for two Butterworth filters in series, popularised in Linkwitz-Riley speaker crossovers. This is often favoured in active crossover filters for PA use, such as the Behringer. There is a –6dB dip between the two sections at the crossover frequency, which theoretically gives a constant acoustic power radiation through the crossover region (where both woofer and tweeter are radiating together). The –6dB dip can also be achieved by decreasing the Q on a fourth-order state-variable. The problem with this approach is that it assumes a flat response and constant radiation angle from the drivers well beyond the crossover frequency, which is rarely the case.

Sometimes high-Q, second-order high-pass filters are used with small reflex speakers to give extra extension and volume in the bass. In this case, the peak of the filter is tuned to coincide with the port resonance where the power handling is greatest. Variable frequency and Q controls are a good way to optimise a given loudspeaker system. These speakers sound impressive but do not sound accurate on acoustic music because of the increased group delay (rate of change of phase with frequency) which spoils the transient response.

Anti-aliasing and recovery filters with very steep slopes are now rarely used in digital audio, because high sample rates, noise shaping and oversampling techniques render them unnecessary. They also sound strange – you can sometimes hear the effect with a smearing of the sound of high-hats on early CD players. For further filter reading, I recommend Don Lancaster's *Active-Filter Cookbook* and Douglas Self's *The Design of Active Crossovers*.

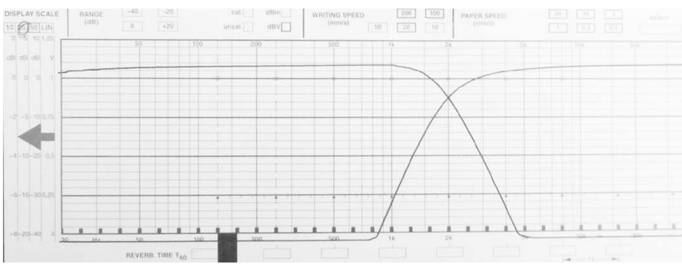


Fig.16. Ashly third-order active crossover. Note the variable Q/overlap function. This often provides a degree of EQ around the crossover point just where it is needed.

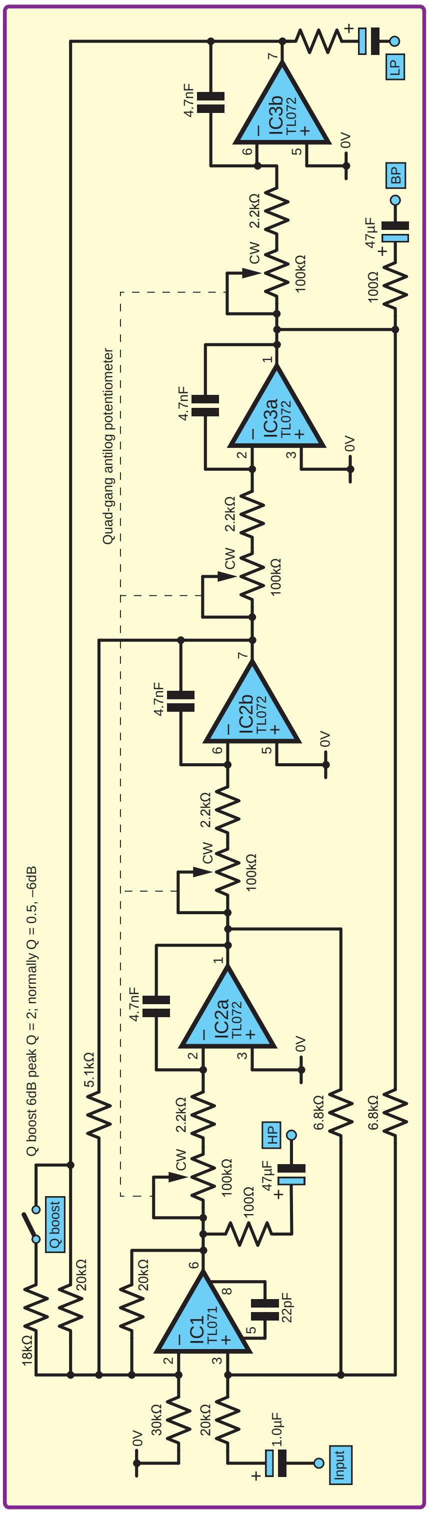


Fig.17.Fourth-order (24dB/octave) SVF, often used in crossovers for PA systems.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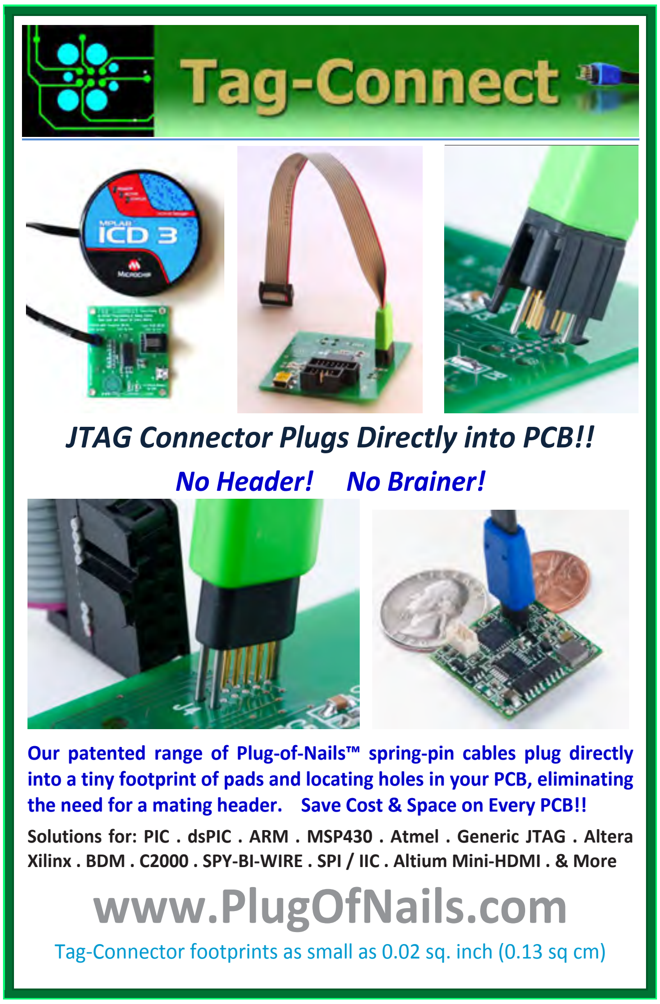
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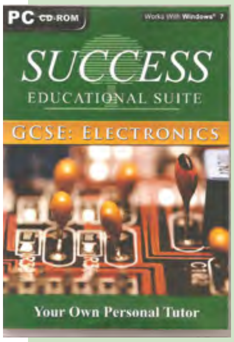

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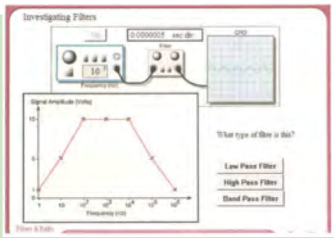

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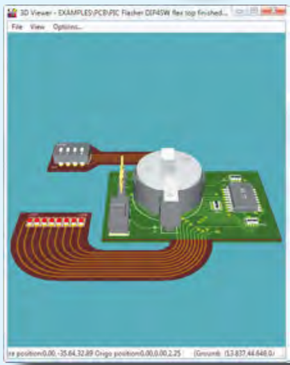

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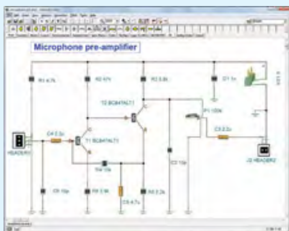

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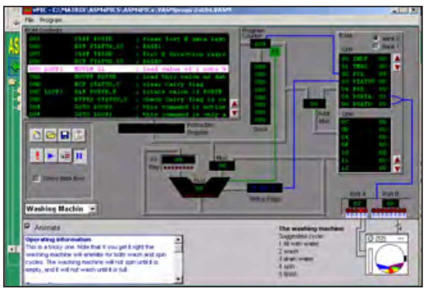

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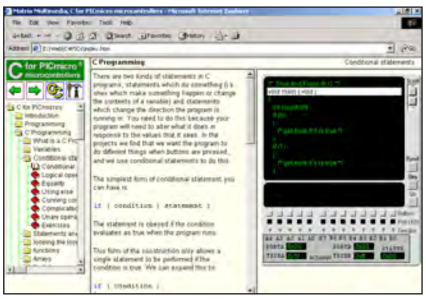

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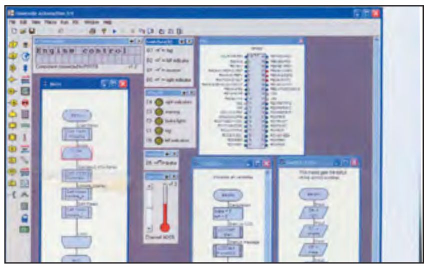

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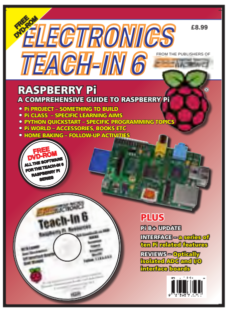

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Teach-In 2017

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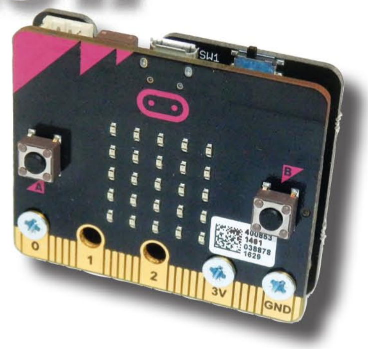

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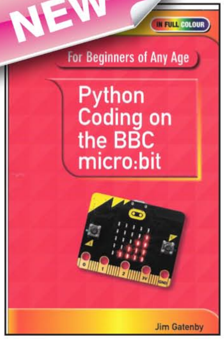

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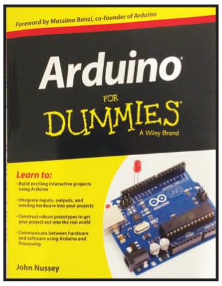
Teach-In 2016

Exploring the Arduino



ARDUINO FOR DUMMIES
John Nussey

Arduino is no ordinary circuit board. Whether you're an artist, a designer, a programmer, or a hobbyist, Arduino lets you learn about and play with electronics. You'll discover how to build a variety of circuits that can sense or control real-world objects, prototype your own product, and even create interactive artwork. This handy guide is exactly what you need to build your own Arduino project – what you make is up to you!

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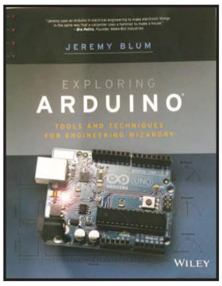

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Jeremy Blum

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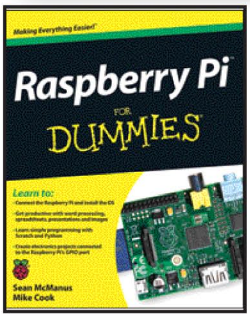

RASPBERRY Pi FOR DUMMIES
Sean McManus and Mike Cook

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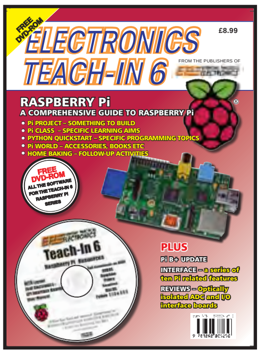

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Turntable LED Strobe	04101161	£7.60
MARCH '17		
Speech Timer for Contests & Debates	19111151	£16.42
APRIL '17		
Microwave Leakage Detector	04103161	£8.00
Arduino Multifunctional 24-bit Measuring Shield	04116011 }	£17.75
– RF Head Board	04116012 }	
Battery Pack Cell Balancer	11111151	£9.00
MAY '17		
The Micromite LCD BackPack	07102122	£11.25
Precision 230V/115V 50/60Hz Turntable Driver	04104161	£19.35
JUNE '17		
Ultrasonic Garage Parking Assistant	07102122	£10.45
Hotel Safe Alarm	03106161	£8.00
100dB Stereo LED Audio Level/VU Meter	01104161	£17.75
JULY '17		
Micromite-Based Super Clock	07102122	£10.45
Brownout Protector for Induction Motors	10107161	£12.90
AUG '17		
Micromite-Based Touch-screen Boat Computer with GPS	07102122	£10.45
Fridge/Freezer Alarm	03104161	£8.00
SEPT '17		
Compact 8-Digit Frequency Meter	04105161	£12.88

PROJECT TITLE	ORDER CODE	COST
NOV '17		
50A Battery Charger Controller	11111161	£12.88
Micropower LED Flasher (45 × 47mm)	16109161	£8.00
(36 × 13mm)	16109162	£5.60
Phono Input Converter	01111161	£8.00
DEC '17		
Precision Voltage and Current Reference – Part 2	04110161	£15.35
JAN '18		
High-Power DC Motor Speed Controller – Part 1	11112161	£12.88
Build the SC200 Amplifier Module	01108161	£12.88
FEB '18		
GPS-Syncronised Analogue Clock Driver	04202171	£12.88
High-Power DC Motor Speed Controller – Part 2 – Control Board	11112161	£12.88
– Power Board	11112162	£15.30
MARCH '18		
Stationmaster – Main Board	09103171 }	£17.75
– Controller Board	09103172 }	
Build the SC200 Amplifier Module – Power Supply	01109111	£16.45
APRIL '18		
Spring Reverberation Unit	01104171	£15.30
DDS Sig Gen Lid	Black	£8.05
DDS Sig Gen Lid	Blue	£7.05
DDS Sig Gen Lid	Clear	£8.05

Back numbers or photocopies of articles are available if required – see the Back Issues page for details. WE DO NOT SUPPLY KITS OR COMPONENTS FOR OUR PROJECTS.

*** See NOTE left regarding PCBs with eight digit codes ***

Please check price and availability in the latest issue.
A large number of older boards are listed on, and can be ordered from, our website.

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Where available, software programs for *EPE* Projects can be downloaded free from the Library on our website, accessible via our home page at: **www.epemag.com**

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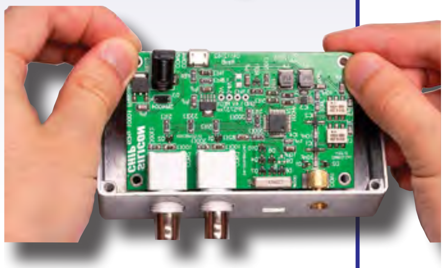
Next Month

MAY '18 ISSUE ON SALE 5 APRIL 2018

Micromite BackPack V2

The best gets better! The *Micromite LCD BackPack* described in the May 2017 issue has been one of our most popular projects. This revised version incorporates the Microbridge described below. It adds a USB interface and the ability to program/reprogram the PIC32 chip while it's onboard.

High Performance RF Prescaler

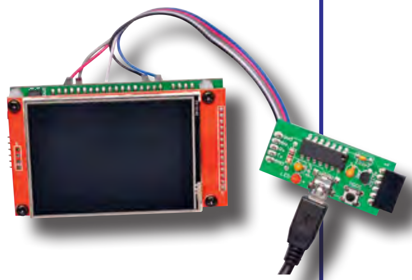


Want to measure frequencies up to 6GHz or more… but your frequency counter is not in the race? Well, if you already have a frequency counter which will measure up to 10MHz or so, you can add this prescaler to provide a dramatic increase in performance. It has selectable frequency division ratios of 1000:1, 200:1, 100:1 or 10:1 to make it especially versatile.

Microbridge

This is a cheap, but very useful universal PIC32 programmer combined with a USB/serial converter. It is primarily intended for use with the Micromite and includes the necessary USB/serial converter. You can manipulate the PIC32 from your PC, program any PIC32 microcontroller and the USB/serial converter can be used with many other processors including those on the Arduino or Raspberry Pi.

Low Cost Electronic Modules – Part 5



Fancy using a 16x2 backlit alphanumeric LCD module? This one comes with a small 'piggy-back' module that provides it with an I²C serial interface. It can be hooked up to any of the common micros via only two wires, letting multiple displays (or other I²C devices) share the same 2-wire bus, while also freeing up some of the micro's I/O pins for other purposes.

Teach-In 2018 – Part 9

Next month's *Teach-in 2018* will look at digital measurements and associated test equipment and our practical project will feature a simple logic probe.

PLUS!

All your favourite regular columns from *Audio Out* and *Circuit Surgery* to *Electronic Building Blocks, PIC n' Mix* and *Net Work.*

Content may be subject to change

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