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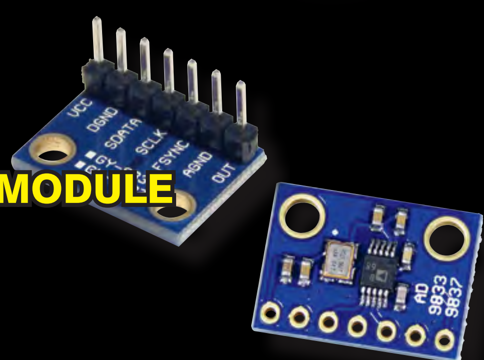

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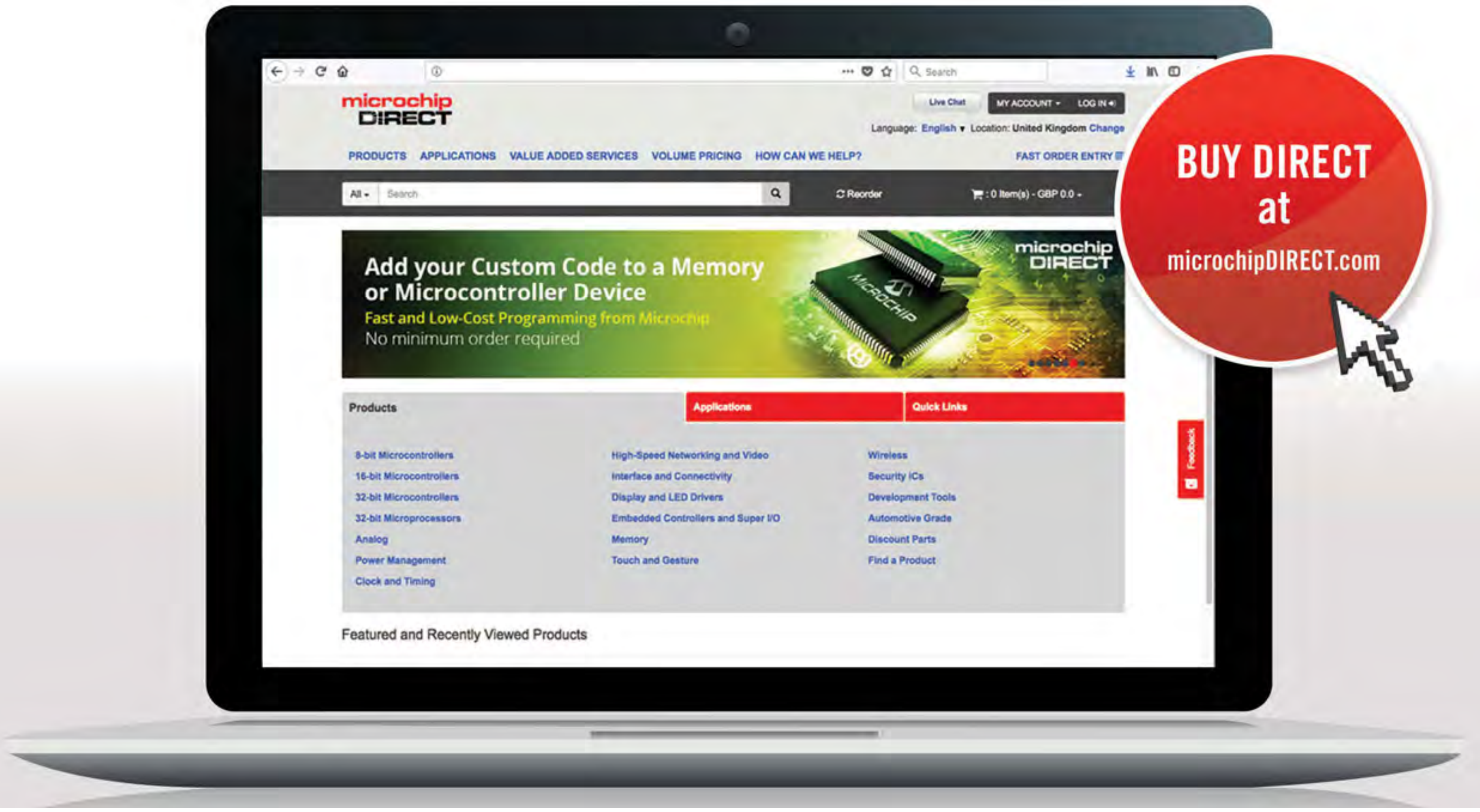

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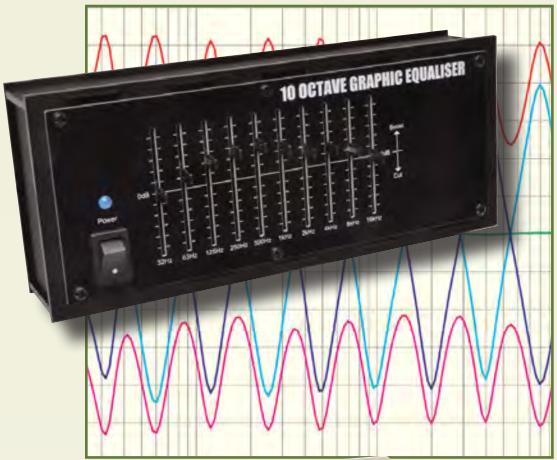

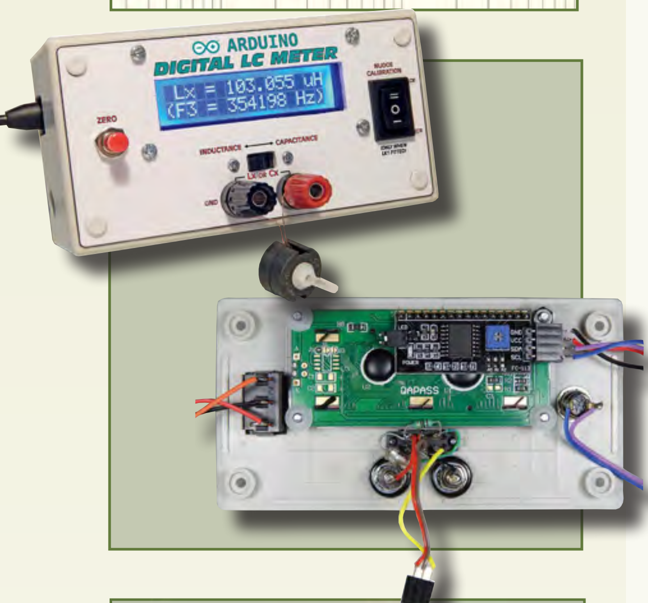

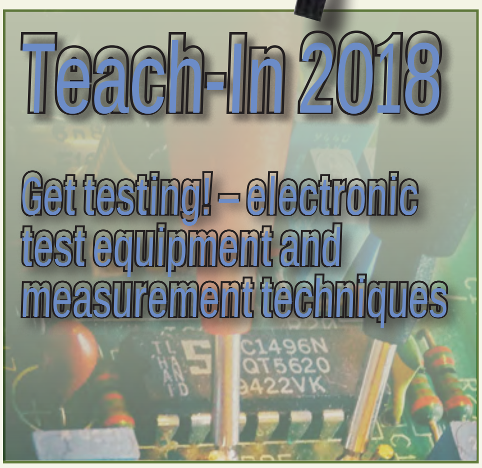

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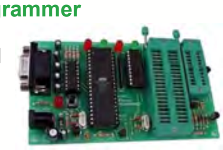

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EVERYDAY PRACTICAL ELECTRONICS

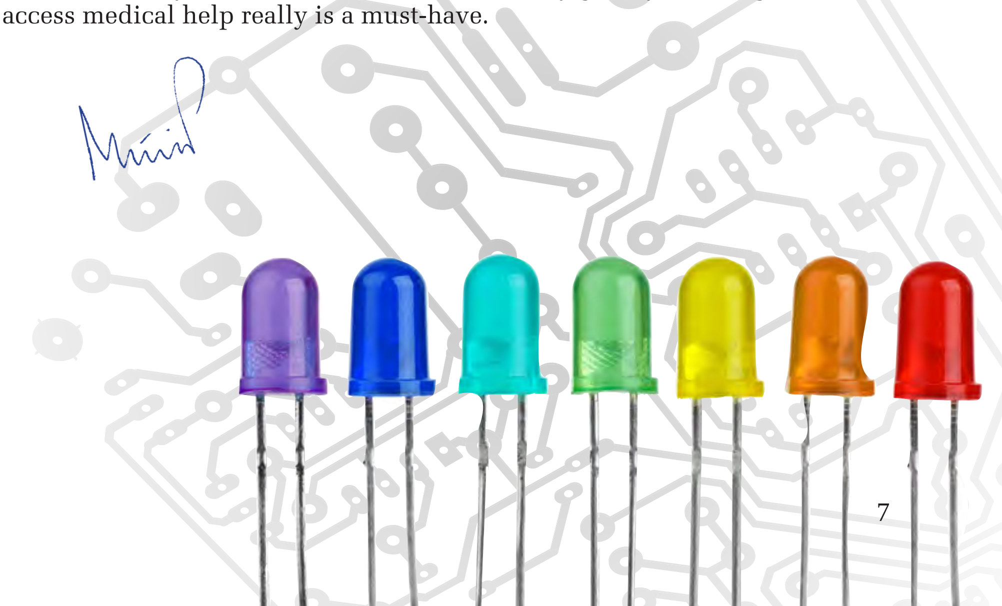
Great answer... what's the question?

Hindsight is a wonderful thing – it gives you the kind of 20/20 vision that makes everything simple and obvious. Looking back at Apple's iPhone launch on 29 June 2007 – and knowing what we know now – the iPhone's success was surely assured. But it was not obvious back then. The iPhone was expensive, ran on low-speed networks and lacked the super-high-resolution screens we now rely on for all those cats videos on YouTube and Facebook.

Nevertheless, Apple's gamble paid off, and a company that previously hadn't sold a single phone became an unassailable market leader, crushing long-established rivals like Nokia, Blackberry and Motorola. The iPhone became a classic example of disruptive technology that transformed the market. Apple understood that people wanted (or even needed) convenient access to the Internet in their pocket, and that is exactly what they delivered. Early limitations were quickly ironed out and their market dominance continues to this day. True, other companies, particularly Samsung, are now providing Apple with serious competition, but from a design perspective it's clear that even if a smartphone isn't actually an Apple product it is almost always unquestionably a direct descendent of Apple's revolutionary 2007 model – and as we all know, imitation is the sincerest form of flattery.

Fast forward nearly eight years to 24 April 2015 and Apple released another new line – the Apple Watch. Although it quickly became the best-selling wearable device, even the most ardent Apple fan would concede that it has struggled to find the kind of market penetration and must-have status of an iPhone or similar smartphone. I have one, and it's fun, occasionally quite useful, but if it disappeared I wouldn't miss it the way I'd miss my iPhone. It just doesn't do much that most any smartphone already does.

Its biggest limitation is that it is a great solution looking for a problem. Well, we are now starting to catch a glimpse of what that 'problem' might be. A report on **diabetes.co.uk** notes that, 'a technological breakthrough means smart watches could pave the way for diagnosing diabetes in the future. A study, based on data from 14,000 users of DeepHeart, a popular Apple Watch app, has shown the wearable technology was able to identify people with diabetes with 85% accuracy.' In other words, and please excuse the unfortunate terminology, healthcare could well be the killer app for smart watches. The market is huge and the potential returns equally astronomical. After all, who would refuse to pay a fraction of the price of a smartphone for monitoring a serious health condition or being given critical advanced notice of a possible heart attack? A watch that tells you the time is nice, but one that alerts you to serious trouble and thereby gives you enough time to access medical help really is a must-have.



NEWS

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

MQA: high-resolution Internet-streamed audio – report by Barry Fox

First, what exactly is MQA? Well for starters, it's not 'Master Quality Audio', as the chair of a panel session recently organised at Metropolis Studios by Warner Music to mark the launch of some Debussy tracks in MQA thought, until publicly corrected by Bob Stuart, the brains behind the MQA system. It's 'Master Quality Authenticated'.

High-quality, low-bandwidth streaming

The main objective of MQA is to stream high-resolution audio over the Internet, from legitimate online sales sites, at a manageable and affordable data rate of slightly over 1Mbps. Also, the MQA decoder displays a blue light when it 'authenticates' the source as genuine Master Quality.

Bob Stuart describes the way MQA works as 'musical origami', because the encoder folds some of the high-resolution detail into bit space that is otherwise largely unused. An MQA decoder then recovers the hidden data and uses it to rebuild the original high-resolution signal. The system is backwards compatible, in that an 'ordinary' non-MQA decoder can still deliver a CD-quality signal.

Data loss?

There is hot debate over whether the MQA process is 'lossless', with the source input and decoded output mathematically identical – or whether it is 'lossy', with some data lost in the process. There is more debate over whether or not any 'lossiness' is or is not audible.

Search under 'MQA' and you will find much highly technical argument on all this – but little if anything that explains in simple, but factual, words how the 'origami' process actually works.

Reading the patents on MQA is of limited help because they describe a wide range of theoretical options.

And it's in the interests of the MQA licensing body not to make things too easy for competitors or for critics to pick holes in the system.

MQA basics

So I talked with some industry insiders and put together – if only for my own use and reference – an idiot's fact guide to what goes on inside the MQA coder and decoder. I then put this to MQA for comment and correction. So far, I've not been corrected, but we can always add anything forthcoming as a future footnote.

CD-quality audio, known as 'Red Book' from the standards set in the 1980s by Philips and Sony, samples analogue sound at 44.1kHz (or 48kHz) and describes each sample in 16-bit words. High-resolution audio leaves the studio as 96k or 192k/24-bit PCM – sampled at 96kHz or 192kHz, and each sample described in a 24-bit word. In its raw state this needs very high data rates, around 10Mbps, which streaming services cannot cope with and would cost users on limited data plans an arm and a leg to receive.

Online sites can already deliver near-CD quality as MP3 at its highest data rate of 320kbps, but MP3 most definitely is lossy and performance can depend on the nature of the music.

MQA origami

MQA starts with 96/24 high-resolution source material, and as a first step filters it into three frequency bands; DC to 24kHz, 24kHz to 48kHz, and 48kHz to 96kHz.

Everything in the 48-96kHz band is then 'origami' folded into digital space below the 24-48kHz band.

The 24-48kHz band, and the embedded 48-96kHz band, are then 'origami' packed into the DC-24kHz main band in the bottom 7 bits of the 24-bit words.

So the bottom 7 bits (least-significant bits, LSBs) of the 24-bit words now carry all audio data above 24kHz, and nothing else. The top 17 bits (most-significant bits, MSBs) carry 'normal' CD-quality audio.

The resultant 24-bit MQA origami stream now has a data rate of around 1.1Mbps, which is regarded as manageable for online streaming. When this stream is played with a 'normal' 16-bit player, the player ignores the bottom 8 LSBs and uses the top 16 MSBs to give CD-quality frequency and dynamic range.

When the 24-bit MQA stream is played with an MQA decoder, it

MQA: high-resolution Internet-streamed audio – *continued*

switches on its blue light to signify that the stream is 'authenticated' and uses the bottom 8 LSBs to rebuild the original 96/24 high-resolution audio.

This begs several questions, apart from the basic issue of lossy or lossless. For instance, if people with limited data plans use MQA high-resolution material for CD-quality listening, they will be paying for around four times as much for data as they would if the source were 320kbps MP3 – will they hear the difference on their playback system?

PicoLog 6 launched

Pico Technology designed and manufactured its first data logger, the ADC-10, in 1991 with a simple scope and logging software package for MS-DOS, so that users could achieve the three fundamental purposes of logging software: set up the logger, view the graphed data and save the capture to disk. These packages later became known as the PicoScope and PicoLog software.

While not much has changed with the three fundamental logging software functions in 27 years, computing and technology have moved on considerably, allowing faster setup, more advanced ways to review the capture and almost unlimited storage capability.

Designed from the ground up to be intuitive from the outset, the newly launched PicoLog 6, available for download on Windows, Linux and macOS, allows you to set up the logger and start recording with just a few clicks of the mouse, whatever your level of data logging experience. Features include:

- Real-time data collection/display
- Virtually unlimited logging capacity to PC
- Robust database format minimises data loss
- Simple and complex programmable alarms
- Up to four independent graph displays
- Channels and graphs can be scaled using lookup tables or equations
- Data can be exported as CSV, clipboard image or PDF
- Supports multiple different Pico data loggers on same PC

PicoLog 6 software is designed for use with all current Pico Technology data loggers, ranging in price from £95 to £459. Further information available at: **www.picotech.com**

Cellular batteries

Researchers from Imperial College London have fused living and non-living cells for the first time in a way that allows them to work together, paving the way for new applications.

The system encapsulates biological cells within an artificial cell. Using this, researchers can harness the natural ability of biological cells to process chemicals while protecting them from the environment.

This system could lead to applications such as cellular 'batteries' powered by photosynthesis, synthesis of drugs inside the body, and biological sensors that can withstand harsh conditions.

New PV cell technology

Few toys have captured the public's imagination quite like the Rubik's Cube. Competitions have been held to find who could solve the Rubik's Cube the fastest by hand, then engineers started building robots to solve the cube at lightning speeds.

MIT students Ben Katz and Jared Di Carlo took up the challenge. 'The gist is that there is a motor actuating each face of a Rubik's Cube,' explains Katz. Custom-built electronics are then used to control each of those motors. The robot also has a pair of webcams pointed at the cube. 'When we tell the robot to solve the cube, we use those webcams to identify the different colors on the face of the cube,' says Katz.

Di Carlo wrote software that identifies the colors of each individual part within the cube to determine the cube's initial state. Then they use existing software to instruct the robot on how to move the cube's faces.

The result? They set a new world record of 0.38 seconds. See the video at: **https://youtu.be/OZu9gjQJUQs**

Mystery machine

TechnoTalk

Mark Nelson

It's a bit of an enigma, but not of an Enigma. Even so, it does use rotors and remains a mystery. Confused? You may be even more so after reading this article!

WITHOUT OUR LOYAL readers, this magazine would be very much the poorer. In this spirit, I thank Roger Morgan for sending in a fascinating observation about an intriguing, little-known electronic device from the 1960s. I had never seen or even heard of it before Roger contacted me, although many of our readers will undoubtedly remember its 'context'.

During the 1950s and 60s, Radio Luxembourg (208m / 1439kHz medium wave) was not only the biggest commercial radio station in Europe, but also had a formative influence on generations of listeners. Pumping out more pop music than the BBC did, its only real competitors were the offshore 'pirate' radio stations – and even then, hordes of listeners still tuned in to 'Luxy'. The major record companies sponsored many of the programmes, and in between the programmes the listeners' ears were assailed by commercials for cosmetics, acne cures, teenage magazines – and the Infra-Draw method.

Earworm extraordinaire

In those days, the expression 'earworm' had not yet been invented but nobody could forget the constant repetition of 'Keynsham, spelled 'K E Y N S H A M' on one of the commercials. This was so memorable (or irritating) an address that it later became immortalised in a Bonzo Dog Doo-Dah Band song called, of course, 'Keynsham'. But what's this got to do with electronics? We'll get to that just after these words from our sponsor.

Keynsham, a town not far from Bristol, became famous or perhaps notorious as the home of a man called Horace Batchelor, who used Radio Luxembourg to promote his 'Famous Infra-Draw Method', a system that he claimed increased the chances of winning large sums on the football pools. And, as Wikipedia explains, before the National Lottery began in 1994, betting on 'the pools' was the only way to win large sums for a small stake. Listeners were invited to submit their stakes to Batchelor, who then determined how the stake was placed, using a secret formula that he called the 'Infra-Draw method'. Punters were paid only if their bet won.

Peter Baker explains on the Whirligig TV website, 'The system, which was devised to pick draws was as follows: various features of a particular game were awarded points. For example, a local derby got points, if the away team were a certain number of places above the home team, more points were awarded, and so on. If I recall correctly, there were six features. The advice given by Horace Batchelor was to use one's own skill to identify 16 games likely to be draws then the use system to identify the eight getting the most points. These eight would then be entered on the coupon. We used to run the system on all 60-odd games (very time-consuming) but to no avail.'

Madness in his method?

Here's where the electronics come in. Although Horace Batchelor never revealed the secrets of his method, most commentators consider it was based on little more than daydreams or imagination. But as Roger Morgan points out, his forecasts may have been developed using sophisticated electronics (if you're prepared to stretch a point). Have a look at: **www.bit.ly/2pE0KyA**, you'll see what purports to be the random selection machine that the great man used. I say 'purports' because someone has commented on the page: 'It seems highly unlikely to me that this is anything to do with Horace Batchelor; where is the proof? I have a copy of the 'Infra-Draw Method' [paperwork] and it gives no indication of using anything like this.'

What's in the box?

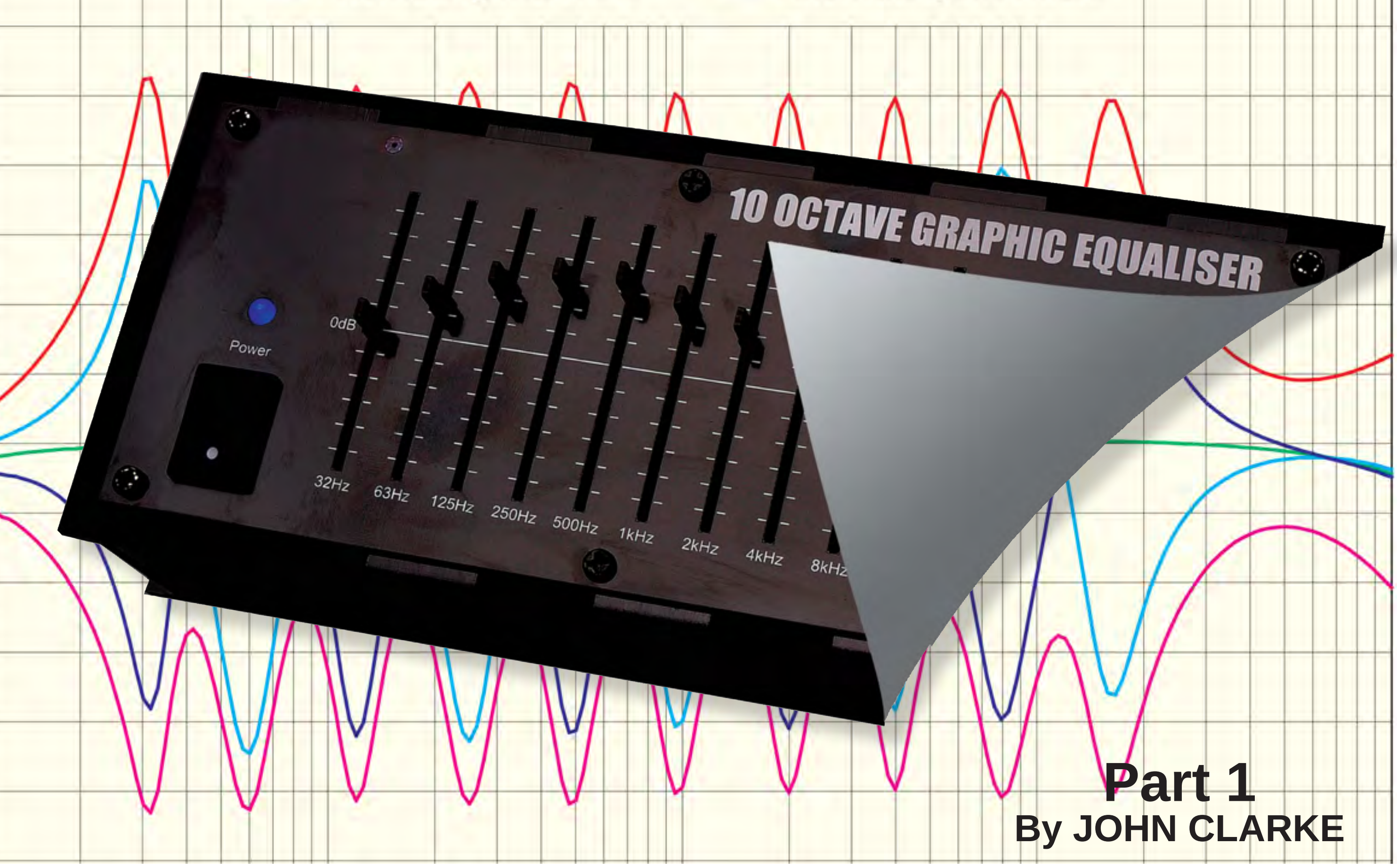
The photo caption says the device is based on electromechanical telecomms machinery, and the face includes a telephone dial, selection switches and lighted displays for 'Home Team', 'Away Team' and for the number of goals scored. That's a good start, but Roger and I found plenty to see in the photos (one has now been taken down). The whole thing is purely electromechanical, seemingly with no electronics as such. That said, the finished job is very stylish in a 1960s kind of way, with a neat wooden case and a control panel featuring some illuminated displays equipped with legends that have been professionally scripted with a Taylor Hobson pantograph engraving machine.

Many of the internal parts are mounted on tag boards, with very tidy wiring. The RS emblem on one of the capacitors dates from the 1960s, and the telephone dial on the control panel is a French one that was first introduced on the S63 type of (French) telephone in 1963. Internal fittings include a power supply, a couple of 3000-type relays, numerous tag boards, a panel connector and a mysterious central mechanism. All told, the machine is clearly custom-made for the task and the design/assembly work is extremely professional. It looks rather like something made by a technical college lab engineer.

Not an Enigma

The 'mysterious central mechanism' employs two rotors labelled H (ome) and A (way) and at least one is lettered A-Z around the rim. The parts are reminiscent of those in the well-known Enigma cipher machines used by the Germans during the Second World War, but some spirited detective work by Roger reveals these rotors originate from a SIGABA cipher machine used by the United States for message encryption from World War II until the 1950s. Says Roger, 'Perhaps this is the first proof of the urban myth that cipher machine rotors could be had in Tottenham Court Road radio dealers post war?

He also reckons the machine surely cannot contain any memory, so presumably past performance did not influence the predictions. One thing did bother Roger: why provide the HOME MATCH and AWAY MATCH indicators as illuminated displays? They weren't variable, so they might just as well have been printed labels. On the other hand, simple labels would not have looked appropriate for a time that was proclaimed to be in the vanguard of 'the white heat of the technological revolution'!



Part 1

By JOHN CLARKE

High performance 10-Octave STEREO GRAPHIC EQUALISER

This stereo graphic equaliser is very compact and quite cheap to build. However, it has the performance to match full-blown commercial models that are far more expensive. Plus, it can be used in a wide range of applications from AC or DC supplies.

This new graphic equaliser was prompted by a reader's suggestion to revise an old 3-band parametric equaliser design. However, when we looked at updating the design we were conscious that parametric equalisers can be quite confusing to use – you never quite know how to vary the controls to obtain a desired effect.

By comparison, graphic equalisers are much more intuitive – you can see which bands you are boosting or cutting and it is quite easy to repeat the settings after a particular listening or recording session. Used carefully, a graphic equaliser can make a considerable improvement to overall sound quality.

It is able to smooth out the frequency response of the reproduced sound, cure peaks, dips or lumps in a loudspeaker's response, or simply and subtly change the program's tonal quality to your liking.

This 10-octave unit uses an individual slider potentiometer for each octave, giving you far more detailed control than is possible with simple bass and treble controls. And of course, the settings of the slider potentiometers provide a visual graph of the equaliser adjustments with the centre position providing a flat response in the respective octave – ie, no cut or boost.

A slider adjusted above centre shows the level of boost and a slider below centre shows the level of cut. This is why it is called a 'graphic' equaliser.

Compact design

Our new *10-Octave Graphic Equaliser* is very compact and can be used as a stand-alone unit or incorporated into existing equipment.

So having decided to produce a new design for a graphic equaliser, we had to concentrate on the problem of reducing the cost, particularly that of the metalwork, the large and complicated PCB with all those op amps and gyrator components, and finally all those expensive slider controls. Yesteryear's approach was not going to work.

The slider control was an easy choice, even though it is a bit of a compromise. Compact ganged sliders with a 45mm travel and a centre detent are now readily available at low cost and their plastic actuators mean that multiple knobs are not needed. By using ganged sliders, we have been able to drastically reduce the cost and the size of the PCB.

So what was the compromise? The sliders we have selected are linear types with a value of 10kΩ and a centre detent. However, for the best noise and distortion performance we would have preferred a value of 50kΩ. Further, we would have also preferred sliders with a 4BM taper instead of a linear resistance characteristic. The 4BM taper has a log/antilog resistance taper; log in one direction, antilog in the other. But, if we had gone to the trouble of sourcing special 50kΩ 4BM slider pots,

Performance of prototype

Gain	Unity
Input signal with no clipping at max boost	up to 2.3V RMS
Maximum input signal with flat response	up to 9.25V RMS; 4.5V RMS with single 15V supply
Frequency reponse (flat)	+0.25,–0.75dB: 10Hz-60kHz (Fig.1)
Maximum boost	±12dB (see Fig.1)
Signal-to-noise ratio	–96dB unweighted with respect to 2V RMS
Total harmonic distortion plus noise	<0.002%, 20Hz-20kHz, 22kHz bandwidth; typically 0.0016% (see Fig.2)
Channel separation	> –60dB 20Hz-20kHz, 90dB @ 1kHz (see Fig.3)
Input impedance	100kΩ || 100pF
Output impedance	470Ω
Supply current	55mA typical; 110mA maximum

the final design would have been very expensive to build.

Suffice to say that we have been able to get the performance up to or better than CD standard, so the compromise is quite satisfactory.

Naturally, we are using a double-sided, plated-through PCB with the 10 ganged sliders on one side and all the rest of the components on the other side (pretty closely packed).

However, it is not a hard board to assemble. First, most of the resistors and some of the capacitors (all with a value of 100nF, used as supply bypass capacitors) are reasonably sized (easy-to-solder!) surface-mount components. The rest of the components are easy-to-solder through-hole types.

Furthermore, all the SMD resistors are clearly labelled with their values; OK, you will need keen eye-sight, a magnifying glass or spectacles! The SMD capacitors all have the same 100nF capacitance so you don't need to worry about identifying those.

The rest of the capacitors are normally sized MKT polyesters. There are 13 low-noise LM833 op amps and again, to keep the PCB size in bounds, we have used surface-mount types. However, they have a pin spacing of 1.27mm, so they are quite straightforward to mount in place.

The resulting combination of 10 ganged sliders and a double-sided PCB with a mixture of surface-mount and through-hole components is a compact assembly and avoids a large, expensive PCB.

But what about the problem of the expensive metalwork and a precision machined, screen-printed front panel with all those slots?

Well, we have dispensed with metalwork altogether!

The front panel is a black screen-printed PCB with precision milled slots – it looks great. And following our recent practice with smaller projects, the case is made of black acrylic which slots together very easily. It looks neat and can be used as a free-standing unit or as part of a larger installation.

If you decide to build the *10-Octave Graphic Equaliser* into a larger piece of equipment, such as an amplifier or recording console, you probably don't need the acrylic case. You can simply mount the unit in a rectangular cut-out, with the front panel PCB over the top.

All the components are on the one PCB and there is no external wiring apart from the supply leads from the on-board connector. Even the RCA input and output sockets are directly soldered onto the PCB. What could be simpler?

Typical applications

Our new *10-Octave Graphic Equaliser* can be connected to a stereo amplifier or receiver in several ways. First, it can be connected in the 'Tape Monitor' loop that's still provided on most amplifiers and receivers. Alternatively, the equaliser can be connected between the preamplifier and power amplifier. Some home theatre/stereo receivers include pre-out/in connectors for this.

If you only have a single sound source that has line level output level (anywhere between 500mV and 2V RMS) then the equaliser input can be connected to that source output and the equaliser output connected to the amplifier input.

For sound reinforcement use, you can connect the equaliser between the sound mixer output and amplifier input. In that case, connectors other than the RCA types may be required and you may need to add a balanced input and balanced output converter on each channel. We published a suitable project to do this in May 2010.

Power supply options

There are three supply options; you can use a DC supply of around 18-20V, a 15-16VAC plugpack supply or a centre-tapped mains-powered 30VAC transformer (or equivalent supply rails in a power amplifier, mixer desk).

Performance

The overall performance is summarised in a separate panel and a number of graphs. Fig.1 has several coloured response curves. The green curve shows the frequency with all controls set to the centre position, giving a ruler flat response which is only 1dB down at 10Hz and 100kHz.

The red and mauve curves show the response with all sliders in the maximum boost setting and all in the maximum cut setting. Finally, two blue curves show the sliders alternately set for maximum boost and cut, and these show the effective octave width of each band.

Note that you would never use a graphic equaliser in these extreme settings – the sound quality would be just weird. Instead, you would normally use comparatively small boost and cut settings for the sliders.

For example, if your loudspeakers are a touch too bright in the 4kHz region, you might apply a slight amount of cut to the respective slider. You could not do this with a normal treble tone control because it would drastically impact the higher frequencies.

If you wanted to lift the bass response below 60Hz, you could apply a significant amount of boost on the 31Hz band and get a much more subtle effect than would be possible with a conventional bass control.

We stated that the overall performance was effectively CD-standard and that is backed up by the figures for signal-to-noise ratio and harmonic distortion. Fig.2 demonstrates that the

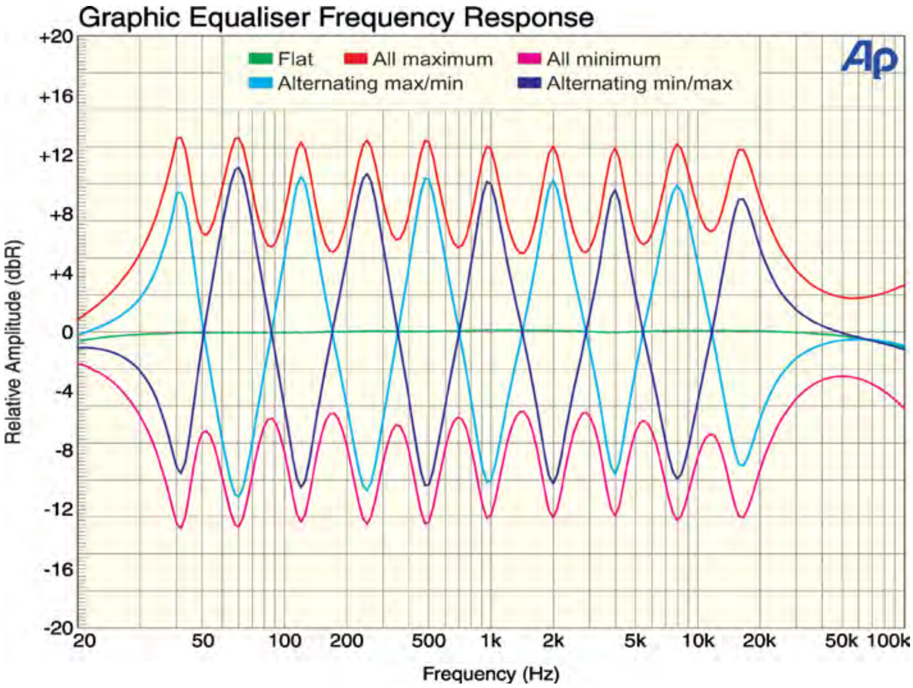


Fig.1: the green curve shows the frequency with all controls set to the centre position, giving a ruler flat response which is only 1dB down at 10Hz and 100kHz. The red and mauve curves show the response with all sliders in the maximum boost setting and all in the maximum cut setting. Finally, two blue curves show the sliders alternately set for maximum boost and cut and these show the effective octave width of each band.

harmonic distortion performance is limited by the residual noise 'floor' of the crucial gain stage in the circuit (that of IC11b and IC12b).

In fact, the actual harmonic distortion is well below our quoted figure of around 0.0016% (typical) but is masked by the residual noise. Suffice to say, the harmonic distortion of this circuit is better than can be achieved by CD and DVD players, so it will not adversely affect the sound quality of signals from such sources.

Finally, Fig.3 shows the channel separation of the graphic equaliser and the two curves show that the separation between the channels is almost perfectly symmetrical.

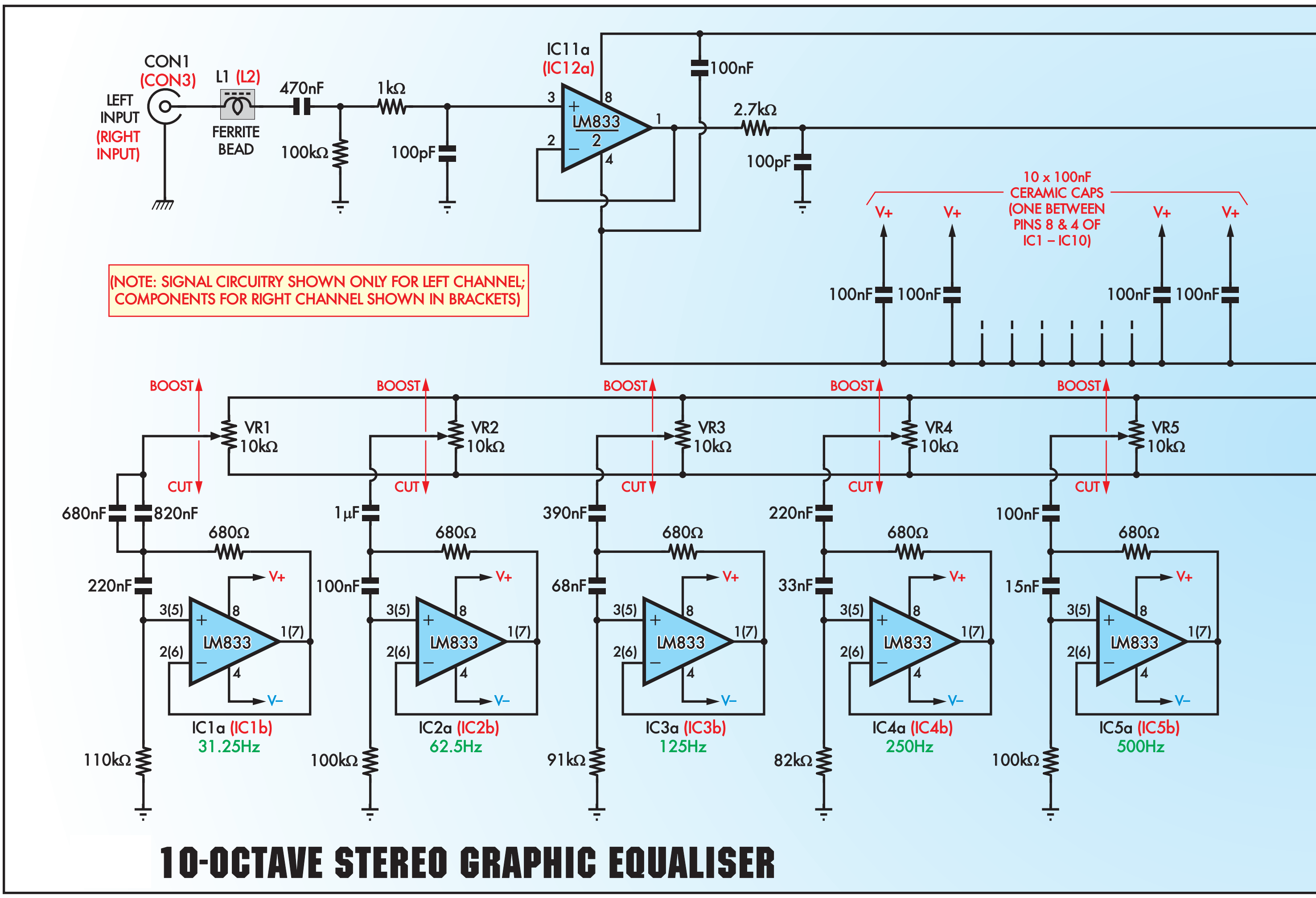
Circuit details

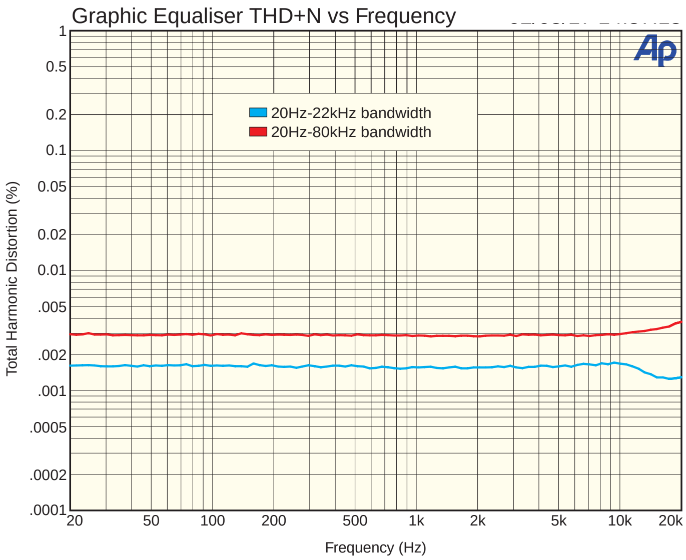
Fig.4 shows the full circuit of the left channel of the new *10-Octave Graphic Equaliser* . The right channel is identical. The IC numbering and pin numbers for the right channel are shown in brackets. We have used dual low-noise/low-distortion LM833 op amps throughout for high performance.

Before going into the detail of the circuit, let's discuss the operating principles of a typical graphic equaliser.

The overall circuit is effectively an input buffer amplifier, op amp IC11a, followed by a non-inverting op amp stage, IC11b, with the 10 slider potentiometers connected in parallel inside its feedback network. Connected to the wiper of each 10kΩ slide potentiometer is a series-resonant LC circuit; one for each octave band.

Inevitably the story is much more complicated than this because there





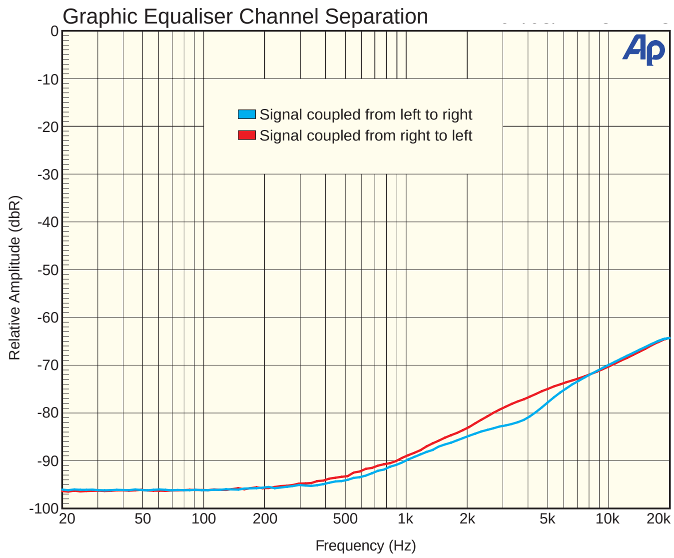


Fig.2: the harmonic distortion performance is limited by the residual noise 'floor' of the crucial gain stage in the circuit. The actual harmonic distortion is much lower.

Fig.3: the channel separation of the graphic equaliser and the two curves show that the separation between the channels is almost perfectly symmetrical.

are no inductors in the tuned LC resonant circuits. Close tolerance, low distortion inductors are very expensive and bulky, as well as being prone to hum pickup.

Therefore, all graphic equalisers designed over the last 50 years use gyrators, an op circuit that performs just like an inductor and can be connected to a capacitor to provide a series resonant circuit.

Series-resonant circuit

Let's break down the graphic equaliser circuit to show just one op amp and one 10kΩ slider and one series-resonant circuit, as shown in Fig.5. Remember that there are actually 10 resonant circuits, but in order to simplify matters, we'll only consider one.

In the simplest case, the 10kΩ slider control is set to its centre setting. In this condition, the op amp stage has unity gain and a flat frequency response and the series resonant circuit hanging off the wiper has no effect, because whatever its impedance at a particular frequency, it affects the signals at the inverting and non-inverting inputs (pins 5 and 6 here) equally.

When the slide pot is set to the boost end, the negative feedback from the output pin tends to be shunted to ground by the low impedance of the

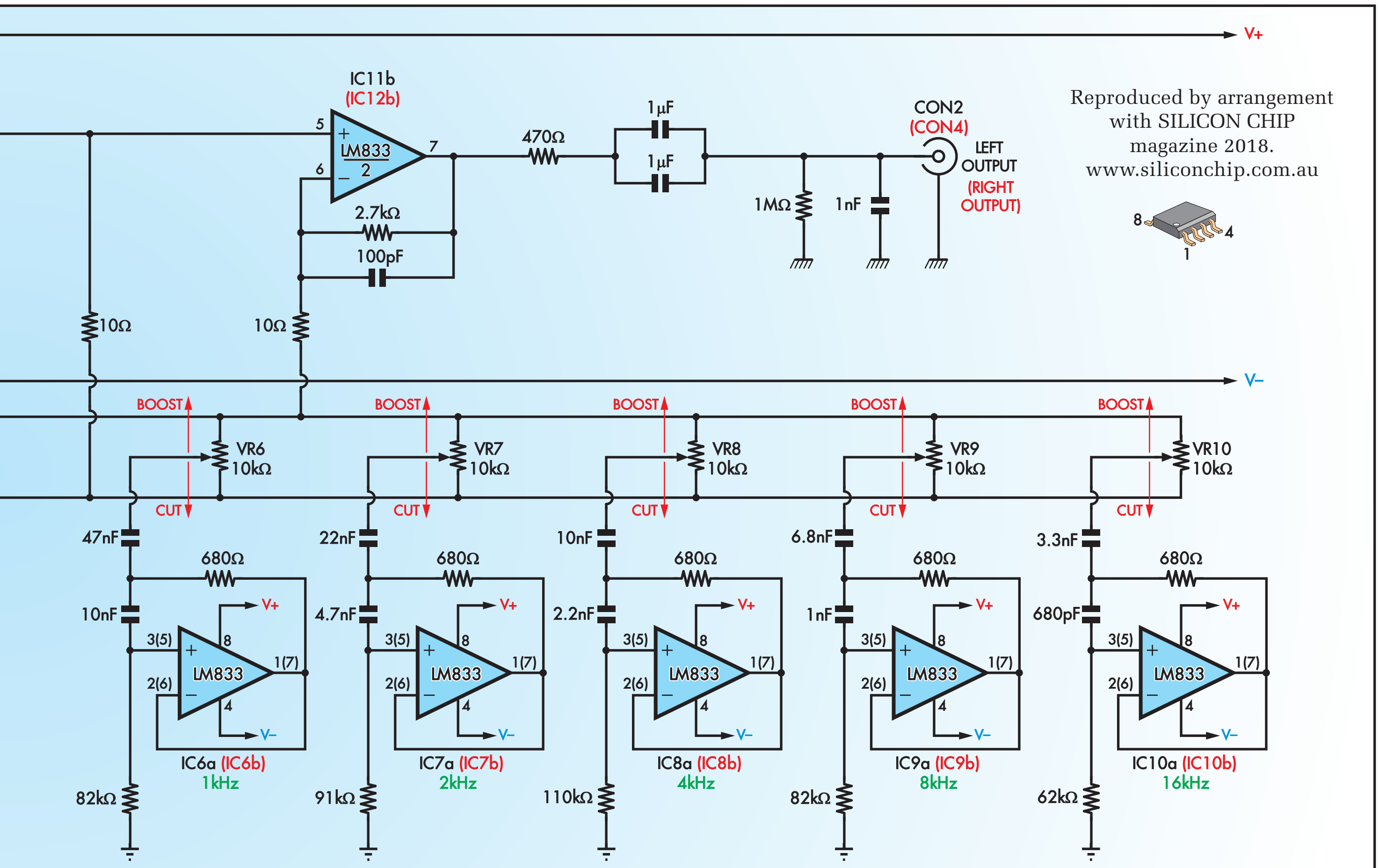


Fig.4: this circuit shows only the left channel – the right channel is identical apart from the IC numbers (shown in brackets).

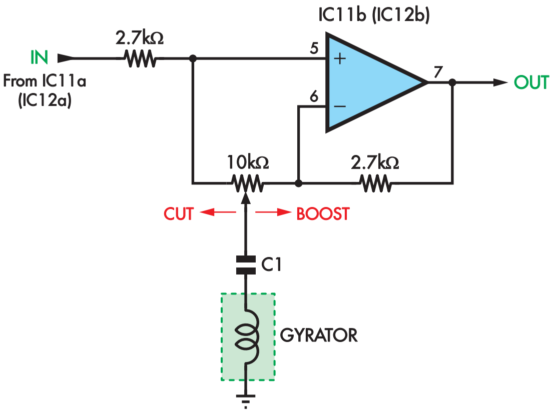
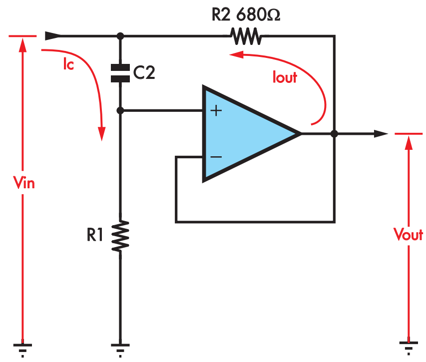


Fig.5: this is the circuit of a graphic equaliser reduced to its basic essentials – with just one op amp, one slider and one gyrator. But do remember there are 10 sliders and 10 gyrators!



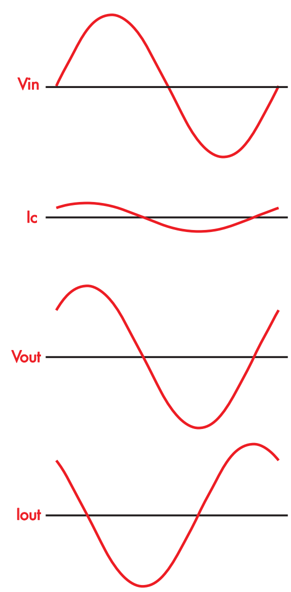


Fig.6: each gyrator in the circuit is essentially capacitor C2 and the op amp and the two together work as if they were an inductor. The accompanying waveforms at right shows how the current I_{OUT} lags V_{IN}, just like it would for a real inductor.

series-tuned circuit at frequencies that it is resonant.

Since its impedance is high at all other frequencies, this means that the feedback is only reduced over the narrow band centred around the resonance of the series tuned network. So frequencies in that band will be boosted while others will be unaffected.

When the slider is set to the other extreme, to 'cut', the negative feedback is at a maximum and the series-tuned circuit actually tends to shunt input signals in its resonant band to ground. This results in a reduction of gain for the frequencies at or near the resonance of the series tuned network. As you'd expect, the amount of boost/cut is proportional to the slider setting, so intermediate settings give an intermediate level of signal boost or cut.

Note that the circuit of Fig.5 does not show an inductor in the series resonant circuit; it shows the equivalent component, a gyrator (mentioned above).

Gyrators explained

Fig.6 shows the circuit of a gyrator made with an op amp. It effectively transforms a capacitor into an inductor. In an inductor, the current lags the voltage (ie, the current is delayed in phase by 90°) while in a capacitor, the voltage lags the current (by 90°), as it charges or discharges. Another way to explain this is that if you apply a large voltage step across a capacitor, a very high current flows initially, which tapers off as the capacitor charges up to the new voltage.

By comparison, if you apply a large voltage step to an inductor, at first the current flow remains the same as it was before, while the inductor's magnetic field charges, but over time the current flow builds as the magnetic field density increases.

To understand how the gyrator circuit behaves like an inductor, consider an AC signal source, V_{IN}, connected to the input of Fig.6. This causes a current to flow through the capacitor and through the associated resistor R1.

The voltage impressed across R1, due to capacitor current I_C, is fed to the op amp (connected as a voltage follower / buffer.

The voltage at the output of the op amp thus tracks the voltage across R1. This then causes a current to flow through resistor R2. This current, I_{OUT}, adds vectorially with the input current I_C and the resultant current which flows from the source lags the input voltage.

As far as the signal source is concerned, the gyrator 'looks' like an inductor, not like an op amp with two resistors and a capacitor connected to it. The inductance is given by the formula:

$$L = R1 \times R2 \times C2$$

where L is in henries, R is in ohms and C is in farads. If you're having trouble understanding how this works, consider again the effect of a large voltage step at the input. Say the input rises suddenly by 1V. This is initially coupled through C2 directly to the op amp and so its output also rises by 1V, keeping the voltage across R2 the same.

Thus the current flow from the input changes very little initially; it is just the current to charge C2 which is normally much smaller than that flowing through R2 (since it's normally a much lower value than R1). However, as C2 charges, the voltage across R1 drops, as does the op amp output voltage, causing the current flowing from the input, through R2, to rise to 1.5mA (1V÷680Ω) higher than it was initially.

This behaviour is very much the same as if an inductor was connected instead of the gyrator.

Building a series resonant circuit

To make the tuned LC circuit in Fig.5, all we need do is to connect a capacitor in series with the input to Fig.6. The result is a circuit with a dip in its impedance around a specific frequency.

The 'Q' of each gyrator is determined by the ratio of R1 and R2. Note from the formula above that if you double the value of R1 and halve the value of R2, the simulated inductance does not change. The same is true for the opposite, ie, halving the value of R1 and doubling the value of R2. But the 'Q' does change.

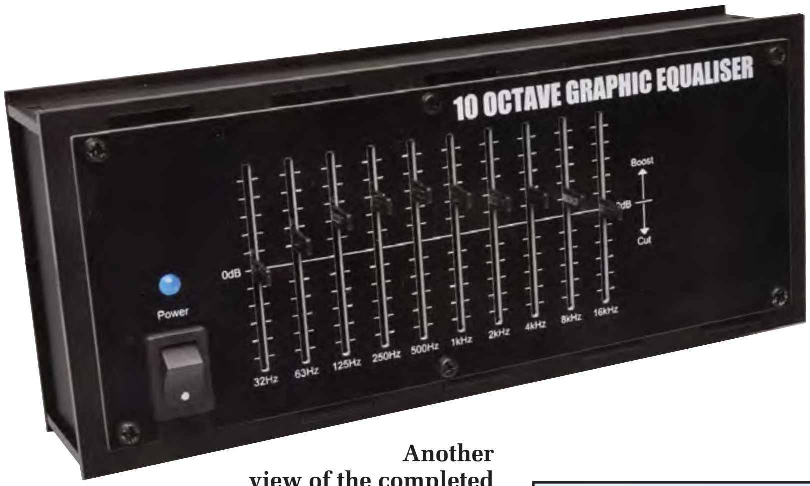
If you view the resonant circuit's impedance as an inverted bell curve, the 'Q' relates to the width of the bell. So if you increase the value of R2 and proportionally decrease the value of R1, you reduce the 'Q' and thus broaden the bandwidth of the filter.

Note that there are limits to this. You don't want to make the value of R1 too low or else the error current through it could overwhelm the current through R2 and the gyrator would no longer be a very good simulation of an inductor.

Nor do you want to make the value of R2 too low, as you'll eventually reach a point where the op amp is no longer able to drive such a low load impedance and it will hit current limiting.

And changing the value of R2 also affects the minimum impedance of the resonant circuit, which may require changes to other circuit components to avoid reducing performance.

The value of series capacitor C1 also controls the 'Q'; you can change the value of C1 without affecting the centre frequency as long as you change the value of the simulated inductor so



Another view of the completed *10-Octave Stereo Graphic Equaliser* in its laser-cut black acrylic case. No knobs are used – the actuators on the slider pots are quite sufficient.

that the product remains the same (by changing any of R1, R2 or C2).

Higher values for C1 result in lower 'Q' and vice versa. However, adjusting 'Q' with R1 and R2 is generally easier.

The values in our circuit set the bandwidth of each slider to approximately one octave. You can see the degree of overlap provided from the red and mauve curves in Fig.1.

We could have provided more overlap by increasing the values of R2 in our circuit, and reducing the R1 values (which differ for each band) proportionally, however this would also increase the interaction between adjacent bands.

Back to the equaliser

Remember that we have one op amp buffer stage IC11b, with 10 slider pots connected inside its feedback loop. The wiper of each slider is connected to one of the series-tuned circuits described above. Each is tuned to a frequency that is double that of the last, to provide octave bands.

Refer to the main circuit diagram in Fig.4. This shows just the left channel of the stereo equaliser, with one gyrator circuit repeated 10 times, with different values for R1, R2 and C2.

Looking at the top left-hand side of the circuit, the input signal is applied to CON1 and passes through a ferrite bead, which acts like an inductance to attenuate any radio signals. A 470nF capacitor blocks any DC voltage, while a 100kΩ resistor provides a charging path for the that capacitor and 'grounds' the signal. An RC filter comprising a 1kΩ resistor and 100pF capacitor provides further high frequency filtering.

Op amp IC11a buffers the input signal, giving it a low impedance, for the following equaliser circuitry comprising IC11b, the sliders (VR1-VR10), IC1-IC10 plus associated components for the gyrators.

The output signal of the graphic equaliser appears at pin 7 of IC11b, and this is fed via a 470Ω resistor and a 2µF DC blocking capacitor (using two parallel 1µF capacitors) to the output at CON2. The 1MΩ resistor to ground sets the DC level for the output signal, while the 1nF capacitor shunts any out-of-band high frequency noise to ground.

The 470Ω resistor sets the output impedance of the equaliser, while the 2µF output capacitor and 470nF input capacitor set the low frequency –3dB point of the entire circuit to about 4Hz.

Potentiometer value does not affect gain

One thing to note about the equaliser circuit, which may not be obvious, is that if you changed the poten-

Parts list – Graphic Equaliser

1 PCB available from the *EPE PCB Service*, coded 01105171, 198 × 76mm
1 front panel PCB 198 × 76mm
1 Acrylic case and hardware to suit (optional)
10 dual-ganged 45mm travel 10kΩ linear slider pots (VR1-VR10)
2 vertical PCB mount white RCA sockets (CON1,CON2)
2 vertical PCB mount red RCA sockets (CON3,CON4)
1 3-way PCB mount screw terminals with 5.08mm spacing (CON5)
2 5mm-long ferrite RF suppression beads (L1,L2)

Semiconductors
12 LM833D SOIC-8 op amps (IC1-IC12)
1 5mm high-brightness blue LED (LED1)

Capacitors
(through-hole 5.08mm pitch, all 5% tolerance except surface-mount types)
6 1µF MKT polyester
2 820nF MKT polyester (Rockby Electronics #32693)
2 680nF MKT polyester
2 470nF MKT polyester
2 390nF MKT polyester
4 220nF MKT polyester
4 100nF MKT polyester
12 100nF X7R ceramic ^
2 68nF MKT polyester
2 47nF MKT polyester
2 33nF MKT polyester
2 22nF MKT polyester
2 15nF MKT polyester
4 10nF MKT polyester
2 6.8nF MKT polyester
2 4.7nF MKT polyester
2 3.3nF MKT polyester
2 2.2nF MKT polyester
4 1nF MKT polyester
2 680pF MKT polyester
6 100pF ceramic

Resistors
(0.25W 1%; # = metal film;
^ = 120 thin-film surface mount)
2 1MΩ# 2 100kΩ# 4 2.7kΩ#
2 1kΩ# 2 470Ω# 4 110kΩ^
4 100kΩ^ 4 91kΩ^ 6 82kΩ^
2 62kΩ^ 20 680Ω^ 4 10Ω#

AC supply
2 2-way pin headers with 2.54mm spacings (LK1, LK2)
2 shorting blocks
1 W04 1.2A bridge rectifier (BR1)
1 7815 positive 15V regulator (REG1)
1 7915 negative 15V regulator (REG2)
2 470µF 25V PC electrolytic
2 10µF 16V PC electrolytic
1 47kΩ resistor^

DC supply
1 2-way pin header with 2.54mm spacing (LK3)
1 shorting block
1 LM833D SOIC-8 op amp (IC13)
1 W04 1.2A bridge rectifier (BR1); 1N4004 diodes may be substituted (see text)
1 7815 positive 15V regulator (REG1) or 7812 12V; or no regulator (see text)
1 470µF 25V PC electrolytic
1 100µF 16V PC electrolytic
1 10µF 16V PC electrolytic (not required if REG1 not used)
1 100nF X7R ceramic ^
1 22kΩ resistor^
2 10kΩ resistor^
2 100Ω resistor^

Acrylic case parts
1 Acrylic case 211 × 89 × 40mm
1 SPST rocker switch (S1)
1 panel mount 2.1 or 2.5mm DC socket to suit supply plug
1 15mm length of 5mm heatshrink tubing
1 20mm length of 10mm heatshrink tubing
4 6.3mm-long M3 tapped spacers
4 25mm-long M3 tapped spacers
4 3mm nylon washers
4 15mm-long M3 screws
6 10mm-long M3 screws
2 M3 nuts

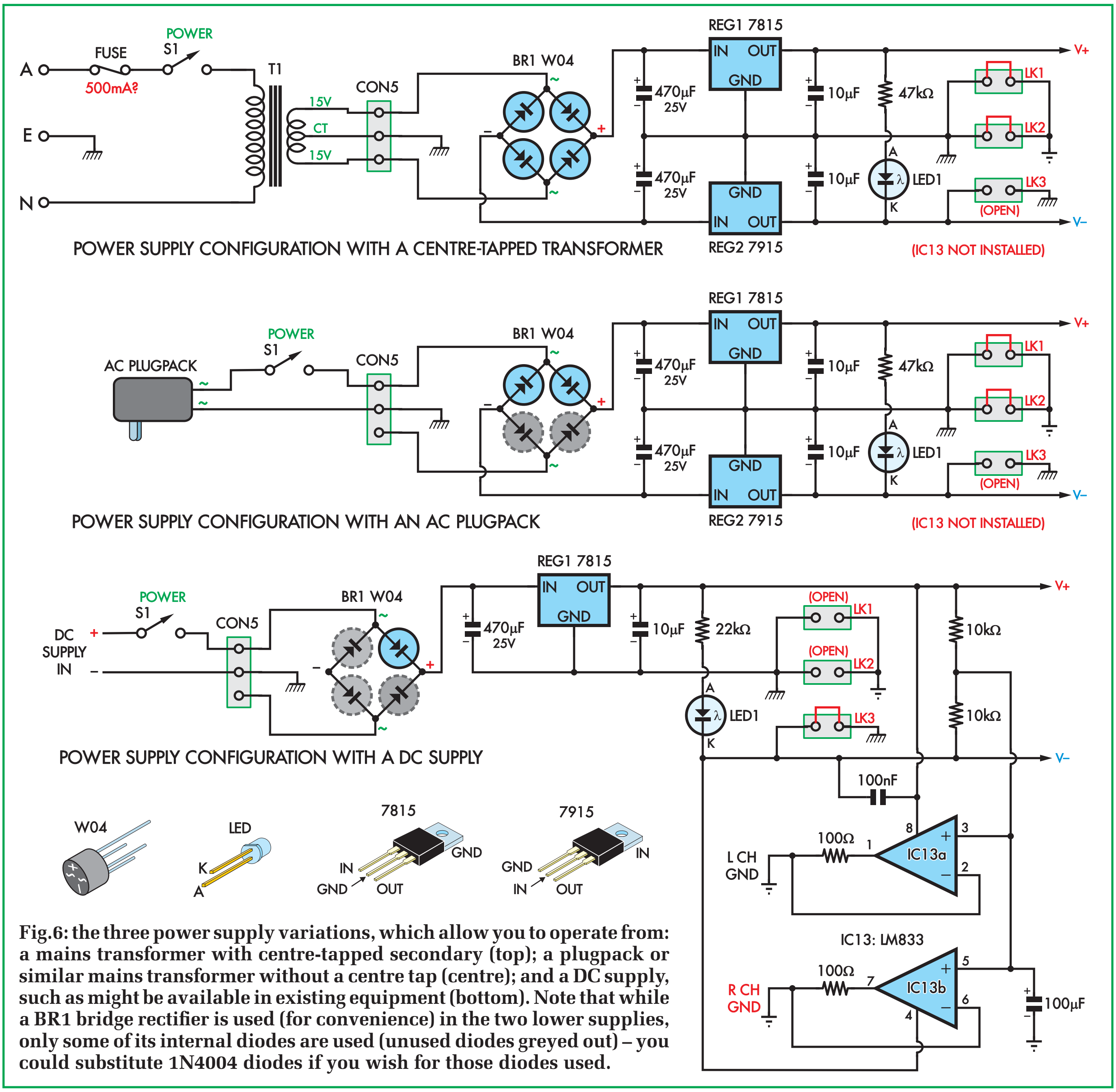
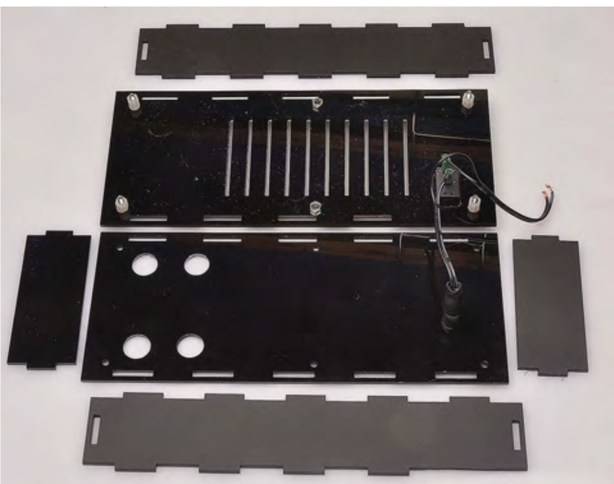


Fig.6: the three power supply variations, which allow you to operate from: a mains transformer with centre-tapped secondary (top); a plugpack or similar mains transformer without a centre tap (centre); and a DC supply, such as might be available in existing equipment (bottom). Note that while a BR1 bridge rectifier is used (for convenience) in the two lower supplies, only some of its internal diodes are used (unused diodes greyed out) – you could substitute 1N4004 diodes if you wish for those diodes used.



Here's a sneak peek at the laser-cut acrylic flatpack 'case' mentioned in the text, which significantly reduces the cost of building the *Graphic Equaliser* – and adds to the professional appearance. The pieces slot together to form a very smart-looking case in piano-finish black with white marking. We'll show how this goes together – and how the PCB fits in place – in *Part Two* next month.

tiometer resistances to another value, the output level and frequency response would not change but the noise performance might.

Imagine that all the slider pots are centred for the moment and consider each tuned circuit as having a low impedance (since white noise exists over a wide range of frequencies). This means that half of each slide pot is effectively connected between pin 5 of IC11b and ground (with a 10Ω resistance in series). The impedance of ten 5kΩ resistances in parallel is 500Ω; add the 10Ω to get 510Ω.

This 510Ω forms a divider with the 2.7kΩ resistor at the output of IC11a, providing a signal attenuation of 0.16 times (510Ω ÷ [2.7kΩ + 510Ω]). Now, IC11b has a 2.7kΩ feedback resistor and it also forms a divider with the other half of all the slide pots in parallel, again 2.7kΩ/510Ω. But because it's in the feedback loop, it provides gain, not attenuation; 6.3 times in fact.

Since 0.16 × 6.3 = 1.0, therefore, the gain from input to output of the equaliser is unity. If you change the potentiometer values to say 50kΩ, then you end up with an attenuation of 0.48 (2.5kΩ ÷ [2.7kΩ + 2.5kΩ]) and a gain of 2.08 times (2.7kΩ ÷ 2.5kΩ + 1), again giving 0.48 × 2.08 = 1.0. So the gain is still unity.

So the lower the slide pot values, the more the input signal is attenuated and the more gain is applied later to compensate.

Unfortunately, though, that gain also applies to any noise in the circuit. Thus, 10kΩ pots result in three times (6.3 ÷ 2.08) as much noise as if we were using 50kΩ pots, or a degradation in signal-to-noise ratio of around 9.5dB.

But as we said earlier, 50kΩ slide pots with a centre detent are more expensive and harder to get. As the performance with 10kΩ pots is pretty good, we feel that this is a reasonable compromise.

Power supply options

There are three power supply options and these are depicted in Fig.7.

You can use a centre-tapped 30V transformer, a 15-16VAC plugpack or a DC supply of up to 20V. There are two ground/earth connections shown on the circuit, with different symbols for each.

One is the ground for the power supply, signal inputs and signal outputs. The second is the ground reference signal for the op amp circuitry. The two are connected directly together when using a ±15V (AC-derived) supply. This is shown in the dual supply section of the circuit, where LK1 and LK2 connect the grounds together.

The power supply ground is connected to the centre tap of the transformer and is also the ground for both REG1 and REG2. These regulators provide the +15V and –15V supply rails and receive voltage from the full-wave rectifier (BR1); the raw rectified DC is filtered using 470µF capacitors. One capacitor is for the positive supply and the other for the negative supply.

Power LED (LED1) lights with voltage applied between the +15V and –15V supplies and is supplied current via a 47kΩ resistor.

You can use a 15-16VAC plugpack instead of a centre-tapped transformer. This connects to CON5 between the 0V and an AC terminal of CON5. The bridge rectifier then half-wave rectifies the input AC voltage. Two of its internal diodes are thus unused, and are shown shaded. The resulting ±15V supply rails then run the circuit.

For a DC supply, the positive voltage is applied to one of the (normally) AC inputs and the negative connection to the 0V terminal of CON5. Bridge rectifier BR1 then operates as if it were a diode, providing reverse polarity protection (the other three internal diodes are unused and thus are shaded in Fig.6).

For an input voltage above about 18V, you can use a 15V regulator for REG1, as with the AC supply options. If the input DC supply is less than this, use a 12V regulator (7812). With a supply voltage below 15V, REG1 should be left out, and its input and output terminals shorted, so that the external supply runs the circuit directly (but via BR1).

When using a DC supply, there is no negative rail available and so REG2 is left off. LK3 is fitted to connect the V– supply rail to the negative side of the external supply (ie, 0V). LK1 and LK2 are left open. As there is no negative rail, all signals to the op amps now must be biased at half supply so that there will be a symmetrical signal swing. The half supply voltage rail becomes the op amp signal grounds.

This is provided by additional op amps IC13a and IC13b. A half-supply rail is derived from two series 10kΩ resistors across V+ and V– that are bypassed with a 100µF capacitor, to remove supply ripple.

Op amps IC13a and IC13b buffer this half supply for the two channels. The signal grounds are separate to minimise crosstalk between channels. IC13 can be left off when using an AC supply.

Construction

Next month, we will go over the details for assembling the PCB and case, putting it all together and getting it up and running.

Win a Microchip Explorer 16/32 Development Kit

EVERYDAY PRACTICAL ELECTRONICS is offering its readers the chance to win a Microchip Explorer 16/32 Development Kit from Microchip. The Explorer 16/32 Development Kit is a flexible, convenient and ready-to-start development, demonstration and testing platform for 16-bit PIC24 MCUs, dsPIC DSCs and 32-bit PIC32 MCUs.

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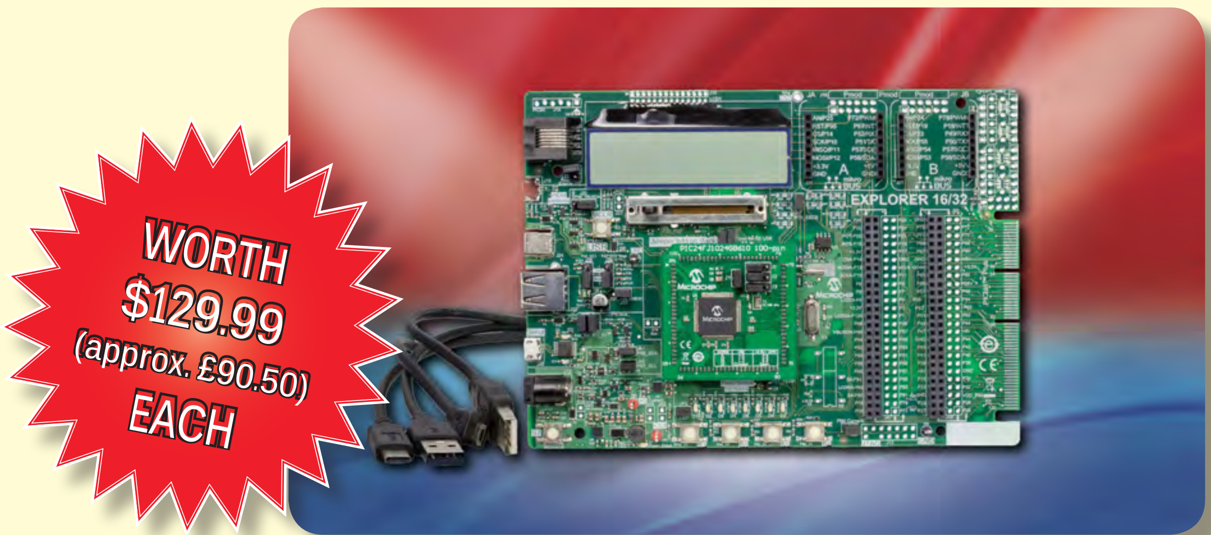
The Explorer 16/32 Development Kit (DM240001-3) includes the main Explorer 16/32 Development Board as well as a PIC24FJ1024GB610 Plug-In-Module (MA240023) and two USB cables. The Plug-In-Module features a 16-bit PIC24FJ1024GB610 microcontroller with 1MB Flash and a superset of memory and functions for the lower-power PIC24F family. This kit, with PIM and cables, is priced at $129.99.

For a fantastic, out-of-the-box experience, the Explorer 16/32 Development Kit comes with all the necessary components to get started right away with evaluation and prototyping!



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By JIM ROWE

Arduino-based Digital Inductance/Capacitance Meter

Do you ever need to check or confirm the values of inductors or capacitors? This Arduino-based LC meter will give you a digital readout and can even measure parasitic inductance or capacitance present in a circuit. It's much more accurate than most DMM-based LC meters.

Many digital multimeters (DMMs) have capacitance ranges, but they are not normally accurate for values below about 50pF. And those few DMMs that can measure inductance are often not very good at measuring inductance in the range of 1-100µH – ones that are typically used in audio and RF circuits.

An inductance meter with a 10µH resolution (typical for DMMs) isn't very helpful if you want to wind a choke of say 6.8µH, for an amplifier output filter.

Professionals tend to rely on digital LCR meters for these types of measurements. They allow you to measure almost any passive component quickly and automatically, often measuring not just their primary parameter (like inductance or capacitance) but one or more secondary parameters as well. However, many of these excellent instruments also carry a hefty price tag, keeping them well out of reach for many of us.

Fortunately, thanks to microcontroller technology, much more affordable digital instruments are becoming available. These include both commercial and DIY instruments like the low-cost unit described here.

Essentially it's an improved version of the *PIC-based Digital LC Meter* we described in the March 2010 issue of *EPE*. This time, we're basing it around an Arduino Uno or equivalent module.

Main features

Our new *Digital LC Meter* is compact and easy to build, since the Arduino board comes pre-assembled. It also has a better LCD readout than the previous version. It fits snugly inside a UB3 utility box and you should be able to build it for under £70.

It offers automatic digital measurement of both inductance (L) and capacitance (C) over a wide range and with 5-digit resolution. Measurement accuracy is better than ±1% of reading over most of the ranges.

It operates from 5V DC, drawing an average current of about 62mA, so it can run from a 5V USB supply (either mains or battery) or from a spare USB port on your PC.

How it works

This meter's impressive performance relies on an ingenious measurement technique developed almost 20 years ago by the late Neil Heckt in the US.

It uses a wide-range test oscillator and its frequency is varied by connecting the unknown inductance or capacitance you're measuring. The resulting change in frequency is measured by the microcontroller and used to calculate the component's value, which is displayed directly on a small LCD panel.

To achieve reliable oscillation over a wide frequency range, the test oscillator is based on an analogue comparator with positive feedback around it, as shown in Fig.1. This configuration has a natural inclination to oscillate, because of the very high gain between the comparator's input and output.

When power (+5V) is first applied, the comparator's positive input is held at +3.3V by the divider formed by the two 100kΩ resistors and the 100kΩ and 4.7kΩ resistors. Initially, the voltage at the negative input is zero because the 10µF capacitor at this input needs time to charge via the 47kΩ resistor.

So with its positive input much more positive than the negative input, the comparator initially switches its output high, to near +5V.

Once it does so, the 10µF capacitor connected to the negative input begins charging up via the 47kΩ resistor and the voltage at this input rises. As soon as it goes above +3.3V, the comparator output switches low and the positive input is brought to 1.67V due to the 100kΩ feedback resistor pulling the 100kΩ divider low.

The low comparator output voltage is also coupled through the 10µF input capacitor to the tuned circuit formed by inductor L1 and capacitor C1. This makes the tuned circuit 'ring' at its resonant frequency.

As a result, the comparator and the tuned circuit now function as an oscillator at that resonant frequency. In effect, the comparator functions as a negative resistance across the tuned circuit, to cancel its losses and maintain oscillation. Once this oscillation is established, a square wave of the same frequency is present at the comparator's output and it is this frequency (F_{OUT}) that is measured by the microcontroller.

In practice, before anything else is connected to the circuit, F_{OUT} will simply correspond to the resonant frequency of the tuned circuit comprising L1, C1 and any stray inductance and capacitance that may be associated with them.

When power is first applied to the circuit, the microcontroller measures this frequency (F1) and stores it in memory. It then energises reed relay RLY1, which switches capacitor C2 in parallel with C1 and thus lowers the oscillator frequency. The micro then measures and stores this new frequency (F2).

Next, the micro uses these two frequencies plus the known value of C2 to accurately calculate the values of both C1 and L1. The equations it uses to do this are shown in Fig.1. Following these calculations, the micro turns RLY1 off again to disconnect C2, allowing the oscillator frequency to return to F1. The unit is now ready to measure the unknown inductor or capacitor (L_X or C_X).

As shown in Fig.1, the unknown component is wired to the test terminals at far left. It is then connected to the oscillator's tuned circuit via switch S1.

When measuring an unknown capacitor, S1 is switched to the 'C' position so that the capacitor is connected in parallel with C1. Alternatively, for an unknown inductor, S1 is switched to the 'L' position so that the inductor is connected in series with L1.

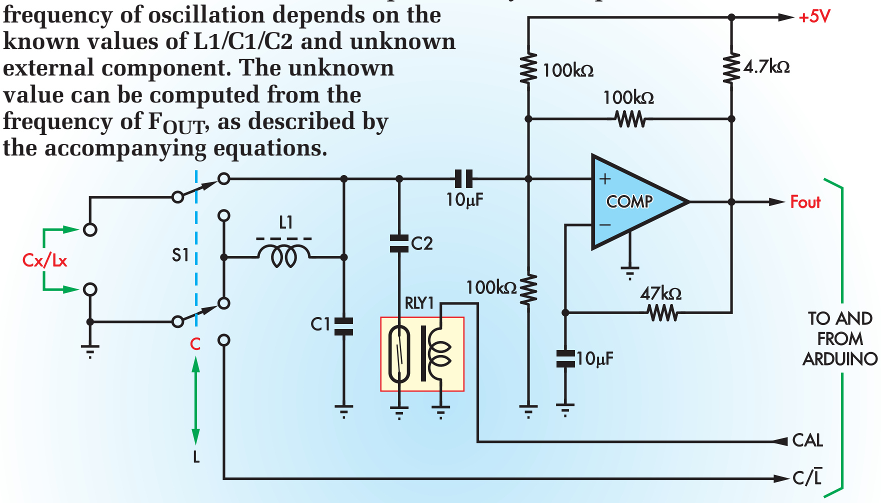
In both cases, the added L_X or C_X again causes the oscillator frequency to change to a new frequency (F3). As with F2, this will always be lower than F1.

So by measuring F3 as before, and monitoring the position of S1 (which is done via the C/$\overline{L}$ line), the micro can calculate the value of L_X or C_X using one of the equations shown in the right section of the equations box in Fig.1.

From these equations, you can see that the micro has some fairly solid number-crunching to do, both in the calibration mode when it works out the values of L1 and C1, and in the measurement mode when it must work out the value of L_X or C_X.

Each values needs to be calculated to a high degree of resolution and accuracy using floating-point maths.

Fig.1: operating principle of the *Digital LC Meter*. L1 and C1 form a tuned circuit in combination with an external capacitance/inductance connected via S1. Feedback for oscillation is provided by a comparator and the frequency of oscillation depends on the known values of L1/C1/C2 and unknown external component. The unknown value can be computed from the frequency of F_{OUT}, as described by the accompanying equations.



HOW IT WORKS: THE EQUATIONS

(A) In calibration mode

(1) With just L1 and C1: $F1 = \frac{1}{2.\pi.\sqrt{L1.C1}}$

(2) With C2 added to C1: $F2 = \frac{1}{2.\pi.\sqrt{L1.(C1+C2)}}$

(3) From (1) and (2), we can find C1:

$$C1 = C2 \cdot \frac{F2^2}{(F1^2 - F2^2)}$$

(4) Also from (1) and (2), we can find L1:

$$L1 = \frac{1}{4.\pi^2.F1^2.C1}$$

(B) In measurement mode

(5) When Cx is connected: $F3 = \frac{1}{2.\pi.\sqrt{L1.(C1+Cx)}}$

so $Cx = C1 \cdot \left(\frac{F1^2}{F3^2} - 1\right)$

(6) Or when Lx is connected:

$$F3 = \frac{1}{2.\pi.\sqrt{(L1+Lx).C1}}$$

so $Lx = L1 \cdot \left(\frac{F1^2}{F3^2} - 1\right)$

NOTE: F2 & F3 should always be lower than F1

Features and specifications

Inductance range: .... 10nH to 100mH+
Capacitance range: .... 0.1pF to 2.7µF+ (non-polarised only)
Measurement resolution: .... five digits in either mode
Range selection: .... automatic
Sampling rate: .... approximately one measurement per second
Accuracy (when calibrated): .... ±1% of reading, ±0.1pF or ±10nH
Supply voltage: .... 5V DC @ <65mA (including backlit LCD)
Supply type: .... USB charger or the USB port on a PC

Circuit details

The full circuit diagram is shown in Fig.2. It mainly consists of the Arduino microcontroller module and the serial I^2C LCD module, together with the oscillator circuit we've already introduced, built using an LM311 high-speed comparator (IC1).

The Arduino controls RLY1 to switch calibrating capacitor C2 (1nF) in and out of circuit, via its IO3 pin. Diode D1 is connected across the relay coil to prevent the Arduino's internal circuitry from being damaged by inductive spikes.

The Arduino senses the position of L-C switch S1 using its IO2 pin, which is pulled high internally when it's not pulled low by S1b (in the L position).

The output of the oscillator at pin 7 of IC1 is taken to pin IO5 of the Arduino via a series 6.8kΩ resistor. It needs to be taken to this pin because this is also the external input pin for the 16-bit timer/counter inside the ATmega328P micro, which forms the heart of the Arduino Uno.

As a result, we are able to use the Arduino to easily measure the oscillator's frequency.

The results of the Arduino's measurements and calculations are displayed on a blue back-lit 16x2 alphanumeric LCD module.

This has a serial I^2C module fitted, so it can be controlled from the Arduino via its I^2C port lines (SCL and SDA).

Its features are fully described on page **31** of this issue.

Calibration functions

The firmware sketch running in the Arduino is designed to perform its 'zero calibration' adjustment just after initial startup.

However, pushbutton switch S2 is also provided to allow zero calibration to be performed at any other desired time as well (to allow for temperature drift, for example).

S2 pulls the Arduino's $\overline{\text{RESET}}$ pin (pin 4) down, so that it is forced to reset and start up again, readjusting its zero setting in the process.

LK1 and switch S3 can be used to nudge or tweak the calibration in small increments or decrements, if you have access to an accurate reference capacitor. When LK1 is fitted, pulling input pin IO7 low, the micro will increase the capacitance reading by about 0.5% each time S3 (a centre-off rocker switch) is pushed to the upper 'INCR' position, or alternatively decrease the reading by the same amount if S3 is pushed to the lower 'DECR' position. So the idea is to push S3 in one direction or the other until the reading is correct.

Each time a change is made, the adjustment factor is stored in the Arduino's EEPROM memory, so it's remembered for future sessions. When link LK1 is not fitted, pressing S3 in either direction has no effect at all.

This is a safety feature, to prevent unintended changes to the meter's calibration during normal use. Although this calibration is normally done using a reference capacitor, it also improves the accuracy of inductance measurements.

Construction

There is no custom PCB used for the LC meter's circuitry; instead, most of the added circuitry is fitted on a prototype shield board which simply plugs into the top of the Arduino PCB.

There aren't that many components involved, so it's a straightforward job to wire it up, as shown in the wiring diagram, Fig.3.

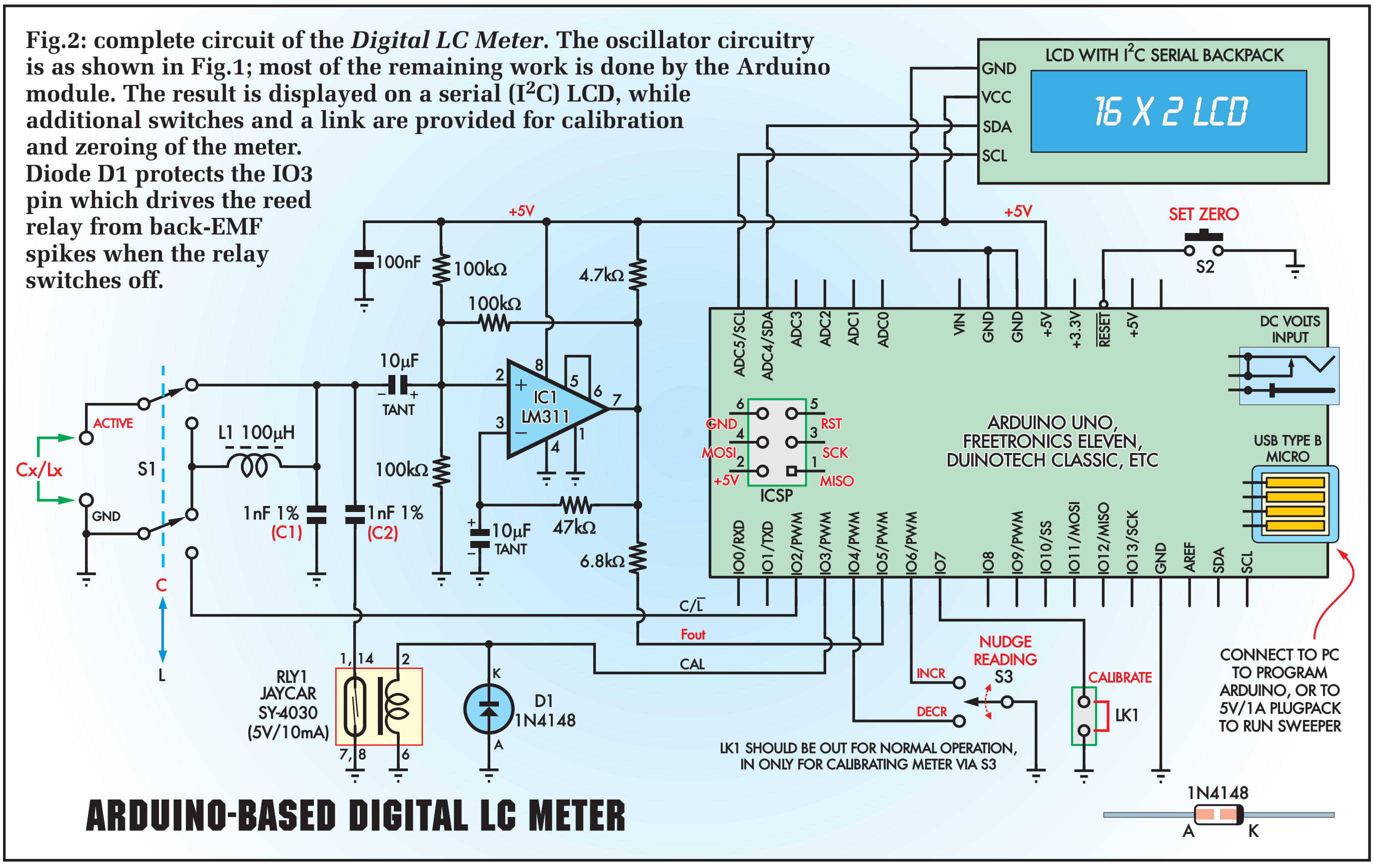
The only components which are not mounted on the ProtoShield are the serial LCD module, switches S1-S3, the test terminal binding posts and reference components L1 and C1.

As shown in Fig.3 and the photos, these are all mounted on the lid of the UB3 box, which forms the *Digital LC Meter's* front panel. These off-board components are all linked to the ProtoShield board via short multi-wire interconnection leads and SIL connector plugs and sockets, which are also shown in Fig.3.

You can get an idea of how everything fits together from the internal cutaway diagram of Fig.4, along with the internal photos.

The Arduino module mounts in the bottom of the box via four 9mm-long

Fig.2: complete circuit of the *Digital LC Meter*. The oscillator circuitry is as shown in Fig.1; most of the remaining work is done by the Arduino module. The result is displayed on a serial (I^2C) LCD, while additional switches and a link are provided for calibration and zeroing of the meter. Diode D1 protects the IO3 pin which drives the reed relay from back-EMF spikes when the relay switches off.



M2.5 machine screws and four M2.5 nuts, with another four M3 or M2.5 nylon nuts used as spacers.

The ProtoShield is plugged into the top of it. The rest of the *Digital LC Meter* circuitry connects via the 90° pin headers on the ProtoShield.

Follow the wiring diagram (Fig.3) and internal photos to build the ProtoShield. Start by soldering the components into place where shown in Fig.3, ensuring you use the correct orientation for polarised components: IC1, diode D1, RLY1 and the two 10µF tantalum capacitors.

Next, add the wiring on the underside, as shown in the underside wiring diagram of Fig.3. Use insulated wire because several of these wires cross over each other.

In cases where adjacent pads are connected, you can simply place a solder bridge between the two pads or alternatively, bend the component leads while fitting them and trim them so that they reach the adjacent pads. For longer connections, use component lead off-cuts, routed carefully to avoid the possibility of shorting anything else, or short lengths of light-duty hookup wire (eg, stripped from a piece of ribbon cable) or bell wire.

Here's our suggested order of fitting the components and wiring the ProtoShield board; check Fig.3 for the exact placement in each case:

1. Fit the four 90° SIL headers.
2. Fit a four-pin vertical header for switch S2.
3. Fit the four SIL pin headers to the underside, along the upper and lower edges of the ProtoShield, which connect it to the Arduino. These comprise a 10-pin header at upper left, two 8-pin headers (one at upper right and the other at lower centre) and a 6-pin header at lower right. Do not fit a 3x2 DIL pin header in the ICSP position at centre right on the ProtoShield board.
4. Fit IC1's 8-pin DIL socket with the notched end to left, then relay RLY1, with its notched end towards the top.
5. Mount the six resistors, the 100nF capacitor and the two 10µF tantalum caps. Note that the last two are polarised, so make sure you fit them with the orientation shown.
6. Fit the 12 insulated wires on the top of the board and any insulated wires required to complete the wiring on the underside. This will require you to strip the insulation from each end by about 5mm or so.
7. Fit diode D1, making sure its end with the cathode band is uppermost and adjacent to pin 2 of RLY1, then plug IC1 into its socket.

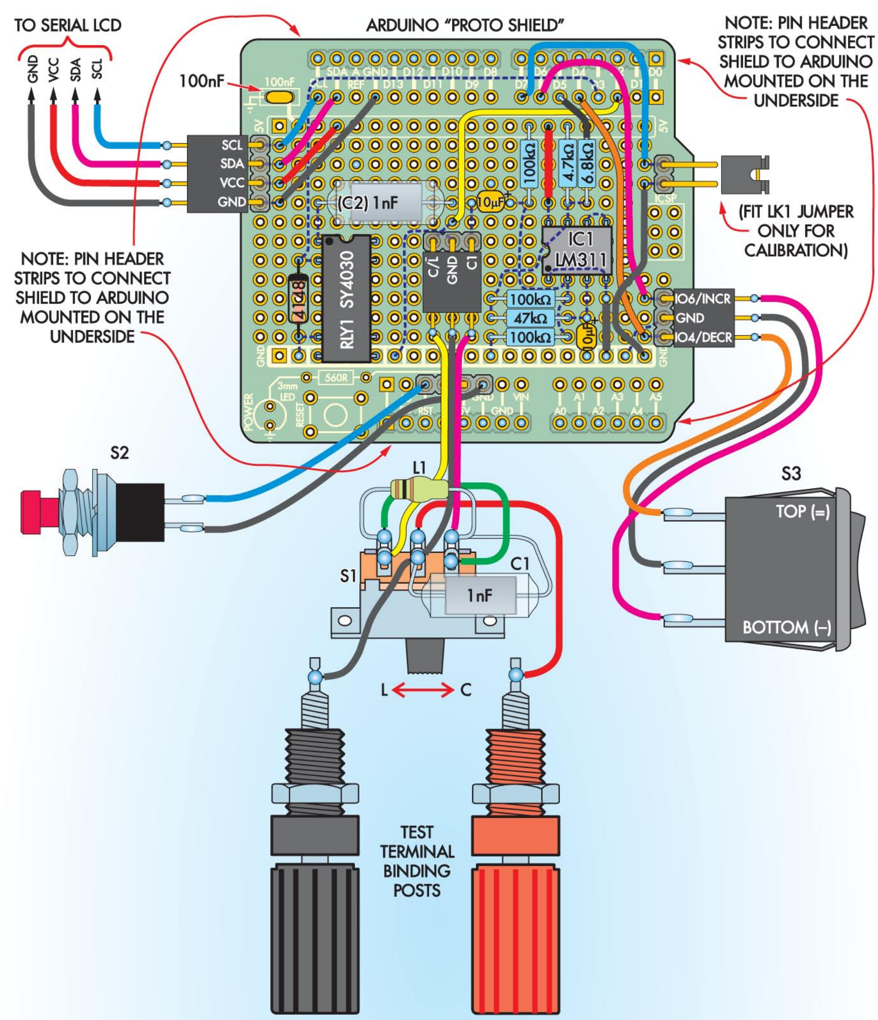
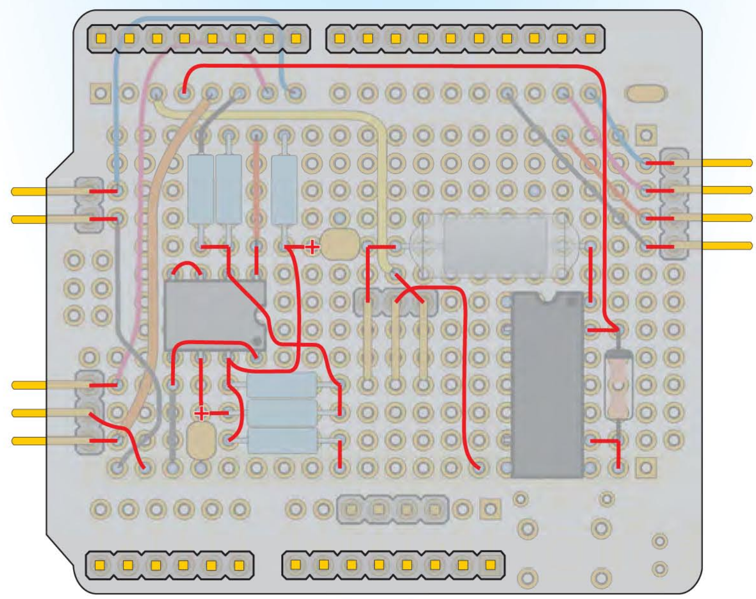
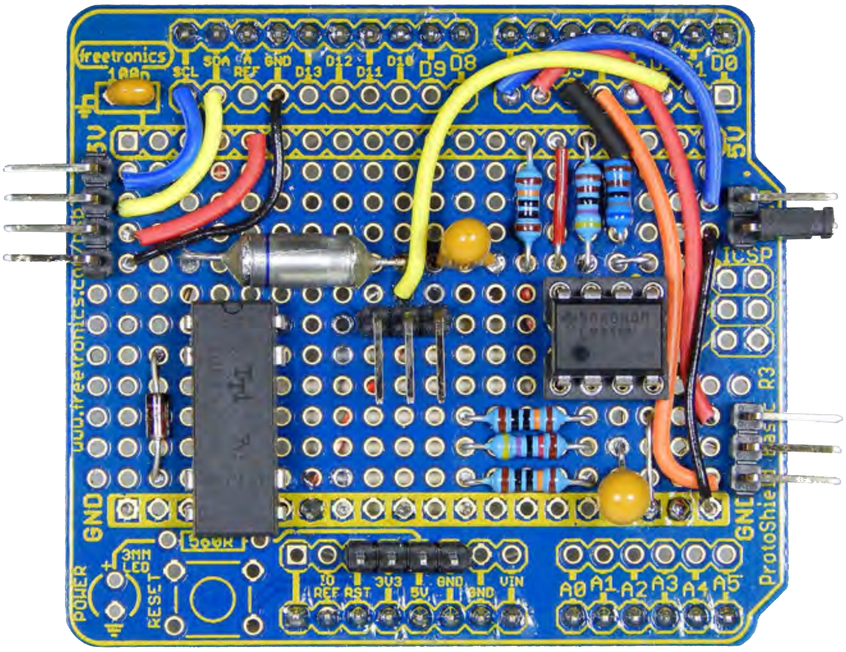


Fig.3: follow these diagrams to fit the components to the ProtoShield and also to wire up all the external connections. The connections made between component pads on the underside are shown below, significantly larger than the 1:1 diagram above, for clarity. These should be made with insulated wire to avoid short circuits.



Box and lid preparation

There are four holes to drill in the bottom of the box for mounting the Arduino module and two larger holes to cut in the left-hand end for the USB plug and alternative DC power plug. The locations and dimensions of all of these holes are shown in Fig.5, the drilling template, while the corresponding information for the holes



'Larger than life' photo of the wiring on the top side of the Freetronics Arduino ProtoShield board (actual size is shown in Fig.3). This board 'plugs in' to the Arduino Uno board via the rows of pin headers on the underside; the I^{2}C LCD board plugs into the ProtoShield board.

to be drilled and cut in the lid/front panel are shown in Fig.6.

For best results, start the larger holes with a smaller pilot drill and enlarge with a stepped drill bit, a series of larger drill bits or a tapered reamer. Rectangular or other non-round holes can be made by drilling a series of holes, knocking out the centre section and then filing the hole to shape.

We fixed four self-adhesive rubber feet to the underside of the box to protect any surface it's placed on.

Making all the required holes in the lid is rather tedious as there are twelve, including three rectangular cut-outs and two holes with flat edges. To save time and guarantee a neat result, you can purchase a laser-cut clear acrylic lid (which replaces the lid supplied with the box) from the Silicon Chip online shop (see parts list).

As the acrylic panel is transparent, the lid doesn't need a cut-out to view the LCD. Note that since the 3mm acrylic is slightly thicker than the lid supplied with the UB3 box, depending on the length of the screws that came with it, you may need to use slightly longer self-tapping screws to attach it.

We have also prepared artwork for the front panel to give it a professional look. You can download this as a PDF file from the *EPE* website.

There are two ways to go here: after you print it, it can be hot laminated, then attach it to the box lid using double-sided adhesive tape or spray glue. After that, you can cut out the holes in the front panel to match those in the box lid using a sharp hobby knife.

Or, for longest life and an even more professional finish, consider fitting the label to the *underside* of the lid – it's more fiddly to fit but doesn't require laminating, nor double-sided tape to hold it in place (the switches and terminals hold it in position; a very light mist of clear spray adhesive will also ensure it stays tight against the lid).

Perhaps it's gilding the lily somewhat, but if you can print the label onto clear film, you can see the 'works' through the label, as we did with the photo on this project's first page.

Just make sure you get the right film to suit your type of printer (eg, laser printer or inkjet printer).

Once the lid/front panel is finished, fit switches S1-S3 to it, along with the two test terminal binding posts and the serial LCD module.

Slide switch S1 attaches to the front centre of the lid via two 6mm-long M2 machine screws, while switch S2 mounts using the spring washer and nut supplied with it, and S3 simply pushes into its rectangular mounting hole until its two barbs that spring outwards to hold it in place.

Just make sure that you fit it with the '=' sign on its rocker actuator uppermost (see photos).

The two binding posts are mounted using the mounting nuts and lock washers provided.

Take care doing so, however, as the upper and lower mounting bushes have D-shaped sections which should mate with the matching holes in the lid/front panel.

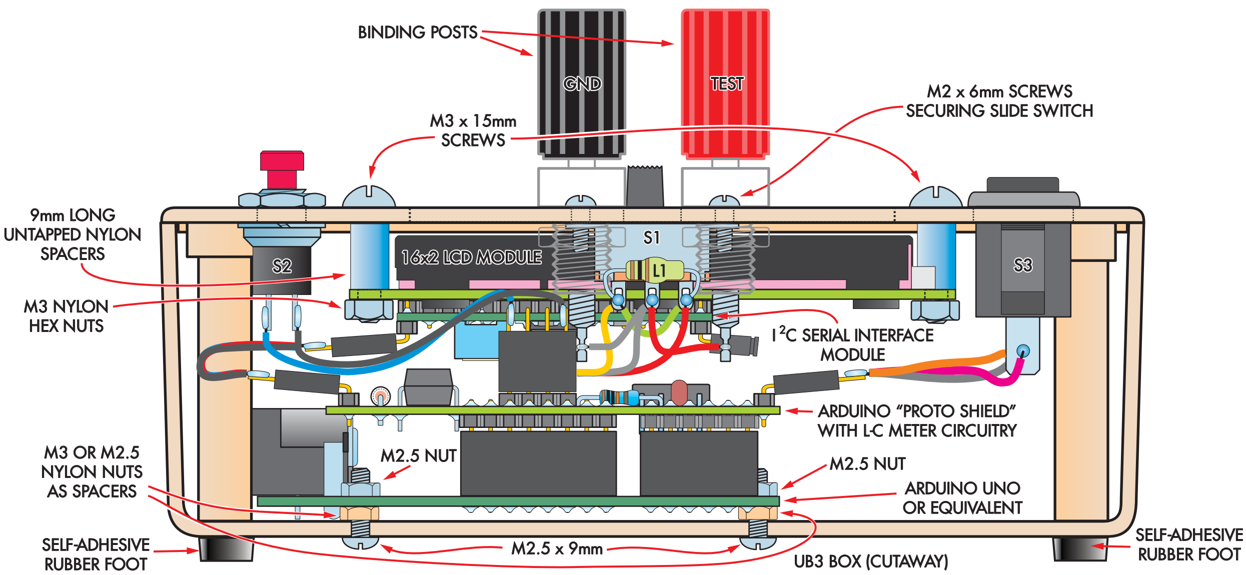


Fig.4: this shows how everything fits together inside the UB3 'Jiffy' box. The Arduino module is attached to the bottom of the case with the proto-board hosting most of the remaining circuitry plugged on top. The three switches, two binding posts and the I^{2}C LCD module are mounted on the lid and connected to the ProtoShield via flying leads.

The Freetronics Eleven (Uno equivalent) board, mounted in the bottom of the case (see drilling template).

The serial LCD module mounts under the lid in the top centre position, using four 15mm-long M3 machine screws passing down through four 9mm-long untapped nylon spacers and fastened using four nylon M3 nuts (under the module PCB).

With the LCD module in position, your front panel assembly is ready to be wired up and provided with its various leads to connect to the ProtoShield board.

Refer back to Fig.3 and the internal photo, following them carefully to make the correct connections between S1, the test terminal binding posts and L1 and C1 in particular.

Note that the leads of L1 and C1 should be kept as short as possible, to keep stray capacitance low (and stable). You can then make up the various short leads which will connect the front panel components to the ProtoShield board.

Note that the lead which connects S1, L1, C1 and the test terminals to the ProtoShield ends in a three-way SIL header socket, as does the lead from S3.

In contrast, the lead which connects to the serial LCD module has a four-way SIL header socket at each end, while the lead to connect zero/reset switch S2 (although of only two wires) ends in a four-way SIL header socket, with the wires connecting only to the pins on each end.

The two pins in the centre of the socket can be either cut short or pulled out, since they are not used.

Rather than using SIL sockets like we did on the prototype, we suggest you simply split a 40-way ribbon jumper cable with individual 'DuPont' sockets on each wire.

This makes the job really easy; you simply pull off the required number of wires and then cut the cable to length and strip the free end, to solder to the switch or connector.

You don't even need to cut the cable for the LCD, you can just plug it in at both ends.

In each case, make sure each wire goes to the correct pin as with individual sockets, it's easy to get them out of order.

Having made up all the required leads, complete the *Digital LC Meter* assembly with the following steps:

1. Mount the Arduino module inside the bottom of the box using four 9mm M2.5 screws and nuts, using four nylon M3 nuts as spacers.
2. Plug the *Digital LC Meter* ProtoShield into the Arduino, making sure you have all four SIL pin headers lined up correctly.
3. Holding the front panel assembly close to the top of the box and oriented correctly, plug the various connection leads into their matching pin headers on the ProtoShield. Be especially careful to get the correct connections between the ProtoShield and the LCD module, as shown in Fig.3.
4. Lower the lid assembly down into the box and fix it into place.
5. Program the Arduino, as described below.

Uploading the firmware

To upload the firmware you need to have the Arduino IDE installed on your PC. The latest version of the IDE can always be downloaded from the Arduino website: **www.arduino.cc/en/Main/Software**

At the time of writing, the latest version is V1.8.5. There are various versions available to suit different operating systems: Windows (32-bit or 64-bit), macOS and Linux (32-bit, 64-bit and ARM).

After the IDE has been installed, download the firmware sketch for the *Digital LC Meter* from the *EPE* website – **Arduino_LC_meter_sketch.ino**

Parts list – Arduino-based LC Meter

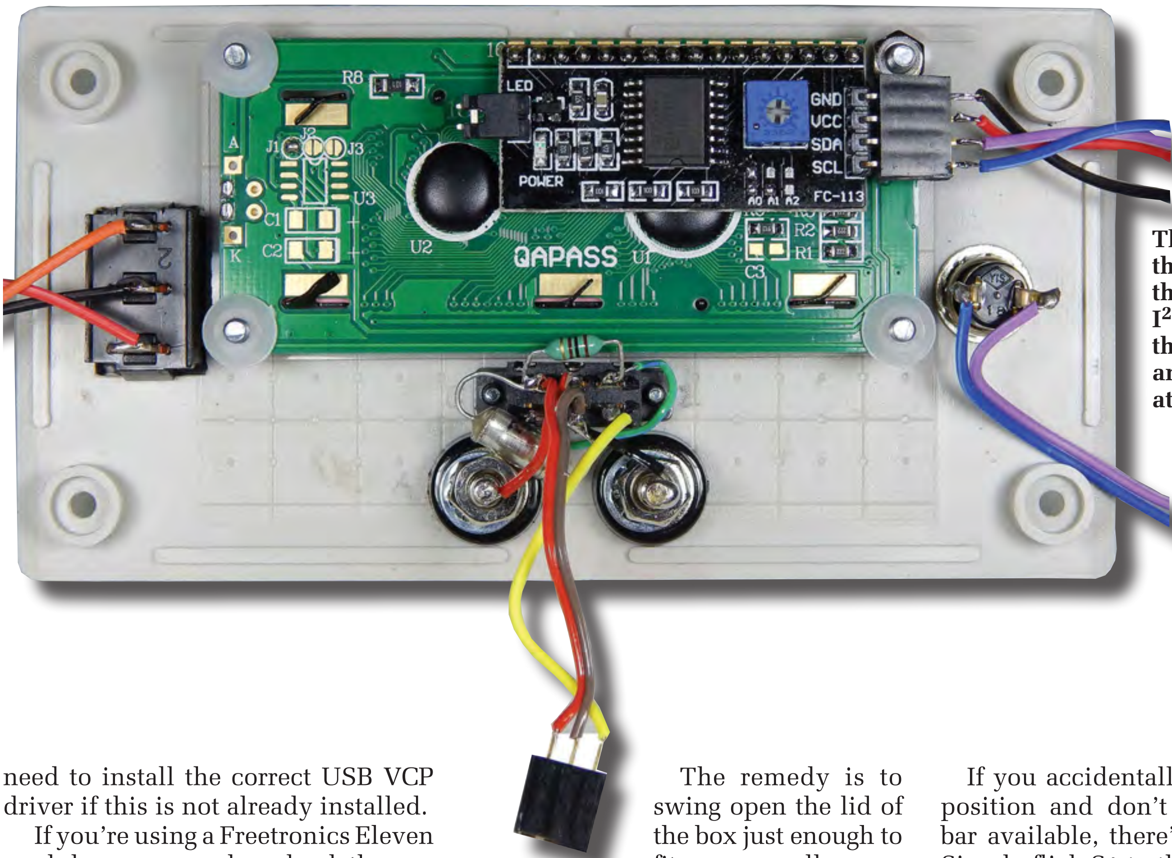
1 Arduino Uno R3, Duinotech Classic, Freetronics Eleven or equivalent microcontroller module
1 Serial I^2C 16×2 LCD module with back-lighting
1 Arduino Uno Prototyping Shield
1 UB3 'Jiffy' box, 130 × 68 × 44mm
1 laser-cut clear acrylic lid for UB3 box [optional but recommended]
4 self-adhesive rubber feet
1 5V/10mA DIL reed relay (RLY1)
1 100µH axial RF inductor (L1)
1 DPDT subminiature slide switch (S1)
1 panel-mount SPST NO momentary pushbutton switch (S2; SP0710)
1 panel-mount SPDT on-off-on momentary rocker switch (S3)
1 8-pin DIL IC socket
1 40-pin header, 2.54mm pitch
1 40-pin right-angle header, 2.54mm pitch
1 150mm socket-to-socket jumper ribbon cable
1 jumper shunt
2 binding posts with integral banana socket (1 red, 1 black)
4 9mm nylon untapped spacers, 3mm inner diameter
4 15mm M3 machine screws
8 M3 nylon hex nuts
4 9mm pan head M2.5 machine screws
4 M2.5 hex nuts
2 6mm M2 machine screws (for S1)

Semiconductors
1 LM311 DIP high-speed comparator (IC1)
1 1N4148 small-signal diode (D1)

Capacitors
2 10µF 16V through-hole tantalum
1 100nF multilayer ceramic
2 1nF 1% NP0 ceramic, mica, MKT, polypropylene or polystyrene

Resistors (all 0.25W, 1%, through-hole mounting)
3 100kΩ 1 47kΩ 1 6.8kΩ 1 4.7kΩ

Now plug your *Digital LC Meter* into one of your PC's USB ports, using a suitable USB cable (usually with a Type A plug on one end, and a micro Type B plug on the other). You may



The underside of the lid, showing the LCD module, I^2C module, the three switches and two terminals attached.

need to install the correct USB VCP driver if this is not already installed.

If you're using a Freetronics Eleven module, you can download the appropriate driver from their website (**www.freetronics.com.au**). All of their drivers are zipped up in a file called '**FreetronicsUSBDr ivers_V2.2.zip**', and there's also a document which explains how to install it.

Once the USB driver has been installed and your operating system confirms that it can communicate with the Arduino in your *Digital LC Meter*, use Control Panel to find out which COM port the meter's Arduino has been allocated (eg, COM5, COM7, or whatever).

Set the port for communication at 115,200 baud with the 8N1 'no handshaking' protocol. The COM port number should be entered into the Arduino IDE's Tools->Port pull-down menu after you start it up.

Now open the *Digital LC Meter* firmware sketch in the Arduino IDE, verify and compile it, and then upload it into the *Digital LC Meter's* Arduino Flash memory. Soon after it has been uploaded, your meter should spring into life, flashing this message on the LCD screen:

Silicon Chip
Digital LC Meter

This should remain visible for two seconds, after which the screen should go black, before the meter begins its initial zero calibration. If you don't see this initial message, this may be because the contrast trimpot on your LCD display module's serial interface PCB is not set to the correct position.

The remedy is to swing open the lid of the box just enough to fit a very small screwdriver or alignment tool into the trimpot's adjustment slot, turn it and then press switch S2 to force the Arduino to reset and start again.

Try changing the pot setting in one direction or the other until the message becomes clearly visible, pressing S2 after each adjustment.

Startup and calibration

At start-up, the *Digital LC Meter* normally expects slider switch S1 to be set in the Capacitance (C) position, and no external capacitor to be connected to the test terminals. If you have done this it will now display the message:

S1 set for C: OK
Now calibrating

But if you have set S1 in the Inductance (L) position instead, you'll see a different message:

Fit shorting bar
Now calibrating

This means that the *Digital LC Meter* has detected that S1 is set to the L position, and is assuming that you want to do the zero calibration in this mode. As a result, it's advising you to fit a very low inductance shorting bar between the test terminals. This can be in the form of a 40mm-long piece of 1.66mm-diameter copper or brass rod between the terminals, or (better still) a 40 × 30mm rectangular piece of 1mm-thick copper or brass sheet. In either case, the rod or sheet must be shiny rather than oxidised.

If you accidentally set S1 to the L position and don't have a shorting bar available, there's no harm done. Simply flick S1 to the C position and then press switch S2 to get the *Digital LC Meter* to reset and begin over again.

Or, if you do want to calibrate in inductance mode, simply fit the shorting bar between the terminals (if you haven't already done so) and then press S2 to reset and begin over again.

In either case, there will be a brief pause, after which the meter will show the values for C1 and L1 it has found from the initial calibration. This will be something like:

C1 = 1084.2 pF
L1 = 91.24 uH

The actual values displayed will depend on the components in your unit, as well as the stray capacitance and inductance. They're shown at this stage mainly as reassurance that the *Meter* is working correctly. The measured values of C1 and L1 will be displayed for three seconds, after which this message will appear:

Calibration done
Ready to measure

This will display for one second, after which the *Digital LC Meter* will begin making measurements. If you have done the initial calibration in C mode and S1 is still in this position but no unknown capacitor is connected to the test terminals, you should now get a display like this:

Cx = 0.004 pF
(F3 = 515838 Hz)

where the value shown for C_x is very

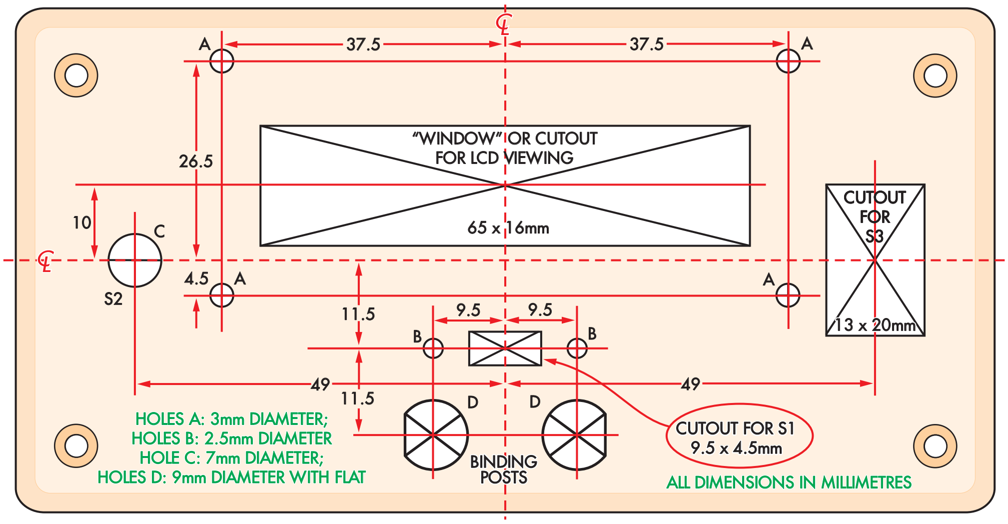


Fig.6: you can either drill and cut the twelve cut-outs required in the lid supplied with the UB3 'Jiffy' box, as shown in this diagram, or purchase a laser-cut acrylic lid from the SILICON CHIP online store and use that instead of the lid that came with the box.

close to zero, while the frequency F3 shown on the second line is for the current oscillator frequency; essentially the same as F1 at the current ambient temperature.

The *Digital LC Meter's* oscillator frequency does drift a little with temperature. This means that after a while, the value shown for Cx with no external capacitor connected may creep up from the almost-zero reading you get initially. At the same time, the reading for F3 would slowly decrease.

If you find the value shown for Cx with no external capacitor has crept up to 0.1pF or more, simply press S2 again to get the Arduino to perform a new zero calibration.

On the other hand, if you've done the initial calibration in L mode and S1 is still in this position but the shorting bar is still connected across the terminals, you should get a display like this:

```
Lx = 0.002 uH
(F3 = 516615 Hz)
```

The value shown for Lx is again very close to zero, and the frequency F3 shows the current oscillator frequency, again very close to F1 at the current ambient temperature. Now, if you remove the shorting bar in this mode, you'll find the display will change to something like this:

```
Over Range!
(F3 = 2 Hz)
```

This simply shows that in this mode, an open circuit between the terminals is equivalent to a very high inductance, because it causes the oscillator frequency to drop to near zero.

When you connect a real inductance between the test terminals, the *Digital LC Meter* will measure its inductance and display it (assuming its value is within its range, which is from 10nH to 150mH).

As before, drift in the *Digital LC Meter's* oscillator may cause the Lx reading for the shorting bar to creep up gradually. So before making a particularly critical measurement, it's a good idea to fit the shorting bar between the test terminals and press S2 again to force the Arduino to reset and perform a new zero calibration.

Optimising accuracy

If all is well so far, your *Digital LC Meter* should be operating correctly and ready for use. If you have been able to procure a couple of 1% tolerance (or better) capacitors for C1 and C2, it should also be able to deliver that order of accuracy without any extra calibration.

However, as mentioned earlier, it is possible to achieve even better accuracy with the meter providing you have access to a reference capacitor whose value is accurately known (because you've been able to measure it with a high-accuracy LCR meter).

Ideally, this reference capacitor should have a value of between 10nF and 100nF, but even one with a value between 1nF and 10nF would be OK.

This is achieved by tweaking or 'nudging' the *Digital LC Meter's* reading for the reference capacitor using switch S3. Here's how you do it:

1. Remove 5V supply from the *Digital LC Meter*.
2. Lift the lid/front panel up from the box and carefully fit the jumper shunt over the pins for LK1, down on the ProtoShield.
3. Close the box and slide S1 to the C position but don't connect your reference capacitor to the test terminals.
4. Re-apply the 5V power and let the *Digital LC Meter* go through its initial zero calibration.
5. Wait a couple of minutes, watching the reading for Cx to see if it drifts up appreciably from the initial near-zero figure. If it does, press switch S2 to force a reset and bring the reading back to less than 0.01pF.
6. Connect your known-value capacitor to the test terminals and note the *Digital LC Meter's* measurement reading. It should be fairly close to

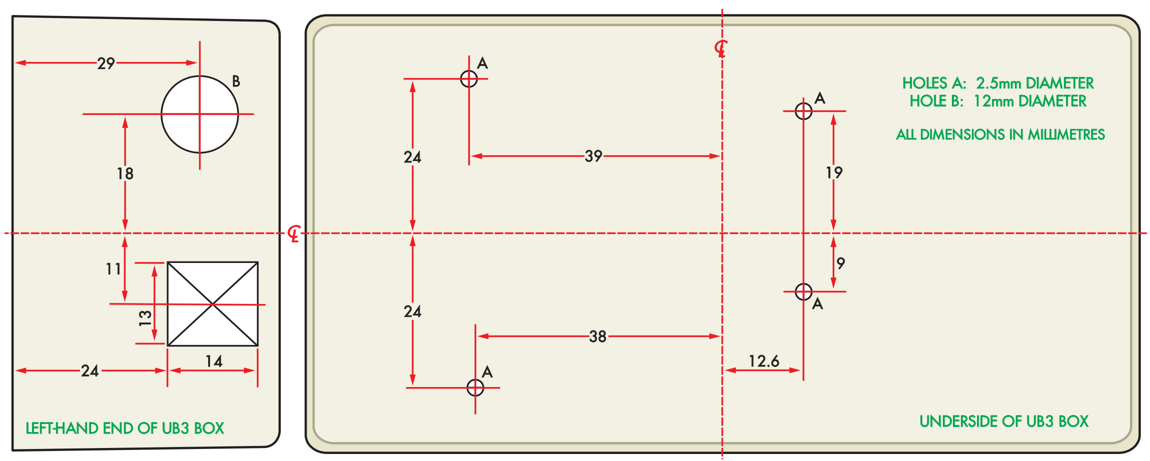


Fig.5: the drilling templates for the four Arduino mounting holes in the bottom of the box, along with the USB and DC power access holes in the left-hand end.

Here's the alternative finish using a paper-printed label fixed to the outside of the UB3 Jiffy box lid, after it has been drilled and cut to suit. (You can, of course, glue a paper label to the laser-cut lid purchased from the Silicon Chip online store). In this case, the meter is measuring a nominal 100µH inductor and showing it is slightly high at 103µH.

the capacitor's known value, but may be a little higher or lower.

7. If the reading is too low, press the rocker of switch S3 at the upper ('=') end for a second or so; if it's too high, press the lower end ('-') instead. The reading should change by about 0.5%. Continue until the reading is as close as possible.
8. Remove power, open the lid and remove the jumper from LK1.
9. Reattach the lid.

The Arduino saves the revised calibration factor in its EEPROM after every measurement during this nudging procedure, so you only have to do the calibration once. Also, when you calibrate the *Digital LC Meter* using a known capacitor, it's calibrated for inductance measurements too.

Credit where it's due

As mentioned earlier, this Digital LC Meter, like our earlier May 2010 design, is based on a 1998 design by the late Neil Heckt, of Washington, USA.

Since then, various people have produced modified versions of the design, including Australian radio amateur Phil Rice VK3BHR, of Bendigo in Victoria. Mr Rice and others also modified the firmware and adapted it to use the PIC16F628 micro with its internal comparator. They also added a firmware calibration facility.

So a significant amount of credit for this latest version of the design must go to these earlier designers. The author is happy to acknowledge their work.

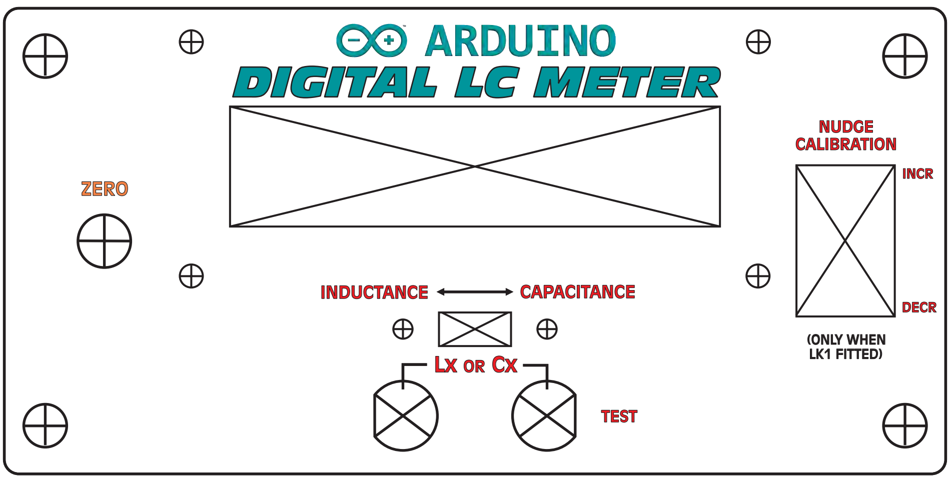


Fig.7: same-size front panel artwork designed to fit a UB3 Jiffy Box. It will also fit the laser-cut acrylic front panel from the Silicon Chip online store. This, along with the two cutting/drilling diagrams, can also be downloaded (as a PDF) from the *EPE* website.

Using cheap Asian Electronic Modules Part 5

by JIM ROWE

The 'New Blue' 16x2 LCD module with piggy-back I²C serial interface

This module combines a 16x2 backlit alphanumeric LCD module with a small 'piggy-back' module that provides it with an I²C serial interface. This allows it to 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.

LCD modules with two lines of 16 characters have been around for many years and we've used them in numerous projects. They are also now much cheaper thanks to their popularity for use with Arduino, Micromite and the Raspberry Pi.

However, many of these Arduino and other microcontrollers are a little limited when it comes to I/O pins, which means that the six or seven pins required to interface to a standard LCD module can leave you with too few pins to interface with other systems and components.

This problem can be solved by using an LCD with a serial interface, or alternatively, attaching a small piggy-back module to a parallel LCD to provide serial/parallel translation.

By using a piggy-back module that communicates using the 2-wire I²C protocol, you end up with an LCD that can be driven using just two wires: one for the serial data (SDA) and the other for the serial clock (SCK). That's apart from the ground and power wires (typically +5V).

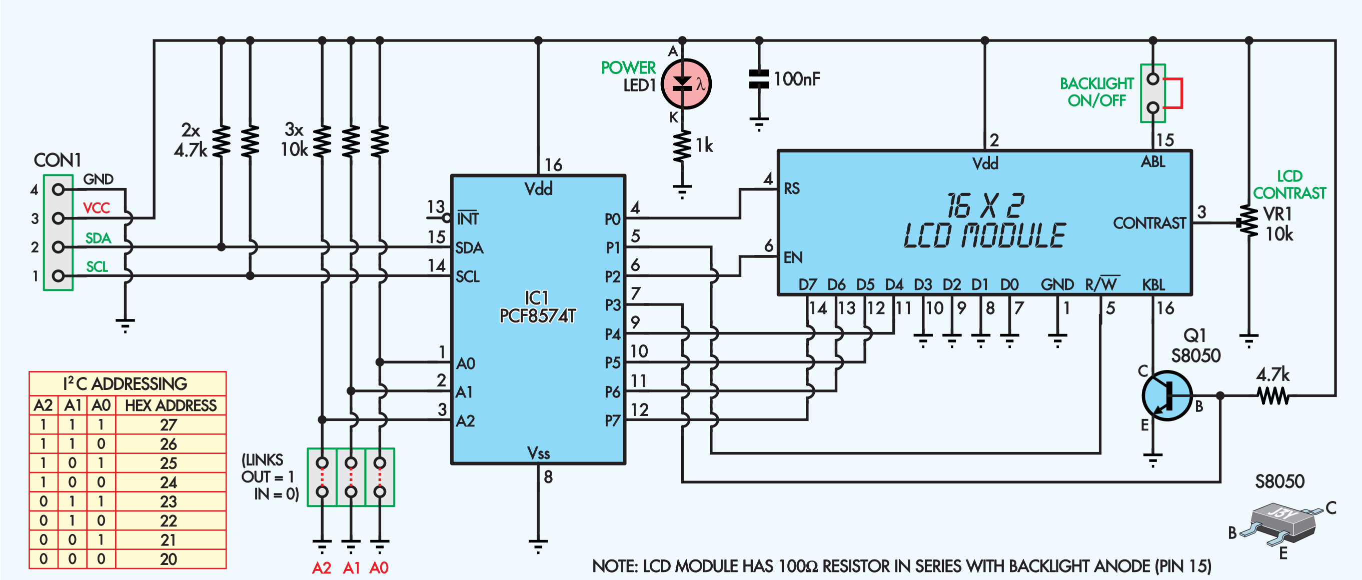
What's inside

The circuit of Fig.1 shows the LCD module at upper right, with the rest of the circuitry – the piggy-back, which connects to the module via the usual 16-pin SIL connector – along the top.

All of the serial-to-parallel conversion is performed by IC1, a Philips/NXP PCF8574T device. This is designated as a 'remote 8-bit I/O expander for the I²C bus'. In other words, this IC accepts serial data over the I²C two-wire bus, via pins 14 and 15, and it makes the data available in parallel format at pins 4-7 and 9-12.

In this case, output pin 4 is used to control the LCD's RS (register select) control pin, while pin 6 controls the EN (enable) pin and pins 9-12 feed the character codes to pins D4-D7 of the LCD. That leaves pin 5 of IC1 to control the LCD's R/$\overline{W}$ pin, and pin 7 to control the LCD backlight via transistor Q1.

What about pins 1, 2 and 3 of IC1? They're used to set the address of IC1 on the I²C bus. All three pins have 10kΩ pull-up resistors connecting them to logic high (V_{CC}) – but do note that the LCD module PCB also provides three pairs of tiny pads so that any of the pins can be tied to ground.



I²C ADDRESSING

A2	A1	A0	HEX ADDRESS
1	1	1	27
1	1	0	26
1	0	1	25
1	0	0	24
0	1	1	23
0	1	0	22
0	0	1	21
0	0	0	20

Fig.1: complete circuit for the piggy-back and LCD module together. Some of these modules use a slightly different chip for IC1 – the PCF8574AT – the main difference being the hex address range is instead from 38h to 3Fh.

This allows the chip's I^2C address to be set to any hexadecimal value between 0x20 (32) and 0x27 (39), just by bridging the pairs of pads, as shown in the small table at lower left in Fig.1.

So the default I^2C address of the piggy-back module (and thus LCD) is 0x27 with all links out, but this can be changed to 0x20 simply by fitting all three links, or to any address in between by fitting one or two links. This allows a number of the LCD-piggyback combinations to be connected to the same I^2C bus, with each one given a different I^2C address so that the micro driving the bus can send data to any one it chooses.

Other I^2C devices can reside on the same bus (eg, temperature sensors, memories, other microcontrollers), as long as you ensure that no two devices have the same address.

There are some serial I^2C LCD modules that use a slightly different chip for IC1, the PCF8574AT. This is virtually identical to the PCF8574T shown in Fig.1, except that the I^2C address range is between 0x38 and 0x3F.

By using a combination of the two chips, up to 16 different serial I^2C LCDs may be connected to the same I^2C bus, provided you use eight with the PCF8574T bridge chip and eight with the PCF8574AT chip.

Fig.1 also shows that the piggy-back has a power-on indicator (LED1), a 2-pin SIL connector and jumper shunt that can be used to disable the LCD's backlight if not required. Trimpot VR1 can be used to adjust LCD contrast in the usual way (via pin 3).

Note that the SDA and SCL lines connecting between pins 1 and 2 of CON1 and pins 14 and 15 of IC1 are each fitted with a 4.7kΩ pull-up resistor, as the I^2C bus uses active-low logic. These resistors can be left in place if the module is the only slave connected to the I^2C bus. However, if you're going to be hooking up other I^2C slave devices to the same bus, all but one should have the SDA and SCL pull-up resistors removed.

Using it

This type of module really needs to be hooked up to a micro, and that turns out to be fairly easy to do with any of the popular options available.

All you have to do is connect the V_{CC} and ground pins to a suitable voltage source (which may be the same one that's powering the micro) and the SDA and SCL pins to the I^2C bus pins on the microcontroller. Fig.2 shows how this is done with an Arduino Uno or a compatible like the Duinotech Classic. It couldn't be much simpler.

By the way, although the SDA and SCL pins of the LCD module are shown in Fig.2 connected to the

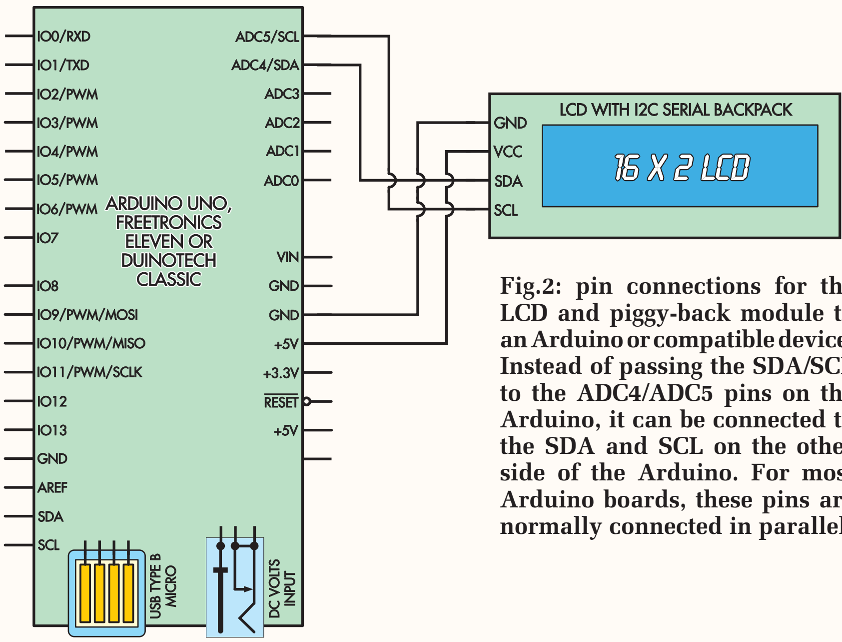


Fig.2: pin connections for the LCD and piggy-back module to an Arduino or compatible device. Instead of passing the SDA/SCL to the ADC4/ADC5 pins on the Arduino, it can be connected to the SDA and SCL on the other side of the Arduino. For most Arduino boards, these pins are normally connected in parallel.

ADC4/SDA and ADC5/SCL pins at upper right on the Arduino, they could instead be connected to the pins marked SDA and SCL on the other side of the Arduino down near the USB connector. On most Arduino boards, these pin pairs are connected in parallel.

It's just as easy to connect the serial I^2C LCD module to a Micromite, as you can see from Fig.3.

Connecting the module up to a micro is only half the story. Then you have to work out how to get the micro to send it the data you want displayed.

The complicating factor here is that quite a few people have written 'libraries' to make it easier to drive this kind of serial I^2C LCD module from an Arduino sketch, by providing a set of simple function calls like:

```
lcd.print("Text");
```

This is all very well, but even though most of these library files have the name **LiquidCrystal_I2C.h**, they are often different in terms of their finer details and compatibility with any particular serial I^2C LCD module.

Rather than you going through the same sort of hassles we did to find a suitable library, we'll simply point you at some that we found to work. These are available at the following Github links:

http://bit.ly/2sHf6Cd
http://bit.ly/2sLrpxm

It's possible that these are actually the same library, because in one place we found the author listed as Frank de Brabander but the maintainer as Marco Schwartz. We found both through the following website: **www.arduinolibraries.info/libraries/**

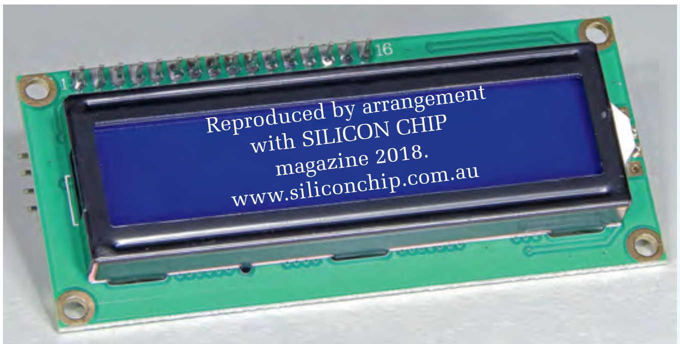
Anyway, these libraries do seem to work with the module shown, as you'll find out by downloading the 'Hello World' sketch (**HelloWorld.ino**) from the *EPE* website (**www.epemag.com**) and running it. We've included a copy of the library (as a ZIP file) within the package. The resulting display is shown in the adjacent photo.

Don't forget to change the I^2C address shown in the sketch (0x27) to 0x3F (= 3Fh), if your piggy-back module is fitted with a PCF8574AT instead of a PCF8574T. You'll also have to change this address if you've changed the address using the three small pairs of pads.

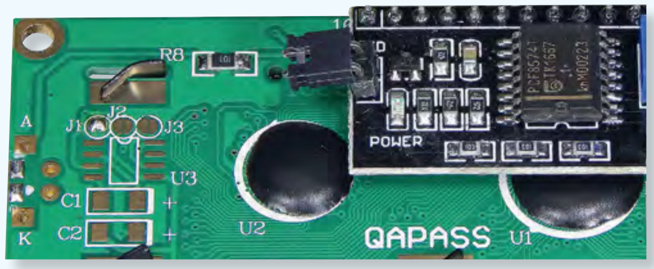
Reading hexadecimal numbers

In this article, values prefixed with '0x' correspond to a hexadecimal number; you might also see values suffixed or prefixed with 'h'. Reading from left-to-right, each character corresponds to a 4-bit long value. With 0-9 being equal to themselves and A-F (case-insensitive) equal to 10-15 respectively. A hexadecimal value is calculated as if each character is appended to the other to form one long string of bits.

Thus, 0x5A (or 5Ah) is equivalent to 01011010 in binary and 90 in decimal form. A string of bits can be read as the sum of each individual non-zero bit, with each bit being equal to 2^{n-1} where n is the index of that bit starting from the right. So, $0101 = 0 + 2^{3-1} + 0 + 2^{1-1} = 4 + 1 = 5$. A longer example BCDEh would just be equal to 48350 in decimal and 1011110011011110 in binary.



The top of the LCD module. The screen is mounted on a PCB measuring 80 × 36mm, while the visible area of the LCD measures 64 × 14.5mm.



The underside of the LCD module's PCB has the piggy-back module (black) located above it. The jumper shunt located on the piggy-back module can be used to disable the LCD backlighting if it's not needed.

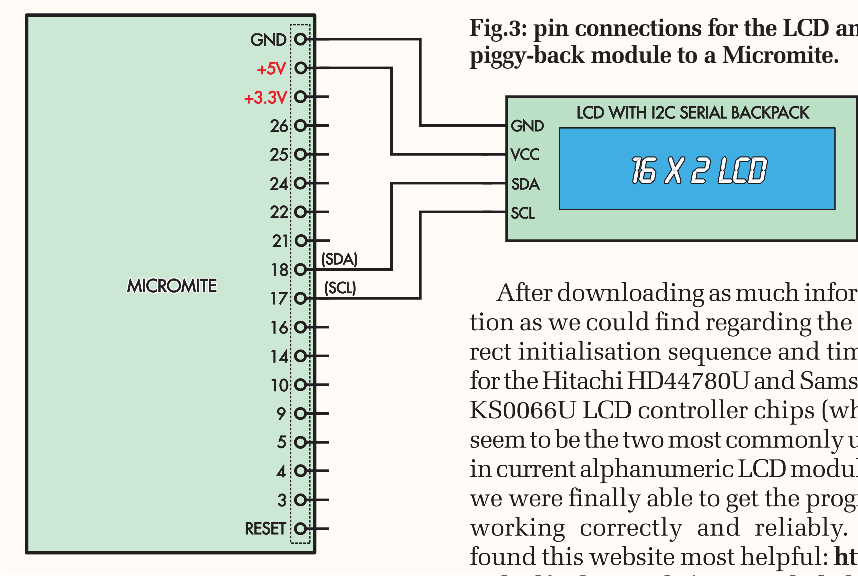


Fig.3: pin connections for the LCD and piggy-back module to a Micromite.

How about a Micromite?

Programming a Micromite to talk to the I^{2}C LCD module is not quite as easy as with an Arduino, as currently the inbuilt MMBasic LCD commands only support the parallel interface. You will find a program called **I2CLCD.bas** in the MMBasic Library, which can be downloaded in zipped-up form from the bottom of this page: **http://geoffg.net/maximite.html#Downloads**

However, this program was written for a piggy-back module with a different configuration than the one which most piggy-backs seem to use (and we have shown in Fig.1). Then there's a further issue in that the I^{2}C command syntax has changed as MMBasic has evolved. As a result, we ended up having to re-write the software completely.

Changing the program's commands to suit the different connections between the PCF8574T bridge chip and the LCD module itself wasn't too hard. The major difficulty was in getting the program to initialise the LCD's controller correctly.

The correct set-up commands have to be sent to it soon after power is applied, and these commands have to be sent in a particular order, with pauses between them to allow the controller to process them before the next command arrives for correct operation.

After downloading as much information as we could find regarding the correct initialisation sequence and timing for the Hitachi HD44780U and Samsung KS0066U LCD controller chips (which seem to be the two most commonly used in current alphanumeric LCD modules), we were finally able to get the program working correctly and reliably. We found this website most helpful: **http://web.alfredstate.edu/weimandn/lcd/lcd/lcd_initialization/**

Basically, our program (called **JRI-2CLCD.bas**) just displays a 'Hello, world!' message over and over on the LCD; just like the one for the Arduino. You can download this from our website, open it in MMEdit and then upload it to your Micromite and you should get the same display as shown in the photos.

As with the Arduino sketch, you may need to change the I^{2}C address given for your display's piggy-back, if it has some of the address links fitted or is using the PCF8574AT chip instead of the PCF8574T. Look for this line near the start:

```
DIM AS INTEGER I2CAddr = &H27
' (A2=A1=A0=1)
```

All you need to do is change '&H27' into the correct address for your module. This program provides a good starting point for writing your own MMBasic programs using an I^{2}C LCD. It's fairly well commented, so you should be able to see how to adapt the program to display other things.

Where to buy

We've stocked some of these modules in the Silicon Chip Online Shop so that you can acquire and experiment with them.

Alternatively, you can find similar units (either pre-assembled or as two separate items) on eBay and AliExpress, and also 20×4 character I^{2}C LCDs which cost very little more than the 16×2 types.

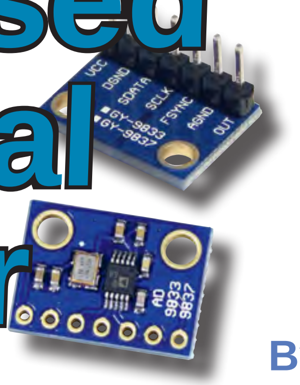
The piggy-back should also work with 20×2 and 16×4 size alphanumeric LCDs; however, these are far less popular than the other two sizes.

Serial USB-UART bridge module – another version

Since writing the third article in this series (for the March 2018 issue), we've become aware of another popular version of the serial USB-UART bridge module based on the CP2102 device. This one is very similar to the one we discussed in the March 2018 article, but differs in two respects. One is that instead of a micro-USB socket on the USB end of the module, it is fitted with a full size type A USB plug – providing a more rugged connection and compatibility with a standard USB type A to type A extension cable.

The other difference (wait for it!) is that the connections to the six pins of the SIL connector on the other end of the module are **NOT** the same as those on the smaller module. So make sure that you allow for the differing SIL pin connections when you connect the module to your micro or other device.

AD9833-based Direct Digital Synthesiser



Using Cheap Asian Electronic Modules Part 6

By JIM ROWE

This little signal generator module uses an Analog Devices AD9833 DDS chip and a 25MHz crystal oscillator. It can be programmed to generate sine, triangle or square waves up to 12.5MHz and it's all controlled via an SPI serial interface.

Direct Digital Synthesiser or DDS chips have been around for well over 20 years now, but for much of that time they were fairly costly.

Until recently, they didn't include an integral DAC (digital-to-analogue converter), so you had to use their digital output to drive a separate DAC to generate the analogue output signal.

In the early 2000s, Analog Devices Incorporated (ADI) announced a new generation of complete DDS devices which did have an integral DAC, as well as offering high performance combined with a price tag significantly lower than what you used to have to pay for a DDS+DAC combination.

Although it's one of the low-cost, lower-performance devices in their range, the AD9833 provides a good example of just what can be achieved nowadays.

When combined with a 25MHz crystal oscillator, it can be programmed to produce any output frequency from 0.1Hz to 12.5MHz in 0.1Hz increments, with a choice of three waveforms: sinusoidal, triangular or square.

All this comes from a chip housed in a tiny MSOP-10 package, running from a supply voltage of 2.3-5.5V, dissipating only 12.65mW and currently with a price tag of around £10 in one-off quantities. That's significantly lower than earlier DDS chips.

As we've seen in previous articles in this series, there has also been a huge surge in the manufacture of many kinds of electronics modules in Asia, especially China, some of them available at surprisingly low prices via internet markets like eBay and AliExpress.

As a result, you can buy the tiny (18 × 13.5mm) AD9833-based DDS module shown in the photos, which includes a 25MHz crystal oscillator, for the princely sum of £4 each – including free delivery to the UK! That's really quite a bargain, which is why we're focusing our attention on it this month.

To get an idea of how a DDS works, take a look at the panel, *DDS in a Nutshell*, at the end of this article.

Inside the AD9833

The block diagram of Fig.1 shows what's inside that tiny MSOP-10 package. There's quite a lot, although some of the elements are mainly involved in giving the chip its flexibility in terms of output waveform and modulation capabilities. The main sections involved in basic DDS operation are those shown with a pale yellow fill.

Down at lower left in Fig.1 you can see the 16-bit shift register where data and instructions are loaded into the chip from almost any micro, via a standard SPI (Serial Peripheral Interface) bus. We'll discuss that in more detail later.

Just above the serial input register is the control register, also 16-bit. This stores the control words, used to set up the configuration of the device, including the output waveform type, which of the two 28-bit frequency registers (FREQ0 or FREQ1) is used to set the DDS output frequency and also whether the phase is shifted by the content of 12-bit phase registers PHASE0 or PHASE1.

The main reason why the AD9833 has two frequency registers and two

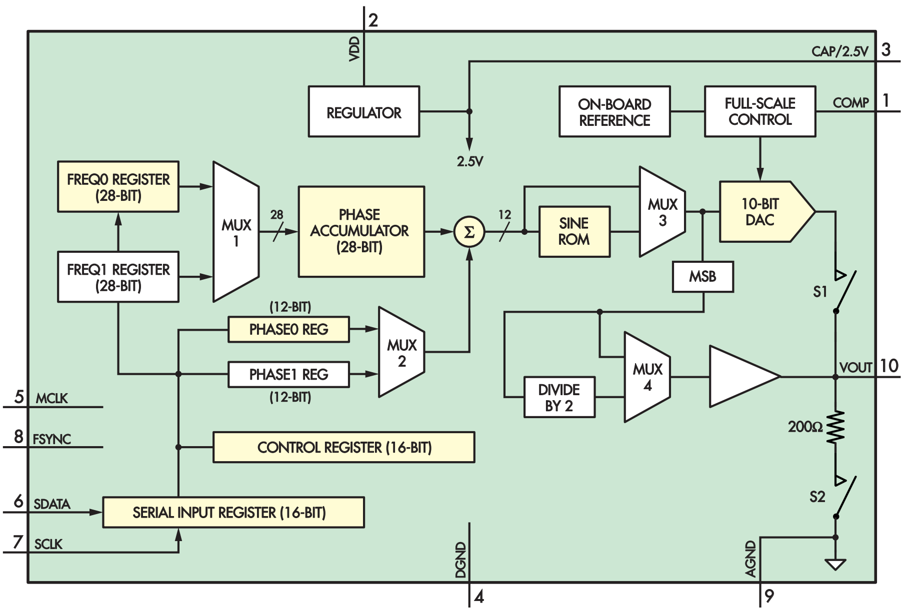


Fig.1: block diagram of the AD9833 DDS IC. The critical blocks are yellow. The phase accumulator generates a series of addresses to look up in the ROM sine table and the resulting values are then fed to a 10-bit DAC which produces the output waveform. Other circuitry allows the output waveform to be changed to a triangle or square wave and also allows for frequency and phase shift keying.

phase registers is to give it the capability of generating signals with frequency-shift keying (FSK) or phase-shift keying (PSK) modulation. Multiplexers MUX1 and MUX2 allow these options to be controlled using bits in the control register.

So how are those 28-bit frequency/phase increment registers FREQ0 and FREQ1 loaded with 28-bit data from the 16-bit serial input register? This is done by sending the data in two 14-bit halves, in consecutive 16-bit words from the micro, with the lower half first and then the upper half.

The AD9833 can be configured to accept the data this way simply by manipulating two bits in the control register. The same bits can also be used to configure it for setting either the lower or higher 14-bit 'half word' alone, which can be useful for some applications (such as frequency sweeping).

MUX3, MUX4 and switches S1 and S2 are all controlled by further bits in the control register. MUX3 simply allows the sine ROM to be bypassed, with the output from the phase accumulator fed directly to the DAC. This is how the AD9833 produces a triangle wave output, since the amplitude of a triangle wave is a linear function of its phase.

For a square wave output, the DAC is disconnected from the chip's analogue output (pin 10) using integrated switch S1, and instead makes use of the MSB (most-significant bit) output of MUX3. This automatically gives a square wave output and MUX4 allows you to divide its frequency by two, if needed.

The integrated 200Ω resistor connected between the analogue output pin and ground via switch S2 is used to convert the DAC's output current into a proportional voltage output. Since S2 is controlled in parallel with S1, this means that when S1 cuts the link between the DAC output and pin 10, S2 also removes the built-in 200Ω output shunt.

This makes the chip's output voltage swing in square-wave mode significantly higher than for the sine or triangular (DAC-derived) options.

To be specific, the square wave output is around 5.2V peak to peak, while for sine or triangular waves the output drops to around 650mV peak-to-peak.

The complete module

Now refer to Fig.2, which shows the complete circuit for the 18 × 13.5mm module shown in the photo opposite.

It simply comprises the AD9833 DDS chip (IC1) and its equally tiny (3 × 2.2mm) 25MHz crystal oscillator. There are six even smaller SMD

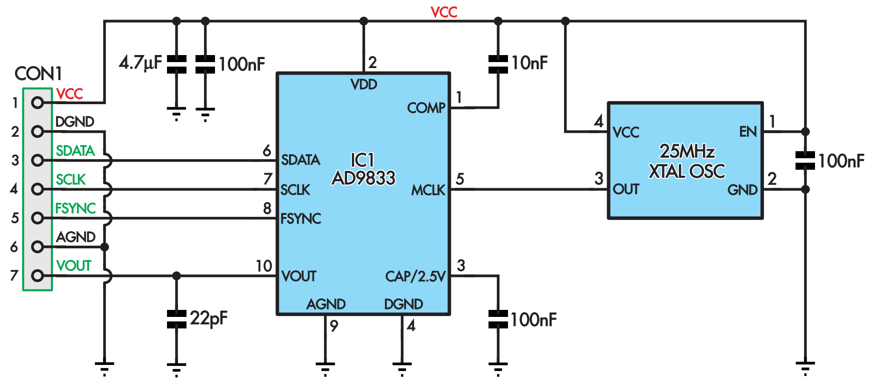


Fig.2: circuit of the AD9833-based DDS module used in this article. The AD9833 IC and 25MHz crystal oscillator plus a few passive components are mounted on a small PCB, with a SIL header to make control and output connections.

capacitors, most of them used for filtering either the power supply rails or IC1's V_{OUT} pin.

Seven-way SIL connector CON1 is used to make all of the signal and power connections to the module. Pins 1 and 2 are used to provide the module with 5V power, while pins 3-5 are used to convey the SPI commands and data to IC1 from the micro you're using to control it. And pins 6 and 7 are used to carry the analogue output signal from IC1 out to wherever it's to be used.

Limitations

Before we talk about driving the module from a micro like an Arduino or a Micromite, we should discuss its limitations.

First, the aforementioned difference in output amplitude for square wave versus sine/triangle waves is a factor of about eight times, or 18dB. So if you want to use the module as the heart of a function generator, you will need to attenuate the square wave output by 18dB, to match the sine and triangle output levels.

You will also need to pass the sine and triangular outputs through a low-pass filter with a corner frequency of around 12-15MHz, to remove most of the DAC switching transients. After this processing, the outputs can all pass through a common buffer amplifier and output attenuator system.

You don't need to worry about any of these niceties if you simply want to use the module as a programmable clock signal source. You can just program it to generate a square wave output and use it as is.

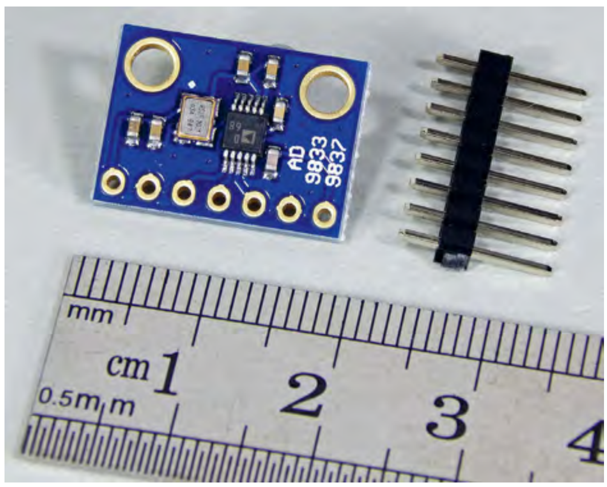
Another limitation, as noted in the *DDS in a Nutshell* box, is that the maximum output frequency is half the sampling clock frequency; in this case, 12.5MHz. But because of the way a DDS works, it can only produce a clean square wave at this maximum frequency.

If you want to get a reasonably smooth sine or triangular wave output, this will only be possible at frequencies below about 20% of the clock frequency, or in this case, a maximum of about 5MHz.

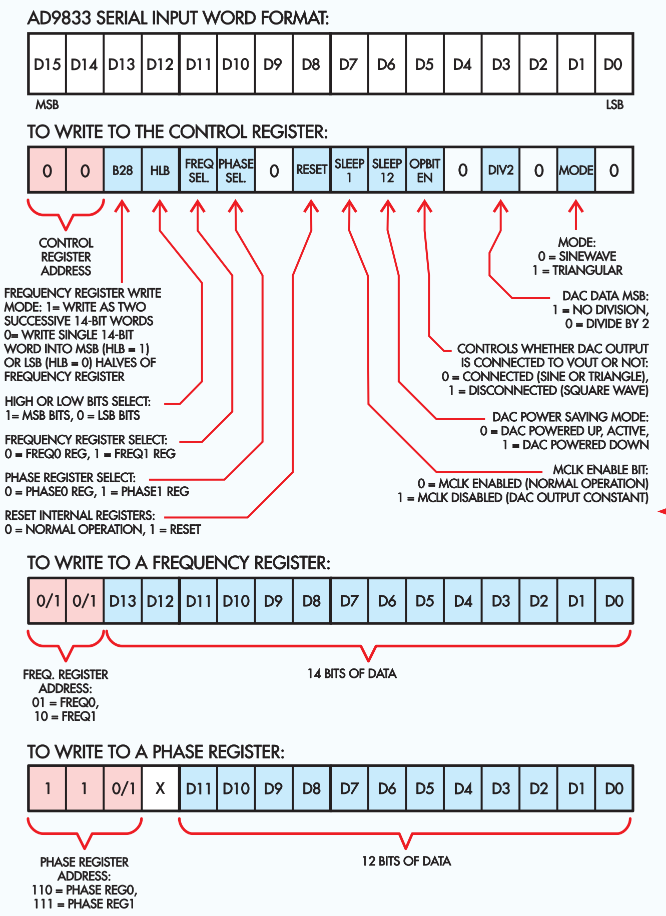
Programming it

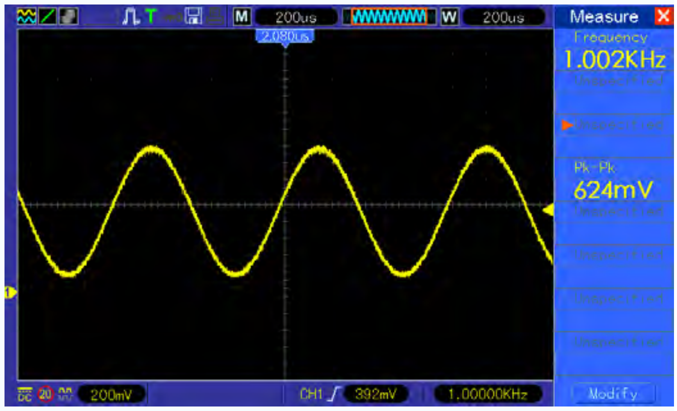
When your program starts up, it will need to carry out a number of set-up tasks. These include:

1. Declare the micro's pins to be used by the SPI interface and set them to their idle state (typically high).
2. Start the SPI interface, configured for a clock rate of say 5MHz, the data to be sent MSB (most-significant bit) first and using clock/data timing mode 2 (10 binary). If possible, it should also be set for the data to be exchanged in 16-bit words rather than bytes.
3. Send initialisation commands to the AD9833 to set up its control register, the FREQ0 register and the PHASE0 register. These involve sending the



The DDS module is shown at approximately twice actual size to provide greater clarity. From left to right, the pin connections are V_{CC}, DGND, SDATA, SCLK, FSYNC, AGND and V_{OUT}.

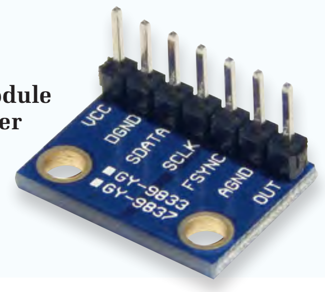




Scope 1 (above): a 1000Hz sinewave generated using an Arduino programmed with AD9833_DDS_module_test.ino

Fig.3 (left): the format of the 16-bit digital control data sent to the AD9833. The top two bits determine whether the remaining 14 bits are used to update the frequency, phase or control registers. The control register is used to change the output waveform type, switch between two different sets of frequencies and phases or go into a low-power sleep mode.

The underside of the DDS module with the 7-pin male header attached. Again, it's shown twice actual size, due to its small size (18 × 13.5mm).



following five 16-bit words (shown here in hexadecimal):

- 0x2100 (resets all registers, sets control register for loading frequency registers via two 14-bit words)
- 0x69F1 (lower word to set FREQ0 for 1000Hz)
- 0x4000 (upper word to set FREQ0 for 1000Hz)
- 0xC000 (writes 000 into PHASE0 register)
- 0x2000 (write to control register to begin normal operation)

With that, the DDS should produce a 1000Hz sinewave. To change to one of the other waveforms, you need to send the correct code to the control register. To change the output frequency, you need to send the appropriate pair of 14-bit words to one of the frequency registers. Note that these are sent lower word first, then upper word.

To make programming the AD9833 a little easier, the basic coding for the control, frequency and phase registers is summarised in Fig.3. I also have written a couple of simple example programs to illustrate programming the AD9833 module; more about these shortly. First, you'll need to know how the module can be connected to one of the popular micros.

Driving it from an Arduino

Fig.4 shows how to connect the module up to almost any Arduino or Arduino clone.

This takes advantage of the fact that most of the connections needed for interfacing to an SPI peripheral are made available on the 6-pin ICSP header that is fitted to most Arduino variants.

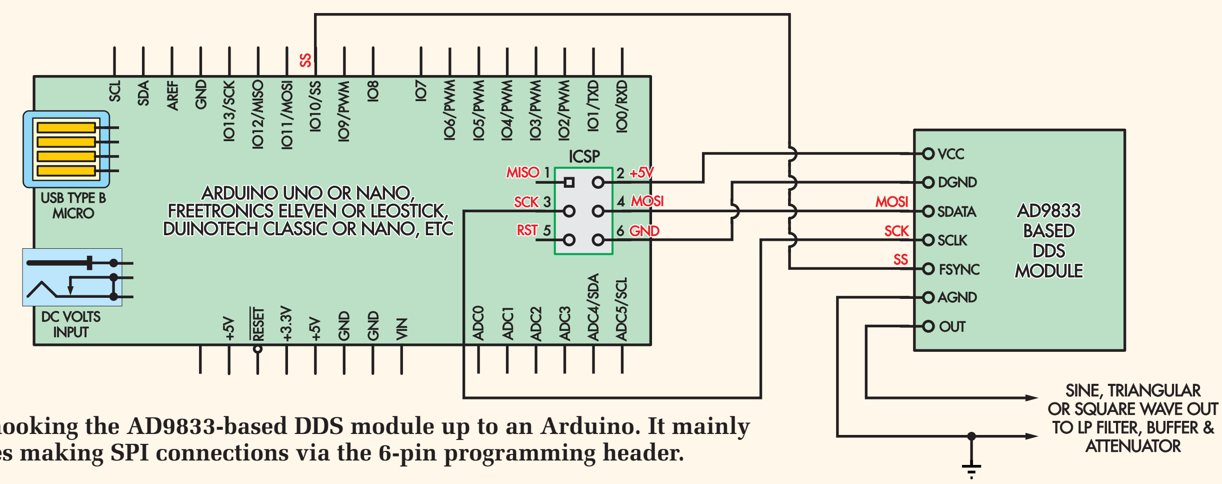
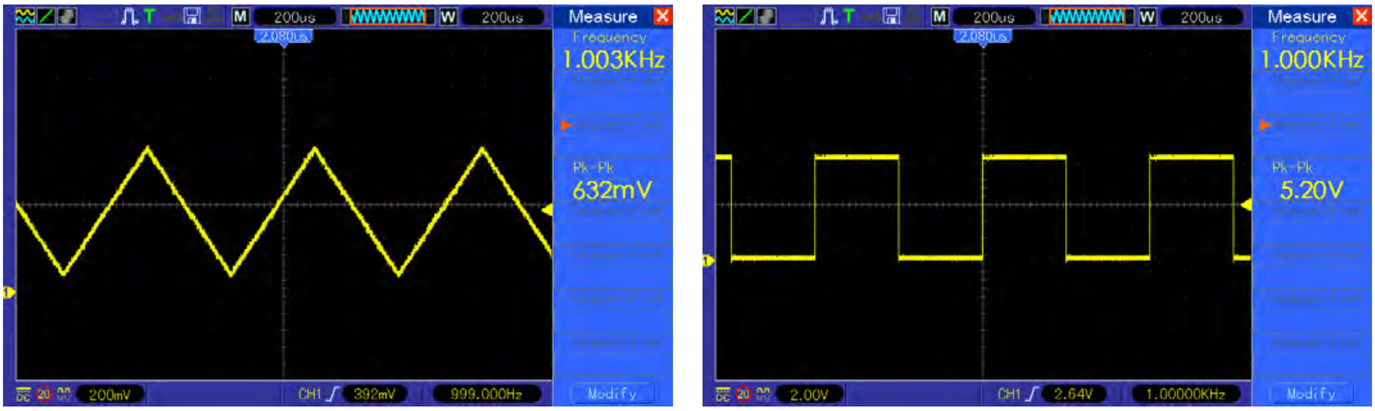


Fig.4: hooking the AD9833-based DDS module up to an Arduino. It mainly involves making SPI connections via the 6-pin programming header.



Scope 2 and 3: a 1000Hz triangular (left) and square (right) wave produced by running the AD9833_DDS_module_test.ino **file on an Arduino or compatible device. Note the higher amplitude of the square wave output.**

The connections to the ICSP header are quite consistent over just about all Arduino variants, including the Uno, Leonardo and Nano, the Freetronics Eleven and LeoStick, and the Duinotech Classic or Nano.

In fact, the only connection that's not available via the ICSP header is the one for SS/CS/FSYNC, which needs to be connected to the IO10/SS pin of an Arduino Uno, Freetronics Eleven or Duinotech Classic, as shown in Fig.4.

With other variants, you should be able to find the corresponding pin without too much trouble. Even if you can't, the pin reference can be changed in your software sketch to match the pin you do elect to use.

Arduino sample program

One of my sample programs written for an Arduino is called, **AD9833_DDS_module_test.ino**

It simply initialises the AD9833, starts generating a 1000Hz sinewave (Scope 1) and then changes the waveform after five seconds, giving you a triangular wave (Scope 2), then a square wave (Scope 3) and finally a half-frequency square wave, before returning to a sinewave and repeating the sequence. If you look at the code, you can see just how easy it is to control an AD9833 DDS module from an Arduino.

Driving it from a Micromite

Fig.5, on the next page, shows how to drive the module from a Micromite. By connecting the MOSI, SCK and SS/FSYNC lines to Micromite pins 3, 25 and 22 as shown, MMBasic's built-in SPI protocol commands will have no trouble in communicating with the module.

One thing to note is that if you want to drive the AD9833 module from a Micromite in a *BackPack* that is already connected to an LCD touchscreen, there's a small complication arising from the fact that the LCD touchscreen also communicates with the Micromite via its SPI port.

To prevent a conflict, your program needs to open the SPI port immediately before it sends commands or data to the module, and then close the port again immediately afterwards.

Micromite sample program

My other program is written for the Micromite, specifically, the Micromite *LCD BackPack*. It's called, **Simple AD9833 FnGen.bas**

This one is a little more complicated and lets you control the AD9833's output frequency as well as the waveform, simply by using buttons and a virtual keypad on the Micromite's touch screen. It's quite easy to drive, and again, should show you how the AD9833 can be controlled via a Micromite.

Both this program and the Arduino program are available for download from the *EPE* website.

Alternative module

In the April issue, we published a Micromite-based *DDS Function Generator* project by Geoff Graham, which also uses the AD9833. Geoff has used a slightly different module which includes the ability to vary the output level and also has a low-impedance buffered output.

The module Geoff has used is shown overleaf, to the left of its circuit diagram, Fig.6. It differs from the simpler module in that it has the output signal from the AD9833 fed to a separate SIL connector.

From there, the signal is routed to an MCP41010 10kΩ digital potentiometer IC via a 0Ω resistor, which acts as a digitally controlled attenuator. The output of this attenuator is fed to an AD8051 rail-to-rail op amp and together these constitute a PGA, or 'programmable gain amplifier' (not pin grid array).

The digital pot also communicates via SPI, and in fact its clock and data pins are wired up in parallel with the AD9833's, so it is on the same SPI bus.

The only difference in communication is rather than pulling the FSY pin low, as you do to communicate with the AD9833, you pull the CS pin low to communicate with the MCP41010.

Like the AD9833, the MCP41010 is controlled by writing 16-bit data words to it. For the MCP41010, the bottom eight bits of the data are the new potentiometer position, while the upper eight bits contain two command bits, two channel selection bits and four 'don't care' bits, which it ignores.

The command bits allow you to select whether you are setting the pot wiper position or commanding the IC to go into a power-down mode.

We suggest you check its datasheet for details; however, you really only need to send one of two different commands to this device when using this module:

0x11xy – set wiper position to 8-bit value xy

0x2100 – shut down potentiometer, saving power and disabling the output signal

For example, the command 0x11FF will set the output level to maximum, command 0x1180 will set the output level to 50% and 0x1101 will set it to the minimum non-zero level.

The attenuated output signal is available at pin 6 and this goes to the non-inverting input of an AD8051 high-frequency op amp, which is configured with a gain of six times, providing an output swing of around 3V peak-to-peak for sine and triangle waves. This signal is fed to both a 2-pin header connector and SMA socket via another 0Ω resistor.

Other variations in this module are that it has 100Ω protection resistors for the four control pin inputs, the

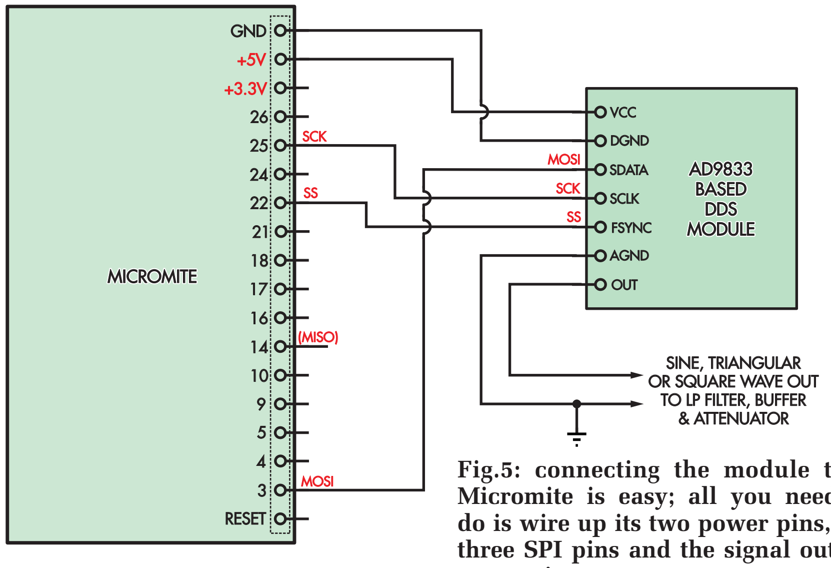


Fig.5: connecting the module to a Micromite is easy; all you need to do is wire up its two power pins, the three SPI pins and the signal output connections.

supply for the 25MHz crystal oscillator has a 10Ω isolating resistor that also forms a low-pass filter in combination with the added 4.7µF ceramic capacitor, and there is a 2-pin header which makes the output of the AD9833 available, before it enters the attenuator.

There are more details on how to use this module in April's article on the Micromite-based *DDS Function Generator*, but besides needing to program the digital pot with the attenuation value, its control is pretty much identical to the description above.

Final comments

Dan Amos has designed a 'Touchscreen Function Generator' using an AD9833 module driven by a Micromite with an *LCD BackPack*, and he has also added a digital potentiometer, an output buffer amplifier and even an incremental encoder for adjusting either the output frequency or its amplitude.

He has provided the MMBasic source code for his program, and a user manual as a PDF file – both of which can be downloaded from the *EPE* website. That project and its software is an excellent example to get you started on using the AD9833 DDS module.

One last comment – as well as being able to generate fixed frequency, FSK and PSK modulated signals, the AD9833 can also be programmed to generate swept-frequency signals.

In fact, the Micromite *DDS Function*

Direct Digital Synthesis (DDS) in a nutshell

This simplified explanation should give you some insight into how a DDS works. A DDS is based around one or more look-up tables stored in read-only memory (ROM). These contain a set of high-resolution digital samples of a single wave cycle.

Let's consider the case where the table contains a sinewave.

The values from the ROM table are fed to a DAC (digital-to-analogue converter), so that for each entry in the table, the DAC will produce an analogue DC voltage corresponding to the value of the sample stored in that address.

As a result, if a counter is used to cycle through the table entries continuously, the DAC output is a continuous sinewave.

Let's say the table contains 1000 entries which represent a single sinewave cycle and the counter which indexes the table is incremented at a rate of 1MHz. This means that the output will be a sinewave at 1MHz ÷ 1000 = 1kHz. By changing the rate at which the counter increases, we can change the output frequency.

Since the DDS chip operates from a fixed external clock, in order to vary the rate at which the DDS runs through its ROM table, a fancier counter configuration known as a 'phase accumulator' is used. This is shown in Fig.7 and it consists of a binary adder feeding an accumulator register.

The important point to note is that the phase accumulator register has 28 bits of precision while the sample table, with 4096 entries, only requires a 12-bit number to index its entries.

Hence there are an additional 16 bits of fractional phase data in the register and these effectively indicate output phase values in-between those represented by the values in the table.

The binary adder has two 28-bit inputs, one of which is the current phase value from the accumulator register. The other input comes from the frequency register at far left, also 28 bits wide.

This is the register which we use to set the DDS output frequency. At each clock cycle, the value in the frequency register is added to the value in the accumulator register and this result is stored back in the accumulator register. As a result, as long as the frequency register value doesn't change, the accumulator register increases by the same amount on each clock cycle.

With a 28-bit phase accumulator register, a value of zero indicates a

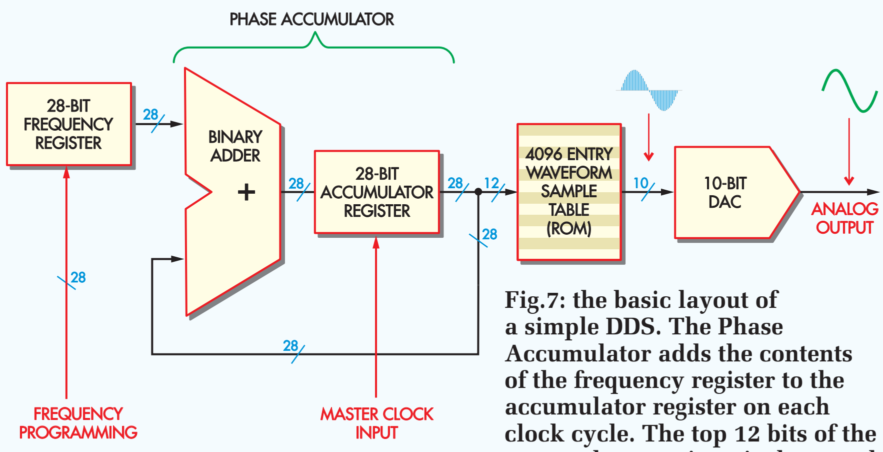
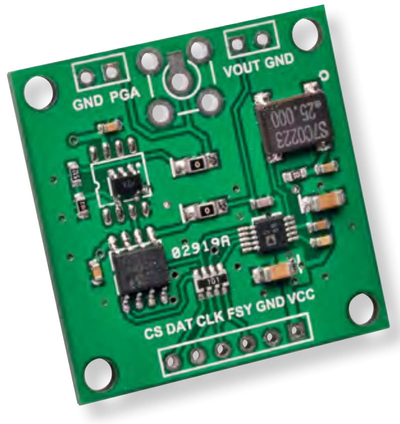


Fig.7: the basic layout of a simple DDS. The Phase Accumulator adds the contents of the frequency register to the accumulator register on each clock cycle. The top 12 bits of the accumulator register is then used to look up an entry in the waveform table ROM, producing a 10-bit digital amplitude value which is subsequently fed to the digital-to-analog converter (DAC) to generate the analog output signal.

Generator in April's issue does just that, so refer to that article for more details on frequency sweeping.



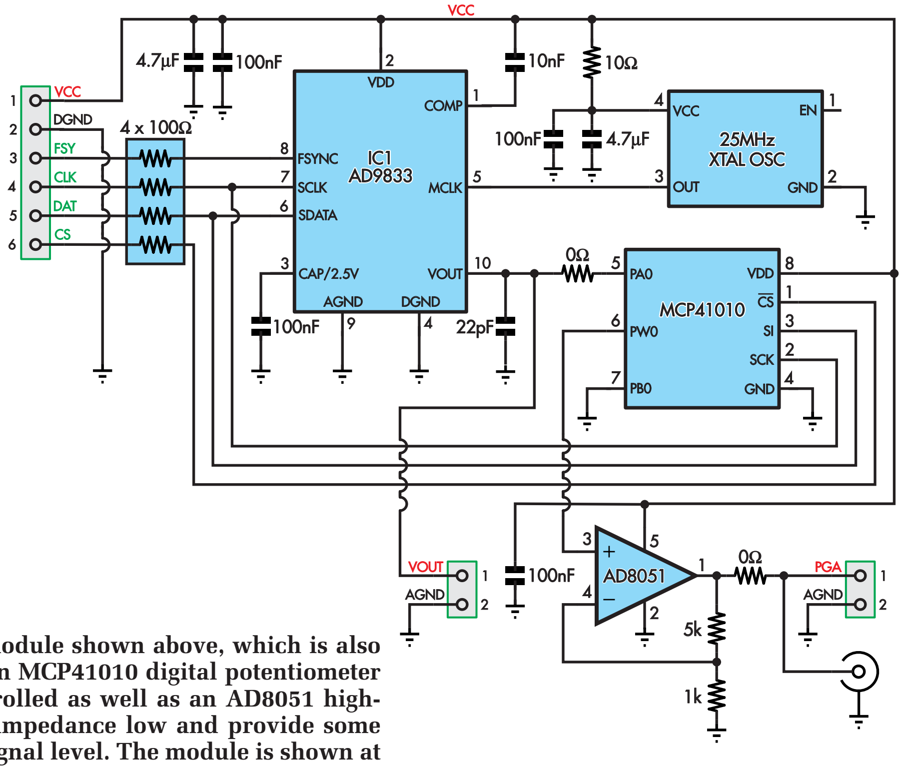


Fig.6: the circuit of the alternative DDS module shown above, which is also widely available. This one incorporates an MCP41010 digital potentiometer to allow the output amplitude to be controlled as well as an AD8051 high-speed op amp buffer to keep the output impedance low and provide some gain to allow a higher maximum output signal level. The module is shown at aproximately 1.5 times its actual size of 32 × 32mm.

phase in radians of zero while its maximum value of $2^{28} – 1$ (268,435,455) represents a phase of just under 2π.

When the value of the accumulator register exceeds $2^{28} – 1$, it rolls back around to zero, hence maintaining $0 \leq \text{phase} < 2\pi$. Each time it 'rolls over', that represents one complete cycle from the output.

With a frequency register value of 1, it will take 2^{28} clock cycles for this to happen. With a master clock of 25MHz, that means the output frequency will be 25MHz ÷ 2^{28} = 0.09313Hz or just under 0.1Hz.

With a frequency register value of 2, it will take $2^{28} \div 2$ clock cycles to roll over, giving an output frequency of 25MHz ÷ 2^{27} = 0.186Hz and so on. So the output frequency resolution with this configuration is just under 0.1Hz.

But how are such low frequencies possible with only a 4096-entry table? Well, only the top 12 bits of the 28-bit accumulator register are used to index the ROM table.

This means with the minimum frequency value of one, it will only roll over to the next entry in the table once every $2^{(28–12)}$ = 65,536 clock cycles. Hence, each value from the table is sent to the DAC 65,536 times before progressing to the next one, giving a very low frequency.

At higher frequencies, in this case above 25MHz ÷ 4096 (6.103kHz), values in the table will be skipped when necessary in order to increase the output frequency. In other words, the counter which indexes the table may increase at a rate of 1, 2, 3, 4 times per clock, or somewhere in-between, by skipping the occasional table entry.

For example, to produce an output of 12.207kHz, every second entry from the ROM table is sent to the DAC (12.207kHz = 25MHz ÷ [4096 ÷ 2]).

Based on the above, we can calculate the output frequency as:

Fo = Dph × Mclk ÷ Rac

For the DDS shown in the diagram, with a 28-bit phase accumulator having a resolution (Rac) of 2^{28} (= 268,435,456) and with a master clock frequency (Mclk) of 25MHz, this simplifies down to:

Fo = Dph × 0.09313

(We're showing the table with 4096 entries, but note that due to symmetry, it is only necessary to store the values representing a quarter of a sinewave.)

The second quadrant of a sinewave is a mirror-image of the first, so this can be achieved by running through a quarter sine table backwards, while the third and fourth quadrant are simply an inverted version of the first and second, and these can be obtained by negating the values from the first two quadrants.

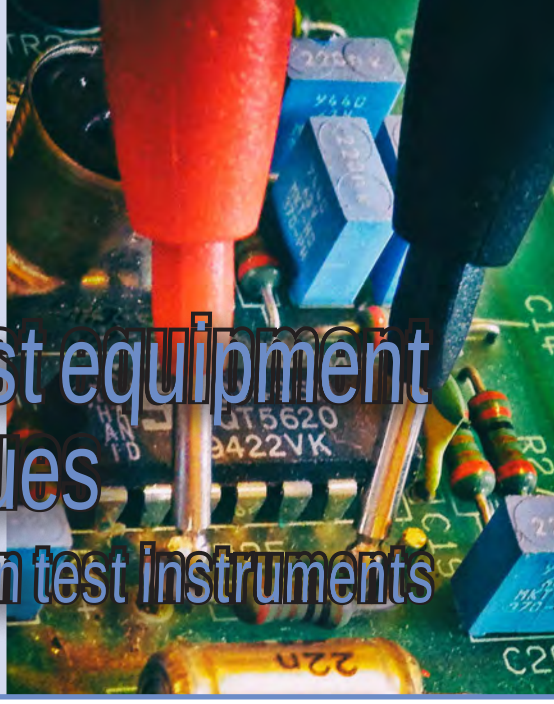
You'll also notice that the samples stored in the ROM are shown as having a resolution of 10 bits, to suit the 10-bit DAC, which can produce an analogue output with 1024 different voltage levels.

One more thing to bear in mind. Because a DDS achieves higher output frequencies by skipping samples in the waveform ROM, at higher output frequencies, the sampling resolution effectively drops.

This continues until you reach the 'Nyquist frequency' of half the master clock frequency (ie, 12.5MHz) above which the output from the DAC actually starts to drop in frequency.

So the theoretical maximum frequency for a DDS is half that of the master clock.

But in practice, because of the above, if you want a reasonably smooth sinewave output that doesn't need too much low-pass filtering, it's a good idea to limit the maximum output frequency to about 20% of the master clock frequency; say 5MHz for a 25MHz master clock.



Teach-In 2018

Get testing! – electronic test equipment and measurement techniques

Part 9: Designing and building your own test instruments

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 final part of our *Teach-In 2018* series we describe the process of designing and building your own test instruments. We also provide some ideas for developing and enhancing several of the practical *Test gear projects* from earlier in the series, and we've also included a comprehensive index to the complete *Teach-In 2018* series.

Designing your own test instruments

Provided that you are not *too* ambitious, it's quite easy to design and build your own test instruments. The process is straightforward, but should always begin with identifying a specific need that's not satisfied within your existing range of test equipment and cannot be met using off-the-shelf instruments, either new or second hand. The following eight stages are involved.

1. Identify a need

It is essential to have a clear definition of what's needed. Without this there's a very real danger that what you develop may not actually be what you require. To put this into context, here are two examples of an identified need.

Example 1: You have a digital frequency meter (DFM) that reads reliably up to 30MHz but you need to accurately measure frequencies in the VHF range. The identified need is then for a device that will bring a signal within range of what the instrument is capable of displaying. Constructing a simple divide-by-ten prescaler could easily satisfy this. With the pre-scalar inserted between the signal source and the DFM input, a signal at 45.200MHz would be displayed as 4.520MHz.

Example 2: You need to count the number of visitors to an exhibition using an interrupted light beam in which the output of a light sensor is amplified and then applied as a pulse to a simple PIC microcontroller. In order to test the operation of the system, you need to apply a series of pulses to the counter, and this could be satisfied using a simple pulse generator in which the pulse repetition frequency can be adjusted over an appropriate range.

2. Develop an outline specification of requirements

Having identified a need it is important to produce a statement of what's required in the form of an outline specification. This should take the form of a list of key parameters, together with an indication of the required performance. Note that each indicator should be stated clearly in terms of a quantity that can be reliably measured.

It can often be useful to refer to the specifications of commercially available test equipment and use those as a basis for developing your own specification. In some cases, a less stringent specification might be adequate, but in others you might need to achieve a specification that is superior to that typically provided by commercially available test equipment. There's no need to be over-specific at this stage and, unless they are critical to meeting the defined needs of your project, you don't need to include features such as dimensions, weight, and supply requirements. These can all come

Fig.9.1. The author's home-built 30MHz DFM would benefit from a pre-scaler for use at VHF

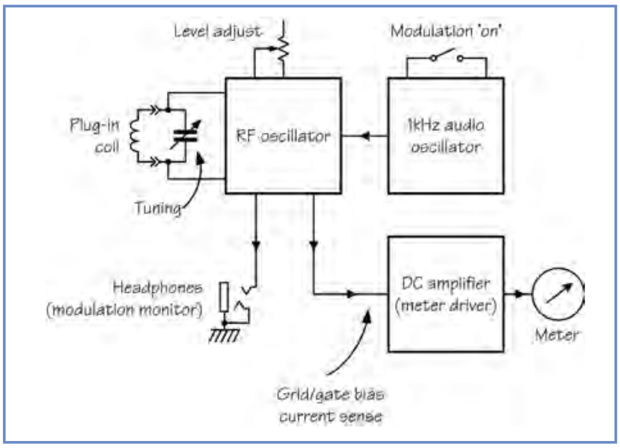


Fig.9.2. A simplified block schematic can be a useful starting point when researching different solutions

later. For the prescaler project mentioned in Example 1, an outline specification might run along the following lines:

- Pre-scale factor: divide by 10
- Input signal frequency: 10MHz to 100MHz
- Nominal output signal: $1V_{pk\text{-}pk}$ into 50Ω
- Input sensitivity: 10mV RMS over the full frequency range
- Input impedance: 50Ω
- Input and output connectors: BNC (50Ω)
- Supply: 6V internal (4 ×1.5V alkaline cells)

For the pulse generator mentioned in Example 2, the outline specification might be as follows:

- Pulse amplitude: 3V (TTL logic compatible)
- Pulse width: Adjustable from 10ms to 10s in three ranges
- Pulse repetition frequency: 0.01Hz to 10Hz in three ranges
- Output impedance: 50Ω
- Output connector: BNC (50Ω)
- Supply: 9V internal PP3 battery

3. Research alternative solutions

It's extremely rare to find that there's a unique solution to a particular design problem, and there's usually several ways of achieving the performance that you require. A great deal can be gained by researching a range of different solutions and a good starting point is reviewing projects carried out by others. Back issues of *EPE* will invariably be a rich source of ideas. An Internet search may also be useful, but can generate solutions of unproven and varying quality.

A web search for 'audio frequency signal generator circuit' will produce a wide variety of different designs (more than enough to keep you busy for several hours!). Some of these can be considered 'classic' and will have appeared in various iterations over the years. Very few solutions are unique. Others may be lacking sufficient information from which to reliably construct a prototype. However, by sifting through designs from reputable sources you should be able to pick out come common threads, which will lead you to a solution that can be built and tested. A simplified block schematic can be a useful starting point when researching different solutions.

4. Optimise and finalise the design

Armed with a technical specification and having researched a range of different design solutions you should now be ready to refine and finalise your design. At this point, it is useful to give some thought to the power supply requirements for your project. If the project is to be self-contained and portable you will need to select a suitable battery, taking into account the supply voltage, load current and the time for which the equipment will need to operate between battery replacement or recharging. Note that there will invariably be a compromise between battery size, weight, cost and service life.

Different circuit technologies impose quite different requirements and constraints on power supplies. For example, operational amplifiers often need symmetrical DC supply voltages of between ±5V and ±15V. This requirement could be satisfied using two 9V batteries or by means of a DC-DC converter operating from a 6V or 12V supply. Alternatively, if an AC mains supply is to be used, a simple full-wave bridge rectifier with split rail outputs and symmetrical three-terminal voltage regulators (eg, 7809 and 7909) would provide a workable solution.

Logic circuits, on the other hand, may require closely regulated supplies of 3.3V or 5V DC. Where mixed logic is used there's a need to ensure that voltage levels are correct and that appropriate logic-level shifting circuitry is used. This can add more complexity to a circuit arrangement but may be necessary where, for example, a 3.3V microcontroller is being interfaced to standard 5V TTL logic-level transducers and sensors.

5. Research and obtain components and materials

Having arrived at an optimised design you will now be in a position to draw up a component list, together with specifications for each component, including required value, tolerance, working voltage and power rating. You might also need to ensure that the style and footprint of the component is appropriate to the available space and method of construction. A detailed component list like that shown in Table 9.1 is important for the next stage,

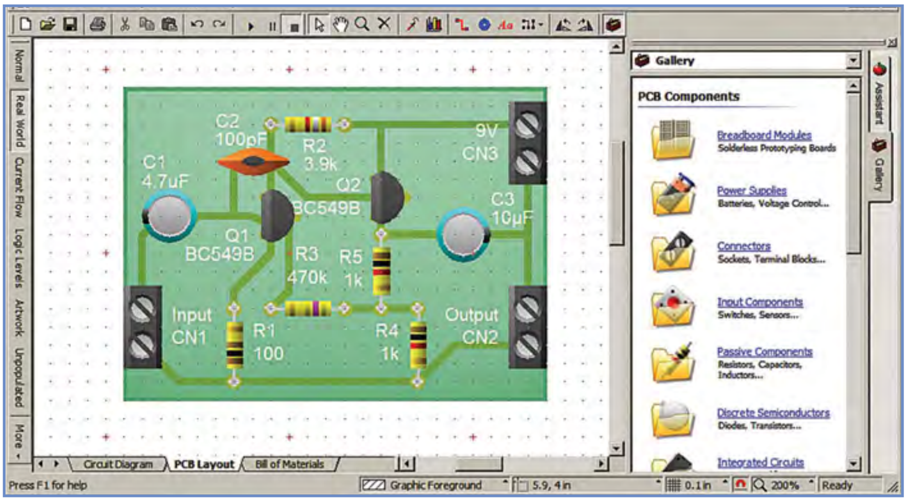


Fig.9.3. Circuit Wizard makes it easy to lay out and visualise a PCB

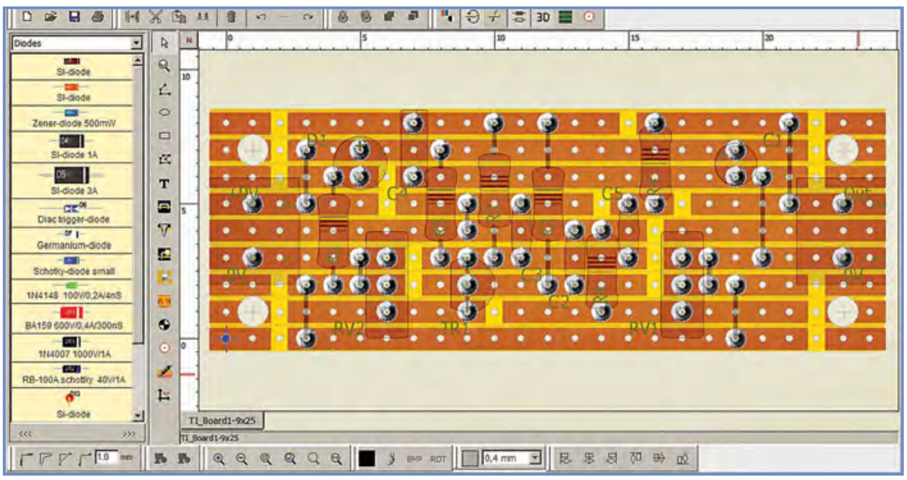


Fig.9.4. LochMaster is an excellent tool for designing stripboard layouts

Table 9.1 A detailed component list is necessary prior to the production of a PCB or matrix board layout

Reference	Specification	Package notes
D1 to D4	1N4001 1A 50V rectifier diode	DO41 package
C1	1000µ 25V axial lead electrolytic capacitor	0.2-inch pad spacing
IC1	5V 1A three-terminal regulator	TO220 package
C2 and C3	100nF 50V 10% miniature polystyrene	0.2-inch pad spacing
R1	270Ω 0.25W 5% carbon film	0.4-inch pad spacing
D5	Standard 0.2-inch dia. LED	0.1-inch pad spacing

Fig.9.5. Ensure that controls and displays are arranged logically and ergonomically with related functions grouped together

producing a physical component layout for PCB or matrix board.

As an example, a simple power supply might be based on the components listed in Table 9.1.

6. Design and construct a prototype

Several excellent electronic CAD packages are available that will allow you to test and also simulate your circuit prior to laying it out on a PCB or stripboard. The author's favourite packages include Circuit Wizard (**www.new-wave-concepts.com**) and Tina Pro (**www.tina.com**) for PCB layouts. For the stripboard layouts used in our *Testgear projects* we used LochMaster 4.0 (**www.abacom-online.de**).

Fig.9.6. A labelling machine can give excellent results. Labels can be printed in both black and white in transparent self-adhesive film

7. Test to specification

Having completed the prototype design, the next stage is testing and measuring each of the key specifications. This will invariably require the services of several items of test equipment, including a good multimeter. It may also be essential to have access to an oscilloscope and/or an appropriate signal source, such as an audio frequency signal generator, pulse generator or waveform generator for calibration purposes. In a school or college, such instruments will already be to hand but, if you don't have them available it is often possible to use virtual computer-based test instruments. Several are free to download and use. One that can be highly recommended is Christian Zietnitz's excellent Soundcard Scope available from **www.zeitnitz.eu/scope_en** (see *Part 2* of *Teach-In 2018*). After checking key specifications, it might be necessary to carry out adjustments and/or modifications to the prototype before carrying out further tests.

8. Acceptance – to what extent has the need been satisfied?

The final stage is that of reviewing the design and asking to what extent the identified needs have been satisfied. If these have only been partially satisfied, it will be necessary to examine each shortcoming and decide upon measures that can reduce or eliminate them.

Project ideas

The *Testgear projects* that we've featured in the *Teach-In 2018* series (see Fig.9.9) have all been kept extremely simple and suitable for complete newcomers to electronic construction. They are all capable of improvement, so here are some ideas that will get you started based on some more advanced projects.

1. An AC bridge

(see Part 4, Fig.4.4)

After calibration using a selection of known-value components, this simple AC bridge is capable of measuring resistance, capacitance and inductance with reasonable accuracy and it makes an excellent project if you need a basic component tester. The variable resistor (VR1) needs to be fitted with a large scale, onto which calibration marks can be placed. A useful improvement to the circuit would be fitting a sensitive DC meter movement in place of the LED indicator (D2) and wiring a variable-sensitivity control (10kΩ to 50kΩ) in series with the meter. Note that C4 to C7 and R4 to R7 need to be close-tolerance components (ideally 2% or better).

2. A transistor tester

(see Part 4, Fig.4.16)

The original circuit was designed for testing conventional PNP and NPN

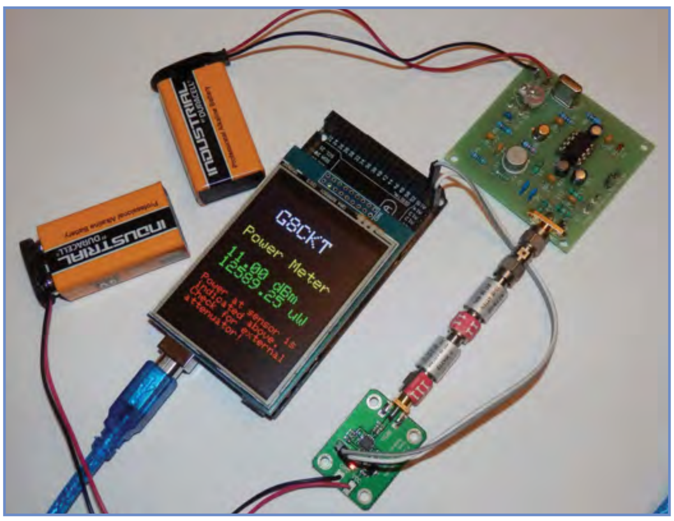


Fig.9.7. An LCD display makes it possible to display text messages and prompts

Fig.9.8. This low-cost touch screen makes it possible to have keypad entry of data on compact hand-held instruments

Get it right when designing and building test equipment

- Don't be afraid to borrow ideas from published designs. 'Classic' designs are easy to spot and are often repeated with minor changes
- Base your designs on readily available 'standard' components. Use common preferred values wherever possible
- Label inputs and outputs clearly and ensure that inputs are adequately protected against the application of excessive voltage and/or current
- Keep a folder of documentation related to your project and ensure that it's kept up to date. The folder should include items such as circuit diagrams, component lists, board layouts, wiring diagrams and specifications
- Ensure that power supplies are adequately rated and, where batteries are used, check that they have sufficient capacity to give you the required service life
- Keep a record of test results and tables of test point voltages and waveforms. These will come in handy if the need for fault finding should arise
- Ensure that controls and displays are arranged logically and ergonomically. Group related functions together and ensure that controls are appropriately labelled
- Use standard input and output connectors (50Ω BNC connectors will normally be preferred)
- To avoid misleading readings, it is often useful to include some form of over-range indication.

bipolar transistors. It measures leakage current (I_{CBO}) and also indicates the approximate value of current gain (h_{FE}). For normal use, R4 should be increased to 10kΩ and the value of R5 should be chosen to act as a shunt resistor so that the meter movement reads 10mA full-scale. To determine the resistance required you can make use of the relationship in *Part 1*. So, if the meter movement (M1) has a full-scale deflection current of 100µA and a coil resistance of 500Ω, the value of shunt resistor (R5) would be calculated as follows:

Using $R_s = \dfrac{I_m r}{I - I_m}$ gives:

$$R5 = \frac{100\times10^{-6}\times500}{10\times10^{-3} - 0.1\times10^{-3}}$$

$$= \frac{50000\times10^{-6}}{9.9\times10^{-3}} = \frac{50}{9.9} \approx 5\ \Omega$$

3. An audio frequency sinewave signal generator

(see Part 6, Fig.6.16)

This is an ideal project for anyone needing a signal source for testing audio amplifiers, tone controls and filters. The oscillator stage is based on a simple Wien bridge network with a dual-gang variable resistor providing frequency control in conjunction with two switch-selected capacitors.

The circuit includes a simple form of amplitude control that maintains the output level constant and undistorted over the full range of operating frequency. A pre-set gain buffer stage is included to minimise the effects of loading, and the output attenuator provides maximum RMS outputs of 10mV, 100mV and 1V in three switched ranges. You will also require access to an oscilloscope to set the gain control (RV1) and an AC meter to adjust the output level control (RV2).

4. A radio frequency signal generator

(see Part 7, Fig.7.13)

A radio frequency signal generator can be a useful instrument if you intend to work with radio equipment such as receivers and transceivers. To be useful, such an instrument needs to be tunable over a fairly wide frequency range (typically 100kHz to 100MHz, or more) and it should incorporate some form of modulation as well as a calibrated output attenuator.

5. A logic probe incorporating a pulse 'memory'

(see Part 8, Fig.8.5)

This is a great project for digital enthusiasts that will save you a great deal of time when checking logic levels on a wide variety of logic systems. The logic probe takes its power directly from the circuit on-test and it incorporates a pulse memory that will help you identify narrow pulses that might otherwise be missed using a basic logic probe.

Fig.9.9. The Testgear projects featured in this series have been kept extremely simple

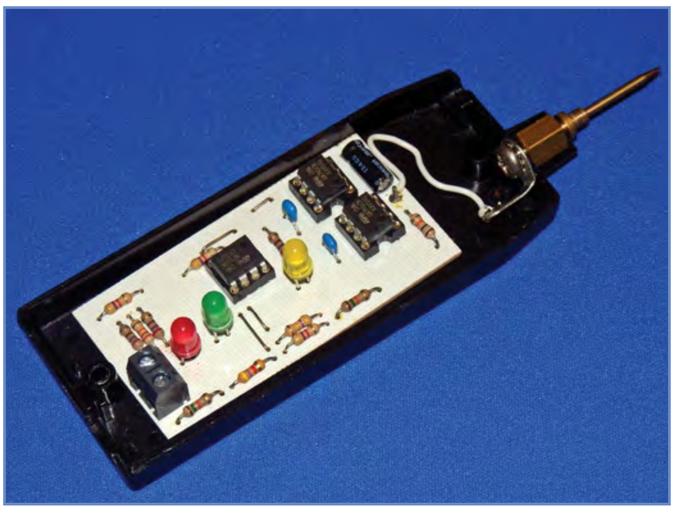


Fig.9.10. Interior of a home constructed logic probe incorporating a pulse 'memory'

6. A logic pulser

(see Part 8, Fig.8.9)

A hand-held logic pulser can often be invaluable when fault-finding digital logic. The device can be built for minimal outlay and can be used separately or in conjunction with a conventional logic probe (see *Part 8*).

The ability to force a change of logic state present at a node in a logic circuit can often reveal the cause of a stuck or frozen logic signal. Furthermore, by generating a train of pulses (rather than just a single pulse) a pulser can be used to simulate the output of digital transducers used, for example, with pulse counters, tachometers and fuel-flow sensors.

The circuit shown in Fig.8.9 produces positive or negative-going pulses of approximately 500ms. If desired, the pulse width can be changed by altering the values of C and R. The relationship between pulse duration (t) and the values of C and R is given by: $t = 1.1CR$

In conclusion

This brings *Teach-In 2018* to a conclusion. Whatever your interest in electronics, analogue, digital, audio or RF, we hope that you've found the series informative and that it has provided you with plenty of food for thought!

Teach-In 2018 index

NET WORK

by Alan Winstanley

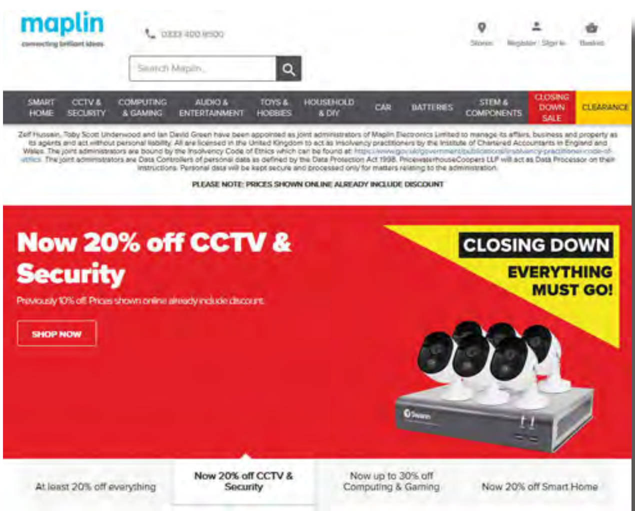
Changing times

SOME years ago I was contracted to a firm that had moved into a wharf-side office block formerly owned by a Scandinavian shipping company. The office walls displayed faded old photos depicting the site in its glory days, with the now-abandoned quay bustling with freighters, cranes, cargo and forklift trucks. I settled down to work in a forlorn old boardroom, dominated by the executive table that was now an electronics dumping ground. They say you can't put new wine in old bottles, and the office block closed down a year or two later.

Many *EPE* readers will be equally saddened to see an old brand like Maplin disappearing from our hobby electronics scene, for reasons that are well documented. At the time of writing Maplin's retail stores are bedecked with 'Closing Down' banners and their website promises 20-40% off retail prices before the shutters are finally brought down.

I visited a store back in 2014, and confirmed the saying that 'young folks think old folks to be fools' – the staffer was surprised that an old duffer like me knew to ask about NFC tags. Maplin's core market has taken flight online (its roots were in 1970s mail order), and the loss of a tech store like Maplin is a sign of the times we now live in. In a recent TV vox pop, a market stall trader complained that people can buy his merchandise online cheaper than he can buy it from a wholesaler, hence he was barely hanging on in business. Retailers in Britain are having a torrid time partly because, as one analyst said, there are just too many mouths to feed on the High Street, and also because in a commodity market, price is seemingly everything. No bricks and mortar retailer can compete with online shopping where choice is virtually unlimited and goods can be delivered effortlessly to your door at a lower cost than the High Street.

The satirical website *The Daily Mash* ran a spoof item, *Dads hold candlelight vigil for Maplin* (have a chuckle at: **https://tinyurl.com/y7hndy2o**) but today's young satirists are probably the ones who will lose out. Barring a last-minute reprieve, Maplin's closure will leave consumers with one less place to touch and feel tech goods before buying and taking them home, and a lack of choice is only ever a bad thing.



Maplin bids farewell with a fire sale

Face Off

A recent edition of the BBC TV weekly debate programme *Question Time* was held in Leeds on 22 March before an audience of people all aged 30 years or less. One audience member posed a topical question: 'Should I delete Facebook?' (BBC iPlayer has the video at: **https://tinyurl.com/yaqvcgf9**). Elon Musk, the founder of Tesla electric cars and the SpaceX rocket scientist who recently propelled into outer space his own roadster, complete with Starman at the wheel (**www.whereisroadster.com**), thought that he should 'delete Facebook', and on Twitter he allowed himself to be bounced into closing down their Facebook pages in protest at Facebook's lax privacy controls.

A BBC TV Question Time *audience debates whether or not to quit Facebook*

Question Time's young audience members had spent the best part of their lives embracing social media, but the nuclear fall-out from sharing personal details was now raining down on them. In March, Facebook became embroiled in a privacy row involving the earlier work of a university researcher and Cambridge Analytica, a data mining and profiling company whose website brags, 'Data drives all we do. Cambridge Analytica uses data to change audience behavior [sic].' Facebook stood accused of playing fast and loose with personal data after a data breach allegedly allowed some 50 million Facebook user profiles to be garnered, which may in turn have been used to target voters and possibly influence US elections, and that's before foreign countries are implicated for interfering as well. Everyone is blaming each other for Facebook's data loss and the row is still reverberating around the UK's Information Commissioner's Office (ICO), which raided Cambridge Analytica some days later. The machinations are examined on **vox.com** at: **https://tinyurl.com/y784wqbv**

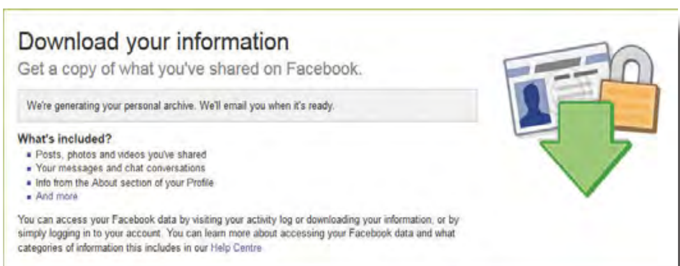
The only surprise is that it has taken so long for naive users to realise what impact the effects of data profiling can have on them. In *Net Work*, July 2004 I highlighted the way that the new Google Gmail was being tested to rival Hotmail. Controversially, it would scan keywords in your emails and pop context-sensitive adverts alongside them. While no one liked the feeling that emails were being 'read', no one seemed to care, because Gmail was free and offered unlimited storage.

It wasn't only US elections that had been harnessing the power and reach of Facebook. According to the intriguing book, *Betting the House: The Inside Story of the 2017 Election* by Tim Ross and Tom McTague (Biteback Publishing, ISBN 978-1-785-90295-6), in the UK's elections Labour (Britain's equivalent of the Democrats) quickly saw how social media was its main battleground and user profiles were therefore its artillery. With the credit reference agency Experian, Labour created 'Promote', a critical piece of profiling software designed to let Labour canvassers decide whose doors to knock on. More to the point, it allowed voters' Facebook accounts to be hooked up to their profiles and be targeted with online adverts. The quick-fire messaging app Snapchat would also prove an ideal vehicle to carry Labour's political message – perfect, because Snapchat users are typically young, earnest and impressionable. The book also revealed how Labour produced a simple digital tool that helped steer one million of its young voters to their nearest polling station on the big day.

Unknown to Labour, the book claims, the Conservatives (our GOP or Republican party) also had access to the same credit reference agency data. The Conservative Party abandoned Snapchat (wrong voter profiles!) and instead hit Facebook with nearly 4,000 different advertisements (ten times more than in the 2015 election), each tuned to appeal to the profile of the Facebook user. They could build your profile by checking your 'likes' and what websites you visit: whether the NHS, Brexit or a political party, an online store or indeed who your friends are: it has since come to light that Facebook could even capture the metadata of mobile phone call history and Messenger or SMS logs.

The controversy seems to highlight a case of the lunatics running the asylum. By running an innocuous personality quiz app on an Android device, 50 million Facebook user profiles were harvested and eventually found their way to an analytics firm. Android is, of course, developed by none other than Google, which is user-analytical to the bone, but rival Apple bans iPhone SMS data from being exploited by apps this way.

The US and UK elections were eye-openers about the way personal data could be exploited to skew the outcome. Just as supermarket shoppers' buying habits can be distorted with bogus '3 for 2' special offers, we are sleepwalking into a state where analytical systems can target you surgically with online adverts, offers, political messages and more, all highly refined to tune into your personal beliefs. As James Cleverly MP pointed out on BBC TV, it's the user who is 'the product' and each must decide how much privacy they want to sacrifice before signing up.



You can download your Facebook history via Settings; the results may surprise you

If readers use Facebook, you can download your entire log – your history, photos, videos, contacts and more – by going to **Settings** and click **Download a copy of your Facebook data** then go **Start My Archive**. Eventually, a zip file (220MB in my case) is available to download. Unzip it, then double-click **index.htm** to open it in your browser, and follow the navigation links. You may be shocked at the results!

Facebook has not emerged from the crisis smelling of roses, and it has been scrambling to re-establish its credibility and trust with users. A half-hearted 'Sorry it won't happen again' apology about the Cambridge Analytica leak eventually appeared in some of Britain's press, but the damage had been done.

Souled out

Question Time then debated whether personal data had in fact been stolen or simply given away. Users invariably agree to the T&C of Facebook, Google Gmail, Ebay, Adobe, Amazon, Microsoft Outlook and many more when they sign up, just as they would by clicking 'I agree' when installing software on their computer or tablet. It takes a lawyer to read and digest ten pages of legalese written by more lawyers, so in truth we all tend to simply accept them and carry on with our lives, otherwise we would have to lead our lives like we were running a business. To prove this point, more than eight years ago the UK retailer GameStation pulled an April Fool's stunt by inserting a clause deep in their online T&C. Customers agreed that GameStation would also own their immortal souls, but they could opt out (and earn a £5 voucher) by ticking a box. Reportedly 7,500 customers (88%) of them did not opt out and therefore had technically sold their souls to the retailer.

We all sign up without giving it much of a thought, hoping the supplier is trustworthy, especially when time is short and, besides, eager buyers have little choice anyway. However, it was encouraging to hear several *Question Time* audience members agree that profile exploitation was ultimately the user's own fault, and they should take responsibility for the consequences of processing such data that came with the service they had agreed to and signed up for, especially free ones like Facebook and GMail.

The founder of the principles of the 'world wide web', Sir Tim Berners-Lee complained in the web's 29 birthday address in March how it had fallen prey to a few dominant brands, 'allowing a handful of platforms to control which ideas and opinions are seen and shared.' He recognises how the web has been 'weaponised', how data is being stolen and even elections are being 'interfered with' (a reference no doubt to foreign states planting influential material online). The text is available at: **https://tinyurl.com/ycup5rc8**

PURE Evoke lives again

Regular readers will recall my praise for the PURE Evoke Flow Internet-based radio. Sadly, my own radio became useless when its OLED display failed, a common problem that renders the radio inoperable. It seems the organic element simply wears out. I had hoped this problem could be fixed when an enterprising individual in the UK sourced replacement displays and offered them for sale, enabling a DIY repair for displays in PURE Evoke Flow, Avanti Flow or Evoke-1S Marshall DAB radios. Unfortunately, I've now been told the vendor was being sold untested lots and the 75% failure rate made it too costly to sell; further details here: **https://tinyurl.com/yc7puv68**. Despite this setback, the author is keen to hear other, viable repair ideas.

That's all for this month's *Net Work*. You can contact the author at: **alan@epemag.net** – and don't forget to visit the **EPE-Magazine** section of **EEWeb.com** where you can share queries with like-minded electronics enthusiasts and engineers of all abilities. To ensure it is not overlooked, please remember to also 'tag' your query with 'EPE-Magazine' (it will autocomplete), which will route your message to our section of the EEWeb forum.

PIC n' Mix

Mike Hibbett

Our periodic column for PIC programming enlightenment

Practical DSP – Part 3

In this third article in the series on DSP (digital signal processing) techniques, we look at taking our reference design from last month's *PIC 'n Mix* article to a practical application – an acoustic guitar tuning aid. This project aims to take a 'clinical' DSP reference design, which used hard-coded 'fake' audio signals held in a source file and complied into the application, and construct a real, useful project.

We are expanding on the hardware and software design introduced last month, but don't worry if you missed that episode as the details are provided in full within this article, and the source files are downloadable from the magazine website. So, let's start by reminding ourselves of the problem statement, what we aim to achieve.

Using a cheap electret microphone coupled to our microcontroller we will 'listen' for the sound of a guitar string and turn on an LED when it is correctly tuned – a very simple guitar tuner, but one which can be easily expanded with added sophistication. The 'A' string of a guitar is normally tuned to 110Hz, so we will look for the peak frequency detected being at that frequency, plus or minus a few hertz. We have most of the components we need assembled from last month's article, all we need to do now is add a microphone and connect it to the ADC input. Then it becomes a software integration activity – enabling the ADC to take periodic samples, formatting them correctly for the FFT (Fast Fourier Transform) function, running the FFT function and then detecting the peak frequency.

Our FFT algorithm expects 512 samples. We will sample this data at a rate of 2kHz – allowing our FFT to resolve signals up to 1kHz. Recall from last month that this means our FFT will output an array indicating the magnitude of the signal at different frequencies, with the resolution of each entry being 3.9Hz. That means we turn the LED on if the 29th index into that array of magnitude values is the highest.

Once again, we creep up on this solution slowly, as there are several different peripherals to deal with, all of which are new, because we have not used this processor family before. The order of progression allows us build up each part of the design without relying on the previous step, so each step can be verified by itself. This minimises the number of variables we must deal with when working out problems during development. 'Change one thing, then test' is a valuable lesson we have learnt over the years – in other words, avoid building upon foundations of sand.

It helps to have all the relevant technical information to hand when beginning development, and we've listed these in the *Reference* section at the end of the article. It's vital when starting on a new microcontroller to look at the errata sheet too, listed in Ref. 4. All processors have errata sheets, documents that describe how the device deviates from the datasheet due to design errors or performance limitations detected when the chip has been manufactured. Normally, these errors are minimal enough that the chip can still be sold; more than once, however, your author has selected a device only to find a critical feature is not working. Always read the errata!

Now we are ready, we can start building up the design. We are going to do so in the following order:

- Microphone circuit set-up
- dsPIC op amp configuration
- dsPIC ADC configuration

Microphone circuit set-up

Let's get started and take a look at the microphone. We are using the cheapest electret microphone we could find, Farnell part number 2066501, shown in Fig.1. It's tiny, and the wires more so – we had to solder stronger solid core wire to them, so the microphone could be connected to the breadboard. The datasheet for this device can be found in Ref. 6.

The construction of electret microphones makes them very high impedance devices, and most are supplied with a JFET transistor integrated into the case on a tiny PCB. Output is about 6mV peak to peak, and they draw about 500µA. Such a

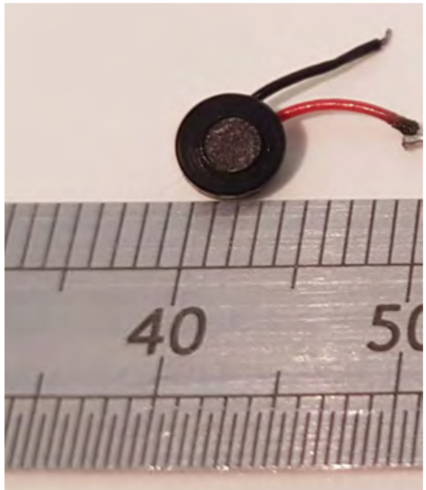


Fig.1. Electret microphone

low-output signal is no use to us; we want the signal to have a peak-to-peak voltage close to the range of the ADC, 0 to 3.3V. So, we will need something like an op amp to boost the signal.

There are many examples of op amp electret interface designs on the Internet, almost all of them associated with the message 'why does this circuit not work'. In this column, we're not analogue experts, so let's go to a company that understands this problem – Texas Instruments. They have provided an excellent application note for designing op amp circuits with electret microphones, listed in Ref. 5. Luckily for us, the electret microphone they use is very similar to ours. The design calls for the microphone signal to be connected to the inverting input of an op amp through a DC blocking capacitor, and then a resistor with a small capacitor connected from the output forms the feedback. A DC block capacitor is fitted on the output, but we can ignore this, as we will be coupling the signal directly to the ADC input. The recommended circuit values must be changed due to our slightly lower performance microphone and a lower supply voltage, but the application note provides clear details on how to calculate these values.

The next question is, 'what op amp shall we use?' We're not looking for a Hi-Fi design, so we don't need a high-

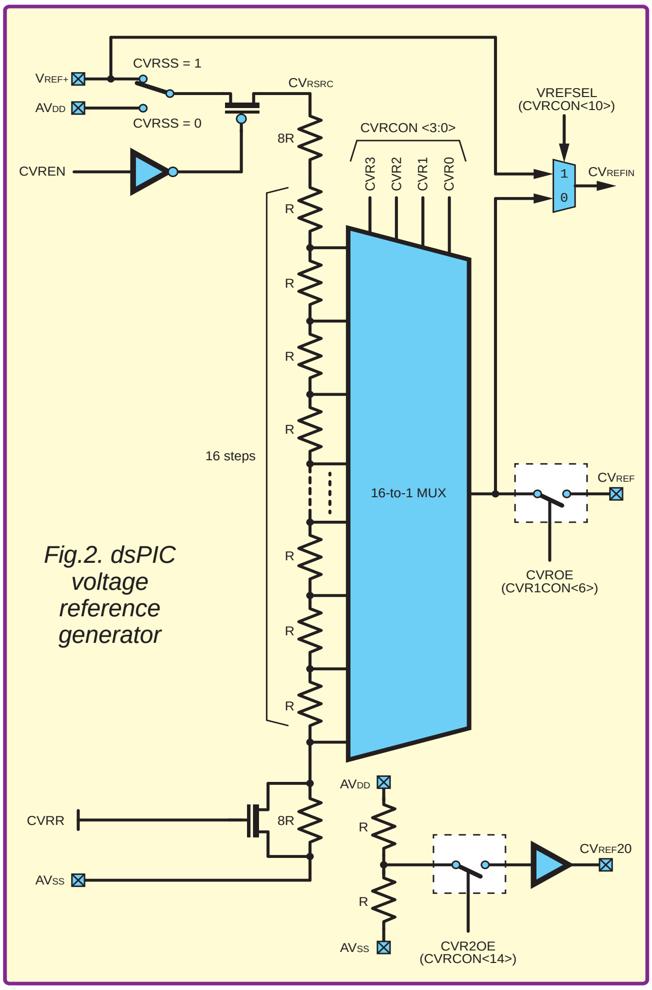


Fig.2. dsPIC voltage reference generator

performance op amp – virtually any device would do. But wait – do we need an additional IC at all? The dsPIC has an op amp peripheral built in. Not only that, it has a reference voltage generator, which can provide the mid-rail bias voltage required for the non-inverting input to the op amp. This will save us two resistors and a capacitor!

So, that's our microphone circuit identified. Let's take a look at how the op amp needs to be configured to complete the circuit.

dsPIC op amp configuration

An op amp is typically followed by an AC coupling circuit to an amplifier, as shown in the Texas Instruments applications note, but with our mid-rail bias supplied by the on-chip voltage source, we can feed the op amp output directly into the ADC. The dsPIC ADC converter, as already mentioned, supports the signed fractional data type required by the FFT function, and so the signal is already correctly conditioned for sampling by the ADC.

Biasing the op amp to mid-rail enables us to make use of the ADC's signed fractional conversion capability (yes, we did look that up first!) which sets the 'zero volt' reference point to mid-rail of the ADC input. Although our signal is not going to be perfectly centred around mid-rail, that will not interfere with the FFT function's ability to resolve the frequency components in the signal. Another advantage of the DSP algorithm. This could be achieved by using two resistors, plus a capacitor to remove noise, but another integrated feature of the dsPIC can be used to achieve this signal – the voltage reference circuit within the op amp, shown in Fig.2. This block provides a programmable voltage reference on pin CV_{REF10}, or a fixed mid-rail voltage on pin CV_{REF20}. We will go with the former, as the ability to adjust the op amp bias voltage may be useful in the future. Although the diagram suggests that this voltage can be routed internally via the CV_{REFIN} signal to the op amp, reviewing the op amp configuration pages shows that this signal is not available when the peripheral is configured as an op amp. Strange, but we must follow the rules – we will simply connect the CV_{REF10} pin (pin 17) to the op amp non-inverting input C2IN1+ (pin 3.) Note that we are using the second op amp within the peripheral, not the first. This is because the input pins for the first op amp are shared with the programming pins – which would have made debugging impossible. Thankfully, that is not an issue for us, as there are two op amps using different pins. Routing the reference voltage off-chip and back in again, using just a simple wire connection, does benefit us – it means we can measure the voltage on the wire to know that we have configured the pins and the reference voltage correctly.

The inverting input C2IN1– is on pin 4 and the op amp output OA2OUT is on pin 2. We wire the feedback resistor, two capacitors and the microphone supply resistor to these two pins to complete the circuit. The final schematic is shown in Fig.3, with the completed breadboard layout (now a little less tidy than last month) shown in Fig.4.

Configuring the op amp is complicated, because there are

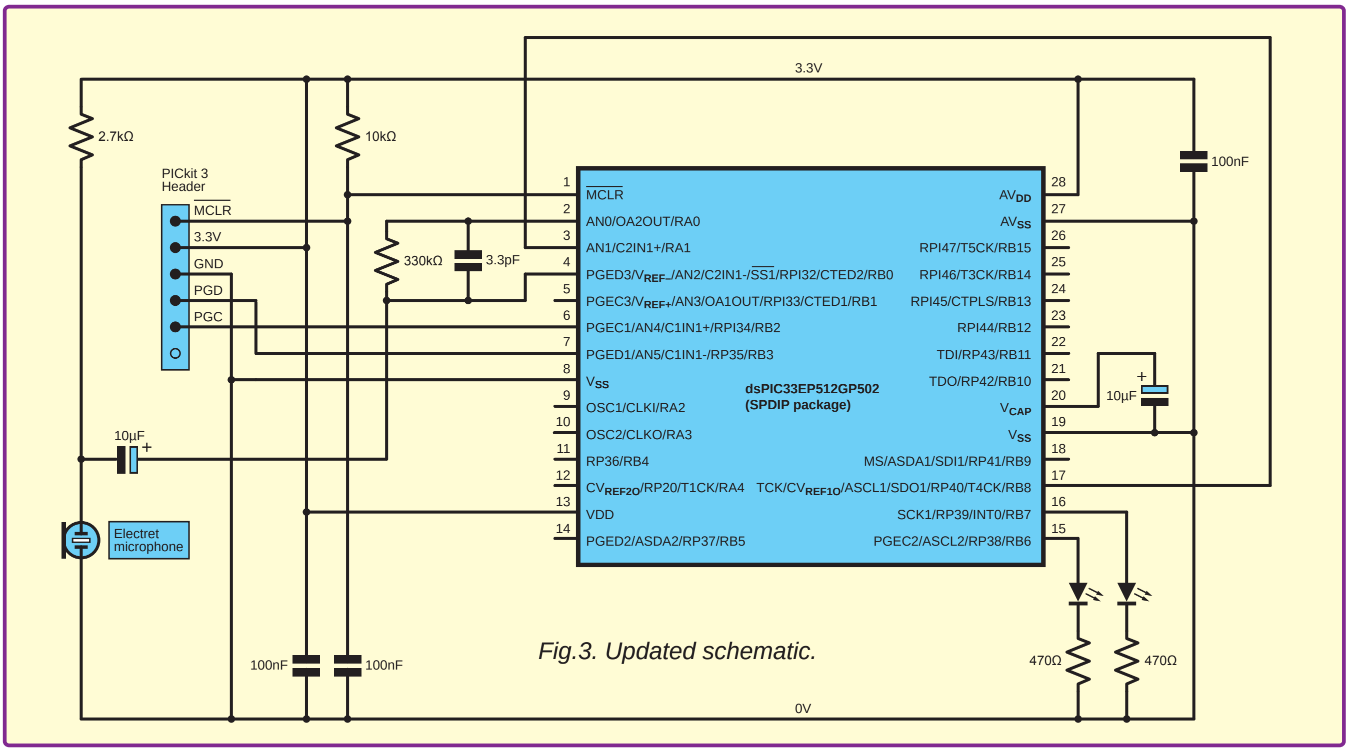


Fig.3. Updated schematic.

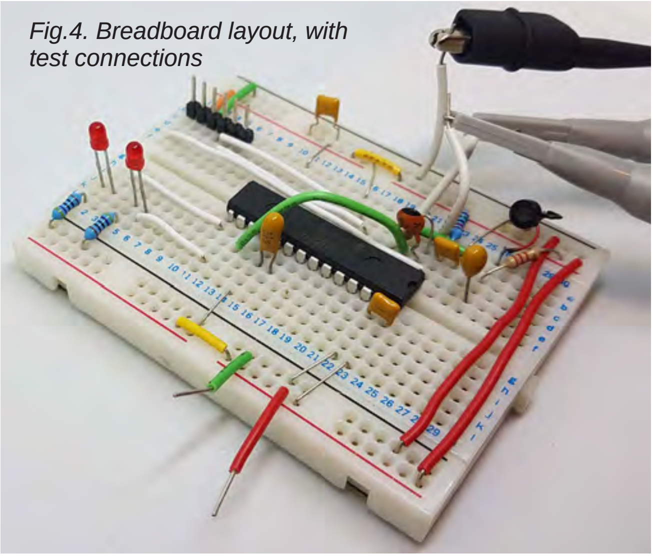


Fig.4. Breadboard layout, with test connections

a number of things we need to do in order to have the op amp circuit working at all. It's worth listing these so that we can go through them step by step, ticking them off so we do not miss anything:

- Configure the voltage reference to mid-rail
- Configure the voltage reference pin to be a voltage reference, not a GPIO
- Configure the C2IN1–, C2IN1+ pins to be op amp inputs, not GPIO
- Configure the second op amp to route the output directly into the ADC
- Configure the OA2OUT pin to be op amp output, not GPIO

You might be wondering about the last one – why output the signal onto a pin, when we are routing the signal internally to the ADC? The answer is simple, it's to allow us to see the signal with an oscilloscope so that we can confirm the input signal is being correctly amplified. This is far easier than looking at ADC data.

We start by configuring the CVRCON register, described on page 371 of the datasheet, using this single line:

```
CVRCON = 0x00EC;
```

We also must set TRISB bit8 to be an input, because despite the datasheet saying the opposite, you have to configure the GPIO pin shared with the reference voltage output signal as an input for the reference voltage to appear on the pin. That didn't appear in the errata sheet!

Configuring the C2IN1 and OA2OUT signals turns out to be straightforward too – a single line of code configures the op amp appropriately:

```
CM2CON = 0x8400;
```

Again, we modify the TRISB register so that the RB0 pin, shared with C2IN1–, is configured as an input. There are no additional changes required to configure the C2IN1+ pin, because this is shared with port RA1, which we had left in its default state of input. Unusually, we are not required to turn off the ADC peripheral selections on the op amp pins; enabling the op amp appears to override the pin selection. We tested this anyway, but writing zeros to the ANSELB register did not change the behaviour of the circuit. Best practice dictates that all unused pins be set as a digital output, which we do. (Leaving a processor configured as an input and leaving it unconnected will generate additional noise in the system and unwanted increased power consumption as the input signals float to mid-rail, rapidly turning transistors on and off briefly due to unavoidable noise.)

Signals and noise

With the above code changes built and running on the breadboard, it was time to do a quick check to see if everything is working. Sure enough, a voltage of approximately 1.7V appeared on the CV_{REF10} pin, and it was with some delight that when monitoring the OA2OUT pin while playing a 440Hz tone over the PC speakers, the oscilloscope trace shown in Fig.5 was recorded. As mentioned before, this is not high-fidelity audio and it is no surprise to see a lot of noise, but the signal is clearly that of a 440Hz sinewave and spanning a good part of the voltage range. Exactly as predicted. Thank you, Texas Instruments, for an excellent application note!

Including an internal op amp with the microcontroller has saved us adding an op amp IC, with all the wiring of power rails and signals which that would have entailed. Instead, we have just two resistors and two capacitors required to couple the microphone to our processor. Great work Microchip!

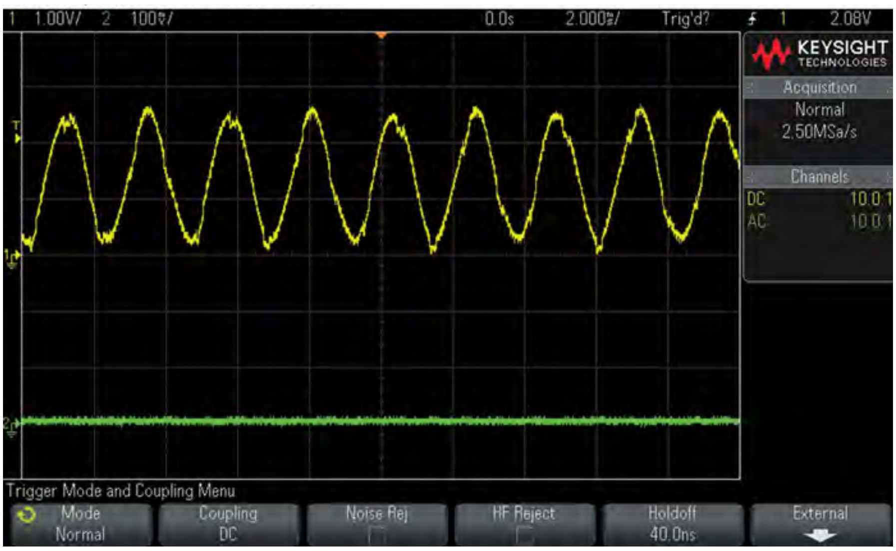


Fig.5. Op amp output signal and input signal

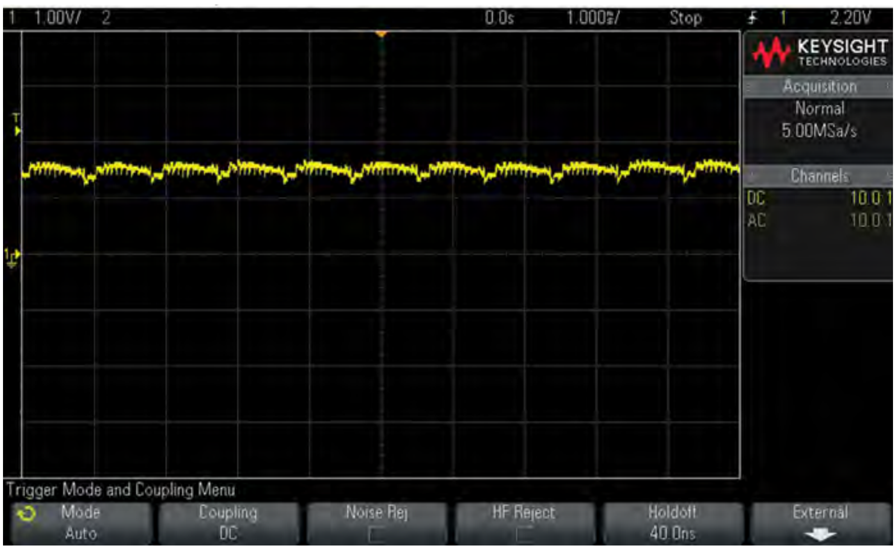


Fig.6. Op amp output with CPU running, with no input signal

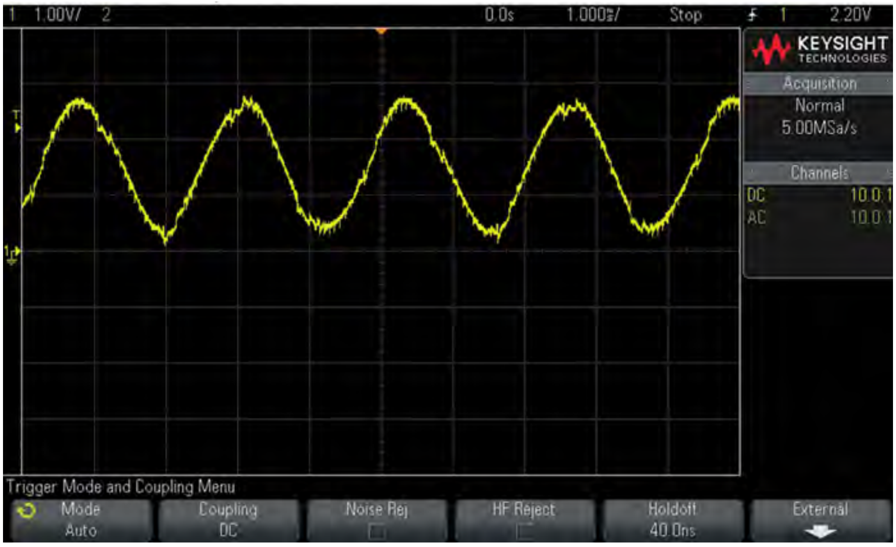


Fig.7. Op amp output with CPU running, with signal present

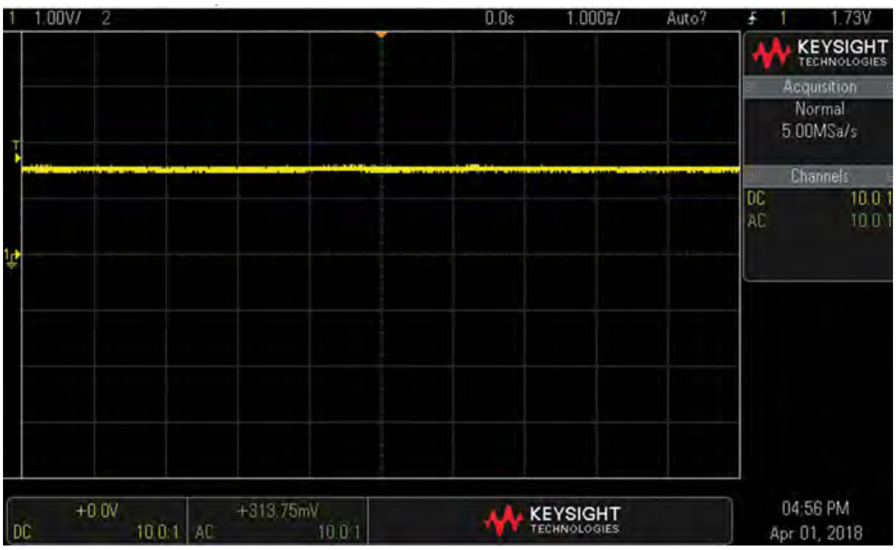


Fig.8. Op amp output with CPU halted, with no input signal

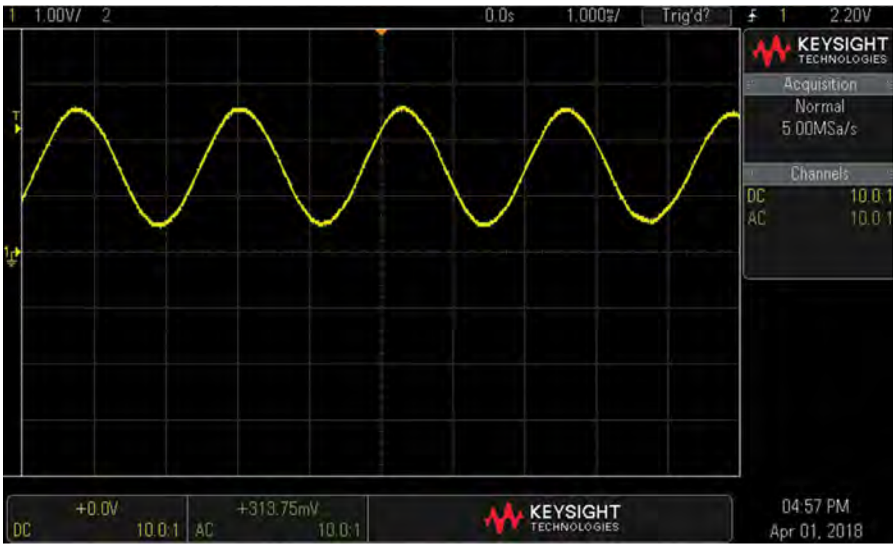


Fig.9. Op amp output with CPU halted, with signal present

Finally, for this phase of the design, we look at how to route the OA2OUT signal (the output of the op amp) directly to the ADC internally within the device, to save us adding a further wire. No mention is made of it in the op amp peripheral section of the datasheet, nor the family reference manual documentation. Sure enough, that is handled within the ADC peripheral – so we will cover that next, as we look at the ADC.

Before we move onto the ADC part, let's take a closer look at the signals around the op amp. The trace shown in Fig.5 is noisy, which is to be expected because we had the PICkit 3 debugger connected and providing power. What do the signals look like under a more representative scenario when the debugger is disconnected, and we are running from batteries, with no audio signal present? Fig.6 shows this. You can clearly see a repeating waveform with a period of about 1ms, or 1kHz. With a strong audio signal present, shown in Fig.7, the distortion is evident. Where is this coming from?

Looking around the lab gave no obvious clues to external interference, and the LED room lights – a common source of interference – were turned on and off to no effect. Then it dawned on us – the 'test' FFT function, implemented in the code in last month's article, was still running. Adding an `Idle();` command at the end of configuring the op amp will stop the processor from executing any further CPU instructions. The result of adding this is shown in Fig.8 and Fig.9, and show a significant improvement in our op amp output signal. Clearly, with low voltages being amplified by an op amp one needs good power and signal routing to reduce noise coupling into the circuit, and a breadboard is not ideal for that – our processor operation is, naturally, generating its own noise. Thankfully, this will not impact us – the ADC peripheral has been designed with this issue in mind and can perform signal sampling and conversion while the processor is in Idle or Sleep mode. We will make use of that feature when setting up the ADC peripheral.

dsPIC ADC configuration

We now reach the complicated stage of the project – getting ADC readings from the op amp, at the correct sample rate and in the correct format. As mentioned before, the ADC supports outputting the voltage measurement in a signed fractional format, matching our FFT's input requirement. This is great, but our task is to correctly configure the ADC to do so, and that's where the challenge lies. The ADC module is highly configurable, and we must navigate our way through the ADC peripheral documentation – in the datasheet and the more detailed family reference document listed in Ref. 2.

ADC peripherals have always been complex, owing to the flexibility demanded by embedded engineers pursuing different solutions. Some designs require many different voltage sources to be read, but slowly – which drove microcontroller manufacturers to introduce multiplexers at the input to a single ADC peripheral, allowing one of several pins to be connected under program control. Other solutions require fast sampling, or operation while in a very-low-current-consumption state. In most cases, the same ADC peripheral would do the job, but bells and whistles were added over time to meet multiple needs, and we now have ADC peripherals that make undergraduates cry (well, my students do!)

So once again, we write down our requirements clearly, individually, and in a list so we can hopefully attack each one in turn, breaking a complex problem into a series of simpler problems:

- Configure the ADC into 12-bit mode
- Select the op amp as the input source
- Configure the data conversion mode to signed fractional
- Sample, hold and take a single value – while keeping the processor in halt state.

We will later use a timer to take repeated samples at the interval we require and use this timer to manage the scheduling of the FFT transformation once 512 samples have been recorded. Using a timer this way means we can also do the sample modification – dividing by two and zeroing the imaginary component – as we are receiving each sample. This will speed up the FFT conversion, recorded last month as under 1ms, which will increase the rate at which we can take measurements and update the display (our two LEDs in this case.)

Next month

That's enough for this month. Next month, we will complete the project by enabling the ADC peripheral and passing the data into the FFT routines, illuminating the LED when a guitar string is correctly tuned. We will also look to integrate a graphical LCD display, so we can see the full range of our audio signal and consider ways to expand on this project in the future. Finally, we look at the performance of the FFT algorithm on some pre-generated noisy input signals to see just how powerful the FFT algorithm is at detecting signals in the presence of noise.

The updated source code for this month's progress on our project is available for download on the *EPE* website – navigate to the current issue's page at: **www.epemag.com**

Looking further ahead

If you think this is a useful design, from which you want to build an actual guitar tuner, here are some thoughts (based on product development experience) to help guide you taking this further.

Is this the kind of user interface you want? This design shows when you are close to the correct frequency, but not whether you are above or below it. Perhaps another LED or three would make sense to enable you to indicate 'above/below correct frequency'. Would it be better to have a graphical display showing the current frequency verses the desired frequency? That would make more sense, although it would complicate and increase the cost of the design. We cover the integration of a graphical display in the next article.

Can you put this breadboard into an enclosure and bring it to concerts? No, a breadboard is only good for development on a bench – and even then, as we have seen, when dealing with very low voltage signals, a breadboard layout encourages noise to enter the input signals. If you want a robust solution, you are going to need to build it up on a PCB – either veroboard (which is OK, but tedious to build more than one) or design your own PCB if you want several copies, or to share the design with friends.

Does this need to be robust? If the software crashes – either through a software bug (surely not!) or electromagnetic interference – do you want the device to automatically recover, or require a power cycle? If the former, then implementing a watchdog solution in software would be a good idea.

Does the device need to be more accurate than this design? If so, you need to look at taking more samples (thus increasing the size of the FFT input from 512 samples) or lowering the sampling frequency.

Are you interested in applying FFT techniques to other problems? If you do, then please reach out to us on the forum – we would love to hear your ideas. The application of DSP algorithms to real-world problems is wide, and a Microchip solution can be applied to many. Measuring vibration from a rotating drive shaft for example or filtering out a specific tone from a noisy audio signal are just two examples.

For an example of just how far DSP algorithms can take us, have a look at this YouTube video: **www.bit.ly/2JciKbe**. It shows the decoding of radio amateur signals on a PC. The display that you can see on the screen, called a waterfall, is nothing more than an FFT – but covering 300kHz of the 14MHz amateur radio spectrum, rather than 1kHz of the audio spectrum. The FFT is running on a fast personal computer; so don't be expecting similar performance from our tiny microcontroller!

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Chopper and auto-zero amplifiers – Part 1

CIRCUIT SURGERY articles are usually inspired by posts on *EPE's* online forum (previously the *Chat Zone*, but now hosted on EEWeb: **www.eeweb.com/forum/**). However, we also cover other topics which will be of interest to readers. This month, I will discuss 'chopper' amplifiers, a topic suggested by *EPE's* editor Matt Pulzer. These circuits are typically available as operational amplifier chips and can be used in many of the usual op amp circuit configurations. Of course, there is something special about them – they exist to deal with a particular problem, which is the difficulty of high-precision amplification of low-frequency signals, with 'low frequency' typically meaning a few hertz and below. Depending on the application, these amplifiers may also be required to handle much higher frequencies, adding to the difficult of creating suitable circuit designs.

Low-frequency signals typically occur in certain sensing applic-ations, where the measured signal changes slowly or contains slowly changing components of importance. Applications requiring high-performance amplifiers at low frequencies include high-precision temperature measurement, electronic weight scales (and other strain gauge systems) and medical equipment sensing electrical signals in the body (eg, ECG and EEG). The design problem is the low-frequency noise and offsets which are inherent in electronic amplifiers, and which render ordinary op amps inadequate for these tasks.

The solution is not simply to design a better conventional amplifier – the unwanted low-frequency signals have to be removed by more sophisticated means, which is in effect a form of calibration. Thus, amplifiers of this type are referred to as 'auto-calibrating', 'auto-zero', or 'zero-drift' amplifiers, as well as 'choppers'. The term 'chopper' refers to the fact that these amplifier make use of switching circuits. There are actually two fundamentally different approaches to implementing these circuits, which can lead to some confusion over the use of the term 'chopper', which is more often associated with one particular method, although both use switches and both achieve an auto-zero functionality. We will discuss the circuit designs in more detail after briefly looking at the problems they deal with and some basic theory needed to understand their operation.

Offsets

A key issue with low-frequency amplifiers is offset. In simple terms, offsets are DC errors due to imperfections in the circuit or components. If an offset was pure DC (zero frequency) it would be, by definition, fixed for all time, and so could be removed by a one-off calibration procedure. But real offsets drift due to changes in temperature, aging and other factors that influence the circuit. These changing offsets are just like low-frequency signals and so get amplified along with the wanted signal – they act as low-frequency noise.

Ideally, with a differential input of zero, an op amp's output should also be zero, but in real op amps there will typically be a non-zero output. The input offset voltage, V_{IO}, is defined as the DC voltage that must be supplied between the inputs to force the quiescent (zero-input signal) open-loop (no feedback applied) output voltage to zero. The definition of input offset voltage is illustrated in Fig.1.

The input offset voltage is defined with respect to the input. The error in the output voltage due to V_{IO} is equal to the circuit's noise gain times V_{IO} (note that noise gain relates to *circuit* gain, not op amp *open-loop* gain). For example, if the datasheet quoted V_{IO} as 500μV max, and a noninverting op amp amplifier had a gain of 500, you could get a 250mV (quarter of a volt) error on the output. Temperature often has a significant effect on offsets – the temperature coefficient of input offset voltage specifies how V_{IO} changes with temperature. The datasheet for an op amp may also have a graph showing offset variation with temperature.

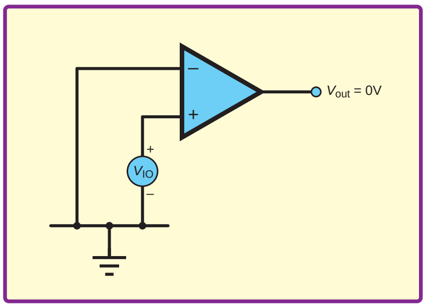

Fig.1. Offset voltage defined

Spectra

When discussing circuit properties that are related to particular frequencies or ranges of frequencies, it is common to use graphs of signal strength versus frequency – the signal's spectra. The frequency axis of a signal spectrum may be linear or logarithmic, with logarithmic axes used when a wide range of frequencies is considered.

Signal strength may be plotted as voltage or current, but is more usually expressed in terms of power or power density, usually with logarithmic scale (typically in decibels). A spectrum is built up from a set of arbitrarily narrow frequency bands. It follows that if we change the width of the band the power in each band also changes. Use of power density, or power spectral density (PSD), avoids this problem. PDS is measured in watts per hertz (W/Hz), but can also use the square root of this (corresponding to the root-mean square (rms) voltage), measured in volts per root hertz (V/Hz^{½} or V/√Hz). However, when spectra are drawn simply to show the general shape of the signals at different frequencies, rather than providing specific numerical data, the units are of less importance.

The sinewave has the simplest spectrum, just a single peak at one frequency. Other simple periodic waveforms, such as square waves, have spectra with peaks at specific sets of individual frequencies. Complex, meaningful waveforms, such as voice signals, contain a wide range of different frequencies, but with stronger components at some frequencies than others, and complex variation of signal strength with frequency. In contrast to all of these, random noise has a smooth continuous spectrum.

As just mentioned, a periodic waveform can be broken down into (or built up from) a set of sinewaves, known as 'Fourier series', after the French mathematician Joseph Fourier (1768 – 1830). Fig.2 shows an idealised spectrum of a square wave. The spectrum has peaks at the fundamental frequency (f)

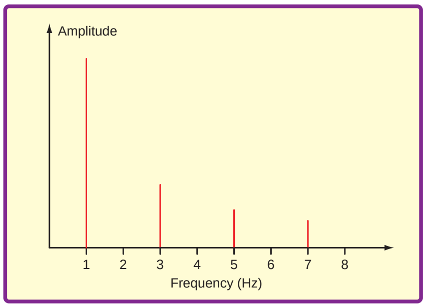

Fig.2. Ideal spectrum of a square wave

(the frequency of the square wave) and at harmonics (multiples of the fundamental) the peak sizes are 1/3 of the fundamental at $3f$, 1/5 at $5f$, 1/7 at $7f$ and so on. The spectrum has zero amplitude at other frequencies. We'll return to the spectrum of the square wave later when we discuss its use as a chopping signal.

Flicker noise

All electronic circuits and devices generate random noise. This is fundamentally due to the discrete nature of electricity at the atomic level, where electric charge in circuits is carried in packets of fixed size – electrons. Random noise may be classed according to the shape of its spectrum (see Fig.3 and Fig.4). White noise has the same power throughout the frequency (f) spectrum, whereas $1/f$ noise (also called 'flicker noise' and 'pink noise') decreases in proportion to frequency.

Standard amplifiers typically exhibit a mixture of pink and white noise, with pink noise dominating at low frequencies. The frequency at which the dominant noise component changes between pink and white noise is called the corner frequency or noise corner (see Fig.5). The fact that flicker noise increases with decreasing frequency makes it particularly important (troublesome) in the design of circuits processing low-frequency signals – hence the need for special chopper and auto-zeroing amplifiers. For audio and video amplifiers, the $1/f$ noise is not a problem as long as the $1/f$ corner is sufficiently low (say 20Hz for audio). This can be achieved with conventional amplifiers; the unwanted lower frequencies can be blocked by capacitive coupling. From an offset perspective, offsets are simply blocked by capacitive coupling.

Flicker noise is an interesting topic in its own right because it occurs in such a wide range of systems, including physical and biological systems as well as human responses and financial markets. There is interest in pink noise as a ubiquitous phenomenon to which some universal theories may apply. Readers interested in pink noise processes outside of electronics may find the Wikipedia page on the topic a useful starting point: **https://en.wikipedia.org/wiki/Pink_noise**

Two approaches

As indicated above, there are two fundamentally different approaches to designing amplifiers with very low offset and flicker noise. The first technique, the original chopping approach, is based on amplitude modulation – the signal of interest is moved to a higher frequency band, similar to how a relatively low-frequency audio signal is transmitted using a much higher frequency broadcast radio signal. After chopping, the signal of interest is contained in the higher frequency band, so it is out of the way of the offset and flicker noise from the amplifier. The amplifier signal is then demodulated to return it to its original frequency band, like a radio receiver recovering the audio from a high-frequency carrier wave. As will be explained later, demodulation moves the amplifier noise to a higher frequency band, so low-pass filtering can be used to remove it.This approach deserves the description 'original' because it has been around a long time. The first chopper amplifiers date from the 1950s and used mechanical chopping and vacuum tube amplifiers. Although the chopping technique removes offsets and low-frequency noise, there is the potential problem of adding noise created by the chopping process.

The second technique, which we will refer to as 'auto-zeroing' is based on a sampling approach. The circuit periodically samples the error (noise and offset) in the amplifier's output and uses this to apply a correction to remove the error, with the correction voltage usually stored on a capacitor. The correction can be applied at the input, output, or intermediate stage of the amplifier. If the amplifier has a fixed offset it will be removed by this approach. Flicker noise will also be reduced (effectively high-pass filtered), but the circuit will introduce additional noise from the sampling process. Next month, in *Part 2*, we will look at auto-zeroing amplifiers.

Modulation basics

Understanding the operation of the chopper amplifier requires some knowledge of amplitude modulation (as used in AM radio), so we will take a quick look at the basics. In the simplest example of amplitude modulation, the amplitude of a sinusoidal carrier wave, of frequency f_c is controlled by a sinusoidal modulating signal of frequency f_s. The carrier is at a higher, usually much higher, frequency than the modulating signal. In the context of broadcast radio, f_c corresponds to the frequency of a radio station and f_s corresponds to the frequency of the speech or music signal being transmitted. For a chopper amplifier, f_c is the chopping frequency and f_s is the frequency of the signal being amplified. The real signals are not necessarily all sinewaves, but a quick calculation with sinewaves provides basic insight to the modulation process.

The carrier signal is represented as a function of time by $A\cos(2\pi f_c t)$, where A is its amplitude. The 2π factor converts

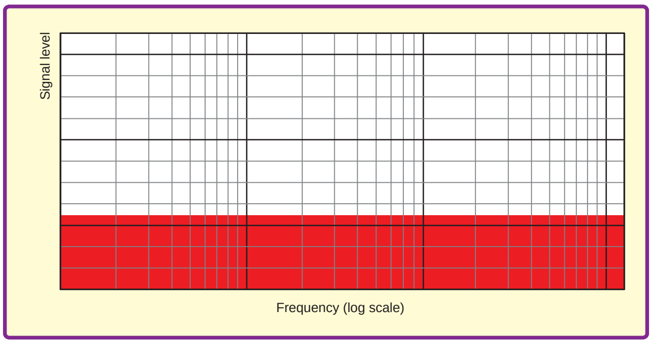

Fig.3. Spectrum of white noise (signal level in dB)

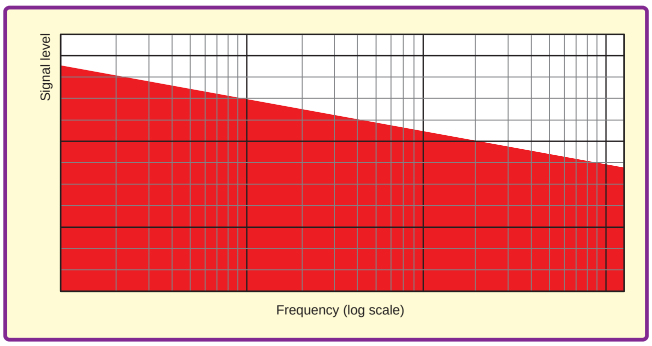

Fig.4. Spectrum of pink noise or flicker noise (signal level in dB)

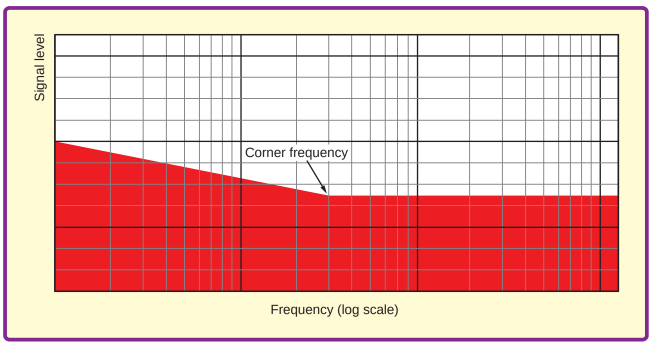

Fig.5. Typical spectrum of standard amplifier noise

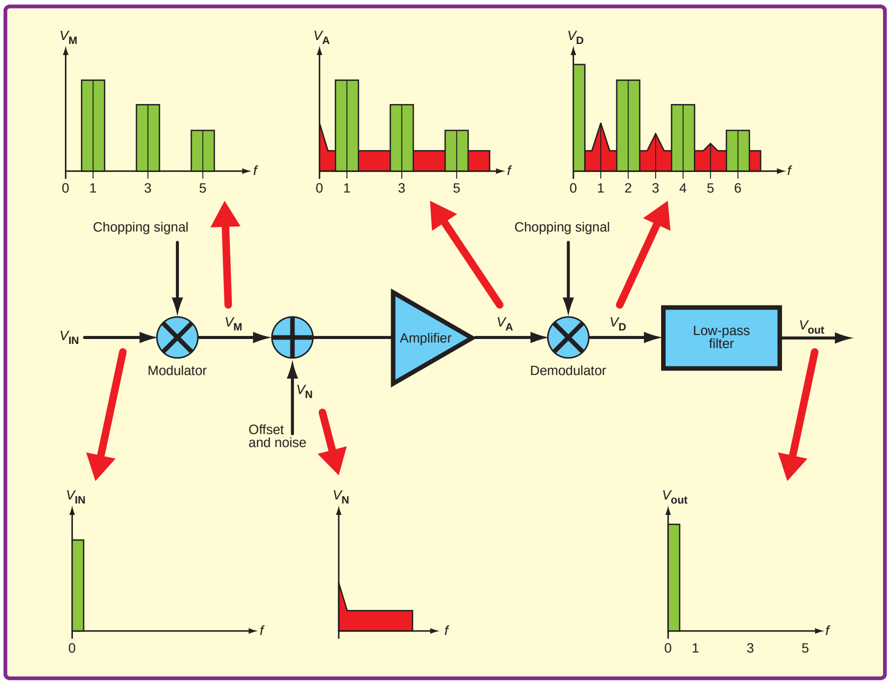

Fig.7. Principle of chopper amplification, showing signal spectra

the ordinary frequency of the signal (f_c) in hertz to an angular frequency in radians. The carrier is multiplied by the modulating signal to create the AM signal. The modulating signal is represented as a function of time as $(1+M\cos(2\pi f_s t))$, where M is the depth of modulation. If $M = 0$ the modulating expression reduces to 1. Multiplying the carrier by 1 leaves it unchanged – there is no modulation. If $M = 1$, given the extreme values of a cosine are +1 and –1, the carrier is multiplied by an instantaneous value ranging from 0 to 2. This is the maximum value of M, which does not cause potential loss of the original signal (over modulation).

Prosthaphaeresis

Multiplying the carrier and modulating signal gives:

$(1+ M\cos(2\pi f_s t))A\cos(2\pi f_c t)$

To see the implications of this clearly in terms of frequencies we need to convert the product of two cosines to individual sine or cosine functions. This problem was solved in in the sixteenth century by people interested in astronomy-based maritime navigation. This lead to the 'Prosthaphaeresis formulas' (see: **https://en.wikipedia.org/wiki/Prosthaphaeresis**) – a set of four trigonometric identities, of which we need:

$$\cos\alpha\cos\beta = \frac{1}{2}\cos(\alpha-\beta) + \frac{1}{2}\cos(\alpha+\beta)$$

Applying this to the AM signal expression gives:

$$A\cos(2\pi f_c t) + \frac{AM}{2}\cos(2\pi(f_c - f_s)t) + \frac{AM}{2}\cos(2\pi(f_c + f_s)t)$$

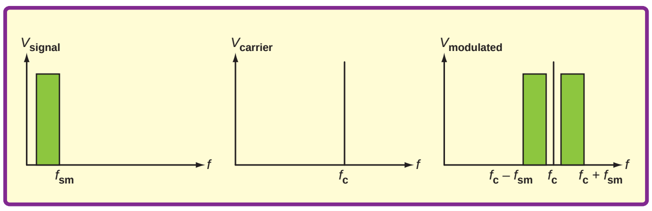

Fig.6. Spectra of amplitude modulation signals

This shows a modulated signal consists of the original carrier (frequency f_c) and two other sinusoids at frequencies above and below the carrier frequency – $(f_c - f_s)$ and $(f_c + f_s)$ – which are referred to as 'sidebands'.

Fig.6 shows the spectra of signals involved in amplitude modulation. The first graph shows the signal (V_{signal}) occupies a range of low frequencies, up to a maximum signal frequency of f_{sm}. The signal frequency does not extend all the way down to DC in this case. The second graph shows the carrier ($V_{carrier}$) at a much higher frequency than f_{sm}. The carrier is sinusoidal, so it has a single frequency (f_c) and is depicted as a narrow peak on the graph. The third graph shows that amplitude modulation moves the signal to two sidebands either side of the carrier.

Chopper amplifier

Fig.7 shows the structure of a chopping amplifier, together with simplified signal spectra at various points in the circuit. The circuit is draw as an abstract model, which shows the key processes, not as a schematic with real components. The amplifier is represented as an ideal amplifier with its noise and offset added at its input from an external source, similar to the depiction of input offset voltage in Fig.1.

Starting at the input in Fig.7, we have the low-frequency signal, which we want to amplify (V_{IN}) – its spectrum is shown to occupy a range of low frequencies, extending down to zero to illustrate our interest in DC or very low-frequency signals. The signal is modulated by a square wave (at the chopping frequency) rather than the sinewave used in the general discussion on modulation above. A square wave is used for the modulation because it is easy to implement, just requiring some switches driven by a digital clock signal. In modern ICs the switches are implemented using MOSFET transistors. As previously discussed, square waves can be broken down into a set of sinewaves at odd multiples of the fundamental frequency of the wave. Referring back to the spectrum of the square wave in Fig.2, and the spectrum of an amplitude modulated signal using a single sinewave carrier, leads to the spectrum of the modulated (chopped) waveform (V_M), shown in Fig.7.

In the system of Fig.7 the input signal (V_{IN}) and the modulation process are assumed to be noiseless. However, the amplifier adds noise and offset, represented by the signal V_N. The spectrum of V_N is shown in Fig.7 and is like that of a typical amplifier, as shown in Fig.5. Therefore, the output of the amplifier (V_A) is the modulated signal (V_M with the noise and offset (V_N) added. Again, the shape of the spectrum is illustrated in Fig.7. For convenience, the spectra are not drawn to scale (the amplifier output has a lager amplitude than V_{IN}) – it is the general shape of the spectra that is of importance in this discussion, not the absolute amplitudes.

Demodulation

The output of the amplifier passes to a demodulator, which is another set of switches, like the modulator, with a clock synchronised to the chopping of the signal. This approach is called a product detector – it multiplies the modulated signal by the original carrier to demodulate it. This differs from the well-known diode-based envelope detectors used in simple AM radio receivers.

We can consider the effect of the demodulator on the noise and signal parts of the V_A signal separately. The noise and offset has not been modulated because it originated in the amplifier after the modulator. The demodulator therefore acts on it in the same way as the modulator does on the input signal – moving the low-frequencies parts to sidebands around the chopping frequency and its odd harmonics. The modulated signal is demodulated, recreating the input signal (now amplified) in its original low-frequency range. A full

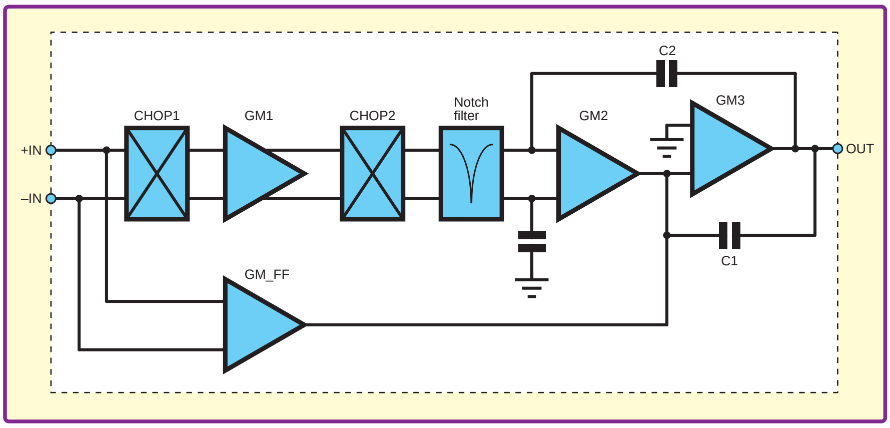

Fig.9. Block diagram of the OPA333 chopper-stabilised op amp (based on Texas Instruments data sheet)

analysis (similar to the modulation calculations above) shows that copies of the input are also created as sidebands around twice the chopping frequency. Similarly, the higher odd harmonics of the chopping frequency have their sidebands moved to the even harmonics in the demodulator output.

The clever thing here, as can be seen in the spectrum of the demodulated signal (V_D) is that the $1/f$ noise and offset has been moved out of the way to higher frequencies, whereas the wanted signal is back in its original range. There is a lot of superfluous content in the higher frequency part of the V_D spectrum, but that can be removed by a low-pass filter. The result is the output signal (V_{OUT}) free from the $1/f$ noise and offset. The output will of course still have some noise (Fig.7 is somewhat idealised) – but this will be 'ordinary' white noise, not the $1/f$ and offsets which are particularly troublesome at low frequencies.

Chopper-stabilised amplifiers

An issue with the chopped amplifier is that the low-pass filter which is required to remove the modulated noise and harmonic sidebands seriously restricts the bandwidth of the chopped amplifier. The solution is to use multiple signal paths to produce what is called a chopper-stabilised amplifier. A basic concept is shown in Fig.8. The input signal is amplified via two paths, one for low frequencies and the other for higher frequencies. DC and low frequencies are handled by a high-gain chopper amplifier like the one shown in Fig.7, followed by another (ordinary) wide-band amplifier. High frequency signals are only amplified by the second amplifier, which is not restricted to having the low bandwidth of the chopper.

A chopper-stabilised amplifier should have a frequency response like that of a compensated op amp – it will typically be used in similar circuits. Thus the gain will be very high at DC and low frequencies, but fall steadily as frequency increases – to ensure that circuits are stable when the amplifier is used with negative feedback. We discussed op amp open-loop response recently in *Circuit Surgery* (April 2018) so we will not go into further details here. The implication for the circuit in Fig.8 is that the gain

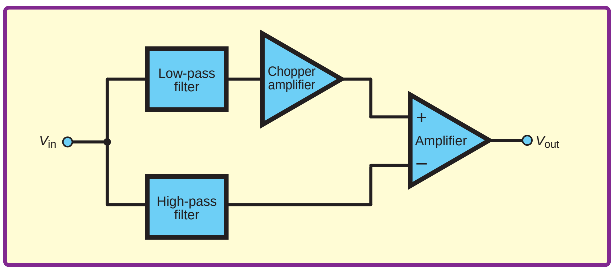

Fig.8. Circuit concept for a chopper-stabilised amplifier

of the second amplifier is much lower than the gain of the chopper. This means that although it will contribute offset to the whole circuit, the effect is relatively small.

The design of a circuit like that in Fig.8 must ensure that the frequency responses of the two amplifiers are configured so that there is a smooth transition between the high and low-frequency paths through the circuit. Modern chopper-stabilised ICs will typically have more complex structures than that shown in Fig.8 (more than two amplifiers – see below) and will achieve very smooth frequency response characteristics.

Fig.9 shows a block diagram of an example commercial chopper-stabilised amplifier – the OP333 from Texas Instruments. The device is described as a 'low-power, rail-to-rail input and output operational amplifier with a zero-drift architecture which provides ultra-low offset voltage and near-zero offset voltage drift'. In Fig.9, the amplifier GM_FF provides the high frequency path (GM refers to mutual conductance – voltage in, current out gain, and FF refers to Feed Forward – the high-frequency path is fed forward past the chopper). The chopper comprises CHOP1, GM1 and CHOP2 and has the same structure as Fig.7. The choppers are simply sets of switches, which reverse the polarity of the differential signal. The amplifiers GM2 and GM3, along with the capacitors are used to combine the signal paths and correctly configure the overall frequency response. The circuit also includes a notch filter to reduce chopping noise.

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AUDIO OUT

By Jake Rothman

Life expired? – Part 1

Fig.1. A skip full of e-waste outside a small mid-Wales electrical retailer. It is very difficult to recycle such a complex mix of materials.

Fig.3. An active loudspeaker – the current trend of putting cheap electronics in expensive loudspeaker enclosures with no circuit diagrams available is resulting in studios losing their monitoring systems for long periods.

I'm downsizing right now, the children are flying the nest and the wife will too if I don't 'de-clutter' several decades of accumulated electronic stuff! I'm looking at a hallway full of equipment which I have now decided is life expired. It's been a hard decision to make. Talk to most retailers and they will say most electronics lasts for around five years. However, in high-end audio and music, around 30 years is a common lifetime. Many a hairy musician comes to my workshop bearing a burnt-out amplifier from the 1970s, expecting it to be fixed. However, I am on the musician's side. The rising mounds of domestic e-waste illustrated in Fig.1 are rapidly becoming an ecological disaster. Reuse and repair is the best way, recycling is inefficient and too much of an excuse for, and enabler of the continuation of the throwaway culture.

This month and next, I am going to take you on a tour of failure – not mine or yours(!) – but an examination of why electronics dies and what, if anything, can be done to save old equipment.

One last point before we dive in. I am almost exclusively an analogue engineer, so I will not be covering very much of the digital side of electronic expiration.

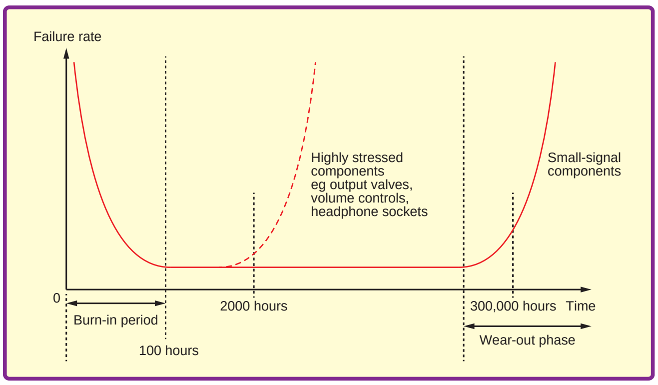

Fig.2. The familiar 'bathtub-shaped' curve for electronic equipment failure rate. Of course the middle bit won't be as flat as shown due to the sporadic failure of switches, connectors and potentiometers.

How does electronics fail?

Most electronic engineers are familiar with the 'bathtub curve' (Fig.2). There is an initial rash of random failures due to manufacturing and material defects, usually in the first 100 hours of operation after the first power up. Following this period, things settle down, until after a few years of reliable operation the first signs of wear appear. It is usually mechanical components that fail first; for example, connectors, pots and switches. Interestingly, the last components to fail tend to be the semiconductors – so long as they don't get hot.

Catastrophic failure

Catching fire is the biggest threat to electronic equipment, the damage is usually so extensive that life expiry is certain (as is the relevant design engineer's employment). Power supplies seem to be the biggest cause of such total failure, and the rise of switch-mode power supplies with stressed MOSFETs and complexity has only made the problem worse.

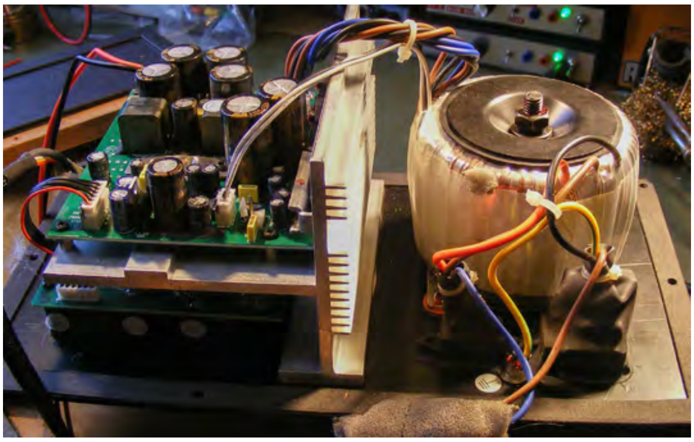

Fig.4. Typical crossover amplifier module (from the system in Fig.3). A couple of power-amp chips costing around £2.00 feed drive units costing a total of £90.00. For long life the cost ratio needs to be much lower.

Design faults

A lot of failures are the result of simple design errors, such as not taking transients into account – a classic example is tantalum capacitors placed across low-impedance power rails. They get badly stressed during turn on.

Poor thermal layout, such as putting a wet electrolytic capacitor above the air stream of a hot wire-wound resistor can also lead to premature equipment death. The up-shot of this kind of problem is an interesting side effect that specific equipment models can become more reliable with time as these defects are ironed out. (The opposite of this is when engineers produce a good reliable design that accountants slowly chip away at. A corner cut here with a slightly cheaper component, or a corner cut there with one less winding on an inductor can degrade an otherwise good design until what was reliable becomes unreliable.)

One problem that has recently been driving me mad is active loudspeaker systems. The mean time between failures (MTBFs) of the internal electronics is a tenth of the loudspeaker drive units. No wonder Hi-Fi shops don't like stocking active speakers when they are much more likely to fail than their passive counterparts. If a passive system fails then they can just give the customer another stereo amp – and any will do the job.

All systems are only as strong as their weakest link. In active speaker systems the electronics are hidden and the drive units are on 'full view', leading to a quality disparity. All too often I come across designs where a £60 woofer is driven by a £2.50 chip. The electronics inside the cabinet is exposed to considerable vibration as well as high temperatures. Even if the cabinet is vented for acoustic purposes there may be little or no regard for convection currents to vent hot air, so the electronics slowly cooks. If the failed speaker shown in Fig.3 had been designed with a port at the top of the cabinet, then sufficient airflow could have taken place. The internal electronics is shown in Fig.4.

Fig.5. Built-in obsolescence or ignorance? This LED lamp has a capacitor that lasts 2% as long as the LEDs.

Built-in obsolescence

There is plenty of evidence for this, although actually 'proving' it can be difficult! Why specify sub-systems with plastic tabs that clip together on assembly, but quickly age and snap apart when, for example, opening an enclosure? Plastic cases are a major cause of scrapping. If something is in a wooden or metal case then it is often worth repairing.

It sometimes feels as if the infamous light-bulb cartel is alive and kicking – why else would 2000-hour rated capacitors be used in LED lamps (Fig.5) when the LEDs themselves last for 100,000 hours?

Forced life expiry

Bureaucracy-led decisions such as health and safety directives or changes to standards can render equipment life expired before it has even been turned-on.

A classic case was the black and white 405-line TV system switch-off around 1982, although to be fair it was reasonably well managed. Nevertheless, it did render hundreds of thousands of TVs obsolete. Ironically, modern DSP techniques now allow 405 line signals to be easily generated, and one can look at restored 405 line

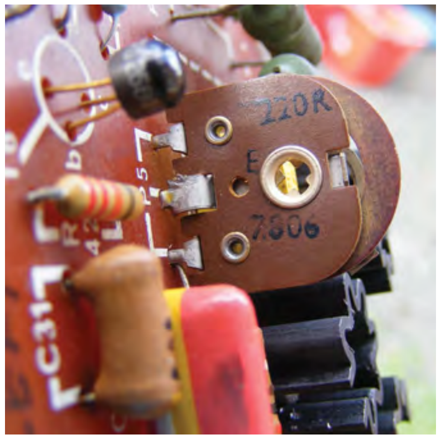

Fig.6. The age of electronic equipment can often be estimated from the date codes on components. In this Leak Delta 70 amplifier the preset is dated year 1978 week 6, putting the actual date around 1979.

pictures today at Vintage Wireless Society events.

Around 45 million radios were destined for landfill because of a stupid decision by the BBC to replace FM with an inferior 1977 digital system called DAB. Luckily, they seem to have seen sense, realising that if people want digital radio they will use the internet. My four domestic Roberts radios are rejoicing at this stay of execution – until public servants dream up another 'bandwidth-saving' idea.

Technological 'progress'

I knew an engineer who decided it was possible to make a television set that would last for 30 years, so he went ahead and built one. Unfortunately, this was just before the advent of the digital switchover and high definition LCD displays. It became 'life expired' 25 years earlier than planned. His wife put it in the skip when he expired, another victim of

Fig.7. This excellently built Tektronix storage 'scope from 1969 will go on for many more years, but is simply too big compared to modern units.

Fig.8. Carbon composition resistors steadily increase in value with time and heat. In most old equipment they have to be replaced, especially ones above 4.7kΩ. This one has been snapped in half to show its granular construction.

changing standards and leapfrogging technologies.

The Leak amplifier I have described in earlier articles in this column as an example of how hand-assembled analogue gear can be fixed forever does present a dilemma. It's high-frequency distortion is subtly inferior to the latest designs due to its quasi-complementary 2N3055 output stage. For some, this would be reason enough to dump it, but I would argue that it's better to modify it to a true complementary output if its 'inferiority' is a problem. (As an aside, using this amplifier as an example, a piece of equipment's age can often be estimated by the date codes on components, as seen in Fig.6.)

Another road to obsolescence is where a venerable old design is simply too big and heavy. My trusty old valve storage scope (Fig.7) is too difficult to move around, unlike a 2-inch deep plastic digital scope, so I've decided it's got to go.

Degraded media

Sometimes equipment is rendered obsolete because the media it is designed for gives up the ghost first. Nothing lasts for ever – so, when floppies and DAT tapes can't be read, reel-to-reel tape has gone sticky, the aluminium has oxidised in CDs and vinyl is too scratched then otherwise sound equipment is rendered useless. I am impressed that 1970s cassette tapes still work, or at least they will until 'irreplaceable' rubber belts perish in cassette desks.

I get most of my music online today, but how long will that last? Maybe the gold record on the Voyager space probe is going to be the longest-lasting medium!

Temperature

This is the main cause of aging in electronic components. Most engineers know that increasing the ambient temperature by 10°C doubles a chemical reaction rate. Extrapolating this from 20°C or a normal room temperature

Fig.9. Often, carbon composition resistors can be measured in situ. This decoupling resistor could be said to have suffered a parametric failure.

Fig.10. Electrolytic capacitors, which lose capacitance over time, can be checked with an ESR meter – this one is fine.

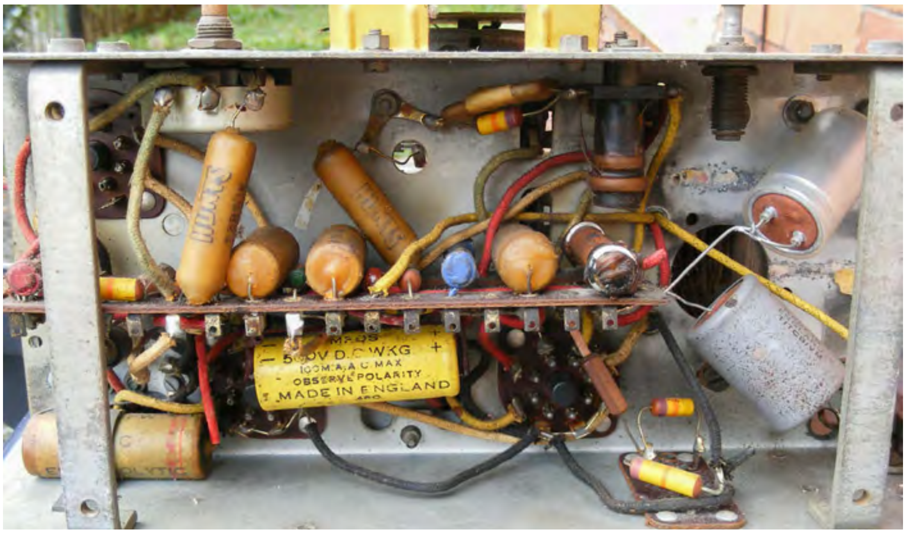

Fig.11. All the resistors and capacitors in this 1943 wartime British-government-produced civilian valve radio will probably be life expired. The wax-covered paper capacitors will be leaky due to moisture ingress. When such equipment is restored but must look original, the wax from the capacitors can be melted and the innards removed and replaced with a new axial polyester capacitor. Anything that says 'Hunts' on it has usually failed – in my experience these are particularly poor capacitors!

Fig.12. A typical lead-free solder joint failure due to heat and brittleness. Note the characteristic jagged edge, often only visible as a thin black line with magnifier. This was on a bridge rectifier in a sound bar.

Fig.13. Plastic, such as the foam on speaker surrounds, often decomposes with time, as shown here on this Vidoton Minimax speaker.

Fig.14. New synthetic rubber surrounds are now readily available, allowing many speakers to gain an extra 20 years of life.

Fig.15. The repaired bass drive unit shown in Fig.13. (The dried-out electrolytic capacitors in the crossover also had to be replaced to make the speaker sound as good as new.)

Fig.16. This mid-range unit could not be repaired since the plastic cone material had hardened and cracked due to prolonged UV light exposure (sunshine).

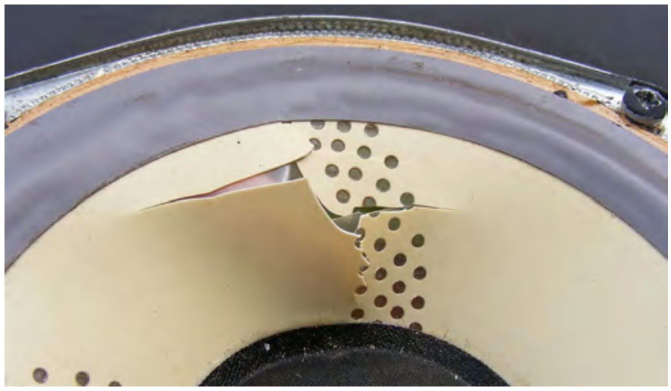

Fig.17. A close-up of the cone cracking. No replacement diaphragm assembly could be found, so more magnets for the children.

to 100°C, a hot semiconductor chip experiences a rise in reaction rates of a whopping 256 times. It is these chemical reactions that drive degradation. Carbon composition resistors are made of carbon particles (Fig.8.) that shrink with time, resulting in an increase in resistance. In the Leak amplifier shown in Fig.9, a 390Ω resistor had increased to 457Ω, nearly the next E12 value. Likewise, electrolytic capacitors dry up with time, resulting in increased ESR. Fig.10 shows how they can be checked – a low value indicates a good capacitor. (see my article in *EPE*, October 2017). Old paper capacitors become electrically leaky through water absorption if they are made in cheap, non-hermetically sealed cases, such as those shown in Fig.11).

Unleaded solder

When unleaded solder was introduced in 2006 all hell broke loose, especially as there was a lot of contamination from leaded solder, which caused a huge increase in dry joints. Unleaded solder does not have the resilience of the softer leaded variety, which leads to the thermally induced cracking shown in Fig.12. The attachment of semiconductor dies to metal tabs and single-sided PCB connections to hot components are especially prone to this problem.

Chemical instability

Many plastics and glues used in loudspeaker diaphragms seem to decompose with time. The worst offender is the foam rubber used for the surround on woofers shown in Fig.13. Luckily, new synthetic rubber surrounds (Fig.14) are available online from vendors such as the Dutch company Good Hi-Fi. The restored unit is shown in Fig.15. Another cause of accelerated chemical degradation is UV light. The highly developed Wharfedale mid-range unit shown in Fig.16 had a white plastic cone that went hard and yellow on exposure to light and then cracked (Fig.17). These units are now all life-expired with no new cones available. In turn, this means the loudspeaker systems incorporating them need to be redesigned to accommodate a different drive unit.

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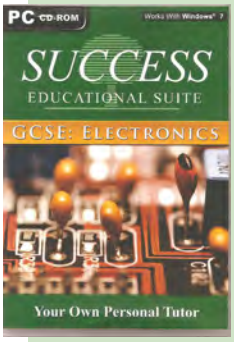

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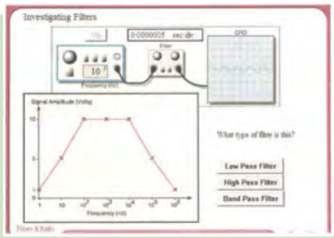

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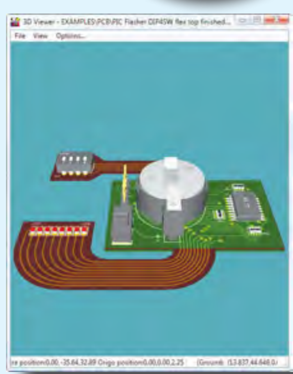

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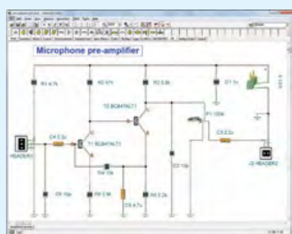

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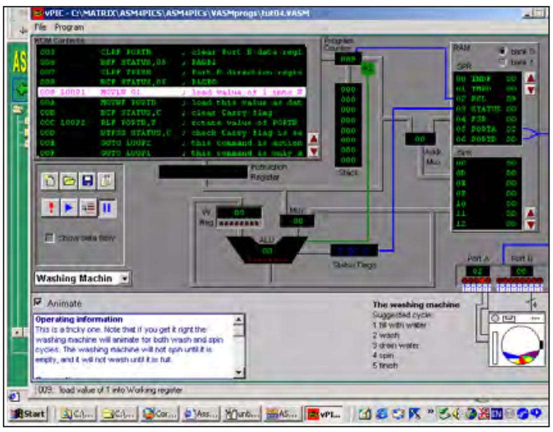

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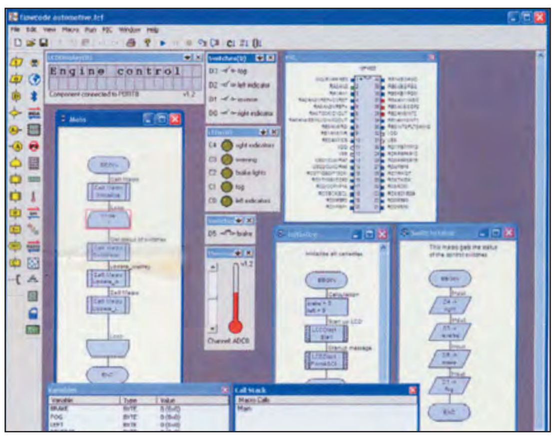

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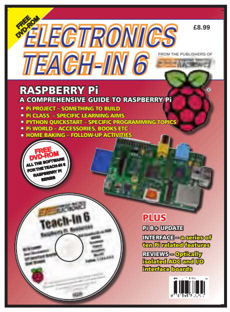

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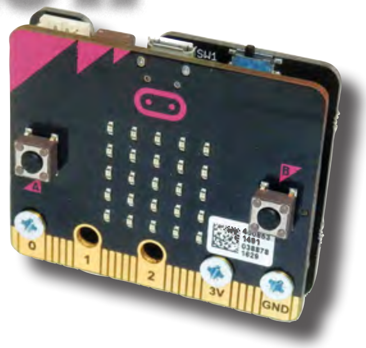

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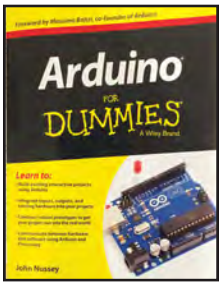

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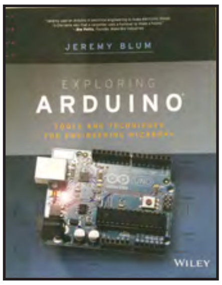

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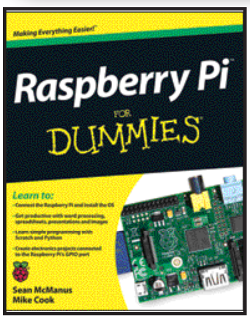

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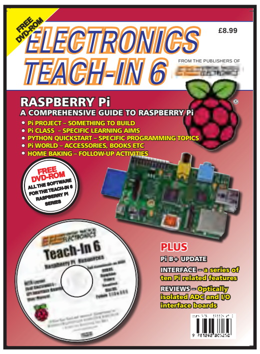

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Electronic Building Blocks

By Julian Edgar

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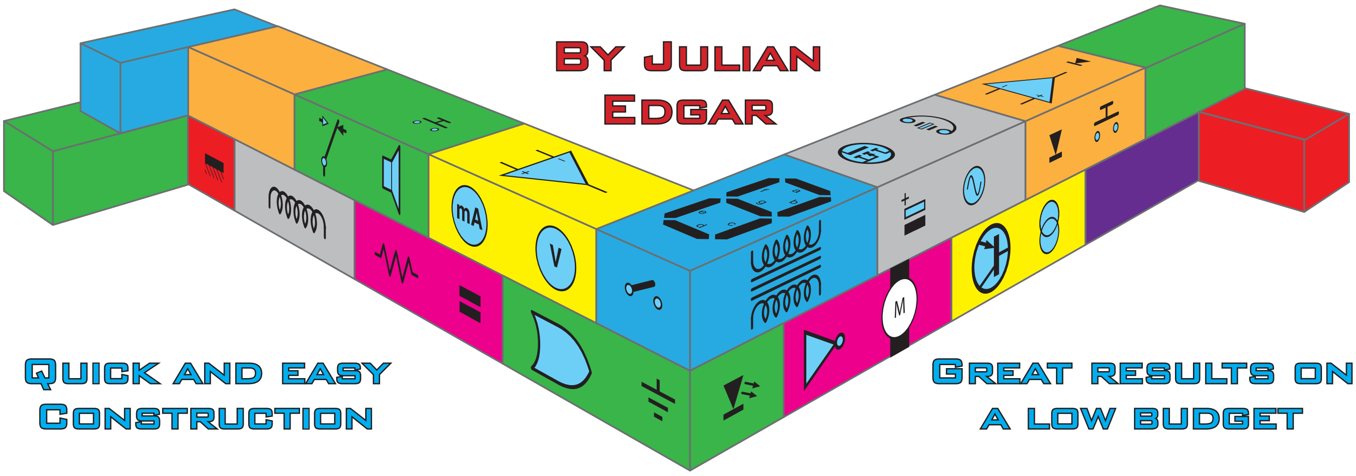


Pre-recorded Sound Module

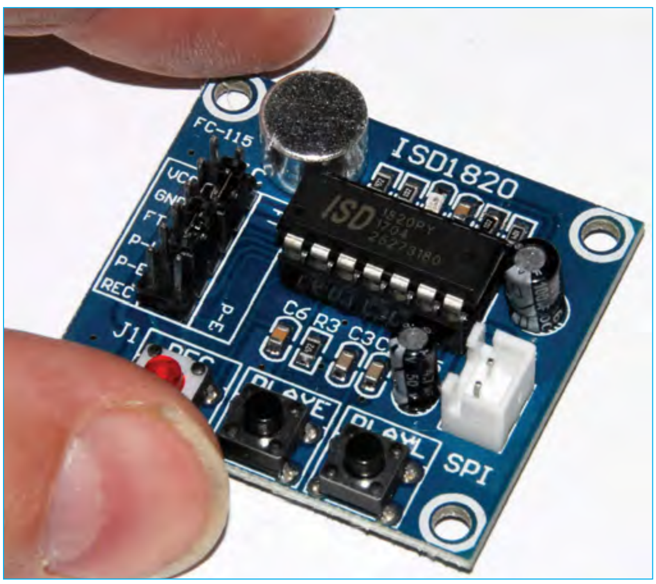
Large complex projects are fun, but they take time and can be expensive. Sometimes you just want a quick result at low cost. That's where this series of *Electronic Building Blocks* fits in. We use 'cheap as chips' components bought online to get you where you want to be... FAST! They represent the best value we can find in today's electronics marketplace!

I'm sure you're all aware that electronic sound modules that can play back messages have been getting cheaper and cheaper. In fact, so cheap that they can be found in throwaway greeting cards singing 'Happy Birthday' and the like. However, those modules are not only very cheap, they're also pretty nasty, with poor sound quality.

Pro sound quality

But the module shown here is different in one key respect. It's still incredibly cheap – less than £1 delivered – but it has excellent quality sound and so can be used in professional applications. So, what sort of application? For my money, the best applications are where warnings need to be given. For example, if you design a piece of equipment with warning lights for different fault conditions, it's now easy to add spoken warnings to accompany each light. A power supply can have a warning that actually *says*, 'Over-current, over-current', or a generator can have a warning that says 'The engine is overheating, please stop the machine'. I once got too near to an exotic car that bellowed at me, 'Step away from the car, step away from the car!'

To find the board, search on **eBay.co.uk** for 'ISD1820 module'; at the time of writing eBay item 183140919412 is supplied with an 8Ω 0.5W speaker for £1.27, including delivery. Note that not all suppliers include a speaker – the module I used didn't have one (that's why it was cheaper), so I simply added my own.



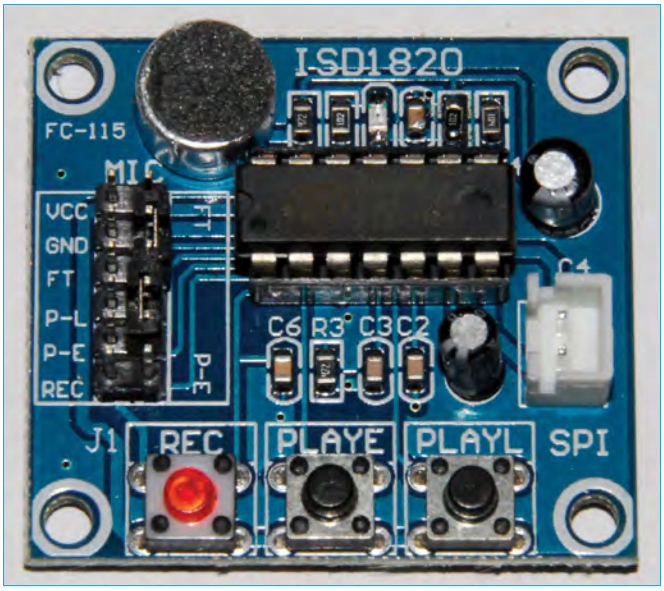
This tiny module can record up to 10 seconds of audio and then play it back on demand. It's perfect for warning messages when equipment fault conditions occur, with the audio quality sufficiently good for use in professional applications.

Specification

So what are the specs on this little module? Measuring only 43 × 38mm, the tiny board is equipped with the ISD1820 chip, three pushbuttons, power connections via header pins (note you need to use 3-5V, *not* the 12V of most modules I cover), a connector for an external speaker, and an on-board microphone. The message can be up to 10 seconds long and, as I have said, you're likely to be positively surprised by the quality of the recording. (Of course, the delivered sound quality also depends on how good the speaker is.)

In use

At its simplest, connect power and a speaker (nominally 8Ω). Press the 'REC' button and speak into the microphone. Best quality occurs if your mouth is close to the microphone. Release the 'REC' button and press the 'PLAYE' button. The message you just recorded will then be played through the speaker. If you press the 'PLAYL' button, the message will be played only for as long as you have your finger on the button – if you release the button, the message will cease playing at that point. Note that in these modes, the message will be played only once. Alternatively, by pressing the buttons, these functions can be triggered by connecting the 'REC', 'P-E' or 'P-L' header pins to the positive supply, as required. The message is retained even if power is removed – an important aspect considering the next function.



The header pins on the left enable the board to be configured in different ways. One mode plays the message on a continuous loop whenever power is supplied to the board.

Continuous loop

Moving the provided link so that the 'P-E' pins are bridged changes the way in which the module operates. In this configuration, the message is played on a continuous loop whenever power is connected to the board. This is a very useful function because it means that the board is drawing no power at all until the message is required – perfect for that fault-condition warning function. When playing a message on a continuous loop, the board draws about 50mA at 4V.

Audio monitor

Moving the other provided link so that the 'FT' pins are joined causes the module to act as a small amplifier – the sound being picked up by the microphone is fed straight to the speaker. (So that makes this module about the cheapest remote audio monitoring device available!)

The level of audio output provided straight from the module is adequate for normal desk-type warning functions. In noisier environments, you'd need to add an amplifier.

I was very impressed by this module – well made, flexible in application and with sound quality good enough to use in professional applications. And of course, for the 'less professional' among us, it's also great fun to play with in a household!

Next time

In my next column I'll be looking at a handy *Machine Tool Tachometer*.

PCB SERVICE

CHECK US OUT ON THE WEB

Basic printed circuit boards for most recent *EPE* constructional projects are available from the *PCB Service*, see list. These are fabricated in glass fibre, and are drilled and roller tinned, but all holes are a standard size. They are not silk-screened, nor do they have solder resist. Double-sided boards are **NOT plated through hole** and will require 'vias' and some components soldering to both sides. **NOTE: PCBs from the July 2013 issue with eight digit codes** have silk screen overlays and, where applicable, are double-sided, plated through-hole, with solder masks, they are similar to the photos in the relevent project articles.

All prices include VAT and postage and packing. Add £2 per board for airmail outside of Europe. Remittances should be sent to **The PCB Service, Everyday Practical Electronics, Wimborne Publishing Ltd., 113 Lynwood Drive, Merley, Wimborne, Dorset BH21 1UU. Tel: 01202 880299; Fax 01202 843233; Email: orders@epemag.wimborne.co.uk. On-line Shop: www.epemag.com**. Cheques should be crossed and made payable to *Everyday Practical Electronics* **(Payment in £ sterling only)**.

NOTE: While 95% of our boards are held in stock and are dispatched within seven days of receipt of order, please allow a maximum of 28 days for delivery – overseas readers allow extra if ordered by surface mail.

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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 ***

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A large number of older boards are listed on, and can be ordered from, our website.
Boards can only be supplied on a payment with order basis.

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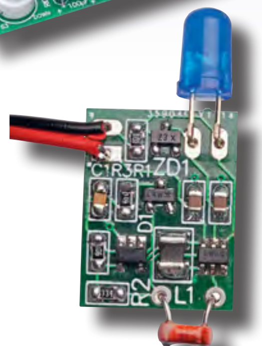
Touchscreen Appliance Energy Meter

How much do your appliances actually cost to run? Are you getting the most bang for your buck? This new *Appliance Energy Meter* will tell you exactly how much they're using, how much they're costing you and the total energy consumed. It can even log the results to your computer.

Automotive Sensor Modifier

With this *Automotive Sensor Modifier* you can change the signal response of many car sensors to improve driveability, throttle response, handling and so on. It allows you to modify and program the response of any voltage sensor in your car, without prejudicing reliability or affecting the ECU in any way.



Ultra-low-voltage Mini LED Flasher

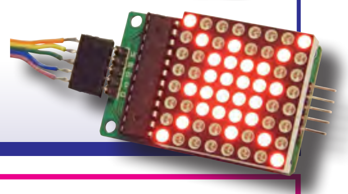
This versatile design uses just a handful of components to flash any colour LED brightly, powered from a single alkaline cell. In fact, it will run off any supply from 0.8V to 3.3V and consumes very little power when the LED is off. It's built on a tiny board, so it will fit just about anywhere and incorporates ambient light monitoring to switch the LED off during the day.

Low-cost Electronic Modules – Part 7

Next month's low-cost module uses a Maxim MAX7219 serial LED display chip and comes complete with a plug-in 8x8 LED matrix display. It is equally capable of driving an 8-digit 7-segment LED display and its SPI interface allows it to be driven by a micro using only three wires.

10-Octave Stereo Graphic Equaliser – Part 2

We conclude this project with assembly details of the PCB and acrylic case with its very smart front panel, which is actually a black screen-printed PCB.



PLUS!

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

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