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OPTO-THEREMIN – PART 2
Construction, testing and adjustment

PLUS
TEACH-IN 2015, INTERFACE, CIRCUIT SURGERY, NET WORK, AUDIO OUT, TECHNO TALK, READOUT, COOL BEANS & PIC n’ MIX
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by Nicholas Vinen
One unit – but ten effects! Build this superb digital processor with every popular effect from echo and reverb to fuzz, flanging and phasing.

OPTO-THEREMIN – PART 2
by John Clarke
Now comes the fun part! We complete construction and describe the test and adjustment procedure for our modern take on a classic electronic instrument.

COURTESY LED LIGHTS DELAY FOR CARS
by John Clarke
A circuit specifically designed to suit LED lamps. It keeps the interior lights of your car lit for a preset time after you shut the car doors.

TECHNO TALK by Mark Nelson
Lateral thought

TEACH-IN 2015 – DISCRETE LINEAR CIRCUIT DESIGN
by Mike and Richard Tooley
Part 9: Bringing it all together

NET WORK by Alan Winstanley
When content is king... Taking Ten Flat Earth society... Out in a Flash

INTERFACE by Robert Penfold
Raspberry Pi PWM

CIRCUIT SURGERY by Ian Bell
Current mirrors and transistor matching

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Passing of the torch and concluding the LPLC Oscilloscope

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RIAA equalisation – Part 4

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EDITORIAL

Putting our ‘regulars’ in the spotlight!

I often focus on projects in my editorials, but that does rather sell short our excellent columnists. So this month, I have decided to focus on our regular features. We have some fascinating articles for you in this issue. Mark Nelson’s Techno Talk provides some illuminating insights into the latest light-based technology – from solar power to novel LED fabrication technology.

Net Work is the usual must-read compendium of news, technology, tips and advice. I do wonder how Alan finds time to cram all his activities into what must be a very busy schedule that includes looking after our website, helping Chat Zone forum members and writing Net Work.

Ian Bell’s Circuit Surgery is a wonderful combination of in-depth analysis coupled with practical down-to-earth design, and this month focuses on current mirrors; helping a reader understand why real-word circuits can depart from their idealised versions. Naturally, LTSpice is on hand to handle the drudgery of endless calculations and error-prone manual modelling. If you have yet to take the plunge with LTSpice then you really should promise yourself to make its acquaintance, it is an absorbing and invaluable tool.

In Audio Out, Jake Rothman has completed his four-part series on RIAA filters – even if audio is not your passion, Jake’s columns are well worth reading for their tips and insight into high-quality analogue design.

Max and his passion for flashing lights receives a boost in Cool Beans, as he explains his latest foray into smart tri-coloured LEDs. The combination of diode and his passion for flashing lights receives a boost in Cool Beans, as he explains his latest foray into smart tri-coloured LEDs. The combination of Arduino controllers and a neat little package called the NeoPixel is a gift for any light-based project designer.

Welcome Mike O’Keeffe

Last month, we had to bid a fond farewell to long-term PIC n’ Mix columnist Mike Hibbett. As promised, this month, stepping into Mike’s 32-bit-sized boots is another Mike – Mike O’Keeffe – who has completed Mike Hibbett’s excellent LPLC Oscilloscope project. This is a most-promising start for our new PIC n’ Mix scribe, and we wish him well as he settles into his new role.

Passive DI project

Our eyes were bigger than our belly page-wise this month, so our apologies for the lack of the promised Passive Direct Injection Box – we just couldn’t squeeze it in! Fear not, it is a good project and we will include it in an upcoming issue.
Cinema innovates with 4K – report by Barry Fox

Slowly but very surely, digital projection has become the norm in cinemas. Only a few diehards still think they prefer 35mm or 70mm film, and occasionally a few specialist cinemas oblige with a screening. Unless the print is brand new, the dirt, scratches, jitter and flicker serve as a useful reminder of how much better digital projection is for everyday cinema.

Widescreen weekend
For serious film buffs, the National Media Museum, in Bradford, is holding a Widescreen Weekend (15-18 October). Bradford is one of only three venues in the world capable of screening 70mm widescreen, three-strip Cinerama and 35mm film formats under one roof. The weekend will celebrate the 60th anniversary of the Todd-AO* 70mm widescreen process with ‘Oklahoma!’ (1955), as well as staging a 70mm screening of Stanley Kubrick’s ‘2001: A Space Odyssey’ (1968) and a three-projector Cinerama screening of ‘How the West Was Won’ (1962) from an authentic Technicolor print.

4K digital cinemas
Back in the real world, cinema chain Vue International recently equipped all its 83 screens in the UK and Ireland with Sony Digital Cinema 4K projection systems, and is using them to screen ‘live’ plays from the UK’s National Theatre, shot and post-produced in 4K.

Finding new markets
After 4K tests with the stage production of ‘War Horse’ in February 2014, the first end-to-end 4K production was ‘Behind the Beautiful Forevers’, The play, by David Hare and based on Pulitzer Prize-winner Katherine Boo’s book on slum life in Mumbai, India, was captured in performance in March 2015 using six Sony F55 cameras.

At a press preview screening in central London, Kevin Styles, UK & Ireland MD, Vue Entertainment, told how ‘Event Cinema’, theatre and sport on a big screen, is now Vue’s ‘fastest growing business, doubling year on year’.

Rufus Norris, of the National Theatre, said that screened theatre had doubled their audiences since the first 2K productions in 2009.

Colour space and quality
Sony says the colour space captured with the Sony F55s is far in excess of the Rec.709 standard for HD TV. The material is mastered in UHD (3840x2160 resolution) and a 4K DCP (digital cinema package) file created for cinema projection (4096x2160).

The 4K image clarity is certainly impressive, and the experience rewarding and enjoyable. But it is very different from attending the National Theatre’s venues on the South Bank of the Thames in London – and not just because few people can afford seats that are close enough to the stage to compare with a screen experience. The use of close-ups and focus-pulling between characters is quite unlike live theatre, even as seen from the front row of the stalls, especially as 4K resolution heightens the defocussing effect.

On a non-technical level, the second half of the play at the press screening was more captivating than the first, simply because Vue had given everyone at the preview event a bag of noisy popcorn. Anyone munching loudly in a South Bank theatre would be quickly shushed and then ejected.

Fast charging
New low-cost smartphones from Motorola boast a feature called ‘TurboPower’, claimed to make them ‘the world’s fastest charging smartphones’, with 10 hours of power stored from a mains charger in just 15 minutes. As Motorola demonstrated, in a comparative test staged at the London end of a teleconference link with Brazil, Turbopower far exceeds the charging speed of a Samsung Galaxy S6.

Operation
Finding anyone at the London event who could explain how the technology works, without risk of damage to the lithium ion batteries – or even fire caused by overheating – proved surprisingly difficult.

But persistence – and a magnifying glass on the microscopically small print on the charger itself – revealed that to use the feature the user must purchase an extra charger (made in China) which has a standard micro USB plug, but stepped DC output. Step 1 sends the normal 5V to the phone, but at high current (2.85A), and Step 3, called ‘Turbo 2’, sends 12V at 2.15A.

Fast-charge-phone models
When connected to a conventional phone, the charger defaults to safe 5V working. However, when connected to a new Moto X Style or X Play phone, the charger handshakes with the phone via the USB data wires, and sends 12V. After partial charging is complete the chargers step down to 9V. Moto X Play will be available from the end of August and will cost £299; Moto X Style follows later this year with pricing as yet undecided. Details of availability and pricing for the Turbo chargers were unavailable.
Antex TCS230 launched

Digital temperature control at the touch of a button from Antex's new TCS230 soldering iron

New Antex digital soldering iron delivers the best of both worlds – Alan Winstanley reports

Every electronics constructor has an electric soldering iron in their toolbox, and many discerning enthusiasts choose a variable-temperature soldering station that offers more flexibility to cope with a wider variety of tasks. Soldering iron manufacturer Antex (Electronics) Ltd has launched a brand new concept in soldering irons that offers users the best of both worlds. The new Antex TCS230 is a mains-powered pencil-style soldering iron with a digital temperature control built into a new ergonomically shaped handle, which totally eliminates the need for a bulky bench-top station.

A backlit digital LCD temperature display and twin push buttons within the handle enable temperatures over the range 200 to 450°C (392 to 842°F) to be programmed directly into the iron at the touch of a button. The 230V, 50W rating is ample for all routine electronic and maintenance tasks, including lead-free soldering. A high-grade ceramic element offers very high insulation properties with leakage claimed as low as 5µA, making the Antex TCS230 suitable for working with sensitive microelectronics. Included as standard is their new-style slide-on, tool-free 2.3mm nickel-plated bit that makes the TCS230 ideal for general purpose use, and a full range of nine bits from a pinpoint 0.12mm to 6mm is available from Antex. A silicone mains cable protects against accidental damage, making the iron ideal for student or trainee use, and Antex offers a choice of UK or European mains plugs. A matching compact bench stand with sponge (the ST6A) is also available.

The new digital soldering iron will be the ideal solution for hobbyists and professionals seeking a powerful space-saving soldering iron with variable temperature and LCD display. The Antex TCS230 retails at £69.99 and is available from Antex and their distributors. Contact Antex (Electronics) Ltd, Westbridge Industrial Estate, Tavistock, Devon PL19 8DE, UK. Tel: +44 (0)1822 613565. Online, visit: www.antex.co.uk or email sales@antex.co.uk for more information.

Breakthrough memory from Intel

Intel and Micron Technology have unveiled their new 3D XPoint technology, a non-volatile memory that has the potential to revolutionise any device, application or service that benefits from fast access to large sets of data. Now in production, 3D XPoint is a major breakthrough in memory process technology and the first new memory category since the introduction of NAND flash in 1989.

XPoint is up to 1,000 times faster and has up to 1,000 times greater endurance than NAND, and is 10 times denser than conventional memory.

The performance benefits of 3D XPoint technology could also enhance the PC experience, allowing consumers to enjoy faster interactive social media and collaboration as well as more immersive gaming experiences. The non-volatile nature of the technology also makes it a great choice for a variety of low-latency storage applications since data is not erased when the device is powered off.

The innovative, transistor-less cross point architecture creates a three-dimensional checkerboard where memory cells sit at the intersection of word lines and bit lines, allowing the cells to be addressed individually. As a result, data can be written and read in small sizes, leading to faster and more efficient read/write processes.

Wireless charging for metal-body devices

Mobile phones that have metal bodies will soon be able to wirelessly charge themselves thanks to a new Qualcomm technology.

Wireless charging technologies that use induction heat metal, making them incompatible with any metal cases. Qualcomm’s approach uses magnetic resonance, which is much more tolerant of metal items. It creates a charge over a small three-dimensional space so that coins, keys and other metal items in that space are unaffected by the charge.

It taps into a technology known as ‘WiPower’, which conforms to a standard called ‘Rezence’. WiPower’s magnetic resonance can operate at a frequency better able to handle metal objects without affecting the charging process. With this latest innovation, Qualcomm’s WiPower can charge an entire device made of metal. According to Qualcomm, WiPower can charge devices that require as much as 22W at speeds that match or exceed those of other wireless charging technologies.

Braille smartwatch

A Korean company has created ‘Dot’ – the first smartwatch for the visually impaired. Dot has four sets of six dots, which raise to produce braille characters at a time. New characters can rise at speeds ranging from 1-100 per second, which should suit all skill levels. See: http://fingerson.strikingly.com
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Solar power, one of the cleanest sources of renewable energy, may be poised to take a fundamentally new direction. Generating just volts may not be the best way of turning sunlight into power. This and other radical new trends in solar electronics are discussed by Mark Nelson.

**Lateral thought**

HAVE YOU EVER STOPPED YOUR car to ask for directions, and received the following reply? “If I were going there, I certainly wouldn’t start from here…” Not very helpful, but it displays a refreshingly alternative line of thought. Precisely the same kind of lateral thinking applies to a remarkable new development in solar power – outputting fuel rather than electricity. The fuel is hydrogen gas that can be employed in the chemical industry or combusted in fuel cells – in cars for example – to drive engines. The principle has been demonstrated in working prototype form in the Netherlands, where it was devised by researchers at Eindhoven University of Technology. Solar fuels are a hugely promising replacement for fuels that pollute the environment, and if the process can be operated economically, the benefits are potentially enormous. So how does it work?

**Nannoo, nannoo!**

Remember this catchphrase from ‘Mork and Mindy’? No! Nor do I, because I never watched the programme. It’s no longer relevant anyway, because the new deal is ‘nano, nano’ – ‘nano’, as in nanowire. Nanowires, about 10 to 500nm (nanometres) in diameter – where 1nm is one thousandth of a micrometre, or a millionth of a millimetre – came up last month in connection with LEDs. Nanowires can also produce fuel from sunlight, by ‘splitting’ liquid water into its constituent hydrogen and oxygen parts using the electricity that is generated photovoltaically. You might assume that there’s nothing special about the electrolysis process, but actually that’s the clever part. To produce hydrogen gas from water, the solar panels use the material gallium phosphide (GaP), specifically in the form of very small nanowires. Using nanowires for this task is novel and helps to boost the yield by a factor of ten – and using ten-thousand times less precious material.

GaP has good electrical properties, but its drawback is that it cannot easily absorb light when a large flat surface is used in GaP solar cells. The researchers overcame this problem by making a grid of very small GaP nanowires, measuring 500nm long and 90nm thick. This immediately boosted the yield of hydrogen by a factor of ten to 2.9%. The university’s professor Erik Bakkers explains the breakthrough is not simply about the yield, conceding there is still plenty of scope for improvement. But, he stresses: ‘For the nanowires we needed ten-thousand times less of the precious GaP material than for cells with a flat surface. That makes nanowire cells potentially a great deal cheaper. In addition, GaP is also able to extract oxygen from the water – so you then actually have a fuel cell in which you can temporarily store your solar energy. In short, for a solar fuel’s future we cannot ignore gallium phosphide any longer.’

**Choose your own colours for LEDs**

Graphene, an atomically thin and perfectly crystalline form of carbon, was featured in last month’s article. This time, the spotlight is on phosphorene, a layer of phosphorus also one atom thick. The similarity ends there, however, because unlike graphene, phosphorene is a semiconductor, just like silicon, which is the basis of almost all of today’s electronic technology. In Australia, scientists studying thin layers of phosphorus have discovered surprising properties that could open the door to ultrathin and ultralight solar cells and LEDs. It also shows very promising light-emission properties, states lead researcher Dr Yuerui (Larry) Lu, from the Australian National University (ANU).

“Because phosphorene is so slender and wispy, it opens possibilities for making lots of interesting devices,” he says. As well as creating semiconductors much thinner and lighter than silicon, phosphorene has light-emission properties that vary widely with the thickness of the layers, which enables much more flexibility for manufacturing. This property has never been reported before in any other material, he told the journal *Light: Science and Applications*. ‘By changing the number of layers we can tightly control the band gap, which determines the material’s properties, such as the colour of LED it would make. Under the microscope you can see quite clearly the different colours of the sample, which tells you how many layers are there,’ concludes Dr Lu.

**Solar Paper is here already**

Dr Lu’s team is still working on his ultrathin solar cells, but a solar design start-up company in South Korea plans to have them on sale by the time you read this article – allegedly, anyway. As the technasia.com website reports, a couple of green entrepreneurs plan to make solar power accessible to smartphone users, and since January of this year they have been working on making a portable solar charger that’s as thin, powerful and aesthetically appealing as possible.

Called ‘Solar Paper’, the product fits snugly between the pages of a notebook and weighs just four ounces. In sunshine it can reliably charge your smartphone in about 2.5 hours, just like a wall charger. The basic model supplies 5W and if you need more power to revitalise a tablet or to charge on a cloudy day, you can click additional panels into place using their built-in magnets. Outdoorsy types can clip it to a backpack with carabiner clips supplied and charge while hiking. Urban adventurers can easily slip the panel into a jacket pocket or notebook. You can even see how the panel performs under different weather conditions, using a built-in LCD ammeter that shows exactly how much current is flowing into your device at any given time.

Retail price for the basic model has been set at around £77 and Solar Paper has its own web page at https://www.kickstarter.com/projects/1398120161/solar-paper-the-worlds-thinnest-and-lightest-solar where you can see how chic it looks, discover its ingenious features and place orders. Alternatively, go to YouTube.com and enter ‘Solar Paper’ (with the quote marks) to watch no fewer than ten videos on this imaginative device.
WANT TO SPICE up your guitar performances? Build this Digital Effects Processor into a guitar amplifier and you will get many different effects to play with, without needing to lug around and wire up many different effects pedals. It can apply the majority of common effects to a line-level signal, and you can adjust them to suit your needs.

We can’t promise that this will replace all your effects units but it certainly gives a lot of different options, which suit a variety of instruments, performers and musical styles. The idea is to build it into a guitar amplifier by connecting it between the preamplifier and amplifier sections. It can be powered directly from an amplifier supply rail, assuming a suitable DC voltage is available, or the supply rail can be derived, creating one very convenient package!

But it is not just intended for use with guitars. It is suitable for use with a large variety of other musical instruments, whether they are keyboards or instruments with pickups. And they can be used to enhance vocals as well.

By NICHOLAS VINEN

This deceptively simple unit provides 10 different musical instrument effects, including echo, reverb, tremolo, fuzz, compression, flanging and phasing. Each effect is adjustable and can be defeated with a foot pedal switch. It’s designed for use with electric guitars but will work with other instruments and vocals too.

Digitally Effects Processor

The available effects are shown in Table 1. For each effect, there are two parameters that can be set using potentiometers VR3 and VR4. Those parameters are also listed in the table and described in the list of effects below.

Note that when one of the enabled effects causes a reduction in signal level (eg, echo or reverb), the level for all effects is reduced, as well as the
The available effects are as follows:

- **Echo**: delays/attenuates the incoming signal, then mixes it back in for the output. VR3 adjusts the delay between (nearly) 0ms and 1200ms with an exponential curve, to make it less sensitive at the shorter end, which is more useful. VR4 adjusts the attenuation level; at higher settings, the echo is louder. Note: as the echo becomes louder, the original signal becomes quieter to prevent overload.
- **Reverb**: the same as echo, except that many extra short echoes are added to simulate reflections from multiple hard surfaces in close proximity.
- **Tremolo**: the output volume is modulated by a sinusoidal waveform. VR3 adjusts the amount of modulation (ie, ‘depth’) while VR4 changes the frequency.
- **Vibrato**: the output frequency is modulated by a sinusoidal waveform. VR3 adjusts the amount of modulation (ie, ‘depth’) while VR4 changes the frequency. Note that this is performed by slightly speeding up and slowing down the audio signal, although the change in delay that this causes should be imperceptible.
- **Overdrive**: this provides adjustable clipping for the signal. VR3 adjusts the gain applied to the signal and once the amplitude is high enough, it clips. VR4 adjusts how progressively the clipping occurs; at minimum setting it is hard, resulting in a square wave, while at higher settings for VR4, the clipping is more progressive and the waveform becomes rounded.
- **Fuzz**: the same as overdrive, except that the gain is applied asymmetrically, in order to inject extra distortion into the signal.
- **Compression**: the gain is slowly increased until the output reaches 90% of maximum. If the output exceeds 90%, the gain is decreased. VR3 sets the rate of increase while VR4 sets the rate of decrease. The scale for VR4 is different for VR3 as the rate of decrease is normally much higher.
- **Noise gate**: similar to but not quite the opposite of compression. When the input signal is below the threshold, there is no output. When the input goes above the threshold, it is sent to the output. VR3 adjusts the threshold while VR4 adjusts the hysteresis, to prevent the output from fluctuating on and off with a signal near the threshold.
- **Flanger**: this mixes the input signal with a version of the signal that has slight vibrato applied, causing a distinctive ‘comb filter’ Doppler effect.
- **Phaser**: similar to flanger but mixes the signal with a version that has a modulated phase shift, causing a ‘rippling’ effect which makes the sound seem artificial.

**Features & Specifications**

- 10 effects to choose from: echo, reverb, tremolo, vibrato, overdrive, fuzz, compression, noise gate, flanger and phaser
- Each effect has two adjustable parameters
- Maximum echo/reverb delay: 1.2 seconds
- Four-position switch selects between three effects and no effect with seamless transitions
- Optional defeat switch (eg, foot pedal)
- Low noise and distortion: THD+N typically <0.02%, signal-to-noise ratio >76dB
- Two power supply options: 3.5-6V DC / 7.5-12V DC; current drain 60-80mA
- Optimal input signal range: 0.5-2V RMS
- Line output signal: typically 1V RMS
- Input impedance: 4-6kΩ
- Optional headphone output
- Optional microphone preamplifier

### Table 1: Effects controls

<table>
<thead>
<tr>
<th>#</th>
<th>Effect</th>
<th>VR3</th>
<th>VR4</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>Echo</td>
<td>Echo delay</td>
<td>Echo fall-off</td>
</tr>
<tr>
<td>2</td>
<td>Reverb</td>
<td>Reverb delay</td>
<td>Reverb fall-off</td>
</tr>
<tr>
<td>3</td>
<td>Tremolo</td>
<td>Amplitude</td>
<td>Rate</td>
</tr>
<tr>
<td>4</td>
<td>Vibrato</td>
<td>Amplitude</td>
<td>Rate</td>
</tr>
<tr>
<td>5</td>
<td>Overdrive</td>
<td>Gain</td>
<td>Softness</td>
</tr>
<tr>
<td>6</td>
<td>Fuzz</td>
<td>Gain</td>
<td>Softness</td>
</tr>
<tr>
<td>7</td>
<td>Compression</td>
<td>Attack</td>
<td>Decay</td>
</tr>
<tr>
<td>8</td>
<td>Noise Gate</td>
<td>Threshold</td>
<td>Hysteresis</td>
</tr>
<tr>
<td>9</td>
<td>Flanger</td>
<td>Amplitude</td>
<td>Rate</td>
</tr>
<tr>
<td>10</td>
<td>Phaser</td>
<td>Amplitude</td>
<td>Rate</td>
</tr>
</tbody>
</table>
Everyday Practical Electronics, October 2015

Fig. 1: the basic Digital Effects Processor circuit. The incoming audio analogue signal at CON1 is digitised by CODEIC2 and then fed to IC1 where it is processed and then sent back across the same digital audio bus to IC2. A DAC in IC3 then converts it back into an analogue signal, which is fed to the output (CON2).

Options

This Digital Effects Processor uses the same hardware as the Stereo Echo & Reverberation Unit (March 2015) and the Dual Channel Audio Delay (February 2015). However, we have removed a number of components which aren’t needed. For example, most musical instruments are not stereo, so components are only fitted for one channel (and indeed, the software only supports one channel).

As with those earlier designs, it is possible to add extra components to provide a microphone input or stereo headphone output. The processed mono signal is sent simultaneously to both headphone output channels.

The headphone output could be useful for monitoring purposes. It’s up to you whether you want to install the few extra components required, which are shown in the circuit diagram at upper-right and on the overlay diagram, labelled in green.

The microphone input is less useful as its signal-to-noise ratio is only average. For a musical performance, you would be better off using an external microphone preamplifier such as our High-Performance Microphone Preamplifier from the June 2012 issue, which can run from the same DC voltage source as the Digital Effects Processor unit.

Software

In adding these new effects to the software, we have made some other changes at the same time. By making it process only a mono signal, this doubles the maximum echo to 1.2 seconds without needing an external RAM chip. This is more than long enough for instrumental work and so we haven’t even bothered to provide the option of extra RAM in the software.

We’ve also gone to some effort to make changes between effects and changes in effect settings ‘seamless’, so that clicks and pops are not generated during a performance, even if settings such as echo delay are adjusted live.

Circuit description

The circuit diagram of the Digital Effects Processor is shown in Fig.1. As stated earlier, this is a simplified version of the circuit for the Stereo Echo & Reverberation Unit from the March 2015 issue, with unnecessary components removed. That’s why there are so many unconnected pins on IC1: those originally used for interfacing with the unused SRAM chip and USB socket are not connected to anything.

A line-level signal, from a guitar preamp or mic preamp is fed into CON1 (connector tip). RF signals that may have been picked up are rejected by a low-pass filter comprising a 1kΩ series resistor and 1nF capacitor to ground, while 5kΩ trimpot VR6 is used to reduce the level to no more than 1V RMS, the limit of what the CODEC can handle.

The signal is then AC-coupled to the right channel input of the CODEC (IC3) via a 1µF DC-blocking capacitor. A half-supply (−1.65V) DC bias for this input is provided by the IC itself.

Alternatively, a microphone signal can be applied to a 3.5mm jack socket connected to pin header CON9 and this is coupled to IC3’s microphone input pin (pin 18) via a 1µF capacitor and optional series resistor (Rmic) which reduces the amount of gain if fitted; otherwise it is linked out. IC3 can supply a bias current for electret microphones and this is fed via a 680Ω series resistor. The associated 220µF capacitor provides some RF filtering for the microphone signal.

The microphone input is selected when the RE2 input of IC1 (pin 62) is pulled low. This is wired to the microphone socket so that the sleeve of the mono jack plug shorts it to ground when it is inserted. When this line is open-circuit, the line input is the active input. If the microphone input is not needed, the components in the pink box at left do not need to be installed.

CODEC operation

Whichever signal is selected, it is digitised by IC3 with a sampling rate of around 40kHz, and the resulting PCM digital audio signal is transmitted to PIC32 microcontroller IC1 via an FSB bus. This appears at pins 3, 5 and 6 of IC3, which are the serial bit clock, sample clock and serial data line respectively. These connect to the audio CODEC-compatible SPI peripheral in IC1.

IC1 reads the digital audio data from the CODEC, processes it to add the selected effect (depending on the mode) and also stores it within its 128KB
RAM, for the echo and reverb effects. Processed audio data is sent back over the same I²S bus, this time to pin 4 of IC3 but timed using the same clock lines. The CODEC then converts this digital stream back to analogue audio data, which it transmits from its line out (pin 13) and headphone out (pins 9 and 10).

These signals are all AC-coupled to the respective output connectors, to remove the 1.65V DC bias, via a 1µF capacitor for the line output and 220µF capacitors for the headphone outputs. The reason the headphone output needs much larger capacitors is that the headphones will have a much lower impedance than the line input of other equipment; 8-600Ω for headphones compared to several kilohms for a line input.

The line output also includes a 100Ω series resistor, both to prevent cable capacitance from causing instability in the output drivers of IC3 and to protect IC3 against a shorted output.

IC3 also contains a digital volume control which adjusts the headphone amplifier output. If VR7 is fitted to the board, IC1 detects this and sends commands to IC3 to set the headphone volume depending on the voltage at VR7’s wiper. If VR7 is not fitted, the headphone outputs are disabled and in that case, the other components in the pink box may be omitted.

**Controls and power supply**
Pots VR3 and VR4 are used to change the effect parameters. These form voltage dividers across the 3.3V supply rail, and the wiper voltage is read by IC1 using its internal analogue-to-digital converter (ADC).

The power supply is quite simple. D1 provides reverse polarity protection while REG1 drops the incoming 7.5-12V rail to a regulated 3.3V, as required by IC1 and IC3. LED1 indicates when power is applied. IC1 and IC3 have 100nF bypass capacitors for each pair of supply pins, plus a 10µF capacitor for IC1’s internal core regulator (on pin 56, Vcap).

CODEC IC3 also has 100nF bypass capacitors for IC1’s internal core regulator (on pin 56, Vcap). CODEC IC3 also has 100µF bypass capacitors for IC1’s internal core regulator (on pin 56, Vcap). CODEC IC3 also has 100µF bypass capacitors for IC1’s internal core regulator (on pin 56, Vcap). CODEC IC3 also has 100µF bypass capacitors for IC1’s internal core regulator (on pin 56, Vcap). CODEC IC3 also has 100µF bypass capacitors for IC1’s internal core regulator (on pin 56, Vcap).

**5V operation**
As with the Echo & Reverb unit, you can change some components to operate the unit from a 5V supply, such as is available from a USB port. This arrangement is shown in Fig.3. Basically, REG1 and its associated components are deleted and an LM3940 low-dropout 3.3V linear regulator is substituted. This is necessary because the LM317 used for higher voltage supplies drops...
too much voltage and can't operate from 5V. Also, D1 is replaced with a 1N5819 Schottky diode, which has a much lower forward voltage.

Construction
Fig.2 shows the PCB (available from the EPE PCB Service, code 01110131) parts layout. If building the 5V-powered version, refer also to Fig.4 for the necessary changes to fit the different regulator and Schottky diode (D1).

Start by fitting SMDs IC1 and IC3 (IC2 is left out). In each case, place the IC alongside its pads, right-side up and identify pin 1 (there should be a depression in one corner but magnification may be required to spot it). A pin 1 dot is also shown on the overlay diagram and PCB.

Apply a very small amount of solder to one of the corner pads. If you are right-handed, it’s easiest to start with the top pad on the right side, or if left-handed, with the top pad on the left side. Avoid getting any solder on the adjacent pad.

Now, pick up the part with a fine-tipped pair of angled tweezers and while heating this pad, gently slide the IC into place. Check the part’s alignment under a magnifying lamp. All the pins must be centred fairly accurately over their respective pads.

If they aren’t, don’t panic, it’s just a matter of re-melting the solder on that one joint and carefully nudging the IC in the required direction, then re-inspecting it. It may take a few attempts to get it correct. Care and patience are a virtue here, the goal being to eventually get it properly aligned without spreading solder onto any more pins or pads and without heating the PCB or IC enough to damage them.

Once the part is in place, solder the diagonally opposite pin, then re-check the alignment under magnification as it may have moved slightly. If it has, you can reheat this second pad and gently twist the IC back into alignment. Once you’re happy, proceed to solder the remaining pins without worrying too much about bridging them (it’s hard to avoid). Remember to refresh that first pin you soldered.

Now spread a thin layer of flux paste along all the pins and gently press down on them with solder wick and a hot iron to suck up the excess solder. If done correctly, this will leave you with neatly soldered pins and no solder bridges. Go over all the pins once with the solder wick, then check under a magnifier for any remaining bridges. If there are any, add a dab of flux paste and go back over them with the solder wick.

With all the joints looking good, you can install the other SMD IC using the

---

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same procedure. Note that a hot-air gun/toaster oven and solder paste can also be used for these ICs.

Once you’ve checked that the ICs are all soldered properly, follow with the SMD ceramic capacitors, using a similar procedure; ie, add solder to one pad, heat and slide the part into place, then solder the other pad and refresh the initial joint. Don’t get the 10µF capacitor mixed up with the others.

You do need to be careful to wait about 10 seconds after soldering one side of a capacitor before applying solder to the other side though. The capacitors are so small that the solder joint can remain molten for quite some time. If you try to solder the opposite pad too early, the capacitor will move out of alignment and it’s frustrating to re-align capacitors when this happens, especially if solder has taken to the other pad too.

So take it slowly and be careful not to short any of the adjacent IC pins when soldering the pads; the capacitors have been placed quite close for performance reasons. A fine soldering iron tip will make this easier.

**Through-hole parts**

Proceed now with the low-profile components such as resistors and diodes – remember to slip a ferrite bead over the 4.7Ω resistor lead before soldering it in place. It’s best to check each resistor value with a DMM before fitting it as the colour bands can be difficult to read. The diodes are all the same type and all have their cathode bands facing to the top or right edge of the board.

For FB2, slip another bead over a resistor lead off-cut and then solder it.

**Parts List**

<table>
<thead>
<tr>
<th>Description</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Capacitors</strong></td>
<td></td>
</tr>
<tr>
<td>2 100µF 25V electrolytic</td>
<td></td>
</tr>
<tr>
<td>6 100µF 16V electrolytic</td>
<td></td>
</tr>
<tr>
<td>1 22Ω 16V electrolytic</td>
<td></td>
</tr>
<tr>
<td>1 10µF 6.3V 0805 SMD ceramic</td>
<td></td>
</tr>
<tr>
<td>1 1µF 50V monolithic ceramic</td>
<td></td>
</tr>
<tr>
<td>11 100µF 6.3V 0805 SMD ceramic</td>
<td></td>
</tr>
<tr>
<td>1 1nF MKT</td>
<td></td>
</tr>
<tr>
<td>2 33pF ceramic disc</td>
<td></td>
</tr>
<tr>
<td><strong>Resistors (0.25W, 1%)</strong></td>
<td></td>
</tr>
<tr>
<td>1 47kΩ</td>
<td>1 120Ω</td>
</tr>
<tr>
<td>5 10kΩ</td>
<td>1 100Ω</td>
</tr>
<tr>
<td>1 1kΩ</td>
<td>1 4.7Ω 0.5W 5%</td>
</tr>
<tr>
<td>1 200Ω</td>
<td>1 3.3Ω 0.5W 5%</td>
</tr>
<tr>
<td><strong>Extra parts for headphone output</strong></td>
<td></td>
</tr>
<tr>
<td>1 10Ω linear potentiometer, panel mount (VR7)</td>
<td></td>
</tr>
<tr>
<td>1 small knob to suit</td>
<td></td>
</tr>
<tr>
<td>2 220Ω F 10V electrolytic capacitors</td>
<td></td>
</tr>
<tr>
<td>2 47kΩ 0.25W resistors</td>
<td></td>
</tr>
<tr>
<td>1 3-way pin header</td>
<td></td>
</tr>
<tr>
<td>1 100mm length 2-core shielded cable or 3-strand ribbon cable</td>
<td></td>
</tr>
<tr>
<td>1 100mm length 3-strand ribbon cable</td>
<td></td>
</tr>
<tr>
<td><strong>Extra parts for microphone input</strong></td>
<td></td>
</tr>
<tr>
<td>1 3.5mm panel-mount stereo jack socket</td>
<td></td>
</tr>
<tr>
<td>1 1µF multi-layer ceramic capacitor</td>
<td></td>
</tr>
<tr>
<td>1 220µF ceramic capacitor</td>
<td></td>
</tr>
<tr>
<td>1 47kΩ 0.25W resistors</td>
<td></td>
</tr>
<tr>
<td>1 680Ω 0.25W resistor</td>
<td></td>
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<tr>
<td>1 2-way pin header</td>
<td></td>
</tr>
<tr>
<td>1 length shielded cable</td>
<td></td>
</tr>
<tr>
<td>1 length light-duty hookup wire</td>
<td></td>
</tr>
<tr>
<td>1 metal case (optional)</td>
<td></td>
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<tr>
<td>Light duty hook-up wire/ribbon cable</td>
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</tr>
<tr>
<td><strong>Semiconductors</strong></td>
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<tr>
<td>1 PIC32MX470F512H-I/PT 32-bit microcontroller programmed with</td>
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<tr>
<td>120914A.hex (IC1) (the software</td>
<td></td>
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<tr>
<td>to programme your own PIC is</td>
<td></td>
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<tr>
<td>available from our website</td>
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<tr>
<td><a href="http://www.epemag.com">www.epemag.com</a>, alternatively</td>
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<tr>
<td>pre-programmed PICs are</td>
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<tr>
<td>available from <a href="http://www.siliconchip.com.au">www.siliconchip.com.au</a></td>
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<tr>
<td>1 WM8731SEDS 24-bit 96kHz stereo CODEC (IC3)</td>
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</tr>
<tr>
<td>3 1N4004 diodes (D1-D3)</td>
<td></td>
</tr>
<tr>
<td>2 1N4148 diodes (D4,D5)</td>
<td></td>
</tr>
<tr>
<td>1 LM317T adjustable regulator (REG1) (refer to text for parts</td>
<td></td>
</tr>
<tr>
<td>required for 5V DC operation</td>
<td></td>
</tr>
<tr>
<td>1 3mm blue LED (LED1)</td>
<td></td>
</tr>
<tr>
<td>1 double-sided PCB, available from the EPE PCB Service, coded</td>
<td></td>
</tr>
<tr>
<td>01110131, 148 × 80mm</td>
<td></td>
</tr>
<tr>
<td>1 12MHz HC-49 crystal (X1)</td>
<td></td>
</tr>
<tr>
<td>1 100Ω±5% axial RF inductor (L1)</td>
<td></td>
</tr>
<tr>
<td>2 10kΩ 5mm horizontal potentiometer (VR3,VR4)</td>
<td></td>
</tr>
<tr>
<td>1 5kΩ mini horizontal trimpot (VR6)</td>
<td></td>
</tr>
<tr>
<td>2 6.35mm PCB-mount stereo switch jack sockets (CON1,CON2)</td>
<td></td>
</tr>
<tr>
<td>1 10-way pin header, 2.54mm pitch (CON5)</td>
<td></td>
</tr>
<tr>
<td>1 5-way pin header, 2.54mm pitch (CON7) (optional)</td>
<td></td>
</tr>
<tr>
<td>1 PCB-mount SPDT right-angle toggle switch (S1)</td>
<td></td>
</tr>
<tr>
<td>1 chassis-mount NO momentary pushbutton switch (S2)</td>
<td></td>
</tr>
<tr>
<td>4-position rotary or slide switch (S3)</td>
<td></td>
</tr>
<tr>
<td>1 3-way pin header, 2.54mm pitch (for S4)</td>
<td></td>
</tr>
<tr>
<td>1 foot switch with cable (S4, optional)</td>
<td></td>
</tr>
<tr>
<td>1 DC plugpack, 7.5-12V, 100mA+</td>
<td></td>
</tr>
<tr>
<td>1 PCB-mount switched DC socket to suit plugpack</td>
<td></td>
</tr>
<tr>
<td>2 4mm ferrite suppression beads</td>
<td></td>
</tr>
<tr>
<td>9 M3 × 6mm machine screws</td>
<td></td>
</tr>
<tr>
<td>1 M3 nut</td>
<td></td>
</tr>
<tr>
<td>4 tapped spacers</td>
<td></td>
</tr>
</tbody>
</table>
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Checking it out

You can then fit crystal X1 and the electrolytic capacitors, of which there are three different values (four if using the headphone outputs). As usual, the longer lead is positive and this should go in the hole marked with the ‘+’ symbol on the overlay, ie, towards the top edge of the board.

Next, fit power switch S1 and the power LED. The latter should have its lead bent at right angles 4mm from the base of the lens and then soldered so that the centre of the lens (and thus this short lead section) is 6.5mm above the top surface of the PCB. This aligns the centre LED with the centre of the switch. When bending the leads, pay attention to the ‘A’ and ‘K’ markings on the PCB – the longer (anode) lead of LED1 must be soldered to the anode pad.

S1 and the power LED could also be chassis-mounted if you wish.

The PCB assembly can now be completed by soldering jack sockets CON1 and CON2 in place. You will also need to wire up a rotary or slide switch (S3) and a momentary pushbutton switch (S2) to a pin header socket, as shown in Fig.2.

If using the foot (defeat) switch, headphone or microphone options, wire them up too. We’ve shown the foot switch connected via a 3.5mm phono socket, but you could use a 6.35mm socket or some other connector instead.

Checking it out

If you used a blank PIC32 chip, program it now. The circuit can be powered from a PICKit3 at 3.3V. In fact, the whole unit will operate normally from this supply, so you can test the audio signal path immediately after programming the chip. Use the firmware for the Digital Effects Processor which is available from our website www.epemag.com named ‘0120914A.hex’.

If you don’t have a PICKit3, you will need to power the unit from a DC plug-pack for testing. In this case, connect a voltmeter across the 3.3Ω resistor next to D1. Small alligator clip leads (or other test probe clips) are very useful for this purpose as you can switch the unit on while watching the meter reading and switch it off quickly should the voltage across this resistor rise too high.

Expect a reading in the range of 0.2-0.3V, depending on the exact resistor value and how you have configured the unit. Much less than 0.2V indicates that there is an open circuit somewhere, while much more than 0.3V indicates a likely short circuit. If the reading is outside the expected range, switch off immediately and check for faults.

The most likely faults would be pins on the SMD chips bridged to an adjacent pin or not properly soldered to the PCB pad, followed by incorrect device orientation (primarily ICs, diodes and electrolytic capacitors) or poor/bridged through-hole solder joints.

Assuming all is OK, connect S3, set it to position 2 (no effect) and feed a signal into the input; if it’s a stereo plug, the left channel will be shorted out. Connect the output to some sort of amplifier. You should hear clear, undistorted audio with no effects. You can then try out the effects to check that they operate as expected.

Using effects

Initially, the effect for switch position #1 is echo, position #3 is reverb and position #4 is tremolo, so you can easily try these out. To adjust the parameters, hold down S2 and then rotate VR3 and/or VR4. Once you release S2, turning VR3 and VR4 will have no effect, so you can’t accidentally change the settings.

To assign a different effect to one of these switch positions, select that position and then give S2 a brief press without turning either VR3 or VR4. The unit will switch to the next effect and will emit a series of ‘pips’ from the audio output; one pip for effect #1 (echo), two for effect #2 (reverb) and so on. If you press S2 when effect #10 (phaser) is selected, it will switch back to #1 (echo).

The settings are remembered even when power is removed; they’re stored in Flash memory. If you press S2 in order to adjust VR3/VR4 and then decide against it, hold S2 down for a short period before releasing it to prevent an undesired change in the selected effect. Any press longer than about half a second will not cause the selected effect to change.
THE OPTO-THEREMIN’S main PCB is housed in a black UB1 plastic utility box measuring 158 × 95 × 53mm. This box is supported on a timber plinth (or base) using threaded rods and three 50mm lengths of 10mm ID aluminium tubing.

The first step is to prepare the box by drilling the various holes. We’ve prepared a template (in PDF format) to make this job easy. This can be downloaded from the EPE website and printed onto plain paper.

While you are there, download the front panel artwork and the drilling templates for the timber plinth and the smaller UB5 case. These can also be printed onto plain paper, but for a better result, print the panel artwork onto photographic paper.

Next, cut the case template sheet into its various sections, then attach the templates to the case (eg, using adhesive tape) and drill the holes to the dimensions indicated. Use a small pilot drill to start the larger holes, then carefully enlarge them to their correct sizes using larger drills and a tapered reamer.

Be careful not to over-enlarge the 10mm-diameter hole for the antenna. The aluminium antenna tube should be a tight fit into this hole.

Once the holes have all been drilled, the main label can be affixed to the lid using silicone sealant or a suitable adhesive. Allow the adhesive to dry, then cut out the various holes using a sharp craft knife. The speaker can then be secured to the inside of the lid by smearing a suitable adhesive (eg, super glue) around its outside metal frame.

Once the speaker is in place, it can be fitted with a short figure-8 connecting cable terminated in a 2-way header plug at the far end.

The main PCB is fitted into the box by first tilting it down at the front, so that the pot shafts and the switch can be slid into their respective holes. The rear of the PCB can then be pushed down into the case, after which the . . .

By JOHN CLARKE

Fig.7: here’s how to install the rear spacer assemblies. No spacers are required at the front of the case, since the PCB is supported along this edge by the two pot shafts.
assembly should be secured in position by attaching the nuts to the pots. Do the nuts up firmly, then mark out the positions of the two rear mounting holes on the base of the case (e.g., by hand-twisting a 3mm drill through the PCB holes).

That done, remove the PCB and drill these holes in the base out to 3mm. There’s no need for corresponding front mounting holes, since the PCB is supported on this side by the pot shafts.

The rear spacer assemblies can now be installed, as shown in Fig.7. First, an M3 × 6mm screw is inserted up through the bottom of the case. This is then secured with an M3 nut, after which an M3 × 9mm tapped spacer is fitted.

Don’t reinstall the PCB yet – that step comes later, after attaching the case to the timber plinth.

**Making the timber plinth**

A piece of 151 × 90 × 19mm DAR (dressed all round) pine timber is used to make the base – see Fig.8. **Note that Fig.8 is not to scale, so you should download the full-size diagram from the EPE website and print it out to use as a template.**

Cut the timber plinth to size, then round off the edges and the corners using sandpaper. The paper template can then be attached to the base and the three holes drilled to accept either M5 or 3/16-inch threaded steel rod (zinc-plated).

Countersink the holes on the underside to allow the nuts to be recessed. Fig.8 shows the cross-sectional view (two rods only shown).

The timber base is now used as a template to mark out the corresponding holes in the bottom of the case. Drill these to suit the threaded rod, then cut the threaded steel rod into three 75mm lengths. You will also need to cut three 50mm lengths of 10mm-diameter aluminium tubing, to serve as spacers.

It’s a good idea to paint the timber base black to match the box colour. After that, it’s just a matter of attaching it to the case using the 75mm threaded rods, 50mm aluminium tube spacers and nuts, as shown in Fig.8.

Take care to ensure that the threaded rod protrudes no further into the box than the nut, otherwise it may later short against the tracks on the underside of the PCB.

The PCB can now be reinstalled in the case and secured to the previously installed rear spacers using M3 × 6mm machine screws. Tighten these screws down firmly, then install the pot nuts and fit the two knobs. If the knob pointers are in the wrong positions, prise the end caps off and refit them so that they are correct.

**Volume control case**

The volume control PCB is housed in a transparent blue UB5 plastic utility box measuring 83 × 54 × 31mm. A rectangular cut-out has to be made in the base (which becomes the top) to accept the distance sensor, while five holes have to be drilled in one end for the external wiring connections and two threaded mounting rods.

As with the larger case, it’s just a matter of attaching the drilling template downloaded earlier and then drilling the holes to the sizes indicated. The rectangular cut-out is made by drilling small holes around the inside perimeter, then knocking out the centre piece and filing to shape.

**Be sure to make this cut-out in the base (not the lid). The case is later...**

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attached to the main case with the base facing upwards and the lid on the bottom.

Refer now to Fig.9 to see how the volume control case is attached to the main case. The first job is to cut two 62mm lengths of M5 or 3/16-inch threaded rod plus two lengths of 10mm OD aluminium tube. These aluminium tube pieces should be 50mm long minus the width of the nuts used (eg, if the nuts are 4mm wide, then cut two 46mm tube lengths).

Once you have all the pieces, attach the two threaded rods to the volume control case, as shown in Fig.9; ie, for each rod, use a nut inside the case and another outside the case. We used nylon lock nuts (metal nuts with a nylon thread insert) because they each have a rounded end that the aluminium rod fits over and because they don’t come undone.

The next step is to fit three straight 88mm lengths of 1mm-diameter steel wire to CON5 on the volume control PCB. These wires are then slid into the holes in the end of the case (between the nuts securing the threaded rods) and the PCB clipped into place (ie, into the integral ribs).

If you can’t get steel wire, use 1mm-diameter tinned copper wire. This can be straightened by clamping one end in a vice and then stretching it slightly by pulling on the other end with a pair of pliers.

Final assembly
The volume control case assembly can now be attached to the main case as follows:

1) Cover the front threaded rod (ie, at the bottom in Fig.9) with a length of 6mm-diameter heatshrink tubing. This heatshrink layer should cover the entire length of the thread and can be trimmed to size after shrinking it down.

2) Cut another length of heatshrink tubing about 3mm shorter than the aluminium tubing and add this to the rod. Push it all the way up.

The PCB is installed in the case by first angling it down at the front and sliding the pot shafts and the switch actuator into their respective holes. The rear of the board is then slid down into position and the PCB secured by doing up the pot nuts and fitting the screws to the rear spacer assemblies.

The speaker is secured to the inside of the lid by smearing super glue or silicone around its outside metal frame.
against the nut at the volume control case end before shrinking it down.
3) Repeat step 2, add more heatshrink layers until the aluminium tube is a firm fit over this threaded rod.
4) With the aluminium tubes in place, insert the three wires and the threaded rods into the main case, with the ends of the wires going into CON2. The heatshrink-covered rod should be a tight fit into its hole.
5) Secure the other threaded rod with a nut on the inside of the main case.
6) Tighten CON2’s screws to secure the three wires in place.

Making the pitch antenna
The pitch control antenna is also made from 10mm-diameter aluminium tubing. You also need an M4 × 10mm nylon (or polycarbonate) screw and two M4 nylon (or polycarbonate) nuts.

First, cut a 450mm length of the tubing and clean up the ends with a file to remove any metal burrs. That done, gently file each corner of one of the M4 nuts until it fits tightly into one end (ie, the top) of the antenna. Once it’s in position, wind the second M4 nut all the way onto the screw and then screw this into the captive nut in the antenna. This translucent ‘top piece’ provides the blue glow at the top of the antenna when lit by LED3 on the main PCB (ie, the blue LED that shines up the antenna tube).

The other three blue LEDs (LEDs1, 2 and 4) light the base of the antenna. As an option, these three LEDs can be covered with a translucent, hemispherical, half of a hollow ball that’s slid over the antenna and pushed down onto the lid of the main case.
A ball salvaged from an empty can of roll-on deodorant is suitable. All you have to do is cut the ball in half using a fine-blade hacksaw, file the ends to

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Constructional Project

An M4 nylon (or polycarbonate) nut is pushed into the top of the pitch antenna, after which a nylon M4 screw with captive nut is fitted. This translucent assembly glows blue when lit by the LED shining up the aluminium tube.

The pitch antenna is pushed into the two fuse clips on the main PCB assembly (usually after the lid is in place). LED3 is between the two fuse clips and shines up the antenna tube to light the translucent screw assembly at the top.

As previously stated, the bottom end of the antenna is connected into the circuit by sliding it into the two fuse clips on the main PCB. It may be necessary to squeeze the lugs of these fuse clips together slightly so that the antenna makes a good contact. Rotating the antenna a few times will also clean the contacts if they oxidise over time.

For the time being, leave the lid off the case and simply support the antenna in its fuse clips. You are now ready for the setting-up procedure.

Setting up
The adjustment procedure is as follows:
Step 1: fit link LK1 (near the equalising coil) to the TEST position and LK2 to the MAX position.
Step 2: connect a 9VAC plugpack or a 12V DC source, switch on and check that all the LEDs light. If they don’t light, check that they are oriented correctly.
Step 3: connect a DMM set to read DC volts between TPS (near IC2) and TP GND and adjust trimpot VR4 for a reading of 1.7V.
Step 4: connect the DMM between TP1 and TP GND and adjust the slug in transformer T1 for a minimum reading (note: do not use a screwdriver as this could crack the ferrite core. Either use the correct plastic alignment tool or grind down an old screwdriver so that its blade is thicker than normal and snugly fits the slot in the slug).
If you are unable to find the minimum, then either coil L1 has been incorrectly wound or its leads haven’t been soldered. Check the solder joints and check also that the nylon washer spacers have been installed to provide the required 2.5mm gap between the two core halves.
Step 5: move your hand very close to the antenna (but don’t touch it) and adjust T1’s slug so that the voltage slightly increases. When it does, move your hand away from the antenna and check that the voltage increases even further.
If the voltage decreases instead, then the slug needs to be rotated the other way. On the prototype Opto-Theremin, we adjusted T1’s slug for 1.1V with the hand close to the antenna and 1.7V with the hand away from the antenna.
Step 6: move jumper LK1 to the NORMAL position, connect the loudspeaker to CON4, set VR1 to mid-position and set VR2 fully clockwise.
Step 7: adjust transformer T2 until a tone is heard and set it for a low frequency. This tone should then change if you move your hand away from T2 (and away from the antenna), so this may take some trial and error.
Step 8: rotate VR2 anticlockwise and check that the pitch can be adjusted to just reach a point where there is no sound. The frequency should then become audible again and increase as a hand is brought close to the antenna. If not, reset VR2 fully clockwise again and repeat Step 7, this time adjusting T2’s slug in the opposite direction.
Note that these adjustments require patience and you may need to repeat the process several times before you get it right.
Step 9: adjust VR1 fully clockwise, then adjust trimpot VR3 to limit the volume so that it isn’t high enough to cause spurious vibrations or noticeable distortion.

Voicing adjustment
Trimmer capacitor VC1 must now be adjusted to set the voicing. It’s just a matter of tweaking it to obtain the required sound from the Theremin.
Note that there will be a point where, at the lowest frequencies, there’s a ‘snap-on’ effect whereby either no frequency is produced or the tone suddenly snaps on and becomes audible with hand movement. This occurs because inter-coupling between the pitch and reference oscillators causes both oscillators to track together, and if there’s no frequency difference between them, there’s no audible output from the mixer. However, as a hand is brought closer to the antenna, the pitch oscillator’s tuning changes and it is eventually ‘pulled’ far enough to suddenly produce a different frequency to the reference oscillator.

Hand volume adjustment
The hand volume adjustments are all done on the main PCB as follows:
Step 1: move jumper LK2 back to the NORMAL position, then check that the volume control has a suitable hand-movement range. The volume should increase as the hand is moved away from the sensor and vice versa.
Step 2: if you want to change the range, connect a DMM between TPS and TP GND and adjust trimpot VR4 for a reading that differs from the 1.7V set earlier. Note, however, that if VR4 is set to give maximum volume too far away from the sensor, the volume will rise again at close range (ie, as the hand is brought below 40mm). This is a quirky effect of the sensor itself and is cured simply by backing off the setting for VR4.
That completes the adjustments. You can now complete the unit by attaching the lid to the main case and reinstalling the antenna, with the translucent dome slid all the way down so that it covers the three LEDs. Take care when fitting the lid to ensure that the four LEDs go through their corresponding holes. You will find it easier to do this if you apply power so that the LEDs are lit.
Finally, fit the lid to the underside of the volume control box and your Opto-Theremin is ready for action.

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June 2015 ISSUE WINNER
Mr Colin Wedgbury, who works at Suetronics, UK.
He won a PIC32 Bluetooth Starter Kit valued at £53.
Most modern cars have a courtesy light delay but older vehicles do not. This new circuit is specifically designed to suit LED lamps but will also work with conventional filament lamps. It keeps the interior lights of your car lit for a preset time after you shut the car doors. The lights will also turn off if the exterior lights or ignition are switched on during the time-out period.

If you build this courtesy light delay unit, you will be able to upgrade your vehicle to LED interior lighting. LEDs give much improved lighting compared to the yellow/orange of incandescent lamps, and the bulb diffusers will not discolour with age.

Examing older systems (designed for filament bulbs) a little analysis shows that they won’t work with LED lighting unless you have at least one filament bulb connected; not the best compromise.

To understand why this is the case, look at Fig.1, which shows a typical filament bulb design concept. It’s based on a MOSFET (Q1), two capacitors (C1 and C2) and a 1MΩ discharge resistor. When the door switch is closed, the interior lamp(s) light and the capacitors are discharged. The instant the door switch opens, the two capacitors charge via the filaments in the interior lighting.

Due to the different values of the two series-connected capacitors, the 47µF capacitor (C2) will charge to a voltage that’s about 10-times higher than the voltage across the 470µF capacitor (C1). So with a 12V supply and taking into account the 0.7V drop across diode D1, the 47µF capacitor will have about 10.2V across it and the 470µF capacitor about 1.02V. The 10.2V across C2 becomes the gate voltage for the MOSFET, which then drives the lamps. After a short time, the gate voltage discharges via the 1MΩ resistor and the lamps go off.

As shown in Fig.1, a few refinements were often included. These include adding a short time delay to prevent MOSFET Q1 from switching on instantly when the door switch opens. This is to allow time for capacitors C1 and C2 to charge sufficiently before the MOSFET switches on and shunts the door switch. This delay is achieved using transistor Q2, which is momentarily switched on at power-up (ie, when the door switch opens) due to base drive through the 100nF capacitor and 10kΩ resistor.

When Q2 switches on, it momentarily shunts Q1’s gate to ground. This prevents Q1 from conducting until the 100nF capacitor charges. The duration is only 1ms and any tendency for the lamp to briefly flicker off as the door switch opens is virtually unnoticeable.

Such designs often included additional circuitry to switch off the MOSFET (and thus turn the interior lamps off) if the tail lights were activated (ie, if the parking lights or headlamps were switched on).

As stated, this circuit doesn’t work with LED lighting. That’s because the circuit relies on current flowing through the lamp filaments, just after they are switched off, to charge capacitors C1 and C2. Typically, a 5W lamp filament will have a resistance of about 29Ω when it is hot and so the 47µF capacitor takes much less than 1ms to charge. However, interior lighting often uses more than one lamp and so the charging resistance is usually much lower than 29Ω.

By contrast, typical 12V LED lamps incorporate two or three white LEDs connected in series with a current-limiting resistor. The voltage drop across each LED is typically 3.5V for a white LED and so the total voltage drop is around 7V with two in series, or about 10.5V with three in series. So there is not much load of the 12V supply to charge the capacitors shown in Fig.1.

When twin-LED lamps are used in this circuit, the resulting gate voltage will be around 2.9V when using a standard diode for D1 and 3.13V when...
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Other uses

This PCB module is not just confined to vehicle use. Its circuit can also be used for timed lighting, such as in a hallway, provided you use 12V LEDs run from a 12V DC supply. A pushbutton momentary switch would be used to switch the lights on and they would then turn off automatically at the end of the preset period.

This would also be ideal for a stairwell with one or several pushbutton start switches (eg, one on each floor). With a 12V supply, up to 36W (3A) of lighting can be controlled and these could be powered from a 3A 12V power brick or similar (or use a 2A plug-pack for up to 24W of lighting).

Note that the pushbutton switch needs to be rated for the total current drawn by the LED lighting.

using a Schottky diode. We do not get the expected 3.8V because of the voltage drop across the current-limiting resistor in the LED lamp.

Now, 3.13V is too low to fully switch on most MOSFETs, including typical logic-level types that can conduct (at least partially) with a 3V gate-to-source voltage but switch off below 2.5V. This means that LED lamps will not be correctly switched on by the circuit of Fig.1. Even if we substitute a MOSFET with a very low on-threshold voltage, it would be difficult to get a consistent delay period due to the low capacitor voltages compared to this threshold.

It's unfortunate that the filament-type circuit doesn't work with LED lighting because it has several desirable features. First, there's no need to connect it directly to the vehicle's 12V supply; you just connect across a door switch (in a vehicle with incandescent interior lamps) and it works. In addition, the circuit will operate regardless as to whether the door switch is on the negative side of the lamp (Fig.2a) or the positive side (Fig.2b). Provided it's connected with the correct polarity across the door switch, the circuit works in exactly the same manner for both 'high-side' and 'low-side' switching.

So how do we design a circuit to operate with LEDs? In this case, we need to connect our new circuit directly to the 12V supply as well as to a door switch. And if we want the interior lamps to switch off when the parking lights or ignition are turned on, then these too need to be monitored by the circuit.

LED version

Our Courtesy LED Lights Delay circuit is shown in Fig.3. Unlike the filament circuit, it also monitors the ignition as well as the exterior lights. Monitoring the parking lights or tail lights is only useful for night-time driving, since you are unlikely to use the lights during the day. By monitoring the ignition line, the courtesy lamps will immediately go out if the car is started rather than having to wait for the delay period to expire.

As with our previous circuit, the Courtesy LED Lights Delay operates with the door switch in either configuration (ie, high-side or low-side). Again, it’s only necessary to wire the circuit to a door switch with the correct polarity. It’s not necessary to know how the door switch is connected in the vehicle; you just have to identify its positive and negative leads.

The other connections to the circuit are to +12V, chassis (0V), ignition and the switched supply for the vehicle’s exterior lights. The lights connection can be regarded as optional; in many cases, it will be sufficient to simply monitor the ignition line to automatically turn the interior lamps off before the delay period has ended (ie, when the car is started). The lights input connects across the parking lights or tail lights (but not the stop lights) and can be connected with either polarity.

If the courtesy lights use a low-side switching arrangement, MOSFET Q1’s...
source terminal will be connected to ground via SWITCH–. But this won’t be the case with high-side switching.

You might expect that this could be solved by driving Q1 with an IR2125 (or similar) MOSFET driver, which could produce a suitable gate drive above the MOSFET’s source voltage, whether that rises to the 12V supply (for a high-side connected MOSFET) or 0V (for a low-side connected MOSFET).

However, in the high-side configuration, this scheme relies on a low-impedance source load such as a light bulb to charge the boost capacitor during the MOSFET’s off-time. This capacitor is subsequently used to generate a voltage above the 12V supply when the MOSFET switches on, so that it remains in conduction.

Once again, using LEDs for the load will mean that the capacitor will only charge to 12V minus the voltage drop of the LEDs. Ultimately, we would still be restricted to only a couple of volts for the MOSFET gate supply, so it won’t work for the same reasons outlined earlier. Another problem is that the IR2125’s quiescent current is rather high, at up to 1.2mA.

To get around this problem, our circuit is based on a PIC12F675-I/P microcontroller (IC1) and this drives MOSFET (Q1) via transformer T1. IC1 produces a 1MHz square-wave to drive the transformer and it provides a timing function to switch off this signal after a set period (the delay). This delay period can be adjusted using trimpot VR1.

In operation, microcontroller IC1 detects when a door switch is opened to start the delay period. It also monitors when the ignition or lights are switched on to cancel the delay period.

Because the circuit is always connected to the vehicle’s 12V battery, it’s vital that microcontroller IC1 has a low quiescent supply current. As a result, IC1 is normally in ‘sleep’ mode and draws negligible current (up to...
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2µA maximum). In fact, most of the quiescent current is drawn by 3-terminal regulator REG1, as described later. IC1’s GP2 input indirectly monitors the door switch, which is wired across the MOSFET. As shown, Q1’s drain connects to the positive side of the switch, while its source connects to the negative terminal.

GP2 is normally held high via an internal pull-up resistor. When the door switch is in the ground side (see Fig.2a), a closed switch pulls GP2 low via a 1kΩ resistor and diode D5. At the same time, transistor Q3 will be off since Q1’s source is at ground and so Q3’s base is held at 0V.

Alternatively, if the door switch is connected in the positive side of the supply, as in Fig.2b, a closed switch drives Q3’s base via a 10kΩ resistor. As a result, Q3 turns on and pulls GP2 low. In this case, diode D5 is reverse biased as Q1’s drain is connected to the positive supply.

So, for either connection of the door switch, IC1’s GP2 input is high when the switch is open and goes low when the switch closes. IC1 is configured to generate an interrupt on a positive edge at input GP2 and when the door switch subsequently opens again, this interrupt wakes IC1 from its sleep mode. The microcontroller’s firmware then starts an internal oscillator and this produces a 1MHz clock output at pin 3. This then drives transformer T1 via a 100nF capacitor.

Diodes D1-D4 rectify the voltage from T1’s secondary and the resulting DC is filtered by a 1µF capacitor. This in turn switches on MOSFET Q1 to drive the interior lights, just as if a door switch was closed. Note that D1-D4 are 1N4148s because a standard bridge rectifier would not work at 1MHz.

The end result is that Q1’s gate is charged sufficiently above its source to ensure it switches on, regardless of whether the source voltage is actually 0V or 12V. This configuration is known as a ‘floating gate supply’. At the same time as Q1 is switched on, IC1’s GP0 output is taken low (to 0V) and this connects a 5V supply across trimpot VR1 (100kΩ). The setting at VR1’s wiper is then read via IC1’s AN1 input. The GP0 output is then taken high to stop the current flowing through VR1 and this is done to minimise the current drain, particularly during sleep mode.

IC1 goes to sleep again at the end of the time-out period, as set by VR1. This stops the 1MHz drive to transformer T1 and MOSFET Q1 then quickly dims the interior lights over a nominal one-second period as its 1µF gate capacitor discharges via a parallel 1MΩ resistor. Basically, the MOSFET’s internal impedances rises in response to decreasing gate voltage, thereby dimming the lights until they are ultimately completely off.

Interrupting the delay
IC1 monitors the ignition and taillights circuits via its GP5 input at pin 2. If either the ignition or lights are switched during the time-out period, the PIC immediately goes to sleep and the interior lights go out.

In greater detail, GP5 is normally held high via an internal pull-up resistor. If the ignition is switched on (eg, when the car is started), it drives the base of Q2 via a 10kΩ resistor. Q2 thus turns on and pulls GP5 (pin 2) of IC1 low to put the micro to sleep.

Alternatively, if the external lights are switched on, the resulting 12V DC supply is fed through bridge rectifier BR1 and drives the LED in optocoupler OPTO1. This in turn switches on OPTO1’s output transistor, again pulling GP5 (pin 2) of IC1 low and putting the micro to sleep.

BR1 and optocoupler OPTO1 ensure that the lights circuit will work regardless of how they are switched in the vehicle. It doesn’t matter whether the lights are ground connected and switched to positive or connected to positive and switched to ground.
Software
Not much is required in the way of software for IC1. As stated, it includes a rising-edge interrupt handler that wakes the PIC from sleep whenever a door switch is opened from its closed position. The PIC’s internal oscillator is then automatically started and it generates the 1MHz clock signal at pin 3.

The delay counter is set from 1-133s, depending on the 8-bit ADC reading from AN1 and this period is timed using the overflow period of the internal 16-bit timer (timer 1), which occurs every 524ms.

When the delay counter reaches zero, the PIC is placed back into sleep mode so that it draws minimal power and the 1MHz clock signal ceases. During the delay period, the GPS input is monitored and if this goes low, the processor is immediately placed in sleep mode and the LED lights quickly dim to off.

Construction
The Courtesy LED Lights Delay is built on a double-sided PCB, available from the EPE PCB Service, coded 05109141 and measuring 71 x 47mm. This clips neatly into the side channels of a UB5 plastic case and there is sufficient room to install a cable gland at the terminal block end.

Fig.4 shows the parts layout on the PCB. Install the resistors first, followed by diodes D1-D6 and zener diode ZD1.

Check each resistor with a multimeter before soldering it in position and make sure that the diodes and zener diode go in with the correct polarity. The zener diode is a 30V type and will probably be marked as a 1N4751.

OPTO1, the 4N25 optocoupler, is installed next, along with an 8-pin DIL socket for IC1. Be sure to orient these parts as shown on the overlay (ie, pin 1 at top left). Transistors Q2 and Q3, regulator REG1 and bridge rectifier BR1 can now go in. Check that the LM2950-5.0 device goes in the REG1 position and check that BR1 is correctly oriented and sits flush against the PCB before soldering its leads.

The capacitors are next on the list. Watch the orientation of the electrolytics and make sure that their tops are no more than 12.5mm above the PCB, otherwise they will later foul the lid of the case. The parts list shows the codes used for the 100nF and 1nF capacitors.

Connector CON1 is made up using one 3-way and two 2-way screw terminal blocks. These should be dovetailed together to form a 7-way block which is then mounted on the PCB with the wire entry holes facing towards the adjacent edge. Make sure that this 7-way connector sits flush against the PCB before soldering the pins.

Table 1: Resistor Colour Codes

<table>
<thead>
<tr>
<th>No.</th>
<th>Value</th>
<th>4-Band Code (1%)</th>
<th>5-Band Code (1%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>1MΩ</td>
<td>brown black green brown</td>
<td>brown black black yellow brown</td>
</tr>
<tr>
<td>2</td>
<td>10kΩ</td>
<td>brown black orange brown</td>
<td>brown black red brown</td>
</tr>
<tr>
<td>1</td>
<td>4.7kΩ</td>
<td>yellow violet red brown</td>
<td>yellow violet black brown</td>
</tr>
<tr>
<td>3</td>
<td>1kΩ</td>
<td>brown black red brown</td>
<td>brown black black brown</td>
</tr>
<tr>
<td>1</td>
<td>100Ω</td>
<td>brown black brown brown</td>
<td>brown black black black brown</td>
</tr>
</tbody>
</table>
MOSFET Q1 is mounted horizontally on the PCB with its metal tab secured to an M3 × 6mm nylon spacer. To do this, first bend the MOSFET’s leads down through 90° about 1mm from its body, then fit the MOSFET in position and slide the nylon spacer into position under its tab. The assembly is then secured to the PCB using an M3 × 12mm screw and nut – see Fig.5.

Note that this screw must be inserted from the underside of the PCB, so that the nut goes on top of Q1’s tab. That’s because the screw head is small enough not to foul adjacent PCB tracks, whereas the larger nut would run the risk of shorting out an adjacent 1µF capacitor. Do the screw up firmly to secure the assembly, then solder the MOSFET’s leads to the PCB.

Winding T1

The PCB assembly can now be completed by winding and installing transformer T1. This transformer consists of two windings on a ferrite ring core, as shown in Fig.4. The first winding consists of nine turns of 0.8mm enamelled copper wire, while the second consists of 24 turns of 0.8mm enamelled copper wire. They are wound on opposite sections of the core and it doesn’t matter in which direction they are wound.

Once the windings are in place, position the toroid on the PCB. The 9-turn winding goes through pads at the lower right-hand side of the PCB, while the 24-turn winding goes to a pad just to the right of Q3 and to a pad at top right. Push the toroid all the way down onto the PCB, then secure it in position using a couple of cable ties. These pass through the centre hole of the toroid and through adjacent holes on either side.

Note that the enamel coating will need to be scraped off the wires before soldering them to the PCB.

Testing

Before installing the PIC micro, connect a 12V supply to CON1 and check that there is about 5V (4.85-5.15V) between pins 1 and 4 of IC1’s socket. If this voltage is correct, disconnect the power and install IC1 with its pin 1 to the top left. If the voltage is incorrect, check the orientation of D6, the value of the series 100Ω resistor and that REG1 is an LM2936-5.0.

If you have a spare 12V LED lamp, this can be used to test the circuit before installing it in the vehicle. Do not use a white LED on its own. It must be a LED lamp with a limiting resistor to keep the current to a safe level for the LEDs.

Assuming you have a spare LED lamp, connect it between the switch minus terminal (pin 2 of CON1) and 0V (pin 4 of CON1). Note that the polarity is important here – the anode or positive side of the LED lamp must go to the switch minus terminal.

Now reapply power – the LED lamp should light for a second or so, then quickly dim to off. If that checks out, momentarily bridge the switch terminals on CON1 (pins 1 and 2). The LED lamp should now light for the length of time set by trimpot VR1 (note: timing begins when the door switch opens; ie, when the door is closed).

Assuming it works as expected, VR1 can now be adjusted to set the required delay. This ranges from 1-133s but note that the circuit’s response to the trimpot setting is non-linear. The fully anticlockwise to mid-position setting has the range of 1-33s, while the next half of the travel is divided into two equal sections. The first section has a range from 33-66s, while the remaining clockwise section sets the delay from 66-133s.

Having set the delay period, you can test that the ignition input works. That’s done by first triggering the delay period, then connecting the ignition input (pin 5 of CON1) to +12V using a wire link. When you do this, the lamp should extinguish after one second or so.

Note that you will need to set a reasonably long delay time for this test, to give yourself time to connect the ignition input to the +12V terminal.

Similarly, you can check that the lights inputs work by triggering the delay and connecting either pin 6 or pin 7 of CON1 to +12V using a wire link. When you do this, the lamp should turn off after a second or so.

If you wish, you can increase this 1s dim-down period by increasing the 1µF electrolytic capacitor value at the Q1 gate. It will be around 10s with a 10µF capacitor. Note that this dimming period is additional to the time-out or delay period. So if the time-out is set at 15s, the overall LED ‘on-period’ will be 15s plus the dimming period. For 1s dimming, the total time-out will be 16s.

Final assembly

If it all works as expected, drill a 12.5mm hole in the end of the UB5 box for the cable gland. This hole should be positioned 13mm down from the top of the box and centred horizontally. Drill a small pilot hole to begin with, then carefully enlarge it to size using a tapered reamer and mount the cable gland in position.

The assembled PCB is now simply clipped into the UB5 box with CON1 adjacent to the cable gland. This gland clamps the external wiring cable to prevent the connecting wires from being pulled out of CON1.

Installation

To connect the unit, you will need to access one of the door switches, +12V power, the ignition line and either the tail light or parking light connections. Alternatively, you may wish to just use the ignition input and not bother with the lights input.

Note that some door switches will have two wires while others have only a single wire connection. In the latter case, one contact is connected directly to chassis at the switch mounting position.

It’s important to get the door switch connections to the unit the right way around. The positive door switch connection must go to the switch positive of the Courtesy LED Light Delay. You can quickly determine which is the positive door switch connection by using a multimeter to measure the voltage across the door switch when it is pushed open.

Note that if there’s only a single wire running to the switch, this will be the positive (assuming the chassis connection is negative).

For the +12V supply rail, you will need to find a source of +12V that remains on when the ignition is off. This +12V supply rail must be protected by a fuse in the vehicle’s fusebox and is best derived at the fusebox itself. The 0V lead can be run to an eyelet connector that’s screwed to the chassis.

The lights terminals on the Courtesy LED Lights Delay are connected across one of the tail lights or parking lights. You can access this wiring either directly at the lights socket wiring, at the lights switch or in the fusebox. It doesn’t matter which way around you connect them, since the bridge rectifier automatically caters for both polarities.

Once you have found the relevant wiring points, it’s a good idea to disconnect the vehicle’s battery before running the wiring, to guard against

Everyday Practical Electronics, October 2015
any inadvertent short circuits. Note that all wiring should be run using proper automotive cable and connectors.

Once the wiring is complete, reconnect the battery and check that the courtesy lights remain on after the door is closed. Now turn the ignition (or the exterior lights) on and the courtesy lights should quickly dim to off (over 1s or so).

Finally, the unit can be mounted in any convenient location under the dashboard. It’s up to you how you secure it, since a suitable position will vary from vehicle to vehicle.

Existing delay circuit
What if your vehicle already has a courtesy lights delay? This may work fine if you substitute LED interior lamps for your car’s original incandescent lamps, but there’s always a possibility that it may not. In that case, you may wish to use the Courtesy LED Lights Delay instead.

One problem here is that the door switches will probably be connected to the existing delay circuit rather than directly to the interior lamps. Bypassing this delay circuit will therefore involve disconnecting all the door switches and wiring them directly to the interior lamps instead.

That’s too complicated (and time consuming) to be practical in most cases, but there is a way around this – keep the original delay circuit and simply add the Courtesy LED Lights Delay unit to the existing installation. That’s done by connecting the delay unit in parallel with the existing unit across one of the door switches.

There’s just one wrinkle to watch out for here – the original delay circuit may pull one side of the door switch to +5V rather than +12V. This should be checked using a multimeter and if it does go to +5V, the 1kΩ pull-down resistor connected to Q3’s base will have to be increased to 10kΩ (otherwise the transistor won’t turn on).

Note that, depending on the circuit used, the original delay period may be added to the delay introduced by the Courtesy LED Lights Delay unit. That won’t be a problem, however, since the Courtesy LED Lights Delay period can be adjusted down to as low as 1s.

Note also that connecting the Courtesy LED Lights Delay in parallel with an existing delay circuit may not work in all cases. It will very much depend on the vehicle and the circuit used.

Troubleshooting
If the courtesy lights are always on, the door switch terminals have probably been connected to CON1 (at pins 1 and 2) with reverse polarity. If that happens, the courtesy lights turn on via the intrinsic reverse diode inside MOSFET Q1 and simply swapping the leads to the door switch will fix the problem.

Finally, if the interior lights switch off immediately after the door is closed (and the connections are correct) check that there is no voltage applied to either the lights terminals or the ignition terminal on CON1 (pin 5, 6 and 7).

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Teach-In 2015
Discrete Linear Circuit Design
Part 9: Bringing it all together
by Mike and Richard Tooley

Welcome to Teach-In 2015. This series is aimed at anyone wishing to develop a detailed understanding of linear discrete semiconductor devices and how they are used in a diverse range of circuits. We hope you will join us on this exciting voyage of discovery!

Introduction
In this, the penultimate part of Teach-In 2015, we shall be bringing together a number of ideas and concepts from previous months. Knowledge Base takes a detailed look at stability, thermal and over-current protection, while Get Real is devoted to the design and construction of a low-cost, high-quality 10W power amplifier that will out-perform most of today’s similarly rated integrated circuit amplifiers. This final practical project is designed for use with previous Get Real circuits and it will help you to build your own good-quality modular audio system.

Knowledge base: Improving amplifier performance

Loudspeakers as loads
Earlier in this series we considered the load present at the output of an amplifier, and the power that would appear in it. However, when we did this we assumed that the load was well-behaved and just acted like a pure resistor and had no reactive components. Unfortunately, in the real world this is very far from the truth since most loudspeakers exhibit an impedance that varies widely over the audio frequency range. To illustrate this, we carried out some measurements on a popular Sharp XL-60H 40W loudspeaker system (see Fig.9.1). The impedance/frequency characteristic of this relatively low-cost system is shown in Fig.9.2a.

There are a few important things to note from Fig.9.2a. First, the distinct and relatively sharp peak in impedance at around 150Hz and second, the steady increase in impedance that occurs above 4kHz. That said, the system manages to maintain its nominal 4Ω impedance from 20Hz to 120Hz and from 250Hz to 4kHz. These variations of impedance are ‘seen’ by the output of the amplifier, but they are some way from the nominal 4Ω that might at first be assumed.

So, how do these undesirable fluctuations arise and what can be done to minimise them? To partly answer this question, we carried out some measurements on the Sharp XL-60H loudspeaker using an accurate universal LCR bridge. Our measurements revealed that the loudspeaker had a DC resistance of almost exactly 4Ω, together with a series inductance of 130µH. This latter value was in very close agreement with 129µH calculated from our previous impedance measurements.

Resonance
The low-frequency peak (around 180Hz) in impedance results from resonance in the low-frequency driver unit. The size of the resonant peak and the frequency at which it occurs is determined not only by the characteristics of the driver (mass, suspension) but also on the enclosure in which the driver is placed. The free-air resonance is often quoted in a driver’s specification and the size of the resonant peak relates to the Q-factor of the unit. The larger the Q-factor, the sharper the peak will be. When a driver is placed in an enclosure, its resonant frequency will increase along with its Q-factor. A typical increase in resonant frequency might be around 30Hz for a small sealed enclosure fitted with a 6-inch driver. At the same time, the Q-factor might increase from around 1.2 to a little more than 1.5. It should go without saying that loudspeaker design is a complex art, and in most cases it can be instrumental in making significant improvements to the quality of the sound produced by an audio system.

Voice coil reactance
The steady increase in impedance above 4kHz is attributable to the inductive...
nature of the voice coil which, as mentioned earlier, was measured at 130Ω. Fig.9.2b shows how this reactance varies with frequency. You should note that this curve rises from zero to about 8Ω at 10kHz and closely resembles the curve shown in Fig.9.2a. In practical terms, the upshot of this is that, due to the increase in impedance the current (and therefore also the power delivered to the loudspeaker) will become progressively reduced at frequencies above 4kHz. This problem can be partly offset by the addition of a tweeter – a speaker dedicated to higher frequencies – which helps to increase the sound pressure at high frequencies.

**Equivalent circuit of a loudspeaker**

Fig.9.3 shows the evolution of a loudspeaker as a load. In Fig.9.3b we have simply replaced the loudspeaker by a resistor having a DC resistance that is the same as the nominal impedance of the loudspeaker. This, as we mentioned earlier, is far from satisfactory in terms of modelling a loudspeaker’s behaviour. Fig.9.3c shows an improved equivalent circuit in which we’ve added the voice coil inductance. The impedance can be calculated at any frequency for this circuit using:

\[
Z = \sqrt{R^2 + X_L^2} = \sqrt{R^2 + (2\pi f L)^2}
\]

Fig.9.3d takes our equivalent circuit one step further. The circuit comprises a series network of \( R_1 \) (the voice coil resistance), \( L_1 \) (the voice coil inductance), and a parallel network comprising \( L_2, C_1, \) and \( R_2 \). These three components are not ‘real’ physical components, but we have shown them here to account for the resonant peak in the impedance characteristic that we discussed earlier. Fig.9.4 shows the complete equivalent circuit of our Sharp XL-60H 40W loudspeaker with the values that we obtained by measurement and calculation.

**Zobel networks**

Fortunately, the steady increase in impedance of a loudspeaker that results from the inductance of its voice coil can be counteracted by nothing more than a simple C-R network connected in parallel with the driver. This network is designed so that it exhibits an impedance which is equal and opposite to that of the voice coil. The result is a load that appears to be almost purely resistive at all frequencies. This clever idea came from the work of Otto Zobel, an engineer who was employed by the American telecommunications company, AT&T. A French electrical engineer, Paul Boucherot, who was then working for the French railway company, Chemin de Fer du Nord, had a similar idea, although his application was intended more for electrical power distribution than for use in the audio/communications sector.

Fig.9.5 shows a Zobel network connected in parallel with a loudspeaker. The values required for \( R_Z \) and \( C_Z \) (which essentially balance the ‘bridge’ circuit) can be calculated from:

\[
R_Z = R_1
\]

and

\[
C_Z = \frac{L_1}{R_1^2}
\]

**Multiple speaker systems**

Good quality loudspeaker systems usually comprise several speakers within each speaker enclosure, and these are dedicated to different frequency ranges. Low-frequency driver units (woofers) generally cover from about 20Hz to 200Hz, while mid-range and high-frequency drivers (tweeters) usually cover the frequency range 200Hz to 20kHz. It then becomes necessary to separate the audio range into different bands so that each of the speakers is presented with an appropriate range of signals. This is usually achieved with the use of a ‘crossover network’ of passive components arranged as simple second- or third-order L-C filters. Fig.9.6 shows how this works in the case of a dual loudspeaker system.
A second-order high-pass filter, \(C_1/L_1\), is interspersed between the amplifier output and the high-range unit, \(LS_1\), while a second-order low-pass filter, \(L_2/C_2\), is interspersed between the output and the low-range unit, \(LS_2\).

The design of the filters used in a cross-over network assumes that the loudspeaker (particularly the low-frequency driver) is well-behaved and has a constant impedance over the frequency range. As we've just shown, this is never the case, so one or more Zobel networks are invariably added, as shown in Fig.9.6. Note that \(C_Z\) and \(R_Z\) are fitted directly across the loudspeaker, not as part of the amplifier circuitry.

**Using Tina Design Suite to model a loudspeaker**

The performance of a loudspeaker, with and without a Zobel network fitted, can be easily modelled using Tina Design Suite, as shown in Fig.9.7. Note that Tina cannot plot impedance directly, so instead you use Tina's Signal Analyzer tool to plot voltage against frequency. In Fig.9.8 we plotted the voltage/frequency characteristic when the loudspeaker (without an added Zobel network) was supplied with 1V from a source having an internal impedance of 4\(\Omega\). In this graph, a voltage of 500mV corresponds to a load impedance of 4\(\Omega\). This is the case at 10Hz and 1kHz but, as expected, the impedance rises to a peak at around 150Hz and also rises progressively towards the high end of the spectrum.

In Fig.9.9 we used Tina’s Signal Analyzer to plot the response of the system with a Zobel network present. In this case, \(R_Z\) was 4\(\Omega\) and \(C_Z\) was 8\(\mu\)F, but please feel free to check these values by calculation (use the formula that we showed earlier). Note that the resonant peak is still present but, above a few hundred Hz the impedance does remain substantially constant at almost exactly 4\(\Omega\), clearly demonstrating the usefulness of Zobel’s work.

**Improving stability**

Earlier in this series we showed how the stability of an amplifier can be improved by adding negative feedback. But, as we’ve just demonstrated, there’s always a degree of uncertainty concerning the load to which an amplifier is connected. This load can often have significant reactance and because of this, additional precautions are usually necessary to ensure that an amplifier remains unconditionally stable.

A typical output network for a high-power audio amplifier is shown in Fig.9.10. R1 and R2 are low-value emitter resistors that add some series current feedback to help maintain the operating conditions of \(TR_1\) and \(TR_2\). They also help ensure that the amplifier has a measure of protection against a short circuit at its output. R1 and R2 must be suitably rated. In the case of an amplifier rated at a fairly modest 10W into 4\(\Omega\), an RMS output current of around 1.5A will be produced and so R1 and R2 will need to be rated for a power of at least 1W.

The series network formed by R3 and C1 connected in parallel with the output is not, as it might first appear, a Zobel network for correcting the rising impedance characteristic of the loudspeaker (see earlier). Instead, it ensures that the amplifier is always loaded at frequencies outside the normal audio frequency range. This is...
sometimes necessary due to an inherent susceptibility of emitter-follower output stages to oscillate at frequencies well outside the audio spectrum. With the values shown in Fig.9.8 the network ensures that the amplifier sees a load of around 5Ω at a frequency of 1MHz. In addition, the parallel network formed by L1 and R4 connected in series with the output helps to increase the impedance of the output at high frequencies (the reactance of L1 is about 14Ω at 1MHz).

**Over-current and short-circuit protection**

With high power amplifiers and due to the vagaries of load impedances and, more importantly, the possibility of an inadvertent short circuit at the output, it is desirable to incorporate some form of over-current protection in the output stage. A typical arrangement is shown in Fig.9.11. TR1, TR2, TR3 and TR4 form a complementary cascaded emitter-follower output stage (see last month). The voltage across the two emitter resistors, R5 and R6, will be directly proportional to the emitter currents of TR3 and TR4 respectively. If the voltage dropped across R5 or R6 should ever exceed the base-emitter voltage threshold (around 0.6V) for TR5 and TR6, the respective transistor will conduct heavily, effectively clamping the base bias voltage applied to TR1 or TR2. This, in turn, will have the effect of limiting the output current to about 2A with the values shown in Fig.9.11. Short-circuiting the output terminals under driven conditions will then only have the effect of limiting the output current and voltage swing (at which point the output waveform will become severely clipped).

**Thermal protection**

Earlier in Part 7 (August 2015, EPE) we discussed heat and heat dissipation, and described some practical heatsink arrangements that can be used with audio power amplifiers, the aim being to prolong component life and avoid early failure. A properly designed heatsink mounted in a ventilated enclosure will help limit the temperature of the output devices but, due to the vagaries of the environment in which an amplifier operates (particularly as regards ambient temperature), it may become necessary to incorporate an electronic means of limiting the rise in temperature experienced by the output transistors. We can do this easily by incorporating thermal feedback into our design. The best place to apply such feedback is in the stage that sets the bias for the output devices.

Thermal feedback is based on sensing the temperature of the output devices in order to regulate the amount of bias voltage applied to them. There are several ways of doing this, but one of the most effective makes use of the positive temperature coefficient that is exhibited by conventional bipolar junction transistors. When the junction temperature increases the base-emitter threshold voltage, V\text{BE}, will fall at a rate of about 2mV per °C.

As an example, a BC847 transistor operating with a collector current of 10mA will exhibit a base-emitter threshold voltage of 700mV at 25°C but, at 75°C this will have fallen to around 600mV (see Fig.9.12).

As temperature increases, the falling threshold voltage will produce an increase in base current. If this sounds a little counterintuitive it is worth recalling that the base connection of a V\text{BE} multiplier is fed from a potential divider (see Fig.9.13). If the base-emitter threshold voltage falls, more base current will flow and, in turn, this produces a corresponding increase in collector current. As a result, the collector-emitter voltage (V\text{BIAS} in Fig.9.13) will fall. The reduction in V\text{BIAS} will result in a decrease in base bias current applied to both the driver and output transistors and, as a result, the power dissipation will be reduced and the temperature of the output devices will fall.

The device used for the V\text{BIAS} multiplier stage should be mounted in very close proximity to one or both of the output devices (ie, it should have close thermal coupling). In some cases the sensing device is attached directly to the case of one of the output transistors, but in others it is simply clamped onto the heatsink between the final output pair. Last, it
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is worth noting that in Fig.9.13, pre-set potentiometer RV1 has been included in the lower section of the VBE multiplier’s potential divider. This helps to ensure that, in the event of this component failing in an open-circuit condition, the output stage is biased completely off. You can easily check the operation of a VBE multiplier as a temperature sensing device using Tina Design Suite, as shown in Fig.9.14. We used Tina’s virtual multimeter to display the output voltage, VBIAS. Fig.9.15 shows Tina’s transistor parameter dialogue in which the ambient temperature can be set. Note the tick in the checkbox which allows the scroll arrows (on the right) to be used for adjusting the ambient temperature. Finally, Fig.9.16 shows the results of our investigation plotted as a graph. This clearly shows how the bias voltage falls from about 1.44V at 20°C to around 1.25V at 80°C.

Get Real: A high-performance 10W audio power amplifier

Our final Get Real project is designed to satisfy the need for a simple, low-cost power amplifier that can be used with other modules in this series to realise a complete modular audio system. When we originally discussed this series with editor Matt Pulzer, I suggested that it would be possible to produce a simple low-cost audio amplifier based on discrete components that would out-perform existing designs based on integrated circuits. This is the culmination of our Get Real practical projects; an amplifier capable of excellent performance at a reasonable cost. Our original minimum design specification for the amplifier is given in Table 9.1.

Circuit description
The complete circuit of the high-quality audio amplifier is shown in Fig.9.17. The amplifier uses eight commonly available low-cost silicon transistors. The input stage, comprising TR1 and TR2, is connected in differential configuration with a constant current source (the ‘tail’) formed by TR3 and associated components. This device supplies the combined emitter current for the differential pair TR1 and TR2. A PNP device, TR4, acts as a common-emitter driver stage, while TR5 and associated components connected in the collector circuit of TR4 act as a VBE multiplier, which defines and stabilises the bias applied to the output stage formed by TR6, TR7 and TR8. The output stage comprises a quasi-complementary configuration in which a Sziklai compound feedback pair, TR7 and TR8, replaces the single PNP device that might otherwise have been fitted as a complement to the NPN emitter-follower stage, TR6. To help you understand the circuit, Fig.9.17 shows the five functional stages shaded together with the negative feedback path from the output stage back to the differential amplifier.

Gain and frequency response
The input signal is coupled into the pre-amplifier by means of C1 but, apart from this, the amplifier is entirely DC coupled. This ensures excellent low-frequency response. DC negative feedback from the output is provided by means of R6. This helps to regulate the DC conditions within the amplifier. AC negative feedback is used to define the overall voltage gain and frequency response. The overall mid-band voltage gain is approximately given by:

<table>
<thead>
<tr>
<th>Device</th>
<th>Emitter</th>
</tr>
</thead>
<tbody>
<tr>
<td>Voltage gain</td>
<td>10 (approx.)</td>
</tr>
<tr>
<td>Frequency response</td>
<td>Better than 10Hz to 20kHz at –1dB</td>
</tr>
<tr>
<td>Input impedance</td>
<td>50kΩ (approx.)</td>
</tr>
<tr>
<td>Output load impedance</td>
<td>3Ω to 15Ω (4Ω nominal)</td>
</tr>
<tr>
<td>Output power</td>
<td>10W into 4Ω at 1kHz at less than 1% THD</td>
</tr>
<tr>
<td>Phase shift</td>
<td>0° at 1kHz (ie, output non-inverted)</td>
</tr>
<tr>
<td>Supply voltage</td>
<td>12V to 18V (15V nominal)</td>
</tr>
<tr>
<td>Supply current</td>
<td>50mA (no signal), 1A (max.)</td>
</tr>
<tr>
<td>Distortion</td>
<td>Better than 0.5% THD at 5W output into 4Ω</td>
</tr>
</tbody>
</table>

Fig.9.14. Testing the VBE multiplier at different temperatures

Fig.9.15. Setting transistor parameters and ambient temperature in Tina Design Suite

Fig.9.16. Variation of output stage bias voltage with temperature

Table 9.1 Outline design specification for the high-quality 10W audio amplifier
The overall mid-band voltage gain, $A_V$, is approximately given by:

$$A_V = 1 + \frac{R_6}{R_5} = 1 + \frac{100\times10^3}{10 \times 10^6} = 11$$

The lower cut-off frequency, $f_1$, is thus approximately:

$$f_1 = \frac{0.159}{C_2 \times R_5} = \frac{0.159}{4.7 \times 10^{-10} \times 10 \times 10^6} = 3.4 \text{ kHz}$$

While the upper cut-off frequency, $f_2$, is thus approximately:

$$f_2 = \frac{0.159}{C_3 \times R_6} = \frac{0.159}{47 \times 10^{-10} \times 100 \times 10^6} = 34 \text{ kHz}$$

Where necessary, an external volume control (for example, a good quality 100k logarithmic-law carbon potentiometer) can be fitted at the input of the amplifier.

**Bias adjustment**

Two adjustments are required; RV1 sets the symmetry and allows the output voltage to be set to zero in the absence of a signal, while RV2 sets the standing bias current in the two output devices, TR6 and TR7. The correct adjustment of these two pre-set controls is crucial in determining the overall performance of the amplifier.

The setting of RV1 determines the collector current of the constant current source, TR3. This, in turn, impacts on the collector voltage of TR1. If this voltage falls, TR4 will conduct more heavily and, as a result, the voltage at the base of TR6 will rise. This, in turn, will cause the output voltage (at the junction of R10 and R11) to increase. Conversely, if TR4’s collector voltage rises, TR4 will conduct less heavily and, as a result, the voltage at the base of TR6 will fall. This, in turn, will cause the output voltage (at the junction of R10 and R11) to fall.

The setting of RV2 determines the collector-emitter voltage of the $V_{BE}$ multiplier formed by TR5 and associated components. This voltage also acts as the base bias voltage applied to TR6 and TR7. This voltage should be nominally around 1.2V to bias the two base-emitter junctions for linear operation but, as we explained earlier, the voltage will be dependent on the ambient temperature and thus will also provide some thermal feedback to protect the output stage. C5 simply ensures that the same signal voltage is applied (in-phase) to the base of TR5 and TR7.

Finally, it’s worth noting that these two low-value resistors have been included to provide some protection for TR6 and TR8. As the current flowing in R10 and R11 increases, the base-emitter voltages for TR6 and TR7 will be correspondingly reduced. So, for example, if the emitter current in TR6 or TR8 was to increase, by say 50mA, the respective base bias voltage would be reduced by 50mV, opposing the original change and helping to stabilise the emitter current.

**Supplies**

The positive and negative supply rails are derived from an external supply of nominally ±15V. In practice, any symmetrical voltage supply of between ±12V and ±17V rated at 1.5A can be used, but the maximum output power will fall to about 8W with supplies of less than about ±13.5V. Decoupling of the positive and negative power rails is provided by C7 and C8 respectively. Note that these two components will effectively appear in parallel with the larger smoothing/reservoir capacitors present in the power supply (more of this next month) and so their use here is primarily to decouple signals that might otherwise appear on the supply rails.

**Components**

- **1 PCB, code 909 available from the EPE PCB Service, size 163mm × 56mm**
- **2 PCB mounting 2-way terminal blocks**
- **1 PCB mounting 3-way terminal block**
- **2 Heatsinks (see text)**

**Resistors**

- 1 100kΩ (R1)
- 1 470Ω (R2)
- 1 3.9kΩ (R3)
- 1 10kΩ (R5)
- 1 10kΩ (R6)
- 1 560Ω (R9)
- 1 1W (R10 and R11)
- 1 4.7Ω (R12)
- 1 500Ω miniature skeleton pre-set (RV1)
- 1 100Ω miniature skeleton pre-set (RV2)

**Capacitors**

- 1 1µF (C1)
- 1 4.7µF (C2)
- 1 47µF (C3)
- 1 220µF (C4)
- 1 220µF (C5)
- 1 100nF (C7, C8)

---

**Fig.9.17. Complete circuit of the high-quality 10W audio amplifier showing functional stages**

**Everyday Practical Electronics, October 2015**
Semiconductors
4 BC337 (TR1, TR2, TR3, TR5)
2 BC327 (TR4, TR7)
2 TIP41 (TR6, TR8)
2 1N4148 (D1, D2)

Heatsinks
The two power transistors, TR6 and TR8, need to be fitted to individual heatsinks rated at 8.3°C/W or better. These heatsinks should be isolated from the rest of the circuit as this will allow the transistors to be bolted directly to the heatsink surface without the need to use insulating washers and bushes (see EPE August 2015). The heatsinks used in the prototype were PP50BPHC units manufactured by ABL. They were supplied pre-drilled for TO220, TO218 and TO247 devices. The heatsinks measure 13 × 35.6 × 50mm and are supported by two solder pins, spaced at 25.4 mm (1 inch). Many other heatsinks are suitable, provided they can be fitted within the available area on the PCB and that they are suitably rated. Under no circumstances should a heatsink rated at worse than 8.5°C/W be used. Please also note that the heatsinks should be of the ‘notched’ variety with a cut-out so that the three transistor leads can be positioned in line with the two mounting pins when the heatsink is mounted flush with the PCB.

Choice of transistor
We selected low-cost BC337 transistors for use as the low-power NPN stages (TR1, TR2, TR3 and TR5) and BC327...
devices for use in the PNP stages, TR4 and TR7. These devices are rated for an absolute maximum collector power dissipation of 625mW, which should be derated above 25°C by 5mW/°C. Many other devices with current gains in the range 100 to 250 can be substituted (such as BC548 and BC557 devices for the NPN and PNP transistors respectively). Note that the BC548 has a maximum collector power dissipation of 500mW and the quiescent power dissipation for TR3 will be around 400mW in normal operation.

The two output transistors, TR6 and TR8, should have similar current gains (hFE) but, as mentioned last month, devices taken from the same manufacturer’s batch will usually be reasonably closely matched. Many other NPN power transistors with similar ratings can be substituted for the specified devices but, as always, it is essential to check on the device pin-out before making any substitution.

**Construction**

Our prototype PCB was designed to be built into a small enclosure or incorporated into a larger enclosure along with other circuitry. It measures just 163mm x 56mm. In common with all of our Get Real projects, the PCB component layout and copper track layout was produced using Circuit Wizard – shown in Fig.9.19 and 9.20 respectively. The printed circuit can be purchased, ready drilled, from the EPE PCB Service, code 909. Our finished prototype, ready for testing, is shown in Fig.9.21. Finally, Fig.9.22 shows our pre-prototype version assembled on matrix board. We used this version to carry out performance checks before finalising the design and committing to the final printed circuit layout.

Assembly is straightforward, but we suggest fitting the components to the printed circuit board in the following order: terminal blocks (3), links (2), fixed resistors (12), pre-set resistors (2), diodes (2), non-electrolytic capacitors (5), electrolytic capacitors (5), NPN small-signal transistors (4), PNP small-signal transistors (2), power transistors (2, pre-mounted on two heatsinks). Note that it is essential to observe the correct polarity of the diodes and electrolytic capacitors and the correct orientation of transistors.

**Setting-up**

Before testing it is worth carrying out a careful visual check of the board and components, checking the underside of the PCB for solder bridges between tracks. You will need a digital multi-meter and a fused (or electronically protected) DC power supply with symmetrical outputs of nominally ±15V at 1A (see earlier) to read the DC voltage appearing across R10 (the negative meter lead should be taken to the terminal that appears closest to TR8). Switch the power supply ‘on’ and readjust RV1 for a reading of less than ±10mV.

1. Set RV1 to mid-position and RV2 fully clockwise.
2. Connect the digital multi-meter (on the 2V DC range) to read the DC voltage appearing across the output at CN2 (the negative meter lead should be taken to the terminal that appears closest to TR8). Switch the power supply ‘on’ and carefully adjust RV1 for a reading of less than ±10mV.
3. Re-connect the digital multi-meter (on the 2V DC range) so that it reads the DC voltage appearing across R10 (the negative meter lead should be taken to the junction of R10 and R11 near C6 on the printed circuit board). Switch the power supply on and adjust RV1 for a reading of between 25mV and 35mV (corresponding to a standing bias current of between 25mA and 35mA).
4. Reconnect the digital multi-meter (on the 2V DC range) to once again read the DC voltage appearing across the output at CN2 (the negative meter lead should be taken to the terminal that appears closest to TR8). Switch the power supply ‘on’ and readjust RV1 for a reading of less than ±10mV and that the standing (quiescent) bias current in the output stage is approximately 30mA.

Finally, our measured test voltages are shown in Table 9.2.

**Next month**

In the final installment of Teach-In 2015, Discover will examine practical aspects relating to measurement, adjustment and fault-finding in power amplifiers. Knowledge Base will look at power supplies, and Get Real will explain the procedure for measuring the performance of our high-quality 10W amplifier and we reveal whether or not it met or exceeded our original design specification.

---

**Table 9.2 Test voltages**

<table>
<thead>
<tr>
<th>Device</th>
<th>Emitter</th>
<th>Base</th>
<th>Collector</th>
</tr>
</thead>
<tbody>
<tr>
<td>TR1</td>
<td>–1.0V</td>
<td>–0.4V</td>
<td>+16.7V</td>
</tr>
<tr>
<td>TR2</td>
<td>–1.0V</td>
<td>–0.4V</td>
<td>+17.1V</td>
</tr>
<tr>
<td>TR3</td>
<td>–16.2V</td>
<td>–15.6V</td>
<td>–1.0V</td>
</tr>
<tr>
<td>TR4</td>
<td>+17.1V</td>
<td>+16.7V</td>
<td>+0.6V</td>
</tr>
<tr>
<td>TR5</td>
<td>–0.6V</td>
<td>0V</td>
<td>+0.6V</td>
</tr>
<tr>
<td>TR6</td>
<td>+0.03V</td>
<td>+0.6V</td>
<td>+17.1V</td>
</tr>
<tr>
<td>TR7</td>
<td>–0.03V</td>
<td>–0.6V</td>
<td>–16.45V</td>
</tr>
<tr>
<td>TR8</td>
<td>–17.1V</td>
<td>–16.45V</td>
<td>–0.03V</td>
</tr>
</tbody>
</table>

---

*Fig.9.21. Completed 10W amplifier ready for testing*

*Fig.9.22 The pre-prototype 10W amplifier built on low-cost matrix board*
BRITISH Telecommunications (BT) claims to be the world’s oldest communications company, tracing its heritage all the way back to 1846. At a local conference some 15 years ago, BT sought to squeeze more from its broadband by pleading with developers and web designers to take full advantage of the then-new ADSL by providing more multimedia content for end users; the network speed was there, but online content that could fully utilise that speed had yet to arrive. Today, with high-speed VDSL now on offer over its phone lines, BT wants a greater stake in building today’s ‘information age’ and it has the financial firepower to do it. In 2013, BT paid nearly £900 million ($1.4 billion) for the rights to stream three years’ worth of live Champions League and Europa League football. BT’s website rubs it in when bragging about its online programming, reminding us that ‘BT Sport is the only place to watch all the UEFA Champions League live. These matches are no longer shown on ITV or Sky Sports.’

Not resting on its laurels, the telecoms titan recently announced a new BT TV Ultra HD service as well, that they claim includes Europe’s first live sports Ultra HD (4K) channel. So now you can watch a football being kicked around with four times the definition of HD.

If BT wants to captivate us with online content then it faces plenty of competition. Amazon continues its very aggressive bid to grab more market share with the news that the former BBC TV ‘Top Gear’ team has been signed up to produce three series exclusively for Amazon Prime. The hugely popular Top Gear motoring programme had been a BBC TV staple and the brand was sold very successfully worldwide, but the series fell apart after presenter Jeremy Clarkson took a swing at a TV producer after a tough day’s filming. The orphaned programme had been looking for a new home and rumours were rife about switching to a streaming service. In an extremely deft move, a new series will premier on Amazon Prime in 2016, and its success is assured. Amazon Prime in the UK currently costs £79 a year and it bundles TV and movies with a one-day delivery service on much of its merchandise. Amazon has also introduced Prime Music, offering a million songs at no extra charge for its Prime members.

Taking Ten
As the joke goes, you can spend more time on Netflix looking for movies than actually watching them, and I have rented my share of duff films that were a waste of money. After an evening of sifting through endless choices of online movies, Microsoft Studios’ *Halo 4: Forward Unto Dawn* finally burst onto my TV screen. Featuring the characters first seen in the Halo Xbox video games, the movie gave the nod to an interactive artificial intelligence called Cortana that conversed with our armour-plated hero. The holographic Cortana first appeared in Halo on Xbox in 2001, but it’s also the name for Microsoft’s digital assistant found in Windows 10 and Windows Phone.
10. Like many users, though, I remain perfectly content with Windows 7, and feel no need to upgrade straight away.

Among other benefits, Windows 10 brings back the Start button and menu, the user interface (UI) first introduced 20 years ago in Windows 95 and foolishly omitted in the unloved Windows 8.0. Windows 10 also introduces Microsoft Edge, their new, leaner, faster web browser that promises to be everything that Internet Explorer is not (fast, secure and reliable). It has a two-dimensional characterless styling first mooted by Google Chrome, but Microsoft Edge integrates with Cortana to provide voice interaction in the same way that Google, Apple's Siri and Amazon's Alexa offer voice control for their users. Users can also 'write' on web pages in Edge and share notes with others.

There is no doubt that the stakes could not be higher for Microsoft's latest operating system, and downloading the free upgrade to Windows 10 within the next twelve months will be the best choice for PC users wanting to move with the times and keep their system as secure as possible. Maybe users may not be ready to upgrade just yet, but Microsoft has graciously allowed a year to think about it. Future upgrades of Windows 10 will be incremental: there will be no new release and instead Microsoft will enhance Windows in the same way that many operating systems and apps update themselves on a rolling basis.

Flat Earth society

Windows users are not alone when it comes to the need to update operating systems. Owners of Android handsets and tablets bear witness to the drudgery of apps, sometimes several dozen of them at a time, being patched to the latest versions. To add insult to injury, the author's Android phone updated itself one evening, which resulted in its Do Not Disturb (DND) function disappearing. The DND feature stops the smartphone from sounding at inconvenient times. Consequently, the newly-updated phone woke the writer at 3am with the sound of new email arriving, and 4am saw him pillaging it under a pillow with moaning to find how to silence the wretched thing – now running Android Lollipop – when it's supposed to stay silent overnight.

For some reason, following the Android update, DND had become ' Interruptions', accessible via the phone's volume controls. The option to silence it automatically on Sundays had scrolled off the bottom of the Android screen, tucked out of sight. Hence the phone would wake the writer on Sunday nights, beeping incessantly with new email until eventually the hidden 'Sunday' tick-box was found on the calendar menu, and weekend bliss was finally restored.

This fashion for flat, bland styling is creeping into UI design on websites, PCs and touchscreens everywhere, and a counter-intuitive and featureless landscape of coloured tiles and oversized text with few usage cues is all too commonplace these days. Traditional Windows-type sliders and controls are being usurped by flat, touchscreen-type navigation, begging you to scroll up or down with your thumb. This lack of a recognisable slider knob is typical of the over-simplified 2D user-interface beloved of the current generation of website and touchscreen designers. Furthermore, the need to re-learn a user interface following an Android update and get to grips with any changes is becoming a very annoying and time-consuming aspect of device ownership.

In recent years – until the touchscreen came along – the vast increases in computer graphics and monitor performance had seen more attractive 3D graphics, alpha blending and transparency, drop shadows and high-resolution icons (eg the Windows Aero interface) that made computer use both intuitive and attractive. The information age has now gone ‘flat earth’ and users are presented with expanses of flat colour tiles and often poorly-contrasting icons. Maybe you will spot the odd ‘X’ to close a window if you’re lucky, but operating systems, software and apps are all heading the same way. Hand in hand with this is the fashion for the so-called ‘hero image’ – a term borrowed from the print design industry for a very large graphic or background image that dominates a page. Full-screen hero images often have no menu or navigation at all, but are only feasible due to improved Internet bandwidth and larger computer screens.

Out in a Flash

Regular Net Work readers will know how I constantly nag about the need to keep software updated, mainly to defend a system against vulnerabilities that could be exploited by hackers. A ‘zero day’ vulnerability is one that is currently unmatched by vendors, which leaves the product susceptible to attack. Most software houses release updates fairly swiftly and users would have to be quite unlucky to suffer any problem directly, but heading my list of likely targets are Adobe Reader and Adobe Flash Player. They offer many opportunities for hackers to exploit, either by opening an infected document on your PC or by visiting a website that exploits weaknesses in your software.

The latest Adobe Acrobat Reader DC (Document Cloud) is best known as a standalone PDF viewer that also has a web browser plugin. It includes some useful features, such as Toner/Ink Saving, which saves up to 15% by economising on the amount of ink or toner printed onto paper. (The Ryman Eco typeface from www.ryman.co.uk/ryman-eco/ can be used in your own documents to reduce by 33% the volumes of ink or toner needed to print them.) The free Adobe Reader DC also tantalises users with useful tools such as rubber-stamping or overprinting PDFs, but some tools such as Convert (files) to PDF head off an Adobe subscription sign-up webpage costing up to £100 a year!
Firefox now blocks by default the Flash plugin unless you activate it

So-called ‘Flash’ is the technology that delivers motion graphics, video and audio over the web. In recent weeks both Firefox and Google web browsers decided to block Flash altogether due to perceived security risks. This is now resulting in web pages failing to load inline content such as Google adverts and banners or multimedia presentations. Users can manually allow Flash to run, but this can be such a nuisance that web surfers prefer to simply ignore the Flash plugin warning and not bother activating Flash at all. The most advanced media websites are gradually adopting the latest web standards – HTML5 – which avoids the need for Flash altogether. For now though, users may have to make their own judgments about whether to allow Flash to run on a particular web page, and you can be sure that Firefox and Google Chrome will put obstacles in the way of displaying these multimedia elements.

That’s all for this month’s Net Work. You can contact the writer at alan@epemag.demon.co.uk

Firefox now blocks by default the Flash plugin unless you activate it

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Everyday Practical Electronics, October 2015
Using the Raspberry Pi to produce a controllable output voltage with the aid of a digital-to-analogue converter (DAC) has been covered in previous Interface articles. This method is the best approach for many applications, but is less than ideal for others. The alternative approach of using pulse-width modulation (PWM) is preferable for some applications, such as controlling the speed of DC electric motors, and the brightness of filament bulbs or LEDs. PWM often requires simpler and cheaper hardware, and in power-control applications it is more efficient.

**PWM basics**

PWM does not provide a variable DC output voltage, but instead delivers a pulse signal that has a variable duty cycle. It is the average output voltage that is controlled, although with suitable filtering a DC voltage equal to the average output potential can be provided. Fig.1 shows some example waveforms that help to explain the way in which PWM operates. In the top waveform there is a 1:1 mark-space ratio, and the output is switched off for 50 percent of the time. The average output potential is therefore equal to half the ‘on’ voltage. In the middle waveform the mark-space ratio is 4:1, and the average output potential is therefore 4/5 or 80 percent of the ‘on’ voltage. The lower waveform is the inverse of this, with a 1:4 mark-space ratio, and an average output potential equal to 1/5 or 20 percent of the ‘on’ voltage. Most PWM controllers use a fixed frequency and a variable duty cycle, as in the example waveforms, but much the same effect can be obtained using a fixed pulse width and a variable frequency.

Efficiency is high with a PWM controller because the output transistor operates as a switch, and it provides either the full output potential or no output voltage at all. In theory, there is no power wasted in the control circuit. The switch passes a high current when it is closed, but the voltage drop is zero, giving zero power loss. No output current at all flows when the switch is open, and the output power is therefore zero. Of course, theoretical perfection is one thing and real-world electronics is another. Although there is a totally insignificant current flow when a switching transistor is turned off, there is a significant voltage drop when a transistor is switched on, especially when high currents are involved. PWM is still much more efficient than using some form of variable resistance, where the power wasted in the controlling element will sometimes exceed that supplied to the load.

Of course, in its basic form with no output filtering, pulse control is not suitable for all applications. Where it is suitable, due care has to be taken when selecting the pulse frequency. Using a very low frequency is unlikely to be successful. LEDs and filament bulbs would be seen to flash on and off rather than dim. Similarly, a DC electric motor would operate in a ‘fits and starts’ fashion rather than operating at a set speed. High pulse frequencies could give problems with some types of load, for example, inductive types. There could also be problems with the generation of strong radio frequency interference (RFI). A frequency in the region of 50 to 200Hz is suitable for most practical applications of pulse control.

**Software PWM**

PWM can be achieved using hardware or software. The hardware method has been covered in previous Interface articles, and one method is to output values to a digital-to-analogue converter (DAC). The output voltage of the converter is fed to a circuit that is comprised of a triangular waveform generator and a voltage comparator. Much simpler and cheaper hardware is required with the software method, because all the hard work is done in the program. This generates the required waveforms on a digital output of the computer, and the hardware simply has to provide voltage amplification and buffering so that the load can be driven properly.

The Raspberry Pi has PWM software available via Python and the RPi.GPIO module. This enables a PWM controller to be implemented very easily using minimal hardware. The software is very simple as well, because the wave generation is handled by the RPi.GPIO module, and the control program simply has to instigate the required changes in the duty cycle via the appropriate Python program instruction. Any of the GPIO input/output pins can be used as a PWM output, and it is possible to have two or more PWM outputs operating simultaneously. Any line used as a PWM output must first be set to the output mode in the normal way.

One problem with software PWM is that the accuracy might be impaired by the processor servicing interrupts and handling other tasks that keep the computer running properly. In practice, Raspberry Pi software PWM works well, but it is probably best to keep the operating frequency quite low. Higher frequencies involve shorter pulse durations that are more vulnerable to brief processing interruptions. Also, there is a limit to the minimum pulse duration that can be provided, and the maximum available resolution of the system is therefore better at low output frequencies.

**PWM instructions**

There are several instructions available for use with a PWM output:

\[
p = 
\]

\[
p.GPIO.PWM(channel, frequency)
\]

Here, \(p\) is the name given to the PWM output, and it is effectively the variable used to control it, and \(channel\) is the number of the GPIO line that is being used to output the PWM signal. The frequency is specified in hertz (Hz).

\[
p.start(duty cycle)
\]

This instruction starts the PWM output signal and sets the initial duty cycle. The value used here is from 0 (always low) to 100 (always high), and the value does not have to be an
integer. This value is setting the percentage of the time that the PWM output is in the high state.

**p.ChangeDutyCycle(duty_cycle)**

Once the PWM output has started operating, this instruction is used to alter the duty cycle. It should be noted that the capitalised letters in the instruction are required. Using (say) changedutycycle will not work.

**p.ChangeFrequency(frequency)**

In a normal PWM application it is not necessary to change the initial operating frequency, but it can be achieved using this instruction. Again, the capitalisation of some letters is required for the instruction to work.

**p.stop()**

The PWM signal should be terminated using this instruction prior to the program ending. The GPIO.cleanup() instruction should still be used at the end of the program in the usual way.

**Software**

The simple Python program of Listing 1 can be used to test the PWM instructions. The first part of the program imports the RPi.GPIO and time modules. The latter is needed because the program uses the sleep function of the time module to provide delays. The main section of the program sets pin 26 of the GPIO port as an output, designates it as a PWM output, sets the initial operating frequency at 100Hz, and then starts the output signal with a duty cycle of 50 percent. After an eight-second delay, the output frequency is increased to 200Hz, and after a further eight-second delay, the duty cycle is changed to ten percent. The PWM signal is switched off after a further eight-second delay, and the program then terminates.

```
import RPi.GPIO as GPIO
import time
GPIO.setmode(GPIO.BOARD)
GPIO.setup(26,GPIO.OUT)
p=GPIO.PWM(26,100)
p.start(50)
time.sleep(8)
p.ChangeFrequency(200)
time.sleep(8)
p.ChangeDutyCycle(10)
time.sleep(8)
p.stop
GPIO.cleanup()
```

Listing 1

```
import RPi.GPIO as GPIO
import time
GPIO.setmode(GPIO.BOARD)
GPIO.setup(18, GPIO.IN, pull_up_down=GPIO.PUD_DOWN)
GPIO.setup(22, GPIO.IN, pull_up_down=GPIO.PUD_DOWN)
GPIO.setup(24, GPIO.IN, pull_up_down=GPIO.PUD_DOWN)
GPIO.setup(26,GPIO.OUT)
p=GPIO.PWM(26,100)
p.start(0)
dutycyc = 0
x = 1
print ("Started")
while (x == 1):
    if GPIO.input(18):
        x = 2
    if GPIO.input(22):
        dutycyc = dutycyc + 1
    if dutycyc == 101:
        dutycyc = 100
    if GPIO.input(24):
        dutycyc = dutycyc - 1
    if dutycyc == -1:
        dutycyc = 0
    p.ChangeDutyCycle(dutycyc)
time.sleep(0.1)
p.stop
print ("finished")
GPIO.cleanup()
```

Listing 2

Ideally, an oscilloscope is needed in order to check the output signal properly. Figs.2a to 2c show the output waveforms obtained with duty cycle values of 50, 80, and 20 respectively. There are simpler ways of checking the output signal. Connecting the output signal to a crystal earphone or the input of an amplifier/speaker will enable the changes in the signal to be heard. An analogue multimeter set to a low DC voltage range should show a reading of about 1.65V initially, or in other words, about half the 3.3V high logic level of the output. The reading should drop to about 0.33V when the signal switches to a duty cycle of ten percent. This will also work with most digital multimeters, but some will not respond to the average voltage of a pulsed signal.

Another way is to use the simple LED driver circuit of Fig.3. This uses a common-emitter switching transistor Tr1..
to drive an LED (D1) via current-limiting resistor R3. The ‘on’ current is about 20mA. Any NPN transistor having a reasonably high current gain will suffice for Tr1, and any LED that operates in the visible light part of the spectrum will do for D1. Connection details for the 40-pin version of the GPIO port are provided in Fig.5. The 26-pin version is essentially the same, but it has pins 27 to 40 omitted.

Variable control
For practical applications it is necessary to have an easy way of varying the output signal. The program of Listing 2, in conjunction with the extra hardware of Fig.4, provides control of the output signal by way of three pushbutton switches. These switches are normally open types that do not latch when operated. The initial part of the program sets pin 26 of the GPIO port as the output for the PWM signal, but it also sets pins 18, 22 and 24 as inputs that have internal pull-down resistors. Each of these inputs is therefore at the low state under normal conditions, but an input can be taken high by operating the appropriate switch. The duty cycle is set at zero initially, or continuously low in other words. The duty cycle of the PWM signal is controlled by the value in the variable called dutycyc, and this is set at a starting value of zero. Variable x is set at a value of 1 initially, and this is used to control a while… loop that operates until x is set to a different value. The word Starting is printed on the screen to confirm that the program is going into the main routine.

The loop checks the states of the three input lines in turn using if… statements. If pin 18 goes high, x is made equal to 2. The loop then terminates and the PWM signal is switched off. Pin 22 going high results in dutycyc being increased by one on each loop of the program, gradually ramping it up to a value of 100. Taking pin 24 high has the opposite effect, with dutycyc being reduced by one on each loop. Two additional if… statements provide error checking that prevent the value in dutycyc going out of range. A potential problem with the program is that it would loop so fast that even momentary operation of S2 or S3 would almost instantly set the output at maximum or minimum. A time.sleep instruction is therefore used to provide a 0.1s delay in each loop so that the speed at which the output changes is more controllable. More on PWM next time, including more powerful driver circuits.
Current mirrors and transistor matching

James’s questions raises some issues not discussed in previous articles, particularly in relation to the accuracy of the current mirroring behaviour of the circuit. The current output (\(I_2\) in Fig.1) is a copy of the reference current (\(I_1\) in Fig.1). If the circuit was a perfect current mirror then these two currents would always be exactly equal, but of course in a real circuit this may not be the case. The circuit’s output (from the collector of Q2 in Fig.1) acts as a current source. If \(I_1\) is constant, which it should be if R1 and the supply do not change, then Q2 acts as a constant current source.

Before looking at James’s question in more detail, we will briefly recap/ introduce current sources and explain the current mirror operation of the circuit in Fig.1. For further background on current sources, please look at Circuit Surgery in the March 2015 issue of EPE.

\[\begin{align*}
R2 &= 1000\Omega, \quad V_{ce} = 4.10V, \quad I_1 = 59.0mA \\
R2 &= 470\Omega, \quad V_{ce} = 3.37V, \quad I_1 = 14.1mA \\
R2 &= 6800\Omega, \quad V_{ce} = 2.22V, \quad I_1 = 11.5mA \\
R2 &= 10000\Omega, \quad V_{ce} = 0.80V, \quad I_1 = 9.2mA
\end{align*}\]

\(I_2\) varies significantly with \(V_{ce}\) and is only similar to \(I_1\) when the two \(V_{ce}\) values are similar. I selected two transistors that have similar values of \(h_{fe}\) measured on a meter with an \(h_{fe}\) measuring facility. I appreciate that \(h_{fe}\) varies slightly with \(V_{ce}\), but this effect is too small to cause the wide variation in \(I_2\) that I am seeing. It appears, at first sight, that using discrete transistors like this is not a practical way to build a current mirror — or am I doing something wrong? Why am I seeing such a wide variation in \(I_2\) as the load (R2) varies? My interest in this area was sparked by Robert Penfold’s use of a discrete current mirror in August’s Interface article.

\[\begin{align*}
\text{Fig.1. James’s current mirror circuit}
\end{align*}\]

Current sources

An ideal current source outputs a particular current irrespective of the voltage across it. Real current sources have internal resistance (see Fig.2), which can be modelled as being in parallel with an ideal current source. The larger the internal resistance, the better the current source. Compare this with a voltage source, where the internal resistance is in series and the smaller the resistance, the more ideal the source.

If we connect a large resistance across an ideal current source, then by Ohm’s law it must develop a large voltage across the resistor. For example, if we connect a 1M\(\Omega\) resistor across a 1mA current source it would have to produce 1000V across the resistor to maintain the required current, impossible in the circuit of in Fig.1 due to the limited supply voltage.

The internal resistance, \(R_{int}\), of an otherwise ideal current source, \(I_0\) limits its output voltage, \(V_{out}\), to \(I_0R_{int}\) (Fig.2), at which point \(I_{out}\) is zero — so this is the open-circuit output voltage. The larger the value of \(R_{int}\), the larger the maximum output voltage. For truly ideal current sources, the open-circuit output voltage is infinite and \(R_{int}\) is infinite — these do not exist in reality, but are a useful mathematical model. If we short-circuit a current source \(R = 0\) in Fig.2) the output current is equal to \(I_0\) and there is zero voltage across the source.

Real current sources also have a limited range of voltages over which they can operate (called ‘compliance’), often determined by factors other than the internal resistance, for example when the transistors in the circuit switch off, change operating mode as their bias conditions change, or reach their operating limits.

Current mirrors

In the circuit in Fig.1 the collector and base of Q1 are shorted together; so we are left with the base-emitter PN junction — the transistor acts as a diode (Fig.3a). We can bias this diode to carry any reasonable forward current by connecting a resistor from the supply, as shown in Fig.3b. The forward voltage drop of the ‘diode’ will be the \(V_{be}\) of the transistor with the collector current set by the resistor.

If we wire this diode-connected transistor to another transistor —
emitter-to-emitter and base-to-base – they will both have the same \( V_{BE} \) (see Fig.3). If the transistors have equal characteristics, then whatever current is set for the first transistor will also flow in the second (\( I_{CM} = I_{c} \) in Fig.4). This is a basic current mirror circuit.

To get a constant current source (rather than a current mirror) we supply a fixed input current through the diode-connected transistor, as in Fig.1. To calculate the value of \( R \) we can assume that the base-emitter voltage of the transistor \( V_{BE} \) is fixed, and in the range 0.6-0.7V, the output current is then:

\[
R = \frac{V_{CC} - V_{BE}}{I_{out}}
\]

If we have more details of the transistor, such as the \( V_{BE} \) vs \( I_{c} \) curve then we can set this more accurately.

Current mirrors are very important in electronic circuit design; they are to be found inside almost every complex analogue IC (such as operational amplifiers) where they have widespread general use for biasing and as active loads.

**Fig.4. The current mirror**

**Fig.5. Currents in the basic current mirror**

**Mirror accuracy**

Fig.5 shows an analysis of currents in the basic current mirror. This starts with the assumption that the transistors are exactly equal in terms of both characteristics and environment. So we can assume they both have the same forward current gain \( \beta = h_{FE} \) and just use \( \beta \) in our calculations, rather than distinguishing, say \( \beta_1 \) and \( \beta_2 \).

Given that the two transistors have the same \( V_{BE} \) (from the way the circuit is wired), they must also have the same \( I_{c} \) (because the characteristics are the same). So currents in the circuit can be determined using \( I_{c} = \beta I_{EB} \) and

\[
I_{C}(1 + \beta)I_{EB} \text{ together with application of Kirchhoff’s current law at the wire junctions. Thus, from Fig.5 we have:}
\]

\[
I_{ref} = \left( \frac{\beta + 2}{\beta + 1} \right) I_{E}
\]

and

\[
I_{out} = \left( \frac{\beta}{\beta + 1} \right) I_{E}
\]

so

\[
I_{out} = \left( \frac{\beta + 2}{\beta + 1} \right) I_{ref} = I_{ref}
\]

That is \( I_{out} \) is approximately equal to \( I_{ref} \) if \( \beta \) is much greater than two. For a ‘rule of thumb’, value of \( \beta \) of 100, the output is about 98% of the reference current.

We don’t know what gain James’s transistors had, but the BC337-25 model in the LTSpice circuit simulator (which we will use shortly) has \( \beta = 292.4 \). Using this in the above equation with \( I_{E} = 9.3mA \) from Fig.1 we get \( I_{out} = 9.2mA \), which is consistent with James’s result for the 1kΩ load. Of course, this does not explain the other measured values.

**Simulation set up**

We can simulate the circuit in Fig.1 in LTSpice using the circuit in Fig.6. Before showing the results we will discuss some details of how this simulation is set up. The simulation we are performing is an operating point analysis. This simply calculates the DC currents and voltages in the circuit – capacitors are treated as open circuits and inductors as short circuits; we do not get any waveforms or frequency response curves. The output from SPICE is just a text listing of the DC currents and voltages.

In Fig.6 we have extended the basic operating point calculation by adding a parametric sweep, that is, we instruct LTSpice to run the same simulation many times with different values of one or more component values. In this case we are using values of R2 from 2Ω to 10000Ω in steps of 5Ω, as defined by the Spice directive, \( .step \) param R 2 10000 5.

\[
\text{The resistance parameter } R \text{ is declared by the SPICE directive, } .param \ R \ 1k \text{, which sets its default value to } 1k\Omega. \text{ The value of } R2 \text{ is set to } R \text{ which is SPICE syntax for referring to a parameter value. Using a parameter sweep allows us to plot a graph of other circuit values (eg. current) against the parameter value.}
\]

We can easily switch between running a simple operating point simulation and the full parameter sweep by right clicking on the .sweep directive and selecting ‘Comment’ rather than ‘SPICE directive’. The .sweep text is shown blue when it is just a comment and R2 takes the default value of parameter R.

**Models**

Another point to note about the schematic in Fig.6 is that we have include a SPICE .model statement directly on the schematic. This model provides characteristics for the transistor (in this case, the BC337 used by James). LTSpice provides this model as part of its download. When you right click a bipolar transistor you can select BC337-25 or BC337-40 (versions with different gains) from the list of real transistor models provided with LTSpice. The models are defined in a text file, standard.bjt, in the lib folder tree, within the LTSpice installation. Typically, the location will be something like: C:\Program\Files (x86)\LTSpiceIV\lib\cmp.

Here, the .model directive was simply copied from the BC337-25 .model directive in standard.bjt, pasted into a SPICE directive, the model name was changed to BC337-25X and the directive placed on the schematic. The transistor type value was set to BC337-25X to match the model name. This approach allows us to change the model parameters in our simulations and quickly have access to the full model. The full model text is not shown on the schematic, it is:

\[
\text{.model BC337-25}
\]

\[
\text{NPN(1S=4.13E-14 NF=0.9822 \text{ ISE=3.534E-15 NE=1.35}}
\]

\[
\text{BF=292.4 JF=0.9 VAF=145.7}}
\]

\[
\text{NR=0.982 ISC=1.957E-13 NC=1.3}}
\]

\[
\text{BR=23.68 ICR=0.1 \text{ VAR2=20 RB=60}}
\]

\[
\text{IRB=2.00E-04 RBM=8 \text{ RE=0.1129}}
\]

\[
\text{RC=0.25 XCT=0 \text{ EG=1.11 XTI=3}}
\]

\[
\text{CJE=3.799E-11 VJE=0.6752}}
\]

\[
\text{MJE=0.3488 TF=5.4E-10 XTF=4}}
\]

\[
\text{VTF=4.448 ITF=0.665 PTF=90}}
\]

\[
\text{CJC=1.355E-11 VJC=0.3523}}
\]

\[
\text{JJC=0.383 XJC=0.455}}
\]

\[
\text{TR=3.00E-08 CJSD=0} \text{ VJS=0.75}}
\]

\[
\text{MJS=0.333 FC=0.643 Vces=0.45}}
\]

\[
\text{Irating=500m mfg=Philips}}
\]

**Simulation**

The results from the LTSpice sweep simulation are shown in Fig.7. Here we see that the currents are generally not equal and the output current varies with the value of the load, R2.

**Fig.6. LTSpice schematic for simulating Fig.1**

**Everyday Practical Electronics, October 2015**
This matches with our calculations in Fig.5 – the two transistors have exactly the same base, emitter and collector currents, but the resistor currents are slightly different. Putting $I_{ref} = 0.0093322$ and $\beta = 292.4$ into the current mirror equation above gives 0.00926880, very similar to the R2 current listed above.

**Mismatch**

We now have a simulation of the current mirror working well, but no explanation for the high currents observed by *james*. One aspect of reality, which is important here, but which has not been included in the simulation is the fact that the two transistors are not identical. In the simulation, the transistors both have the same model (BC337-25X) and hence exactly the same characteristics. We can fix this by giving them different models. We can cut and paste the existing .model directive on the schematic to create a copy, and then edit the two to have different names (e.g., BC337-25X1 and BC337-25X2), then we edit the transistor type names so each has a different one of the two new model names. After this we can change the parameters for the individual transistors and resimulate.

If we set the gain for Q2 to 100 (BF=100 in the model) we get the following results:

--- Operating Point ---

| V(c):     | 1.47173   | voltage |
| V(b):     | 0.66763   | voltage |
| V(supply):| 10        | voltage |
| Ic(Q1):   | 0.00926463| device_current |
| Ib(Q1):   | 3.37854e-005| device_current |
| Ie(Q1):   | -0.00929842| device_current |
| Ic(Q2):   | 0.00852827| device_current |
| Ib(Q2):   | 8.9692e-005| device_current |
| Ie(Q2):   | -0.00861796| device_current |
| I(R1):    | 0.00933237| device_current |
| I(R2):    | 0.00852827| device_current |
| I(Vcc):   | -0.0178606| device_current |

Here we see that the transistors’ base, collector and emitter currents are no longer equal and the R2 current is not so close to the R1 current. Observant readers may have realised that we could contrive to make the resistor currents equal by selecting exactly the right gain for Q2, but this is not really a constructive exercise because we do not have such precise control over the gain of our real transistors.

**Getting hotter**

Another observation from experimenting with varying Q2’s gain is that we cannot achieve the levels of current observed by *james* with realistic values of gain, so it is not gain mismatch which is the cause of the high currents he observed. We stated earlier in our analysis of the current mirror that we assumed that the transistors are exactly equal in terms of both characteristics and environment. Usually, the most important environment parameter is temperature.

SPICE sets a default temperature of 27°C for all components unless otherwise instructed. We can set the temperature for all components using the SPICE directive .OPTION temp=value but in this case we need to set the temperature of an individual transistor. To change the temperature of a transistor we right click the type/model name text and add temp=value after the model name, for example, for Q2, with model BC337-25X2 at 65°C the transistor type label becomes BC337-25X2 temp=65.

If we set the transistor gains back to their default value, but keep the Early voltage at zero and resimulate using a sweep of R, as before, but with Q2 at 65°C we get the results shown in Fig.9. There are a couple of key things we need to discuss regarding these results.

**Results**

First, we can see that the current values are not exactly the same as *james’s* values, but they are consistent with
In practical terms, with discrete transistors, we could both at the same temperature the mirror works correctly. If the transistors are not held in thermal equilibrium, then the mirroring breaks down, and they are free to self-heat and increase their currents independently within these limits. If are they forced to have the same temperature then this cannot happen and the current in Q2 will be set by the current in Q1.

**Matched**

Matched transistors are often used in IC designs in current mirrors and other circuits. IC designers use layout techniques to ensure the transistors are closely matched, both in terms of characteristics and temperature. Two transistors close together on the same chip are likely to be at similar temperatures, but gradients in temperature across the chip are still possible. One approach to counteracting this, which is not applicable to discrete transistors, is to split each transistor into two parts and place them on the silicon so that they have a common ‘centre of gravity’: this is known as common centroid layout. Using this approach, any linear gradients (e.g., in temperature and dopant levels) will have, on average, equal effect on both transistors, ensuring a close match.

Even if the transistors are well matched, the circuit in Fig.1 does not produce a very accurate copy of the reference current. However, there are better circuits available, for example the Wilson current mirror (see Fig.11), which uses three transistors and has better matching and better output characteristics (higher current-source internal resistance).

---

**ByVac introduces ByPic and Internet of Things**

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Passing of the torch and concluding the LPLC Oscilloscope

This month we say farewell and good luck to Mike Hibbert, who has written the PIC n’ Mix column for over a decade. Mike has kindly passed the torch to me, and I am truly delighted and honoured to be taking over his position. I’m sure it’s not the last we’ll hear from Mike – I’m hoping to persuade him to do a guest article or two for us in the future. As a close friend, I wish him every success in his new ventures.

So, who am I? I’m an electronic engineer and I have loved building gadgets since the age of nine – building my very first radio with a lollipop stick attached to a ferrite rod, which moved up and down the centre of a poorly wound coil made from a Wotabix box. I have built, fixed and wrecked many gadgets in the pursuit of knowledge and fun. One of my latest designs is a Teslatronix Tesla Coil, which illuminates LEDs and energy saver bulbs up to a foot away using a 9V PPC battery.

I’ll be completing Mike Hibbert’s PIC-based LPLC Oscilloscope, using the LPLC board developed from his very successful Kickstarter campaign.

Let’s get a quick reminder of what the project involved. First, we looked at developing our own microcontroller circuit with a custom-made PCB back in January 2014. After a month’s request, the idea of starting a Kickstarter project was conceived, resulting in a very successful campaign. A total of 430 LPLC boards were sold and shipped to those who supported the campaign (including me).

I used my LPLC board for a test system in manufacturing. It had a number of ADC inputs to monitor signals and voltages under varying conditions. When everything was good, a number of green LEDs were illuminated, when anything was bad, the corresponding red LED would light – nice, simple and effective.

Back to our PIC Oscilloscope

What’s next? We’ve hooked up the screen, power, ground, our SPI interface, the reset and other signals. We’ve programmed the PIC and we see the grid on the screen, with a green horizontal line representing our signal.

We want to keep our LPLC Oscilloscope simple, but at the same time, as realistic as possible – so, the aim is to clean up the layout of the screen and add more, genuinely useful functions. I’ve changed the layout of the screen to be more symmetrical, there are now eight squares along the horizontal and four in the vertical, giving a nice even 32 squares. I’ve changed the maximum voltage to 4V, so we can always see the top of the waveform and to keep it nice and even. I’ve also adjusted the ADC values in the code to match 3V on the screen. This can all be seen in the code available on EPE’s website.

So, the features we are going to add are:
1. Ability to measure voltage level on screen
2. Rising-edge/falling-edge triggers
3. Automatic sampling/normal sampling
4. Stop sampling and HOLD current sample on screen
5. Infinite persistence

Buttons first

Let’s get started, in order to change values on the fly without having to program our LPLC board every time, we’re going to need a few buttons – four is sufficient. This allows us to cycle through various options. One button can be used solely for stopping the sampling, and holding the current trace on the screen. The next button can be used to select the various items around the screen; we can change each item to yellow to indicate it is selected. The next two buttons are ‘up’ and ‘down’ for the current selected item. Now we must connect some push buttons. Ideally, we want to have our inputs on the PIC toggle between high and low to indicate the button has been pressed. The best way is to connect each of our inputs to ground using 1kΩ resistors. Then we connect one side of our push buttons to the input pin of the PIC and the other side of the button to our 3V supply. We expect logic 0 at all times – except when a button is pressed and then we see logic 1.

Now we have the hardware sorted out, we need some code to act on the button presses. For our input pins for the buttons, we’re going to use the following ports on the PIC: PORTC6, PORTC7, PORTB0 and PORTB1. We initialise these ports as inputs in our existing function Hardware_Init() in the hardware.c file.

TRISBbits.TRISB0 = 1;    // Setup as Input
TRISBbits.TRISB1 = 1;    // Setup as Input
TRISCBits.TRISC6 = 1;    // Setup as Input
TRISCBits.TRISC7 = 1;    // Setup as Input

Next, we create a function called CheckforButtonPress() in the hardware.c file (Listing 1), which checks to see which button has been pressed; it will either toggle between the options or increase/decrease a value. PORTC6 will be used to halt or stop the current sampling, PORTC7 will be used to select between the various on-screen options, PORTB0 and PORTB1 will increase and decrease the selected on-screen value. Only the code for ‘HOLD’ and ‘SELECT’ are shown – the rest of the code can be downloaded from the EPE website. We add the delays (500); command as a debounce for our buttons, so that they are not pressed more than once in a given (short) period.

Fig.1. Push-button schematic
Now we look at what we want on the screen. We'll need an on-screen trigger level. We create a function `LCD_DrawTriggerLine()` and a global variable `TRIGGER_LINE` to draw the line on the screen. This function will be called every time a new capture is available. Now all we have to do is increment/decrement the `TRIGGER_LINE` variable and the change will be seen next time a capture is available. Listing 2 is the code for the function. In the function, we must remove whatever line we had previously. I've also added the trigger-level value to the bottom-right of the screen in mV. So, now we can set and see the trigger-level value.

It's also nice to have a measurement line, which will indicate the voltage level where this line sits. This is just the same code as in Listing 2, except we use the `MEASURE_LINE` variable instead of `TRIGGER_LINE`.

### Rising/falling-edge triggers

Now we want to see a transition from high to low or low to high (also known as rising and falling edges). This is one of the most useful aspects of any oscilloscope, as it allows us to see if a signal is behaving as it should. From the code above, we now have a `TRIGGER_LINE` variable, which we can use to see if a value has gone from high to low.

Let's create a variable called `EDGE_SAMPLE`, which indicates we want a rising edge when it is 1 and we want a falling edge when it is 0.

Now, in order to find a rising or falling edge, we need to look at the ADC values we're capturing. We edit the function `handle_int_ADC` in `lplc_adc.c` and add the following lines inside the `if(!samples_ready)` statement. What we want to do here is look at the previous ADC value, compare it against the new ADC value and see if it is either a rising or falling edge. If either are seen, we assert the `RISING_EDGE` and `FALLING_EDGE` global variables – see Listing 3. We will come back to this later to bring it all together.

### Automatic/normal sampling

Another useful feature of the oscilloscope is the Automatic or Normal sampling features. Automatic is just a free running sampling mode, which updates the screen every time a new sample is ready. However, the Normal sampling feature is much more useful, in that

```c
void CheckforButtonPress(void) {
    if (PORTCbits.RC6) { // Hold Sampling Button
        // Assert HOLD_SAMPLING if not already asserted
        // and stop writing to screen
        // Clear Screen and start sampling again if
        // HOLD_SAMPLING is already asserted
        if (HOLD_SAMPLING) {
            LCD_Clear(BLACK);
            POINT_COLOR = GRAY;
            LCD_DrawRectangle(10,10,239,319);
            LCD_DrawScreenWords(BUTTON_SELECT);
            HOLD_SAMPLING = 0;
        } else {
            HOLD_SAMPLING = 1;
        }
        delayms(500);
    }
    // Select Screen words code
    if (PORTCbits.RC7) {
        // Checks Button Select
        // Increment button selected, LCDDrawScreenWords
        // function chooses colour of on screen selection
        if (BUTTON_SELECT == 3) {
            BUTTON_SELECT = 0;
        } else {
            BUTTON_SELECT++;
        }
        LCD_DrawScreenWords(BUTTON_SELECT);
        delayms(500);
    }
    // Check if Up Button has been pressed and increment
    // values if needed and debounce button
    // Check if Down Button has been pressed and decrement
    // values if needed and debounce button
}
```

```c
void LCD_DrawTriggerLine(void) {
    // Initialise variables for function, calculate mV line
    unsigned char trig_char[16];
    unsigned long int trig_temp = TRIGGER_LINE-11;
    trig_temp = trig_temp * 190;
    trig_temp = trig_temp / 10;
    sprintf(trig_char, "%d mV", trig_temp);
    // Check if TRIGGER_LINE has been selected,
    // if selected colour is yellow, red if not
    // Write our new line to the screen
}
```

```c
// Take the last sample value and store it in prev_sample,
// Increment the sample_index to store new adc value
prev_sample = samples[current_buff][sample_index++];
// scale the sample to our range of 11-238, or 11 + 0-227,
// Take latest sample and store in next_sample
next_sample = samples[current_buff][sample_index];
// Compare previous sample and current sample versus our
// TRIGGER_LINE global variable
if((prev_sample <= (TRIGGER_LINE)) && (next_sample >= TRIGGER_LINE) && (!RISING_EDGE)) {
    FALLING_EDGE = 1;
} else if((prev_sample >= (TRIGGER_LINE)) && (next_sample <= TRIGGER_LINE) && (!FALLING_EDGE)) {
    RISING_EDGE = 1;
}
```

### Trigger level

Now we look at what we want on the screen. We'll need an on-screen trigger level. We create a function `LCD_DrawTriggerLine()` and a global variable `TRIGGER_LINE` to draw the line on the screen. This function will be called every time a new capture is available. Now all we have to do is increment/decrement the `TRIGGER_LINE` variable and the change will be seen next time a capture is available. Listing 2 is the code for the function. In the function, we must remove whatever line we had previously. I've also added the trigger-level value to the bottom-right of the screen in mV. So, now we can set and see the trigger-level value.

It's also nice to have a measurement line, which will indicate the voltage level where this line sits. This is just the same code as in Listing 2, except we use the `MEASURE_LINE` variable instead of `TRIGGER_LINE`.

### Rising/falling-edge triggers

Now we want to see a transition from high to low or low to high (also known as rising and falling edges). This is one of the most useful aspects of any oscilloscope, as it allows us to see if a signal is behaving as it should. From the code above, we now have a `TRIGGER_LINE` variable, which we can use to see if a value has gone from high to low.

Let's create a variable called `EDGE_SAMPLE`, which indicates we want a rising edge when it is 1 and we want a falling edge when it is 0.

Now, in order to find a rising or falling edge, we need to look at the ADC values we're capturing. We edit the function `handle_int_ADC` in `lplc_adc.c` and add the following lines inside the `if(!samples_ready)` statement. What we want to do here is look at the previous ADC value, compare it against the new ADC value and see if it is either a rising or falling edge. If either are seen, we assert the `RISING_EDGE` and `FALLING_EDGE` global variables – see Listing 3. We will come back to this later to bring it all together.

### Automatic/normal sampling

Another useful feature of the oscilloscope is the Automatic or Normal sampling features. Automatic is just a free running sampling mode, which updates the screen every time a new sample is ready. However, the Normal sampling feature is much more useful, in that
we only want to update the screen when a trigger event has happened. Something like our rising_edge and falling_edge triggers mentioned earlier.

Ideally, we would like our transitioning signal to occur in the middle of the screen, but I think we will just stick to triggering anywhere on the screen. We need some boolean algebra here – we want to enter this if statement if the following conditions are true:

1) It is not the first sample
2) And auto sampling is not enabled
3) Or a falling edge has been seen when edge_sample is a logic 0
4) Or a rising edge has been seen when edge_sample is a logic 1

if ((!first_sample) && ((!auto_sample) || ((falling_edge && !edge_sample) || (rising_edge && edge_sample)))) {
    ...
}

Stop Sampling and Hold current sample on screen
Suppose we want to capture a signal doing something on the screen, but we keep losing it. What we need is the ability to press ‘stop’ on a signal we’ve just captured and have a look at it. Using a unique button dedicated to holding a signal, we can easily capture our on-screen signal. First, we grab ourselves a global variable, hold_sampling. When it is 0, we let the sampling run free under whatever conditions it’s in. When it is 1, we stop writing to the screen. We don’t necessarily have to stop the ADC from running (we could if we wanted to and thus save some power). We can implement this change in the exact same place as earlier when we change from automatic to normal sampling. We just add a boolean and to our if statement and we’re good to go. When we press the button again, the hold is released and we continue writing to the screen, see below:

if ((!first_sample) && (hold_sampling) && (!auto_sample) || ((falling_edge && !edge_sample) || (rising_edge && edge_sample))) {
    ...
}

I also thought it would be nice to remark on the fact that we are not running, by printing the word ‘hold’ in big bold red letters, top right of the screen.

Infinite Persistence
Infinite persistence is another useful feature. It’s where the sample line persists on screen or doesn’t go away. This is actually very easy to do. At the moment, we draw over our last samples in black and draw our new samples in green (in effect deleting our last line). Instead of deleting our previous lines, we can just leave them there and draw our new samples wherever they may lie. In the main loop, when we write over our existing green line, we add a simple if statement and we have our infinite persistence.

When is infinite persistence actually useful? Imagine we want to see how a voltage rail into our device behaves? May be we’ve added a Wi-Fi or GSM module to our device. Both of these technologies cause regular instantaneous 2A current draws when trying to connect. If there isn’t sufficient current or it doesn’t provide the current fast enough, then we will get a voltage drop. This is when infinite persistence becomes particularly useful. We can capture our voltage minimums, maximums and any spurious voltage spikes.

if(!infinite_persistence) {
    LCD_DrawLine(old_samples[lpy-1], 10+lpy, old_samples[lpy], 11+lpy);
}

Wrapping it all up
So, we’ve connected buttons to allow us to interface with our board – plus, we’ve written the code to allow us do

Fig.2. LPLC Oscilloscope showing a stopped rising-edge trigger
something with those buttons. We’ve added edge-sample triggering for free running (auto sampling) and triggered running (normal sampling). We’ve also added a stop button and infinite persistence. We bring it altogether and now we can look at various signal inputs connected to our PIC. Fig.2 shows a poor rise time from the output of a function generator. Alternatively, we could output a pulse-width modulation signal from one of our CCP (Capture/Compare/PWM) pins and connect it up to the input PORTA0.

To improve on this design, there are a few things we could add. One of the inherent problems with the current PIC is we assume we know the value and consistency of the PIC’s VDD input voltage – after all, this is what our ADC values are based on. If this changes, then our measurements will be incorrect. In order to measure our own voltage without using any external measuring circuits, we could use the band gap reference that other PIC microcontrollers use (like the PIC24F family). This is a stable, internally regulated voltage reference at 1.2V. We could use this reference to measure input voltages.

We could also add improved circuit protection against ESD and damaging voltage spikes. We will come back to our DIY LPLC Oscilloscope design later in the year with a bigger and better PIC and use it to increase the sample rate into the MHz range.

Next month
In November’s issue we’ll take a look at MikroElektronika’s mikroMedia for PIC32. This board uses the same size screen as our LPLC Oscilloscope, but includes a touch screen interface. It uses the PIC32MX460F512L and a stereo MP3 codec chip that allows us to develop multimedia apps with ease.

Not all of Mike’s technology tinkering and discussion makes it to print. You can follow the rest of it on Twitter at @MikePOKeeffe, and from his blog at mikepokkeffe.blogspot.com

Everyday Practical Electronics, October 2015
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RIAA equalisation – Part 4

Testing RIAA amplifiers
A convenient way to test an RIAA input is to pass a signal from a signal generator through an inverse RIAA network. This has the opposite frequency response of an RIAA pre-amplifier’s equalisation curve, which hopefully results in an overall flat response when combined. If it deviates, it indicates a problem somewhere, such as bad design or poor component tolerance. A suitable test set-up is given in Fig.28. Fig.29 shows an inverse RIAA circuit based on a design by Reg Williamson, then modified by Lipshitz and Jung in January 1980 in The Audio Amateur – followed by a few tweaks of my own. It’s a passive network, with high ‘insertion loss’ (loading) effect, but this is advantageous since RIAA inputs have very high gain. This enables a better match to the (relatively) high voltage output of signal generators. The curve of the inverse RIAA filter is in Fig.30. The circuit can be built up onto a phono plug, as shown in Fig.31. The whole circuit can then be plugged into left and right inputs in turn to check channel matching.

Amplifier check
I was recently asked to look over a Sansui A-80 amplifier brought into the workshop with the fault, ‘sounds louder on one side’. I checked the phono stage RIAA response, which gave a normal curve shown in Fig.32. In conjunction with the inverse RIAA filter a minor deviation of +0.5dB broadly centred on 2kHz was observed, shown in Fig.33. The ‘fault’ turned out to be a 30dB 4kHz dip in one of the septuagenarian owner’s ears. The small 2kHz rise was enhancing, filling the slight droop most cartridges exhibit between the inductive droop and stylus resonance. Was it deliberate ‘voicing’ on the designer’s part?

System synergy
I checked the Sansui amplifier’s moving coil (MC) input with an Audio-Technica AT32E cartridge, which sounded hissy with high-frequency loss. The MC input stage was simply a common-emitter circuit tacked onto the moving-magnet input. I then tried using audio-matching transformers via the moving-magnet input and it sounded fantastic! An accidental synergy had occurred, a serendipitous event all audio system designers experience sometimes. The hiss was subjectively lower than with a 5534 RIAA input using the transformers. The noise spectrum was also biased more towards the extreme high frequencies, making it much less noticeable. The transformers used were standard industrial audio transformers from Vigortronics (Fig.36) costing about £9.00 each from Rapid Electronics – not the crazy expensive ‘Hi-Fi’ variety. Investigation found the Sansui’s RIAA stage had a JFET on the input. Such JFET designs usually give lower noise with high source impedances compared with bipolar transistor versions. It appeared this was the case here, with the inductive load of the moving-coil cartridge reflected through the transformer accidently providing excellent impedance matching. As every electronic circuit designer knows, when two circuit blocks connected together give unexpectedly good results, a new design is in the making.

The circuit of the Sansui input stage is given in Fig.35, with my transformer connections shown in Fig.37. It is a fairly simple four-transistor arrangement, similar to many small power amplifiers. The standard ‘ring-of-three’ feedback system, comprising input voltage to current (transconductance) stage, voltage amplifier stage (VAS),

---

Fig.28. Using an inverse RIAA network to test RIAA inputs – the output should be flat

Fig.29. Inverse RIAA network – it uses some funny resistor values. The Farnell 1% MSR25 series resistors are recommended

Fig.30. Inverse RIAA network curve
common-emitter stage and output stage. There are some variations to the standard theme, the input transistor is a JFET, the VAS stage is ‘upside down’ using a PNP device, enabling a more common N-channel JFET to be used. Finally, the output stage is a class-AB push-pull, emitter-follower design. This will not generate crossover distortion on normal signal levels, but will have the current capacity to deal with nasty scratches without overloading by going momentarily into class B. This will sound less objectionable than a hard-clipped class A.

**FET frenzy**

Unfortunately, as with most Japanese audio circuits, the Sansui used unusual semiconductors. The 2SK163, (marked K163) JFET from NEC was difficult to source. The other devices were all easily substituted. The K163 and other rare audio JFETs are available as NOS ‘new old stock’ from Donberg Electronics in Ireland. I measured the transconductance ($g_m$) of the K163 with the Peak DCA75 and found it to be in accordance with the data sheet at 17mA/V. This is much higher than standard JFETs, such as the 2N5457, which typically achieve 3-4mA/V. This would reduce the open-loop gain of the circuit by around four times and increase the noise level, since the second stage would have to do more work. The

---

**Fig.31.** Photo of 'hard-wired' inverse RIAA network – it plugs into an RIAA amplifier input

**Fig.32.** Sansui A80 RIAA response curve

**Fig.33.** Sansui RIAA input fed by all inverse RIAA network. Note slight response bump at 2kHz

**Fig.34.** Sansui A-80 amplifier

**Fig.35.** Sansui A-80 RIAA input stage – a simple discrete JFET input design, with a few minor changes and transformers added. The switch on the transformer ground allows the left minus right signal to be listened to, which allows the anti-skating force on the pickup to be set to optimum by minimising distortion on vocal sibilants
K163 is one of those rare 1980s audio JFETs that uses a finely interdigitated channel (see Fig.38) to achieve the high transconductance and low noise. Other suitable types are the 2SK117 and 2SK170. If a P-channel JFET is used, such as the 2SJ74, the circuit could be made to work on a single +48V rail. One side effect is high capacitance, but this is not a problem in this application. Another advantage of FETs is that no input capacitor is needed in this circuit.

Toshiba still make some audio FETs, albeit only in surface mount packs, such as the 2SK209-Y which is designed for condenser microphone head amps. These cost around 25p from Mouser UK and meet all the specs. I expect to pay commodity prices for my parts and I have no truck with the ‘antique-dealer’ pricing endemic in the audio parts industry. The K163 has an unusually high maximum channel voltage of 50V, but the device is only exposed to 17.5V if the rails are ramped up slowly. This allows 30V TO92 package devices to be used. For those interested in all things JFET, Erno Borbely is the circuit designer to watch.

**Wind up**

Finally, to end on a ‘sad’ note, a friend of mine gave me his defunct £500 moving-coil cartridge that had only lasted for ten plays. In an attempt to prevent ‘extraneous resonance’ from a casing, the designer had left the coils exposed – typical Hi-Fi madness. The inevitable happened, one of the fine wires got caught and broke. In a failed attempt to repair the unit, I was able to take a photo showing the inner workings of a moving-coil in Fig.39, and just for interest, there is a ceramic cartridge internal structure shown in Fig.40.

---

**Fig.36. Vigotronix 003 transformers used for converting moving-magnet to moving-coil inputs. Note the essential mumetal screening can to prevent hum pick-up**

**Fig.37. Wiring for a Vigotronix VTX-101-003 moving-coil step-up transformer (pin view)**

**Fig.38. An example of an ‘interdigitated’ JFET – an early low-noise audio design, the BFW10 – from Field-Effect Transistors, Mullard Ltd, 1972; Plate A p13**

**Fig.39. Sumiko Blackbird moving-coil cartridge with broken wire. The 4-coil armature can be seen**

**Fig.40. Ceramic cartridge – the two bi-morph elements are visible**
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Tri-colour LEDs – Part 3

As I always say: ‘Show me a flashing LED, and I’ll show you a man drooling.’ This phenomenon is even more pronounced in the case of tri-color LEDs. In my first column on these little scamps (EPE, July 2015), I considered a traditional tri-colored device with a common-cathode terminal and three anode terminals – one each for the red, green, and blue sub-LEDs. As part of this, we noted that if we limited ourselves to simply turning these LEDs on and off, then we can achieve $2^3 = 8$ different colour combinations (assuming the all-off + black combination is a colour). These colours are black, red, green, blue, yellow (red + green), cyan (green + blue), magenta (red + blue), and white (red + green + blue).

The big problem with these little rascals is that the fact that each tri-color LED requires three microcontroller (MCU) pins to drive it.

In Part 2 of this mini-series (EPE, August 2015), we discussed driving each LED terminal using a pulse-width modulated (PWM) MCU output. In the case of the Arduino, we have 8-bit PWMs, which theoretically allows us to achieve $2^8 = 256$ different brightness values, from 0 (fully off) to 255 (fully on). Since we can do this for each sub-LED, this means that we can theoretically achieve $2^8 \times 2^8 \times 2^8 = 16,777,216$ different colours. This is all jolly nice, but we are still faced with the fact that each tri-color LED requires three microcontroller (MCU) pins to drive it, which does tend to limit you as to what you can achieve. Happily, there is a way by which we can drive hundreds of tri-coloured LEDs using a single MCU pin... so, how do we do that? – read on!

Introducing the WS2812

A couple of years ago, a company called Worldsemi introduced a device called the ‘WS2812’. This little beauty is presented in a 5mm × 5mm square package that contains super-bright red, green, and blue LEDs, beauty is presented in a 5mm × 5mm square package introduced a device called the ‘WS2812’. This little

Fig.1. A WS2812 tri-colour LED

The great thing about these devices is that they have four pins: 0V (ground), 5V (power), Data-In, and Data-Out. This means that they can be daisy-chained together, with the MCU driving the Data-In of the first element, the Data-Out from the first element driving the Data-In of the second, and so forth, as illustrated in Fig.2.

Different companies have used these devices as the basis for a wide variety of products. For example, the folks at Adafruit have a whole range of WS2812-based products that they call NeoPixels. These range from strips to rings to panels to ... all sorts of things – see: [http://bit.ly/1gr71a3](http://bit.ly/1gr71a3).

I’ve been using NeoPixels a lot in my hobby projects, such as my [Bodacious Acoustic Diagnostic Astoundingly Superior Spectromatic (BADASS) display](http://bit.ly/1I5ncTc) and [http://bit.ly/195JKWR](http://bit.ly/195JKWR). The great thing about Adafruit is that they give you a NeoPixel software library along with a bunch of examples on how to use it.

One thing that can confuse beginners is that the Adafruit library includes a function called [setPixelColor()](http://bit.ly/195JKWR), but this doesn’t actually update the physical pixel itself. Take my BADASS display, for example. Suppose we use the Adafruit library to declare a string of 256 NeoPixels called myNeos (these will be numbered from 0 to 255). Now suppose I use the following statement, where the three parameters of 255 are associated with that pixel’s red, green, and blue channels, respectively:

```c
for (int i = 0, i < 256; i++)
  myNeos.setPixelColor(i,255,255,255);
```

What do you think will happen on the display? The answer is ‘not a lot.’ This is due to the fact that, when we first declare our NeoPixels, the system creates an array in the Arduino’s memory. This array contains one main element for each NeoPixel, and this main element is comprised of three byte-sized sub-elements to contain that pixel’s red, green, and blue values. The [setPixelColor()](http://bit.ly/195JKWR) function actually modifies the values in this array. When we are ready to update the display, we would use the [show()](http://bit.ly/195JKWR) function; for example:

```c
for (int i = 0, i < 256; i++)
  myNeos.show();
```

The [show()](http://bit.ly/195JKWR) function streams the entire contents of the array from the Arduino’s memory into the NeoPixel string. In the above example, this means that all of the pixels will turn fully on simultaneously. Now, assuming that all of the pixels are already on, how would we turn them off one at a time with a tenth of a second (100ms) delay between each one? Well, we could achieve this as follows:

```c
for (int i = 0, i < 256; i++) {
  myNeos.show();
  delay(100);
}
```

As you can see, we’ve moved the [show()](http://bit.ly/195JKWR) function inside the [for()](http://bit.ly/195JKWR) loop, and we’ve also added a [delay()](http://bit.ly/195JKWR) function inside the loop so that we can see the pixels being extinguished one after another.
How much power?!

Now, one thing you have to consider is how you’re going to power the little ragamuffins. Your knee-jerk reaction may be along the lines of ‘LEDs are very efficient and they don’t consume much power.’ Well, that’s as may be, but the power starts mounting up when you use a lot of LEDs. Adafruit’s website states that each NeoPixel can consume up to 60mA when its red, green, and blue channels are all fully on. In the case of my BADASS display, I have a $16 \times 16 = 256$ array of NeoPixels, which equates to $256 \times 60mA = \text{(approx) } 15.5A$ (I actually use a 20A supply to give myself some ‘wriggle room’).

You may decide to power your NeoPixels using a separate power supply to that used to power your Arduino. In this case, it’s very important to ensure that the ground (GND, 0V) of this power supply is connected to the Arduino’s GND pin. It’s also a good idea to mount a $1,000\mu F$ electrolytic capacitor across the power supply rails, as illustrated in Fig.3.

Furthermore, it’s a good idea to insert a resistor in series with the data signal coming from the Arduino. This resistor, which is used to dampen overshoot and undershoot on the signal, should be mounted close to the first element in the NeoPixel chain (ie, not at the Arduino end of the signal path). Adafruit’s website says the value of this resistor should be $300\Omega$ to $500\Omega$. I typically use a $390\Omega$ resistor. (NeoPixels are also prone to being damaged by electrostatic discharge (aka ESD) so you should take appropriate earthing precautions while using them.)

In the case of my projects, I typically modify my Arduino so that it works from a 5V supply (the unmodified Arduino requires a 7V to 12V supply, which is subsequently regulated down to 5V), and then I use the same regulated supply to power both the Arduino and the NeoPixels. However, this involves removing and bypassing the Arduino’s on-board regulator and tweaking a few resistor values, so it’s not for the faint of heart.

Using the Arduino 5V pin

The other alternative is to power the NeoPixels from the Arduino’s own 5V supply pin, but this means you are limited as to how many NeoPixels you can use. For example, I’m teaching a young lad called Jacob about using microcontrollers – he comes into my office for an hour each week. Not surprisingly, Jacob would like his own BADASS display, so we built him a mini version, as shown in Fig.4 (check out this YouTube video http://bit.ly/1UalHbP).

Now, Jacob powers his Arduino from his PC’s USB port, which is only specified to provide 0.5A (500mA). When running at full pelt, an Arduino Uno R3 draws around 45mA, and the spectrum analyzer shield only consumes about 3mA, so if we assume a worst-case of 50mA, this leaves us with 450mA to play with. Although Adafruit says that each NeoPixel can draw up to 60mA when fully on (20mA each for the red, green, and blue sub-pixel LEDs), we’ve actually put a string of NeoPixels on the test bench and measured them as drawing an average of 43mA full on, which we rounded up to 45mA (or 15mA per color). So... in the case of Jacob’s mini display, we calculated that if we have three columns per channel – one each for the base, middle, and treble – and if each column has five NeoPixels – and if each NeoPixel is used to display only one colour (red, green, or blue), then we end up with $6 \times 5 \times 15mA = 450mA$ (I love it when a plan comes together).

‘Ah ha!’ I can hear you say, ‘but Fig.4 shows you using red, blue, and yellow (which is composed of red and green) – how can this be?’ The answer is really simple – so simple, in fact, that I’ll leave you to work it out for yourselves!

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This Teach-in series of articles was originally published in EPE in 2008 and, following demand from readers, has now been collected together in the Electronics Teach-in 2 CD-ROM.

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Also included are 29 PIC N’ Mix articles, also republished from EPE. These provide a host of practical programming and interfacing information, mainly for those that have already got to grips with using PIC microcontrollers. An extra high part breadboard layout for using the C programming language for PIC microcontrollers is also included.

The CD-ROM also contains all of the software for the Teach-in 2 series and PIC N’ Mix articles, plus a range of items from Microchip – the manufacturers of the PIC microcontrollers. The material has been compiled by Werbome Publishing Ltd. with the assistance of Microchip Technology Inc.

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Everyday Practical Electronics, October 2015
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```c
} delay(30);
// wait for 30 milliseconds to see the dimming effect
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9W Stereo Amplifier

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! These projects range from around £15 to under a fiver… bargains!

9W Stereo Amplifier

Need a cheap audio amplifier – perhaps as a test bench amplifier or for a kid’s MP3 sound system? Well it doesn’t get much cheaper than this – try under £4 for the prebuilt amplifier module, including postage to your UK letterbox!

So what do you get for your measly £4? Well, it isn’t exactly ‘Hi-Fi’ – but then again, neither is it anywhere near as nasty as you might first think! (See Fig.1) The module uses a Tripath TA2024 integrated circuit that delivers 9W per channel into a 4Ω load at a total harmonic distortion of 0.04%. Crank it up to 15W per channel and distortion rises very substantially to 10% – so you don’t want to overdo it! Dynamic range is 89dB. The module comes fully built and is 92mm x 53mm in size.

Connections

At one end of the module is a terminal strip to allow connection of the two speakers, and power and ground (Fig.2). The board is spec’d at 12V, but it will work down to 7V. Note that it is not reverse-polarity protected.

At the other end of the board is a three-terminal connector for left, right and ground audio inputs (Fig.3). A short input cable is supplied with an appropriate plug and lead. At 10W audio output power, the manufacturer rates the efficiency of the IC at about 78% (with 4Ω speakers), so the chip does not need a heat sink. In fact, in our testing, it stayed relatively cool even when running continuously at full power on audio material. The cheapest and easiest approach to powering the module is to use a discarded mains-powered plugpack. Any design with a DC output of around 8-12V will work fine.

Driving the amp

Many people will want to run the module from a phone or other MP3 player. Cheapest is to get an old pair of earphones (or any other discarded cord) that has a 1/8-inch stereo plug already wired into place. Just cut off the lead and make the connections to the input cable supplied with the module (Fig.5).
The conductor that connects to the tip of the plug goes to the red input wire of the module; the central conductor on the plug to the white input lead; and the base of the plug to the black input. Connect speakers to the appropriate screw terminals, switch on power and listen to music! The output level of an iPhone that we used to test the module worked well – the amp input was not overloaded at even full phone output.

Battery powered?
What about running the module off a 9V battery? At 9V and driving both channels from an iPhone, we measured an average current draw of just 60mA at a low listening level, 80mA at a moderate level, and 185mA at a loud level. That makes battery operation entirely feasible.

Conclusion
As I said at the beginning, don’t even pretend that this is a full-blooded Hi-Fi amplifier that will shake the house. But where you need a utility low-powered amplifier, at just £4 plus an old plugpack, it's pretty hard to go wrong.

Sourcing
At the time of writing, a typical online unit is eBay item 351092217368 at a very reasonable £3.88 incl delivery.

Next month
Fancy an electronic stethoscope? Here is a really neat device that will allow you to listen to all sorts of strange noises in machines and other mechanical devices. Just add some headphones and shielded microphone cable. All this is in our next Electronic Building Block article.

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Valve-based hearing aid

Dear editor,

There’s no need to ‘...imagine a valve hearing aid!’ (EPE, June 2015, Audio Out, Ge-mania – Part 3, p.52) because here’s a picture of me modelling one (my thanks to Christine Mlynek for the photo). The Medresco (MEDical RESearch COuncil) OL35A was one of the first NHS hearing aid patterns. The OL15A is slightly earlier, but the differences are subtle. I haven’t an ear moulding so the earpiece is hooked loosely round my left ear. The dangling leads are for the heater and HT (‘high tension’) batteries, again I have no originals, but they were probably larger than the aid itself. It does work, an AA cell with a touch of series resistance provides the heater supply and a 23A compact 12V battery is sufficient for the HT, even though slightly lower in voltage than specified. Remember that the zinc-carbon cells of the day dropped from their rated potential difference quite quickly, and so circuits had to be tolerant of a lower supply. Also shown is a photo of the circuitry.

The sub-miniature valves are wire-ended and originally intended for guidance/fusing in late wartime ordinance. This is far smaller than the more commonly accepted miniature B7G and B9A valves. The remaining picture shows one of the wire-ended valves (second from right) with an OC59 (still in original package), as also seen in Jake’s Ge-mania, Part 3 article in Fig.26 – the transistor replaced valves in smaller aids. Valve designers were aiming to make valves smaller still, one more place to the left is a 6CW4 Nuvistor in its shiny metal can.

Jake mentions Sinclair’s transistors rescued from disposal by Plessey. Clive Sinclair claimed to have involvement in the novel microalloy devices – they were actually the poor-spec Plessey throw-outs with a crude paper label bearing his own non-standard MAT type numbers (a MAT100 still in its packet is seen at the far left of the photo). Transistors didn’t start this way, of course. The very first offerings were point contact, as exemplified by the rare GET1 seen at the far right of the picture – sadly, non-working (do not confuse with GE numbered devices).

Tumbling audio caps!

On a completely different tack, I’ve just had to replace the motor-run capacitor in my tumble dryer (which stopped tumbling when the capacitor failed for no obvious reason!). What’s that to do with audio? These are non-polarised (not electrolytic) caps in a range of values in the several µF class. ESR measurements are very low. Lateral thinking: could they be used for audio? They’re big, but not massive. Available in a metal screening can which is isolated from the terminals.

Godfrey Manning G4GLM, Edgware

Jake Rothman replies:

I thank Godfrey for his extra information, and I have appreciated his letters to Practical Wireless in the past. The photo of the GET1 point-contact transistor is an especially useful addition, since I couldn’t get one. I did try making a point-contact transistor with a couple of needles probing a germanium slice recovered from a shattered OA10 diode. I think I got...

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Getting back into electronics

Dear editor,

I have worked in many different electronic disciplines, but have not had ‘hands on’ experience, including hobby projects, for over 40 years. Somehow, the advent of integrated circuits removed the magic of construction.

Recently, however, having almost reached the age of seventy-nine, I was struck with an urge to build something again. Quite by chance I discovered a copy of your January 2015 issue, on a market stall, containing an article on constructing the Tiny Tim Amplifier; this seemed the ideal project to satisfy my construction urge.

How things have changed! I’ve spent countless hours trying to understand, what is new to me, the various acronyms, trying to source individual components, as firms supplying kits of parts, common in my past, seem to have disappeared – but the biggest stumbling block has proved to be a source for the specified heat-sinks. Neither Farnell nor Radiospares, the two suppliers of Element14 parts, recognise the part number or description quoted in the parts list! Unless you can supply me with your source for the experienced audio enthusiast who would need to make a substitution. These are only small TO-220 type power transistor heatsinks and appear to be nothing special.

In my experience they can be freestanding, clip-on or they have extra PCB pillars. A wide range is also listed on eBay.

As the Tiny Tim Amplifier evolved from the earlier Hi-Fi Stereo Headphone Amplifier project (October 2014 issue et seq.) I wanted to point out that it is probably much less straightforward to build the Tiny Tim ‘out of the box’. This is not really a cut and dried project, but one that I cannot remember seeing or knowing of it before. I do, however, remember Practical Electronics and other titles. Is Everyday Practical Electronics descended from these?

GJ Gillson, by email

EPE Online Editor Alan Winstanley replies:

Most of our projects are sourced from Australia’s Silicon Chip magazine, so component ID queries crop up from time to time. Unfortunately, EPE cannot promise 1:1 close technical support, as this would require immense resources on our part. The best place to ask for technical help is in the EPE Chat Zone forum – www.chatzones.co.uk – where many readers are very happy to offer support. A section called ‘Shop Talk’ highlights any late-breaking news or component-sourcing Q&As.

It is very common these days to source parts by mail order/Internet from suggested sources in Australia or eBay for example. Sometimes, full kits are available online (e.g. see Jaycar or Altronics websites). Delivery to the UK is never a problem.

The exact TO-220 12°C/W heatsink appears to be available from Jaycar Australia. Jaycar is a reputable, recommended mail order supplier, who ship constantly to the UK. Go to www.jaycar.com.au and search for Cat No HH8504.

Drawings are available from Jaycar to enable users to make a substitution. These are only small TO-220 type power transistor heatsinks and appear to be nothing special. In my experience they can be freestanding, clip-on or have extra PCB pillars. A wide range is also listed on eBay.

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EPE is indeed 50 years old and our November and December 2014 issues contained special supplements that I wrote, charting our heritage from the 1964 Practical Electronics issue No. 1 to the present day. Everyday Electronics was spun out of PE to appeal to newcomers. For well-documented reasons both titles went their separate ways before merging to become EPE, owned by the present publishers.

Back issues of EPE magazine can be bought from us online or over the phone, please see www.epemag.com for details. Your local WHSmith should be able to order a copy for you, but the best (cheapest) way is to order a subscription from us for direct delivery to your door. I’m afraid it’s a sign of the times – it is ever harder to get a niche title onto the newsagent shelves. Please contact us if you have difficulties with your local newsagent.
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November’s Teach-In 2015 is the final part of the series. We will examine practical measurement, adjustment and fault-finding in power amplifiers. The Knowledge Base section will examine power supplies, and Get Real will explain the procedure for measuring the performance of our high-quality 10W amplifier.

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