Circuit ideas: Class-A power buffer

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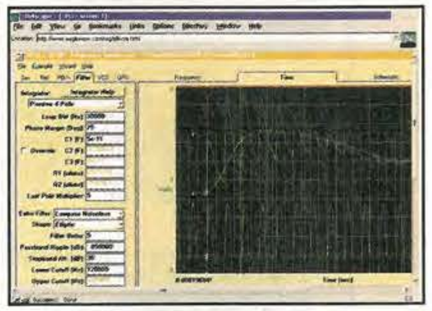
Mobile video – just one new feature that the new third-generation mobile phone technology makes possible. But when? Read Richard Wilson's report on page 373.

Photography : Mark Swallow

Michael Slifkin's novel micro-controlled inductance meter measures from 20μH to 200mH with minimal errors due to ohmic loss – page 424.



A phase-locked loop simulator especially for rf design designers. If you are involved with PLLs, have a look at page 392.

When you're using an oscilloscope, there's a right way and a wrong way to probe a circuit like this. Find out how to do it properly on page 384.

June issue on sale 5 May

Mobile Internet – who wants it?

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REED BUSINESS INFORMATION

As the tortuous auction for 3G licences went on and the prices rose into the billions, did anyone ask themselves: "Who wants mobile Internet?" Already people talk about 'm-commerce', meaning mobile commerce, as the successor to e-commerce.

Will people really transact business deals and compose messages using the tiny keypads of mobile phones? Will people access the Internet using wireless connections which frequently cut off?

Many wonder. Not least Sir Alan Sugar, chairman of Amstrad/Viglen, at one time the parent company of the Danish mobile phone company DanCall. "Who wants to send e-mails from the middle of a field?", asks Sir Alan.

Many agree with Sugar that fiddling around with tiny buttons is not going to be a popular pastime. Those who say speech recognition, not tiny buttons, will be the input mechanism for portable Internet access aren't familiar with the inaccuracies of speech recognition. These require the frequent intervention of buttons for correcting the many errors.

With the technology at its current state of development, is Mobile Internet feasible?

Undoubtedly billions are being spent pursuing it. Marketing 'data' – often pretty spurious – abounds for supporting the proposition that 'by 2003 or 4 or even 5 more half the people accessing the Internet will be doing it from mobile terminals'.

How can these projections be believed? How can the people spending the billions sleep at nights? The reason could be that a very large number of people have a vested interest in making mobile datacommunications catch on. Not least the mobile phone manufacturers. How else can they escape the fate of the four-function calculator?

Back in the 1970s, the four-function calculator became a commodity. Its price tumbled from hundreds of dollars to ten dollars in a few years. The technical reason for that was the effect of the IC – the integrated circuit – then a ten-year old invention but doubling in transistor count every year.

The mobile telephone could easily go the same way as the four-function calculator. The technical reason for that happening is the effect of SOC, or system-on-chip, which is fast becoming possible in most types of consumer product. When you can put the electronic innards of a mobile phone on one chip, the cost of a mobile phone will plummet.

There is, of course, a way of avoiding the nightmare. You can add functionality to the mobile phone. When the makers of calculators were threatened by commoditisation, they added a host of increasingly sophisticated and exotic functions ending up with such esoteric capabilities as div, grad and factorial.

So complicated were some of HP's calculators that, it was said, even PhDs only used 30 per cent of their functionality.

And PhDs do not make a mass market. The mass market wanted only the original four functions: add, subtract, multiply and divide. The attempt to move the mass market to accept – and pay for – more sophisticated functions failed.

The PC fared better. Terrified that commoditisation would happen to it, the PC industry succeeded for a couple of decades in avoiding price declines by having Microsoft produce ever more code-heavy operating systems. These required Intel to produce faster and faster processors and the DRAM-makers to produce denser and denser DRAMs.

Those strategies ensured that, for 20 years, the PC industry successfully averted the commoditisation process as the consumer bought more and more of this increasingly sophisticated – and sometimes functionally superfluous – kit. Mainstream PCs were $1000 twenty years ago and are $1000 now. That's a very clever trick to have pulled off.

Now it is the mobile phone that stands at the cross roads. One road leads to simplicity, commoditisation and declining profits, the other to complication, added value and high profitability. I wonder which will it be? ■

David Manners

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UPDATE

Athlon outdoes Pentium III in 800MHz benchmark tests

The processor war between AMD and Intel has intensified following benchmark results that claim AMD's Athlon is up to twice as fast as Intel's Pentium III in certain situations.

The tests on 800MHz processors, carried out by Bert McComas of Inquest Market Research in the US, also show that virtual-channel DRAM is far better in many cases than standard synchronous or Rambus DRAM.

McComas tested several system configurations based on AMD's and Intel's 800MHz processors. Chip sets were from either Intel or VIA Technologies, the former supporting Rambus, 133MHz SDRAM (PC133) or virtual channel (VC133), the latter PC133 or VC133.

Using LinPack (see diagram) and StreamD benchmarks, McComas showed Athlon to be 50 per cent or more faster than Pentium III.

"Athlon's consistently high DRAM throughput is related to its CPU bus characteristics. At 200MHz it is much more capable of extracting the untapped performance (headroom) of PC133 and Virtual Channel SDRAM," the report stated.

Rambus and PC133 showed little difference on those metrics.

The Inquest report also examined the bottleneck caused by the graphics accelerator.

In his report, McComas said:"As 3D benchmarks are used to evaluate CPU, chipset and DRAM performance, it has become common practice to crank down the screen resolution, colour depth and image quality features to their minimum. On the surface, this may seem logical, but in effect the hottest, fastest PCs on earth are being benchmarked at 1997 screen resolutions."

Therefore he tested the different systems using a high-end Creative Labs board based on nVidia's GeForce graphics chip.

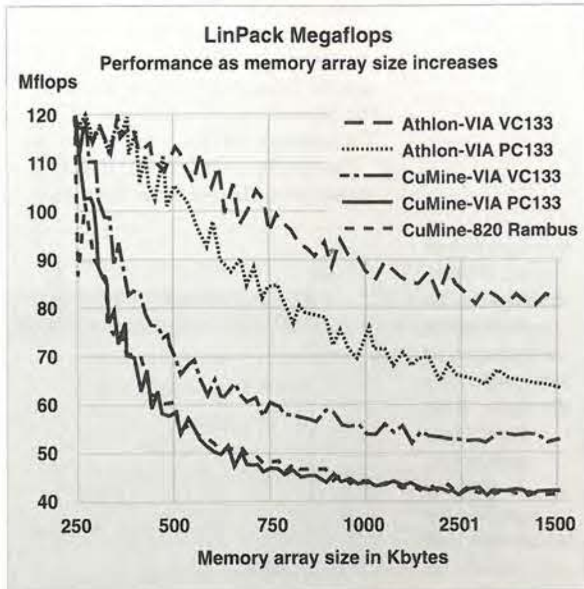
The 3D tests used a Quake3 test program, setting resolution and image qualities to both their highest and lowest settings.

At the lowest settings, AMD comes out worst, due said McComas to its lower speed level two cache. At the highest resolution though, there was no difference between the various systems. Why? Because the bottleneck that is the graphics card completely chokes the system, says McComas.

Indeed, clocking either processor at just 400MHz caused graphics frame rates to drop by just two per cent. Overclocking the graphics chip – up to 140MHz from 120MHz – saw performance rise by 15 per cent, even though the CPUs were running at half speed, proof that graphics is a bottleneck.

The moral of this story – spend your cash on the graphics card and chipset, not the processor.

***Richard Ball** Electronics Weekly*



In the LinPack benchmark, which tests the combination of floating point performance, bus bandwidth and memory performance, Athlon outdoes Pentium III (CuMine) by a considerable margin – a factor of nearly 2x for larger data sets. Virtual-channel memory, on either processor, outperforms PC133 and Rambus. Intel's speed drops quickly as data array sizes increase, AMD is less affected.

New substrate for high-speed processing

A group of ex-Intel workers has developed a method of reducing problems with electrical noise, clock skew, bus speed, and soft errors in high speed microprocessors.

Their start-up firm, Primarion, plans to split a processor's power grid into small independent sections. The die is then bonded to another silicon substrate which provides power regulators for each section of the circuit.

Such a system would improve the quality of the power supply, reducing problems such as ground bounce. Response times to current changes could improve by a factor of 100, the firm claims.

Clocks would also be sent to the processor chip via the substrate, reducing clock skew, noise and crosstalk. Noise could also improve by 100.

Circuits on the substrate will be built using heterojunction bipolar transistors (HBTs) to improve the quality of power and clocks, which are essentially analogue.

Primarion expects its technology will be in widespread use within five years.

In order to develop the technology it has garnered $13m in venture funding, mainly from Intel's $8bn venture fund. The firm has also bought a bipolar fab from Lansdale in Arizona.

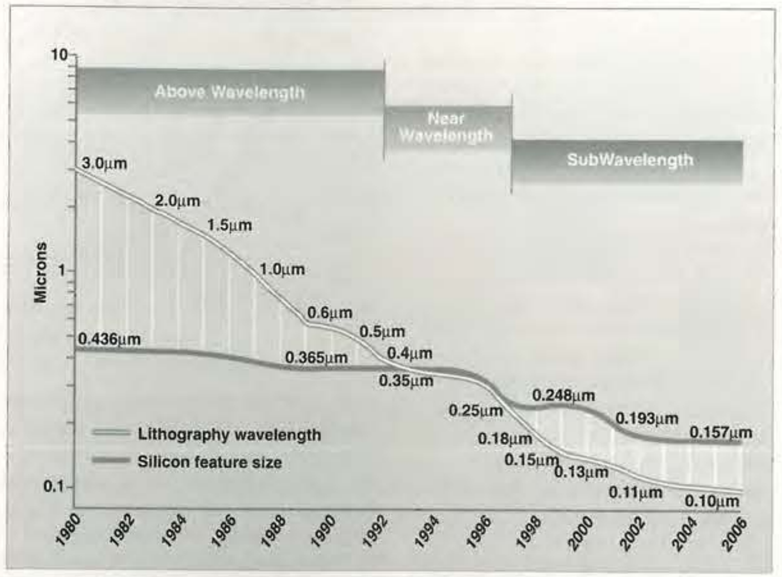
Chip transistors shrink to 50nm in the bid for faster, cheaper silicon

The incredible shrinking transistor has done it again, this time at the hands of Silicon Valley firm Numerical Technologies (NumeriTech) which has produced 50nm (0.05μm) transistors using conventional 248nm optical lithography.

The devices, claimed to be the smallest ever fabricated using optical lithography, were made at the Massachussetts Institute of Technology (MIT).

Using phase-shift mask (PSM) techniques, the previous limit for 248nm lasers was thought to be 100nm feature sizes. Getting to 70nm and below was the realm of 157nm lasers.

"People have been predicting the end of optical lithography for several years, saying that it can't extend beyond 100 nanometers," said Y.C. Pati, president and CEO of NumeriTech. "The MIT Lincoln Laboratory results prove that with prudent use of phase shifting, optical lithography can be extended much further than anyone ever thought was possible."



Sub-wavelength era... Feature sizes used to be larger than the wavelength of light patterning the wafer. With the 0.35μm generation, features equalled the wavelength, but today, at 0.18μm, features are smaller than the 248nm laser light used.

Beyond the 100nm mark, it was expected that new techniques such as ion-beam lithography would be used to pattern silicon wafers.

If NumeriTech's 50nm technique can be commercialised, it could lead to faster and cheaper chips without the need for developing advanced chip manufacturing techniques to create smaller transistors.

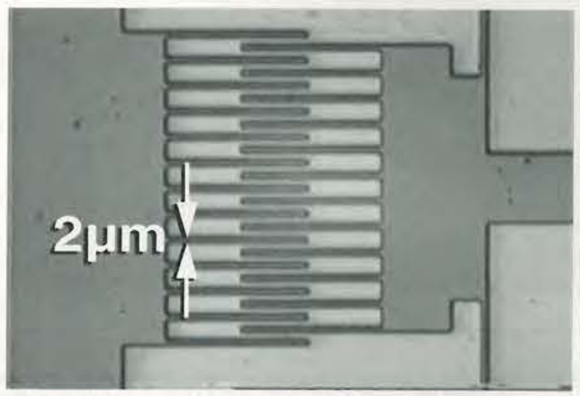
However, PSM cannot get features much closer together because of the interference fringes close to the desired transistor, so you get fast, but not necessarily compact, devices.

PSM relies on the fact that a coherent imaging system will create interference as the laser is diffracted as it passes through the mask. The effect can be used to reduce the spatial frequency of a given object or to enhance its edge contrast.

Both resolution and depth of field of the image can be improved.

To shift the light phase, an extra layer of transmissive material with a different refractive index is needed on the mask.

NumeriTech designs the software that modifies the chip layouts to make use of optical proximity correction and PSM. It is already being used by firms such as Lucent and Motorola to produce features down to 120nm, or 0.12μm.



Diamond micromachines: Sandia National Laboratories in the US has created what it believes are the world's first diamond micromachines. The comb drive is constructed from etched amorphous diamond – the second hardest substance behind crystalline diamond. It is a better material than silicon because it is wear resistant. In addition, it is not rejected by the human body so it could be used for medical biochips. "Micromachines, for their marvellously tiny size, are still machines. They're subject to wear, even if it's only at the micro level," said researcher Tom Friedmann. He believes diamond MEMS could last 10 000 times longer than silicon devices.

Electronic tagging for consumer goods?

RF tagging chips will be taking an active part in the fight against crime if a government Home Office initiative takes off.

"We must use every means at our disposal, including the latest technology, to see how it can best be used or developed to cut crime and make our communities safer," said Home Office minister Charles Clarke at the recent 'Chipping the Goods' seminar.

The plan is to electronically tag equipment by embedding a passive RF tagging device into consumer goods such as TVs, computers and cameras. The chip would carry information about the equipment and its owner. The information stored in the device can be changed in situ using RF technology and can be read remotely when required.

The technology is referred to as RF ID and devices are being developed by several companies including Texas Instruments, Philips, Oxley and Gemplus of France.

The technology should also allow companies to track goods more effectively and so reduce stock losses as the information written to it will specify its last official location. "This is new technology which the Home Office is looking to encourage industry to make use of in protecting its property," said Martin Swedlow, chief executive of IT consultants Integrated Product Intelligence which is involved in organising trials of the tagging.

● The first UK licence for spectrum for an Asset-Tracking Mobile Data Network has been awarded by the Radiocommunications Agency to QNL (UK).

The licence is for the 866-868MHz range, which is causing concern to industry body The Low Power Radio Association, which fears it may cause interference with equipment offered by its members.

Melanie Reynolds *Electronics Weekly*

Researchers develop self-assembling micro-wires for multi-chip modules

Self-assembling micro-wires could form the basis of a new interconnection system for multi-chip modules.

Researchers at Leeds University formed the wires from stacks of disc-shaped molecules in a 'discotic' liquid crystal. These stack themselves under certain conditions to form columns like piles of plates, as in the diagram below

Stacking on any flat surface can be provoked predictably, forming a forest of independently-conducting wires like bristles on a brush. "The molecules will generally orient on anything," said Professor Richard Bushby of the Leeds team developing the technology.

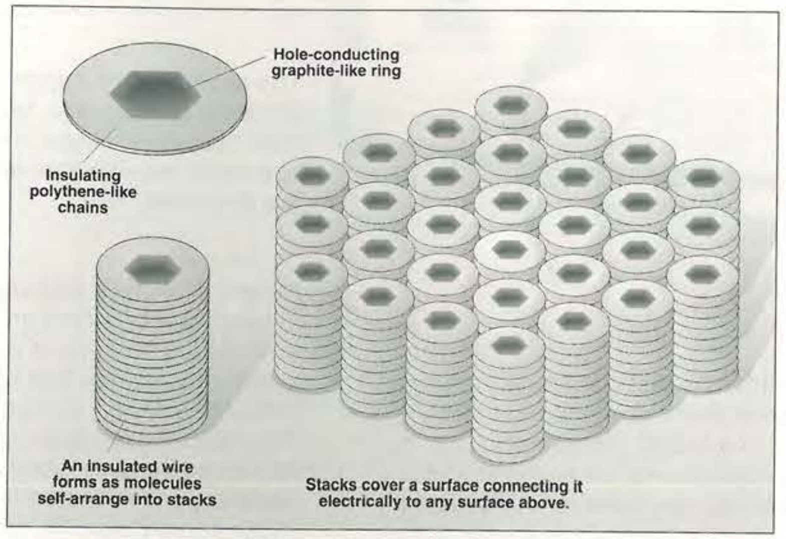
Conductivity along the wires comes from a graphite-like six-atom carbon ring in the centre of each molecule. These can lose an electron leaving a hole which, once the molecules are stacked, can migrated up the centre of the column. Surrounding polythene-like hydrocarbon chains form the wire's insulator.

If two chips are positioned face to face with a small gap between them, any conducting pad on one chip will be connected to a corresponding conducting pad on the other once the space is filled with the liquid crystal. Insulating layers on the chip surface prevent conduction where it is not wanted.

At the moment, pads 10μm across have been connected. Bushby expects this to be reduced to sub-micrometre levels in the future.

Resistance is a problem. "We have improved it by a couple of powers of ten over the last two years and need a couple more powers of ten," said Bushby.

The Leeds team is to receive a share of a £1.5m grant from the EU as part of a 3D-chip programme. Other members are: the University of Delft, which is developing ways to produce conducting vias through chips, CNRS in Paris, which is working on a conductive polymer interconnection system to achieve the same ends as Leeds, and University College London which is looking at some related visual image processing techniques.



Steve Bush *Electronics Weekly*

Digital watermarking finds its way into maps

Signum Technologies has done a deal with Ordnance Survey to apply its digital watermarking technology to the OS's library of digital maps.

"Some people think they can just copy a map, use it commercially and get away with it," said Carl Calvert, from the OS's intellectual property division.

While the OS is self funded, any shortfall in revenue if made up from taxpayer's money.

"It's one thing to photocopy a map so someone can find you easily. It is quite another to use the information that costs us millions of pounds every year to provide and then use it commercially and unlicensed," said Calvert.

Signum's 'watermarks' can show when an image has been copied, even if all the colours are changed and content is removed or added. ■

Now you see it ... Similar technology from Signum to the one it is using in the Ordnance Survey watermarking technology is seen above being applied to a speeding biker (allegedly). The left-hand image shows how changes to the bike's registration plate and speed, and removal of the car to the left have been spotted by the software.

This programmable load tester for 12V sealed lead-acid and NiCd batteries covers twenty of the most popular capacities from 0.8 to 7Ah. Designed by Dave Sawford and Ken Duggan, it features an automatic test procedure, it has a load tolerance of better than 2% and it includes a DVM function with 0.3% measurement error.

Verify batteries to 7Ah

The function of backup batteries is to provide power to the system in the event of mains supply failure. In UPS, telecomms and alarm systems, sealed lead-acid batteries are the most popular because they offer good combinations of size, energy density per unit volume and cost.

However, batteries and related charging circuits remain common sources of failure – especially in alarm systems. Charging circuits are statistically more reliable than sealed lead-acid batteries, as correctly applied modern semiconductors typically last at least three times longer than even the most advanced batteries.

Testing such batteries gives multiple benefits, especially in the alarm industry; fewer call-outs for the installer, police, site owners' etc., and fewer claims for the insurers. Testing identifies both weak and healthy batteries, reducing the chances of failure. It also allows you to make the best use of what is both a resource and, at the end of its service life, a source of pollution. Most batteries now being classified as special waste.

There are only two methods of testing sealed lead-acid batteries. Firstly you can measure the battery's internal resistance. By inference, a battery with a high internal resistance cannot deliver power.

Alternatively, you can evaluate a lead-acid battery using load testing. Here, the battery delivers power to a load and the voltage is monitored. Demonstrating that a battery can deliver power for the required time increases confidence that it can perform to an almost identical level in the immediate future.

A testing problem

Taking the example of a 12V, 7Ah battery, such a battery could theoretically be tested by connecting a $12 \div 7 = 1.71\Omega$ resistor, which would allow 7A to flow; the 1C rate of this battery, hence a '1C' load test. If the battery voltage fell below 12V during the test, the battery would be failed.

Any size of sealed lead-acid battery can be tested in this manner, the load resistor value simply being calculated by Ohm's law. However, for each capacity of battery, a different resistor is required. Also, a nominal '12V' battery may have a start-of-test voltage of 13.8V causing, in this example, an initial current flow of $13.8 \div 1.71 = 8.07$A.

Taking in to account the 5-10% tolerance of the power resistor, the result is a worst case figure of 8.97A.

A PIC based solution

The tester described here enables twenty of the most popular capacities of sealed lead-acid battery to be tested by applying an accurate constant current load. Yuasa recommends a minimum test of 1 minute at a 1C rate, and specify that the voltage should not drop below 12V.

The circuit is based on the cheap and versatile Microchip PIC16C74A. This device allows you to build a sophisticated but easy to use test instrument using very few components, thanks to features such as the on-board five-channel eight-bit analogue to digital converter. The block diagram of the tester is shown in **Fig. 1**.

As well as controlling the load, the PIC also directly controls four 7-segment multiplexed displays and three indicator LEDs to show what is being displayed – current, voltage or time. The controller also responds to select, stop and start controls, and monitors the voltage of the battery under test. In addition it checks its own battery and keeps an eye on the load temperature while it is active.

The adjustment links allow a 2.5% increase or decrease in the load current to counteract inaccuracies in the current measuring circuit. A buzzer gives an audible indication of the end of a test, or a problem with the load temperature. It is also used to make a click when a control key is pressed or if a held down key is auto-repeating.

Important safety issues

There are safety concerns that influenced the design of the circuit, and these are due to the nature of the batteries being tested. Although they might be tested at currents of up to 7A, the batteries involved here can deliver many times more.

Uncontrolled current flow could cause arcing, which might damage the test lead clips. In a worst-case scenario, the power dissipated could be a fire hazard.

For these reasons, three considerations were taken into account. Whenever the unit is switched off, the load must be inactive. This applies whether the test lead clips are connected before the unit is turned on, or if the unit is turned off during a test.

At the first sign of a bad connection to the battery under test, the load must be swiftly de-activated.

Accidentally connecting the test leads with the wrong polarity must not cause any damage to either the tester or the battery under test.

Specifications of the battery monitor

Battery types tested	12V sealed lead-acid and NiCd
Battery capacities tested (C)	0.8, 1.0, 1.1, 1.2, 1.8, 2.0, 2.1, 2.3, 2.5, 2.8, 3.0, 3.2, 3.5, 4.0, 4.5, 5.0, 5.7, 6.0, 6.5, 7.0 Ah
Test type	60s partial discharge at 1C
Fail threshold	11.75V
Voltage measurement tolerance	0.3%
Voltmeter range	3-13.8V
Voltmeter resolution	0.054V
Load voltage range	9-13.8V
Load tolerance	Better than 1.7% @ 7A, typically ±5-15mA)
Power consumption, typical	40mA, approx.
Internal battery low level warning	7.6V

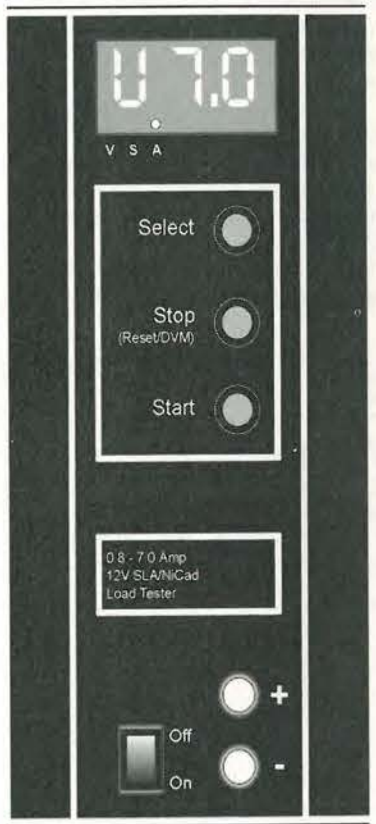
Constant-current loading

The principle of the constant current load used for the tester is shown in **Fig. 2**.

Resistor R_1 is the main load resistor, and Tr_1 actively controls current drawn from the battery. The comparator shown represents the way the PIC interacts with the circuit and is not physically present. Switch S_1 is electronic, controlled by the PIC.

Voltage across R_1, namely V_1, is directly proportional to the current flowing through it. This allows the PIC software to make a comparison with a reference value, corresponding to the desired V_1, hence the desired current load.

By simply closing S_1 if V_1 is below the reference and opening it if V_1 is greater than the reference, the PIC generates a pulse width modulated output through S_1. This output is integrated by the low-pass filter R_2 and C_1 to provide a reasonably smooth analogue voltage



Front panel of the prototype battery tester.

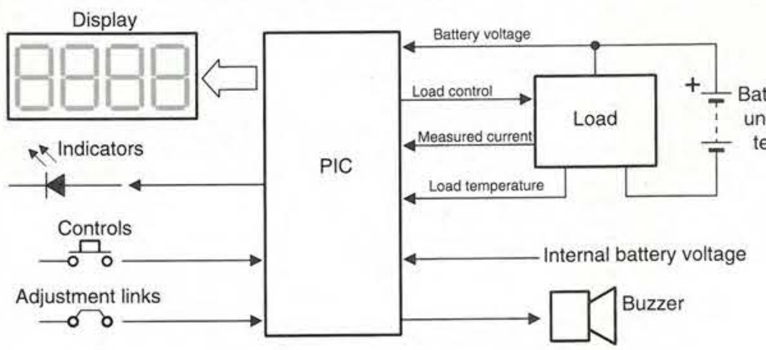


Fig. 1. At the heart of the tester is a PIC microcontroller, which evaluates the battery being tested. Having a five-channel analogue-to-digital converter built in, this controller greatly simplifies the monitor's design.

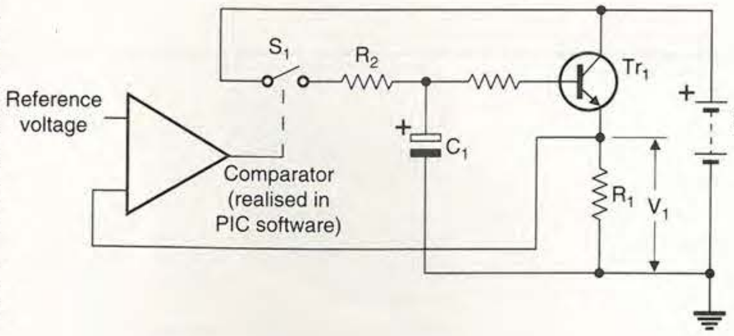


Fig. 2. Principle of the constant-current battery load circuit. The PIC compares voltage over R1 against a reference and switches the transistor on or off depending on the reading, resulting in a pulse-width modulated drive.

to the base of Tr_1. Resistor R_1 provides negative feedback to essentially make Tr_1 a unity voltage gain buffer. For example, if R_1 has a value of 1Ω, a current range of 1 to 7A will mean a V_1 of 1 to 7V, requiring $1+V_{BE}$ to $7+V_{BE}$ volts at the base of Tr_1.

Circuit description

Circuit details for the tester are shown in **Fig. 3**. In a practical circuit, a number of extra components needed over the ones shown in Fig. 2.

Although the load transistor can be thought of as a unity-gain voltage buffer, it has to have a huge current gain. To achieve this, a Darlington pair configuration is used with Tr_{10} providing the power handling capabilities and Tr_9 providing most of the gain.

Using the ZTX690B and the 2N3715 in this arrangement gives a total h_{FE} of at least 14000. Alternatively, a power MOSFET could be used in place of Tr_9 and Tr_{10}.

Comparing the actual circuit to Fig. 2, transistors $Tr_{7,8}$ form the switch for the PWM signal from the PIC, with R_{23} and C_4 performing the filtering.

Output from a PIC can easily drive a MOSFET directly, but as Tr_8 is a p-channel device, the result would be a load that was activated by a low output from the PIC. If the test battery were connected before the unit was powered up, Tr_8 would turn hard on, so the load would be turned on too. Using Tr_7 as an inverter to make it active high makes sure that this does not happen.

Diodes D_2 and D_8 protect against current flow through the load transistors in the event of a reverse voltage being applied to the test leads. Diodes $D_{3\text{-}6}$ make sure that the voltage applied to the PIC's analogue input pins never goes more than 0.4V above V_{DD} or 0.4V below V_{SS}. The PIC data sheet specifies that the voltage on any pins should not exceed 0.6V above V_{DD} or 0.6V below V_{SS}.

With this circuit, there is an added problem that a pre-power on voltage on one of the pins can work its way through the PIC's internal circuits to the output pins. This is not desirable as RE0 turns on the load. Transistors $Tr_{6,5}$ stop this happening by disabling the input to RA1 until the program enables it by putting RE1 high. RA0 is already protected from a pre-power on voltage because the initial state of Tr_{10} is to be off.

Diode D_1 is used to monitor the temperature of the load components. If the load is sinking a current of 7A at 13.8V, the heat dissipated by it will be almost 100W, so it is essential that adequate heat sinking is provided for the components that do the dissipation.

Most of the heat is dissipated by R_{30}, R_{31}, Tr_{10} and D_8. We recommend that they are mounted together on a separate heat sink with a good thermal connection to D_1 to allow the PIC to accurately monitor their temperature.

An LM4040 AIM-4.1 voltage reference is used by the PIC's a-to-d converter, which produces a voltage of 4.096V with 0.2% accuracy. This IC comes in a range of tolerances and packages but as all measurements made by the PIC are referenced to this voltage, using one with a different tolerance will affect the accuracy of all aspects of the tester.

All four 7-segment displays are common cathode types. This allows them to be multiplexed, with the output from the PIC sourcing the current to drive them. The PIC can source up to 20mA, but we found that a drive current of 9mA per segment was sufficient with the type of displays used here.

The buzzer used is a piezo-ceramic type that needs only 10mA to provide a surprising amount of noise!

List 1. A-to-D conversion routine.

```
doatod
        bcf INTCON, GIE       ;Disable all interrupts
        movwf ADCON0          ;Set up the port
        movlw 0x04            ;Delay for 24µS (4 loops)
        movwf temp
dalp1
        nop
        nop
        decfsz temp, F        ;Keep looping until temp=0
        goto dalp1
        bsf ADCON0, GO        ;Start the converting
dalp2
        btfsc ADCON0, GO      ;Check to see if it is done
        goto dalp2            ;Nope. Loop
        movf ADRES, W         ;Store the result in W
        bsf INTCON, GIE       ;Enable interrupts again
        return
```

The microcontroller

The PIC16C74A has 33 i/o pins, arranged as five ports. Any of the pins on a port can be configured as an input or output. Many can also be configured for a special function such as external interrupts (RB6, RB7) or asynchronous communication (RC6, RC7). In this circuit though, they are only used as straightforward i/o pins.

Data sheets for the PIC, available on the Microchip web site (www.microchip.com), give the full details but a brief summary of the ports and how they are used in this design is given in **Table 1**.

Pins not shown in the table are concerned with the workings of the PIC

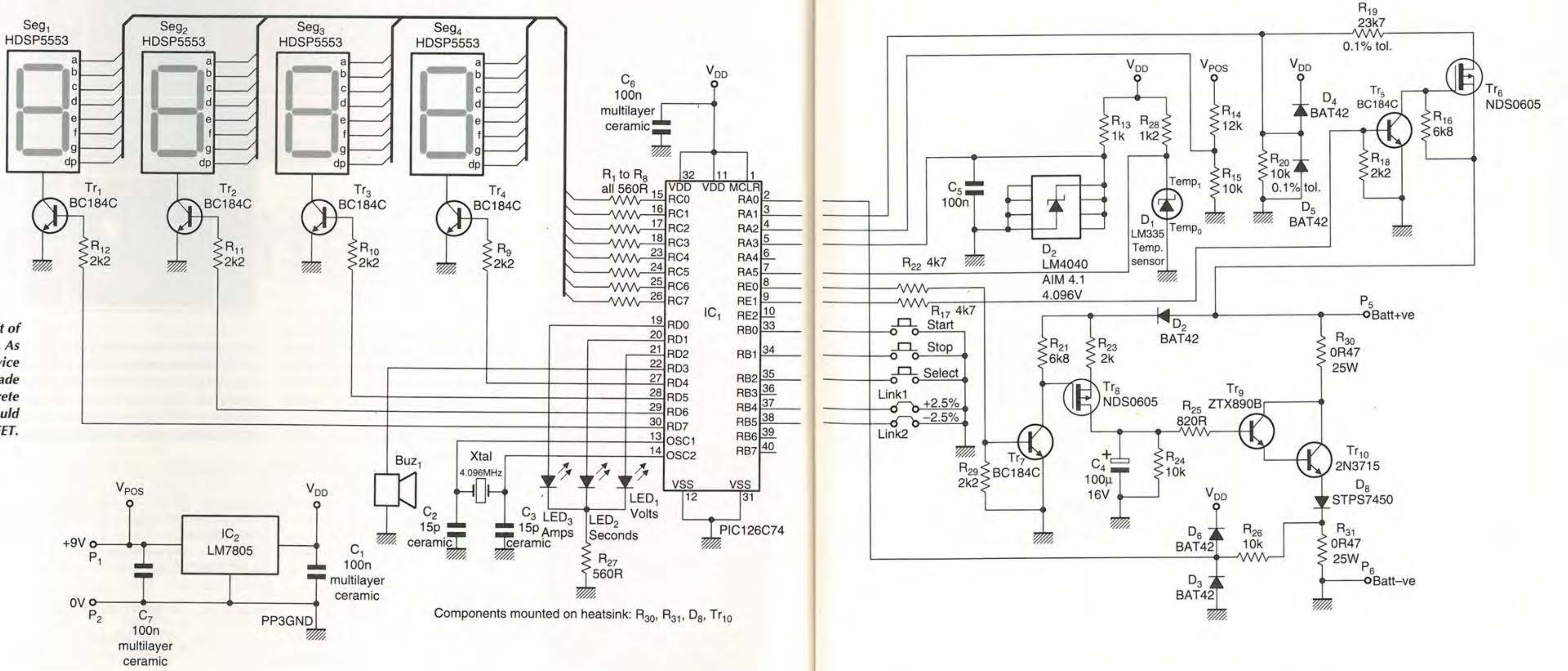


Fig. 3. Full circuit of the battery monitor. As shown, the load device is a Darlington made up of two discrete devices, but you could use a MOSFET.

Table 1. The i/o pins of the PIC controller are designated as follows.

Pin	Name	Description
	Port A	*All used pins configured as analogue inputs*
2	RA0	Load current. Voltage is 0.47 x the load current.
3	RA1	Voltage of the battery under test.
4	RA2	Internal battery voltage monitor.
5	RA3	The voltage reference for the PIC's a-to-d module.
6	RA4	Not used
7	RA5	Load temperature monitor.
	Port B	*All used pins configured as digital inputs with pull-up resistor enabled*
33	RB0	'Start' key
34	RB1	'Stop / Reset / DVM' key
35	RB2	'Select' key
36	RB3	Not used
37	RB4	+2.5% Load current adjustment
38	RB5	–2.5% Load current adjustment
39	RB6	Not used
40	RB7	Not used
	Port C	*All pins configured as TTL outputs*
15	RC0	Display segment 'a'
16	RC1	Display segment 'b'
17	RC2	Display segment 'c'
18	RC3	Display segment 'd'
23	RC4	Display segment 'e'
24	RC5	Display segment 'f'
25	RC6	Display segment 'g'
26	RC7	Display segment decimal point
	Port D	*All pins configured as TTL outputs*
19	RD0	'Current' indicator LED
20	RD1	'Time' indicator LED
21	RD2	'Voltage' indicator LED
22	RD3	Buzzer
27	RD4	'Units' LED display
28	RD5	'Tens' LED display
29	RD6	'Hundreds' LED display
30	RD7	'Thousands' LED display
	Port E	*All pins configured as TTL outputs*
8	RE0	Load PWM output
9	RE1	Input enable
10	RE2	Not used

itself. The system clock is a 4.096MHz crystal, chosen so the processor executes 1024 instructions a millisecond. This means that timings can be calculated reasonably easily using binary arithmetic. The MCLR pin is the reset pin and should be kept high.

The software

The software can be broken down into four main areas.

The 'pre-test' routine. Referring to **Fig. 4**, this routine is where the interrupts and ports on the PIC are set up, variables initialised and the program goes into a loop where it waits for the user to press a key. Pressing the 'Select' key changes the test current, the 'Stop' key toggles to the 'DVM' mode where the voltage measured is continuously displayed and updated every quarter of a second, and the 'Start' key starts the test. The internal battery is also tested at this point.

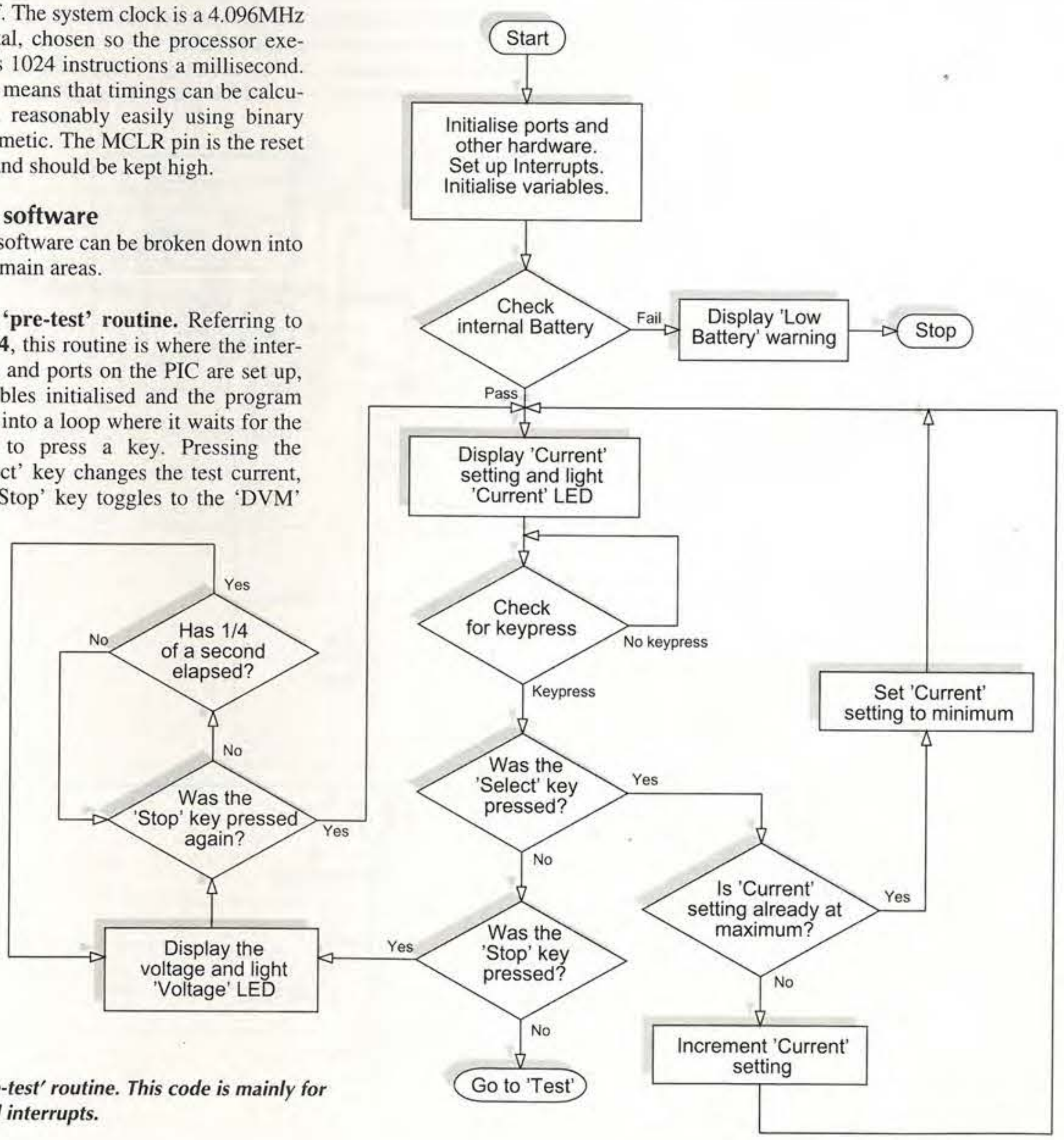


Fig. 4. Flow chart for the 'pre-test' routine. This code is mainly for setting up the PIC's ports and interrupts.

Table 2. ISR timing functions. The main program can use the bits of the interrupt register for its own timing requirements. 'Astable' bits toggle between 0 and 1 while 'monostable' bits need resetting by the main program routines.

Bit	7	6	5	4	3	2	1	0
Function	S	a	b	c	R1	R2	R3	R4

s: 1 second astable
a: 0.25 second astable
b: 0.25 second monostable 1
c: 0.25 second monostable 2
R1-R4: 0.25s bit rotation in the sequence R1 → R2 → R3 → R4 → R1...

The 'test' routine. This is the actual test, where the current specified by the user in the 'pre test' routine is drawn from the battery, **Fig. 5**. While this is happening the voltage, load temperature, time and keys are monitored. The user can change the display between 'battery save' mode where the only thing displayed is a moving dot (resulting in the internal battery drain of about 15 mA), display the elapsed time, the voltage of the battery under test (updated every quarter of a second), and the load current (set in the 'pre-test' routine) by pressing the 'Select' key.

The 'post-test' routine. The result of the test – 'PASS' or 'FAIL' – is displayed and the user can recall the time the test took for the battery to pass or fail, the final voltage measured before the battery passed or failed, and the load current, **Fig. 6**.

The interrupt-service routine. This routine, **Fig. 7**, provides time-critical routines. These allow other routines to operate without worrying how they will affect factors such as the load PWM signal or the overall timing of the test.

A 'clock' is provided by the interrupt-service routine. Routines in the main program can access this clock through a register, individual bits of that register making 0.25s and 1s monostables, a 0.25s astable and four bits that have a '1' rotated through them every second, **Table 2**.

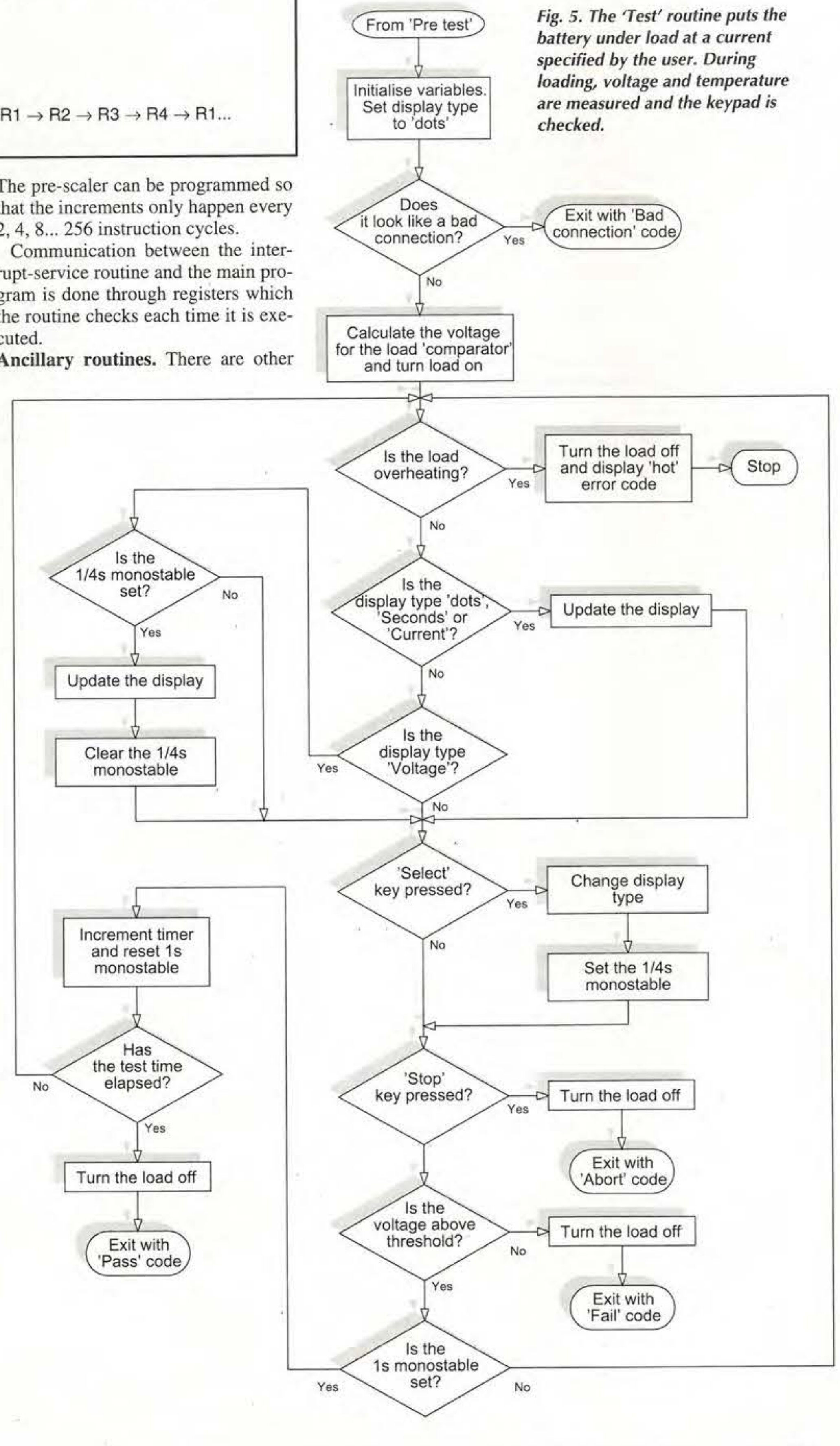
The seven-segment displays are also driven by the interrupt-service routine. It switches between them 62 times a second, outputting the correct digit to each of the four displays in turn. The interrupt-service routine is also responsible for controlling the load.

Interrupts come from the PIC's internal TMR0 timer. This increments a register every instruction cycle (without pre-scaler). It produces an interrupt when it overflows from FF_{16} to 00_{16}. The pre-scaler can be programmed so that the increments only happen every 2, 4, 8... 256 instruction cycles.

Communication between the interrupt-service routine and the main program is done through registers which the routine checks each time it is executed.

Ancillary routines. There are other

Fig. 5. The 'Test' routine puts the battery under load at a current specified by the user. During loading, voltage and temperature are measured and the keypad is checked.



common routines used by different parts of the main program. These perform tasks such as doing the a-to-d conversion, **List 1**, and reading the keys, **Fig. 8**.

The complete code for the tester is too large to reproduce here, but is available on the internet from

`http://www.the-shed.demon.co.uk/tester`

or by post as a hard copy or on floppy disk. Details of how to obtain technical support are given at the end of the article.

At just over 1500 lines of assembly code, the program is fairly large. When writing programs of this size there can be problems which do not occur in smaller programs.

Controller usage notes

Stack depth. The stack of the PIC 16C74A is only eight levels 'deep'. Each time a 'CALL' instruction is executed one of these levels is used up, meaning that calls can only be nested seven times before things start going wrong.

Bear in mind that an interrupt uses the stack to store its return address. Also, if the interrupt-service routine performs an a-to-d conversion another stack level is used when the a-to-d subroutine is called.

These factors reduce the maximum number of nested calls the main program can use to five. It is very easy to accidentally go beyond this without noticing.

Page boundaries. Although the PIC 16C74A has a 13-bit wide address bus, only the lower 8 bits can be addressed when using computed look-up tables via the 'ADDWF PCL, F' command. This command adds the contents of the W register to the program counter.

The other 5 bits are loaded in from the PCLATH register when the 'ADDWF PCL, F' command is executed. For look-up tables, this effectively divides the memory into 256-byte pages, which they must not cross.

Probably the easiest way to avoid crossing these boundaries is to put all the look-up tables at the beginning of the code, before the main program. This makes it easier to check that no boundaries are crossed.

For look-up tables that are after the first 256 bytes of program memory, the PCLATH register must be used to set the upper 5 bytes of the program counter.

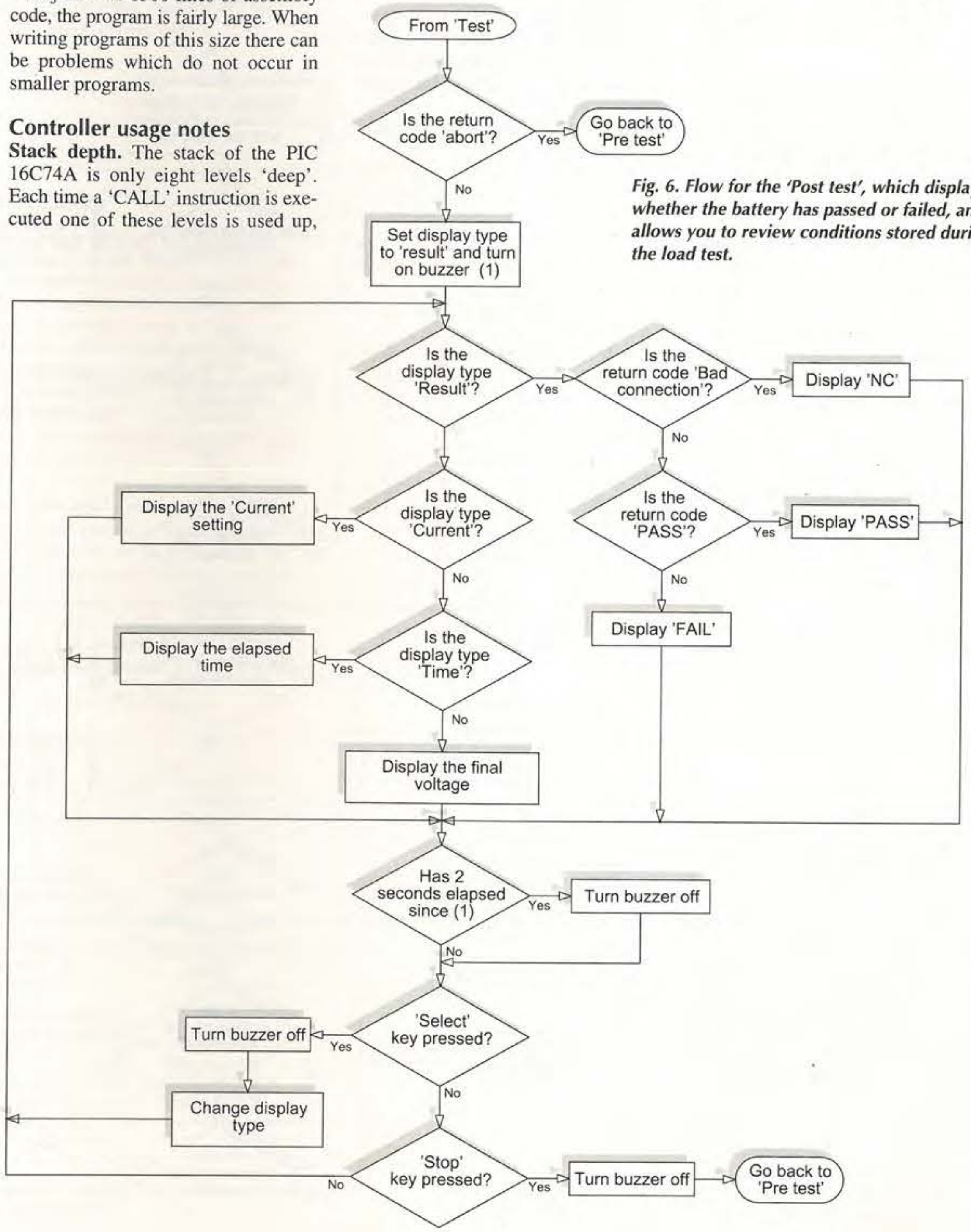


Fig. 6. Flow for the 'Post test', which displays whether the battery has passed or failed, and allows you to review conditions stored during the load test.

More page boundaries. A 'GOTO' or 'CALL' in the program only loads the program counter with the lower 11 bits of the address, meaning that calls only have a 1024-byte range. The other two bits needed to complete the 13-bit address are loaded from bits 3 and 4 of the PCLATH register.

If your program is over 1024 bytes long and uses calls to routines that are beyond the 1024/2046/3069 boundaries you must set the PCLATH 4:3 bits accordingly. The Microchip application note AN556 – available from the Microchip web site – has more details about these problems.

Table 3. The ADCON0 register is used to manage a-to-d conversion.

Bit	7	6	5	4	3	2	1	0
Function	ADCS1	ADCS0	CHS2	CHS1	CHS0	GO/DONE	–	ADON

Mnemonic	Description
ADCS1, ADCS0	A-to-d clock select
CHS2, CHS1, CHS0	Analogue channel select
GO/DONE	A-to-d conversion status bit. Setting this bit starts the A-to-d conversion. On completion this bit is cleared by the a-to-d module.
–	Not used
ADON	Setting this bit turns on the a-to-d conversion module

The analogue to digital converter

One of the most useful features of the 16C74A is the on-board a-to-d converter. Although it only has a resolution of 8 bits, used in conjunction with a precision voltage reference it allows accurate monitoring of many different parts of the circuit.

The reference voltage for the a-to-d – i.e. the voltage that a digital value of 255 from the a-to-d converter represents – can either be set up in software as the V_{SS} supply voltage to the PIC or an external source.

As the tolerance of most 5V regulators is only 5%, an external 4.096V, 0.2% tolerance voltage reference, namely D_2, is used in this circuit for improved accuracy in voltage measurements.

To convert a voltage to its corresponding digital value the following can be used,

$$D=\frac{V_{sample}}{V_{ref}}\times\left(2^n-1\right)$$

where D is the digital value, V_{sample} is the sampled voltage and n is the number of bits of resolution of the a-to-d converter.

For the PIC 16C74A in this application,

$$D=\frac{V_{sample}}{4.096}\times 255$$

This circuit uses four of the PIC's a-to-d channels to monitor the following parameters:

RA0: Current flow through the load. For example, at 7A the voltage at RA0 is,

$$7A\times 0.47\Omega=3.29V$$

So the value returned by the a-to-d conversion will be,

$$\frac{3.29}{4.096}\times 255=204.8$$

Note that as the PIC can only use integers, the actual value returned would be 204.

RA1: The voltage of the battery under

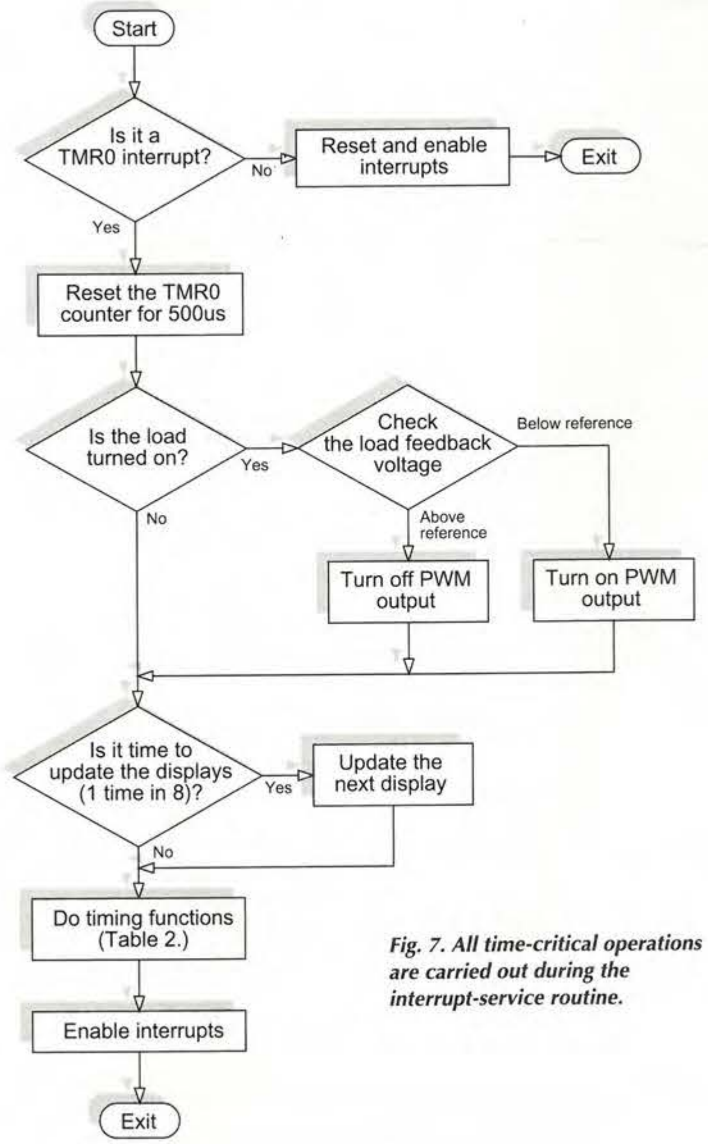


Fig. 7. All time-critical operations are carried out during the interrupt-service routine.

test. To provide a measuring voltage range of up to 13.8V, R_{19} and R_{20} (precision 0.1% tolerance components) are used as a potential divider,

$$13.8\text{V}\times\frac{10\text{k}\Omega}{23.7\text{k}\Omega+10\text{k}\Omega}=4.095\text{V}$$

at RA1.

RA2: The voltage of the tester's own battery. Resistors R_{14} and R_{15} are used to give a measured voltage range of up to 9V.

RA5: Drive-transistor temperature. The LM335 used here provides a voltage output of 10mV multiplied by its temperature in kelvin, so for a temperature of 55°C there would be a voltage of,

$$(273+55)\times 0.01=3.28\text{V}$$

present at RA5. Line RA3 is the input for the external voltage reference.

Code for the a-to-d conversion routine used by this application is shown in **List 1**. On entry to this routine, the W register contains the details needed by the converter. These details specify which port to read and which conversion clock to use – either a division of the clock frequency or an internal *RC* oscillator. They also specify whether the a-to-d module should be turned on. If a conversion is about to be initiated, it must be turned on.

Information from the W register is loaded into the ADCON0 register, **Table 3**. On exit from this routine the value returned by the a-to-d conversion – from the ADRES register – is copied into the W register.

Flags ADCS1 and ADCS0 select which clock is to be used by the a-to-d module. This clock may be the main clock frequency, F_{OSC}, divided by 2, 8 or 32, or an internal *RC* oscillator.

Using F_{OSC} means that the time taken by the successive approximation a-to-d module can be calculated and minimised. But it also means that changing the clock speed necessitates changing the software to bring the a-to-d clock into the required range of values.

Although the *RC* clock option means we cannot predict the time that an a-to-d conversion will take, this does not matter so much if timing functions are performed by an interrupt-service register.

After the a-to-d unit has been set up, a 24µs delay is needed. This delay makes sure that the voltage on the pin to be read has stabilised and that the capacitor in the sample-and-hold circuit of the a-to-d unit has been fully charged.

The lines following the delay subroutine start the conversion and then loop round until the conversion status bit is cleared, indicating that the conversion is complete.

As the program uses a-to-d conversion routines within interrupts, it is vital that an interrupt does not occur while this routine is being executed.

For example, an interrupt routine that used the a-to-d converter while the conversion program was acquiring a reading would over-write the value being read. Similarly, an interrupt during the set-up and delay sections could change which channel is read by the code that follows. For these reasons, interrupts are disabled at the start of the conversion routine and re-enabled at the end.

Technical support

A complete set of instructions on how to use the tester is available from http://www.the-shed.demon.co.uk/tester, where the complete code is also available. The instructions are also distributed with hard copies of the code, pre-programmed PICs and kits of parts.

To obtain details about how to get the code sent by mail, buy a pre-programmed PIC or a complete kit of parts (excluding case and heat sink), send a stamped self-addressed envelope to:

David Sawford,
177 Sturton Street,
Cambridge CB1 2QH.

Or email: david.sawford@the-shed.demon.co.uk

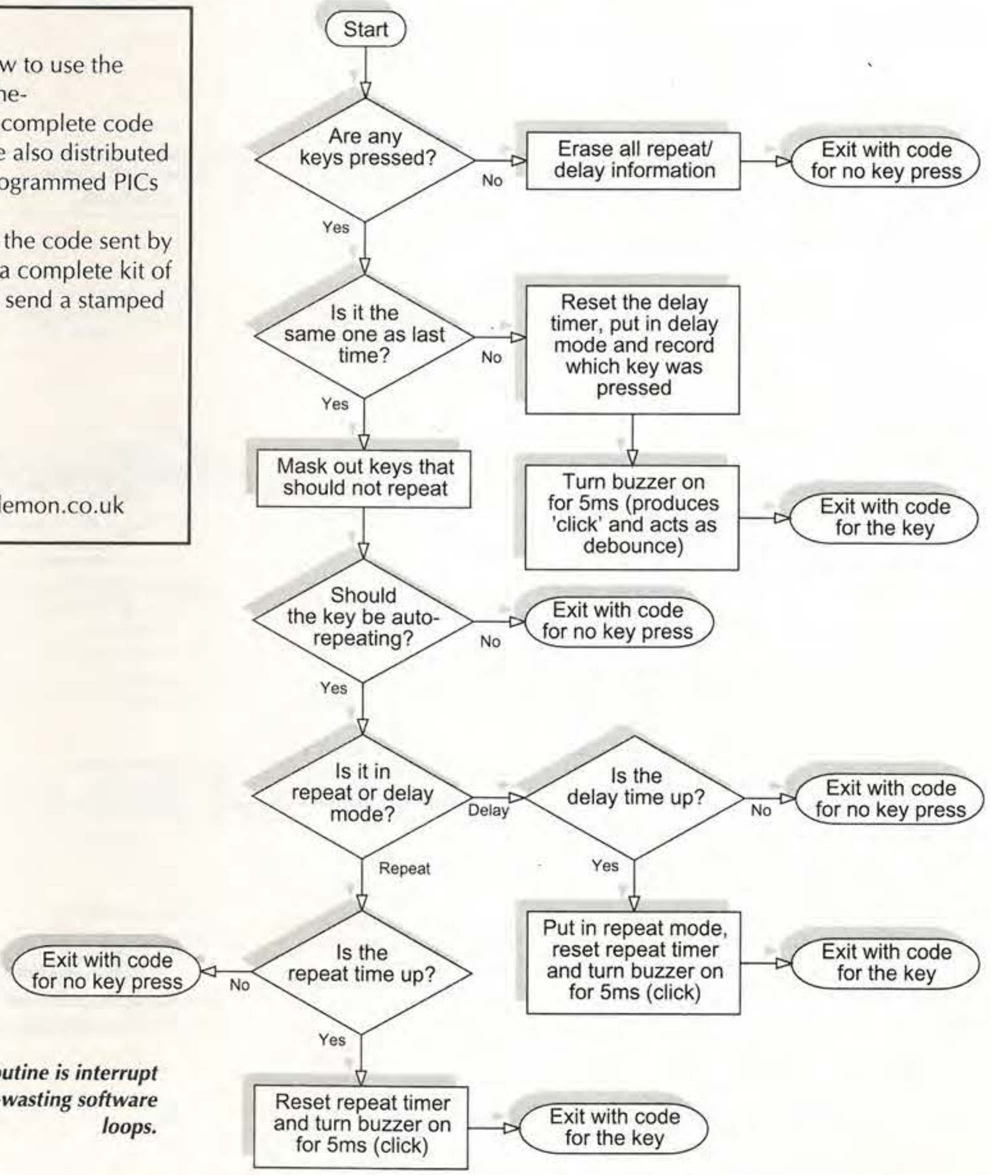


Fig. 8. The keypad interface routine is interrupt based so there are no time-wasting software loops.

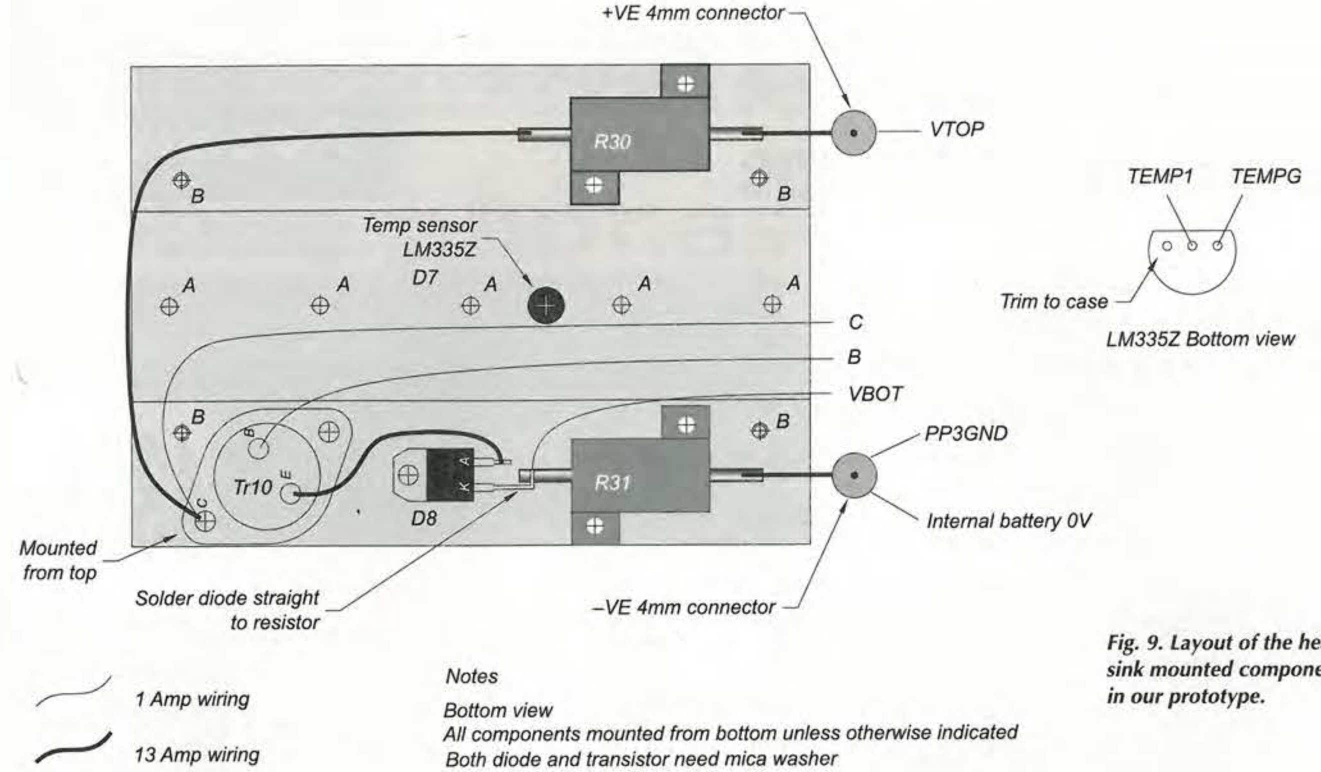


Fig. 9. Layout of the heat-sink mounted components in our prototype.

The keypad interface

In order to create a tactile 'feel' to the tester, the keypad reading routine does a bit more than just report which keys are being pressed.

Although the physical keys are no more than connections from the pins on the PIC to 0V, using software makes it possible to implement debouncing, auto-repeat of held-down keys with adjustable delay and repeat speed and an audible click when a key is pressed. Audible clicks are also produced while a held-down key is auto-repeating.

Figure 8 is the flowchart for the keypad interface routine. Timing functions, such as incrementing the counters for the delay/repeat of keys that are held down, are performed by the interrupt-service routine.

Using interrupts means that the routine does not have to have any loops in it to detect key statuses. As a result, a minimum of time is spent in the routine, allowing the calling routine to attend to more time-critical matters such as monitoring voltages and updating the display.

Implementing the tester

Our prototype battery tester was constructed in an extruded aluminium case, with the heat-producing components mounted on a piece of sheet aluminium, **Fig. 9**. They were bolted to the case using the holes marked 'A' via an aluminium block, shown in dotted lines, to form an effective heat sink.

Because there was a direct thermal path from the heat sink to the case, we chose 55°C as the temperature at which to suspend the test with an error message. The components on the heat sink can stand higher temperatures, but the operator holding the case might not be able to!

In this configuration, we found that three successive 7A tests could be performed before the unit started giving the 'too hot' error message. If a different build technique was used to allow a higher running temperature, the maximum test temperature could be raised by changing a value in the PIC code.

We stuck the temperature sensor into a hole in the heat-sink block using high-temperature epoxy to give a good thermal contact. All other components were mounted on a PCB that was bolted via the 'B' holes to the heat sink with insulating pillars.

There's a rectangular slot in the front of the case for the display. The keys, 4mm test-lead sockets and on/off switch were also mounted on the front. The battery compartment was in the base of the unit.

The choice of test leads is an important factor when it comes to accuracy. Not only must they be rated at 7A or more, but they must also have a known resistance.

Our tester has leads with a resistance of 0.054Ω/m, which made for a total resistance of 0.027Ω with 250cm lengths for both positive and negative connections.

In the code there is a look-up table of the voltage drops across the leads for each test current. This is subtracted from the 'fail' voltage digital equivalent by the software. For different leads, a different set of values will need to be calculated using,

$$value=\left(\frac{R\times I\times 10\text{k}}{10\text{k}+23.7\text{k}}\div 4.096\right)\times 255$$

Good quality test clips are essential to minimise the voltage drop across them, and hence the voltage measurement error.

Setting up

The only adjustment that needs to be carried out is via the two links that add or subtract 2.5% from the load current. This is due to the fact that power resistors such as those used for the current sensor, R_{31}, typically have a tolerance of 5%.

Cutting link 1 adds 2.5% via a look-up table to the reference voltage that the pulse-width-modulated load signal generating routine will use for each current setting. Cutting link 2 subtracts 2.5%, and if both are cut there will be no net change.

Unless you have a way to accurately measure the resistance of R_{31} directly, the easiest way is to use a current of, say 2A from a power supply through the resistor and measure the voltage across it. As the resistance is voltage divided by current, it is easy to calculate R_{31} and cut the appropriate link if need be.

On a final note, it should be easy to apply the loading method to other current ranges and voltages. ■

Third-generation mobile phones

The consumer will probably not get the benefits of 3G mobile phones for another two or three years, but what exactly will be on offer when the services are rolled out? Richard Wilson takes a look

As you can see from the table on the right, the UK, like the rest of Europe, is about to get its first third-generation mobile phone operators. But if like me you're wondering what third-generation mobile communications will offer in sexy new services then we should start to find out before the end of the year.

By Christmas 2000 each of the new third-generation (3G) operators will only be starting to build their network infrastructure. True 3G services are probably two years away at the earliest.

Some even suggest 3G will not happen in a significant way until 2002/3. One reason for this is that mobile-phone users will not have to wait until 3G is ready to get some of the anticipated new services like wireless Web browsers and video.

If GSM is second generation and UMTS is third generation mobile communications, then somewhere in between comes the 2.5 generation system known as general packet radio service (GPRS). This is more than a half-way house mobile technology to full 3G. In fact its capability to carry packet-switched data at aggregated rates up to 384kbit/s may even persuade operators to put off investment in full-blown and expensive 3G networks a little longer.

Evidence suggests that UK operators are racing to upgrade their GSM networks to support GPRS and we could see first commercial general packet radio services by the end of the year.

"There is a strong desire for the rollout of GPRS," says Lance Hiley, a strategic marketing manager for wireless at Lucent Technologies Microelectronics. "Every operator will have commercial GPRS services in 2000 – most likely the second half of the year."

But how does GPRS improve on existing GSM mobile systems? The key is data transmission.

The GSM mobile phone is essentially a voice communicator. GSM data transfers are literally in the dark ages being limited to 9600bit/s. This is less than 20 per cent of the rate of the modem in your PC, or 1000 times slower than the slowest office LAN.

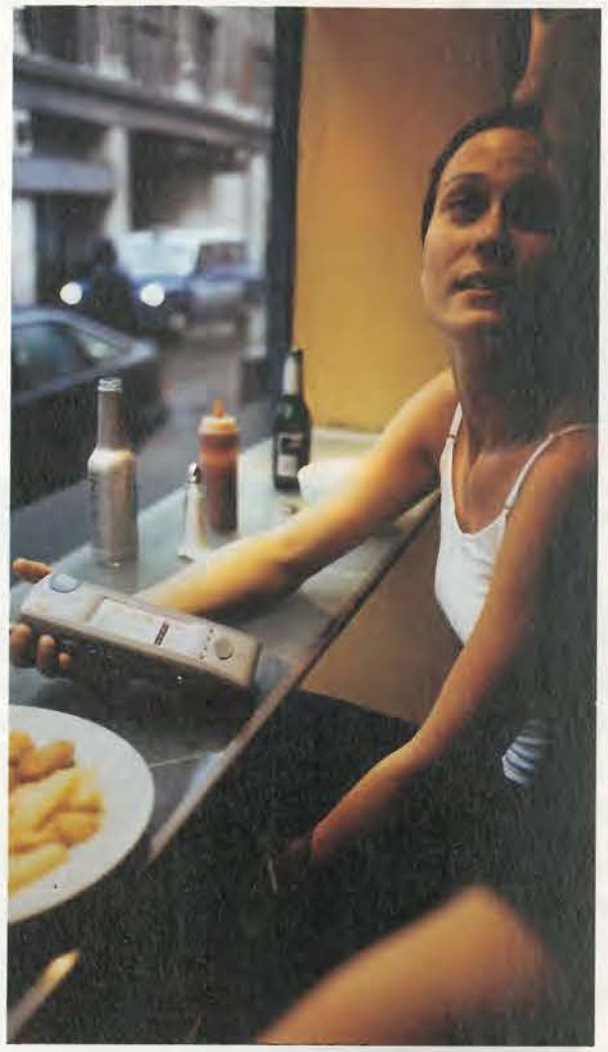
The experts tell us that the 3G mobile phone will be defined as a

3G licences in Europe

Country	Number of licences being awarded	Scheduled award date
Finland	4	awarded March 1999
UK	5	March 2000
Spain	4	March 2000
Netherlands	4-6	Q1 2000
Germany	4-6	Q2 2000
Norway	3-4	Q2 2000
Switzerland	3-4	?
Portugal	4	?
Denmark	4-5	?
France	4	Q3 2000
Belgium	3-4	Q3 2000
Sweden	?	Q3 2000
Italy	5	Q3 2000
Austria	4	Q4 2000
Ireland	3-4	Q4 2000
Greece	4	Q1 2002

Source: Netcom Consultants, UK

Mobile video. Samsung shows what a 3G handset will look like.



Mobile phone operator Orange has already shown us what a 3G video phone could look like

mobile data terminal, with voice communications thrown in, perhaps for free.

"If 3G is only deployed for voice it will be like killing a bee with an F-14 fighter plane," asserts Thierry Laurent from Philips Semiconductors.

Never mind the technology, this will change the nature of mobile communications and present the mobile firms with a new concept to sell. "We must remember that on GSM mobile 99 per cent of the traffic is voice," says Stephen Wall, technical director of The Smith Group. "That is a straightforward sell for the operator; you sell two phones and you talk."

For this switch away from voice to happen, the radio network must be able to handle high-speed data channels. UMTS, the technology behind 3G, will do this. But GPRS will also be able to extend the data rates on GSM networks.

GPRS works by transmitting data in packets. Each packet has an 'address' for where it is being sent so that all the data packets that make up a particular message do not necessarily have to be transmitted in sequence on one radio channel, as is the case with GSM at present.

With GPRS, data transmission can be increased above the capability of a single radio channel because packets can be sent on any of a number of different radio channels set up by the network.

A GPRS system allows a data rate of up to 14.4kbit/s to be transmitted in each of GSM's eight time slots. This gives an aggregated system data rate of 115.2kbit/s. A ten-fold increase on today's GSM modems.

However we as users are only starting out on the 3G road, despite how quickly manufacturers may wish to push us along. A full 115kbit/s GPRS mobile data terminal requires eight radio slots, for both transmit and receive. That requires a minimum of 16 slots and is known as a Class 18 GPRS system.

Class 18 mobiles offering 115kbit/s data rates will not be available until next year. The first GPRS mobile we will see on the market this summer will be Class 8. This class uses four 14.4kbit/s channels so should offer the radio equivalent of a 56k modem.

So if you want to use your mobile to transmit data files to the office, or even to download information from the Internet, the data rates you can expect are as follows – 56kbit/s by the end of the year – 115kbit/s by the summer of 2001. And it does not stop there.

A further GSM data enhancement technology known as EDGE is being developed by Nokia and Ericsson amongst others. This could potentially extend mobile data rates to 384kbit/s. But it will not happen until late 2001 or early 2002.

And that should be just in time to get the market ready for the first full 3G services in 2002/3. The data capabilities of 3G will stretch up to 2Mbit/s by which time the mobile phone will be seen as a mobile data terminal and not simply a voice communicator. ■

WAP

Wireless Web browsing will complicate considerably the business of selling mobile communications so operators and manufacturers have come up with the wireless applications protocol (WAP) standard to help companies through the minefield of the wireless Internet.

WAP defines a set of hardware and software interfaces which when implemented on mobile handsets and Web site servers should allow mobile phones to be used much like a PC, to send and receive e-mail messages and even to browse the Internet.

But the performance of the WAP applications will inevitably put demands on the bandwidth of the mobile's radio channel. So the development of WAP-based services will coincide with the rollout later this year of higher bandwidth radio systems like GPRS (general packet radio service) which theoretically offers data rates up to 115kbit/s.

WAP itself incorporates data compression to make the most of the lowly data transfer rate of today's GSM networks. It also uses error correction techniques to ensure that browsing and data transmissions are not severely affected by changes in the quality of the radio link.

The other important feature of the WAP is it is designed to run applications on different radio network architectures, such as GSM, CDMA, DECT and even TETRA.

The telecoms industry is notorious for its use of acronyms, so when third generation mobile systems became known as 3G systems hardly an eyebrow was raised in disbelief. However, 3G is just the start. Already there are 2G+, 3GPP and 3GPP2 to tax our powers of understanding. To help the inquisitive and perhaps confused amongst you Steve Bush presents a listing of some of the more useful acronyms associated with the next generation of mobile phone technology. Perhaps it will give you a chance to impress someone in the office but don't bank on using the knowledge at the local pub quiz night.

2G+ Second generation mobile communications with enhancements to add high speed data capability. See EDGE, HSCSD and GPRS.

3G Third generation. 3G is used as a collective noun for the forthcoming generation of mobile phone systems. These are characterised by high speed data capability alongside voice services.

3GPP Third Generation Partnership Project. An organisation producing technical specifications for third generation mobile system based on GSM core networks. ARIB (Japan), CWTS (China) , ETSI (Europe), T1 (USA) , TTA (Korea) and TTC (Japan) are members. According to 3GPP, the 3GPP standards will be forwarded to the ITU and will support inter-working between IMT-2000 family members. The project is expected to accelerate IMT-2000 standardisation. www.3gpp.org

3GPP2 CDMA2000 equivalent to 3GPP, members are ARIB (Japan), CWTS (China) , TIA , TTA (Korea) and TTC (Japan). www.3gpp2.org

ARIB Association of Radio Industries and Businesses. Japanese radio standards body. ARIB and the European UMTS specifications were developed to be interoperable at an early stage. www.arib.or.jp

CDMA Code division multiple access. A spread spectrum communication technique, the basis of most 3G systems.

CDMA2000 3G air interface standard, a Qualcomm trade name. Comes in 1X (1.25MHz channel) and 3X (3 x 1.25MHz channel) types.

CWTS Chinese Wireless Telecommunications Standards organisation. www.cwts.org

EDGE Enhanced Data-rates for GSM Evolution. GSM with a different modulation scheme boosting data throughput up to 384kbit/s using existing GSM infrastructure. Was known as 2G+ in Europe, now officially part of 3G and known as IMT-SC.

EFR Enhanced Full Rate. A speech codec for better speech quality codec, also said to be more tolerant to interference.

ETSI European Telecommunications Standards Institute.

FPLMTS Future Public Land Mobile Telecommunication System, the ITU's single global 3G standard, abandoned when no agreement could be found, replaced by the IMT-2000 family of standards.

FDD frequency division duplex, part of 3G W-CDMA with separate Tx and Rx frequencies for up to 2Mbit/s short range data services. Will be implemented after TDD.

GCF GSM Certification Forum. An industry body defining a set of tests to supplement European self-certification type-approval tests to ensure functional compatibility to GSM networks. A 3G equivalent of this is expected to be formed.

GPRS General Packet Radio Services. Part of the GSM Phase 2+ extension. Packet switching on GSM with data rates of up to 115kbit/s carrying Internet Protocol, TCP/IP and X.25 Generalised Packet Radio Service.

GSM Global System for Mobile Communications, formerly Group Systeme Mobile

GSM Phase 2+ GSM with high speed data capability.

H.323 Real-time audio and video and data over packet switched network - multimedia over IP.

HSCSD High-speed circuit switched data, data over GSM up to 57.6kbit/s by concatenating up to four consecutive GSM time slots. Part of GSM Phase 2 development.

IMT-2000 International Mobile Telecommunications for the year 2000. The ITU's family of 3G mobile phone systems. Replaced ITU's original concept of one single 3G system called FPLMTS which was abandoned when none of the vested interests could have it their own way. See: IMT-DS, IMT-MC, IMT-TC, IMT-SC and IMT- FT

IMT-DS Frequency division duplex version of UMTS.

IMT-MC CDMA2000, narrow-band CDMA for the USA. 3X CDMA 2000 using three 1.25MHZ channels.

IMT-TC Time division duplex version of UMTS and time division SCDMA (TD-SCDMA).

IMT-SC The successor to IS-136 mobile standard. This is EDGE [the GSM phase 2+ standard].

IMT-FT The DECT cordless standard.

IP Internet Protocol

IPv6 Forum An organisation promoting Internet access on 3G mobile networks.

IS-95 Second generation narrow-band CDMS system, used predominantly in the USA.

IS-136 Enhanced IS-95

IS-2000 CDMA2000 air interface standard. A successor to standard TIA/EIA-95-B

ITU International Telecommunication Union. Officially the final arbiter of telecommunications standards, but its power lies only in the agreement of its members. Has finally approved the radio interfaces for IMT-2000 third generation mobile phones. www.itu.int

NTT DoCoMo Japanese company, similar to BT in the UK, implementing a 3G system in Japan, due in April 2001, it will be the first third generation system.

Packet switching Communication using a channel shared by multiple users, silent periods between a sender and receiver are filled with data from other user pairs. Very efficient use of system resources leads to more data through a given network, but it is difficult to guarantee the arrival time of time-critical data (voice, music or video) during busy periods. Highly suitable for time-tolerant data like Internet pages.

PDH Pleisynchronous Digital Hierarchy, here only because it includes the word pleisynchronous.

T1 Industry body that creates network interconnection and interoperability standards for the United States. www.t1.org

TAA Korean telecommunications standards body. www.tta.or.kr

TCC Japanese telecommunications standards body. www.ttc.or.jp

TCP/IP Transmission Control Protocol/Internet Protocol. Protocol necessary for Internet data transmissions.

TDD time division duplex, part of 3G W-CDMA. Transmission and reception are at the same frequency, at different times. Data rates will be lower than FDD, but TDD has a longer range. TDD is planned to be operating before FDD.

TDMA Time division multiple access.

TD-SCDMA Time-Division Synchronous CDMA

TIA The Telecommunications Industry Association, a US standards development organisation. www.tiaonline.org

TIA/EIA-95-B Third generation CDMA air interface standard. Replaces IS-95

TIA/EIA-136 ANSI version of the TDMA air interface standard. Replaces IS-136, second generation TDMA air interface standard

TIPHON Telecommunications and Internet Protocol Harmonisation over Networks

TSG Technical Specification Group. Part of the 3GPP and 3GPP2 processes.

UMTS Universal Mobile Telecommunications Systems. The 3G initiative for Europe. Promising to increase bandwidth to between 384kbit/s and 2Mbit/s, depending on the distance to base stations and whether user is travelling or not. Uses W-CDMA. The UMTS specifications and the Japanese ARIB specifications were developed to be interoperable at an early stage.

VoIP Voice over Internet Protocol.

WAP Wireless Application Protocol. If you write your Web site in WAP, it can be read on a suitable mobile phone.

W-CDMA Wideband code division multiple access.

X.25 An ITU defined general purpose packet switching protocol

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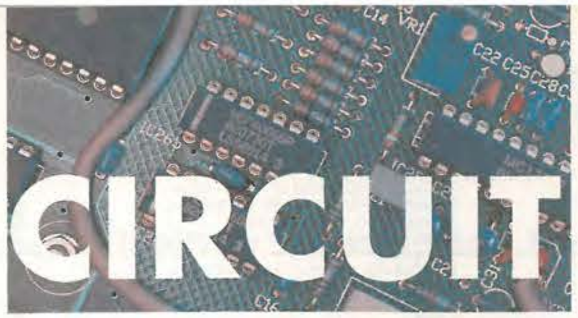


CIRCUIT IDEAS

Fact: most circuit ideas sent to *Electronics World* get published

The best circuit ideas are ones that save time or money, or stimulate the thought process. This includes the odd solution looking for a problem – provided it has a degree of ingenuity.

Your submissions are judged mainly on their originality and usefulness. Interesting modifications to existing circuits are strong contenders too – provided that you clearly acknowledge the circuit you have modified. Never send us anything that you believe has been published before though.

Don't forget to say why you think your idea is worthy.

Clear hand-written notes on paper are a minimum requirement: disks with separate drawing and text files in a popular form are best – but please label the disk clearly.

RS232-to-parallel data conversion

Convert RS232 format data to 8-bit parallel data cheaper than with a UART.

Here, standard ICs convert RS232 format to eight-bit parallel data with one start and stop bit. Total cost and power consumption is less than that of any standard chip available.

The circuit can be used as part of a PC-based serial interface through the COM1 or COM2 port.

A general purpose timer, IC_1 is used as the clock generator, producing clock pulses only on receipt of serial data. Data is shifted into a serial to parallel shift register, and along with each data byte (DATA) a latch pulse (LATCH) is generated for the parallel destination port.

No initialisation other than the bit-rate setting from the PC is required. The rate is equal to the clock, CL, generated by timer IC_1. Setting the divisor value for bit rate 'B' sets the baud value of the PC.

$$CL = \frac{1.4}{(R_A + 2R_B) \times C}$$

The divisor value for the PC is,

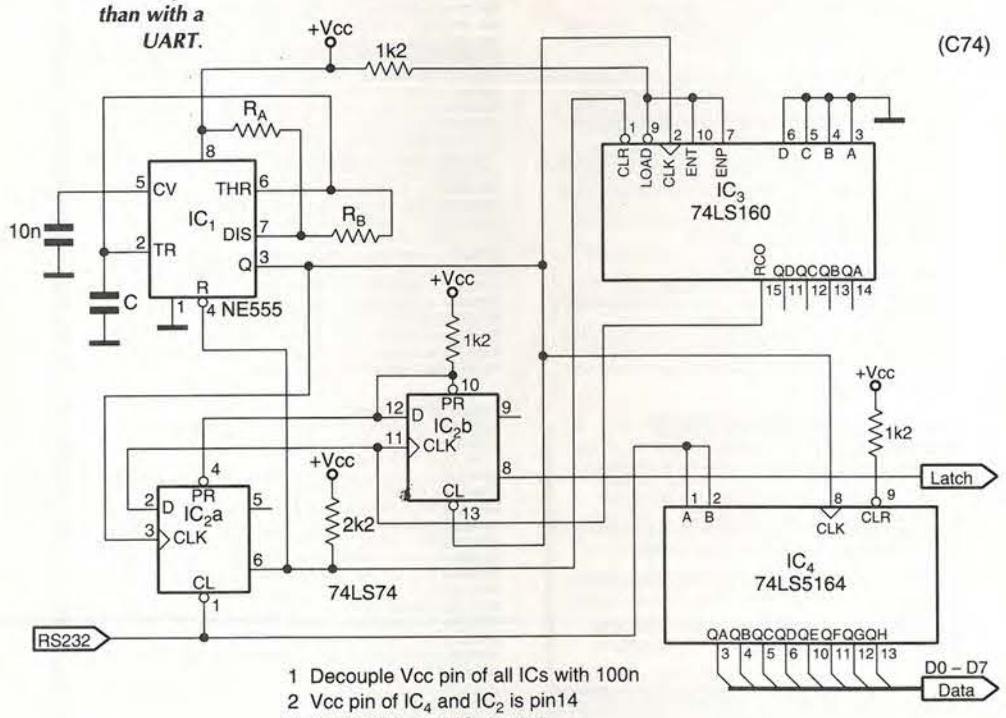
$$\frac{1843200}{16B}$$

It is possible to program the PC for any bit rate by putting the corresponding divisor value in 16-bit format using statements in turbo-C as given below, for COM1,

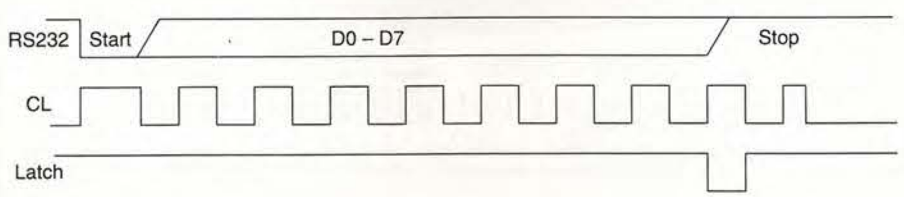
```
    outport(0x2FB,131)
;/*2F8 for Baud setting*/
    outport(0x2F8,0xLSB)
;/* LSB byte of divisor*/
    outport(0x2F9,0xMSB)
; /* MSB of divisor*/
    outport(0x2FB,3)
; /* 2F8 for transmission */
```

Any data in eight-bit format can be sent to pin 3 of the nine-pin D-type connector using statement 'outport(0x2F8,word)'. For COM2 use 3F8, 3F9 and 3FB.

S. Vijayan Pillai
Kerala
India
C74



1 Decouple Vcc pin of all ICs with 100n
2 Vcc pin of IC_4 and IC_2 is pin14
3 GND of IC_4 and IC_2 is pin7
4 Pin16 and pin8 of IC_3 are VCC and GND respectively



FDNCs using current conveyors

Using two current conveyors, the circuit of **Fig. 1**. realises the function $Y_{1-1}(s)=s^2D$, where $D=C_1C_2R_1$. The equivalent circuit consists of a parallel connection of the ideal frequency-dependent negative conductance, D, capacitance C and conductance G.

Parasitic components C and G are due to the non-ideal nature of the current conveyors. The stable condition of the circuit requires positive signs for C and G. **Figure 2** shows computer simulations of the negative conductance $-\omega^2D$ and C dependencies versus frequency. These were carried out for current conveyors constructed according to the Huertas concept[1] using two TL081s, four 100kΩ resistors and two 1kΩ resistors for each conveyor. Component values used were C_1=0.1μF, 1μF, 10μF; C_2=1μF and R_1=1kΩ.

In Fig. 2, the stable operation region of the circuit is marked by continuous lines. If capacitance C is negative the circuit becomes unstable.

The frequency-dependent negative conductance circuit forms a resonant circuit with resistance and can be used in resonant bridges capable of measuring capacitance with high losses, like the frequency-dependent negative resistance circuit used for measuring inductances with high losses[2].

L. Tomawski, M. Manka, M. Slawiec
Katowice
Poland
C75

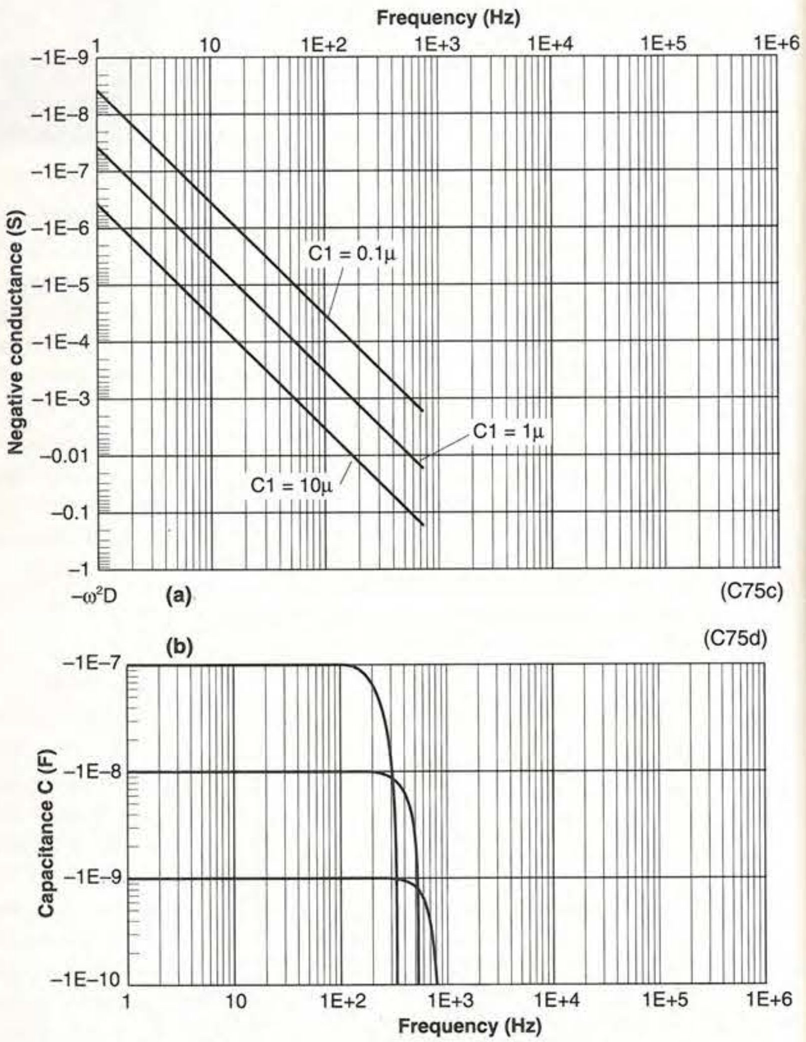


Fig. 2. Computer simulations versus frequency of a) negative conductance $-\omega^2$D and b), capacitance **C.**

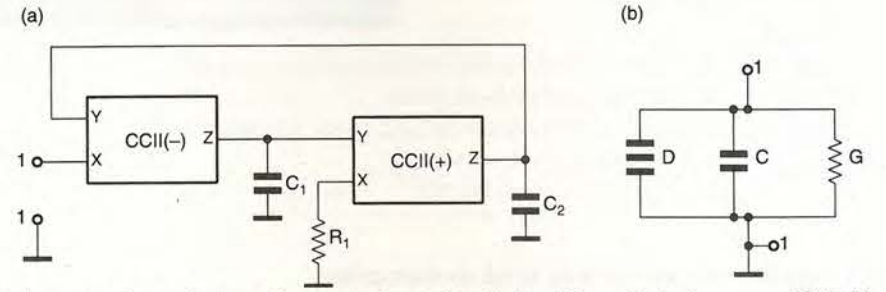


Fig. 1. A frequency-dependent negative conductance circuit, a) and its equivalent circuit, b).

(C75a/b)

References
1. Huertas, J L, 'Circuit implementation of current conveyor,' *Electronic Letters* No. 16, 1980, pp. 225-226.
2. Tomawski, L, 'Resonant bridge with Frequency Dependent Negative Resistance,' IEEE Trans. On Instrum. and Meas. No. 37 March 1988, pp. 45 - 48.

Efficient battery auto-charger

This efficient automatic battery charger commences charging when the battery voltage reaches the set lower limit and ceases charging when the upper level is reached.

Adjustments are provided for both limits, which are best set up using a variable dc power supply. With the mains supply disconnected, the 'fully charged' voltage of 14.5V dc is applied at the circuit's battery terminals, and VR_1 adjusted so that the relay is energised, cutting off the charging supply.

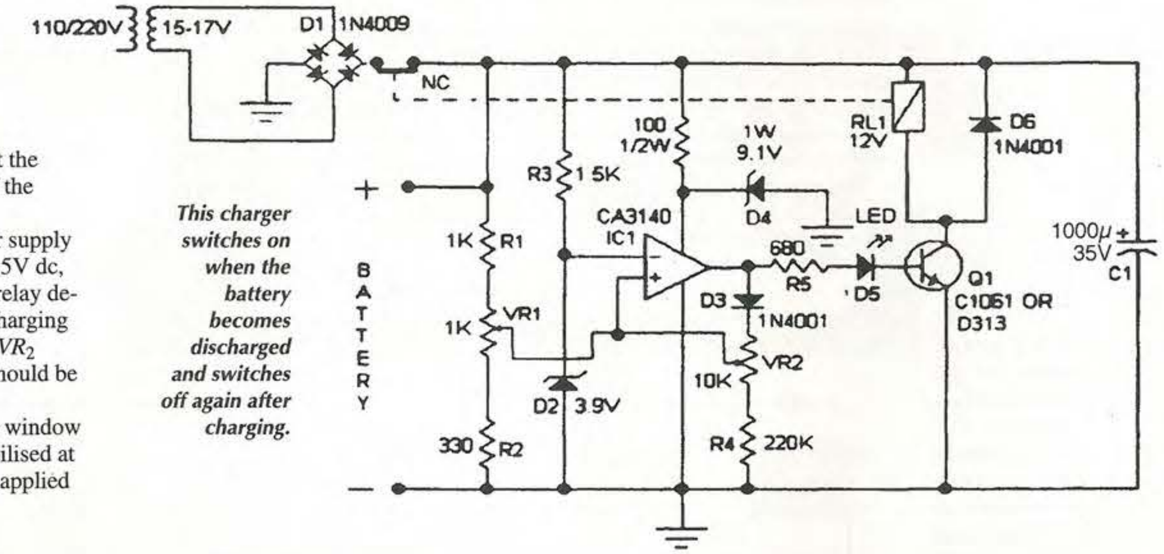
Next, the variable dc power supply is set to the lower limit of 12.5V dc, and VR_2 adjusted so that the relay de-energises, reconnecting the charging supply. Settings for VR_1 and VR_2 interact, so the adjustments should be iterated as necessary.

An op-amp forms a pseudo window comparator. Its supply is stabilised at 9.1V and a 3.9V reference is applied to its inverting input.

Battery voltage is sensed by divider chain R_1, VR_1, R_2 and applied to the non-inverting terminal. Positive feedback to the non-inverting terminal sets the lower limit at which charging recommences. The circuit can be used for charging various battery types.

Ejaz Ur Rehman
Islamabad
Pakistan
E5

Check the charge rate after implementation as it depends on output voltage and regulation of the transformer – Ed.



This charger switches on when the battery becomes discharged and switches off again after charging.

£75 Winner!

30W Class-A power follower

At first glance, this circuit may seem familiar. Essentially, it is a typical source follower working in pure Class-A with a constant-current generator forming the load. Features that differentiate it are its negative power supply and small power supply reservoir capacitor.

The current generator has a DC feedback through the BD139 to prevent thermal effects on quiescent current. Note that this circuit works only in class A, so it needs enough bias current for the required output power.

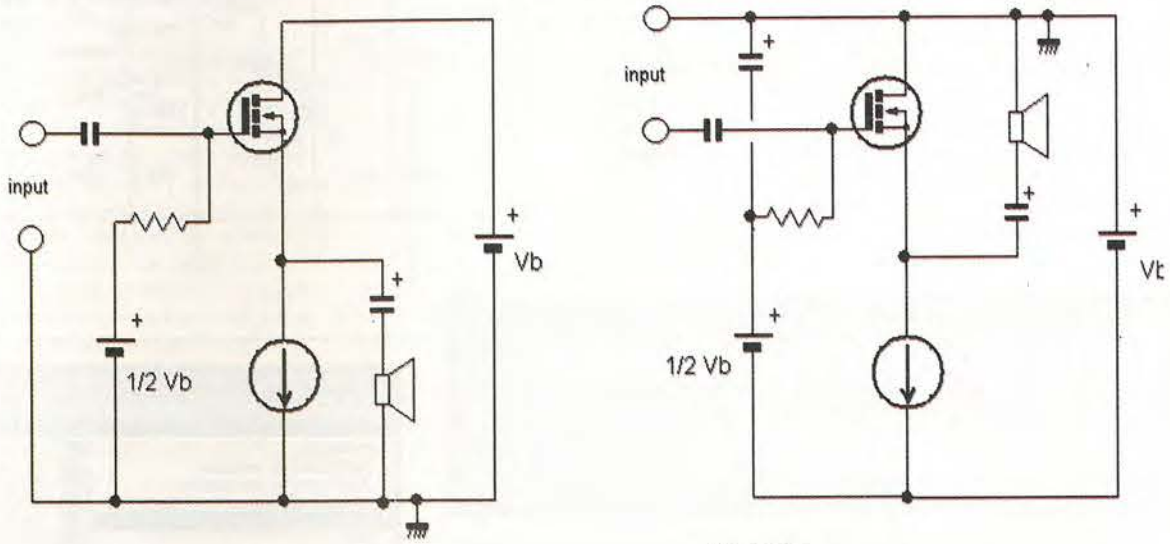
To keep the power dissipated under static and dynamic conditions the same, I have used identical devices for the follower and current source.

Class-A power amplifiers normally involve a large reservoir capacitor. This causes very high current peaks in the diodes, which is undesirable.

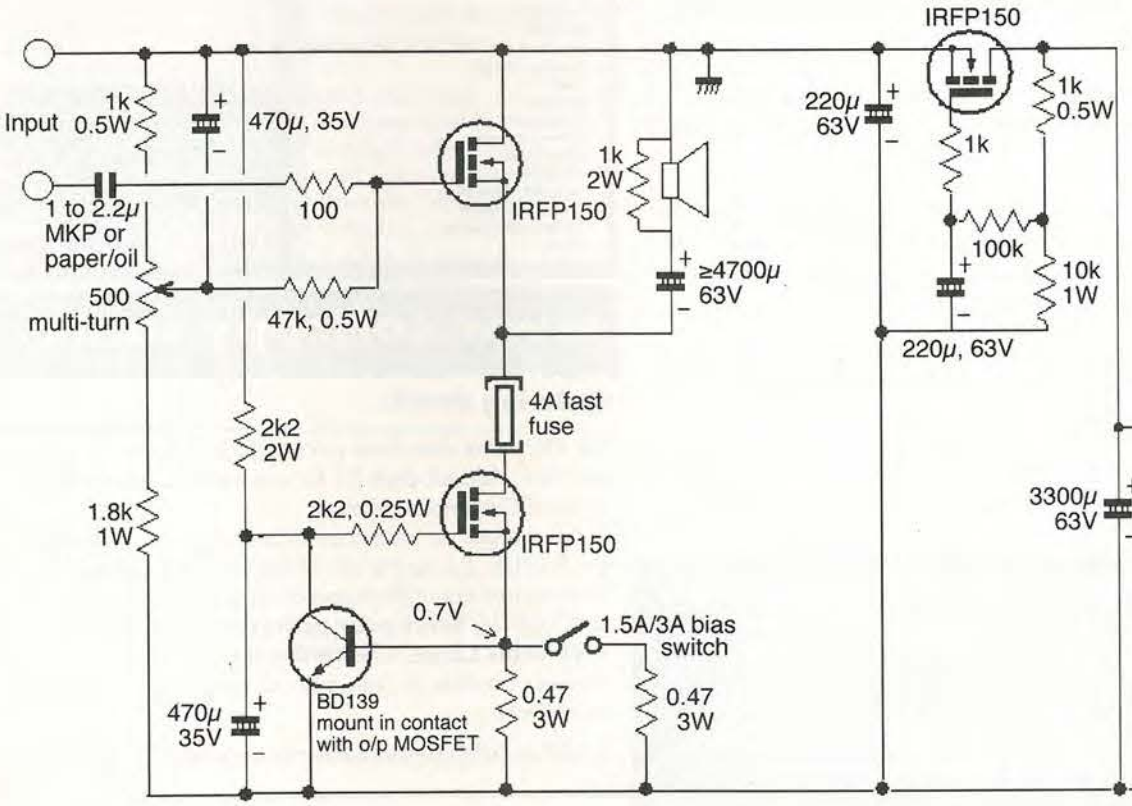
My circuit has a relatively small capacitor after the bridge. It is followed by a MOSFET voltage regulator featuring a very low frequency *CR* filter of 220μF/100kΩ.

Components used in the power supply have been simulated and optimised for a 3A quiescent current and with a capacitor value of 3300μF.

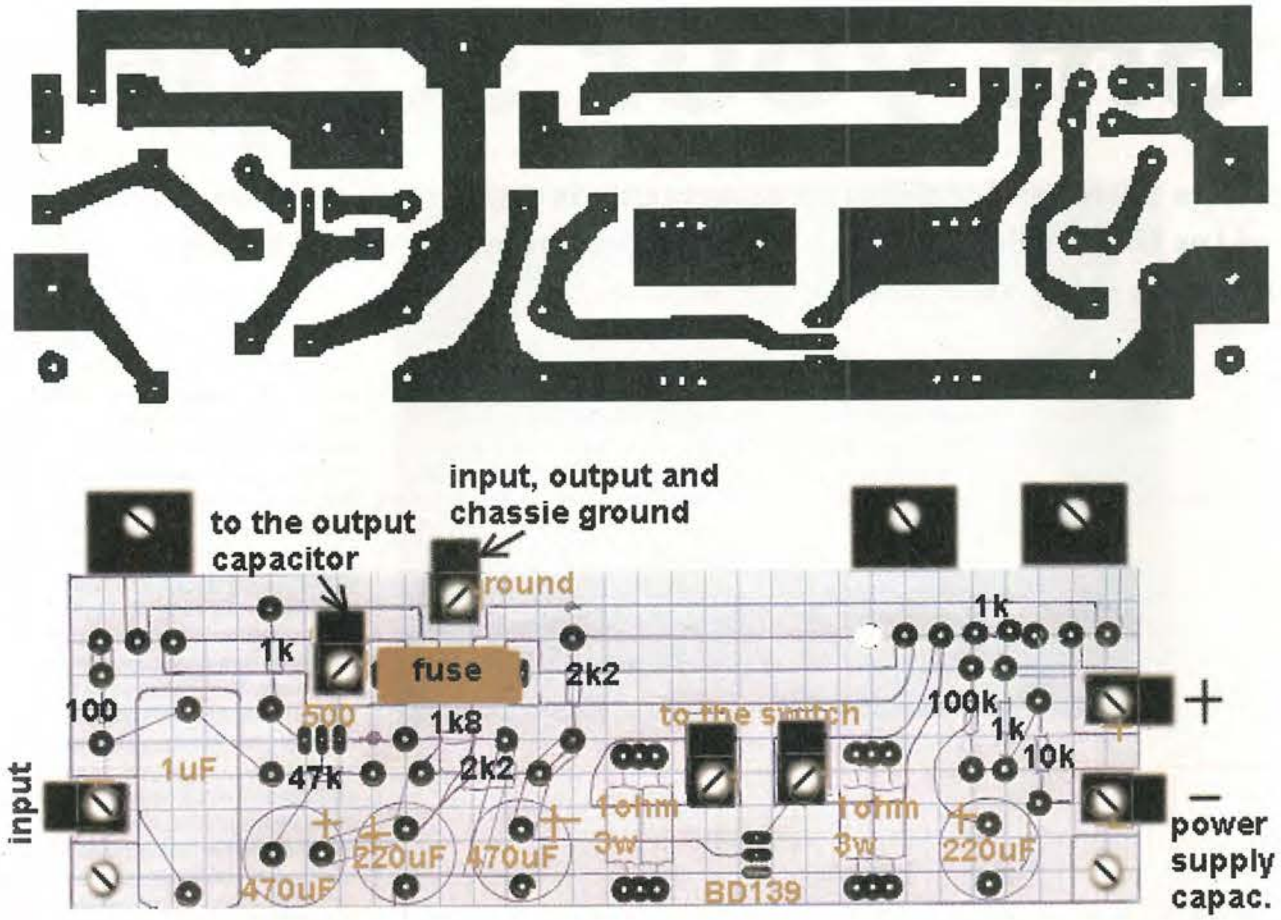
The 500Ω trimmer should be set to balance the output fuse voltage to half the regulator output. This single trim could be done after the switch-on; it simply allows you to get the maximum available output swing.



Class-A amplifiers normally use the configuration on the left. This power follower uses a negative power supply, as in the right-hand diagram, and the speaker is in parallel with the driving mosfet instead of the current source.



Complete Class-A power follower and its regulator/smoothing circuit. Note the bias switch to save power at low listening levels, and the relatively small reservoir capacitor.



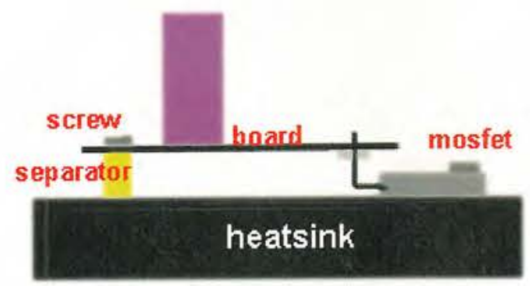
Quiescent current is set by the resistor at the current generator FET's source. It can be changed using I_q=0.7/*R* where 0.7 is the base-emitter voltage of the BD139.

I added a switch that allows the bias current to be changed. Sometimes I listen to music at low volume so the switch can be used to save power dissipation.

The power supply transformer can have a secondary of anything between 24V an 33V, but you need at least 250VA per channel. If you use a smaller transformer, you will have to reduce the bias current.

With a bias current of 3A you can get about 30W continuous into a 6Ω load. Power can be increased within the limits of the heat sink and power devices.

My prototype's PCB foil pattern, its layout, and a suggestion for mounting the circuit board on the heat sink.



I suggest using a heat sink of at least 30 by 42 by 16cm, or a smaller one in conjunction with a fan.

In my view, the input capacitor, the 470μF input bypass and the output capacitor affect the sound quality. I prefer to use ELNA Cerafine whenever possible.

Since the circuit has no voltage gain it has to be preceded by a voltage amplifier with an output swing not lower than 10V rms. It should also have an output impedance of less than 1kΩ.

Input impedance of the buffer is 47kΩ/1450pF. Its resistive value can be adjusted over a pretty wide range by just using a different input resistor.

The input capacitance of 1450pF may seem high but a gain stage with a 1kΩ output resistance gives a cutoff frequency of more than 100kHz.

A TO3 equivalents or other similar MOSFETs, such as IRF250s, IRFP250s, IRF240s or IRFP240s, can be used to replace the IRFP150s with a minimal impact. ■

Andrea Ciuffoli
Italy
by e-mail

Getting the most from your scope

Oscilloscope noise, and making measurements in noisy environments are the topics of Les Green's third article explaining how to make more reliable measurements using your scope.

The noise levels of DSOs and real-time oscilloscopes are not easy for the user to compare. All else being equal, one could reasonably expect a basic real-time scope to be quieter than a DSO because there are no noisy microprocessors and a-to-d converters to corrupt the trace.

So how do you measure the noise of your scope? If it is a DSO, it is really easy; select a slow time-base, turn on the max-min function and measure the noise band.

If you are the unlucky owner of one of those bandwidth-reducing scopes that I mentioned in my last article, make sure that you compare the noise level with a scope of the reduced bandwidth.

To measure the noise of your real-time scope you can use the Tektronix 'tangential noise' measurement technique. This involves putting the oscilloscope in add-mode and applying a small amplitude quiet square wave to the other channel. The trigger is set to auto so that the untriggered square-wave-plus-noise pattern causes two bright bands to appear on the screen.

By adjusting the square wave amplitude the two bands can be made to merge in a very repeatable manner – when the dark band in the middle just disappears – and the square-wave amplitude is a measure of the noise of the first channel.

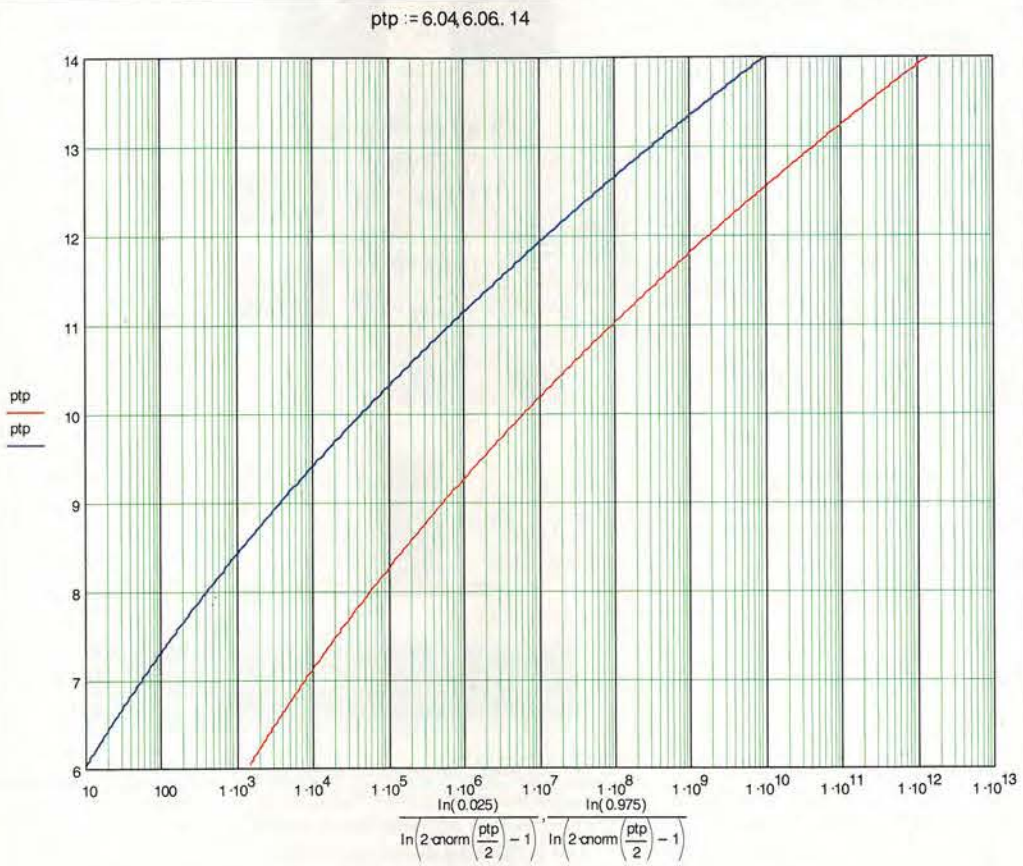
At first glance, anyone would complain that the DSO was very noisy compared to the real-time scope; the trace is much narrower on the real-time instrument. The problem here is the conversion from rms to peak-to-peak. Estimating the peak-to-peak value of a real-time scope trace by eye you may well obtain a value that is twice the rms value. This is $\pm\sigma$ as far as statistical tables are concerned.

Now look at the DSO result: it may be taking 100MS/s over a one-second measurement period and displaying the max-min values. Tables of the Normal/Gaussian distribution do not generally go above 5σ. However, even for this value the tabulated answer is 0.9999997133.

This says that on average 1 acquisition in 3.5 million will exceed the 5σ positive peak. I have used MathCad to generate a graph of rms to peak-to-peak conversion factors for Gaussian noise, **Fig. 1.** In the case of 100MS/s for 1 second the number of samples is 1E8. Reading the graph it says that 95% of the time the peak-to-peak noise level will be greater than 11σ and less than 12.7σ: σ being the rms value.

Clearly, for a DSO running at this speed a more reasonable conversion factor from rms to peak-to-peak is 12 – a conversion factor of x6 for general purpose work is often used, i.e. $\pm3\sigma$.

Conclusion: a DSO with the same rms noise level as an equal bandwidth real-time scope can have a max-min trace width six times as large!



Number of samples required to exceed ptp noise value 2.5% and 97.5% of the time
Hence for 95% of the time the ptp value will lie between the two curves.

CNORM is the MathCad cumulative Normal distribution function.

Probability of being less than +x is	cnorm(x)
Probability of being greater than +x is	(1 - cnorm(x))
Probability of being greater than +x or less than -x is	2(1 - cnorm(x))
Probability of being less than +x and greater than -x is	1 - 2(1 - cnorm(x)) = 2 . cnorm(x) - 1
Probability of being <+x and >-x after n trys is	p = (2.cnorm(x) - 1) ^ n
Re-arrange in terms of n;	n = log(p) / log(2.cnorm(x) - 1)

Fig. 1. MathCad generated graph of rms to peak-to-peak conversion factors for Gaussian noise.

Noisy measurements

Regardless of the type of scope used, there is often a problem measuring the actual noise of the circuit. This is due to the connection between the scope and the equipment being probed.

Unless you are using an isolated-input scope, the BNC connectors on the front panel have to be earthed. In order to meet safety standards like EN61010-1 the BNC outers have to be able to take 25A at the supply frequency and present an impedance of less than 0.1Ω.

As soon as you connect your scope probe earth-lead to the system being tested, a common-mode current flows down the outer braid of the coaxial cable. If the circuit being probed has much high speed logic on it, or a switched-mode power supply, there will be a significant high-frequency content to this common-mode current.

Before connecting your probe to an actual circuit node, try this simple test; clip the probe tip to the earthing crocodile-clip and touch the pair onto a convenient 0V point in the circuit to be probed.

If the noise measured on the scope becomes too large for comfort then you have a noise problem to resolve. This could be due to common-mode current alone or there could be an element of magnetic pickup, as the path from the coaxial cable screen via the probe earth-lead back to the probe tip has quite a large area.

Fortunately both pickup mechanisms are handled by the same technique. Removing the probe signal clip reveals the probe tip and a metal earthing ring. If the probe earth-lead can be unclipped, do so. If not then just clip it to the probe lead out of the way, so that it does not dangle around and short-out parts of the circuit.

Wind several turns of tinned copper wire – 6 turns of 22SWG for example – around the metal earthing ring of the probe and leave a short tail at the end – not more than an inch ideally. You can now solder the free tail to a signal earth and probe nearby points, **Fig. 2**.

It is also possible to use a finer coil of wire to contact the probe tip in order to make a hands-free connection to the circuit. Oscilloscope manufacturers know all about this but the information does not always get through to the user. Tektronix actually sells a little socket to plug probes into, which makes a nice semi-

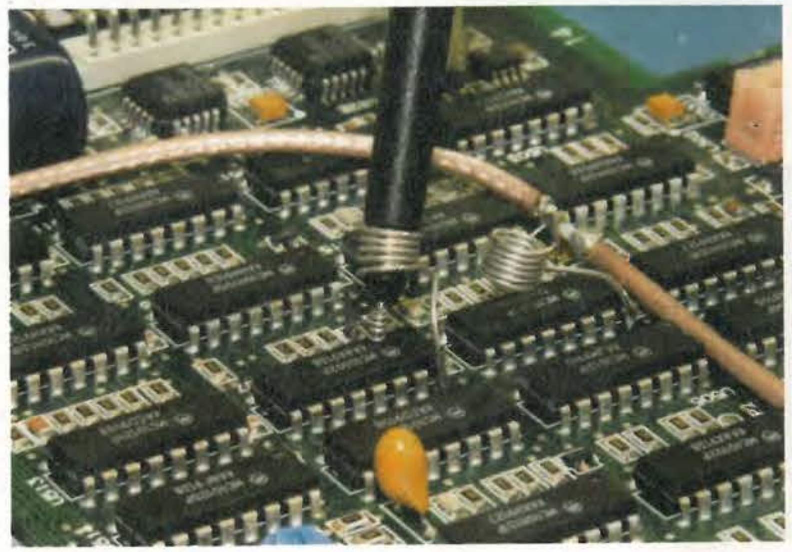


Fig. 2. Measurement technique for high-speed logic and for low-noise measurements. Note the other earthing coil so that pins on the other IC can be manually probed. Use 24SWG for the probe tip coil and 22SWG for the earth coil.

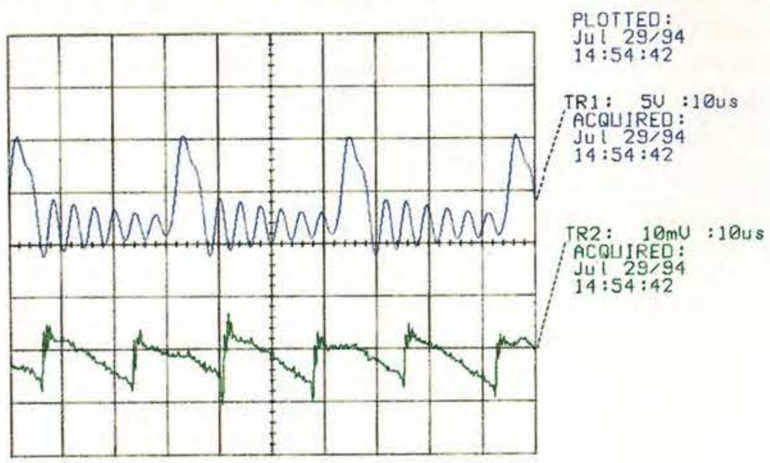


Fig. 3. Two readings of the same signal. Trace TR1 is the measurement with a 10:1 probe negligently placed near a line-output transformer, as in Fig. 5b). Trace TR2 is the same point measured with a coax cable soldered to the underside of the PCB, as in Fig. 5c). Measurements and plots were made on a Gould 475 with 20MHz bandwidth limit selected.

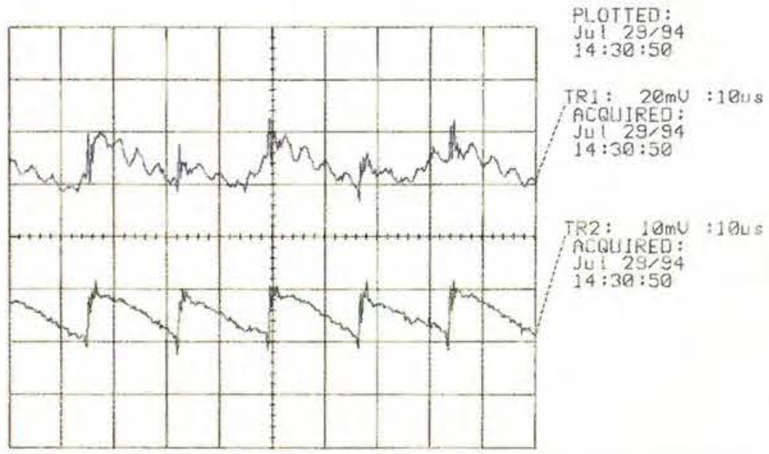


Fig. 4. Trace TR1 is measured with a 10:1 probe in its optimum position, as in Fig. 5a). The second trace is the same point measured with a coaxial cable soldered on the underside of the PCB, as in Fig. 5c). Measurements and plots were made on a Gould 475 with 20MHz bandwidth limit selected.

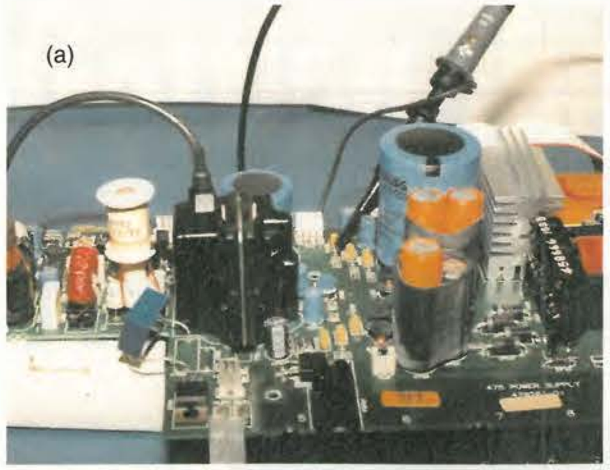


Fig. 5. Power supply and video drive assembly under test. The supply is a flyback converter and the line-output transformer in the centre of the picture is another flyback converter. In a), the 10:1 probe is reasonably well positioned, but in b) it is too close to the line-output transformer. The best solution is to solder a coaxial cable on the underside of the board, c).

permanent probing point for development work. It is manufactured by Specialty Connector Co., of Franklin, Indiana.

There is also such a thing as a spring-loaded probe head that plugs onto the end of a probe with an ATE type of sprung pin for the earth contact.

Even connecting between pieces of equipment via BNC-BNC leads can cause additional noise. Different types of BNC cable can significantly affect common-mode noise performance; the only answer is to try a few different ones and see which is the best in your application.

Another trick is to use a connection to an unused scope input – or even that funny earthed terminal on the front/back of the scope whose purpose eludes mortal man – as a parallel common-mode current path. The idea is to reduce the common-mode current flowing in the signal coaxial cable, to reduce the measured noise.

For low-noise measurements below 1MHz it is often convenient to do away with the probe entirely and just wire a piece of coaxial cable from your circuit to the scope, **Figs 3, 4** and **5**.

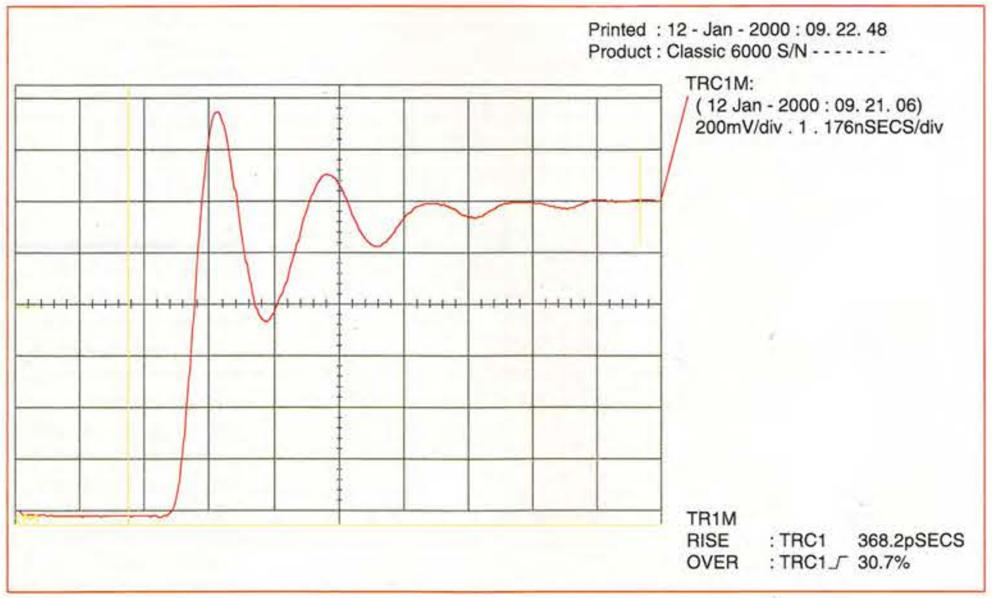
Use a proper BNC plug to connect to the oscilloscope and at the circuit end make sure that the connection to the coax braided screen is keep short, at less than an inch.

Note that the input capacitance of this arrangement may be around 100pF, depending on the length of the coaxial lead, so it may be necessary to put a 1kΩ resistor in series with the signal input to the coaxial cable. Do *not* put a resistor in series with the screen connection. The 1kΩ resistor will only give an attenuation of 0.1% into a 1MΩ input and the bandwidth will be nicely rolled off to around the 1MHz area.

There is another point to be made about measurements made in the presence of significant amounts of radiated field; it is not necessarily correct to move the probe to the position of minimum noise displayed on the scope screen. It is possible that the radiated noise can be made to cancel out some of the actual noise that exists on the point being measured. The measurement technique should be such that this type of cancellation is avoided. ■

The best of both worlds

This is the pulse response of an amplifier in development. It is being driven with a 200ps edge and being measured on an analogue sampling oscilloscope system, namely a Tektronix S-2 sampling head in a 7S11 sampling unit. To get cursor readouts and a stored trace, the analogue output of the 7S11 is fed into a Gould 6100 storage oscilloscope. Using averaging and measurement scaling, the 6100 is calibrated to give the correct rise time.



Room for improvement II

The listening room has a surprisingly significant effect on the loudspeaker. Last month, John Watkinson considered its effects at low frequencies. Here, he discusses how the room affects the rest of the spectrum.

SPEAKERS' CORNER

A real sound source in a real environment produces a combination of direct sound and reflected sound.

The human hearing mechanism has evolved to locate the sound source accurately, despite the reflections, and to assess the nature of the environment from the reflections. If either the source or its reflection is absent, there will be a lack of realism.

In principle, a stereo recording can capture sound from any direction, but the conventional stereo pair of speakers can only create virtual sound sources over an arc between the speakers. The 360° sound field will be mapped into that arc in the way the recording engineer thought best.

Clearly, listening to a sound field of restricted angle is not the same as listening to a 360° sound field. Fortunately the reverberation of the listening room can to some extent make up for the loss. Thus reverberation is essential in the listening room to create a natural result.

Note that the listening room can never create the same reverberant field as the original venue with only a stereo signal pair; but then this is not the goal. Reverberation is chaotic and provided it comes from all around the listener and has no gross defects, it will be accepted as plausible by the ear and the reproduction will appear more realistic.

In the case of surround sound, more channels are available. In principle, sound can be reproduced from any number of directions. But the argument still applies that with a finite number of channels, reverberation in the listening room is still necessary to create a continuous sound field from a set of discrete sources.

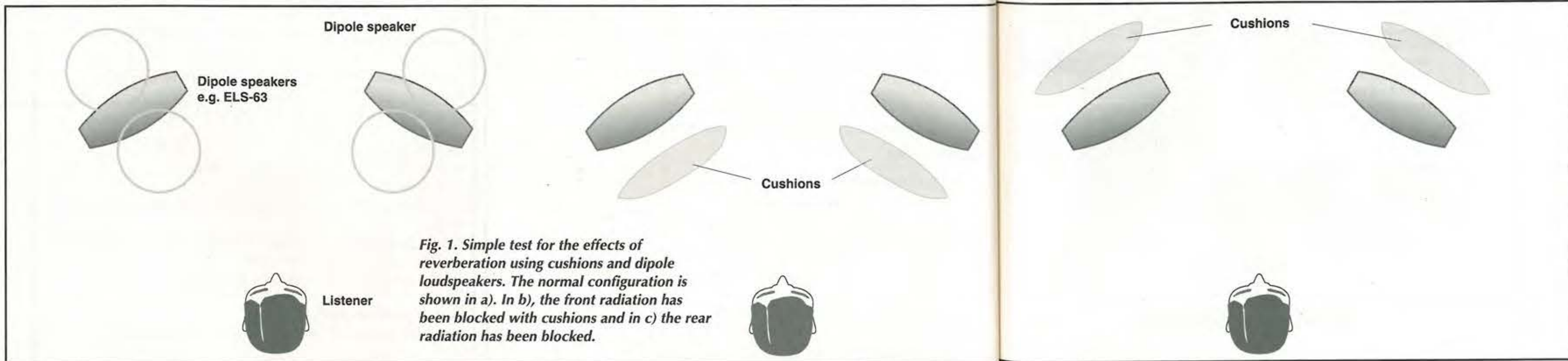


Fig. 1. Simple test for the effects of reverberation using cushions and dipole loudspeakers. The normal configuration is shown in a). In b), the front radiation has been blocked with cushions and in c) the rear radiation has been blocked.

See for yourself

In a domestic listening environment it is surprising how much of the sound we hear is reverberant. This is very easily demonstrated with any dipole loudspeaker – a Quad electrostatic for example.

In **Fig. 1a)** the speakers are in a normal listening position. In Fig. **1b)** a large pair of cushions has been placed in front of the speakers and in Fig. **1c)**, they have been placed behind the speakers.

The results are non-intuitive. In practice position Fig. 1b) sounds better than position Fig. 1c). This is because blocking the front radiation allows the rear radiation to excite the reverberation in the room. Blocking the rear radiation makes the speaker sound dull.

One of the great advantages of the electrostatic speaker is that it is inherently a dipole and the excitation of the room comes free. However, the same results will be obtained if the speaker contains moving-coil units facing forward and backward. Certain Bose satellite speakers contain two tweeters that can be swivelled into such an arrangement.

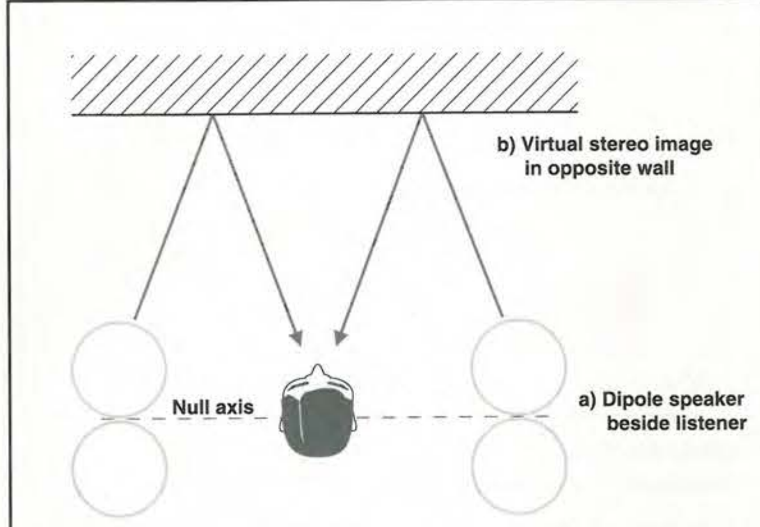


Fig. 2. Listening only to reverberation by sitting on null of dipole speakers.

Tweeters fore and aft

For experimentation, the moving coil speaker with tweeters facing forwards and backwards has the advantage that it is very easy to switch between a dipole and an omnidirectional radiation pattern simply by swapping the wires on the rear driver.

Despite the theories put forward about the massive differences in the excitation of the room between these two configurations, the simple fact is that in practice the skilled listener is hard pressed to tell which configuration is in use. There is no effect at all on the stereo image heard between the speakers; in fact this is still retained if one of the rear speakers is anti-phase and the other is in-phase.

For a moving coil system, having the tweeters working in phase as an omnidirectional source has the advantage that the low-frequency roll off due to cancellation is avoided. This means that a lower and less audible crossover frequency can be used.

Another surprising result using dipole speakers is shown in **Fig. 2a)**. Here, a pair of dipoles has been placed at either side of the listening position and adjusted so the listener's ears are in a null.

No direct sound at all reaches the listener from the speakers. A virtual stereo image will be heard in the opposite wall as shown in Fig. **2b)**.

Both of these examples illustrate just how much of the sound we hear is reverberant. As a result, it follows that an additional criterion for a precision loudspeaker is that as well as creating an accurate waveform on a direct path to the ear, it must also create a sufficiently accurate reverberant field.

This critical aspect of loudspeaker design has been obvious from psychoacoustic considerations for many years but the traditional loudspeaker industry, with a few notable exceptions, has passed it by.

Why do loudspeakers sound so different?

It is well known that loudspeakers with identical specifications sound different. There is only one conclusion to be drawn from that, which is that the specifications are incomplete.

In short, the traditional ways of testing loudspeakers only give part of the picture. It's like testing cars in a straight line and then wondering

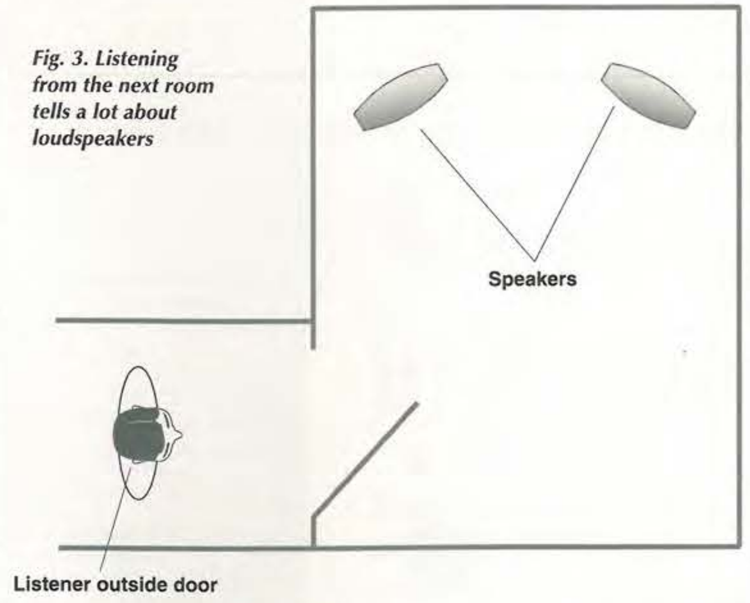


Fig. 3. Listening from the next room tells a lot about loudspeakers

why they corner differently.

In any other industry, a disparity between theory and practice results in research to advance the theory and close the gap. Why the audio industry should make itself a conspicuous exception could be the subject of a long discussion.

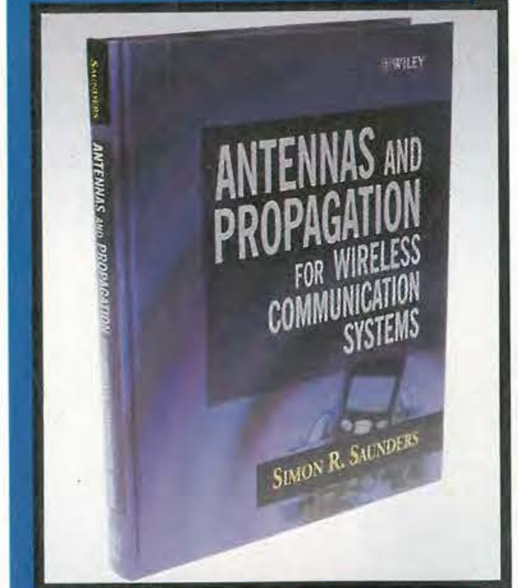
There is a trivially easy way of demonstrating the aspect of loudspeakers which makes them sound different. This is to listen from the next room via an open door and compare the tonal quality or timbre outside the room with that at the ideal listening position. **Figure 3** shows the principle.

Without exception, the loudspeakers that sound the most tonally balanced in the next room are the ones that sound most realistic in the ideal listening position. Also, without exception, the loudspeakers that perform best are those that have had some thought given to the quality of the off-axis sound that they produce.

The constructional philosophy of the speaker seems to be more important here than the amount spent or the transduction principle.

The importance of off-axis sound quality in a loudspeaker cannot be underestimated. In my next article, I will explore the consequences of getting it wrong ■

BOOK TO BUY



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- Methods that overcome and transform channel impairments to enhance performance using diversity, adaptive antennas and equalisers.

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- Extensive worked examples
- End of chapter questions
- Topical and relevant information for and about the wireless industry

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Phase-locked loops

Revealing useful web sites along the way, Cyril Bateman explains the phase-locked loop and discusses the options and design tools available for those involved with this invaluable building block.

Three years ago while researching for a phase-meter design, I surveyed the Internet for information on phase-locked loops. Since then, the drive to provide ever faster computer systems[1] has resulted in an explosion in the use of phase-locked loop integrated circuits. This is reflected in the much increased amount of phase-locked loop information and design aids now available for downloading.

A search of the Philips Semiconductors site alone brought up 407 matches for the phrase 'PLL'. But why search Philips first? Perhaps a brief look at the history of phase-locked loops will explain.

Phase-locked loop roots

The phase-locked loop concept was first proposed in 1922.

Probably the first ever general use of a phase-locked loop was in a 'homodyne' radio receiver in the 1930s. It was in the 1950s though that the rapid expansion of television receivers saw the universal adoption of phase-locked loops. PLLs were used in tv sets to synchronise the 'frame' and 'line' oscillators to the incoming signals – the so-called 'flywheel sync' circuit.

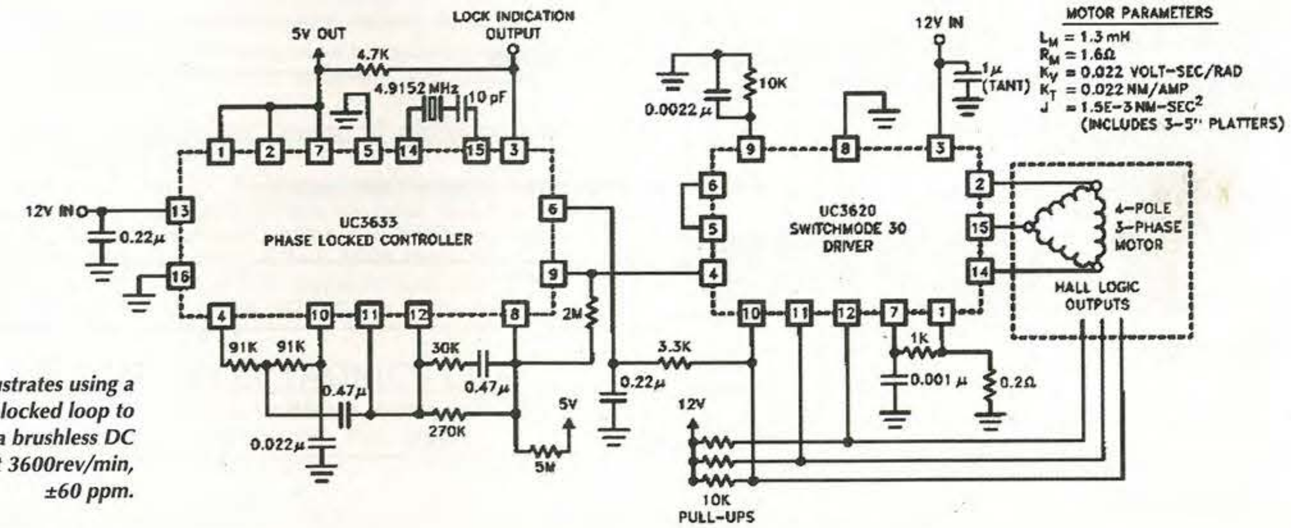


Fig. 1. Illustrates using a phase-locked loop to spin a brushless DC motor at 3600rev/min, ±60 ppm.

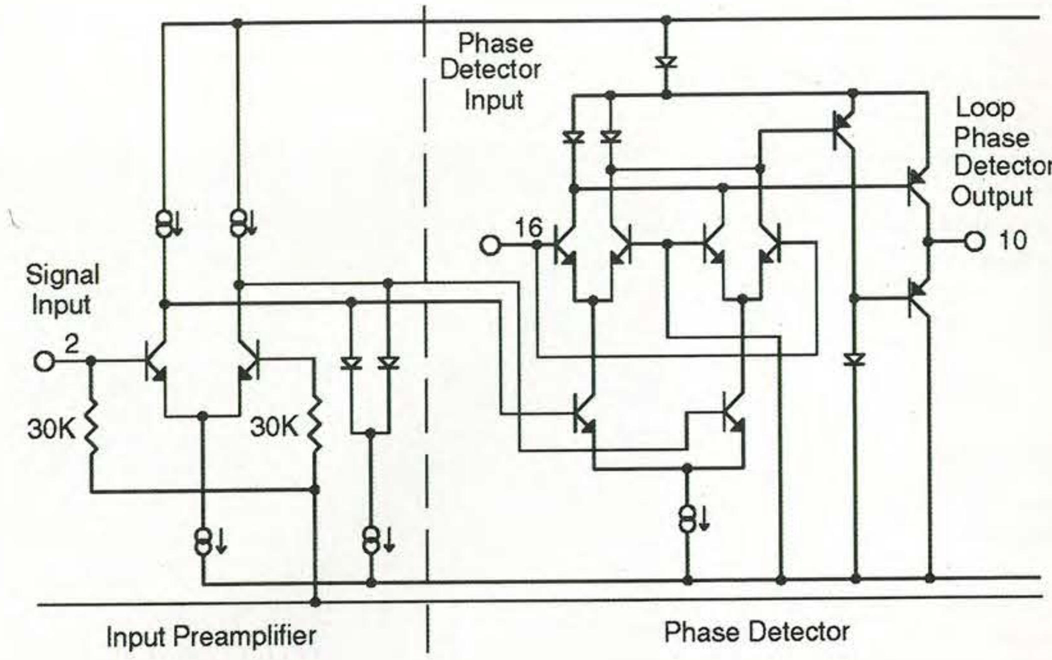


Fig. 2. Preamplifier and phase-locked loop schematics as used in the EXAR XR-2212 precision PLL integrated circuit. This illustrates use of a double-balanced mixer, or analogue phase locked loop.

Signetics claims to have produced the first ever dedicated phase-locked loop integrated circuits in the spring of 1970. These were selected as one of the hundred most significant technical products by the Industrial Research Institute in 1971.

Two other notable firsts that I recall were a phase-locked stereo decoder[2] and a phase-locked-loop fm tuner.[3] Both were published by *Wireless World.*

Today, phase-locked loops can be used to modulate and demodulate AM, FM, FSK and BPSK signals. They can detect the presence of specific frequencies buried in noise and produce voltage controlled frequencies or precise clock signals.

Phase-locked loops can also restore signal waveforms that have been distorted using fibre optic links and they can precisely control the speed of electric motors. In addition, PLLs are useful for restoring the timing waveforms when reading floppy and hard disk drives... the list is endless. **Figure 1** is an example of a precise motor drive.

Designing with PLLs

The 'flywheel' analogy is used to explain phase-locked loop behaviour in application note AN177. Entitled 'An overview of the phase-locked loop,' this note downloads relatively quickly.[4] I commend this as a first read for those of you wanting a simple, readable introduction to the topic.

A much longer 123 page technical article, which is complete with all needed equations and called 'Dean's Book', can be downloaded from National's web site.[5]

The phase detector used in the original Signetics phase-locked loop designs was based on the double balanced mixer circuit. When this method of phase detection is used, the complete integrated circuit is labelled as an 'analogue' phase locked loop, even though it may also include many digital stages.

Analogue phase detectors can be used with both small and large input signals and at very high frequencies. A good detailed description of the analogue phase detector, entitled 'Modelling the PLL double-balanced mixer' can be downloaded as application note AN178.[4] **Figure 2** is an example of an analogue phase-locked loop from inside an Exar XR2212.

Many of the most popular integrated phase-locked loops use a digital phase detector. For this reason, they are labelled as 'digital' phase-locked loops, even though analogue stages may be included.

Perhaps the best known and widely used is the CMOS 4000 series 4046 integrated circuit. While the maximum operating frequency of this particular circuit is restricted, other designs can be used at 100MHz and above.

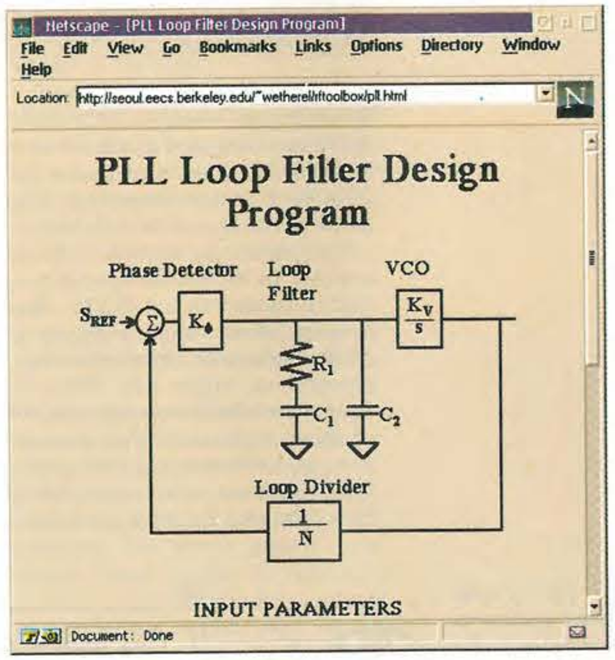
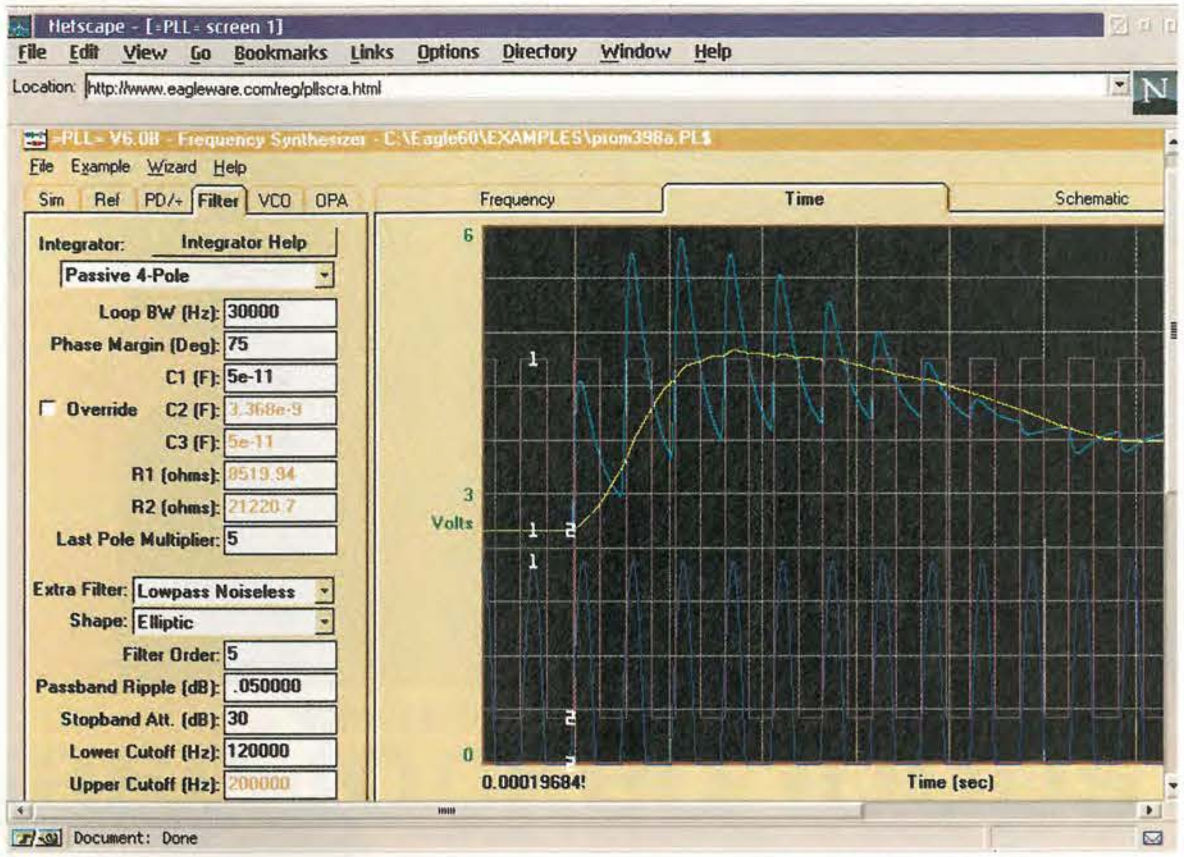


Fig. 3. An easy to use, on-line design aid that designs the loop filter for you, in response to your required parameters.

Three types of digital phase detector

Three common types of 'digital' phase detector circuits are in use.

Fig. 4. A screen shot of the phase-lock-loop simulator, one module in the EagleWare Genesys software package.



Sometimes all three types are included in the one integrated circuit. Since it includes all three, I will use the 74HC4046 as an example here. This device is usable to some 15MHz.

The simplest digital phase detector is based on the exclusive-or circuit, and is similar to a 74HC86. This detector, which should be driven by 50-50 mark-space input waveforms, provides an output zero when the input waveforms are exactly 90° out of phase. Its output is then a square wave at double the frequency of the input signal and with a duty cycle of 50%. The actual duty cycle of this output waveform varies according to the phase difference of the input signals.

The second style of phase detector is effectively a digital memory network comprising four bistable devices, some gate logic and a tri-state output. These bistable devices act only on the rising edges of the input signals, so this detector works independently of the input signal's duty cycle.

When both input signals have the same phase, the tri-state output becomes a high impedance. Its output becomes a logic high or logic low only while the signal phases differ.

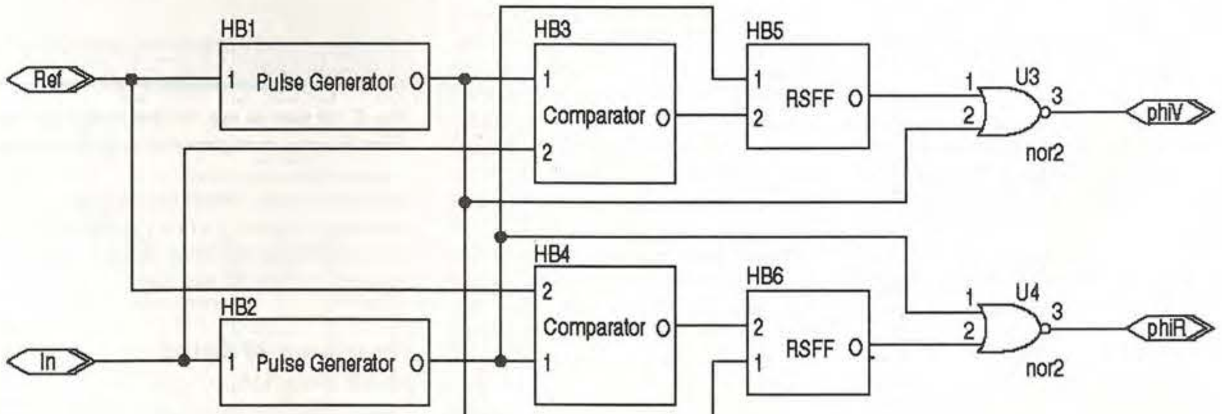
The third type of phase detector is again positive-edge triggered, so is also independent of the input signal duty cycles. Based on an RS flip-flop, it outputs a square wave at the input signal frequency with a duty cycle that varies with phase difference.

When the input signals have a 180° phase difference, this detector output waveform has a 50% duty cycle.

Designing loop filters

You can see from the above that for most applications, the phase detector output must be filtered. This filter may be active, or more simply can be based only on passive components.

Fig. 5. Digital phase detector as modelled for the 'MC145170 PSpice modelling kit'. This detector can be assembled and simulated using only standard PSpice model elements.



Where to browse

1.	High-speed PLL chips ease transition to faster processors.	http://devel.penton.com/ed/Pages/magpages/sept0198/analog/0901ao2.htm
2.	Phase-locked Stereo Decoder, R.Portus & A.Haywood, 'High fidelity designs,' pub. *Wireless World* 1974	
3.	A phase-locked loop fm tuner, J. Linsley-Hood, *Wireless World* annual 1975.	
4.	Philips Semiconductors	http://www-us.semiconductors.philips.com
5.	National Semiconductors	http://www.national.com
6.	Electronic Design/Development Software	http://mypage.uniserve.ca/~gludwig/electron/software.htm
7.	Electronic Design	http://www-inst.eecs.berkeley.edu/~wetherel/rftoolbox/rfdesign.html
8.	PLL Loop Filter Design Program	http://seoul.eecs.berkeley.edu/~wetherel/rftoolbox/pll.html
9.	Interactive Digital Phase Locked Loop Design	http://www-users.cs.york.ac.uk/~fisher/mkpll
10.	Eagleware Corporation	http://www.eagleware.com
11.	SPICE Tools Provide Behavioral Modeling of PLL.	http://sites.penton.com/mwrf/librarypenton/archives/mrf/November1997/501.htm
12.	MC145170 PSpice Modeling Kit	http://www.mot-sps.com/lit/html/PLL.html

Correct filter design however is most important for successful circuit operation. Filter design can prove time consuming if done iteratively. Solving the mathematical design equations can prove tedious.

Three years ago I found only one design aid available for downloading. This was version 2 of the 'HCMOS phase-locked loop design program,' which I downloaded from the Philips site.

When I looked again this January, I was unable to locate it, but I did manage to find an earlier version.[6] This is a DOS-based program that allows you to explore 'what if' designs for the Philips HC/HCT4046/7046 and HCT9046 phase locked loops by answering a few simple questions.

With little effort, I managed to find several other design and simulation aids. One particularly helpful page I found was on a university site.[7] It covered many other topics besides phase locked loops. In fact, it was so useful that I bookmarked it.

On-line design aids

Some sites now offer on-line design aids for phase-locked loops. I found the first one on the Berkeley site. It provided an easy to use on-screen representation of a phase-locked loop, especially suited to first time users, **Fig. 3.**[8]

I also found an interesting and interactive on-line package on the UK's University of York web page. Written in Java code, this provides performance graphs in addition to calculating design values.[9] Many designers though will prefer the convenience of having their design tools stored within their own computer.

National Semiconductor[5] offers three software packages as part of its series of application notes for the design of wireless systems. The first is a Windows program, titled 'Windows Loop Filter Design'. This 3Mbyte download helps with the design of loop filters.

The second is a smaller DOS program, entitled EasyPLL, which assists with component selection and designs and simulates the loop filter performance.

National's final offering is a number of miscellaneous phase-locked loop models to aid loop design and simulation. Many of these can only be run via Mathcad 7. Since I only have an older version of Mathcad, I was unable to try these out.

High-frequency applications

While the various Spice time-domain based simulators are extremely popular, many have restricted capability when performing analysis of high-frequency circuits. Historically high-frequency design has been based on frequency domain and harmonic balance simulators.

One phase-locked loop design module dedicated to rf design is Genesys from the EagleWare Corporation[10]. The EagleWare system is a complete system of design tools, targeted to RF designers' needs. These can be purchased either as separate modules or a package.

There are eleven modules in the full Genesys package. These include schematic entry, simulation, RF and conventional layout, filter synthesis, as well as the dedicated PLL module.[4]

Using PSpice analogue behavioural modelling techniques, a method of simulating the performance of digital phase-locked loop systems running at several hundred megahertz has been devised by Spanish RF engineer Orlando Pena.[11] This routine was developed using the PSpice simulator in version 7.1 of the Microsim DesignLab.

I have found this version of PSpice is particularly useful. While aimed at Windows 95 or Windows NT, provided the Win32S v1.3 extensions have been installed, it will run on a modest 486PC with Windows 3.1x operating systems.

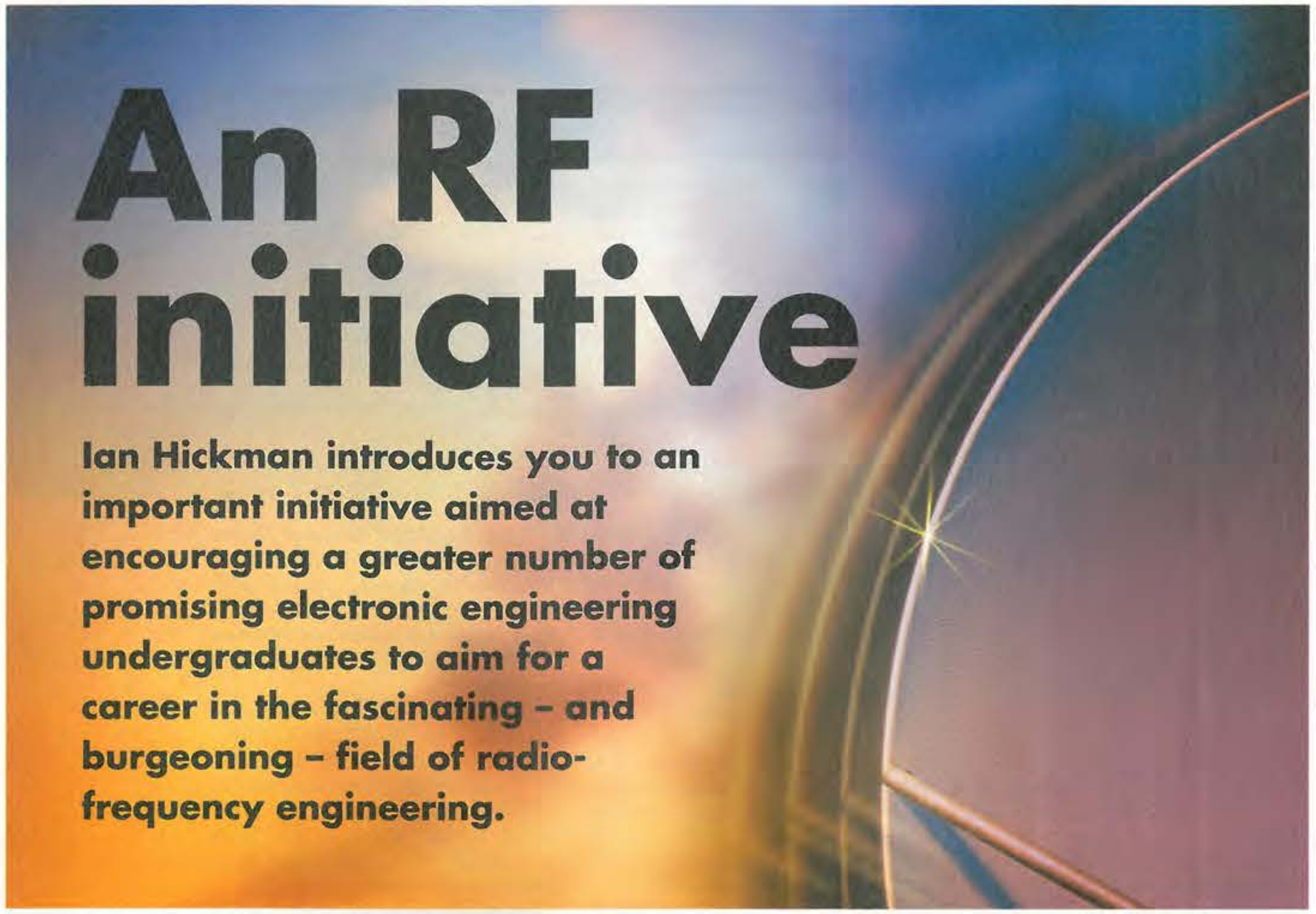
The particular phase-locked loop he modelled was based on a digital phase-frequency detector, as used in the Motorola MC4044 integrated circuit. It has an analogue charge pump, which Orlando based on the Spice models for 2N222 and 2N2907 transistors, as the output.

Potentially even more useful is the 'MC145170 PSpice Modelling Kit' because the full Spice models used have been included. This is described in application note AN1671, which can be downloaded from the Motorola web site.[12] While the models included in this note are specific to this Motorola part, it does include sufficient background information to help you develop models for other integrated circuits.

Frequency synthesis

The MC145170 is a PLL frequency synthesiser integrated circuit complete with serial communications, reference oscillator, phase detector and counters. The phase detector and charge pump circuits are similar to those assumed for the Pena article, except for the models used for the charge pump. These simulate the actual devices used in the MC145170, **Fig. 5.**

The note also includes details of model changes needed to enable these simulations to be made using only the evaluation version libraries supplied with PSpice. I particularly liked the section that advised on providing for component and circuit board parasitics. When these parasitics have been overlooked, the measured results of a practical circuit and its Spice simulations will not agree. ■



An RF initiative

Ian Hickman introduces you to an important initiative aimed at encouraging a greater number of promising electronic engineering undergraduates to aim for a career in the fascinating – and burgeoning – field of radio-frequency engineering.

The Institution of Electrical Engineers started life in 1871 as the Institution of Telegraph Engineers. What must they have thought, I wonder, when along came others, pioneering the transmission and reception of messages without wires?

Naturally, this new technique was called Wireless Telegraphy, WT or just wireless; télégraphie sans fil in France, or just TSF. Later the term 'wireless' was used also for wireless telephony and later still broadcasting, as also TSF – téléphonie sans fil.

But the term wireless, used regularly by my grandparents, fell into disuse, probably in the 1930s. It was replaced by 'radio' in both English and French, and even sometimes in German, in place of the more common word, 'rundfunk'.

Now, the term 'wireless' has been unearthed, dusted off and pressed back into service, in its most literal sense. Short-range UHF or microwave links are in vogue for the exchange of digital traffic such as data, for example the Bluetooth standard. What term could be more appropriate than 'wireless'?

Applications of wireless communication are booming in a dozen other fields – telemetry, remote meter reading, etc., and of course, most notably, mobile phones.

Table 1. Organisations participating in the RFEEI.

- Aerial Group
- Cellnet
- Defence Evaluation Research Agency
- Department of Trade and Industry
- Ericsson
- Hewlett Packard
- IFR
- Lucent Technologies
- Metapath Software
- International
- Motorola
- NEC
- Nokia
- Nortel Networks
- NTL Telecommunications
- Orange
- Plextek
- Racal
- Radio Communications Agency
- Simoco
- Sony
- Vodaphone

Radio engineering

All these current applications, together with others forthcoming, such as UMTS – universal mobile telephone system – are creating an ever increasing demand for engineers skilled in radio frequency engineering. This is a demand that at present cannot be filled. The situation can only get worse, unless suitable steps are taken.

It is with this need in mind that the Department of Trade and Industry has launched the RF Engineering Education Initiative, or RFEEI. Under this, the electronic-engineering departments of universities receive a three year award aimed at encouraging interest in a career in RF engineering.

The scheme is being sponsored by a number of companies and Government departments.

Ride the wave

Organisers of the RFEEI have produced a pack of literature called 'Ride the wave'. It contains a booklet about careers in radio engineering and leaflets describing the activities of the various companies supporting the Initiative are included.

I am privileged to be assisting with the Initiative at one of the participating universities, the University of Portsmouth, just a dozen miles from my home. The university has made a room in one of the Electronic Engineering labs available for use by a 'radio club'. This is an informal arrangement whereby interested students meet one evening a week in their spare time.

My spot is on a Monday evening with second year students, a colleague running a similar arrangement for the first year students on another evening. The club even lays on snacks for the delectation of the students (and yours truly).

Starting with meetings late last autumn, the usual arrangement has been for me to think up a simple but interesting project for the students each week. Working alone or in small groups, they construct, complete and get the project working in a single session of a couple of hours or so, with a little help and advice if need be.

Initially, the projects were very simple. Few of the students had much experience of constructing circuits from the ground up. They had of course done lab work as part of their courses, but this tends to be very structured. Few opportunities had arisen for them to encounter the problems that arise in development work of any sort. These problems are often trivial, but baffling until you have learned how to sort them out.

Most of the students had had no active engagement in electronics prior to embarking upon their degree course. This is in contrast to young electrical engineering students of my generation, who had mostly been exposed to wireless or radar servicing during National Service, or had constructed amateur radio kit from the copious amounts of ex-government electronic gear on the post-war surplus market.

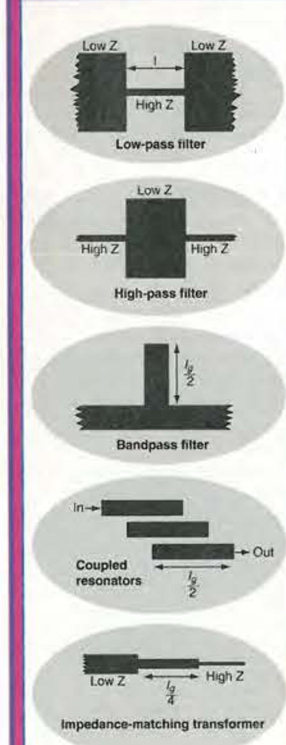
Expanding the scheme?

To date, five universities are the lucky recipients of funding under the scheme. It is to be hoped that it is successful – and seen to be so – leading to its future extension.

In the mean time, there is nothing to stop other universities starting their own home-grown version of the scheme. In fact, I know that one or two have already done so, but at least one such scheme has fallen by the wayside. This was due to lack of enthusiasm, *not* on the part of the students or of the organisers, but of their superiors.

A university wishing to start such a home-grown scheme need not incur any great expense. While a bespoke club room is a nice luxury, the after-hours activity can always be accommodated in one of the electronic engineering labs. And the projects can use inexpensive commonly available parts, many of which will in any case already be in stock.

As an assistance to any department wanting to run such a scheme, I hope – given the Editor's welcome enthusiasm for the Initiative – to publish simple projects, suitable for such a venture, in these pages in the forthcoming months. ■



Microstrip – an example of RF design technique. Using microstrip allows you to make components and even circuits using nothing more than profiles etched on laminated PCB material. These diagrams are extracted from Geoff Lewis's Communications Technology Handbook'.

NEW PRODUCTS

Please quote *Electronics World* when seeking further information

2mm extender

The Robinson Nugent Metpak Two range of 2mm connectors includes an extender board option that allows butt mating (edge-to-edge facing) of PCBs. The 2mm interface and footprint comply with IEC1076-4-001, 48B and EIA/SP-3179 specifications. The design enables inverse female modules with integral guide pin to interface with standard male modules. When mated the connector bodies are 23mm long including the shroud and guide pin. PCB tails are available in press fit and solder, and plastic force fit pegs ensure retention to the PCB during soldering. Available pin counts are 30, 150 and 210 but other multiples of 30-position five row mono-blocks are possible.
Robinson Nugent
Tel: 01227 794495
Enquiry No 501

622MHz VCXO

A voltage-controlled crystal oscillator for time-multiplexing applications in SDH STM-4 and Sonet STS-12 synchronous digital trunk lines has been launched by C-Mac. The 622.08MHz CFPV-2365 uses bulk-acoustic-wave crystal technology. Frequency stability is ±20ppm from 0 to +70°C or ±30ppm from –25 to +85°C, with frequency pullability between ±80 and ±120ppm. Frequency pulling locks the oscillator onto an incoming signal at variance with the output frequency. The oscillator circuit runs at 155.52MHz, but its fourth harmonic is selected and amplified to provide the 622.08MHz output frequency. It operates from a 3.3V power supply. A sine-wave output is provided.
C-Mac Frequency Products
Tel: 01460 74433
Enquiry No 504

1.1W transistors

Zetex's ZXT13 transistors use the firm's Supersot4 technology in a SOT23-6 package. Power dissipation is 1.1W and continuous collector current up to 5A. There are six devices – three n-p-n and three p-n-p – with ratings from 12 to 50V. To allow switching with low base currents, the current gain of each device is from 300 to 900. Typical saturation voltages are down to 45mV at a collector current of 1A, with a base current of 10mA.
Zetex
Tel: 0161 622 4422
Enquiry No 503

Programmable SLIC

Mitel Semiconductor has launched the MT91600 programmable subscriber-line-interface circuit. It provides the interface between a switching system and a subscriber loop for short-range applications. The device provides the normal interface functions between the codec or switching system and the analogue telephone line, including battery feed, programmable constant pulse detection, user-definable line and network balance impedance, and off-chip audio gain programming. It can accommodate a programmable current range between 18 and 32mA.
Mitel Semiconductor
Tel: 01793 518000
Enquiry No 505

Synchronous rectification IC

International Rectifier has introduced the IR1175 synchronous rectification IC (SRIC) which is designed to provide optimised gate drive for the MOSFETs configured as synchronous rectifiers in the secondary side of an isolated DC-to-DC converter. The IC programs the gate drives to ensure that the losses through the synchronous rectifiers are minimised and the parasitic diodes do not conduct. The IR1175 SRIC boosts converter efficiency to 90% at 3.3V output, claims the supplier.
International Rectifier
Tel: 01883 732020
Enquiry No 506

16-bit d-to-a converter

The LTC1657 from Linear Technology is a parallel 16-bit voltage output d-to-a converter for 5V single supply systems. It includes a rail-to-rail output buffer and an internal 2.048V reference. This device includes a 16-bit or 2-byte double-buffered architecture that allows glueless interface to micros and DSPs. Differential non-linearity is ±1LSB maximum. The device consumes 650µA typical and includes an internal reference and rail-to-rail output buffer that can drive capacitive loads. Power-on reset clears the output to zero. The package is a 28-pin SSOP or PDIP.
Linear Technology
Tel: 01276 677676
Enquiry No 507

Audio DSPs

Toshiba Electronics has released digital signal processors supporting audio compression formats in silicon audio players, DVD players and TVs. The TC9446F-004 single-chip DSP supports decoding AAC (two channel) and MP3 compression formats and

the TC9446F-003 supports decoding for Dolby Digital, Dolby Prologic and other audio-visual systems including DVD. Using a 24-bit by 24-bit processor, which incorporates 12k word data RAM and 12k word program ROM, both have a 128 word program RAM. Up to 1Mbit external SRAM can be used as data RAM. Operating at 3.3V, they come in a 100-pin QFP.
Toshiba Electronic
Tel: 00 49 211 52960
Enquiry No 508

LCD card for PII/III

An LCD CPU board from Anders comes in a half-size format supporting Pentium II and III processors. Using industrial specification LCDs, these boards have applications in information kiosks, PoS advertising, games machines, medical equipment

Multi-rate Ethernet transceiver

National Semiconductor claims to have the industry's first single-chip 10/100/1000Mbit/s Ethernet transceiver, which is backward compatible to existing Ethernet infrastructures. The DP83891 Gig Phyter is a full-feature physical layer transceiver with integrated PMD sublayers to support 10BASE-TX, 100BASE-TX and 1000BASE-T Ethernet protocols, and is compliant with IEEE 802.3 specifications, including: IEEE 802.3ab, IEEE 802.3u, IEEE 802.3 compliant Auto-Negotiation, IEEE 802.3u MII and IEEE 802.3z GMII.Gig PHYTER is designed for implementation of 10/100/1000Mbit/s Ethernet LANs. It uses standard single Quad TX-Transformer interface and performs error-free operations at over 140m, 40m greater than IEEE specifications. The intention is to support deployment of the Gigabit technology within an existing 10/100Mbit/s environment using Cat. 5 cabling. Other features include fully IEEE 802.3u-compliant Auto-negotiation at all three speeds of 10, 100 and 1000Mbit/s.
National Semiconductor
Tel: 00 49 870 24 0 2171
Enquiry No 502

Please quote *Electronics World* when seeking further information

and avionics. The HS-6200 supports CPU speeds above 650MHz and data bus speeds of 100MHz. Able to drive TFT, DSTN or EL LCDs up to 800 x 600, with 2Mbyte video memory, the product includes a power connector. The board can be used without a backplane. Although the single board computer is available as a component, the firm can also supply a fully configured product with display, touch screen, bios and cabling. Able to support 100Mbit/s Ethernet, it has USB, IrDA and two serial ports. There are also a bidirectional parallel port, two IDE ports and Dimm support for SDRAM up to 1Gbyte. The device can be configured with two floppy drives. Disc-on-chip to 144Mbyte is possible. The board includes a PISA (PCI+ISA) connector for backplane expansion.
Anders Electronics
Tel: 0207 388 7171
Enquiry No 509

RF power transistor

Ericsson has introduced the PTF10149 RF transistor for the 920 to 960MHz band. It uses Goldmos technology. With an output power of 70W, it is suitable for GSM mobile phone base station transmitters. Gain is 16dB, linearity ±0.25dB and efficiency 50 per cent. It operates from a 28V supply and has a minimum drain-source breakdown voltage of 65V. The n-channel enhancement mode FET has a load mismatch tolerance of 5:1. Class AB two-tone third order intermodular distortion is –37dBc at 20W peak envelope power. It uses gold metallisation, ion implantation and surface passivation.
Ericsson
Tel: 00 46 8 757 4700
Enquiry No 511

LCD controller boards

DM Electronics produces two controller boards for character and graphics LCD modules. The DM4-2120 character controller interfaces with character modules based on Hitachi's HD44780. It can be used with popular display formats up to four lines of 40 characters through an RS232 9600 or 19 200baud input device. Features include input buffer, constant current source for LED backlighting, contrast adjustment potentiometer and patch panel to allow LCD connector reconfiguration. The DM4-2140 graphics controller interfaces with graphics modules based on the Toshiba T6963C. It caters for 128 x 64, 128 x 128, 240 x 64 and 240 x 128 display formats through an RS232 9600 or 19,200baud device.
DM Electronics
Tel: 01235 811880
Enquiry No 512

Time up to 3 hours

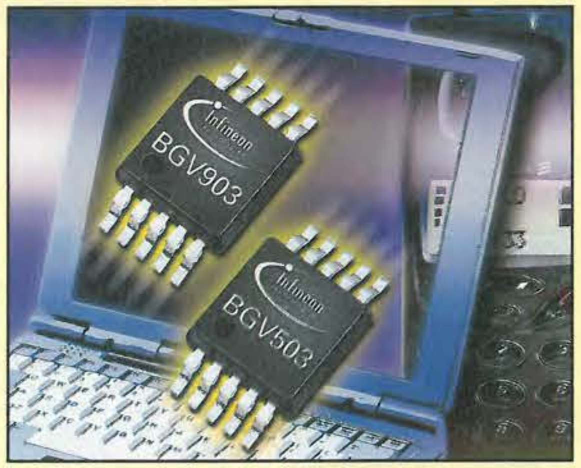
With a three hour timing range, the S1DX timer from Matsushita has four operating modes – power on-delay, flicker, one-shot and one-cycle. Status is shown by colour LEDs on the front face. The timer comes in five relay output and solid-state output versions. Relay output versions cover AC and DC operation, and the solid-state alternatives as transistor and triac types. Accessories include mounting frames, leaf holding clips, protective covers, sockets, caps and DIN terminal sockets in various configurations. This UL and CSA approved timer is suitable for timer control panels, production machinery and sequence operations.
Matsushita
Tel: 01908 231555
Enquiry No 513

Power supply IGBT

Intersil has announced a 600V IGBT for switch-mode power applications requiring off-line power conversion at frequencies of 100kHz or more. It delivers currents to 20A and has a current fall time of 70ns typical at 125°C. They are for use in computers and other switch-mode power applications, including AC-to-DC supplies, DC-to-DC conversion and power factor correction in uninterruptible supplies. The HGTG20N60A4 has a collector-to-emitter saturation voltage of 1.6V typical at 125°C. It is MOS gated.
Intersil
Tel: 01344 350250
Enquiry No 514

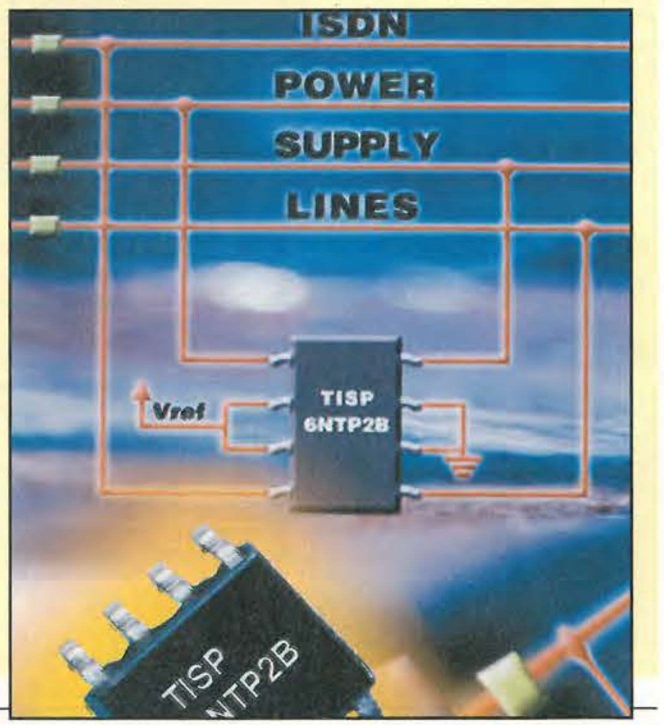
Two-stage voltage inverters

The BGV503 and BGV903 are one and two-stage voltage inverters from Infineon. They convert a positive supply voltage between +2.7 and +5.0V to a typically corresponding negative output voltage of –2.1 to –4.6V and –4.4 to –9.6V, respectively. The BGV503 has an integrated regulator for biasing GaAs FETs and depletion high electron mobility transistors. An integrated oscillator has a typical frequency of 230kHz. The p-p output voltage ripple is typically 25mV at output currents of 3mA (C_{out}=1μF). Shutdown input supply current is less than 5μA between –25 and +100°C. Both come in TSSOP-10-1 packages.
Infineon Technologies
Tel: 00 49 89 234 22767
Enquiry No 110



120V ISDN/Slic protector

Power Innovations has introduced the TISP6NTP2B surface-mount ISDN protector. The 120V maximum rating makes it suitable for providing overvoltage protection on up to four ISDN DC power supply lines or two Slics. It combines four programmable and independently triggered protectors in a SOIC package, with each protector having international protection ratings including 25A ITU-T K20/K21 10/700 and 20A GR-1089-Core 10/1000. The trigger voltage is automatically programmed by reference to the system main supply voltage. In operation, negative surges are initially clamped close to the power supply rail, followed by a fast crowbar action, reducing voltage stress for downstream electronics. Positive surges are clamped to ground by an integrated inverse parallel diode for each line. It comes in tube or tape-and-reel format. Offstate capacitance is 60pF at –50V to ensure transparency in normal use and it has a 150mA holding current. Trigger current is 5mA.
Power Innovations
Tel: 01234 223022
Enquiry No 515



Please quote *Electronics World* when seeking further information

Piezoelectric actuators

Tokin piezoelectric actuators are available from Semicom. Using the piezoelectric longitudinal effect to transform electrical energy into mechanical energy, these devices are suitable for mechatronic systems with movement in increments down to 1μm. The AE resin-coated components deliver a typical clamping stress of 3500N/cm². They can be driven to a third of the self-resonant frequency and have solid-state construction. A metal-sealed variant, the ASB, has also been introduced. Applications include pickup and tracking control of magnetic heads for hard disc drives and positioning of mirrors or prisms in laser systems.
Semicon
Tel: 01279 422224
Enquiry No 516

SVGA colour STN display

Hitachi has announced an SVGA colour STN display for PC and internet browsing. The 25.4cm display has a contrast ratio of 50:1, brightness of 200cd/m² and power

consumption of 2W. It is suitable for point-of-sale, point-of-information and internet terminals. The SX25S003 uses multi-line drivers. The transmissive CSTN display has a resolution of 800 by 600 pixels with 265 000 colours and operates from a +3.3V supply. Response time is 270ms. The CFL is enclosed in an evacuated glass flask, which ensures the heat generated is used to raise the temperature of the fluorescent tube. This means less power is needed to operate the display and backlighting.
Hitachi
Tel: 01628 585163
Enquiry No 517

Security processor

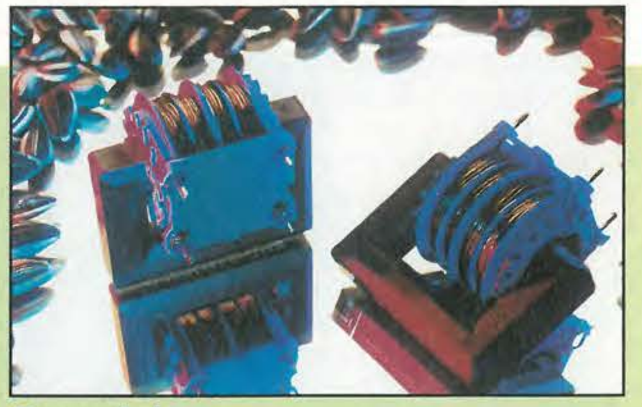
Philips has introduced the VMS747 single-chip security processor that can handle higher than 256Mbit/s Ipsec processing (3DES and MD5/SHA-1 hashing). It is based on transportable IP blocks that are part of the company's Velocity RSP7 security tool set. Applications include virtual private networks for OC3 and higher and cryptographic security in financial servers and point-of-sale terminals. It has an integrated Arm7 32-bit Risc microprocessor subsystem. Security features include a dual-state processor architecture, encrypted program code execution, anti-spoofing technology and an integrated Ipsec accelerator. The processor integrates bulk data encryption, hashing, public key acceleration, random number generation and battery backed memory. A software development kit will be available in the spring.
Philips Semiconductor
Tel: 00 31 40 272 2091
Enquiry No 518

Inverse multiplexer

The Mitel MT90220 is an eight-port ATM physical layer inverse multiplexer providing a multi-vendor PCM interface and is for systems that implement ATM access over existing trunk interfaces, such as central office access concentrators and edge switches. It lets users set up multiple ATM multiplexing groups and can handle eight T1 and E1 interfaces. In mixed mode, T1 and E1 lines can be programmed to operate as inverse multiplexed or Uni lines for ATM over T1 and E1 IMA and Uni services operating simultaneously. It can be used for asymmetrical applications, such as internet access, where more bandwidth is required in one direction. The ability to route clocks from T1 and E1 lines allows the use of various phase lock loops. Supplied in a 208-pin PQFP, it provides a Utopia level two phy interface that lets up to four devices be cascaded, delivering up to 32 T1 and E1 interfaces.
Dip International
Tel: 01223 462244
Enquiry No 519

Enclosures

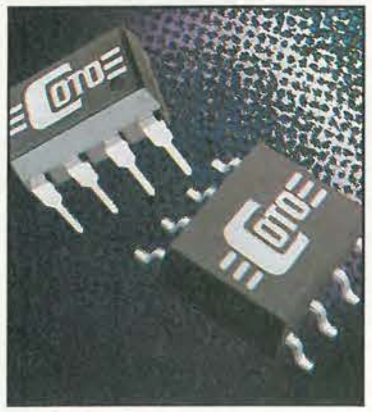
APW has introduced the VEcase 700 EMC-screened desktop, 19in rack-mount and tower enclosures for CPCI and VMEbus systems, with applications in networking, traffic, transportation and industrial control. There are 18 versions – 3 and 6U heights in 42, 60 and 84HP widths and depths of 240, 280 and 340mm. They have an extrusion-based design. For CPCI systems, the front tiebars are compatible with IEEE1101.10, letting the injector and extractor handles operate correctly. For CPCI computer telephony, versions compatible with the IEEE1101.10 rear-plug module are available. The deepest 340mm standard unit will accept two 160mm deep cards back-to-back in the same slot position. For VME and general-purpose use, IEC297-3 tie bars let cards be mounted vertically or horizontally. RFI attenuation is typically greater than 45dB at 100MHz for the standard version. A further 10dB attenuation can be gained by fitting an optional EMC upgrade kit. Thermal management options include rear panels fitted with extractor fans, ventilated covers and sides and, in the extended height units, slim-line fans in the base below the active card area. The units are stackable.
APW Electronics
Tel: 01489 780078
Enquiry No 520



D-core choke

Epcos has introduced the B82731-M D core choke, a current-compensated double choke for interference suppression in electronic power supplies for lighting ballasts and TV sets. Inductance is between 3.3 and 47mH. Pin grid size is 12.5 by 10.0 ±0.2mm. Pin size is 0.6mm. With rated voltages of 250V AC or DC, the EMC component covers rated currents from 0.5 to 1.8A. Weight is about 8g and the device incorporates a closed rectangular ferrite core and a closed polycarbonate coil former with four sections. The coil former is flame-retardant to UL94V-0. The choke is made without encapsulation or adhesive materials.
Epcos
Tel: 01344 396689
Enquiry No 521

Please quote *Electronics World* when seeking further information



SM relay

Rhopoint Components has introduced a solid-state relay in a surface mount package for modems and data communications applications. The G2-DA06 is a dual-form A device with 18Ω maximum on resistance – 10Ω typically – that can handle continuous load current of 180mA and peak load voltages of 250V. Input to output isolation is 3750V RMS. Maximum height is 0.22cm. No EMI or RFI is generated by the relay and its solid state construction eliminates arcing and contact bounce. No snubbing circuits are needed. UL and CSA recognised, the relay is certificated to BABT and VDE.
Rhopoint Components
Tel: 01883 717988
Enquiry No 526

PCB applicator

Donprint has developed a device for printing and applying PCB labels. The Apply-Mark 610 uses the Zebra 90xill printer with 600dot/in resolution. It can be linked or retrofitted to most production line equipment such as conveyors. It can also be combined with the firm's barcode software, data capture products and customised programming services for work in progress tracking. Cycle times are less than 3s and accuracy repeatability 0.5mm. The firm is developing three and four-axis applicators for multiple label pick-and-place application in one pass.
Donprint
Tel: 01355 249191
Enquiry No 524

Flash for GPS

Unique Memec is sampling three silicon storage x16 flash products. The 8Mbit 39VF800A, 4Mbit 39VF400A and 2Mbit 39VF200A have read access times of 45ns and use a 2.7 to 3.6V voltage supply for read, program and erase. They are suitable for program code storage in GPS navigation systems, MP3 players, PDAs, DSL, DVD, CD-RW, modems and disk drives. They use a uniform sector architecture of 4kbyte or 2kword and have a uniform block erase capability of 32kword or 64kbyte. Erase speed is 20μs per word for programming, 25ms for sector and block erase and 100ms for chip erase. They come in 48-pin TSOP and 48-ball TFBGA packages.
Unique Memec
Tel: 01296 397396
Enquiry No 522

Application specific tools

Engineers and scientists can create measurement applications in Microsoft Visual C++, Visual Basic, and Labwindows/CVI using Measurement Studio from National Instruments. The set of integrated measurement tools includes acquisition, analysis and visualisation for text-based programming languages. The package includes user interface controls, mathematical analysis, signal processing routines and high-level data acquisition and instrument control I/O libraries for Visual C++ and Visual Basic. It integrates tools for data acquisition, data analysis and data visualisation directly into Visual C++.
National Instruments
Tel: 01635 523545
Enquiry No 523

Instrument catalogue

A new test and measurement catalogue is available from Professional Instrument Distributors. 100 pages describe equipment for electrical, electronic, safety and environmental testing. Recorders, data loggers and analysers are among the 1000 items in the catalogue. There are five regional sources of the catalogue covering the whole of the UK. Ask for your regional outlet via the number below.
Professional Instrument Distributors
Tel: 01756 799737
Enquiry No 527

USB protector

A dual-channel, high-side switched LSI IC measuring 5.0 by 6.2 by 1.5mm provides power system protection for equipment with a built-in USB hub. Rohm's BD6510F is for PCs, scanners, monitors, printers, disc drives and cameras. Features include under-voltage lockout, overcurrent protection and thermal shutdown circuitry that incorporates an output latch to ensure complete power shutdown. Typical on resistance is 100mΩ and it can operate with voltages from 3.0 to 5.5V making it suitable for notebook computers and other applications with mixed 3.3 and 5V power circuits.
Rohm Electronics
Tel: 01908 282666
Enquiry No 528

24-port 1U patch panels

Series 500 ITT Networks 24-port 1U patch panels from AX Distribution provide 96 outlets in 4U cabinet space. They use Link 5E (cat 5E) RJ45 data jacks and come with shielded or unshielded 808 connectors, cable ties and fixing kits. The panels can be mounted into the rack or cabinet first, letting the terminated 808 mark-two connectors be individually located and secured. Also available are cord management

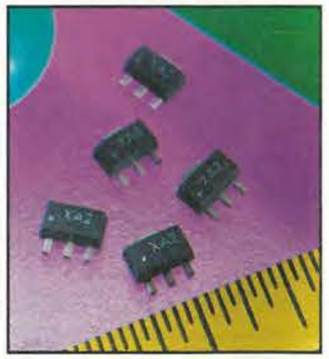
panels and clips for routeing patch cords within cabinets
AX Distribution
Enquiry No 529

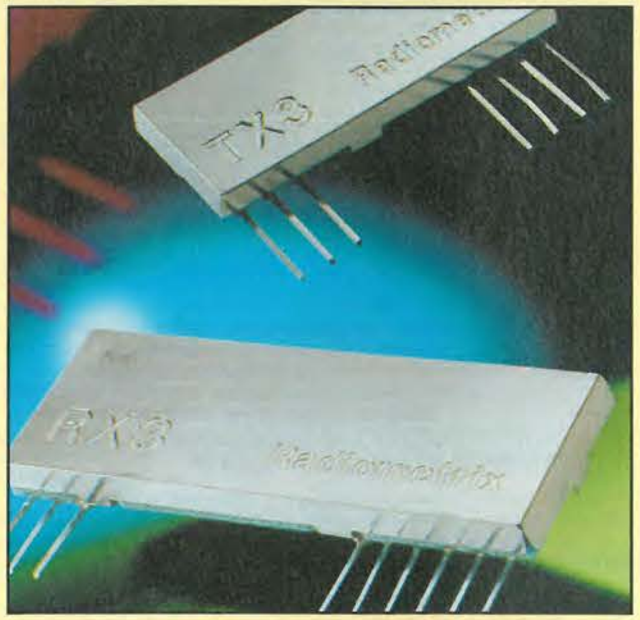
Electrolytic capacitors

The GY electrolytic capacitors from Nichicon are for industrial power supplies in ballast applications. Working voltage is 160 to 250V (–40 to +105°C) and 315 to 450V (–25 to +105°C) and capacitance 39 to 2200μF. Tolerance is ±20 per cent. Ripple current covers 0.65 to 2.9A RMS. They have snap-in terminals, diameters of 22, 25, 30 and 35mm and lengths from 25 to 50mm.
Nichicon
Tel: 01276 685393
Enquiry No 530

GaAs amplifier

The SXA-289 monolithic GaAs power amplifier from Link Microtek is for use as a driver stage in base station equipment for cellular, ISM and narrowband PCS applications. Made by Stanford Microdevices, it uses Darlington pair topology to provide broadband performance from 50MHz to 2GHz, while drawing 110mA from a +5V DC supply. Typical third-order intercept point is 42dBm across the full frequency





FM transmitter and receiver

For the 868MHz band, Low Power Radio Solutions has introduced the Radiometrix FM TX3 transmitter and RX3 receiver. These modules operate at 869.85MHz. Internal filtering gives the receiver selectivity, providing immunity against the introduction of Tetra systems at 870 to 876MHz. Data rate is up to 64kbit/s. The ETS300-220 compliant transmitter provides a 1mW power output, which, with the receiver's –100dBm sensitivity, provides a line-of-sight range up to 200m or around 50m in buildings. Both units will operate from a lithium cell supply. Minimum voltage requirements are 2.2 and 2.7V respectively for the transmitter and receiver, making them suitable for portable, battery-powered applications such as hand-held terminals. They are screened to ensure EMC compliance. They suit products for European and North American markets. Pin compatible versions are available operating at 916MHz and conforming to FCC part 15.
Low Power Radio Solutions
Tel: 01296 397396
Enquiry No 525

NEW PRODUCTS

range. Noise is 5dB in the 800 to 900MHz and 1800 to 2000MHz bands. Power gain is typically 20dB in the cellular band and more than 14dB in the PCS band. Fabricated using GaAs heterojunction bipolar transistor technology, it is housed in a SOT-89 surface-mount plastic package. Evaluation boards, development kits and application circuits are available.
Link Microtek
Tel: 01256 356771
Enquiry No 531

66W switching regulator

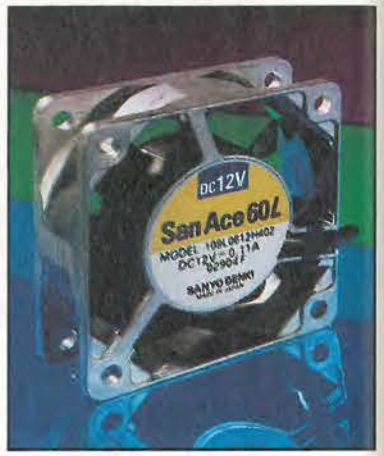
UR has available the STR-G6651 offline quasi-resonant flyback switching regulator from Allegro Sanken. The switch-mode power-supply regulator is for industrial, telecoms, datacomms, white goods, lighting, metering, charger and security applications. Capable of handling up to 66W, the device comes in a five-pin over-moulded isolated TO-220 package and incorporates the primary control and drive circuit with a discrete avalanche rated power MOSFET. The MOSFET has a Vds of 650V and is constant voltage soft gate driven. It can be used at switching frequencies of 20 to 300kHz and has built in cycle by cycle programmable current protection, latched overvoltage protection, latched thermal shutdown, internally programmed soft start and under voltage lock out.
Ultimate Renaissance
Tel: 01793 439310
Enquiry No 532

3U Eurocard controller

HM Computing has launched a 3U Eurocard combining a Pentium processor with functions and interfaces for real-time embedded systems. The card may be expanded using CompactPCI or PC/104 implemented as a mezzanine bus. It provides a foundation for controlling machinery and instrumentation and will run any PC operating system, as well as real-time operating systems including QNX, PSOS, VxWorks, Linux and Pharlap. The CPU-7310 card uses a Pentium or K6 processor operating at up to 400MHz. Also onboard is 64Mbit DRAM, 0.5Mbit synchronous L2 cache, 1Mbyte solid-state disk, watchdog and real-time clock and floppy disk and keyboard controllers.
HM Computing
Tel: 01299 250997
Enquiry No 533

60mm square fan

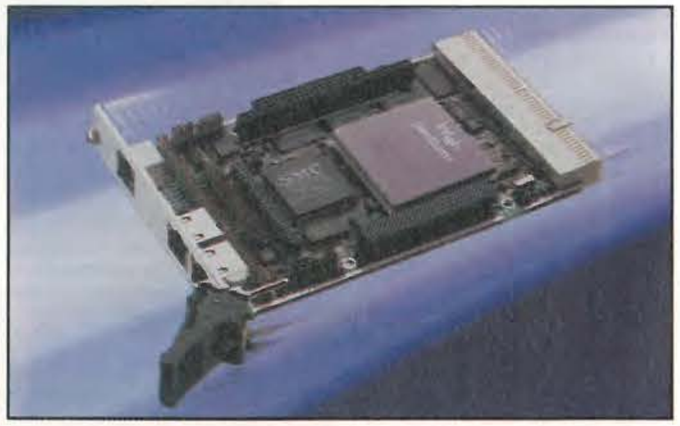
The San Ace 60L fan is available from EAO. The fan measures 60mm diameter by 25mm thick, and has a claimed operating life of 100 000 hours at 60°C ambient temperature. The unit is available in 5, 12 and 24V versions. Nominal acoustic noise level is 28dB(A). Alarm functions are available for remote



monitoring, and locked-rotor protection is provided by an overcurrent limiting circuit.
EAO
Tel: 01444 236000
Enquiry No 534

Configurable memory device

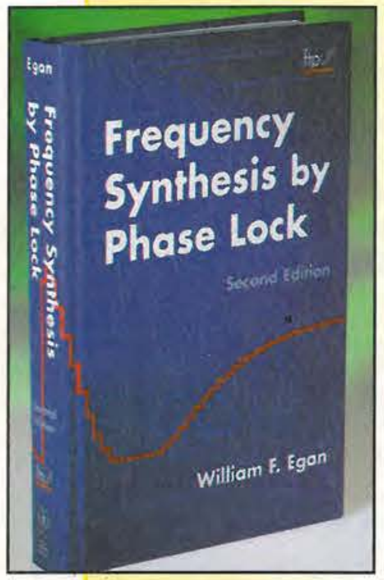
A wafer-scale integration Easyflash programmable system device from Insight Memec provides in-system programmability of blank devices via a four-pin ISP port, based on the JTag standard, which lets code bypass the MCU blank and flow directly into the device's memory banks. The WSI PSD9xx can be soldered onto the circuit board and programmed at the end of the assembly line. The device provides concurrent memory arrays to support in-application



BOOK TO BUY

Frequency Synthesis by Phase Lock

Frequency synthesis is an important element in the design of all communications equipment, but has taken on new life recently with the advent of new hand-held wireless devices. This technology not only allows wireless transmitters to change frequencies quickly, but also gives high reliability and security in transmissions. Thus, mobile devices such as cell phones can utilise this technology to change frequencies until a suitable one is found for the location in which it is being used.



- Emphasises the fundamentals of frequency synthesis
- Based on a course that Dr. Egan has been teaching for over 20 years at Santa Clara University
- Provides a link to the Wiley ftp site for the use of associated MATLAB exercises
- Taken together with Phase Lock Basics by the same author, the two books provide readers with complete coverage of the field.

CONTENTS Introduction; The Elementary Phase-Locked Synthesiser; Modulation, Sidebands and Noise Spectrums; Frequency Dividers; Phase Detectors; Higher Order Loops; Sampling Effects; Architectures; Large-Signal Performance, Natural Acquisition; Acquisition Aids; Spectral Purity; Computer Aided Engineering.

UK Price: £64.00 Europe £66.50 ROW £69.00

** Price includes delivery and package **

Return to Jackie Lowe, Room L333, Quadrant House, The Quadrant, Sutton, Surrey, SM2 5AS

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NEW PRODUCTS

re-programmability, which allows execution out of one memory array while the other is being updated. It integrates up to 2Mbit of flash, 256kbit of concurrent flash or EEPROM, up to 64kbit of SRAM, additional i/o, power management and security. PSDsoft Express software lets external memory and accompanying logic be added to the designs without logic design effort or expertise. It is programmable via the ISP port, the MCU's serial port or a third party device programmer.
Insight Memec
Tel: 01296 330061
Enquiry No 535

Piezo speakers

Megacera's piezo speakers are available from Deltron Roxburgh. The speakers house a 50μm thick ceramic resonator in a stainless package. The driving voltage is less than 6V. They have an operating frequency band up to 20kHz. The largest model, the MS-4048U, measures 40 by 48 by 2.5mm and develops 85dB (1m) sound pressure. The smallest measures 19 by 23 by 2.5mm, weighs 2.3g and achieves a sound pressure of 76dB (1m).
Deltron Roxburgh Electronics
Tel: 01724 281770
Enquiry No 538

Rad-hardened MOSFET

Intersil has introduced a radiation hardened MOSFET for protecting against single-effect effects in satellite and aerospace applications. The 60V p-channel MOSFET delivers power conditioning and control in satellite applications, and has a typical on resistance of 16mΩ. The device is suitable for applications exposed to radiation environments, such as switching regulation, switching converters, motor drives, relay drivers and drivers for high-power bipolar switching transistors. The FST9055 acts as a switch to control or supply power to an entire system or to a single object such as a solenoid or motor. The device is an enhancement-mode silicon-gate power field effect transistor with a vertical double-diffused metal-oxide semiconductor structure.
Insight Intersil
Tel: 01344 350250
Enquiry No 537

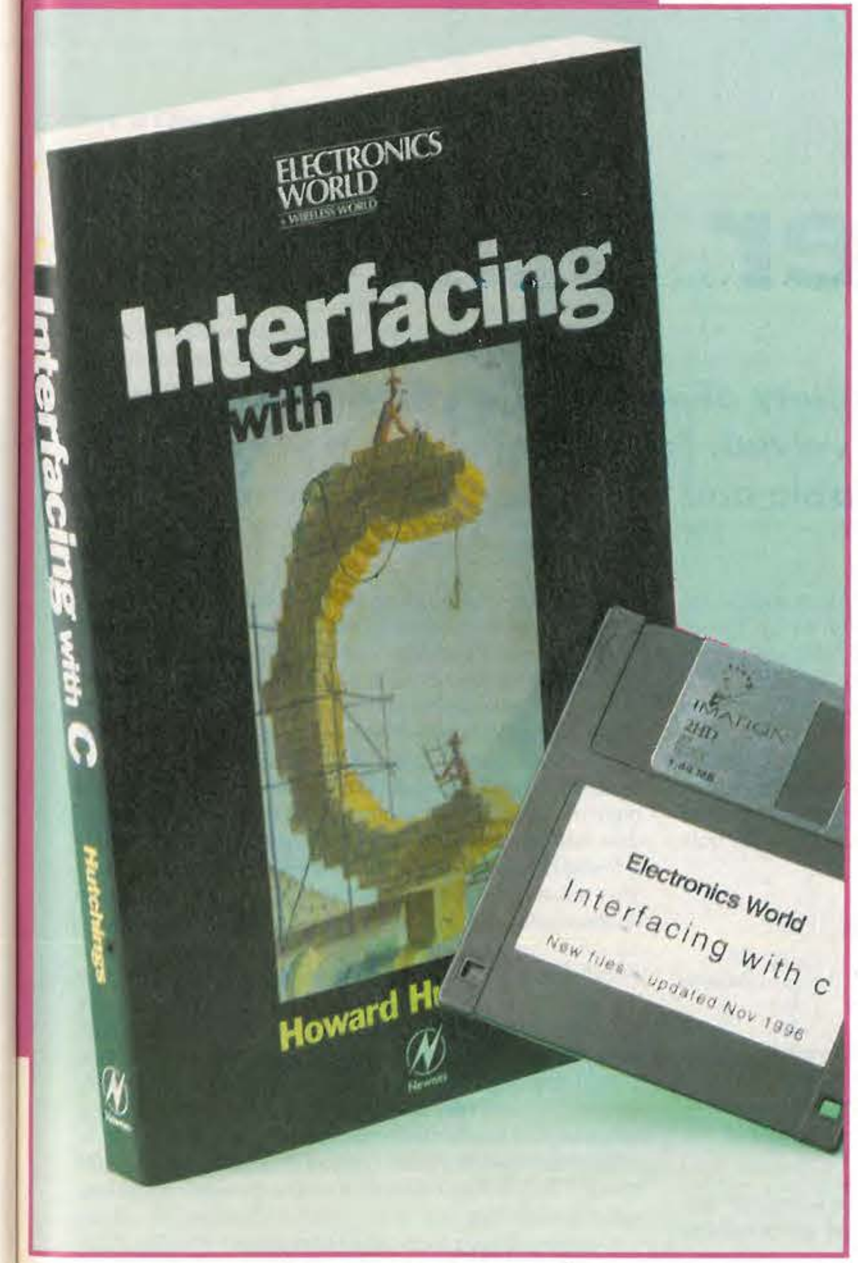
LCD controller with SRAM

The SED1375 from Epson is a colour or monochrome LCD graphics controller with an embedded 80kbyte SRAM display buffer. Applications include office automation equipment, mobile communications devices and palm-sized PCs. Features include hardware portrait mode and virtual and split screen.
It is aimed at Windows CE applications but works with other CPUs and operating systems. Resolution is 640 by 480 at a colour depth of 2bits/pixel, 640 by 240 at 4bits/pixel or 320 by 240 at 8bits/pixel.
Epson Electronics
Tel: 00 49 89 14005 363
Enquiry No 539

DC-to-DC converter with 12V inputs

Ginsbury has launched a dc-to-dc converter with 12V inputs for battery power architectures in portable equipment. The Powercube TD is available in half-brick 30, 50, 75, 100 and 150W units and full-brick 250 and 300W devices. It meets standards for distributed and board mounting power systems in telecoms and server applications. Features include over-temperature shutdown, overvoltage protection, power-good signal output, built-in current monitor, case ground pin and positive or negative logic shutdown (remote on-off). Other inputs are 300, 150, 48 and 24V.
Ginsbury Electronics
Tel: 01634 298900
Enquiry No 536

BOOK TO BUY



Without an engineering degree, a pile of money, or an infinite amount of time, the revised 289-page Interfacing with C is worth serious consideration by anyone interested in controlling equipment via the PC. Featuring extra chapters on Z transforms, audio processing and standard programming structures, the new Interfacing with C will be especialy useful to students and engineers interested in ports, transducer interfacing, analogue-to-digital conversion, convolution, digital filters, Fourier transforms and Kalman filtering. Full of tried and tested interfacing routines.
Price £14.99.

Listings on disk – over 50k of C source code dedicated to interfacing. This 3.5in PC format disk includes all the listings mentioned in the book Interfacing with C. Note that this is an upgraded disk containing the original Interfacing with C routines rewritten for Turbo C++ Ver.3
Price £15, or £7.50 when purchased with the above book.

Especialy useful for students, the original Interfacing with C, written for Microsoft C Version 5.1, is still available at the price of £7.50. Phone 0181 652 3614 for bulk purchase price.

Post your completed order form to:-
Jackie Lowe, Electronics World, Room L333, Quadrant House, The Quadrant, Sutton, Surrey, SM2 5AS

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		Total	£.......

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Fax:__________ Date__________
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IF filter options

Over the years, a variety of intermediate-frequency filter arrangements has evolved. Joe Carr looks at a variety of the techniques available and discusses their merits.

A number of different types of IF filters are used in radio receivers today. In addition to *LC* filters, there are various types of crystal filter, monolithic ceramic filters, and mechanical filters to consider. In this article, I take a look at the various types of IF filter, their characteristics and their applications.

Before delving into the topic, it is helpful to consider how these filters are used. **Figure 1** shows a generic IF amplifier with the filters in place. The IF amplifier provides most of the gain in a superheterodyne receiver, as well as the bulk of the selectivity of the receiver.

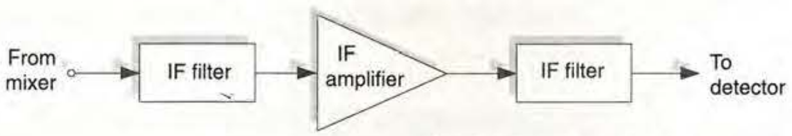


Fig. 1. Intermediate-frequency filtering arrangement, as used in a typical superheterodyne receiver.

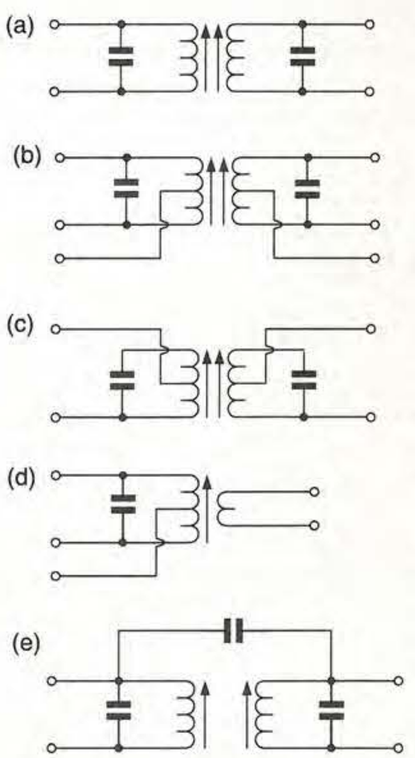


Fig. 2. Various LC IF filter circuits. Filters a) to c) are all tuned on either side of the transformer and vary primarily in drive impedance. Filter a) used to be the conventional IF filter, but it is no longer widely used. Filter b) has low-impedance taps on both sides to increase its usefulness in solid-state circuits; filter c) is identical, except that the high-impedance taps are no longer available to the outside world. Filter d) is single-tuned and its low-impedance output is use for impedance matching. In filter e), the filter coils are not magnetically coupled. Instead, the signal is transferred by common reactance. Here, the reactance is a capacitor, but it could equally well be an inductor.

Selectivity is the function of the IF filter. It has the narrowest bandpass of all the filters in the receiver. Its main function is to supply the selectivity of the receiver by narrowing the bandpass down to accommodate the type of modulation being used.

The IF amplifier may or may not use two – or more – filters, depending on the type of design. In cases where only one filter is used, then the filter will usually be placed at the input of the amplifier in order to eliminate the mixer products that can affect the IF amplifier performance.

Noise produced by the IF amplifier can be significant. This means that an output IF filter is indicated in order to eliminate that noise.

LC IF filters

The basic type of filter, and once the most common, is the *LC* filter, which comes in various types **Fig. 2**. The type shown in Fig. **2a)** contains two parallel tuned *LC* sections. Although it is not apparent here, the input and output sides of the *LC* network need not be the same impedance, although that is usually the case.

Once the most common form of IF amplifier filter, this type has largely been eclipsed by other types, except in certain IC amplifiers.

A more common form today is shown in Fig. **2b)**. This form has a low impedance tap for transistor or IC applications, with the high impedance portions still available. In Fig. **2c)** is a common form of IF filter in which the low impedance tap is available to both input and output sides, but one side of the high impedance portions of the transformer is not.

In Fig. **2d)** you can see a single-tuned IF filter. It has a standard IF filter input side, but it has only a low impedance link on the output side. The IF filtering is performed by the tuned *LC* circuit, whereas the low impedance link is for impedance matching.

In Fig. **2e)** you can see a double-tuned IF filter that is not magnetically coupled (so it is not a transformer), but rather is coupled through a common impedance. In this case a small value capacitor is used as the common impedance, but inductors can also be used as well.

Crystal filters

The quartz piezoelectric crystal resonator is ideal for IF filtering because it offers high Q – resulting in narrow bandwidth – and it behaves as an *LC* circuit. Because of this feature, it can be used for high quality receiver design as well as filter type single-sideband transmitters.

Quartz has a dielectric constant of about 3.8. A schematic symbol for a crystal is shown in **Fig. 3a)**, while the equivalent *LCR* circuit and the device's impedance curve are shown in Figs **3b,c)**. The equivalent circuit shows that there is a series inductance, L_s, and capacitance, C_s, as well as a series resistance R_s.

Series inductance and capacitance of a crystal are sometimes called the motional reactance values. There is also a parallel capacitance, C_p.

The parallel capacitance resonates with the inductor to form a parallel resonance, while the series capacitance resonates with the inductor to form a series resonance.

Figure **3c)** shows capacitive and inductive reactance against frequency. The frequency marked F_s is the series resonance point. The anti-resonance is 'officially' the parallel resonance, but in practical terms there is a range of parallel resonance.

Figure 4 shows a typical crystal package and its mounting on a printed circuit board. The basic package is shown in Fig. **4a)**. Note that the pins might be actual pins, or they may be wires. A cut-away view is shown in Fig. **4b)**. This gives you a better idea of how the crystal works.

Figure **4c)**. shows how the crystal is usually mounted on a double-sided printed circuit board. An insulator is placed on the board to prevent short circuiting the board tracks with the crystal package.

Figure 5a) shows a simple crystal filter that has been around since the 1930s in one form or another. Figure **5b)** shows the attenuation graph for this filter.

There is a 'crystal phasing' capacitor, adjustable from the front panel, that cancels the parallel capacitance. This cancels the parallel resonance, leaving the series resonance of the crystal.

Although the 1930s vintage filters did not use it, this filter is built in trifilar form. This means that the windings of T_1 are wound together, interlaced with each other.

Figure 6a) shows the circuit for a half-lattice crystal filter, while Fig. **6b)** shows its attenuation curve. This type of crystal filter is used in lower cost radios. Like the simple crystal filter described above, this version uses a trifilar coil for T_1. But instead of the phasing capacitor there is a second crystal in the circuit.

The frequency relationship between the two crystals, Fig. 6b) shows that they have overlapping parallel and series resonance points. This makes the parallel resonance of crystal No 1 is the same as the series resonance of crystal No 2.

You can use the half-lattice filter to build a cascade half-lattice filter, **Fig. 7**, and a full-lattice crystal filter, **Fig. 8**. The cascade half-lattice filter has increased skirt selectivity. It also gives fewer spurious responses compared with the same passband in the half lattice type of filter. It is a back-to-back arrangement on a bifilar transformer, T_1. In practice, close matching is needed to make the cascaded half lattice filter work properly.

The full lattice crystal filter of Fig. 8 uses four crystals like the cascade half-lattice, but the circuit is built on a different basis. It uses two tuned transformers, T_1 and T_2, with the two pairs of crystals that are cross-connected across the tuned sections of the transformers. Crystals Y_1 and Y_3 are of one frequency, while Y_2 and Y_4 are the other frequency in the pair.

The principal advantage of the full-lattice filter is the use of two different frequencies instead of one frequency for the crystals. It is more difficult to match four crystals than two.

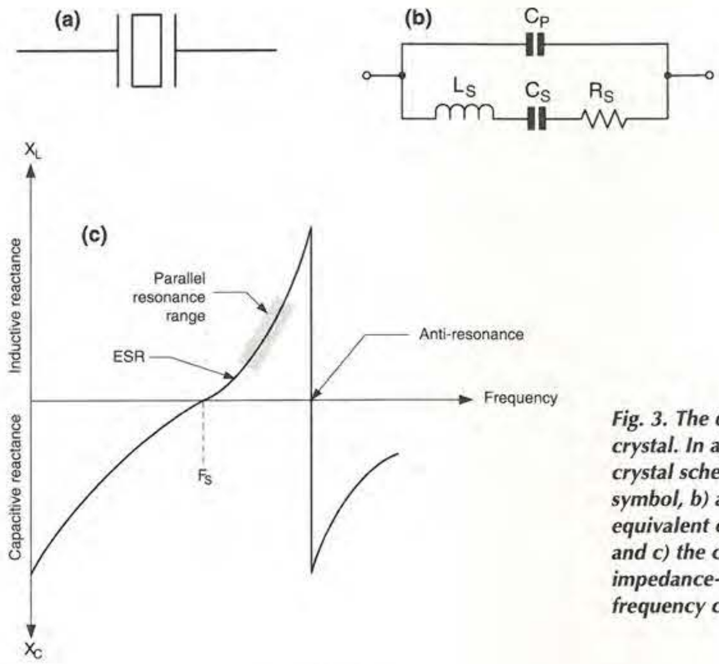


Fig. 3. The quartz crystal. In a) is a crystal schematic symbol, b) an equivalent circuit and c) the crystal's impedance-versus-frequency curve.

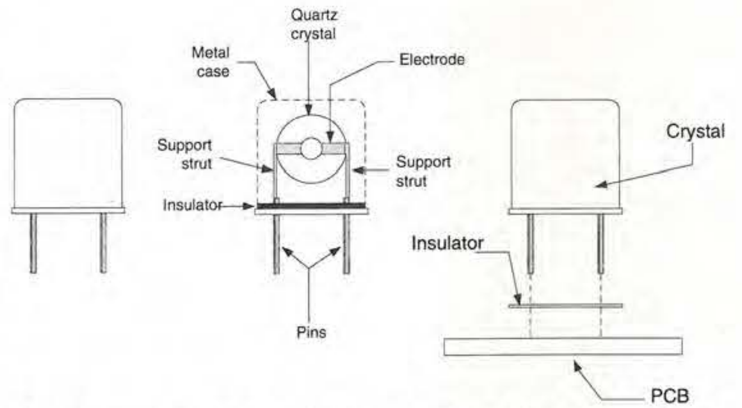


Fig. 4a) Crystal package; b) in cut-away form showing inside; c) Mounting on a printed wiring board.

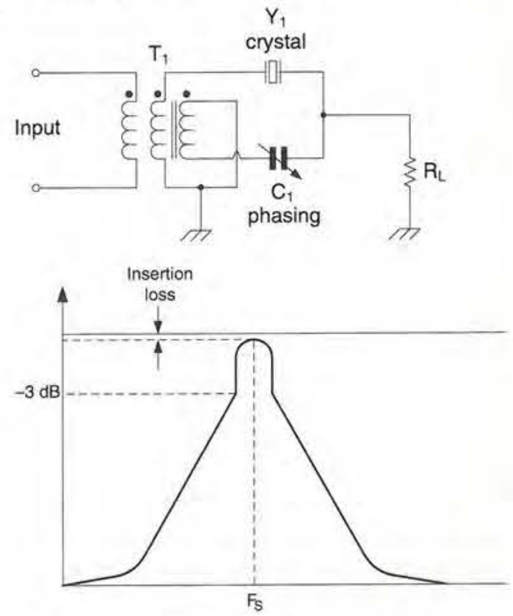


Fig. 5a). Simple crystal intermediate filter and b), its resonance curve. The transformer is trifilar. Capacitor C_1 is for cancelling parallel capacitance.

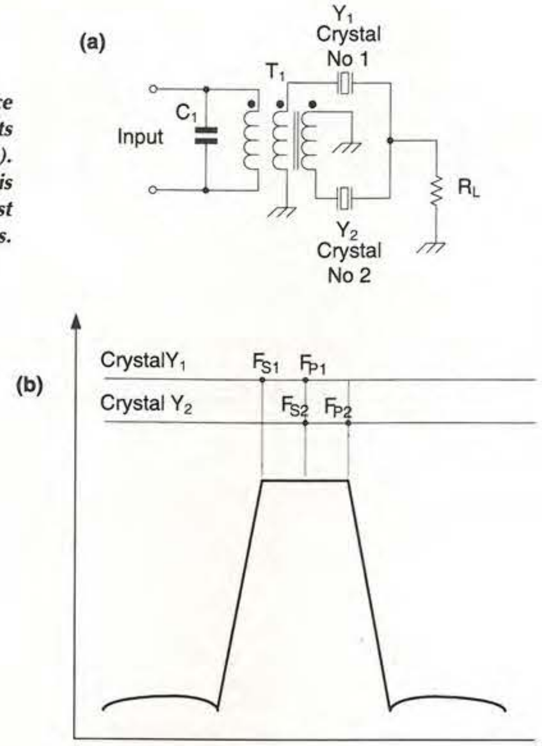


Fig. 6. Half-lattice crystal filter, a), and its resonance curve, b). This type of filter is used in low-cost radios.

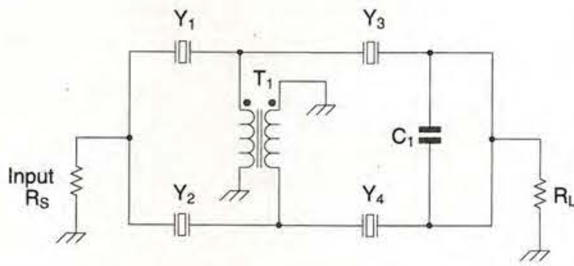


Fig. 7. Cascade half-lattice crystal filter has increased skirt selectivity over the standard half lattice.

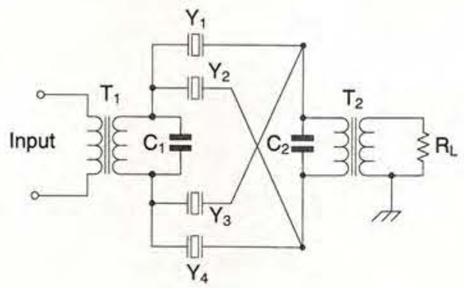
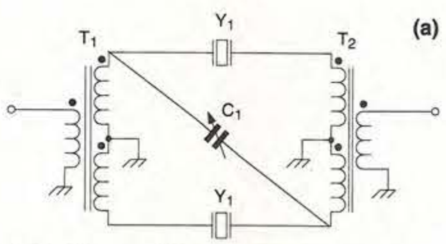


Fig. 8. Full lattice crystal filter involves two tuned transformers.



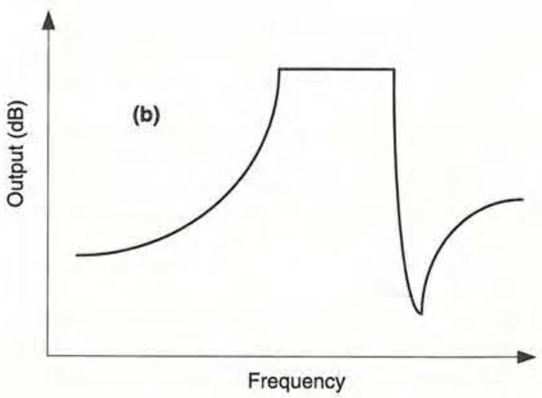


Fig. 9a). This type of filter, a), uses two crystals of the same frequency and results in a pass curve with asymmetrical slopes, as is evident in resonance curve b).

A different sort of filter is shown in **Fig. 9a)**, with its asymmetrical attenuation curve shown in Fig. **9b)**. This filter has a more gradual fall-off on one side than on the other, Fig. 9b).

The filter has the advantage that the frequencies of crystals Y_1 and Y_2 are the same frequency. Capacitor C_1 is included in the circuit to allow tuning of the desired pass band. The bandwidth of this filter is only half what is expected from the half-lattice crystal filter above.

Crystal ladder filters

Figure 10 shows a crystal ladder filter. This filter has several advantages over the other types.

- All crystals are the same frequency – no matching is required.
- Filters may be constructed using an odd or even number of crystals.
- Spurious responses are not harmful – especially for filters over four or more sections.
- Insertion loss is very low.

Both Butterworth and the equi-ripple or Chebyshev responses can be created using this design. Ideally, in Chebyshev designs the number of positive peaks should be the same as the number of crystals. Response should also be of equal amplitude over the passband of the filter.

In reality, fewer peaks than that are found, some being merged with each other. Additionally, the amplitude of the ripple increases towards the edges of the band.

Designing this filter can be simplified by using a test fixture to dope out the problem first. The value of the end capacitors is,

$$C_{END} = \left[\frac{1.59\times10^5}{R_s F_o}\right] \times \left[\sqrt{\frac{R_s}{R_{END}} - 1}\right] - 5$$

The value of the coupling capacitors is,

$$C_{JK} = 1326\left[\frac{\Delta f}{B k_{JK} F_o}\right] - 10$$

And the value of R_{END} is,

$$R_{END} = \left[\frac{120B}{q\Delta f}\right] - R_S$$

Where:

B is the bandwidth in hertz.
C_{END} is the end capacitance in picofarads.
C_{JK} are the shunt capacitors in picofarads.
F_O is the crystal centre frequency.
Δf is the bandwidth measured in a test fixture.
k_{JK} is the normalised values given in tables 1 and 2.
q is normalised end section Q given in tables 1 and 2.
R_S is the end termination of the filter, $R_S > R_{END}$.
R_{END} is the end termination to be used without matching capacitors.

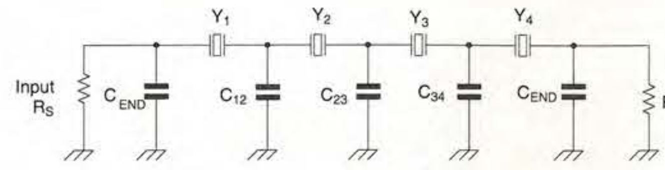


Fig. 10. Crystal ladder filter has a number of advantages over the other types, one of them being that its spurii are not problematical.

A special version of the crystal ladder filter is the Cohn filter or 'minimum-loss' filter of **Fig. 11**.

This filter rotates the end capacitors, and makes the shunt capacitors equal value. It preserves a reasonable shape factor, while minimising loss when built with practical resonators.

Like the crystal ladder filter, the Cohn filter uses same frequency crystals throughout. The error in frequency between the crystals, ΔF_O, should less than 10% of the desired bandwidth of the filter.

The design procedure given by Hayward (1987) is simplified:

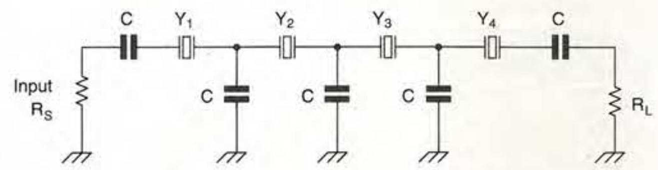


Fig. 11. Cohn ladder filter, also known as minimum-loss filter.

- Pick a crystal frequency in the range 2 to 12MHz. Hayward used 3.579MHz colour burst crystals.
- Pick a capacitance for the filter. A good start is 200pF – higher capacitance yields narrower bandwidth but higher insertion loss.
- Vary the end termination impedance to obtain a ripple-free pass band while providing sufficient stop band attenuation.

Table 3 gives various Cohn filter bandwidths, termination impedances and capacitor values for a three-crystal filter.

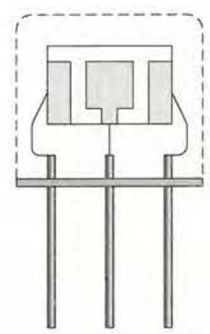


Fig. 12. Outline of a monolithic crystal filter's construction.

Table 3. Cohn three-crystal filter.

Bandwidth	C (pF)	R_{END} (Ω)
380	200	150
600	130	238
1000	70	431
1800	30	1500
2500	17	3300

Monolithic ceramic crystal filters

Figure 12 represents a monolithic crystal filter. The principal benefit of this type of filter is its price, which is low enough to allow its use in low-cost consumer radios. Only a few monolithic filters have the shape factor needed for high-performance receivers though.

These filters are often made with synthetic piezoelectric resonators, rather than quartz. They are made in small packages, with some being made in crystal packages and some being made in special packages. Some of the special packages are smaller than crystals, and some are larger.

Mechanical filters

Considerable improvement in filter action is possible with the use of the mechanical filter. These filters were once used in Collins high-end radio receivers and SSB transmitters. They are now more widespread, although the Rockwell/Collins company still makes the filters.

The basic principle of operation is the phenomenon of magnetostriction, that is the length or circumference of a piece of material will change when it is magnetised. Nickel changes in this way when magnetised, but only by about one part in about 20 000.

Other materials, such as ferrites or powdered-irons (types 61, Q1 or 4C4), provide much stronger magnetostriction effects. In addition, these materials have a high electrical resistivity, so eddy current losses are minimised, and they

Table 1. Normalised k and q values for Butterworth response.

N	q	k_{12}	k_{23}	k_{34}	k_{45}
2	1.414	0.7071	–	–	–
3	1.0	0.7071	0.7071	–	–
4	0.785	0.8409	0.4512	0.8409	–
5	0.618	1.000	0.5559	0.5559	1.0

Table 2. Normalised k and q values for Chebyshev response.

N	q	k_{12}	k_{23}	k_{34}	k_{45}
2	1.638	1.6362	0.7016	–	–
3	1.433	0.6618	0.6618	–	–
4	1.345	0.685	0.5421	0.685	–
5	1.301	0.7028	0.5355	0.5355	0.7028

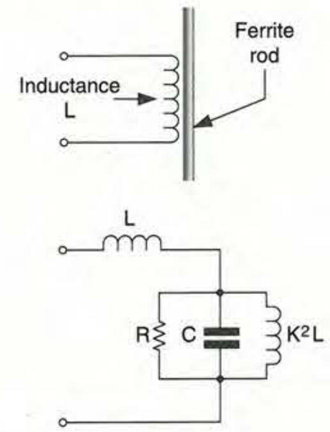


Fig. 13. Mechanical resonator, a), and its equivalent circuit, b). The coil is wound on the rod in such a way that it is free to slip, allowing the rod to expand and contract when subjected to a magnetic field.

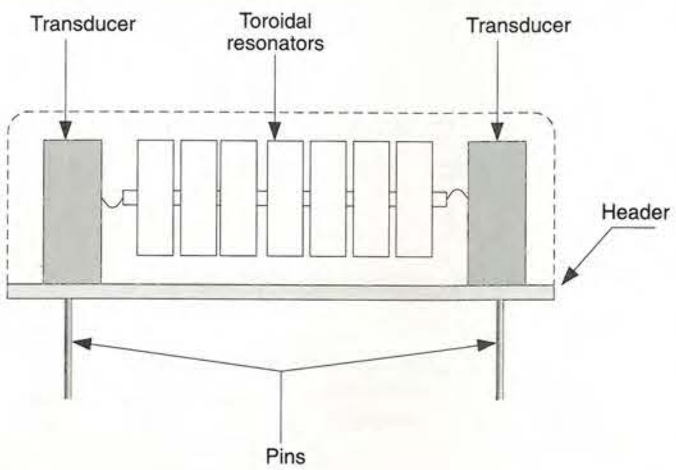


Fig. 14. Mechanical filter structure built using toroidal resonators.

have a mechanical Q on the order of several thousand. They make far better filters than nickel!

The typical ferrite material is formed at 1300 to 1400 degrees Celsius, and has a Q determined by the proportions of oxides used in the formation of the ferrite material.

Figure 13a) shows a magnetostrictive resonator, while Fig. **13b)** shows the equivalent circuit. The ferrite rod is wound with a coil such that it is a slip fit. It is biased magnetically with either a permanent magnet or a DC component to the electrical signal applied to coil *L*.

When alternating current flows in the coil, it adds to or subtracts from the magnetic field of the bias, causing the ferrite length to oscillate. But with the parallel resonant component, Fig. 13b), there's a sharp peak in the response at a frequency equal to the mechanical resonance frequency. When the ferrite rod is wound into a toroidal shape the circumference, hence the radius, of the toroid shape varies as it is magnetised.

Figure 14 shows a mechanical filter built using toroidal resonators. Various mechanical filters are available in frequencies between 60kHz and 600kHz. A pair of transducers are located at either end of the filter to translate electrical energy to mechanical energy, and *vice versa*.

The resonators supply a sharp shape factor of up to 1.2:1 (60-to-6dB), with Qs of 8000 to 12000. This is up to 150 times the Q of crystal filters.

Over a temperature range of –25°C to +85°C, the change of resonant frequency is as little as 1.5 parts per million. In one test, the frequency shift for eight months was one part per million.

The mechanical filter consists of three basic parts: a) the transducers, b) the mechanically resonant discs, and c) disc coupling rods. The transducer is a magnetostrictive device that converts electrical energy to mechanical vibrations, and *vice versa*. The resonant discs form parallel resonant circuits, so increasing the number of discs decreases the bandwidth of the circuit. ■

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JLH a lifetime in electronics

The coming of the transistor – this month John remembers the emergence and evolution of early transistor audio power amplifiers including the Lin design and his own 75 watter.

During the 1950s, a new type of amplifying device emerged in fits and starts – and in some strange shapes – from the semiconductor manufacturers' laboratories. By the end of the decade, reasonably reliable and workmanlike components were readily available.

I had begun to introduce low-power dual-transistor gain blocks of the type shown in **Fig. 6** in my domestic 'hi-fi'. These blocks were low-noise small-signal circuits with a low distortion of typically less than 0.01% at 1kHz and 1V RMS.

In this application, the freedom from heater-circuit wiring, the relatively low noise and the almost complete absence of mains-frequency hum in filter and frequency-correction circuits was welcome. Also, due to the fact that the devices gave off little heat in low-power use, one could, at last, make drift-free FM tuners.

However, there was no doubt that many reasonably competent electronics engineers found it difficult to come to terms with the 'solid-state' revolution.

The transition to transistors

Junction temperatures in excess of about 200°C for silicon devices – and even less for germanium ones – could cause permanent damage to a transistor, even after brief overheating. This snag was made worse because the devices were polarity sensitive. Excessive currents could flow if they were inadvertently connected the wrong way round.

By comparison, valves mostly didn't object to even severe overloads – provided that they were brief. Nor, for that matter, did valves mind being connected the wrong way round. They just didn't work.

If one complicates the issue by noting that transistors could also burst into unexpected – and sometimes damaging – parasitic oscillation, and consequently expire before your eyes, it is easy to understand why many would-be users retired hurt after their first encounters with solid-state components.

From a design point of view, the transistor differed from the valve because it was a current, rather than a voltage, operated device. All its electrodes presented internal conductive paths, whereas in a valve such inter-electrode paths were usually virtually open-circuit.

However, there were electronic circuit design engineers who were at home with these components, and they soon turned their attention to audio applications.

Transistor audio amplifiers

From the late 1950s, circuit designs for transformer-coupled audio amplifiers, based on germanium p-n-p diffused

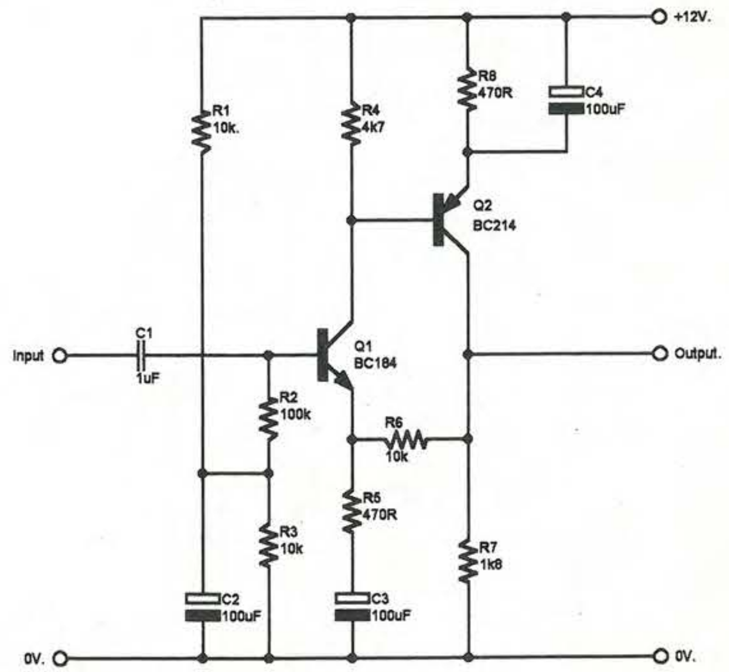


Fig. 6. Low distortion gain block.

junction transistors had been presented by transistor manufacturers. These germanium devices were the only ones commercially available at the time.

However, as in the case of valve-operated audio output stages, the challenge that remained was to eliminate the inter-stage coupling transformers.

These transformers were sources of distortion and they limited the frequency and phase linearity of the system. Moreover, even loudspeaker coupling output transformers shouldn't be needed with low impedance output loads such as a loudspeaker.

Fig. 7. Audio amplifier due to Lin.

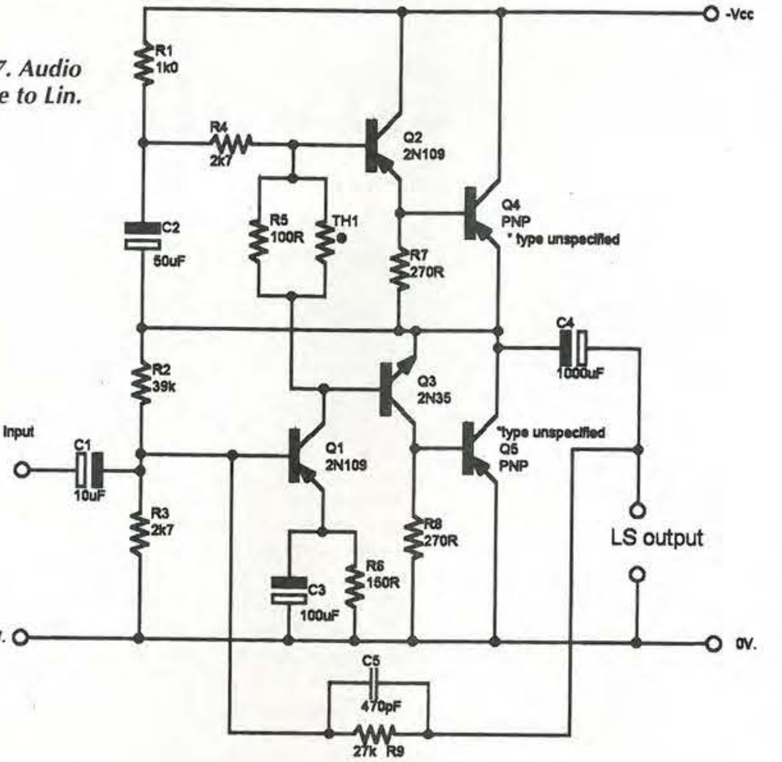
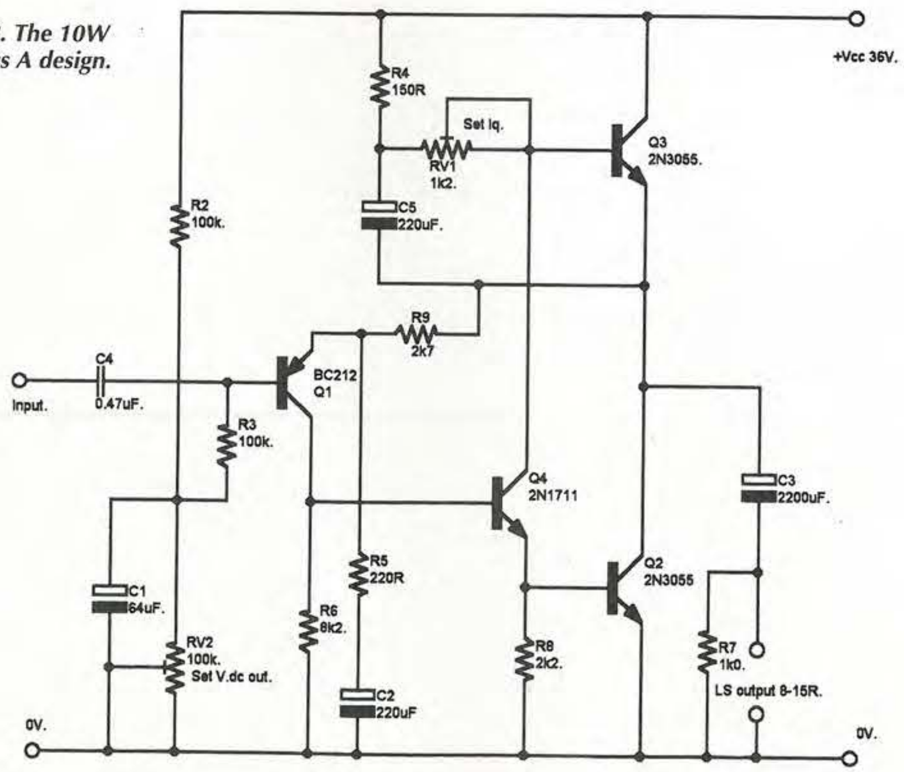


Fig. 8. The 10W class A design.



The Lin design

An almost complete answer to the demand for a transformerless audio power amplifier was proposed by H C Lin in 1956. At that time the only available power transistors were p-n-p by the way.

Lin's design is principally remembered for the introduction of the 'quasi-complementary' output stage layout. This involves a p-n-p/p-n-p and n-p-n/p-n-p arrangement used to simulate a symmetrical polarity pair of emitter-followers as the output group.

However, Lin's real contribution to amplifier design lay in his use of a low-power voltage amplifying stage followed by a low or unity-gain impedance conversion stage. This layout has been adopted for almost every power output stage design since then – not forgetting some thousands of op-amps. Lin's design is shown in **Fig. 7**.

This circuit gave some 6W output with around 1% distortion, and covered the audio range of 30Hz-15kHz within ±1dB. At that time, for reference, the better high power valve amplifiers could give 25 watts at less than 0.05% and with a 5Hz-50kHz bandwidth. However, these were early days, and I was quite content with my Williamson mono set up – the snag lying in the 'mono' bit.

Turntable trauma

When my old friend, the Seascale carpenter, made the cabinets for our two radiograms, he fitted each with a Decca turntable and its XMS sapphire-stylus pick-up. This seemed quite a good choice in 1952, before the advent of stereophonic LPs, even though the playing weight of the spring counter-balanced head was 40g or thereabouts.

Sadly though, there was no way that the Decca system would play a 'stereo' disc without jumping up and down during the loud bits.

Replacing the turntable and pick-up would have presented little difficulty, though any new system would probably not have fitted as snugly as the original joiners installation.

The real problem was what to do about the single Williamson power amplifier. Quite apart from the cost and difficulty of buying and installing another Williamson, with its separate 450V power supply, there simply wasn't room in the cabinet to accommodate it.

So my thoughts turned to constructing two equivalent transistor power amplifiers. This solution would certainly save space, but it had to sound as good too.

I treated the requirement as if it were for a typical industrial control system. The emphasis would be on simplicity – on the premise that 'what you don't put in won't go wrong'. I designed and made up the experimental four transistor power amplifier circuit shown in **Fig. 8**.

Why was it n-p-n throughout?

At that time, 1965, silicon p-n-p transistors were not very good, so the design used only n-p-n power devices. The opposite was the case with germanium devices, where it was the n-p-n ones which were relatively poor in performance.

This design operated in class A, which removed any problems which might arise with class 'AB' output bias level settings.

On being tested with a 35V supply line, this amplifier worked very well. It had an output power of a little over 10W, a THD figure rather better than 0.1%, and a bandwidth of 10Hz to 100kHz, ±0.5dB. This was very encouraging – especially when I compared its sound quality against the Williamson, and concluded that it was at least as good.

On seeing these results, I built a tidied-up stereo version as a Christmas present to myself in 1967. Some time later, I replaced the output transistors with Motorola 'epitaxial base' 2N3055s.

With an increase in the supply voltage to 45V, the new transistors allowed an output power of more than 15W. This was equivalent to the Williamson, though I am unconvinced that I ever needed or used more than two watts.

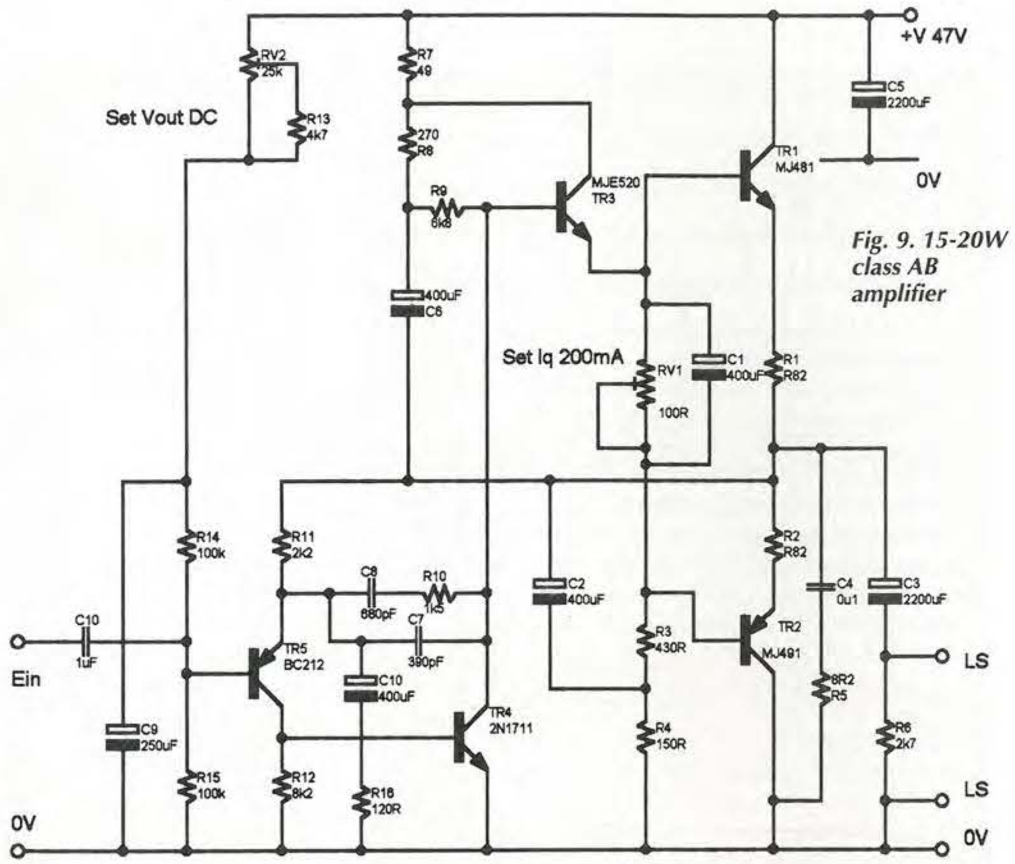


Fig. 9. 15-20W class AB amplifier

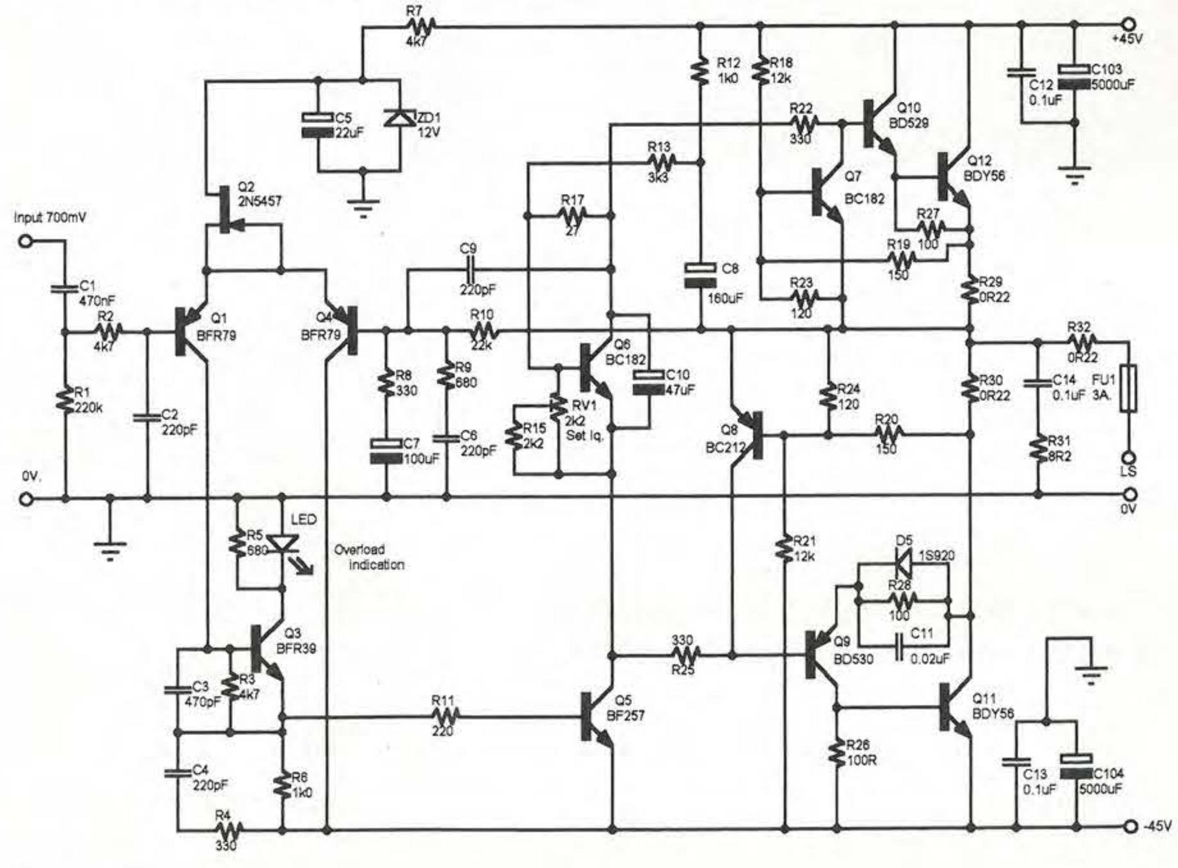


Fig. 10. 75 watt amplifier

Unlike the Lin design, my 10W Class-A circuit did not use a push-pull pair of output devices to provide the required low output impedance. Instead it used a 'Darlington pair' connected amplifier stage, comprising Q_1 and Q_3, driving Q_2 as an active load. Transistor Q_4 gave increased loop gain for the AC and DC feedback loops.

My first audio article

It had not occurred to me to seek to publish my design. But two of my friends urged me to do so.

John Greenbank, an assistant editor on *Wireless World* at the time, greeted my contribution with enthusiasm. Unknown to me, the topic of Class-A *versus* Class-AB was one of current hi-fi debate. As a result of this publication, I suddenly found myself to be a hi-fi guru.

I was urged to offer more audio circuit designs. Two worthy of recalling were the simple 15-20W class AB circuit shown in **Fig. 9** and a 75W design subsequently published in *Hi-Fi News* in 1972.

Of these designs, the 15-20W was aimed to give the same quality as the original 10W amplifier, but with more power at a much lower heat dissipation.

An unexpected tribute to its performance came at a lecture from a representative of a most eminent loudspeaker manufacturer. He mentioned that the company liked my design a lot – so much in fact that it had been adopted as an 'in-house' design.

The 75W design did not owe its inception to any feeling on my part that I needed more output power. It arose from a chance discussion with the managing director of a large speaker manufacturing company at one of the then popular annual Olympia or Earls Court Hi-Fi shows, where I had given a lecture.

The subject of our discussion was the incorporation of power amplifiers into the body of the speaker cabinet. This idea would allow the frequency response equalisation and cross-over networks to be provided at low power and high impedance – an arrangement which we both felt to be a good thing.

The question I then asked was what amplifier power should be provided. I was told '75W into 8Ω at 0.01% THD'. This was an interesting challenge, and I spent the next few months of my spare time trying out various likely circuit arrangements.

High-power audio

The final amplifier circuit, shown in **Fig. 10**, was demonstrated at a BKSTS lecture in London. The chairman was Mike Jones, technical editor of *Hi-Fi News*, where the circuit design subsequently appeared.

This design was offered in kit form by several manufacturers and proved enormously popular. At that point I resolved I would do no more audio amplifier designs. They took up too much time.

Alas, I did not keep to my resolve. There were two reasons for this – technical curiosity and requests from friends. ■

In the next article, John looks at the coming of the IC.

What is Bluetooth?

Soon it will be possible to connect peripherals to your PC via an rf link using a wireless local network technology called Bluetooth. Geoff Lewis outlines how it works.

Bluetooth is the title used for an open-specification system designed to provide a fast and secure short range radio interconnection between portable devices and a local area network (LAN) through the use of spread spectrum techniques. The services provided for include internet, e-mail, image and data transmission together with voice applications extending to three simultaneous 64kbit/s PCM channels.

The concept has been adopted by about 2000 diverse user organisations and is supported by many of the major semiconductor chip manufacturers. These include Intel, Toshiba, VLSI, Nokia, Texas Instruments, Ericsson, Thomson (ST), Lucent and Siemens. In addition, the Bluetooth standards have gained the approval of the FCC (Federal Communications Commission), ETSI (European Telecommunications Standards), the IEEE (Institute of Electrical and Electronic Engineers) and the IrDA (Infrared Data Association) for global use.

The 'user portable terminal devices' involved can include mobile phones, headsets, printers, digital cameras, PDAs (personal digital assistants) and lap/palm top and notebook computers, The emphasis is to provide a robust worldwide system capable of acting as a cable type of replacement at low cost with plug and play capabilities.

The transmission system

The system operates in the license-free Industrial, Scientific and Medical, or ISM, band from 2.4GHz to 2.5GHz and employs frequency-hopping spread-spectrum (FHSS) technology.

For this application, FHSS is preferred to the more common direct-sequence spread-spectrum (DSSS) system.

While FHSS is somewhat slower in operation, it is much simpler to implement through the use of a frequency synthesiser. The power output of the transmitters ranges from 1 to 2.5mW (0 to 4dBm) to provide a coverage of at least 10 metres – a range that can be increased to 100 metres by the use a radio-frequency power amplifier.

A prime number (79) of sub-carriers, spaced by 1MHz, occupies the band 2.402MHz to 2.480MHz. Each sub-carrier is frequency modulated with a deviation of either –140kHz to –175kHz for binary 0 or +140kHz to +175kHz for binary 1. This modulation method is referred to as Gaussian frequency-shift keying (GFSK) because the base-band digital signal is filtered to provide a Gaussian bell-shaped response with at least 90% of the pulse energy contained within the 3dB bandwidth. The resulting signal has a good spectrum efficiency and resilience to co-channel interference (CCI).

Each sub-carrier can be switched at the rate of 1600 frequency hops per second to provide a spread spectrum signal with a hop period of 625μs. The maximum bit rate is currently set to 1Mbit/s but because each sub-carrier can carry multiple bits during

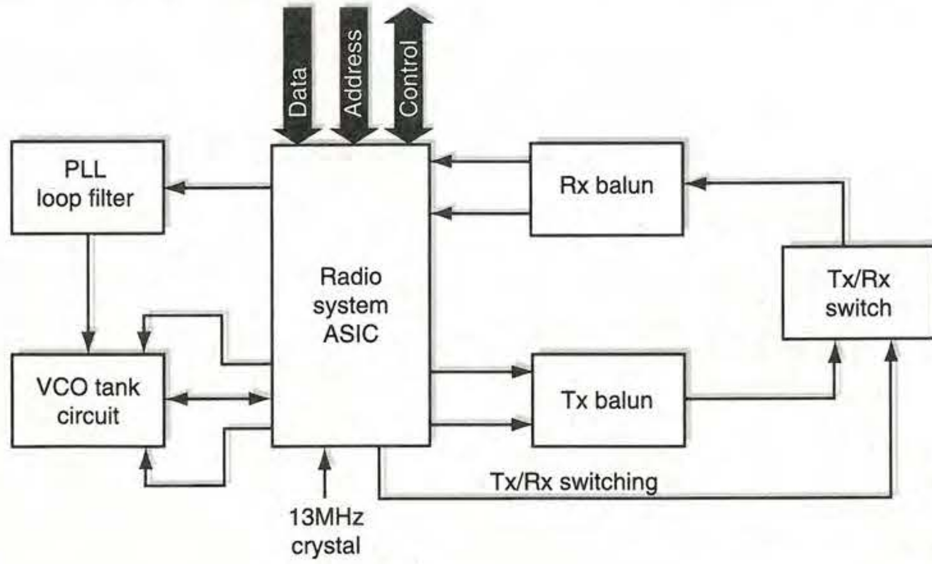


Fig. 1. At the heart of a Bluetooth radio module is a radio-system asic. This chip carries out all functions associated with sub-carrier generation and control.

each time slot, this figure might be doubled in the near future.

A bandwidth bit period ($Bw.T_{bit}$) product of 0.5 has been set for Bluetooth. For a maximum bit rate of 1Mbit/s, this yields a signal bandwidth of 0.5×10^6, which is 500kHz. Under these conditions, the system has an adjacent carrier interference (ACI) ratio of better than –20dBm. If the bit rate is increased to 2Mbit/s then this ACI protection ratio will be reduced.

Bluetooth's radio module

As currently configured, each Bluetooth terminal is constructed around two highly integrated modules. Each contains an application-specific integrated circuit (ASIC) designed to handle the radio functions and the baseband signal processing.

The basic module is constructed on a laminated substrate consisting of seven layers of metal interleaved by six layers of ceramic. This is then connected to a PCB via ball grid array (BGA) mounts.

The lower metal layer acts as a ground plane while the component side carries a pair of metal screens to complete the electromagnetic shielding. In order to maintain good interference rejection, the various earths points are distributed around the ground plane.

As **Fig. 1** illustrates, the radio system ASIC carries out all the functions associated with the generation and control of the sub-carrier frequencies. This is achieved through the use of a voltage-controlled oscillator, or VCO, linked to a phase-locked loop circuit. In turn, the PLL is locked to the output of a 13MHz crystal oscillator.

A VCO tank circuit forms part of the PLL circuit. Its tuned load is laser trimmed for accuracy. A loop filter is used to remove any ripple from the output of the PLL circuit. For receive purposes, this IC employs a heterodyne technique to generate a low IF at 3MHz.

Spread-spectrum frequency-hopping modulation is performed directly on the VCO. A pair of balun circuits convert the balanced signals used within the ASIC into the unbalanced form used externally for transmission and reception.

An antenna switch diverts the signals between the baluns and the antenna input/outputs for the transmit/ receive functions. The antenna filter reduces the harmonic radiation from the terminal device during transmit and helps to minimise the unwanted interference effects during receive. The antenna feed is designed to match a 50Ω impedance load.

Baseband signal processing module

Figure 2 is a simplified block diagram of the baseband signal processor. The ASIC in this stage contains an embedded reduced instruction-set computer core that controls the operation of the radio section and the inputs from either of the input/output interfaces.

A flash reprogrammable memory carries the software that provides the overall control of the system. The supply voltage is well regulated and filtered to provide power for both the signal interfaces and the transceiver. The 13MHz clock signal from the radio section is used to synchronise the action of the processor ASIC.

The universal asynchronous receiver transmitter, or UART, interface operates as a 'data circuit equipment', or DCE, device under the control of the request to send or clear to send signals RTS and CTS. In turn, these handshaking signals control the data flow lines TxD (transmit data) and RxD (receive data). This interface can handle all the standard data rates from 300bit/s up to 460.8kbit/s.

Apart from the universal serial bus standard bi-directional ports D+ and D–, two additional inputs are provided for the control of a lap/notebook computer. The 'wake-up' signal advises the host computer that the terminal has become operative and the 'detach' signal indicates a suspend operation mode. While the USB port is capable of running at 12Mbit/s, the Bluetooth operation is limited to 1Mbit/s.

The PCM interface operates at the standard sample of rate of 8kHz and can handle linear PCM from 13 to 16 bits, plus both μ-law and A-law 8-bit companded samples.

Software and protocol

The code stored within the flash memory provides control for both host controller interface and the local link manager for the USB, UART and PCM ports. The operational protocol provides for link set up and configuration together with authentication.

To maintain a secure transmission system and a high data rate in what might well be a noisy environment, Bluetooth operates with a packet switching protocol that includes both forward error control and encryption.

Networking Bluetooth

Bluetooth terminals are organised into small groups referred to as 'piconets'. These contain up to eight peer devices with one acting as a master for the group. The simplest network is just two devices linked by a virtual cable.

Piconets may be linked together via a terminal device that is common to two groups, but this must not be one of the masters. Each piconet has a different frequency hopping sequence and a collection of piconets is known as a 'scatternet'.

Finally, I should like to acknowledge the help received from Ericsson Components AB in preparing this article. The company's Bluetooth data sheets and extensive information on the web site,

WWW.ericsson.com/bluetooth

have been particularly helpful. ■

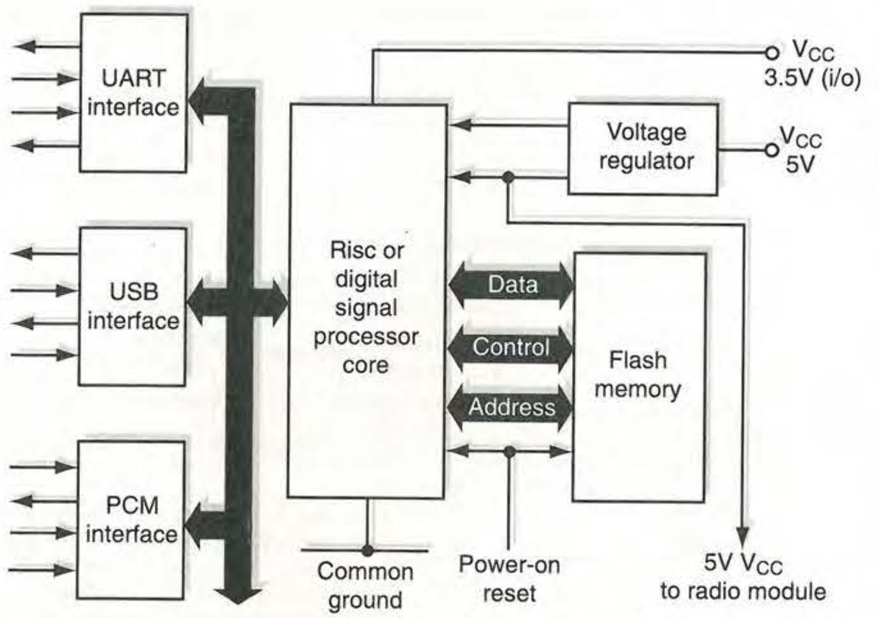


Fig. 2. Baseband signal module controls operation of the radio section and inputs from either of the i/o interfaces using a RISC or DSP core.

A novel inductance meter

Combining high-speed phase detection circuitry with microcontroller data processing, Michael Slifkin and Alexander Gornstein have developed an inductance meter capable of displaying accurate measurements from 20μH to 200mH on an LCD module.

There are many different ways of measuring reactance. Before the era of digital techniques, the usual methods were to use bridge circuits and to find the null point by adjusting resistors. Often the null point was very broad and it could take some time and manipulation to find the reactance value.

Reactance differs from resistance in that it is frequency dependent. Furthermore, inductance will contain some resistance, known as ohmic loss, and this will affect the measurements.

Nowadays reactance measurements can be made automatically, digitally and accurately. There are different options for carrying out such measurements and it is not always possible to say which is the best general method. Among the methods for measuring reactance are:

- Via a bridge. This method is outdated now, but you could automate the technique using a quasi bridge method to bring it up to date. This involves measuring the out-of-balance voltage or current rather than looking for a null, as is normally the case.
- Phase-shifting. Here you would use the reactance as the variable element in a phase-shifting circuit and then measure the change in phase of an oscillator working at fixed frequency. There are different ways of measuring phase change. You could use a pulse-counting technique as described in our article on a power factor meter in the February issue. Alternatively you could use the phase detector described in the April 1999 issue on page 312. It would be possible to devise a range of methods for such measurements.
- You could use the reactance as a frequency determining element in a tuned circuit and obtain the new frequency or change in frequency caused by the reactance. If the change in frequency were very small, you would probably find a linear relationship between the change and the value of the reactance.

Michael Slifkin and Alexander Gornstein, Jerusalem College of Technology, Israel.

Specifications

Displays three digits together with units. Overload message shown. The display is continuously updated

Measurement from 20μH to 200mH in four switchable ranges:

Range 1	20μH to 200μH
Range 2	200μH to 2mH
Range 3	2mH to 20mH
Range 4	20mH to 200mH
Accuracy	±(1.5%+single digit)
Voltage (V_C)	8V

Methods exist that are specific to the type of reactance being measured. You could measure capacitance by charging it to a known voltage and then find the discharge time. Or you could try to measure inductance by finding the back EMF when a pulse is applied to it.

Standard radio techniques can be used to measure reactance. This is not as far-fetched as it sounds. For example, a radio receiver with a BFO would detect the shifting of a carrier wave due to reactance added to the tuned circuit. You would not need to use an actual radio receiver to do this, although I once saw a radio receiver used in this manner about 45 years ago.

Nowadays you can buy chip sets comprising a transmitter and receiver in frequency and phase modulation modes that can be readily adapted to reactance measurements. This is not the subject of this article, but perhaps of one in the future.

Finally, you could try to measure the reactance directly by finding the current through the reactance at some voltage. This is the method we chose.

Although this would appear to be an obvious method, it does not seem to have been used before, probably because of the 90° phase shift between the current and the voltage introduced by the reactance. The ohmic loss also causes phase shifts of unknown value.

The instrument in detail

Figure 1 is a block diagram of our instrument and **Fig. 2** is its full circuit diagram.

The meter involves a simple analogue circuit in which the inductance to be measured is placed in series with an alternating constant current generator at the input to a high impedance FET op-amp. A voltage is produced across the inductance. This voltage is amplified. By using suitable values of gain, the amplified voltage is found to be numerically equal to the inductance.

At this point, the voltage is converted to a digital form. A processor now calculates the value of inductance according to the range and sends the reading to a 16-character by 2-line liquid-crystal display module.

The best form of detection of a sinusoidal voltage signal is the phase-sensitive detector, as mentioned above. As explained in the earlier article, the phase-sensitive detector is just a multiplier and a low-pass filter.

Our implementation of the phase-sensitive detector here is very different from that described in the previous article. In this case, we need only a fixed frequency but at the higher value of 1MHz as compared to a low frequency but variable over a wide range. The reason for using this high frequency is that the inductor has a small inherent resistance – ohmic loss – but as the reactance is given by ωL, this can be swamped by using a relatively high value of ω. Inductance of the component in henries is L, and ω is the angular frequency, i.e. $2\times\pi\times f$ where f is the frequency in hertz.

Phase-sensitive detector

Digital switches together with a high quality op-amp form the phase-sensitive detector. To understand how it works, have a look at **Fig. 3**, which is taken from our 1999 article. The effect of the inductor is to add a 90° phase shift between the current and the voltage. Adding a 90° phase shift in the reference channel compensates for this.

In this design, the 90° phase shifter is the

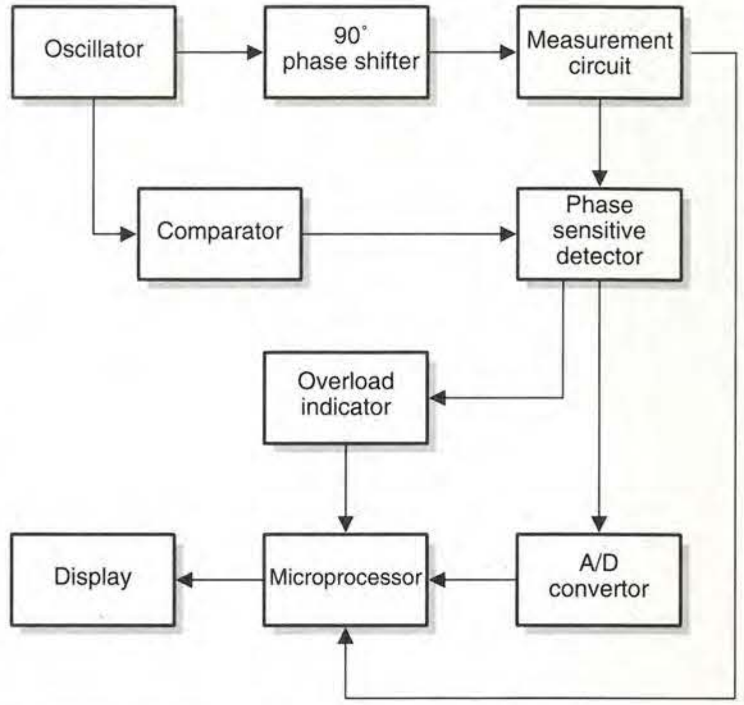


Fig. 1. Combining analogue and digital circuitry, the inductance meter. This design applies a constant-current sine-wave to the inductor under test and measures the resulting voltage.

Hexadecimal listing of 8051 code for the inductance meter. Locations not shown are filled with zeros.

```
000000  02 00 03 75 81 50 12 01 50 79 00 12 00 25 12 02 37 80 f8 e5 90 54  ...u.P..Py...%..7.¯Â.T
000016  08 60 0b 12 01 68 90 05 4c 12 01 8b 80 ef 22 e5 90 b2 e4 54 10 60  .`...h..L....Ô"Â.≤‰T.`
00002c  05 78 01 12 00 5a e5 90 b2 e5 54 20 60 05 78 02 12 00 95 e5 90 b2  .x...ZÂ.≤ÂT `.x....Â.≤
000042  e6 54 40 60 05 78 03 12 00 d5 e5 90 b2 e7 54 80 60 05 78 04 12 01  ÊT@`.x...'Â.≤ÁT.`.x...
000058  15 22 e9 98 60 0b e8 f9 12 01 5b 90 05 00 12 01 8b 11 13 e5 a0 75  ."È.`.Ë˘..[.......Â†u
00006e  f0 64 84 12 01 58 12 01 68 12 01 7f e5 f0 75 f0 0a 84 12 01 58 12  d...X..h...Âu....X.
000084  01 7f e5 f0 12 01 58 12 01 7f 90 05 35 12 01 8b 22 e9 98 60 0b e8  ..Â..X.....5..."È.`.Ë
00009a  f9 12 01 5b 90 05 0d 12 01 8b 11 13 e5 a0 75 f0 64 84 12 01 58 12  ˘..[........Â†ud...X.
0000b0  01 68 12 01 7f e5 f0 75 f0 0a 84 12 01 58 12 01 7f 74 2e 12 01 7f  .h...Âu....X...t....
0000c6  e5 f0 12 01 58 12 01 7f 90 05 35 12 01 8b 22 e9 98 60 0b e8 f9 12  Â..X.....5..."È.`.Ë˘.
0000dc  01 5b 90 05 19 12 01 8b 11 13 e5 a0 75 f0 64 84 12 01 58 12 01 68  .[........Â†ud...X..h
0000f2  12 01 7f 74 2e 12 01 7f e5 f0 75 f0 0a 84 12 01 58 12 01 7f e5 f0  ...t....Âu....X...Â
000108  12 01 58 12 01 7f 90 05 35 12 01 8b 22 e9 98 60 0b e8 f9 12 01 5b  ..X.....5..."È.`.Ë˘..[
00011e  90 05 24 12 01 8b 11 13 e5 a0 75 f0 64 84 12 01 58 12 01 68 12 01  ..$.....Â†ud...X..h..
000134  7f e5 f0 75 f0 0a 84 12 01 58 12 01 7f e5 f0 12 01 58 12 01 7f 90  .Âu....X...Â..X....
00014a  05 40 12 01 8b 22 c2 92 d2 92 12 01 c7 22 24 30 22 c2 90 75 80 01  .@..."¬í"í..«"$0"¬.u..
000160  d2 91 12 02 2e c2 91 22 75 80 c0 c2 90 d2 91 12 02 2e c2 91 22 c2  "ë...¬ë"u.¿¬."ë...¬ë"¬
000176  90 d2 91 12 02 2e c2 91 22 f5 80 d2 90 d2 91 12 02 2e c2 91 22 e4  ."ë...¬ë"ı."."ë...¬ë"‰
00018c  93 b4 24 01 22 f5 80 d2 90 d2 91 12 02 2e c2 91 a3 80 ec 22 75 80  .¥$."ı."."ë...¬ë£.Ì"u.
0001a2  80 c2 90 d2 91 12 02 2e c2 91 22 75 80 0c c2 90 d2 91 12 02 2e c2  .¬."ë...¬ê"u..¬."ë...¬
0001b8  91 22 75 80 0f c2 90 d2 91 12 02 2e c2 91 22 c2 90 7f 20 7e ff de  ë"u..¬."ë...¬ë"¬.. ~ˇﬁ
0001ce  fe df fa 75 80 38 d2 91 12 02 2e c2 91 7f 14 7e ff de fe df fa 75  ˛ﬂ˙u.8"ë...¬ë..~ˇﬁ˛ﬂ˙u
0001e4  80 38 d2 91 12 02 2e c2 91 7e 64 de fe 75 80 38 d2 91 12 02 2e c2  .8"ë...¬ë~dﬁ˛u.8"ê...¬
0001fa  91 75 80 38 d2 91 12 02 2e c2 91 75 80 08 d2 91 12 02 2e c2 91 75  ëu.8"ë...¬ëu.."ë...¬ëu
000210  80 01 d2 91 12 02 2e c2 91 75 80 06 d2 91 12 02 2e c2 91 75 80 0c  .."ë...¬ëu.."ë...¬ëu..
000226  d2 91 12 02 2e c2 91 22 7f 28 7e ff de fe df fa 22 7f ff 7e ff de  "ë...¬ë".(~ˇﬁ˛ﬂ˙".ˇ~ˇﬁ
00023c  fe df fa 22 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00  ˛ﬂ˙"..................
000252  00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00  ......................
...
0004e6  00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00  ......................
0004fc  00 00 00 00 55 70 20 74 6f 20 32 30 30 20 6d 48 24 55 70 20 74 6f  ....Up to 200 mH$Up to
000512  20 32 30 20 6d 48 24 55 70 20 74 6f 20 32 20 6d 48 24 55 70 20 74   20 mH$Up to 2 mH$Up t
000528  6f 20 32 30 30 20 6d 69 63 72 6f 48 24 20 6d 48 65 6e 72 79 20 20  o 200 microH$ mHenry
00053e  20 24 20 6d 69 63 72 6f 48 65 6e 72 79 24 4f 75 74 20 6f 66 20 72   $ microHenry$Out of r
000554  61 6e 67 65 24 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00  ange$.................
00056a  00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00  ......................
...
000fe6  00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00 00  ......................
000ffc  00 00 00 00                                                        ....
```

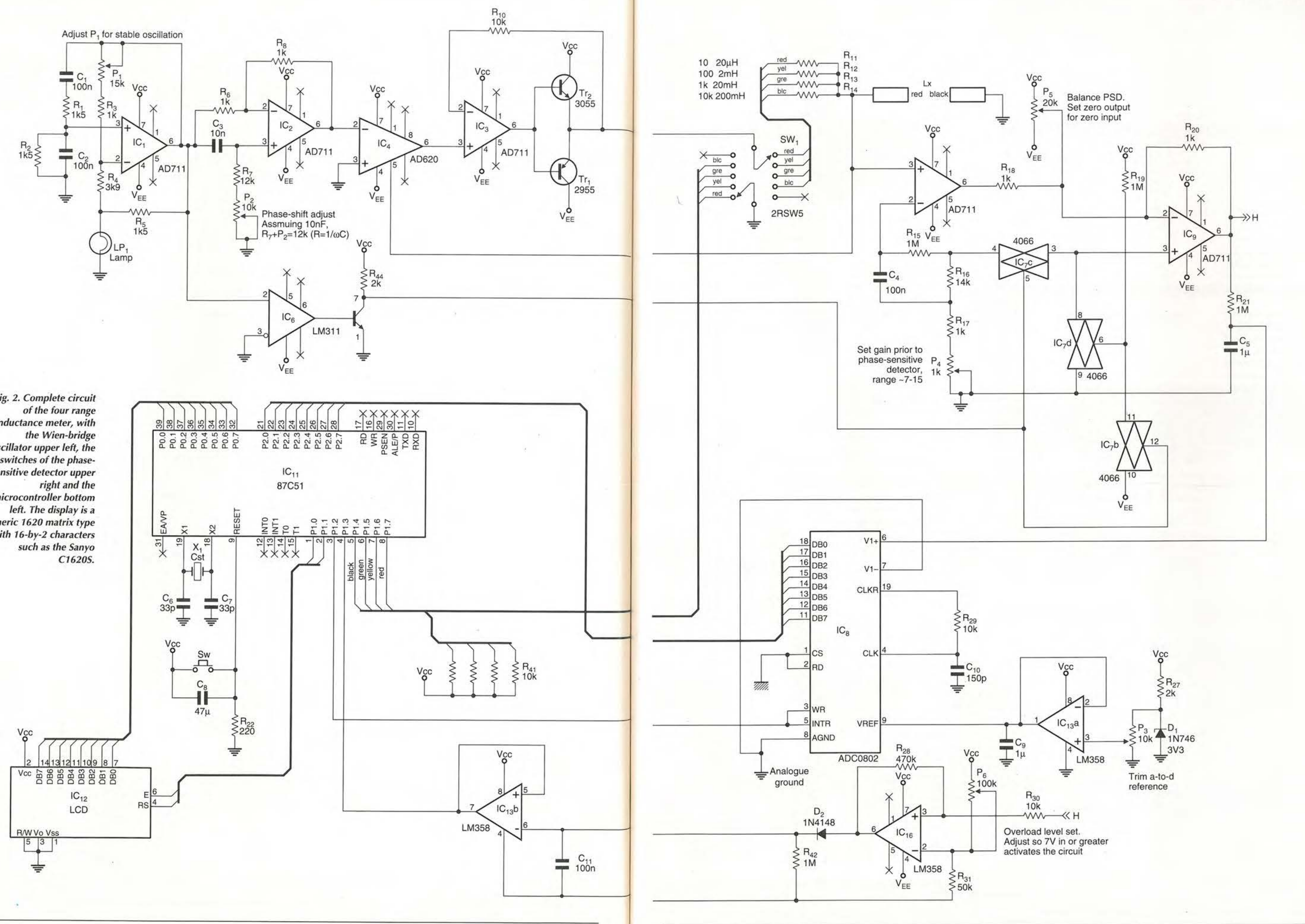


Fig. 2. Complete circuit of the four range inductance meter, with the Wien-bridge oscillator upper left, the switches of the phase-sensitive detector upper right and the microcontroller bottom left. The display is a generic 1620 matrix type with 16-by-2 characters such as the Sanyo C1620S.

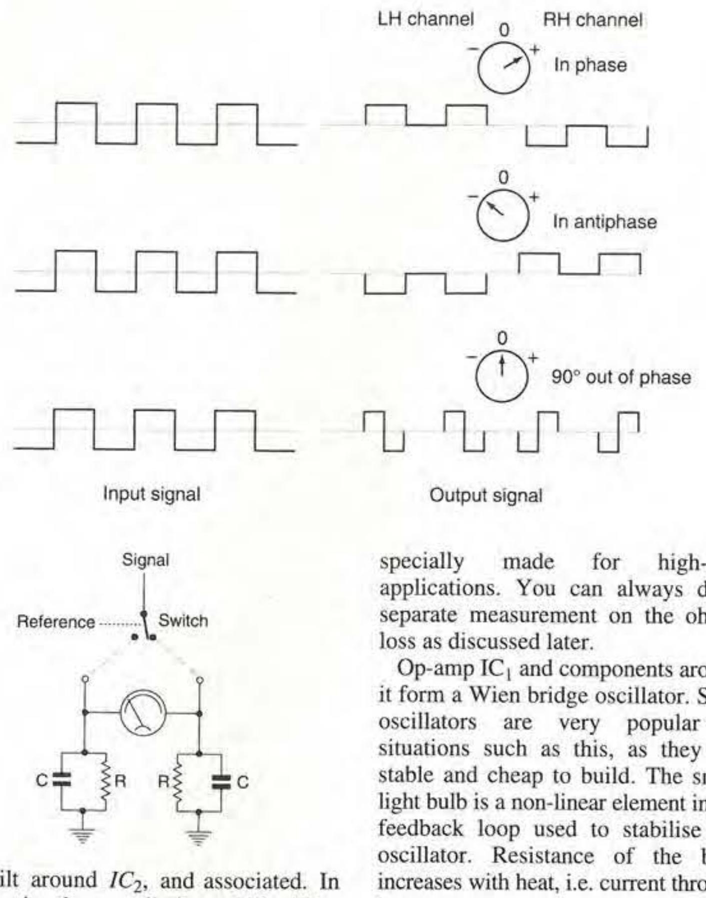


Fig. 3. Phase-sensitive detector concept. Taken from the April 1999 issue, this diagram shows how a switch operated by a reference signal feeds the signal being detected to two RC networks alternately.

built around IC_2, and associated. In practice the overall phase shift will be slightly less than 90° due to the ohmic loss. However, as the reactance should be very much higher than the resistance, the difference will not be great.

A phase shift error of as much as 10° will cause an amplitude error of only 1.5%. Modern inductors are most unlikely to introduce any greater error due to the ohmic loss, except for those specially made for high-loss applications. You can always do a separate measurement on the ohmic loss as discussed later.

Op-amp IC_1 and components around it form a Wien bridge oscillator. Such oscillators are very popular in situations such as this, as they are stable and cheap to build. The small light bulb is a non-linear element in the feedback loop used to stabilise the oscillator. Resistance of the bulb increases with heat, i.e. current through it.

A crystal-controlled oscillator would be easier to implement than the Wien bridge, albeit not as cheaply. However, the stability of the Wien bridge is perfectly adequate in this application. Comparator IC_6 converts the sine wave produced by the Wien bridge oscillator into a square wave suitable for driving the phase-sensitive detector.

This instrument has four manually-selected ranges. Each range is indicated to the microprocessor on a unique i/o pin.

Overload indication involves IC_{16} and IC_{13B} and a separate pin on the microprocessor. The display is a 16 by 2 alphanumeric LCD display made by Sanyo. **Figure 4** details its connections.

The microprocessor is an 87C51. It has an on-board EPROM into which the control program is loaded. We provided the 8V V_{cc} supply rail using a separate power supply. Power drain is rather high for battery operation.

Programming for this instrument is not particularly difficult and only occupies a small amount of memory. The object code is given and the source code in assembler is available from *Electronics World*.

Enhancements

Only a small portion of the EPROM in the microprocessor is used, so there's plenty of scope for enhancements. For example, you could add autoranging by monitoring the measurements and then switching in the appropriate range using electronic relays. You would not need to get the gain exactly right in the instrument; it should be possible to provide a scaling factor in software.

You could easily make a very accurate instrument by also measuring the ohmic loss, *r*, under software control. This would be done by evaluating the DC voltage across the inductance for a given DC current or *vice-versa*.

If *X* is the apparent inductance value then the true value is given by $L=X\sin\theta$ with $\cos^{-1}\theta=r/\omega X$. Of course you could measure the ohmic loss with a simple voltmeter and then apply the correction.

In summary

This instrument allows you to measure a wide range of inductance to a good accuracy without the problem of ohmic loss errors. Extra programming space is available to upgrade the instrument. You could improve accuracy even further by measuring the ohmic loss separately. ■

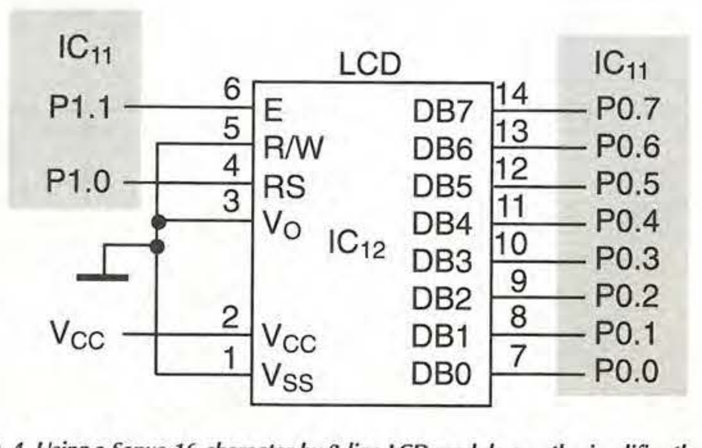


Fig. 4. Using a Sanyo 16-character by 2-line LCD module greatly simplifies the inductance meter.

Technical support

Michael Slifkin is in the Department of Electronics at the Jerusalem College of Technology. Alexander Gornstein is a former student at JCT. The authors have opened a web site at: http://optics.jct.ac.il/~slifkin/nim.htm. It contains URLs to all the devices used in this instrument.

If you want the object code in electronic form, e-mail jackie.lowe@rbi.co.uk and it will be forwarded to you as text embedded in an e-mail. Please use L meter as the subject heading.

If you want the object code on disk, again free of charge, send us a PC formatted 3.5in floppy with a protective self-addressed envelope and enough stamps to allow us to return the disk to you. Post it to the Quadrant House address below.

The source code is available on disk for £15, or as an e-mail attachment for £10 fully inclusive. E-mail the above address with your credit card number, expiry date and card-holder address or fax 020 8652 8555 with the details. Alternatively, send your order to L meter, Electronics World Editorial, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS.

Everything but the sink

Two significant breakthroughs in isolated modular dc-to-dc converters have been made by a relative newcomer in this field. The company, SynQor, has introduced synchronous rectification – increasing efficiency to such an extent that the new converters are the first of their kind to do away with heat sinking. Secondly, planar magnetics have been used, reducing weight and lowering board height.

In the interests of increasing reliability, and to ease the problem of conveying power to electronic equipment in racking, it's common to have regulators on cards distributed throughout the racking system. Each individual regulator is fed by a relatively high voltage, allowing the cross section of power-distributing buses to be reduced.

Distributing the power supply in this way means that heat lost through regulator inefficiency is spread around the cabinet. This is a great advantage relative to having one large central regulator, but it does mean that valuable card space is lost to accommodate the individual regulators.

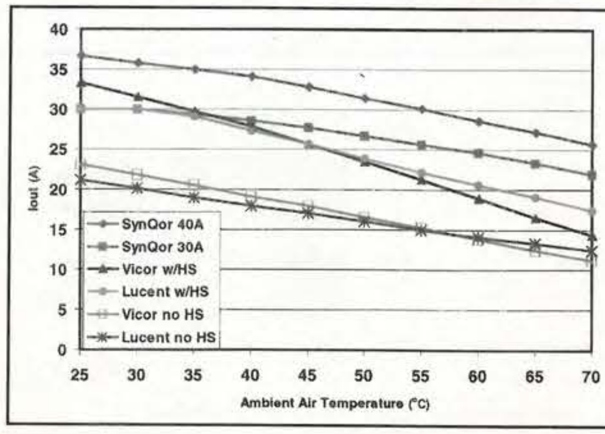
There has been little innovation in the design of such regulators over the past few years. Almost all the inefficiency in modern regulators is still caused by the voltage drop of a Schottky diode – evidence that current products are making the most of traditional switching circuitry.

A typical 5V, 30A regulator will have an efficiency of 83%. At full output, it would dissipate just over 25W. Clearly, such a regulator needs some form of heat sink on top of it, increasing the amount of valuable inter-board space needed for the card, its regulator and the regulator's heat sink.

While 5V regulators have been the norm, and alternatives to Schottky diodes have been impracticable, designers have had to tolerate regulators that need additional heat-sinking. But now that supplies lower than 5V are becoming common, the dissipation problem is compounded.

Efficiency of a traditional 3.3V Schottky-diode-based regulator drops to 79% at 30A – an equivalent 1.5V product is only 64% efficient. This fall in efficiency is mainly due to the fact that the diode's voltage drop is becoming a greater proportion of the output voltage.

So when everything else is shrinking, the regulator assembly is actually increasing in size.



Derating is another important issue in DC-to-DC converter design. As ambient temperature increases, the amount of power you can get out of a DC-to-DC converter falls. These curves show the new SynQor 40A product against some of its competitors for a 200 linear feet-per-minute air flow.

Output	Full load	Conventional converter	New converter	Heat reduction
15V	10A	88%	90%	1.2×
12V	12.5A	86%	90%	1.5×
5V	30A	83%	89%	1.7×
3.3V	30A	79%	89%	2.2×
2.5V	30A	74%	87%	2.4×
2.0V	30A	68%	85%	2.7×
1.5V	30A	62%	83%	3.0×

Reduction in dissipated heat offered by SynQor's DC-to-DC converters over conventional Schottky-diode based converters.

So efficient they don't need a heat sink

Now, a new player in the DC-to-DC converter market, SynQor, has developed a range of products that need no heat sinking. Being more efficient, these regulators output more power for a given volume. And needing no potting compound or heat sink, they are significantly lighter. Eliminating the heat sink also removes the unreliability of the thermal interface between the traditional converter and its heat sink.

The company has replaced the traditional Schottky diodes with synchronous rectifiers, which has made a significant impact on efficiency. In addition, planar transformers are being used, offering benefits in both weight and board height. Windings of these transformers are actually pcb tracks, and their ferrite 'core' passes through the pcb, wrapping around heavy copper tracks.

These are not the first designs to be based on synchronous rectification, but they are the first to incorporate a patented method of improving rectifier efficiency. The rectifiers in this case are MOSFETs. To achieve a low 'on' resistance, these FETs have to be large-area types. But large area is synonymous with large gate capacitance.

Normally, driving a high gate capacitance at a few hundred kilohertz would waste considerable power, but SynQor is using a patented method of re-using the gate driving energy.

Resulting efficiency improvements are good at 5V, rising from 83% for a typical conventional converter, to 89% for the new products. But at 1.5V and 30A, the efficiency is 83% against 62% for a conventional converter. Work this out in watts at 30A and you will get an idea of what a difference the efficiency gain makes. ■

For more information, see www.synqor.com

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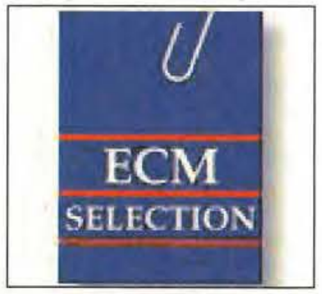

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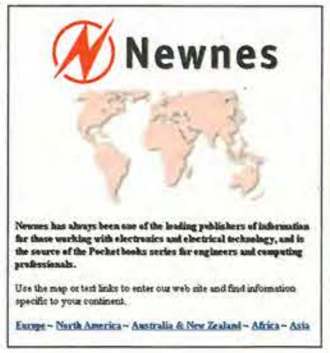

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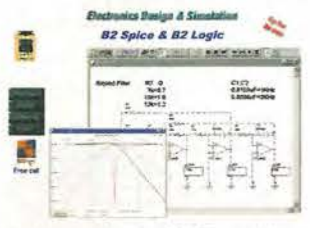

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Letters to the editor

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e-mail jackie.lowe@rbi.co.uk using subject heading 'Letters'.

TID: RIP?

Over the last thirty years there has been a continuing debate as to what might affect the sound quality of transistor power amplifiers.

In the early seventies, the argument was fired up by Lohstroh and Otala's transient intermodulation distortion (TID) theory[1] but over the last few years this appears to have been slowly falling out of favour.

Self has argued strongly for the Miller capacitor as a means to solve amplifier stability and distortion problems.[2] On the face of it, this seems to be a useful way forward. However, this Miller capacitor causes slew-rate overloading when the input signal exceeds the slew-rate limit.

I thought it useful to determine this limit so that it need not be exceeded. The calculations I have made contained a surprise that does indeed link the amount of feedback to the maximum acceptable input rate of change of voltage as I will now explain.

Consider a typical transistor stage in the diagram on the right. The input differential pair operate at a nominal current of i_s and the current source supplying them is $2i_s$. The voltage-amplifier stage, or VAS as Self calls it, shares any current between the base current, i_b, and the Miller capacitor current i_m. The Miller current is proportional to the collector voltage,

$$i_{in} = i_b + i_m \tag{1}$$

where,

$$i_m = C.\frac{dV_c}{dt} \tag{2}$$

This Miller-capacitor current would have appeared as a voltage across the emitter resistors of the input transistors, which are usually used to linearise the input stage to reduce distortion (see Self, *op. cit.*). In order to determine the maximum drive to this capacitor to prevent the input stage from hard switching, and thus cause a high level of distortion, it is necessary to consider what the limit might be.

Limited-slew criterion

I propose that the worst-case limit should be where the input transistors are able to respond to an input, and thus retain control of the amplifier. If the input-pair currents are altered by up to half their DC levels, this would leave a current of a further 50% still available to them. But with such a change this would already be 'large signal' and the distortion levels would be rising. However, it seems to be a reasonable starting point.

In this circumstance, i_1 might be $1.5i_s$, while i_2=$0.5i_s$. Note that the emitter-base

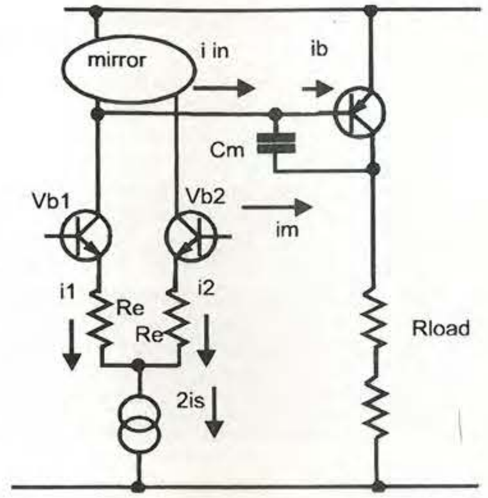

Fig. 1. Input stage of an audio amplifier in which current through each device of the input pair is i_s.

voltages are quite different, and are,

$$dV_b = V_{th}(\log(1.5) - \log(0.5)) \tag{3}$$

where V_{th}, Is the thermal voltage, 25.3mV at 20°C, and the logs are natural. (I prefer the programmer's 'log' rather than the mathematician's 'ln'), dV_b is 27.8mV here.

With rather low values, in my opinion, of 22Ω emitter resistors, as Self originally suggested, there is an additional increase of 22mV where i_s=1mA. This gives a total margin between the bases of about 50mV. It would be 130mV for 100Ω resistors.

When the output voltage is limited by the slew rate to,

$$\frac{dV_c}{dt} = \frac{i_s}{C_m} \tag{4}$$

where the feedback point (V_{b2} the base of the second output transistor) rises at a rate of,

$$\frac{dV_{b2}}{dt} = \frac{i_s}{C_m}\alpha \tag{5}$$

where α is the feedback attenuation factor, assuming that the output stage follows the VAS voltage closely.

If you plot this rising voltage with time, a straight line is observed as shown by (1) in the graph on the next page, where the maximum voltage reached corresponds to the maximum input voltage after the rise time t_r, as observed at the feedback point.

I said earlier that to achieve this voltage, the input must not exceed the 50mV drive

There's a mouse on my knee

In the article 'Mouse holes' in the May 1999 issue, I mentioned the use of a proprietary pad on the upper leg for working the mouse. Using it there reduces muscular strain.

At the time, although these things were being used in the USA, there was no product available in the UK, so I was unable to nominate a supplier or give a review. Now a firm called NMI in Newcastle-upon-Tyne is making a rather good design that has won an Innovation 99 award. I have tried it out and it seems fine. You have to adapt your technique somewhat, but it certainly relieves the aches and pains of continuous mousing.

Rod Cooper
Sutton Coldfield
West Midlands

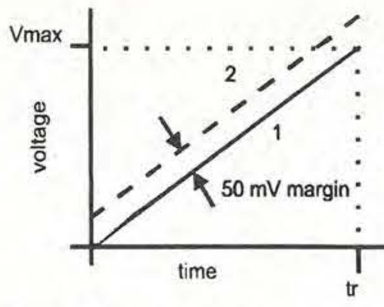

Fig. 2. Voltage curve up to rise time t_r.

margin, or the currents in the transistors will be in danger of cuffing off. This margin is shown by the dotted line on the graph.

Input voltage, V_{b1}, can be drawn to drive V_{b2}+0.05 by line (2). This allows the input transition to change very slightly faster than the feedback, but is in essence virtually the same.

For a typical amplifier operating with i_s=1mA, a Miller capacitor of 100pF, and 50W 'rms' output, the peak output voltage is 28.3V for 8Ω, the maximum slew rate guideline is 10V.μs^{-1}.

With a feedback factor of 0.033, i.e. a gain of 30, the input slew limit is 0.33Vμs^{-1}. This corresponds to a frequency of

$$\frac{333\times10^3}{2\pi0.7}Hz$$

or 75kHz, where the input voltage is now a sinewave of 700mV, which has a peak of 1V.

This slew rate can be ensured by including an input filter in series with the input of, say, 10kΩ and 220pF shunt to ground.

If the situation is ever likely to occur where a square wave input of 1V is applied, the input time constant should be the slew limit, requiring the capacitor to be increased to 330pF.

If the input resistors are made much larger, such that the differential margin becomes equal to the maximum input voltage, then the margin is not exceeded even for a square wave, **Fig. 3**. Here, 1kΩ resistors have been used as these will give a 1V margin at 1mA.

This is the reason why I recommend much larger resistors than Self – even at the expense of reducing the open loop gain and worsening the overall distortion figures. Of course, a different slew-rate criterion could

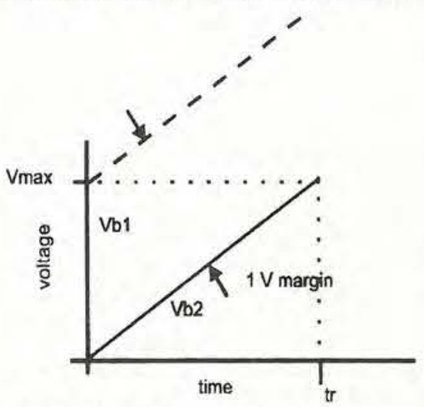

Fig. 3. Input margin for 1kΩ resistors.

be applied other than the 50% change in current, perhaps based on a certain increase in distortion figure.

Now for the surprise. In equation 5, the amplifier feedback attenuation ratio was included. Otala *et al.* (*op. cit.*) conjectured that a high level of feedback may increase transient intermodulation distortion. Equation 5 shows a simple but significant result – that for more feedback, where α is greater, corresponding to a lower gain, a *faster* input signal can be tolerated.

Perhaps I was slow to appreciate this point before, because this is really just another way of saying that for more feedback, the bandwidth is greater! This would appear then to agree with Cherry's comments about TID[3].

The sound of a power amplifier might have been affected by high-speed preamplifiers, which would undoubtedly influence low slew-margin amplifiers. If a simple input filter is used, it might, with the power amplifier limit, have adversely affect the overall frequency response of the system.

The simplest way out of this would be to use a faster amplifier. Another approach might be to use a two or three-order filter, to try to leave more of the upper audio frequencies untouched before the cut-off point.

It is not a new idea to include a low-pass filter on the input of an amplifier to prevent slew-rate limiting, but I hope this has shown a method for determining the optimum.

Alternatively, large input resistors solve the problem, admittedly with some other drawbacks. Perhaps, though, with either approach, we can lay the ghost of TID to rest.

J.N. Ellis
Tavistock
Devon

References

1. Otala, M, 'Transient Intermodulation Distortion In Transistor Audio Power Amplifiers', IEEE transactions, AU-IS, Sept.1970, pp. 234-239. (See also Lohstroh and Otala, 'Audio Power Amplifier for Ultimate Quality Requirements', IEEE Transactions on Audio and Electroacoustics, AU-21, Dec.. 1973, pp. 546-651.)
2. Self, D, 'Distortion in Audio Amplifiers,' series, *E&WW*, August 1993-March *1994*
3. Cherry, EM," Ironing out Distortion", *EW+WW*, January 1995, pp. 14-20.

Free PCB layout tool

A 'lite' version of the Eagle layout tool from Cadsoft is available off the web for free. Its limitation is that it can produce a maximum board size of 4in by 3in, double sided. I've successfully built a board for the 'Five-chip logic analyser' from *EW* February 1999 using the package.

I printed the layout for both sides on a laser printer and used blue 'iron-on' transfer material to make the board.

I succeeded first time and easily produced line thicknesses down to 0.016in with many tracks passing between IC pads – particularly at the FPGA.

Rob Graham
Via e-mail

Early radar

In the very interesting history article on Blumlein in the March issue of *Electronics World*, mention is made of 'chain-home' radio direction finding, later to be called radar. Apart from the later airborne aircraft interception, or AI, fitted to Beaufighter night fighters, from the earliest days there were 'chain-home low-flying', or CHLF, stations and later the ground-controlled interception, or GCI, version.

Chain-home low flying worked at 200MHz with 25kV on the transmitter valve anodes. The range was up to 20 miles, if the open feeder wire SWR could be reduced to around 1.01:1, but on occasions of temperature inversion, second-trace reception was possible with ships plotted at up to 400 miles.

Domestic thermocouples?

Can someone tell me how thermocouples work? From time to time I change our domestic ones, but I do wonder how the minuscule voltage of around 12mV can do the work of opening and holding open a gas valve?

I have asked a couple of professional installers of heating equipment. One said he didn't know the other that it transmitted heat.

Graham Cox
Bexhill-on-Sea
West Sussex

I'm not an expert on this, but the 'thermocouples' you find used to control gas appliances, etc., are not thermoelectric. I've always assumed that they rely on expansion, or state change, of some material within their copper tubing. The only ones I've seen have a plunger that activates an electrical push switch or displaces a mechanical valve. Does anyone know how they work? Are thermocouples ever found in domestic appliances? Ed.

Dangerous protection

On page 776 of the September 1999 issue, the 'Overvoltage load protection' circuit betrays a fundamental and dangerous lack of knowledge of MOSFETs.

Including D_1 in the gate circuit of the MOSFET ostensibly to provide 'protection' against reverse voltage is totally wrong. Firstly, MOSFETs of the type shown are characterised for 20V on the gate. There is no restriction as to polarity – only voltage magnitude – making the diode redundant.

More importantly, the inclusion of the diode ensures that the MOSFET cannot be turned off quickly as the charge cannot be removed from the gate except via the leakage current of D_1: this may take hours.

Another point is that in higher voltage circuits, there is nothing to prevent the gate being charged via the drain-gate capacitance to very high voltages – enough to puncture the very thin insulation layer. MOSFET gates must have a resistive gate-source path under all conditions and should be clamped at ±18V.

Come to think of it, I have never seen anything in your journal that describes in detail the behaviour of switching MOSFETS.

Are you lot totally besotted with ancient linear bipolar circuitry and endless rehashes of low-noise amplifiers and such? How about something that is relevant to modern power electronics?

Dale Butler
Dir. Research and Development.
EnerTec Pty Ltd

Not quite sure who "you lot" are Dale, but I would like to say that the topic of power electronics is well covered by power electronics magazines. You're obviously not a long-term reader either, otherwise you would have seen the numerous articles we have carried on power mosfets since their introduction. I am on the look-out though for someone to do an in-depth piece on how the latest generation mosfets perform though. They are quite remarkable relative to the early devices for both linear and switching applications. Ed.

Can anyone throw light on this?

I believe I have managed to overcome some of the problems associated with the electrical replay of cylinders, but I would like to achieve better quality replay by the use of a light beam.

To this end, I am looking for a Philco photo-electric reproduction pickup head, model 41429, produced around 1946 for the replay of 78rev/min discs. It was a light-beam device in a 78 pickup head reflecting off the recording groove back into a mirror also in the pickup head. A jewel stylus was used for the purpose of guiding the pickup.

Does any reader have such a pickup lying on a shelf somewhere? Alternatively, is there anyone out there working on light beam replay? Allied to the feed-screw drive on cylinder machiners, such a device could well give surprising results.

Joe Pengelly
Plymouth

On 'defective colour vision'

In Letters, February 2000, R N Soar asks if 'colour-blind' males find it easier to read the time on a (green fluorescent display) video recorder than a (red LED display) clock-radio. Could the answer depend on the nature of their deficiency?

While it is true that approximately 8% of men suffer from 'defective colour vision' – often incorrectly called 'colour blindness' – these statistics are made up from 2% who suffer from abnormal 'red' perception, and 6% who suffer from abnormal 'green' perception. By comparison only about 0.5% of women are affected by red/green perception problems.

Most colour deficiency is congenital, following the male line. Abnormal 'Blue' perception is very rare, affecting only about 0.01% of the population, both men and women equally.

My son has green perception difficulties, inherited from my father-in-law, but we reckon that the colour is not the main factor. He doesn't see the 'greenness', only the 'brightness' of the green display. Many of these are blue-green anyway, and he can perceive blue OK.

The over-riding factors that determine clarity are the brightness, contrast and character-shape of the display. Colour does not seem to present him with a major problem either way.

A more telling question would have been 'how many people can manage to programme a video-recorder', which in my experience seems to be linked to 'age' and 'gender' rather than colour perception?

Graham Field
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