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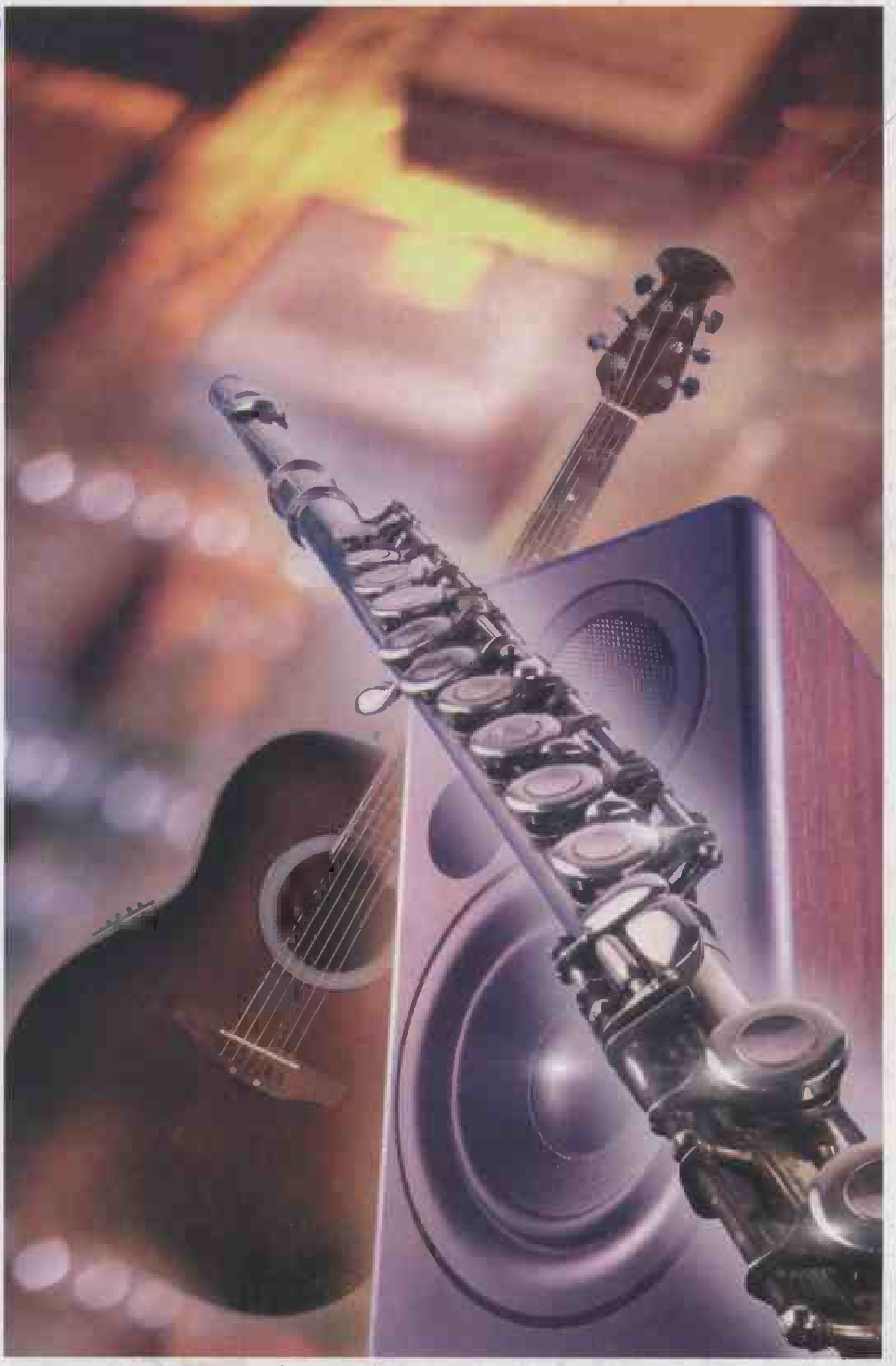

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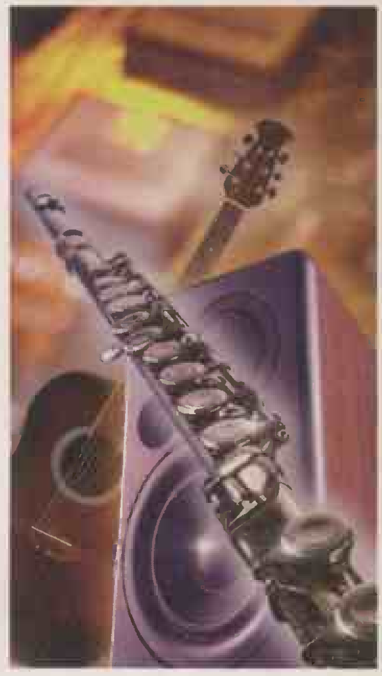
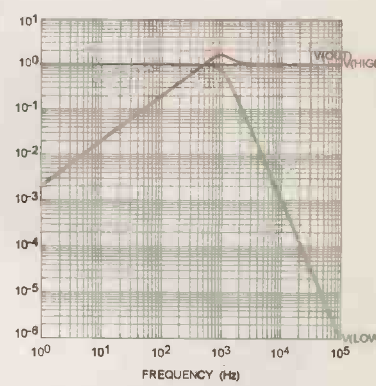
September issue on sale 22 July



Cover photography: **Mark Swallow**

Exclusive EW offer: at last – a simulator with a bread board that looks like a bread board. You can buy it, together with an innovative drag-and-drop system simulator for just £112, fully inclusive. See page 636.



A high-performance active speaker crossover whose components are easy to work out is part of a bumper letters section starting on page 679.

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REED BUSINESS INFORMATION

Electronics in Yugoslavia – 8 years on

I wrote my first piece on this topic for *EW* in the summer of 1990 when Slovenia was still part of Yugoslavia and I could not express myself freely for the fear of the consequences.†

In the meantime Yugoslavia has disintegrated. The Slovenes, Croats, Bosnians, and Macedonians as well as the minorities in Kosovo and Vojvodina expected a compromise arrangement in our common State. The Serbs on the other hand wanted a system that would be optimal for them exclusively. Under such circumstances no sound agreement was possible, so Yugoslavia fell apart.

The economic struggle between the Serbs and all the other nations affected electronics in former Yugoslavia to a great extent. On top of this the communist system favoured reliable leaders over competent ones.

It was almost impossible for anyone who was not in the communist party to get a responsible leading position. The party had spread its damaging influence in our industry and economy like a cancer. For example the Iskra Commerce, a subsidiary of the Slovenian enterprise Iskra – then with over 30 000 employees – had two objectives. Firstly it should sell our products at home and abroad, and secondly it should purchase electronics material and equipment abroad.

Unfortunately, purchasing materials became the main goal, because of the hefty commissions paid in hard currency.

Anyone who criticised such policy or some other grave mistakes of the leadership was punished in an unusual way. Since under our 'workers' self-management' policy it was almost impossible to fire anyone, dissidents were technically employed, but they were given no work to do. In my case this meant two years of being regularly on the job from 6 am to 2 pm, but without being given any work to do.

Although Slovenia was only slightly affected by the war during the disintegration of Yugoslavia, the electronics industry in Slovenia suffered greatly during this period. Big enterprises like Iskra, which was based in Slovenia, fell apart, remaining only some small subsidiaries to survive.

Three years ago I testified before the *Special Parliamentary Committee for Investigating the Reasons for the Disintegration of Iskra*. As with other big frauds in Slovenia this investigation remains unresolved and the lack of urgency for its resolution does not appear of great concern.

However, the Gorenje company in Velenje, which focussed on R&D, has survived and is flourishing – especially in producing and selling domestic appliances both here and abroad. Young and entrepreneurial people are establishing new firms and introducing new products. Unfortunately they do not get much help. The State turns a blind eye to the high interest rates by the banks.

The leading people in the Government and the banks are those who had erstwhile sworn on Karl Marx, but now have in mind only 'Das Kapital'.

The destruction since 1991 due to the war in Yugoslavia has made the situation in other former Yugoslav republics worse than in Slovenia. As the NATO planes continue to pound Serbian industry into dust and ash, at least one generation will probably be needed to restore their national economy and electronics industry to the status of 1990.

Peter Starič Ph. D., Ljubljana, Slovenia

†*Electronics World* January 1991

In the former Yugoslavia money was printed in Belgrade. Whenever the Serbian economy needed more money, they simply printed it. The rest of Yugoslavia, mostly the more hard working republics Slovenia and Croatia, kept covering it. After the break-up of Yugoslavia, the Serbs could not break this habit fast enough. The result was that the bank note shown was not worth a box of matches at the peak of the inflation. In 1993 the Serbian 'banker' Avramovič stopped this nonsense by fixing the exchange rate at 1 Dinar (YUD) for 1 German Mark (DEM) and issuing new bank notes. Before NATO bombing, the real rate of exchange grew already to some 10 YUD/DEM, but it must certainly be worse by now.

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Spectrum allocated for catching crooks

The government plans to free up a chunk of radio spectrum for tracking systems that will locate stolen cars and industrial equipment.

The Radiocommunications Agency has issued a consultation document following an approach from companies keen to use asset tracking for new applications such as home security.

Ram Mobile Data, which already uses its network for tracking fleet vehicles and industrial equipment, welcomed the news. "If there is more spectrum enabling different forms of solutions and combining different technologies then obviously there will be more opportunities for everyone," said Nick Mclean, Ram Mobile Data's director of marketing.

Mclean believes additional spectrum will also help promote telemetry applications such as the transmission of household meter readings over a wide area network.

Regarding the tracking of household goods, his view is that this is still some way off. "It's all about achieving consumer price points – at the moment we are focused on corporate users," added Mclean.

Internet component sales threaten traditional distribution

The Internet poses a direct threat to the component distribution industry, according to Dick Skipworth, chairman of Memec.

"It's a real threat," says Skipworth, "it makes isolated dealing with individual customers more difficult. If companies offer components on the Web they're offering them worldwide."

"We're going to speed up our effort in the area," added Skipworth, "the reason we've not done anything is because we've dealt in advanced components rather than commodity. But, for anything that can be sold by a part number, the Internet is a real threat."

Agreeing , John Dickinson, executive vice-president at TTI said: "The combination of electronic data interchange (EDI) and the Internet is formidable."

But according to Steve Kaufman, president and CEO Arrow: "Some customers will always want to conduct an auction for their order."

The first distribution start-up company aimed at selling components exclusively on the Internet, Microcom Technology, received \$20m funding last month to set up its Partminer database on a Web site.

Chip manufacturers have set up Web sites for direct sales. Micron is selling modules on the Web and this month Samsung starts doing the same.

At Hitachi, Matthew Trowbridge, said: "The difficulty with Internet trading is that there are no standards, which is why we use EDI, where there are standards. But the Internet has to be watched."

Intel said it found EDI expensive to set up and had moved its major accounts to an Internet-based e-commerce system, which currently accounts for \$1bn worth of sales a month.

"Anyone who's a global player has to have availability on the Internet," said Skipworth.

Free Internet calls from BT?

BT may be on the verge of offering free local calls for Internet users.

The company is believed to be in talks with Oftel, the telecoms regulator, with the aim of reducing Web access costs.

Various options are under discussion, but one possibility is a flat monthly fee, giving unlimited Internet access in return. BT refused to comment on the reports.

Internet service provider America Online is already testing a free call service. It uses free-phone 0800 numbers and charges a flat fee of around £15 a month.

So far the company has not decided whether to offer the service to all its subscribers.

The option of paying for calls but no monthly fee was pioneered in the UK by High Street retailer Dixons with Freeserve. Dixons' move quickly established it as the UK's biggest service provider with 1.5 million users.

NetBuy eyes EU

Online component sourcing site NetBuy, which already boasts over 300 000 products, has its eyes on the European market.

"It's in our plans," said Jim Moriarty, the company's v-p of sales and marketing.

NetBuy has agreements with over 60 distributors to sell their parts. Each night they up-load their available inventory onto the site.

Engineers and buyers register their requirements and get the cheapest quote. The order is then processed within 24 hours through the distributor.

Moriarty said there is now over \$2bn worth of inventory on the site. Average order value at present is \$500 "and growing", he added.

Garden project... Hitachi has teamed up with UK-firm Friendly Machines to produce what they claim is the world's first fully robotic lawnmower. An H8/3040 microcontroller supplies the brains and with the help of ultrasonic proximity and impact sensors the robot cuts an entire lawn avoiding obstacles such as flower beds, ponds and garden furniture.

Mobile phone health study shows emissions differ by 20 times

Radiation absorbed by the human brain from different makes of mobile phone can vary by a factor of 20, according to a recent BBC Panorama programme.

Panorama commissioned a study from the National Physical Laboratory to measure radiation absorbed by the brain from eight models of mobile handset.

Specific absorption rate (SAR), measured in watts per kilogram, varied dramatically, from 0.02W/kg for Motorola's Star Tac 70 to 0.44W/kg for a Nokia 2110 – over 20 times higher.

The different results are attributed to the positioning of the mobile's antenna.

Ericsson also fared relatively poorly in the tests. Its GA628 phone had an SAR of 0.26W/kg. An Ericsson spokesperson said:"We can't comment on research that hasn't been published yet."

All the phones tested were well within the UK safety limit of an SAR of 10W/kg. "How one phone differs from another is not important," said Tom Wills-Sandford of the Federation of the Electronics Industry, speaking on behalf of all the mobile manufacturers.

Research has so far found little to link mobile phone use with adverse health effects. Dr Alan Preece from the Bristol Royal Infirmary has carried out the UK's only independent study into short term effects of mobile phone use. His results showed no apparent side effects from mobile phone use.

"If people want to reduce exposure, they should use a hands-free device," said Preece.

Lead ban hits PCB makers

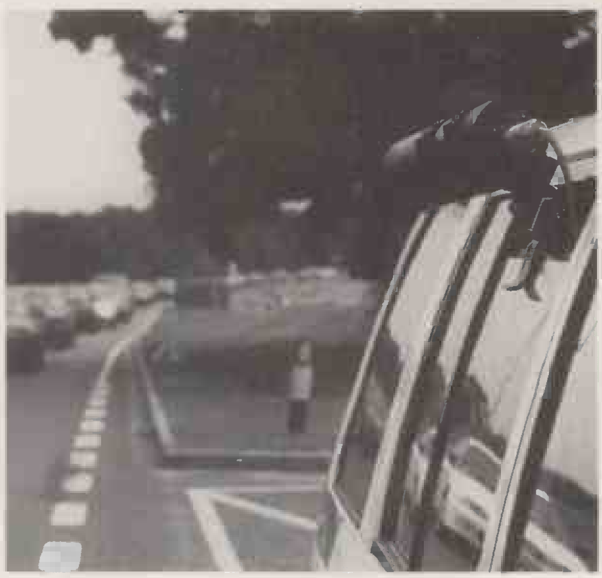
The proposed European ban on lead in solder has been attacked by the PCB industry which is calling for a delay in implementation.

The PCB industry body the European Federation of Interconnection and Packaging wants an outright ban on lead to be delayed and the dumping of scrap electronic products in landfills to be outlawed instead. It wants a co-ordinated international investigation to find a "truly viable lead-free alternative."

The EU ban is to come into affect in 2004.

A DTI report analysing the current status of lead-free soldering confirms that while it is technically possible there is currently no drop in replacement for tin-lead.

Long claw of the law... Racal Messenger will unveil next month its latest generation of Talon, the automatic licence plate recognition system. Using pattern recognition and neural networks, Talon can identify number plates 24 hours a day in a variety of weather conditions. It is said to be 95 per cent accurate. In other words, one in 20 plates are read incorrectly, which is nice to know while being arrested for ram raiding.



Engineers' salaries rise

Engineering pay settlements are rising despite evidence of some pay freezes, according to the latest survey from the Engineering Employers' Federation (EEF).

The average settlement level for the three months to the end of April was 2.7 per cent, a rise of 0.1 per cent over the previous averaged quarter. Pay freezes made up over 14 per cent of the 223 reported settlements.

Meanwhile, the EEF has called on the government to review its plans for an energy tax. After consulting its member companies, the EEF claims that a majority of manufacturers – even after making energy savings – will pay far more on the tax than they will receive back in National Insurance contributions.

"The energy tax will leave manufacturers at a substantial competitive disadvantage,"
ß

Micromirror tv screen deal

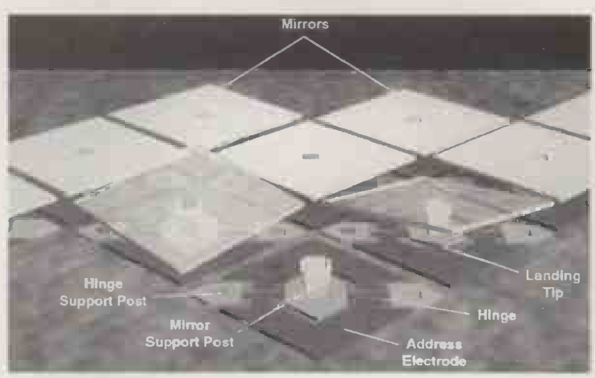
Mitsubishi has signed an agreement with Texas Instruments to use digital light processing (DLP) technology in large screen TVs.

"We have been watching the development of DLP technology for many years and have been impressed with the dramatic performance improvements that have been made," said Yoji Otani, chief executive of Mitsubishi Digital Electronics America.

The firm expects to start manufacturing products based on DLP before the end of next year. TI's DLP technology is based on an array of mirrors fabricated using micromachining onto a silicon chip. Each mirror can be moved, and so acts as a switch. Light hitting the chip is reflected to either the front or rear of a screen.

Mitsubishi will use the DLP technology in high definition rear projection screens.

TI signed a similar deal recently with Hitachi.



Mirror, mirror on the wall... TI first developed its micromirror device technology back in 1994. The technology is based on an array of mirrors fabricated using micromachining onto a silicon chip. Other licensees have included Nokia.

Robots that mimic insects

David Attenborough eat your heart out. A whole new insect world is evolving that consists of tiny robots, made using recent developments in silicon chip production. Tom Foremski keeps his fly swat handy

US researchers are working on the development of tiny robots modelled on insects as part of a Pentagon research project.

Called biomimetic robots, the idea is to create small robots that mimic actual insects in structure and behaviour. Researchers believe that such small robots can be cheaply manufactured in large numbers and will be substantially more compliant and stable than larger robots.

Such small robots can be constructed using recent developments in silicon chip production and silicon based micro-electro-mechanical systems, or MEMS. This research into biomimetic robots is being sponsored by the US Office of Naval Research with obvious military applications. Tiny flying robotic insects could be used to gather battlefield information, but there are also many commercial spin-offs in agriculture, pollution monitoring and detection, and other uses not yet imagined.

Researchers at the universities of Berkeley, Stanford, Harvard and Johns Hopkins are working on the biomimetics robot program. One goal of the program is to gain enough biomechanical insight and manufacturing expertise to allow the construction of a 'robotic cockroach' suitable for autonomous operation in various environments. Unlike larger robots that have slow and very precise movements, insect robots would move more quickly, but also more erratically, making them well suited to move over difficult terrain at high speed.

Another focus is to study and try to understand the mechanism employed by geckos which can adhere to smooth surfaces such as glass, using millions of tiny hair-like structures. One theory is that these small hair-like structures employ weak van der Waals forces at the micro-scale to produce strong adhesion at the macro-scale. Understanding this mechanism would allow researchers to improve the stability of micro-mechanical insects.

Biomimetics combines work in many different disciplines. Biologists are studying the structure of small insects to determine how their wings flap and how they move around. While electronics engineers are developing tiny sensors and actuators, small power supplies and solar cells, and motor control systems.

The initial goal is to use off-the-shelf technologies and integrate newer technologies as they become available over the course of the research project. One of the key focuses is on development of actuators that can move insect-sized legs and wings.

Ron Fearing, associate professor at the University of California at Berkeley is leading one of the biomimetic project teams constructing a micromechanical flying insect. "We've made some progress in terms of actuators with the use of single crystal piezoelectric activators. This is an approach other teams are pursuing with some success."

Fearing is basing his project on the blow fly, a relatively large fly that offers some hints on how a robotic fly might work. "We are studying the structure of insect bodies but we cannot copy them, they are too complex. We've simplified the design to make it achievable," says Fearing.

So far, Fearing's team has constructed a large model of the wings and is working on developing MEMS and activators that will flap the wings. The power source will be a small solar cell that will gather enough light to produce powered flight. The wings, constructed of polycrystalline silicon will be about two centimetres from wing tip to wing tip.

Fearing's team is using computer based finite-element methods (FEM) in combination with simplified fluid models to guide the design of wing shape and actuator placement. Mock-up models using conventional actuation such as voice coils are used to provide experimental verification of the aerodynamic effects of the wing shape and stroke. Because these aerodynamic computations are very complex, Fearing's team is also studying tethered flying fruit flies to gain some insight into the mechanisms used in actual insects.

In terms of powered flight, Fearing estimates that if the robotic fly system can generate 100μW, that should be enough to allow it to fly. A tiny gyroscope would be used for attitude control and stability along with a rudimentary optical sensing system.

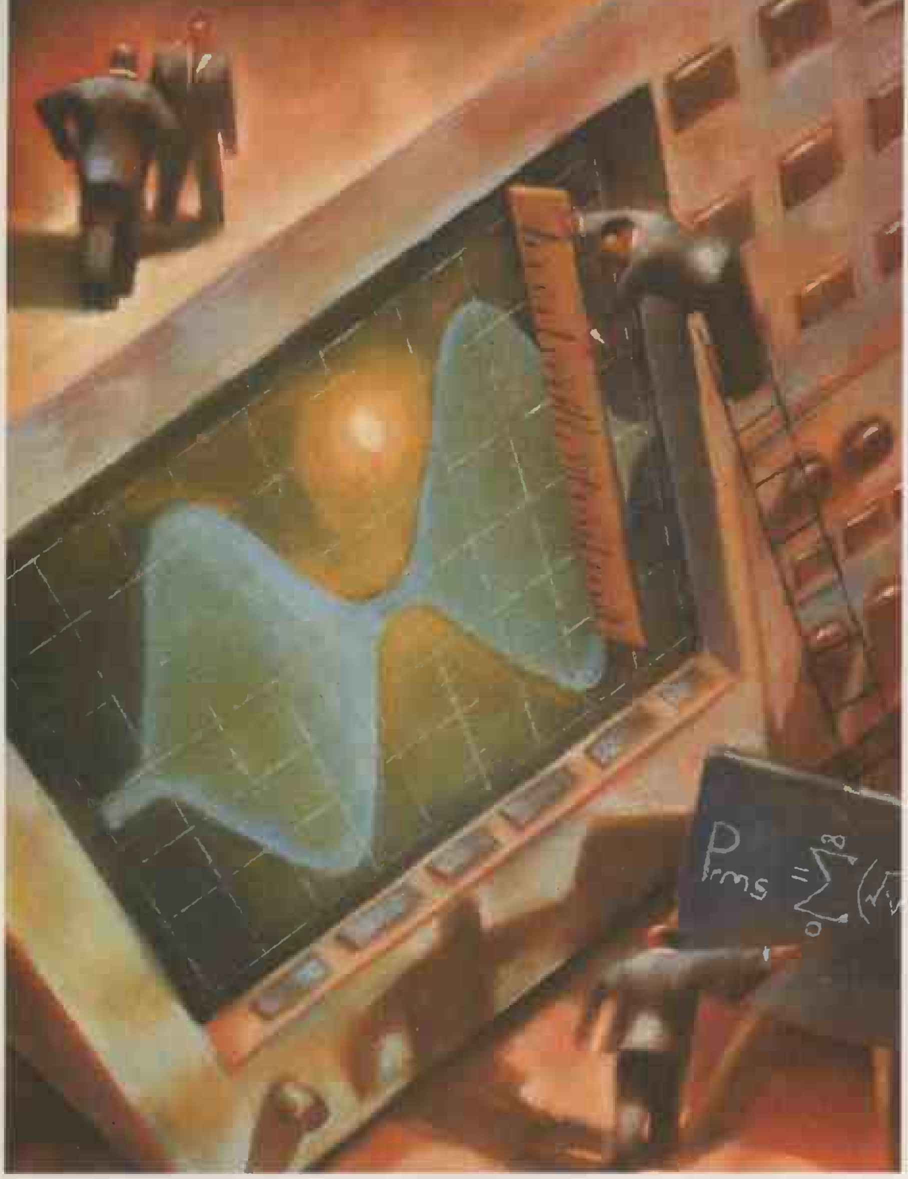
Other teams are using birds and mammals as models for their flying robots. Stanford researchers, for example, are working on a flying robotic bat, while others are studying humming birds and dragonflies.

"We are still several years away from a working robotic fly," notes Fearing. "But we will get there and there will be many applications. The key thing is that you could make these robots very cheaply, say $10 each, which means you could use them in large numbers and collect information over large areas."

The Pentagon would love to have tiny fly-like robots for battlefield intelligence collection. Even if the enemy employed a low-tech fly swatter counter measure, the ability to launch swarms of such devices would still prove effective.

Fearing also envisages uses in agriculture. Robotic flies could gather information about soil conditions, pests and other factors. And they could be used to monitor areas for pollution offering an early warning system, and also tracking down sources of pollution.

And there are also potential applications in space exploration. Planets such as Mars, with a thin atmosphere and low gravity would provide the robotic flies with enough lift to be able to explore large areas without risking humans, and reducing payloads in exploratory probes. ■



Conventional total harmonic distortion meters often suffer from a measurement range limited by the noise floor of the instrument. Ian Hickman's new design provides a noise floor in a 20kHz bandwidth of 0.0009% for THD measurements – and a noise floor of typically 0.000025% when used in conjunction with an audio-frequency spectrum analyser.

Measure thd below 0.001%

Over the years I have made a number of distortion meters and low-distortion oscillators. Measuring the latter with the former can be a rather chicken and egg situation. Is the distortion meter actually measuring the distortion of the oscillator, or is that so low that what you are seeing is the distortion in the distortion meter itself?

So when in 1993 I developed a very low distortion oscillator, later published in these pages,[1] a radical approach was adopted.

My then distortion meter was connected to the output of the oscillator via a special filter box. This contained a passive twin-tee filter, followed by an active second-order elliptic high-pass filter. The cut-off frequency and Q of the high-pass filter were dimensioned so that its stop-band contributed further attenuation of the fundamental, while the combined response was sensibly flat from twice the notch frequency upwards. Thus all harmonics were equally passed to the distortion meter, which, relieved of the presence of the fundamental, itself contributed nothing to the measured distortion.

The only problem was that a separate twin tee and associated high-pass filter were needed for each frequency at which a distortion measurement was to be made, since each twin tee and filter was not tuneable.

The conventional method

The usual sort of total harmonic distortion meter uses a tuneable notch circuit. It is often based on the Wien Bridge or on an all-pass circuit. These types of notch circuit are convenient, being tuneable with just a two-gang component such as a capacitor or resistor.

At frequencies far from the notch, the transfer function is unity, i.e. the response is flat. However, as the notch frequency is approached, the attenuation starts to rise appreciably, so that for example the response of a twin tee circuit at twice – and half – the notch frequency is –9dB.

A notch circuit based on the Wien Bridge is rather better, being about 7dB down at the second harmonic. But it is obviously still necessary to include the notch filter in a negative feedback loop, to bring the response at the second harmonic up to roughly the same as at the higher harmonics. This has the unfortunate effect of greatly increasing the amplitude of the fundamental at a point internal to the filter. Therefore, the signal amplitude that can be applied to the circuit must be limited to quite a low level, to avoid distortion.

A new distortion meter

Recently, my eye was caught by details in the technical press of a very low distortion quad op-amp from

Burr-Brown – the *OPA4134*. Single and dual versions of the same device are also available.

It was clear that in view of its very low distortion, the device would be most appropriate for the design of a new distortion meter to replace my existing instrument. As you can see from the data in the panel entitled 'The *OPA134* family', the device's distortion is typically 0.00008% over the greater part of the audio range, when driving a 2k load at unity gain. Even at 20kHz, its THD is still a respectable 0.0004%.

The THD meter design presented here makes use of this device in a state-variable filter configuration. In particular, the state-variable filter is arranged not as a conventional notch filter, but as a two-pole elliptic high-pass filter. With this arrangement, the circuit can operate at a reduced internal Q. This in turn means that a much larger signal can be applied to the circuit, increasing the dynamic range, since the noise floor is then much lower relative to the fundamental.

Design of the new instrument commenced with modelling to determine the best parameters of a two-pole elliptic high-pass stage for the purpose.

A special kind of filter

The specification for the filter differs from that of a conventional filter. In a conventional elliptic filter, the parameters of interest are A_s, the minimum stop-band attenuation, and Ω_s, the frequency at which this is reached, relative to the last point Ω_p at which the pass-band response passes through the ripple depth.

For this application however, the final attenuation A_s is of no interest. The crucial factor is that the zero in the response should lie at $0.5\Omega_p$. Thus when the fundamental is notched out, the response has already returned to the 'flat' level by the second harmonic.

I used Mathcad for the modelling, the only parameters available as variables in a two-pole filter being A_s and the Q of the second-order section. These were refined – or, to be frank, juggled – to obtain the flattest response from twice the notch frequency upwards.

Figure 1 shows that I finished up by combining 0.6 of the low-pass output with the full high-pass output; i.e. A_s is 0.60 or –4.44dB. Combining this with a damping D of 0.76

THD meter using HP elliptic notch

$w := .01, .02 .. 10$ $\qquad x := 1.29098$

$$f1(w) := \frac{.6 + \left[\left(\frac{w}{x}\right) \cdot j\right]^2}{1 + 0.76 \cdot \left[\left(\frac{w}{x}\right) \cdot j\right] + \left[\left(\frac{w}{x}\right) \cdot j\right]^2} \qquad \text{notch response}$$

$z1(w) := |f1(w)|$ take modulus $\qquad r1(w) := 20 \cdot \log(z1(w))$ convert to dB

$$f2(w) := \frac{1}{1 + 0.76 \cdot \left[\left(\frac{w}{x}\right) \cdot j\right] + \left[\left(\frac{w}{x}\right) \cdot j\right]^2} \qquad \text{lowpass response}$$

$z2(w) := |f2(w)| \qquad r2(w) := 20 \cdot \log(z2(w))$

$$f3(w) := \frac{\left[\left(\frac{w}{x}\right) \cdot j\right]^2}{1 + 0.76 \cdot \left[\left(\frac{w}{x}\right) \cdot j\right] + \left[\left(\frac{w}{x}\right) \cdot j\right]^2} \qquad \text{highpass response}$$

$z3(w) := |f3(w)| \qquad r3(w) := 20 \cdot \log(z3(w))$

$$f4(w) := \frac{.6}{1 + 0.76 \cdot \left[\left(\frac{w}{x}\right) \cdot j\right] + \left[\left(\frac{w}{x}\right) \cdot j\right]^2} \qquad \text{60\% of lowpass response}$$

$z4(w) := |f4(w)| \qquad r4(w) := 20 \cdot \log(z4(w))$

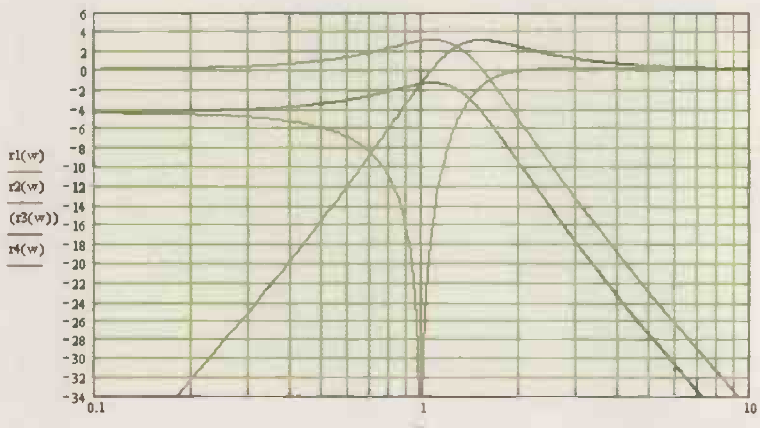


Fig. 1. Design of an elliptic high-pass filter with minimum stop-band attenuation A_s of 4.4dB.

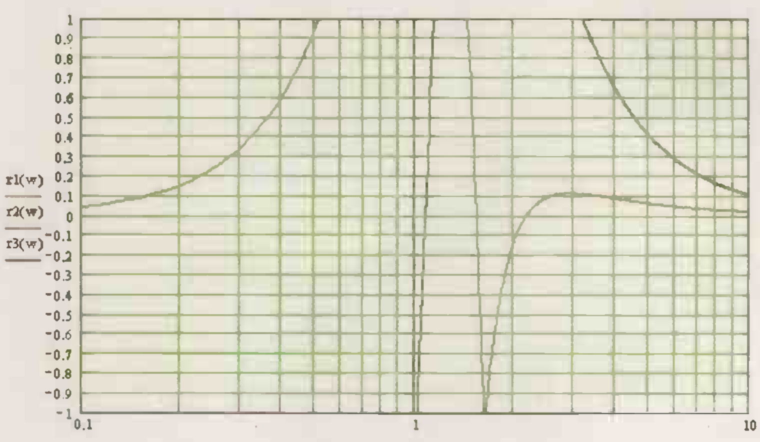


Fig. 2. Expanded view of plot of Fig. 1, showing pass-band gain of within ±0.15dB from twice the notch frequency upwards.

Traditional THD meter

$w := .01, .02 .. 10$ $\qquad x := 1$

$$f1(w) := \frac{1+\left[\left(\frac{w}{x}\right)j\right]^2}{1+0.33\cdot\left[\left(\frac{w}{x}\right)j\right]+\left[\left(\frac{w}{x}\right)j\right]^2} \qquad \text{notch response}$$

$$f2(w) := \frac{1}{1+0.33\cdot\left[\left(\frac{w}{x}\right)j\right]+\left[\left(\frac{w}{x}\right)j\right]^2} \qquad \text{lowpass response}$$

$$f3(w) := \frac{0+\left[\left(\frac{w}{x}\right)j\right]^2}{1+0.33\cdot\left[\left(\frac{w}{x}\right)j\right]+\left[\left(\frac{w}{x}\right)j\right]^2} \qquad \text{highpass response}$$

$$f4(w) := \frac{0+\left[\left(\frac{w}{x}\right)j\right]^1}{1+0.33\cdot\left[\left(\frac{w}{x}\right)j\right]+\left[\left(\frac{w}{x}\right)j\right]^2} \qquad \text{bandpass response}$$

$z1(w) := |f1(w)|$ take modulus (amplitude response)

$z2(w) := |f2(w)| \qquad z3(w) := |f3(w)| \qquad z4(w) := |f4(w)|$

$r1(w) := 20\cdot\log(z1(w))$ convert modulus to dB

$r2(w) := 20\cdot\log(z2(w)) \qquad r3(w) := 20\cdot\log(z3(w)) \qquad r4(w) := 20\cdot\log(z4(w))$

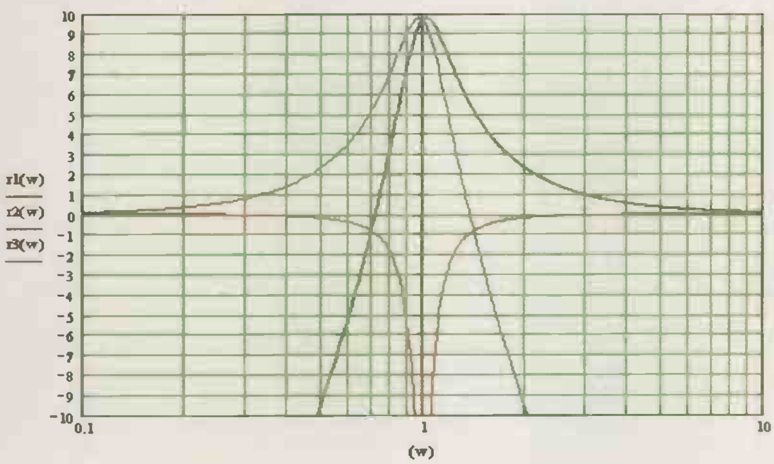


Fig. 3. In a design where the notch coincides with the resonant frequency, A_s=0dB, large internal peaking is needed for a flattish response from twice the notch frequency upwards.

(Q=1/D=1.32) resulted in the notch response shown in red in Fig. 1.

Also shown in the same diagram are the high-pass response (black), the low-pass response (blue) and 0.6 times the low-pass response (black). You can see that the notch occurs where the attenuated low-pass response crosses the high-pass response. At this frequency, the two responses are equal in amplitude, and at all frequencies the low-pass and high-pass responses are in antiphase, hence cancellation occurs.

Figure 2 shows an expanded scale version of the response, showing that at the second harmonic the response is in error by less than –0.15dB, and never exceeds +0.11dB before settling to unity at high frequencies. Note that, for the convenience of the plot, the normalised frequency variable has been offset to make unity frequency coincide with the notch.

The true normalised frequency for the second-order stage is where the high-pass and low-pass responses cross, i.e. the maximum response of the unused band-pass output.

Figure 3 shows that using a conventional notch, a Q of 3 would be necessary to achieve even –0.2dB at the second harmonic, a higher Q still being necessary to equal the –0.15dB performance of the chosen circuit.

The circuit design

Armed with the required filter response, the design of the circuit commenced.

The first problem was some kind of input attenuator, so the instrument could handle a range of different amplitude inputs. As a stand-alone

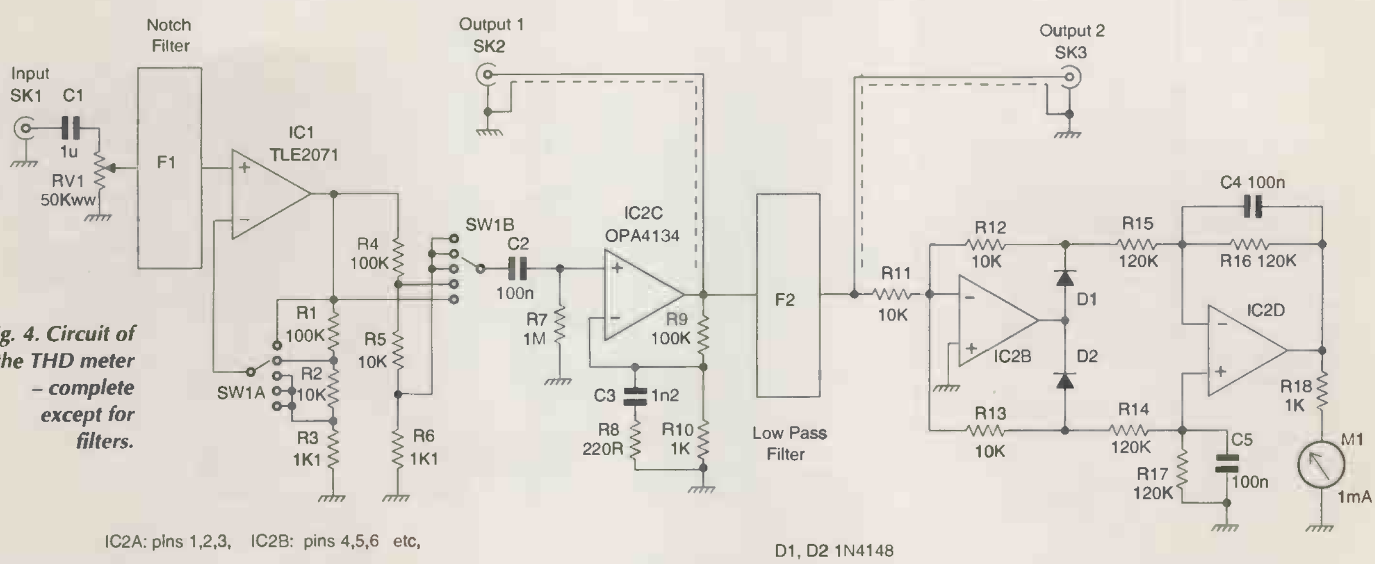


Fig. 4. Circuit of the THD meter – complete except for filters.

The *OPA134* family

The *OPA134* and the dual and quad members of the family are low noise low distortion unity gain stable FET input op-amps. They additionally feature a high slew rate of 20V/μs, a high open-loop gain of 120dB and operate on supplies of up to ±18V.

With an output that can swing to within a volt of either supply rail, they provide exceptional headroom of +23.6dBμ re. 0.01% THD+N, and thus a remarkable dynamic range. **Figure A** shows the THD+N (total harmonic distortion plus noise) at two values of closed loop gain, for two different values of output load.

At unity gain, the distortion is less than 0.0001% or below –120dBc, and measurements at this level are rightly considered as demanding, and requiring considerable care. However, an ingenious circuit arrangement is possible, whereby the op-amp under test is arranged to amplify its own distortion by a known factor relative to the fundamental, considerably easing the measurement problem.

This is shown in **Fig. B**, where setting R_1 open circuit with a 100:1 ratio between R_2 and R_3, the relative level of the distortion and noise is increased by a shade over 40dB. This brings the required distortion measurement to around the 0.01% level, a much more readily measurable figure.

The excellent distortion level at 3V rms output shown in Fig. A applies at ±15V. Elsewhere the data sheet shows the small signal THD+N as decreasing with increase in supply voltage, up to the permissible maximum of ±18V.

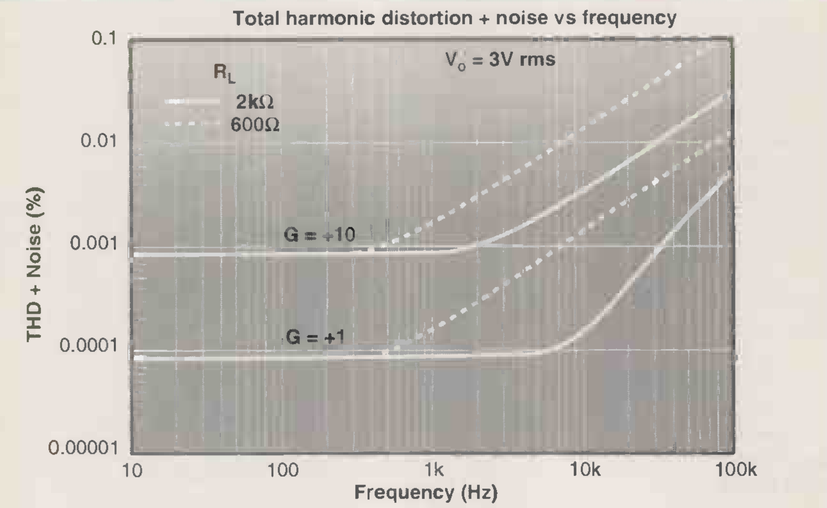
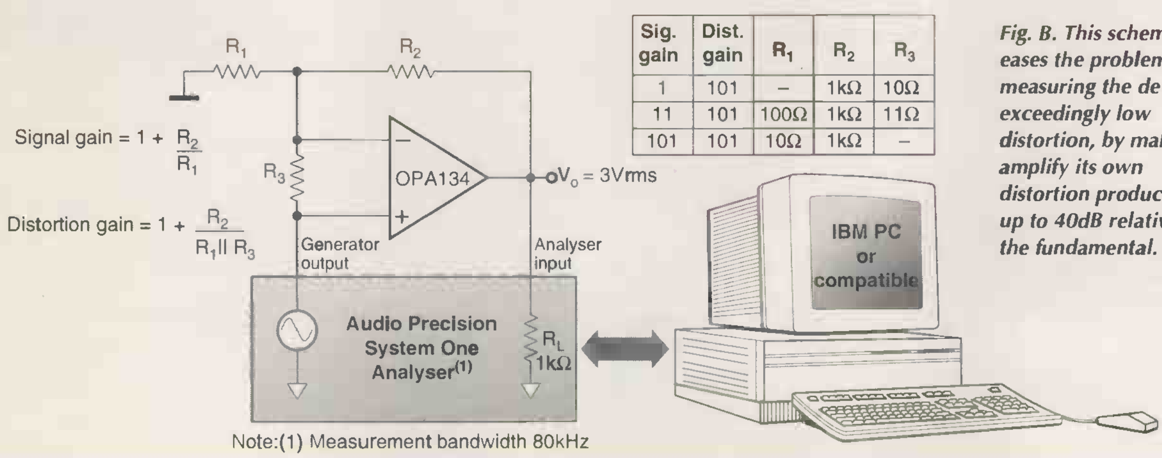


Fig. A. Distortion performance of the Burr-Brown OPA4134 *at 8.5V peak to peak output, for two values of closed loop gain at two different levels of output load.*

While not specifically stated in the data sheet, it seems likely that a lowered level of distortion would also be observed at the 3V rms level of Fig. A when using ±18V supplies.

An interesting corollary of the method of measuring the THD+N of the device as in Fig. B is that the result includes the distortion due to the common-mode signal present at the input. A common-mode component is of course present whenever the op-amp is used in the non-inverting mode. It can be the source of additional distortion, as the P-channel JFETS in the input stage exhibit an input capacitance which varies with the applied input common mode voltage. Burr-Brown therefore recommends that if the source impedance exceeds 2kΩ at either input terminal, the impedance at the other be padded up to match.

This distortion mechanism is of course completely absent when the devices are used in the inverting configuration. For this reason, the four op-amp state variable notch filter used in the distortion meter which forms the subject of this article, uses op-amps in the inverting connection only.



Sig. gain	Dist. gain	R_1	R_2	R_3
1	101	–	1kΩ	10Ω
11	101	100Ω	1kΩ	11Ω
101	101	10Ω	1kΩ	–

Fig. B. This scheme eases the problem of measuring the device's exceedingly low distortion, by making it amplify its own distortion products by up to 40dB relative to the fundamental.

Fig. 5. Circuit of modified elliptic high-pass filter to suppress fundamental (top), 22kHz Sallen and Key low-pass filter (lower left) and power supply arrangements. Non-standard values, e.g. R_{22}, C_{12} were made up by series or parallel combinations of standard values.

THD meter, it would be necessary to adjust the meter to read exactly full-scale deflection when the notch was not present, so a continuously variable control was indicated.

Variable controls can be noisy, so a stepped attenuator was considered. However, this would still require a variable control somewhere, to set full-scale deflection, and there would be more than enough knobs on the front panel anyway. So a I chose a good quality 50kΩ wire-wound potentiometer, giving stepless control and no noise.

The circuit as finally developed is shown in **Figs 4** and **5**. The input attenuator is followed by the notch filter F1, which provides ranges from 100Hz, 1kHz and 10kHz downwards, each tuned by two-gang ten-turn 2kΩ wire-wound potentiometer RV_2.

Stage F1 is followed by two op-amps that provide a gain block, switchable in 20dB steps from 0dB to +80dB. As SW_{1A} is rotated from the anticlockwise 100% distortion range, the gain of IC_1 is increased to +20 and then +40dB, giving the 10% and 1% full-scale deflection distortion ranges. Two further steps progressively remove the attenuation from ahead of IC_{2C}, with its fixed 40dB of gain, giving the 0.1% and 0.01% full-scale distortion ranges.

Op-amp choices

An attempt to design the whole instrument using just two *OPA4134* quad op-amps was, fairly predictably, doomed to failure. It proved impossible to get two +40dB gain stages, a 20kHz low-pass filter and an ideal rectifier circuit to cohabit peaceably in a single quad op-amp package – especially given that, for convenience, 0.1in strip-board construction was used, instead of laying out a PCB.

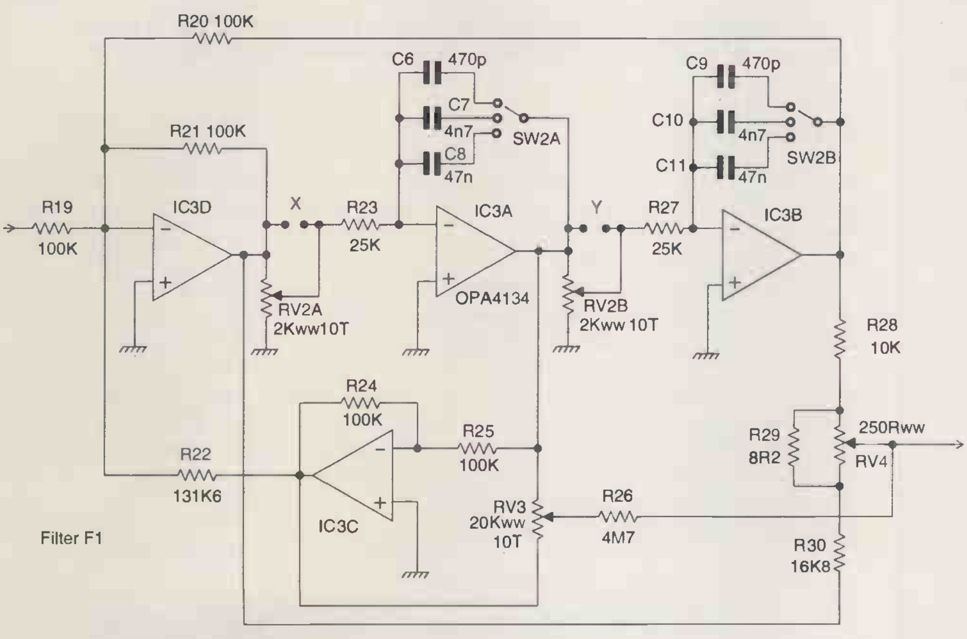
The first +40dB of gain following F1 were therefore off-loaded to a separate op-amp; a *TLE2071* was employed as it was both suitable and to hand. Note that following F1, distortion in following stages is not a consideration, the fundamental having by that stage been suppressed.

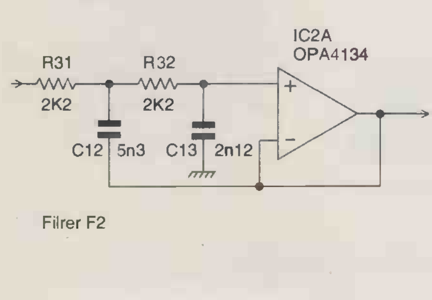
Following the *TLE2071* is a second +40dB gain stage. A second *OPA4134* being available, this was used, although a small amount of peaking (C_3R_8) in the feedback network was needed to maintain a sensibly flat response to 50kHz at the fixed gain of +40dB. There follows a 20kHz low-pass filter F2, using another section of the second *OPA4134*. Both low-passed and unfiltered versions of the residual signal are made available via BNC sockets at the front panel.

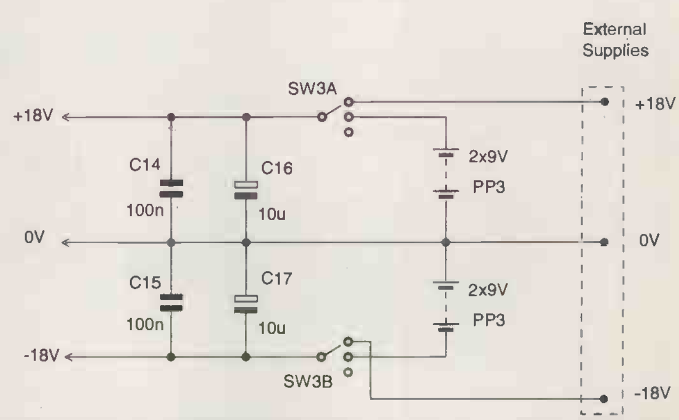
The low-passed residual is applied to an ideal rectifier circuit comprising IC_{2B} and IC_{2D}. This provides a ripple-free dc signal to drive the meter, an essential precaution as the meter was located on the front panel, unavoidably in close proximity to various controls, including those associated with the notch filter.

The 20kHz low-pass filter presented something of a problem, as did signal routeing generally. The bandwidth of the signal ahead of the filter is flat to beyond 50kHz, and with up to 80dB of gain available, the gain-band product is in excess of 500MHz. A vertical screen of tin plate was fitted diagonally across the circuit board, shielding F1 and the controls – other than SW_1 and SW_3 – from the rest of the circuitry.

To avoid high-level signal runs to and from the front panel, the internal meter circuit is driven directly from F2, with no provision to bypass it. This ensured that the instrument was completely stable even at the highest gain setting on the 0.01% full-scale







deflection distortion range, and provided a noise floor for THD measurements of just below 10% full-scale deflection, i.e. about 0.0009%.

Implementing the circuit

The prototype was housed in a metal case of about 20cm wide by 15cm high by 15cm deep. Circuitry was built up on 0.1in matrix copper strip-board, as sold by all the usual electronics suppliers. While useless for RF work, this medium can with care be used at audio frequencies – even for a critical application like this – provided that sensible precautions are taken.

To avoid unintentional coupling between different sections of circuitry, runs should be broken beyond the required connection points, in each direction. The remaining lengths, and any unused strips, can be earthed as and where necessary, to provide an approximation to a ground plane.

Only good quality components should be used – particularly in the F1 section. Some years ago I standardised on 1% metal-film resistors for my component stocks, in view of the small price differential versus other styles. Metal-film resistors were thus of course used throughout. Similarly, wire-wound potentiometers are recommended for RV_{1-4}. The other crucial components are C_{6-11}. These must be polystyrene types.

Initially, I used polyester MKS types at C_8 and C_{11}, but the quadrature-trim control RV_3 ran out of adjustment range on the 0 to 100Hz frequency range. Given the ratio of R_{26} to R_{28}, the quadrature-trim control can only cope with an error of about 0.1°. Hence a tanδ of well under 0.001 is required for both capacitors on each range.

A lower value of R_{26} would cope with poorer capacitors, but the quadrature trim would then not have sufficiently fine resolution to permit distortion measurements down to the 0.001% level or lower.

Using the instrument

Considering its performance, the meter is very easy to use, covering input frequencies from 10Hz to 10kHz in three ranges.

With a 10kHz notch setting, only the second harmonic falls within the audio range, although measurements up to and including the fifth harmonic of 10kHz may be made using Output 1, as described later. After connecting the input signal, the next stage is to set the level corresponding to full scale deflection. There are two ways of doing this.

The first is to set the notch frequency to half the frequency of the input. The input can be measured, if not known already, by first adjusting the main tuning controls SW_2 and RV_2 to set the notch onto the input. The frequency can then be read directly from the ten-turn digital dial operating RV_2.

There will be a very small parabolic error at the half-way setting of RV_2, as the source impedance of the potentiometer appears in series with R_{23} (R_{27}), whereas with the wiper at the top end of the track, R_{23} (R_{27}) see the low output impedance of an op-amp directly. The error can be turned into an even smaller cubic error by fitting 25kΩ resistors at the points marked X and Y in **Fig. 5**.

Knowing the frequency of the input, the notch frequency is set to half the input frequency, so that the latter is on the flat part of the high-pass response. The input attenuator RV_1 is then adjusted to give full-scale deflection on the 100% range. The notch frequency is now set to the same frequency as the input, increasing the post notch gain as necessary with SW_1.

Fine tuning

In addition to the main tuning controls RV_2 and SW_2, a 250Ω wire-wound potentiometer, RV_4, provides an extra fine 'in-phase' tuning control. Additionally, to achieve maximum rejection of the fundamental, a quadrature trim is necessary, and a ten-turn wire-wound pot provides this, RV_3.

The in-phase and quadrature trims are completely independent, and do not interact. This avoids an infuriating 'sliding balance' which is found in some applications, where the trims need to be repeatedly adjusted alternately due to interaction. This frequently results in one or other trim running out of range, necessitating readjustment of the main controls.

If the amplitude of the input signal is known to be independent of frequency, for example the output of an audio oscillator under test, an alternative procedure exists. Here, the notch frequency may be set immediately to the desired test frequency, and the input to twice this frequency. Full-scale deflection is then set as before, and the input frequency then reduced to sit it in the notch.

In the event that the signal under test does not produce full-scale deflection even with RV_1 at maximum, two possibilities remain. Full-scale deflection may be set on the 10% range, the most sensitive range then being 0.1% distortion full-scale deflection, rather than 0.01%. Alternatively, a low distortion preamplifier such as an *OPA134* may be used ahead of the THD meter.

Stability issues

The greater the degree of suppression of the fundamental demanded of the notch, the narrower is its bandwidth. So when measuring very low levels of distortion, a problem may be experienced if the frequency of the signal under test is not sufficiently stable.

There may be a little long-term instability or drift due to the effects of supply voltage or temperature on the sig-

What is total harmonic distortion

Total harmonic distortion is defined in terms of the effective value of the distortion components relative to that of the fundamental, where 'effective' means rms, i.e. root mean square. A general sinewave may be denoted by $E\cos(\omega t)$, where E is the peak voltage and the frequency $f=\omega/2\pi$. A set of harmonically related frequencies can be denoted by,

$$E_n\cos(n\omega t)$$

where n=1 gives the fundamental, n=2 the second harmonic, etc. Then the total harmonic distortion is defined as,

$$THD = \sqrt{\frac{E_2^2 + E_3^2 + E_4^2 + \ldots}{E_1^2}}$$

In practice, the measurement is often made as,

$$THD = \sqrt{\frac{E_2^2 + E_3^2 + E_4^2 + \ldots}{E_1^2 + E_2^2 + E_3^2 + E_4^2 + \ldots}}$$

since the full-scale deflection level is set with the signal itself, rather than just the fundamental component of the same.

If the distortion is not extreme, there is in practice virtually no difference. For example, suppose the distortion is as bad as 10%, consisting solely of second harmonic. Then if E_1=1, then

$$\sqrt{E_1^2 + E_2^2} = 1.005$$

a not very serious error of just 0.04dB, and for lower levels of THD, the error is quite negligible.

So measurements made with a THD meter as described in this article incur no noticeable error due to operating according to the second equation above, rather than the first.

nal source. Measurement is still possible, if a little tedious, as the signal is 'chased' with the fine controls.

Short-term instability, or phase noise, is another matter. As the frequency 'shuffles about', the signal peeps out first from one side of the notch, then the other. I did not observe this particular problem with the low-distortion oscillator used to evaluate the instrument. This was due to the fact that the oscillator was also constructed from good quality components.

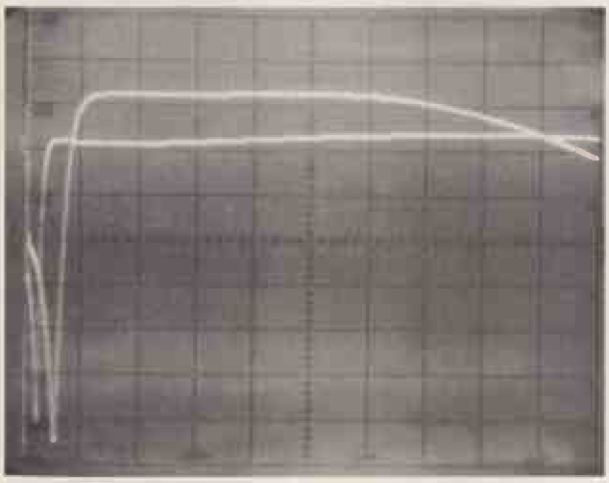


Fig. 6. Frequency response of the unit, with notch set at 1kHz. Upper trace, a), response at Output 2, via the low-pass filter. Span 0 to 20kHz, 100Hz resolution bandwidth, post-detector smoothing minimum, 1dB/division vertical. Lower trace, b), response at Output 1, offset for clarity. Settings as a) except span 0 - 50kHz.

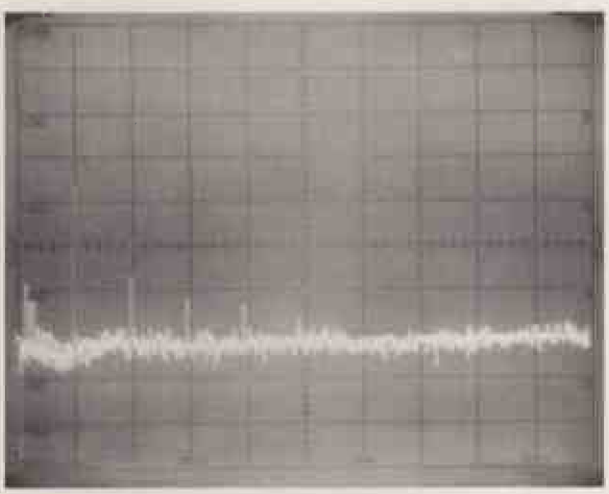


Fig. 7. Spectrum of the output of the low-distortion oscillator of ref. 1. Span 0 - 10kHz, 3Hz resolution bandwidth, post-detector smoothing minimum, scan speed 200s/div., 10dB/div. vertical, THD meter set to 0.1% full-scale deflection range so reference level (top of screen) corresponds to –60dBc. Noise floor –132dBc.

Use of the distortion meter with a spectrum analyser is described later. By connecting an oscilloscope or DVM to Output 2 via a suitable low-pass filter, waveform observations and distortion measurements lower than 0.001% can be made. For instance, for low audio frequencies, a 2kHz cut-off filter can precede the oscilloscope or DVM, extending the measurement range downwards by another 10dB.

Performance?

I evaluated the performance of the prototype as thoroughly as possible by various means. The principle method involved the use of the very low distortion oscillator mentioned earlier.[1] But prior to using this, the frequency response of the prototype was recorded, using an *HP3580A* audio frequency spectrum analyser.

The analyser's built-in tracking generator was connected to the input of the prototype and the frequency response recorded, **Fig. 6**. The upper trace shows the response via Output 2 (SK3), i.e. via the 20kHz low-pass filter.

The analyser settings were 2kHz/div. horizontal, 1dB/div. vertical, 100Hz resolution bandwidth with post-detector smoothing set to minimum. You can see that the notch frequency was set to 1kHz, and that the response is less than 1.5dB down at 20kHz, rolling off at 40dB/decade thereafter. The lower trace – offset for clarity – is via Output 1 (SK2), the analyser settings being as before, except for 5kHz/div. horizontal. The response is flat to within about a quarter of a decibel up to 50kHz.

Next, the output of the low-distortion oscillator, set to 2kHz, was applied to the instrument. The full-scale deflection level was set with the notch at 1kHz, and the oscillator frequency then adjusted to this value. After careful adjustment of the main and fine tuning controls and the quadrature trim, the meter reading on the 0.01% full-scale deflection range was just below one tenth of full scale, corresponding to say 0.0009%.

The 1kHz distortion level of the oscillator had previously been shown to be somewhere below 0.0003%, so it came as no surprise that on switching off the oscillator, the THD meter reading was unchanged. This shows that for THD measurements in a 20kHz bandwidth, only its internal noise floor limits the measurement capability of the instrument.

Extending the meter's range

To see what extension of the measurement range was possible, Output 1 of the instrument was connected to the *HP3580A* spectrum analyser. The analyser was set to 1kHz/div. horizontal, 10dB/div vertical, and the input level set using its input attenuator and 'variable' control to set the response to the fundamental at the top-of-screen 'reference level'.

With the fundamental notched out, the THD meter was set to the 0.1% range. This increased the gain by 60dB, so that the analyser's reference level was now –60dBc. The 0 to 10kHz spectrum was then recorded, **Fig. 7**, with a 3Hz-resolution bandwidth and the post-detector smoothing set to minimum. A sweep speed of 200s/div. was used – definitely a 'come back after lunch' measurement.

The 1kHz fundamental is hiding behind the 1kHz vertical graticule line, but is actually about 36dB below reference level, i.e. 60dB more than this below the fundamental or –96dB. More careful adjustment could have reduced it further, but was not necessary. As long as the residual fundamental level does not exceed the reference level, the analyser will not be overloaded.

The measured levels of harmonics can be seen to be –117dBc second harmonic, –123dB third and –124dB fourth. The fifth harmonic is also just visible, the others being lost in the noise floor, which is at about –132dBc. Other settings of the spectrum analyser can provide an even lower noise floor.

Interpretation of these results brings us back to the chicken and egg dilemma. Summing the contributions of harmonics up to the fifth in Fig. 7 gives a figure of 0.00018% for the distortion level of the low distortion oscillator. This is not greatly different from my estimate, made in 1993, using the twin-tee method outlined earlier. Clearly, in addition to its stand-alone performance, the THD meter forms an exceedingly useful prefilter for use in conjunction with an audio frequency spectrum analyser.

Next, I performed a test to see whether the meter circuit introduces serious error, when using the THD meter as a stand-alone instrument. The meter circuit is average responding, as opposed to true rms responding. Total harmonic distortion is defined in terms of the *effective* or rms value of the distortion components relative to that of the fundamental.

The sine output of a function generator, which I designed some years ago, was used as the test signal. Considered as an audio generator, the performance of this function generator is quite modest, although at 0.03% – as measured on the THD meter – it is distinctly better than the 'sinewave' output usually provided by this class of instrument.

On sine output, the function generator basically filters a squarewave, but using an arrangement that outphases the third harmonic. This leaves a lower level of fifth as the largest single harmonic, **Fig. 8**.

A THD meter with true rms response?

Figure 9 shows the actual waveform of the residual, after the fundamental has been suppressed. It is clearly anything but a sinewave, so would a THD measurement using a true rms responding meter be different?

The sinewave level was set to full-scale deflection on the instrument's internal meter, and the Output 2 connected to a Philips *PM2521* digital multimeter, rms responding on the ac ranges. The reading was 1.2652V.

Next, the notch was tuned to the fundamental on the 0.1% full-scale deflection range, the digital meter's reading then being 0.4071V. Thus the THD calculates as 0.032%, as compared with the 0.3% read on the THD meter.

As a further test, the above was repeated using the function generator's triangular output waveform. The instrument's internal meter gave a distortion reading of 20%, while the DVM read 268mV as against 1.273V at set full scale. These figures correspond to a 21% reading – remarkably close considering that the residual waveform, after the fundamental of the triangular wave was suppressed, looked nothing like either a sine wave or a triangular wave.

Since in both measurements the difference is so small, I decided that the simple average responding meter circuit was adequate. If needed, a true rms derived answer could always be obtained using an external rms responding meter, such as the *PM2521*.

Power supply options

While the *2521* is mains operated, many small bench rms responding DVMs run from internal batteries. Using an average responding meter circuit saves the additional current drain which would have been incurred by incorporating an rms converter chip, thus extending the useful life of the THD meter's internal batteries.

Battery operation using the internal meter avoids any complications due to earth loops, as the earthy side of the THD meter is connected to nothing other than the unit under test.

That said, no great problems were experienced with earth loops; all the results reported here having been taken with the instrument run from ±15V stabilised supplies. However, some hum related components can be seen at the extreme left-hand end of the trace in Fig. 7.

Clearly if one were trying to measure the distortion of a signal at say 250Hz or lower, then battery operation could be the best option. Current consumption of the complete instrument is 35mA, which is within the capabilities of the ubiquitous PP3. If you intend using the meter frequently though, four PP9s would be a better option. Alternatively, four PP3 size rechargeable NiCd batteries, with their 8.4V nominal rating each, could be used.

In summary

The THD meter can be used in a number of ways, for various different purposes.

It can be used as a stand-alone THD meter, run from either internal batteries, or from external ±15 to ±18V supplies. Alternatively, it can be used in conjunction with an external rms responding DVM. Many of these are operated from internal batteries, so this does not introduce any earth loop problems. The meter may also be used in conjunction with an audio frequency spectrum analyser, to greatly extend the measurement range of the latter.

While a THD measurement gives no information as to which is the largest harmonic(s) and a spectrum analyser may not be available, the unit can still provide useful information on this point. The residual from Output 1 or 2 may be connected to an oscilloscope, giving valuable insight to the nature of the distortion.

For instance, in **Fig. 9**, counting the number of positive – or negative – peaks in one cycle of the residual clearly shows that the fifth harmonic is very strong, but is obviously not the only one present. Viewing the residual is particularly revealing when adjusting the bias of a class B audio amplifier for minimum crossover distortion.

The unit may be used as a general purpose notch filter, to suppress e.g. 50Hz contamination on a wanted signal. With the notch set to below the bottom of the audio band, for example 10Hz, the unit may be used as a high gain low noise audio preamplifier.

In summary, the circuit is not overly complicated, and it should be possible to build it for a sum which is not excessive, considering the level of performance offered.

Finally, my thanks to Burr-Brown for supplying data and samples, and for permission to reproduce two diagrams from the OPA134 data sheet. ■

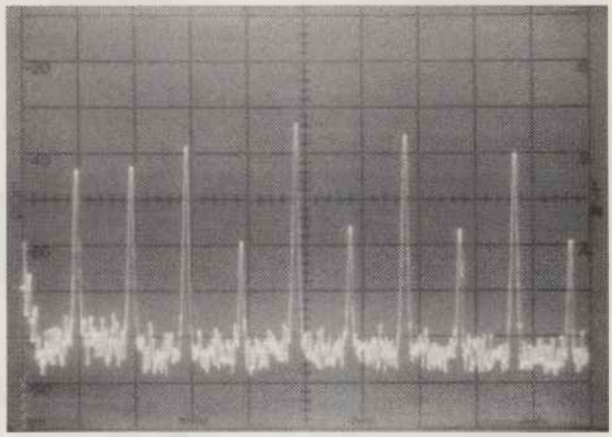


Fig. 8. Spectrum of the sine output of a function generator. Span 0 - 10kHz, 30Hz resolution bandwidth, post-detector smoothing minimum, scan speed 2s/div., 10dB/div. vertical, THD meter set to 1% full-scale range so Ref. level (top of screen) corresponds to –40dBc.

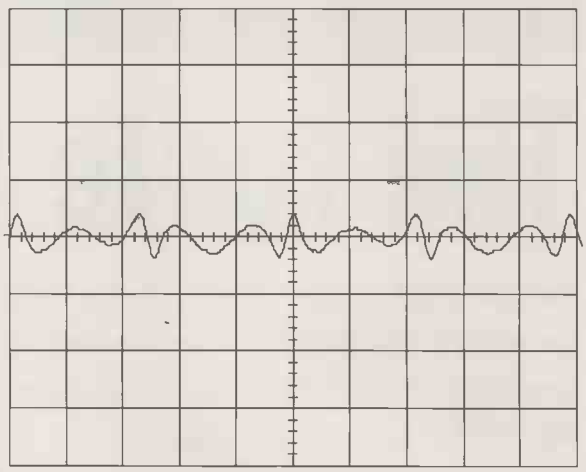


Fig. 9. The residual of a function generator's sinewave output, representing the distortion components left after the fundamental has been suppressed.

Reference

1. Ian Hickman, I, 'Low-distortion audio oscillator,' *Electronics World and Wireless World*, p. 370, May 1994.

Like its modern counterparts, this early attempt at simulation was difficult and expensive to implement – and it didn't always do what its creator wanted it to.

The route to simulation II

Following further advice on what to watch out for when choosing a low-cost circuit simulator for your PC, Rod Cooper investigates version 6 of *CircuitMaker*.

Realising that simulators are used by engineers with experience of the knobs and dials on everyday instruments such as oscilloscopes and signal generators, some simulation programs utilise symbols resembling real instruments.

Both *Electronics Workbench* and *Tina* fall into this category. In this type of program, Spice and its quirks are to some extent hidden from view. The commendable intention is to give the simulator a less hostile face, one which practical engineers can easily identify with.

In some programs, simulations are often arranged by connecting the virtual instruments into the schematic diagram by wiring them in, adding a further touch of realism.

This works extremely well. Programs using these virtual instruments are much easier to understand and operate, although some would say that the ability to tweak and tinker with such a program to get the best out of it has been reduced. It's a bit like the Windows versus Dos argument. Windows is more pre-packaged and digestible, Dos more nuts-and-bolts.

In contrast, other programs have retained the abstract concepts of early simulation programs, and generally speaking these are more difficult to learn and to operate.

Designing on screen

Schematic drawing programs can be divided into three categories by the method they use for handling symbols. The first type extracts symbols one-by-one from the main library and puts them on the screen's drawing-sheet. This is the way several pcb-CAD programs that use schematic capture operate.

Secondly, there is the type that enables the user to assemble a temporary parts-bin of the symbols expected to be needed. These symbols are then readily transferred to the drawing sheet. Again this is best suited to the pcb-CAD type of program.

In both these program systems there is not much to attract the designer who wants to compile experimental circuits on-screen, because going back to the main library for new symbols is a relatively slow process – parts bin or no parts bin. It is most unlikely that the designer will know in advance what symbols will be needed.

However, if the parts-bin type of program permits you to permanently pre-

Review subjects

This first review covered ***Electronic Workbench*** version 5.12, whose maker is IIT Ltd of Canada. Workbench's UK supplier is Adept Scientific plc, tel 01462 480055. ***Electronic Workbench***'s price is £199.

Rod looks at ***CircuitMaker*** later in this article. Subsequent reviews will cover ***Tina Pro*** from Designsoft, Labcenter's ***Lisa***, which is part of *Proteus IV*, and ***Pulsar and Analyser*** from Number One Systems, which are modules from the *Easy PC* package.

stock the bins with the symbols you most often use, by saving it the same way as a template, the situation for experimenting on-screen is much improved. In fact, in some situations this could be the best type of program. You could save several pre-stocked examples with symbols to suit particular types of work – for example, power supplies, audio filters, etc. Labcenter's *Proteus* is a good example of this sort of program.

A third type has permanent parts-bins filled with the most common components, usually held in generic form. This type of program is also very suitable for on-screen designing. The content of these bins can usually be varied to suit the individual user, and the designer can quickly assemble an experimental circuit from these generic components, without having to bother with selection from a main library. If everything is satisfactory, he can then go back and edit each symbol to give it a real-world identity, such as *2N3055* or *LM741, etc. CircuitMaker, Workbench* and *Tina* fall into this category.

This system usually originates from companies with a simulator background, but it can be suited to pcb production. Most automated pcb layout programs do not mind if the net list has been generated from a simulator-type schematic-capture program or a pcb-CAD-type schematic program. The one *proviso* is that a library of connectors – a non-Spice item – should be present.

This assumes that the net list is compatible between programs, but in fact many simulators offer net-list export to a pcb program not just in one format but in several, either as an integral part of the program or as an inexpensive add-on.

Some simulator companies have taken this opportunity a stage further by adding complete third-party pcb-layout programs to their products. However, the user then has to learn one style of operation for the schematic drawing and simulator, and another completely different style of operation for the pcb layout.

Many users would say that learning one program style is bad enough – learning two is beyond the pale. It is indeed much better if the simulator and pcb-layout programs come from the same programming team.

Mixed-mode simulators

As many real-world circuits now have both analogue and digital elements, which seems to be a continuing trend, a simulator capable of mixed-mode operation is highly desirable.

While analogue and digital simulators are confined to their own specific circuits, simulator designers have employed various software tricks so that the two can run together on circuits containing both. The program can then justifiably be claimed to be able to run a mixed-mode simulation. It is usually obvious from using the demonstration disk that a software device joining the two is present. Such simulators are referred to as 'glued'.

While these glued simulators have their uses, life is much easier if there is no apparent joint between analogue and digital parts of the mixed circuit simulation. This especially so if you are starting out in simulation.

Analogue and digital simulators work in different ways. In theory, it may be possible to run an analogue simulation of a circuit that had digital models in it, but it would very, very slow – far too slow for practical use. But if the program were to insert a-to-d and d-to-a converters in front of and behind the digital models in a mixed circuit, the simulator would then see only analogue nodes and it would be possible to run the analogue simulator at a reasonable speed. Simulators that do this are described as native or true mixed-mode simulators.

There are other proprietary ways of achieving a similar result, depending on the program maker. How these work is usually explained in the user manuals.

Related books

Prices and availability of the books listed here are quoted by Waterstones.

The SPICE Book, A. Vladimirescu, ISBN 0471609269 £22.95, delivery 2/3 weeks.
Inside SPICE, R.Kielkowski, ISBN 0079137121, £49.99, in stock.
Computerised circuit analysis with Spice, T.W.Thorpe, ISBN 0471551643, £85.00, delivery 2-3 weeks.

What you can expect for your money

Unlike a pcb-CAD program, it is not easy to say what constitutes a good circuit simulator, because every designer has very different needs and expectations depending on the field that they are working in. The required end product is hardly a concrete object like a printed circuit board, but is ill defined and variable.

I would suggest that the analogue section of a general-purpose budget-price program should offer, as a minimum;

- transient simulation i.e. what you would normally see on a two-channel oscilloscope.
- signal source giving at least sine, square and triangular waveforms, plus a user-constructed or piecewise waveform.
- methods of showing several ac and dc voltages and currents, either by probe or by virtual instrument and preferably simultaneously.
- in the AC analysis section, graphs of the following parameters versus frequency .

  a) amplitude
  b) phase
  c) input and output impedance
  d) noise
  e) distortion
- dc analysis – i.e. a curve plotter.

In the digital section, you should expect to have;

- a logic analyser of at least 16 channels.
- control over the threshold time of glitches.
- step-by-step control over digital simulation .
- a digital signal source, word generator or data sequencer

If a program offers other analyses that are useful to your field of operations, such as temperature sweep, Fourier, worst case, Monte Carlo, these are, at this level, a bonus.

In reality, most budget-price simulation programs do not meet these basic criteria. They exceed basic analyses handsomely in some areas but neglect others entirely.

Regrettably, the designers of simulators see the market differently from the users. They often cram programs full of esoteric features, sometimes to the detriment of overall utility, probably for marketing reasons rather than technical requirement. You can often see this same philosophy at work on the remote control of your video recorder!

Analysis of input and output impedance versus frequency is just such an instance. Several programs lack this – so what are the users of such a program supposed to do when it comes to measuring these impedances? Go back to the bench? Devise their own work-around?

Continued on page 699

Special offer exclusively for *Electronics World* readers

Buy the Edison 3 circuit simulator and The Idea Factory system simulator together and save £38*.

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Supplied by Quickroute Systems Ltd, the new Edison 3 is a unique environment for exploring and analysing electronic circuits. It makes designing circuits fun, but don't be fooled by its ease of use. Edison 3 includes a powerful circuit analyser based on Tina.

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- Virtual instruments include: voltmeter, ammeter, ohmmeter, multimeter, signal generator, signal analyser, and oscilloscope
- High quality diagrams can be plotted after simulation is complete. Diagrams can be annotated, labeled and printed or copied into the clipboard.
- Symbolic analysis of linear circuits show how the results of a circuit are derived or mathematically described.
- Schematic analyser included with Edison lets you draw and analyse schematics directly. Select from a catalogue of real analog and digital integrated circuits as well as a range of passive and mixed-mode devices and perform DC, AC and transient analysis. Results are displayed in the diagram window.
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*Offer ends 31 December 1999. Overseas readers interested in this offer contact Quickroute for details. Quickroute Systems Ltd, Regent House, Heaton Lane, Stockport SK4 1BS. Tel. 0161 476 0202, fax 0161 476 0505.

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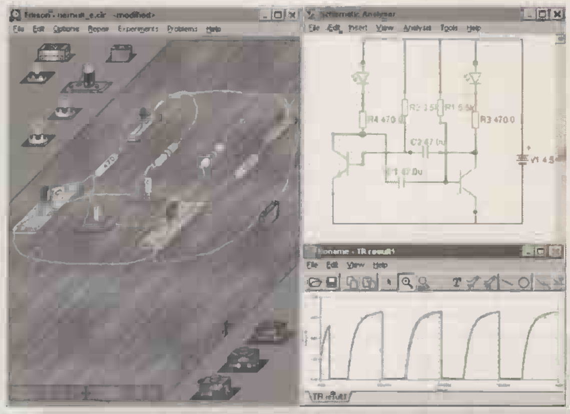

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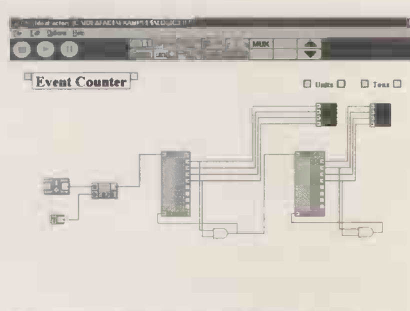
Features

- Very easy to use with an uncluttered set of controls
- Drag and drop objects from palettes onto the workspace and link using the fully automatic orthogonal drawing tool.
- Fully automatic junction placement, with links redrawn automatically when objects are moved.
- Electronic devices include; LEDs, 7 segment displays, counters, sequencers, switches and push-buttons
- Mathematical modeling objects include plus, minus, divide, sin, cos , exp, 1/x, power, summation and polynomials.
- Multimedia objects include bitmaps, animated bitmaps, moving/animated bitmaps, sounds
- Random number generators, number comparators and triggerable counters
- Run models in the same workspace in parallel
- Requires a 386DX or better processor, Windows 3.x, 95, or 98 and 4MB or more memory.



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The 48 hour week: who wins?

With its maximum working week of 48 hours, the EU's Working Time Directive is aiming to bring the UK manufacturing workplace in line with the rest of Europe. Richard Wilson finds out what it all means

If dealing with a sluggish economy was not enough, electronic equipment manufacturers may have to re-appraise their approach to production as the full costs of complying with the government's new Working Time Regulations come to light. The new legislation, which came into effect in October 1998, aims to improve the health and safety standards of employees in the workplace, bringing the UK in line with the rest of Europe.

The EU Working Time Directive restricts the working week to a maximum of 48 hours and enforces the provision of guaranteed rest periods, all of which must be documented for possible inspection by Health & Safety Executives.

Under the directive, employees of three months plus standing are entitled to a minimum of three weeks paid holiday per year. In addition to this, night-shift workers are entitled to free health checks, and all non-employed trainees must be regarded as workers for the purposes of the directive.

But how many manufacturers are prepared for the changes? Ron Biggs, group general manager at contract manufacturer Remploy believes the directive has far reaching implications for all businesses. "Yet the perception we have is that few manufacturers have taken full stock of its inevitable effects," warns Biggs.

Even those employers already working within the new regulations cannot be too complacent. They may have workforce agreements that vary the limits set by the directive, but they are still required to record hours and break periods. That will inevitably add to administrative costs.

"It will undoubtedly bring additional costs to all UK businesses," adds Biggs who is responsible for Remploy's 20 UK factories. "Particularly manufacturers with large, and temporary workforces. Coupled with the introduction of the minimum wage, a lot of manufacturers will be faced with an unexpected increase in labour and administrative costs." ■

WORKING TIME DIRECTIVE AT A GLANCE

- A maximum average 48 hour working week.
- A rest break after 6 consecutive hours of work, for a minimum of 20 minutes.
- A rest period of at least 11 hours in any 24 hour period, and of at least 24 consecutive hours in each seven day period.
- After a qualifying period of 3 months, workers are entitled to a minimum of 3 weeks paid annual leave each year, rising to 4 weeks from 23 November 1999.
- Night workers should not work for more than eight hours in 24 hours (usually averaged over 17 weeks, except for those whose jobs involve special hazards or heavy physical or mental strain).
- Night workers have the right to free health checks.
- Non-employed trainees are to be regarded as workers for the purposes of the directive.

Exemptions:

- Special exemptions apply to some workers in the transport, security, hospital, postal, media, utility and agricultural sectors.
- Employers can make collective or workforce agreements with union or worker representatives to vary the limits on night work, rest periods and other provisions.
- These workforce agreements can apply to specific groups of workers, rather than the whole workforce, for example, groups at a single site, or within a single function.

CIRCUIT IDEAS

Smart switch

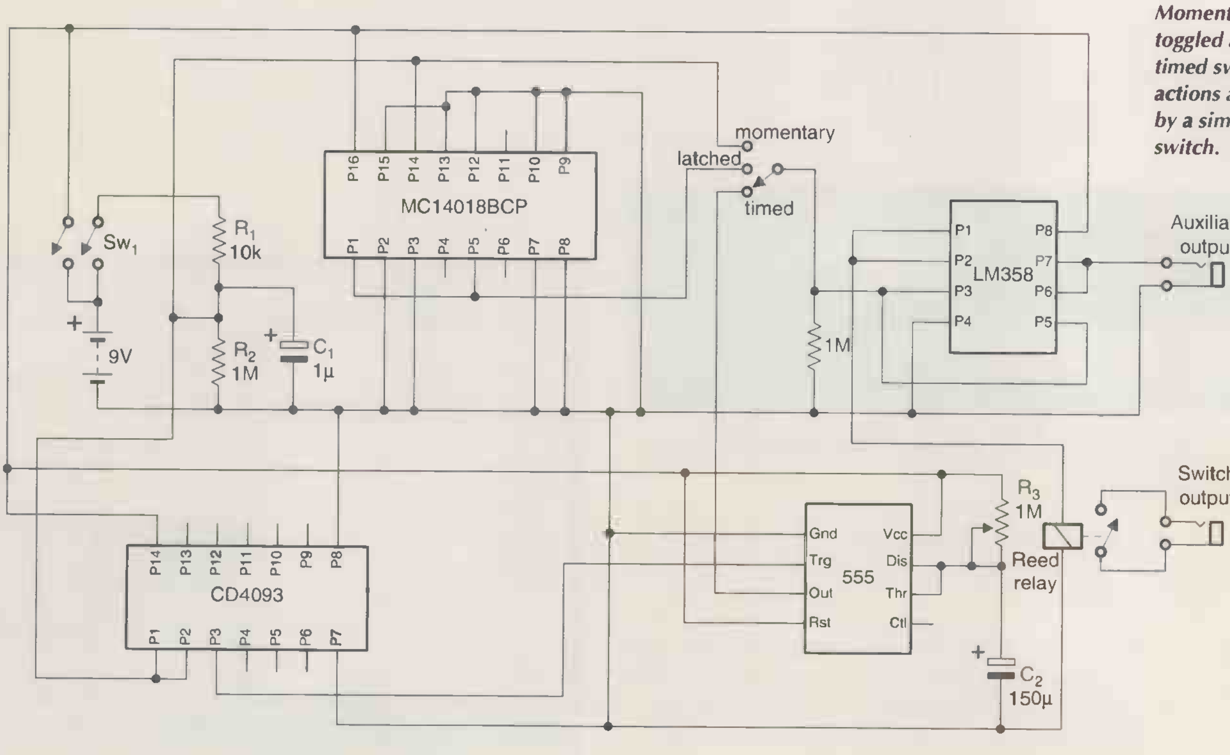
A simple switch – in this case a jack plug and socket – performs the selectable functions of momentary, toggle and variable on-time switches.

Sw_1 is the double-pole jack; inserting the jack plug connects the battery and $R_{1,2}C_1$ debounce the connection. The voltage is taken three ways.

Firstly, it goes straight to the "momentary" input of the rotary selector switch. Secondly, it is connected to the *4093* Nand to square the pulse and invert it to trigger the *555* monostable flip-flop, whose time constant $T=1.1R_3C_2$, the values shown giving a maximum on time of about three minutes.

Finally, the *MC14018BCP* in divide-by-two form, which has an internal Schmitt at the clock input, takes the pulse and produces the toggle action at pins 5 and 6. Half of the dual op-amp *LM358* takes the output of the selector switch and drives the reed relay to provide the output switching action. The second op-amp may be used to give an output for use in other circuitry.

Tomas Ward
Glenties
County Donegal,
Ireland
D2



Momentary, toggled and timed switching actions are given by a simple switch.

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

Like life, *Electronics World* may seem surreal at times, but it is certainly not exclusive. Clearly, 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.

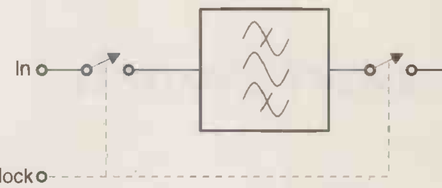
Clear hand-written notes on paper are a minimum requirement: disks with separate drawing and text files in a popular form are best – but please label the disk clearly.

Amplifier for small direct voltages

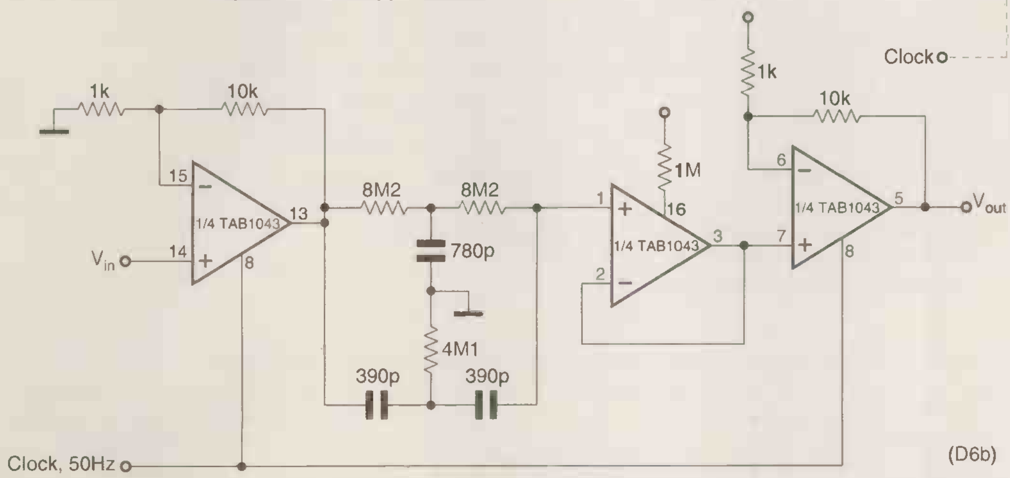
Four op-amps in the *TAB 1043* are programmable by current injected into the bias pins, pin 8 programming three of them and pin 16 one. Three of the op-amps are used here as choppers and a filter in an amplifier with a gain of 100.

Input is chopped, filtered and amplified and chopped again at the same chopping frequency and phase, the advantage of this arrangement being that the two choppers are controlled by the one input. The filter notch is 50Hz, the chopping frequency.

Kamil Kraus
Rokycany
Czech Republic



(D6a)



(D6b)

Three programmable op-amps in a chopper amplifier for direct voltages.

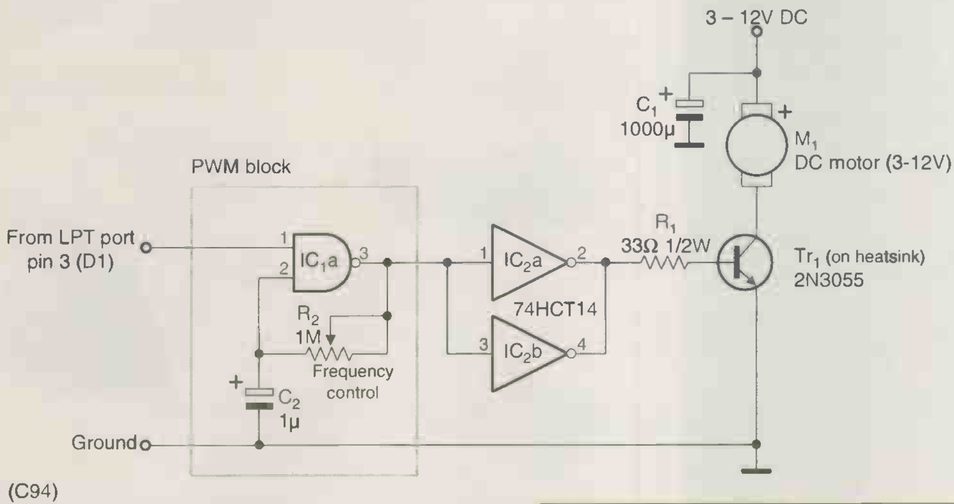
Motor speed control by pc

Mobile robot machines require small batteries and lightweight speed-control circuitry. This arrangement is a pulse-width modulator controlled by a Basic program in the pc.

In the diagram, the Schmitt inverter and associated components form a pwm oscillator, which is controlled by the Basic program, its lowest frequency being determined by the values of R_2C_2. Data bit D1 on pin 3 of the pc LPT1 port controls the pulse width, the motor turning off when D1 is low and on when D1 is high. The relative durations of the two states sets the motor speed; a continuous high level supplies a 50% duty cycle pulse. Motor reversal is not possible with this layout, but an H-bridge control would allow that facility.

The 3055 n-p-n power transistor is able to supply up to 10A peak to enable the robot to start and stop rapidly.

Shyam Sunder Tiwari
Robotics Software Pvt. Ltd
Gwalior
India
C94



Pulse-width control for a robot motor, controlled by the parallel port of a PC. High power allows sudden stops and starts.

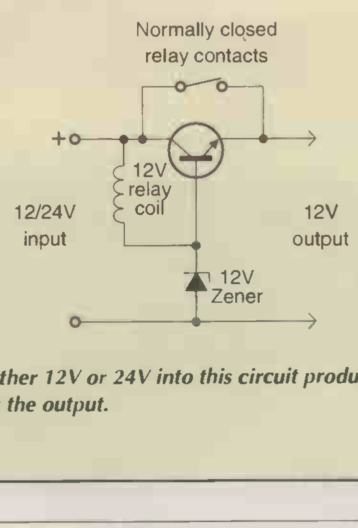
Basic listing for the motor speed controller.

```
REM PWM DC MOTOR SPEED CONTROL PROGRAM
REM Developed by Dr Shyam Sunder Tiwari
REM February 9, 1999
REM You may only use this program for non-commercial applications
CLS
REM Input from user the speed control parameter
start:      INPUT "motor speed power factor (scale 10 to 80) =", P
            IF P > 80 GOTO start
            IF P < 10 GOTO start
period:     INPUT "motor power on time (range 1-1000 seconds) =", S
            IF S > 1000 GOTO period
            IF S < 1 GOTO period
            H = 20 * P
            REM H is high level PWM control output
            L = 20 * (80 - P)
            REM L is low level PWM control output
            CLS
            LOCATE 10, 5
            PRINT "                                             "
            LOCATE 11, 1
            FOR i = 1 TO 8
            PRINT ":";
            FOR j = 1 TO 9
            PRINT ".";
            NEXT
            NEXT
            LOCATE 10, 1
            FOR i = 1 TO P
            PRINT "-";
            NEXT
            PRINT ">": k = 0
            ON TIMER(1) GOSUB time
            TIMER ON
repeat:     IF k > S GOTO done
            GOSUB control: REM Endless loop
            GOTO repeat
done:       STOP
time:       LOCATE 12, 25
            k=k+1
            PRINT "(power="; P; " time="; k; "seconds)"
            RETURN
control:    FOR i = 1 TO H
            OUT 888, 255: REM 888 is the address of LPT1 data port
            NEXT i
            FOR i = 1 TO L
            OUT 888, 0
            NEXT i
            RETURN
```

12V or 24V supplies accepted

If the input to this circuit is 12V, the relay is closed and output is 12V. If the input is 24V, the relay coil energises, the contact opens and the series transistor passes 12V again.

DM Bridgen
Camberley
Surrey
D11



Either 12V or 24V into this circuit produces 12V at the output.

Ppm and pcm demodulators

Following on from the pulse-width demodulator described in the March issue, p. 237, minor changes will produce demodulators for other types of pulse modulation,

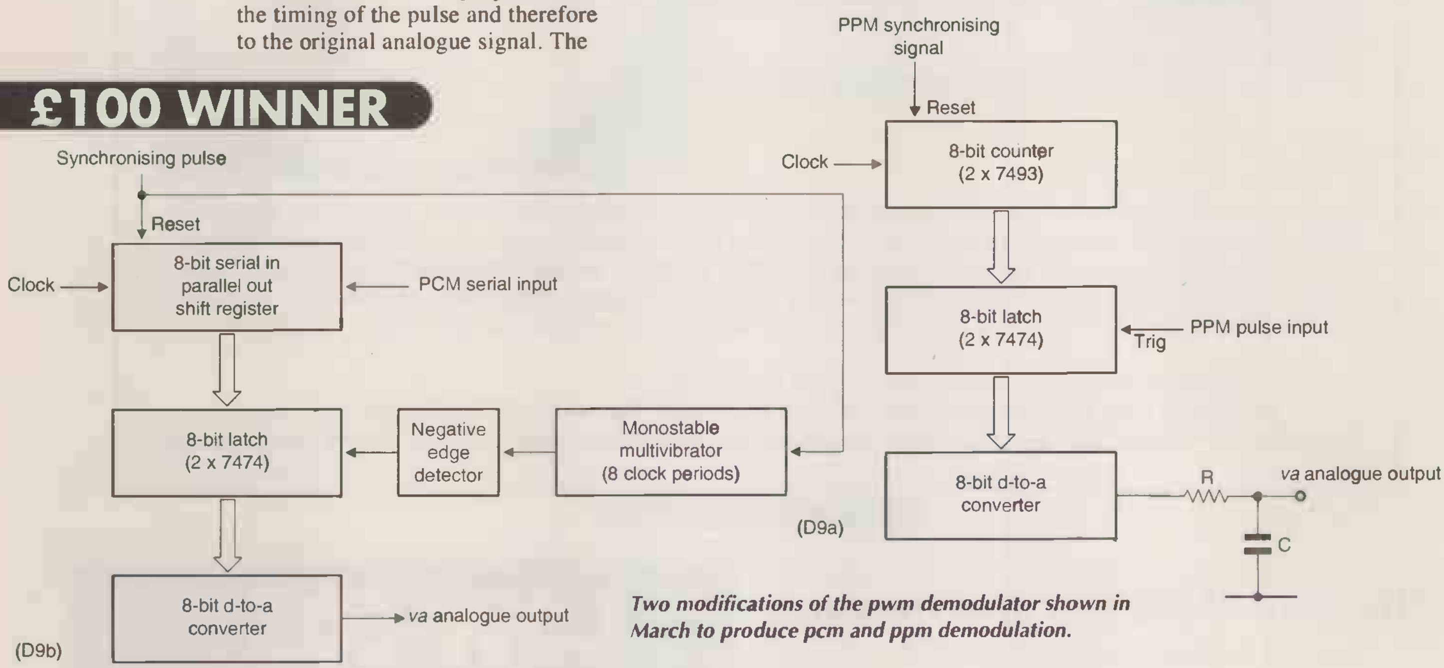
Pulse-position. Instead of the edge detector used in the pwm demodulator, the sync. pulse of ppm resets the counter to start a new count, so that the counter will contain data, at the next reset, proportional to the timing of the pulse and therefore to the original analogue signal. The pulse also latches the counter output and the d-to-a converter produces the analogue output to an *RC* filter.

Pulse code. For this, the modification shown in **Fig. 2** may be suggested.

Each sync. pulse clears the 8-bit serial-in, parallel-out shift register, serial bits from the pcm input then being clocked in. After shifting eight bits, this period being timed by a monostable triggered by the sync. input and a negative edge detector, the data latches into the d-to-a converter, which produces the analogue output. The *RC* filter will again be required.

K Balasubramanian
European University of Lefke
Lefke
Turkish Republic of
Northern Cyprus

£100 WINNER



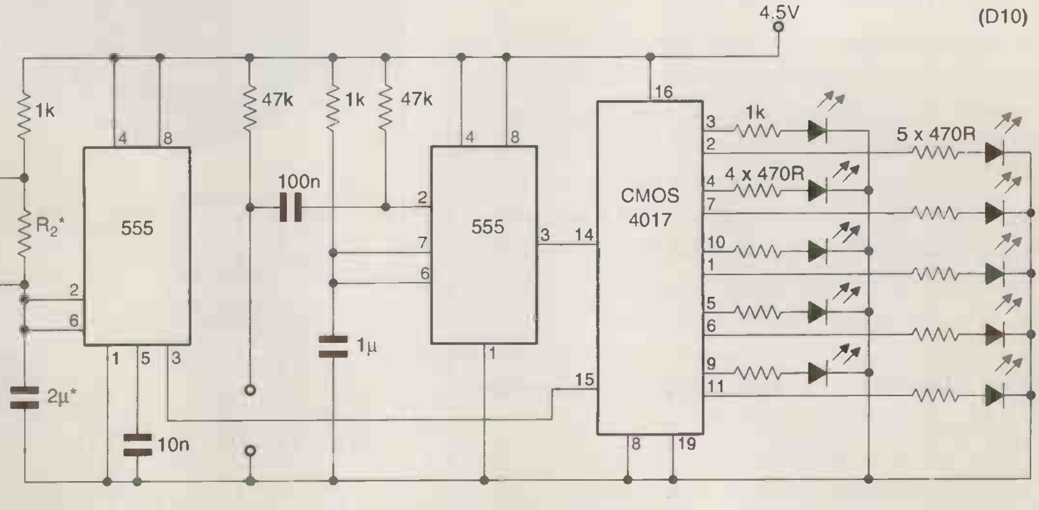
Two modifications of the pwm demodulator shown in March to produce pcm and ppm demodulation.

0-10 led display for digital input

Needing an easily read and robust display of speed and wind strength for use in a dinghy, I had to discount microammeters as being too fragile and a 10-led, continuously illuminated type too current-hungry.

In the circuit shown, the input from the digital sensor, which is a reed switch, is debounced by 0.1μF and 47kΩ and used to trigger a *555*. This 555, the one in the middle, supplies cleaned-up pulses to the counter, which counts up for a time set by the other 555 running at 1-5Hz. Output leds of the counter light up one after the other and are extinguished, so that only one is on at a time; led 1 stays on with no counter input as a confidence-inspirer.

Josef Holoåek
Prague
Czech Republic
D10



Sturdy and easy-to-read digital led display for use in occupations such as sailing a dinghy. The R and C marked with an asterisk set the 1-5Hz frequency

Digitally controlled 555 oscillator

Xicor's *X9315W* digitally controlled, solid-state potentiometer replaces R_a and R_b in the usual *555* oscillator configuration to allow computer control of both frequency and duty cycle. Discrete values of these resistors give frequency and duty cycle as,

$$f_o=\frac{1.44}{(R_A+2R_B)C} \quad \text{and} \quad DC=\frac{R_A+R_B}{R_A+2R_B}$$

If the two sections of the digital potentiometer from the wiper to the ends are kR and $(1\text{-}kR)$,

$$f_o=\frac{1.44}{(2-k)RC} \quad \text{and} \quad DC=\frac{1}{2-k}$$

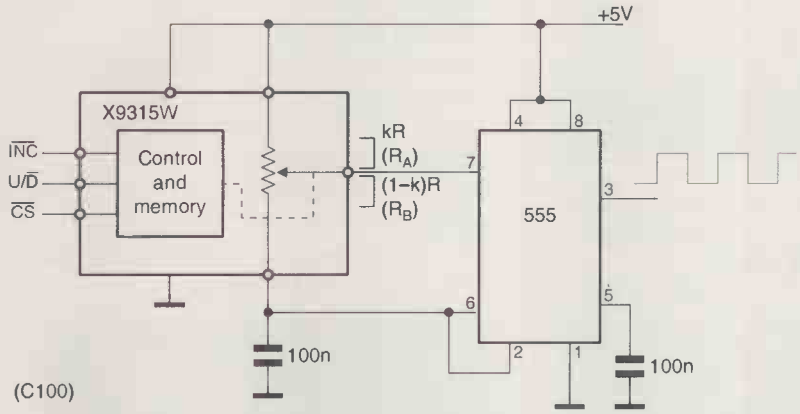
Where k varies between 0 and 1 to indicate the wiper position between one end and the other. Programming the pot. gives a frequency of $1.44/2RC<f<1.44/RC$ from one end to the other, duty cycle being $0.5<DC<1$.

In the circuit shown, frequency is 725Hz-1.35kHz, the potentiometer being a 10kΩ type having 32 taps and a three-wire interface.

Wiper setting may be stored in the digital potentiometer's non-volatile memory. Other types, with differing numbers of taps and resistance values, are available.

Chuck Wojslaw
Xicor Inc.
Milpitas
California
C100

Digital potentiometer allows computer control of 555 oscillator's frequency and duty cycle, storing the setting in memory.



(C100)

Hybrid bridge rectifier

Using both semiconductor and thermionic diodes in a bridge rectifier for valve audio equipment has the advantages of a slow start to reduce cathode stripping in the equipment and a softer characteristic to assist compression effects due to a rather poorly regulated supply.[1] A bridge of semiconductor devices would require the use of a standby switch to allow the equipment's valve heaters to warm up.

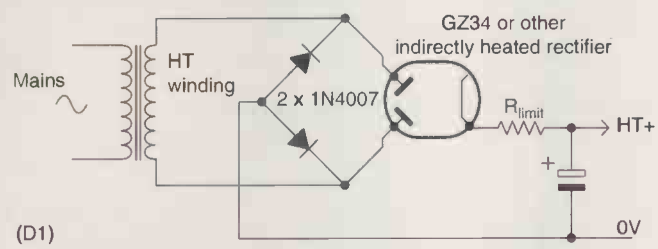
Instead of the centre-tapped transformer needed by a full-wave thermionic rectifier, the two semiconductor diodes shown allow the use of bridge layout and an untapped transformer, while retaining the benefits of valve rectifiers.

Macolm Watts
Wellington
New Zealand
D1

Reference

The Cool Sound of Tubes. *IEEE Spectrum*, August, 1998, p.28.

Using both valve rectifiers and semiconductors in a power supply for valve audio amplifiers confers several benefits.



(D1)

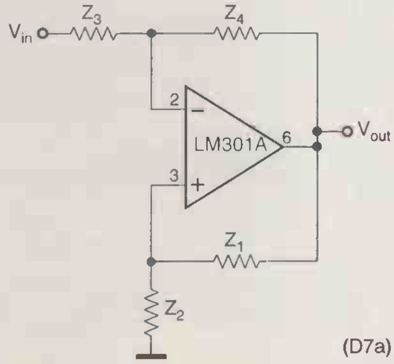
Two single op-amp circuits

By interchanging resistors with capacitors, the circuit of **Fig. 1** becomes a high-pass or a low-pass filter. Similar in arrangement to a Sallen and Key filter, this one uses four components instead of six, four being the smallest possible number.

In **Fig.** 2, assuming that $R_1=R_3$, and that $R_2=R_4$, the gain is $A=(1+R_1/R_2)^2$, the square of that obtained by driving the op-amp at its inverting input.

Kamil Kraus
Rokycany
Czech Republic
D7

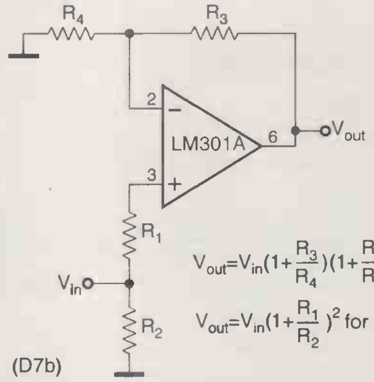
Two useful op-amp circuits. Figure 1 can be either a high or low pass filter and the circuit of Fig. 2 is driven at its non-inverting input to give the square of the gain when fed at its inverting pin.



Low pass:
$Z_1=R_1$ $Z_3=R_3$
$Z_2=\frac{1}{sC_2}$ $Z_4=\frac{1}{sC_4}$

High pass:
$Z_1=\frac{1}{sC_1}$ $Z_3=\frac{1}{sC_3}$
$Z_2=R_2$ $Z_4=R_4$

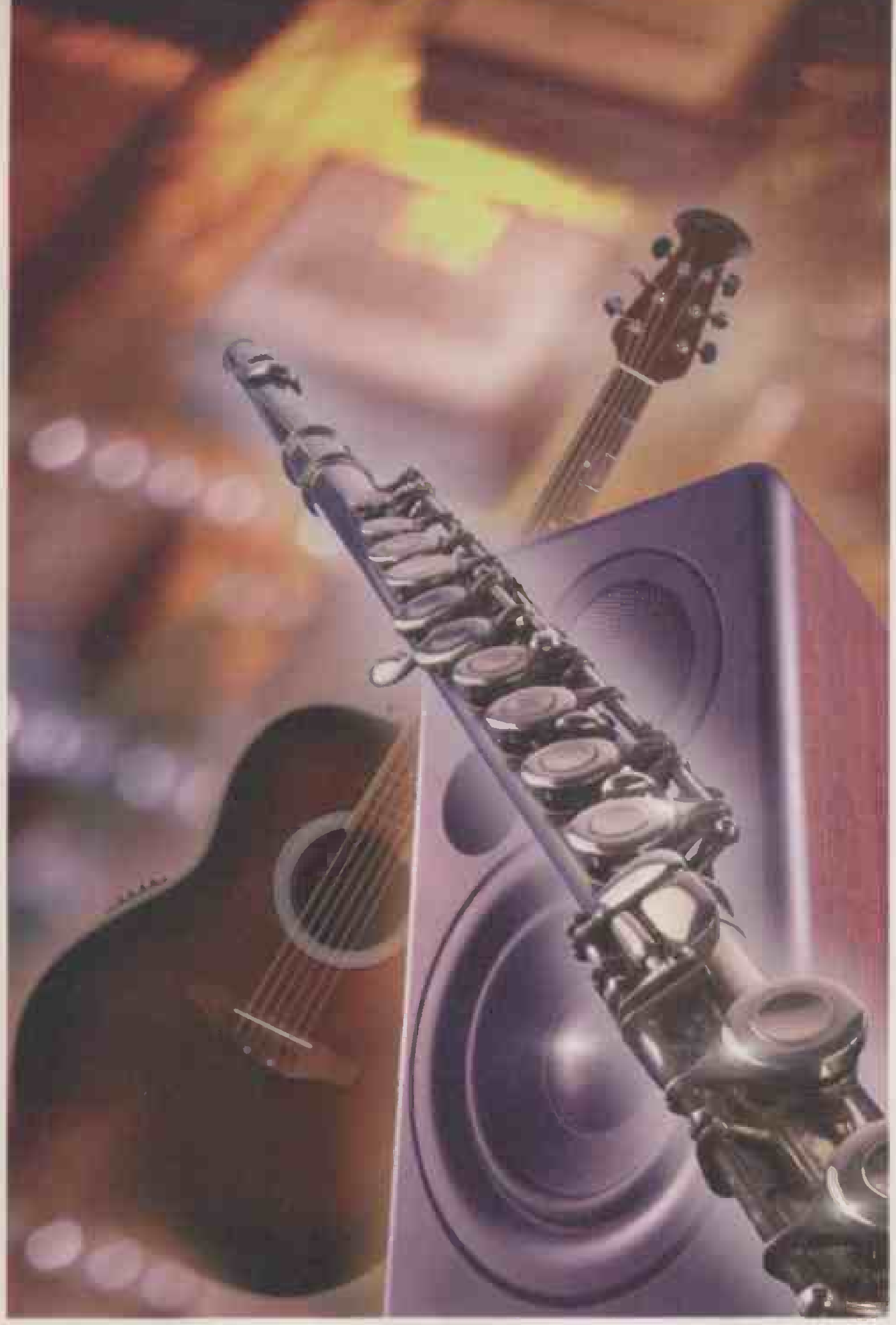
(D7a)



$V_{out}=V_{in}(1+\frac{R_3}{R_4})(1+\frac{R_1}{R_2})$

$V_{out}=V_{in}(1+\frac{R_1}{R_2})^2$ for $R_1=R_3$ and $R_2=R_4$

(D7b)



With the right loudspeaker combination, Bill Hardman's new active loudspeaker crossover design gives clean and precise reproduction – or for those of you of technical bent, it achieves narrow in-phase transition with low group-delay distortion, expanding on the work of Linkwitz and Riley. Bill presents a fully-worked design example for 1.5kHz.

Precise active x-over

The moving-coil loudspeaker lives on in spite of many attempts to find alternatives. It is rugged, reliable and can be made in as many styles as there are needs, to supply milliwatts of audio in a speaker-phone to air shaking kilowatts for a rock concert in a stadium. But there is a major limitation. It is not possible to produce high quality audio from a single drive unit over the whole audio range.

Aside from the problem of trying to get a single drive to have an adequately flat frequency response there are more significant areas that affect the general enjoyment of listening. A loudspeaker large enough to move significant volumes of air at low frequencies emits sound as an increasingly narrow beam as the wavelength becomes shorter.

To avoid this effect the sound handled by the loudspeaker must have a wavelength greater than the drive unit's diameter. For this reason almost every loudspeaker system consists of two or more drive units, each operating over a limited frequency range.

Dividing the audio into frequency bands appropriate for each loudspeaker is done by a combination of two or more band limiting filters, collectively called the crossover filter.

Why use an active crossover?

For 70 years the loudspeaker has generally been regarded as a passive device. The need to plug it into the mains is seen in some way as a negative feature. Moreover, hi-fi enthusiasts like the freedom of being able to choose amplifier and speakers from different manufacturers.

Except at the high end of the speaker market, there is resistance to the integration of loudspeaker, amplifier and crossover. This is a pity because it is in the low and mid price market that active speaker systems would have most impact.

When the crossover is part of the loudspeaker, coming after the amplifier, it must satisfy the following requirements:

- Pass the full power of the amplifier with minimal loss, and distortion.
- Present a non-reactive impedance to the amplifier.
- Present zero impedance to the loudspeaker drive unit.
- Split the signal into low and high frequencies with a steep transition.
- Avoid the introduction of phase or time distortion.
- Maintain sufficient accuracy to guarantee matching of phase between loudspeakers.

The first four affect the sound but the last two have a major impact on stereo imaging. To achieve all six in a top-end speaker is challenging, but for the mid-price market it is impossible. This is the very market that has to manage with amplifiers that cannot drive large powers into reactive loads and use drive units with a less than ideal response needing rapid attenuation of out of band signals that cannot be supplied by low-cost passive filters.

Loudspeaker manufacturers should beware that the computer market has no prejudice against active speakers.

A quick tour of the pole-zero map

Every analogue design engineer should be familiar with the pole-zero map – more formally called the S-plane – because it makes it easy to visualise the mathematics that govern the behaviour of filters. Most of us would rather deal with pictures than strings of numbers and equations. Once you understand this simple representation, it becomes very easy to customise standard filters by moving or adding poles or zeros.

All filters can generally be defined as one polynomial – i.e. a numerator – divided by another – a denominator. For filters, the denominator and the numerator involve frequency. There will be certain frequencies at which each will have zero value, these frequencies are the roots.

The roots of the numerator are the zero locations, the roots of the denominator are the pole locations. Where there is a pole, the response will become infinite; where there is a zero the response will become zero.

The roots of the numerator or the denominator may be complex, that is, they involve the square root of a negative number. We have no way with our number system of dealing with these numbers as a single quantity. But we can manage it as two numbers that always move through the mathematics, as if they were one.

These are termed complex numbers and they have a real part and an imaginary part. The real part is an ordinary number, the imaginary part also appears as an ordinary number but it represents multiples of the square root of –1, and by convention is usually prefixed with 'j' or an 'i'.

The pole-zero map is used as an aid to visualising these numbers. The real part of the complex number is represented horizontally, and the imaginary part vertically. It follows that the horizontal and vertical axes of the pole-zero map are called respectively the real and imaginary axes.

The pole-zero map contains an enormous amount of information. From it can be obtained the amplitude, the phase, the impulse and step response, using no more than the rules of geometry.

The poles are always to the left of the imaginary axis. The closer a pole is to the imaginary axis the higher its Q. The Q becomes infinite when it is on the imaginary axis. Moving a pole vertically will increase its frequency. Zeros can appear any where in the S-plane.

A full description of the use of the pole-zero map is more than be covered here, but any good book on filter theory will cover the basic principles. A simple example will show the relation between the pole and zeros and resulting filter response.

Figure A shows a pole-zero map with a complex pole at –1000, j1000. It has a companion at –1000, –j1000, which is not shown. It represents a second-order filter providing a low-pass response shown on the right.

Figure B shows the effect of moving the pole towards the imaginary axis; the response peaks at a frequency governed by the vertical location of the pole.

Figure C shows what happens when a zero is placed upon the imaginary axis. A notch appears in the amplitude response at the frequency corresponding to the zero location on the imaginary axis. Custom filter design comes no easier than this!

Fig. A. Pole-zero map with a complex pole at –1000, +j1000. It has a companion pole at –1000, –j1000, not shown. This represents a second-order filter providing a low-pass response shown on the right.

Fig. B. Effect of moving the pole towards the imaginary axis. Response peaks at a frequency governed by the vertical location of the pole.

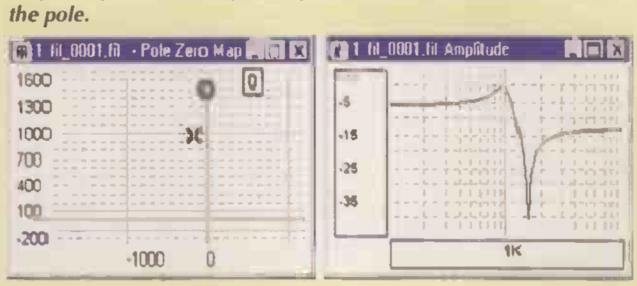


Fig. C. When a zero is placed upon the imaginary axis, a notch appears in the amplitude response at the frequency corresponding to the zero location on the imaginary axis.

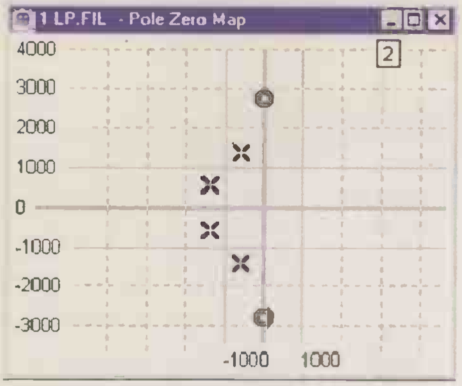


Fig. 1. Pole-zero map for the low-pass filter with two zeros at infinity and a complex zero at 2775Hz.

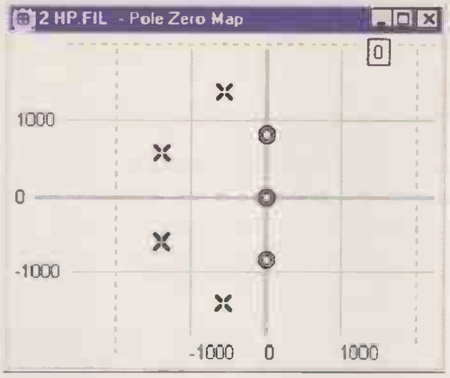


Fig. 2. Pole locations of the high-pass filter are identical to those of the low-pass filter, Fig. 1. The difference is that here, there are two zeros at the origin, i.e. 0Hz, and a complex zero at 807Hz.

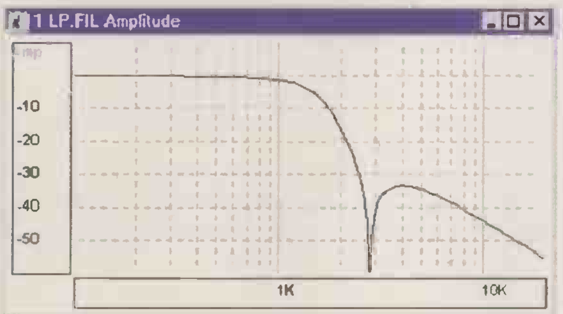


Fig. 3. Amplitude characteristics of the low-pass filter showing the notch at 2775Hz.

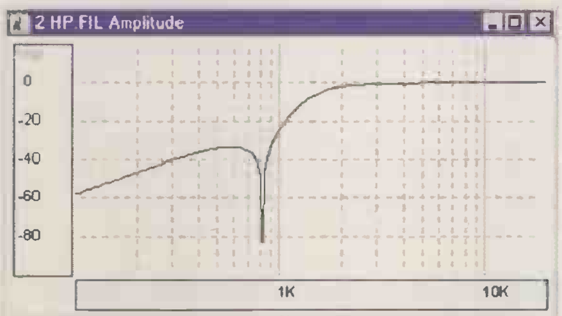


Fig. 4. Amplitude characteristics of the high-pass section.

How long before the technology being used here will start to move out of PC multi-media into the more general audio market – especially if they can bring a new young customer with them free of the more traditional ideas of how an audio system should be constructed?

Active crossover filters in practice

Removing the crossover filtering from between the power amplifiers and loudspeaker drive unit has the following benefits:

- The filters can utilise op-amps and low cost precision components to produce relatively complex designs, or be implemented entirely in the digital domain where even more complexity is possible.

- Although two power amplifiers are required per channel both can have a lower slew rate than the equivalent full range unit because the LF amplifier operates to a lower maximum frequency and the HF amplifier operates at a lower power.

- Driving the loudspeaker units directly from the power amplifier avoids the losses of a passive crossover filter thereby increasing efficiency and allowing the amplifier to have greater control over the motion of the drive coil.

Deriving the design

The aim of this design is to produce an in-phase crossover with –6dB amplitude for each filter at the crossover, a low group delay and narrow transition. There's more on the relevance of 'phase' in the panel entitled 'Crossover phase response' on page 693.

Normally crossover filters are composed of all-pole filters. These are filters that have complete attenuation at zero or infinite frequency and nowhere else. In other words, they have no notches in the frequency response.

Bessel, Butterworth and Chebyshev all belong to this class of filter, but Cauer elliptic and inverse Chebyshev do not. This is because they have notches in their stop band.

All-pole filter designs achieve increasing rates of cut-off by utilising more poles. Each pole contributes another 6dB per octave, but also increases the group delay and the potential for group delay distortion.

For filters with sharp transitions the group delay becomes very non linear around the cut-off. All this is bad news for loudspeaker crossover design, because the cut-off for two speaker systems usually occurs in the range 1-3kHz – that point in human hearing which is particularly sensitive to the effects of phase inconsistency.

Small errors in phase between the high and low-frequency drive units in this region can destroy stereo images and are probably the cause of many loudspeakers being regarded as fatiguing, because the listener's auditory system is continually trying to make sense of an inconsistent sound field.

Another reason for avoiding high-order systems is that the application of group delay equalisation, to make the system linear phase, becomes increasingly complex. The stored energy in the high-Q all-pass filter sections required, leads to considerable pre-echo and ringing.

The design I propose here starts with low-order all-pole low-pass and high-pass filters and adds a notch to the stop band of each that steepens the rate of cut-off in the transition while maintaining a good composite amplitude response throughout.

This technique is used in Cauer or

Phase and auditory perception

The importance of phase response and our perception of it is a controversial subject – but it cannot be ignored. In fact the controversy started over 100 years ago with Rayleigh and his observations on the conclusions of Helmholtz on the inaudibility of phase in his experiments.

Our abilities to perceive and analyse sounds are the result of our biological development in an environment where time delays and reflections are the norm, but non-linear phase shifts are not. The exception would be sound that has been distorted by atmospheric effects, but this would be over considerable distances. The acuteness of our hearing has developed for close range, for example, communication, the identification of sources of danger – or food.

To get a good background to the subject the papers by Preis[4] and Lipshitz, Pocock, and Vanderkooy[7] are well worth reading.

Although non-linear phase response has audible effect on asymmetric wave forms – common with acoustic musical instruments – a more significant effect is likely to be upon that of the transient response. Woszczyk[8] sites this as having considerable effect on our ability to localise and identify sounds, or to put it in terms of our interests here, our perception of the accuracy of stereo image and how faithful the copy is of the original sound source.

elliptic filters. These types of filters have a notoriously non-linear group delay, but it is not due to the presence of the notches, but solely to the placement of the poles.

The notches are produced by complex zeros on the imaginary axis in the S-plane. These zeros contribute fixed phase shifts of plus or minus 90°. Because these phase shifts do not vary with frequency they do not contribute to group delay. It is only the poles that determine the group delay.

It is easy to demonstrate that this is so, using an elliptic filter as an example. If the filter's zeros are removed, an all-pole filter remains. Although the amplitude response is altered, the group delay is not.

Like Linkwitz,[9] the starting point is a Butterworth design, with the advantage that low-pass and high-pass designs of the same order and –3 dB frequency have the same group delay. This is because the poles of the two filters share the same locations.

If the filters are even order, the phase differences between them will be multiples of 180°, which can be corrected by a phase inversion in the low or high-frequency path, where necessary.

By adding a complex zero to the stop band of each Butterworth filter we can narrow the transition, and increase the crossover attenuation from –3 to –6 dB as required for good composite response, without increasing the group delay. The notch in the filter supplying the tweeter can ensure very high attenuation at its resonant frequency allowing a lower crossover frequency and making more use of the tweeter low-frequency response rather than the high-frequency response of the bass drive unit.

A practical design application

Here is an example of the crossover design procedure, completed in a few minutes using *HSPS Filter Designer.** The target for application was a pair of small loudspeakers bought as a kit.

Since most people will be apprehensive of buying conventional speakers and then pulling them apart to by-pass the crossover, a kit seemed a better starting point. This route is also likely to give you better value for money. Also if disaster strikes and a replacement driver is required, doing so for a kit is likely to be easier.

One point to take note of is that it is fairly easy to blow up tweeters. I play safe and fit 60μF of non-polarised capacitors between the tweeter and power amplifier.

The kits I used for the prototypes were for a bookshelf size loudspeaker, using a Morel MW142 bass unit and MDT29 dome tweeter, bi-wiring providing the four terminals required. The tweeter has a resonant frequency of 900Hz. Some experimentation and listening tests indicated that a 1500Hz crossover gave good results.

The starting point for each of the filters is a fourth order 1.5kHz Butterworth. Zeros are added on the imaginary axis and positioned to give a 6dB attenuation at the 1500Hz crossover point for both the low-pass and high-pass filters

Figure 1 shows the pole-zero map for the low-pass filter, while **Fig. 2** is the pole-zero map for the high-pass filter. In both cases the pole locations are the same, ensuring that they both exhibit the same group delay.

All the zeros lie on the imaginary axis. The high-pass filter has two zeros at the origin – 0Hz – and a complex zero at 807Hz. The low-pass filter has two zeros at infinity and a complex zero at 2775Hz.

Figures 3 and **4** show the amplitude characteristics of the filters and **Fig. 5** the all important composite response. The two filter outputs sum with less than 1dB error.

Figure 6 shows the phase difference between the filters. Throughout the crossover region there is no phase error between the filter outputs. At the frequencies of the zeros – the response nulls – the difference becomes 180°, but this is not a serious problem because it is occurs in the filter stop band where there is more than 30dB of attenuation

Continued on page 691...

Importance of good polar response throughout the crossover transition

At the crossover, or transition point, both drivers are pushing and pulling the air. If they do this exactly together an air pressure wave will emanate from the front of the speaker giving the impression to a listener face on, of sound from a point source mid way between the two drivers.

This co-operative effort of the two drivers will be upset by phase errors between them. Small phase errors will cause the pressure waves to be bent up or down – assuming the low and high-frequency speakers are aligned vertically one above the other.

The nasty thing about this effect is that it takes place only in the narrow band of frequencies encompassed by the transition. Although the amplitude response will be degraded a far more serious result is the loss of stereo image. This is not difficult to explain.

For any single point source of sound a room will have a consistent pattern of reverberation, and as I stated earlier, evolution has given us considerable abilities for making sense of this. But a sound source that has a sudden frequency-dependent shift in pattern of reverberation – often in the middle of the most acute part of our hearing – is outside those abilities. Phase errors in the transition will come from the drive units, the crossover filters and/or acoustic misalignment.

Phase errors from the drive units are outside the range of this article, but errors from the crossover and acoustic misalignments can be dealt with. There's more on this in the panel entitled, 'Time delay equalisation for non coincident drivers.

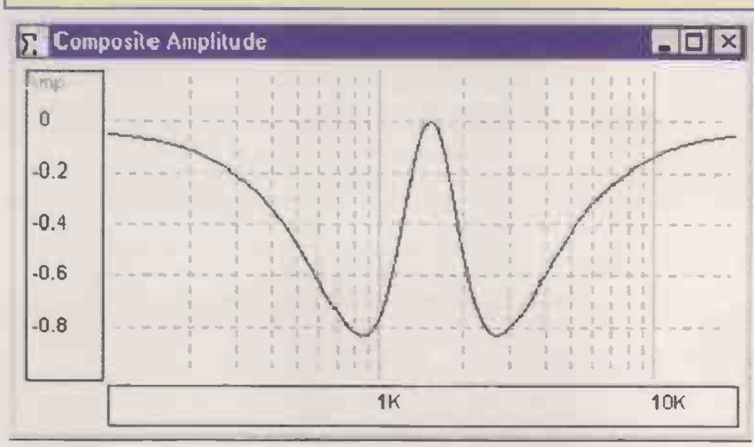


Fig. 5. Amplitude response of the high and low-pass filters combined shows that the two outputs sum with less than 1dB error.

*www.dialspace.dial.pipex.com/hsps/

Cool audio power

There was a time when dissipation in an audio power amplifier was simply a cost or heatsink-size issue. Now, with audio amps proliferating in portable and power-efficient equipment, dissipation is becoming a prime consideration. Doug Self explains exactly where the power goes.

There are several important power relationships in designing an output stage. Both the average and peak power dissipated in the output devices must be considered when determining their type and number. The average power dissipated controls the heat-sink design.

In most amplifier types the power dissipation varies strongly with output signal amplitude as it goes from zero to maximum, so the information is best presented as a graph of dissipation against the fraction of the available rail-to-rail output swing – i.e., the output voltage fraction.

Consideration of average power allows the output devices to be made thermally safe; but it is also essential to consider the peak instantaneous dissipation in them. Audio waveforms have large low-frequency components, too slow for peak currents and powers to be allowed to exceed the DC limits on the data sheet.

For a resistive load the peak power is fixed and easily calculable. With a reactive load the peak power excursions are less easy to determine but highly important because they are increased by the changed voltage/phase relationships in the output device. Thus for a given load impedance modulus the peak power would need to be plotted against load phase angle as well as output fraction to give a complete picture.

Average power drawn from the rails is also a vital prerequisite for the power-supply design; since the rail voltage is substantially constant this can be easily converted into a current demand, which must be known when sizing reservoir capacitors, choosing rectifiers, and so on.

The voltage rating of these components is a much simpler business, requiring simply that they withstand the off-load voltages at the maximum mains voltage, which is usually taken as 10% above nominal. The only thing to decide is how big a safety margin is required.

Power drawn depends on signal-level and is again conveniently displayed with voltage-fraction as the X-axis.

The mathematical approach

When dealing with power amplifier efficiency, most textbooks use a purely mathematical method as shown in Fig. 1, which was produced with the aid of *Mathcad*. The calculation gives only the dissipation in the power devices.

Figure 1 gives the familiar information that maximum device dissipation occurs at 64% of maximum voltage, equivalent to 42% of maximum power. These specific numbers are a result of the sine waveform chosen and other waveforms give different values.

To make it mathematically tractable, the situation is highly idealised, assuming an exact 50% conduction period, no losses in emitter resistors or $V_{ce(sat)}$s, and so on. Solving the problem for Class AB, where the conduction period varies with signal amplitude, is considerably more complex due to the varying integration limits.

Simulating dissipation

Alternatively, the power variations in real output stages can be simulated and the results plotted; the circuits simulated in this article are shown in **Fig. 2**.

For concision, and by analogy with logic outputs, I have called the upper transistor the source and the lower the sink. In simulation, losses and circuit imperfections are included, and the power dissipations in every part of the circuit, including power drawn from the supply rails, are made available by a single run.

It is an obvious choice – which I duly took – to use a sine waveform in the simulations. This allows a reality-check against the mathematical results. Reactive loads are easily handled, so long as it is appreciated that the simulation often has to be run for ten or more cycles to allow the conditions in the load to reach a steady state.

All simulations were run with ±50V rails and an 8Ω resistive or reactive load. The output emitter resistors were 0.1Ω. The drawback to this approach is that it is rather labour intensive. With my current simulation software, PSpice 6.0 for DOS, the steps are:

- Simulate the output stage over a whole cycle, for each input voltage fraction; 5% steps give enough points for a presentable curve; the •STEP command automates this.
- Display simulation results in the graphical post-processor. (In PSpice this is called PROBE) This assumes it can display computed quantities, e.g. $V_{ce} \times I_c$ to give instantaneous device power. Peak and average results can be read from the same display as PROBE. There is a function called AVG, which – unsurprisingly — yields the running average over a cycle. This stage can be automated as a macro, which is just as well, since it has to be performed at least 20 times, once for each input fraction value.
- The awkward bit. The computed

Class B amplifier power dissipation for sine waves

F_n is the fraction of full output swing.

$n = 0...10 \quad F_n = \frac{n}{10} \quad \omega = 1 \quad t = 0$

Rail volts (± rails) V=50

Instantaneous power dissipation $$P_n = \frac{V^2}{R}\left[F_n\left[\sin(\omega t) - F_n(\sin(\omega t))^2\right]\right]$$

Integrate over one half cycle $$Pdiss_n = \frac{V^2}{R} F_n \frac{1}{\pi}\int_0^{\pi} \sin(\omega t) - F_n(\sin(\omega t))^2\, dt$$

Note the $1/\pi$ due to integration from 0 to π.

Since only one device conducts at once, dissipation for one is total dissipation

Load resistance R=8

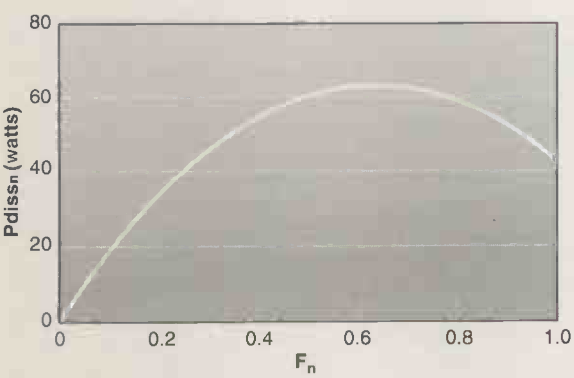
$$Pout_n = \left[\frac{V}{\sqrt{2}}(F)_n\right]^2 \times \frac{1}{R}$$ Output power

$$Poutmax = \left[\frac{V}{\sqrt{2}}\right]^2 \times \frac{1}{R}$$ Max. output power

$Poutmax$=156.25W

$$PF_n = \frac{Pout_n}{Poutmax}$$ Output factor

$$DF_n = \frac{Pdiss_n}{Poutmax}$$ Dissipation factor



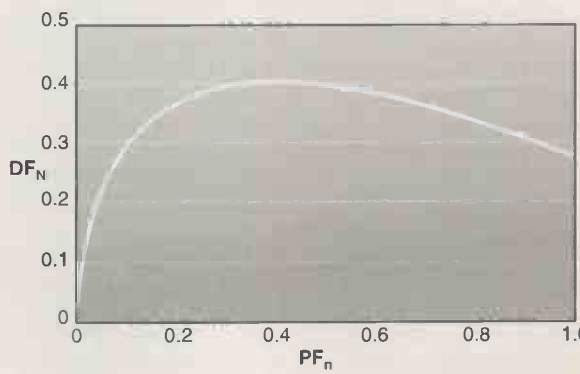


Fig. 1. The standard mathematical derivation for Class-B. Maximum dissipation occurs at 64% voltage output, equivalent to 42% of maximum power output.

peak and averaged power dissipations at the end of the cycle are read out from the PROBE cursor and recorded by hand, for each value of input fraction. There seems to be no other way to extract the information.

- The data from the third step is typed into Mathcad, to produce the graphs shown in this article. Once the data has been entered, Mathcad can manipulate it in almost any way conceivable.

Power-partition diagrams

The graph in Fig. 1 gives only one quantity, the amplifier dissipation.

I suggest a more informative graph format that I call a power partition diagram, which shows how the input power divides between amplifier dissipation, useful power in the load, and losses in drivers, etc.

Power dissipations are plotted against the input voltage fraction; this is not quite the same as the output voltage fraction as these are real output stages with gain slightly less than one. The input fraction increases in steps of 0.05, stopping at 0.95 to avoid clipping. The x-axis may linear or logarithmic.

Figure 3 shows the power-partition diagram for a Class-B complementary feedback pair stage as in Fig. 2, which has a low quiescent current. Line 1 plots the P_{diss} in the sink (lower) device. Line 2 is source plus sink power. Line 3 is source plus sink plus load power.

The topmost line 4 is the total power drawn from the power supply, and so the narrow region between 3 and 4 is the power dissipated in the rest of the circuit – mainly the drivers and the output emitter resistors R_e. This power increases with output drive, but remains negligible compared with the other quantities examined.

The diagram shows immediately that the power drawn from the supply increases proportionally to the drive voltage fraction. This is partitioned between the load – represented by the curved region between lines 2 and 3 – and the output devices. Note how the peak in their power dissipation accommodates the curve of the load power as it increases with the square of the voltage fraction.

Figure 4 shows the same diagram for a Class-B emitter-follower output stage. The quiescent current of an emitter follower output stage is significant – here 150mA – and pushes up the

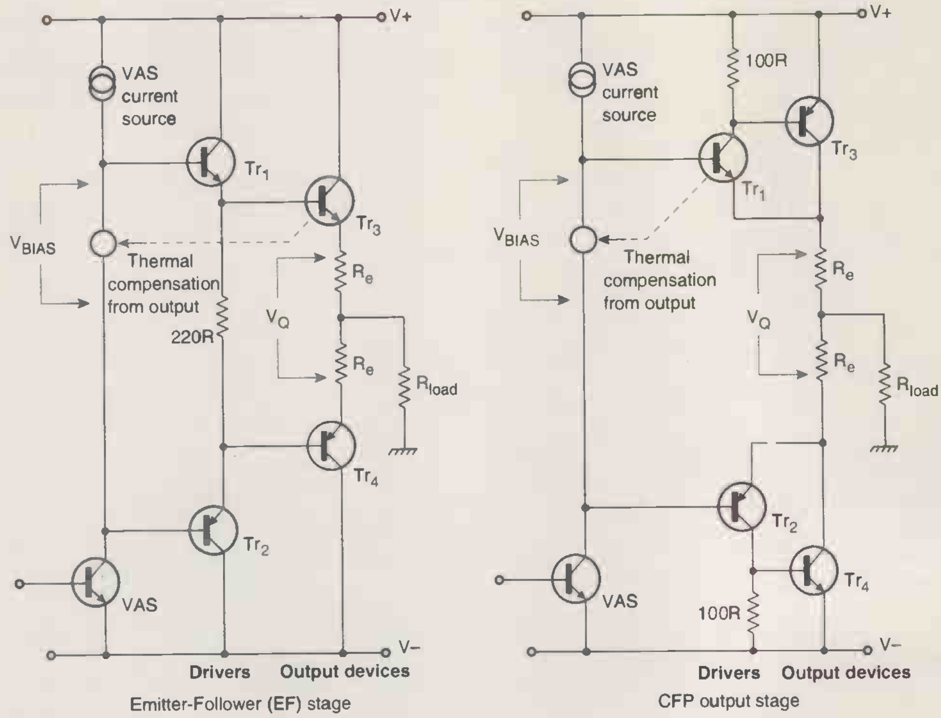


Fig. 2. The standard emitter follower and complementary feedback pair output stages. In optimal Class-B the emitter follower version takes about 150mA of quiescent current while the complementary feedback pair draws only 10mA.

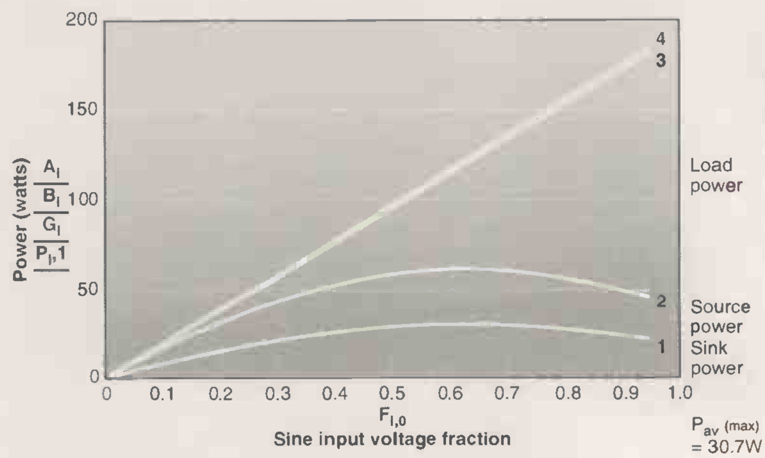


Fig. 3. Power partition diagram for a Class-B complementary feedback pair driving an 8Ω resistive load with a sinewave averaged over a cycle.

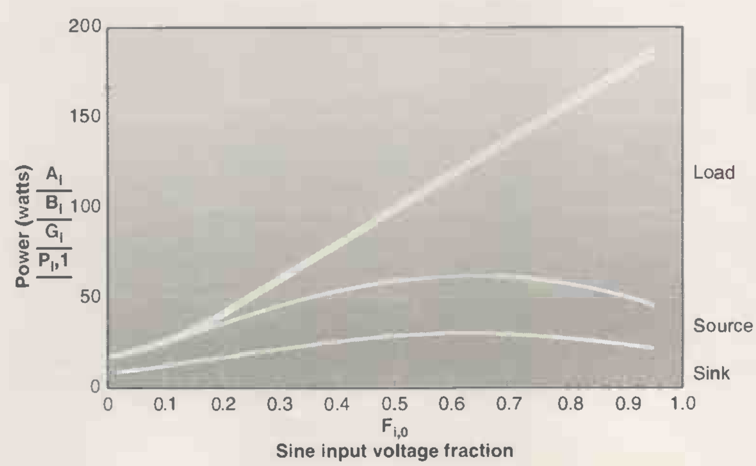


Fig. 4. Class-B emitter follower power partition into 8Ω resistive load. Sinewave drive. Significant quiescent dissipation at zero output.

Table 1.		
Angle (°)	$P_{diss(max)}$ (W)	Voltage fraction
0	60	0.64
10	63	0.65
20	67	0.70
30	70	0.75
45	95	0.95
60	115	1.00
90	185	1.00

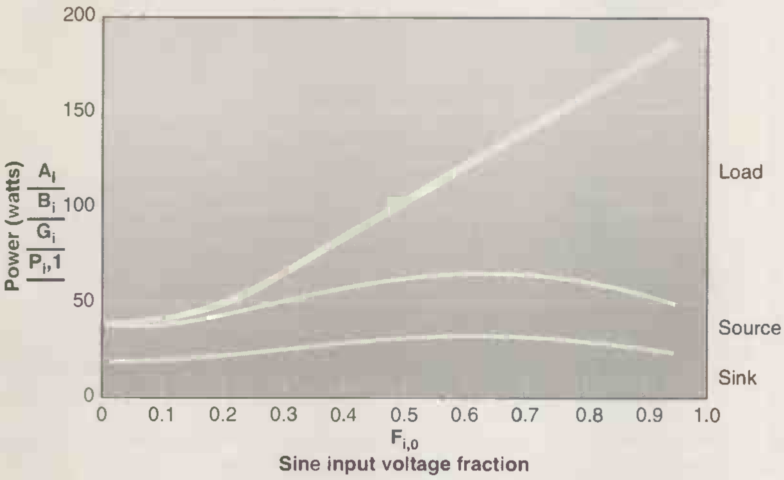


Fig. 5. Class-AB sinewave drive. If the range of Class-A operation is extended, the area below level=0.1 advances upwards and to the right.

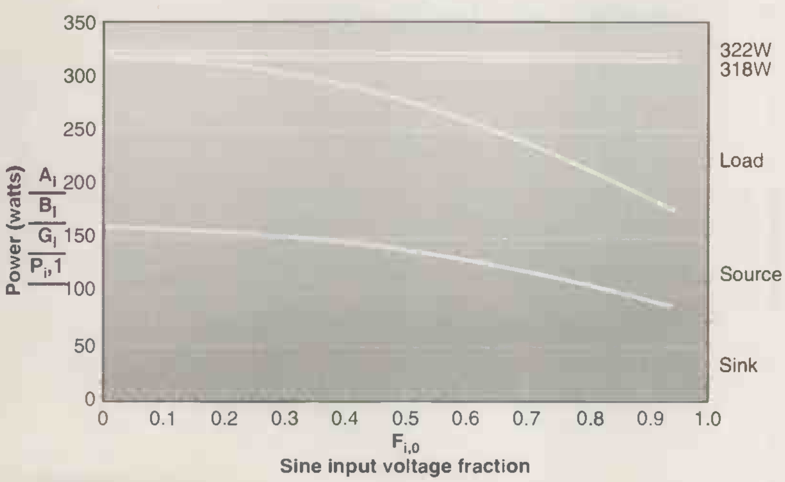


Fig. 6. Class-A push-pull sinewave drive. Almost all the power drawn is dissipated in the amplifier, except at the largest outputs.

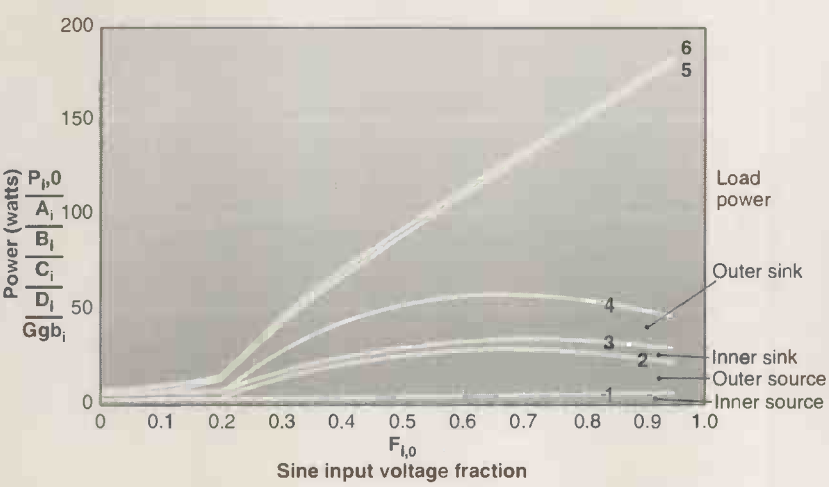


Fig. 7. Class-G with lower supply rails set at 30% of upper rails, with sinewave drive. Compare Fig. 4; amplifier dissipation at low levels is much reduced.

power dissipation around zero output, but at higher levels the curves are the same. There is no need for extra heatsinking over the complementary feedback pair case.

Effects of increased bias

Figure 5 shows Class-AB, with bias increased so that Class-A operation and linearity is maintained up to 5W RMS output.

The quiescent current has increased to 370mA, so quiescent power dissipation is significantly higher for output fractions below 0.1 Device dissipation is still greatest at a drive fraction of around 0.6, so once again no extra cooling is required to deal with the increased quiescent dissipation.

A push-pull Class-A amplifier draws a large standing current, and the picture looks totally different; see **Fig. 6**. The power drawn from the supply is constant, but as output increases dissipation transfers from the output devices to the load, so minimum amplifier heating is at maximum output.

The significant point is that amplifier dissipation is only meaningfully reduced at a voltage fraction of 0.5 or more, i.e. only 6dB from clipping. Compared with Class-B, an enormous amount of energy is wasted internally.

Single-ended or constant-current versions of Class-A have even lower efficiency, worse linearity, and no corresponding advantages.

Class G

Hitachi introduced the Class-G concept in 1976 with the aim of reducing amplifier power dissipation by exploiting the high peak-mean ratio of music.[1] I have recently explained its operation in ref. 2.

At low outputs, power is drawn from a pair of low-voltage rails; for the relatively infrequent excursions into high power, higher rails are drawn from. Here the lower rails are ±15V, 30% of the higher ±50V rails, so I call this Class G (30%).

This gives a discontinuous power-partition diagram, as in **Fig. 7**. Line 1 is the dissipation in the low-voltage inner source device, which is kept low by the small voltages across it. Line 2 adds the dissipation in the high-voltage outer source; this is zero below the rail-switching threshold.

Above this are added the identical – due to symmetry – dissipations in the inner and outer sink devices, as Lines 3 and 4. Line 5 adds the power in the load, and 6 is the total power drawn, as before.

Power consumption and amplifier dissipation at low outputs are much reduced; above the threshold these quantities are only slightly less than for Class-B. Class-G does not show its power-saving abilities well under sinewave drive.

Class-B and reactive loads

The simulation method outlined above is also suitable for reactive loads. It is however necessary to run the simulation not just for one cycle, but sometimes for as many as twenty. This is to ensure that steady-state conditions have been reached.

The diagrams referred to below are for steady-state 200Hz sinewave drive; the frequency must be defined so the load impedance can be set by suitable component values, but otherwise makes no difference.

Figure 8 shows what happens in Class-B emitter follower when driving a 45° capacitive-reactive load with a modulus of 8Ω. Comparing it with Fig. 4, the power drawn from the supply is essentially unchanged, and is still proportional to output voltage fraction.

The larger areas at the bottom show that more power is being dissipated in the output devices and correspondingly less in the load, because the phase shift causes the voltage across and the current through the output devices to overlap more. The amplifier must dispose of 95W of heat worst-case, rather than 60W.

Average device dissipation no longer peaks, but increases monotonically up to maximum output. 45° phase angles are common when loudspeakers are driven. It is generally accepted that an amplifier should be able to provide full voltage swing into such a load.

When the load is purely reactive, with a phase angle of 90°, it can dissipate no power and so all that delivered to it is re-absorbed and dissipated in the amplifier. **Figure 9** shows that the worst-case device dissipation is much greater at 185W, absorbing all the power drawn from the supply, and therefore necessarily increasing monotonically with output level; there is no maximum at medium levels.

This is a very severe test for a power amplifier. It is also unrealistic, as no assemblage of moving-coil speaker elements can ever present a purely reactive impedance; 60° loads are normally the most reactive catered for. **Table 1** shows the worst-case cycle-averaged dissipation for various load angles, showing how the position of maximum dissipation moves towards full output as the angle increases.

This is best displayed in 3D, as **Fig. 10**, which plots power vertically; the slight hump at the front – non-reactive load – disappearing as the load becomes more reactive.

The dissipation hump is of little practical significance. An audio amplifier will almost certainly be required to drive 45° loads, and these cause higher power dissipations than resistive loads driven at any level.

Figure 11 shows the same plot for peak power, which increases monotonically with both output fraction and load angle. **Figure 12** summarises all this data for design purposes. It shows worst-case peak and average power in one output device against load reactance.

Peak powers are taken at 0.95 of full output, average power at whatever output fraction gives maximum dissipation. Therefore to design an amplifier to cope with 45° loads, note that average power is increased by 1.4 times, and peak power by 2.7 times, over the resistive case. This can mean that it is necessary to increase the number of output devices simply to cope with the

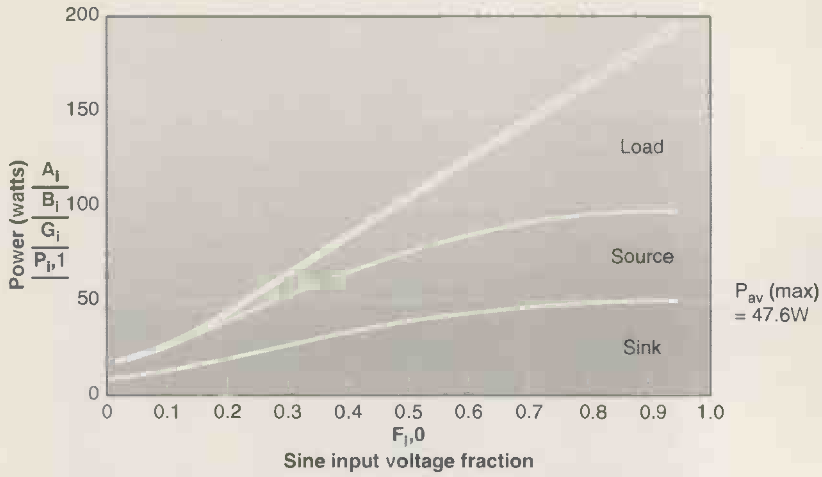


Fig. 8. Class-B emitter follower reactive 45°, sinewave drive. The load is 11.3Ω in parallel with 71μF. Impedance modulus is 8Ω at 200Hz. Amplifier dissipation is increased, power delivered to the load decreased.

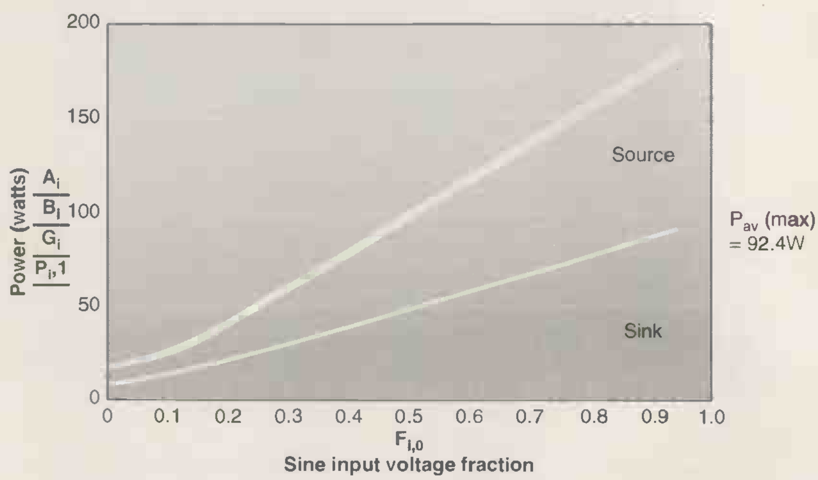


Fig. 9. Class-B emitter follower reactive 90°, sinewave drive. Load is a 99.5μF capacitor. Impedance modulus still 8Ω at 200Hz. All the supply power is now being absorbed by the amplifier, and none by the load.

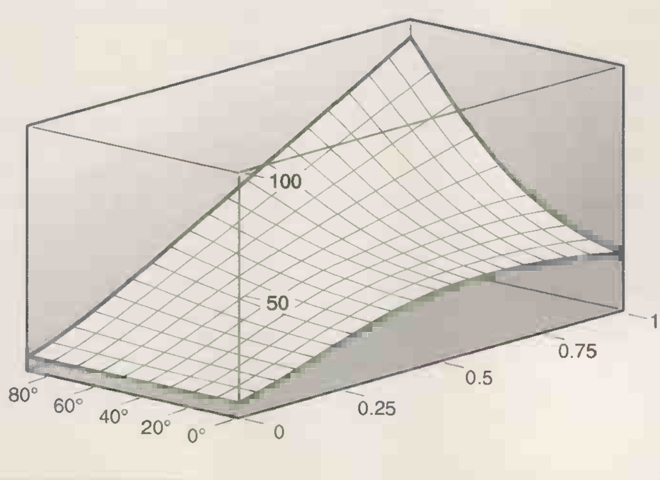


Fig 10. The average power (vertical axis) against load angle (left-hand horizontal axis) and output fraction (r.h. horiz. axis)

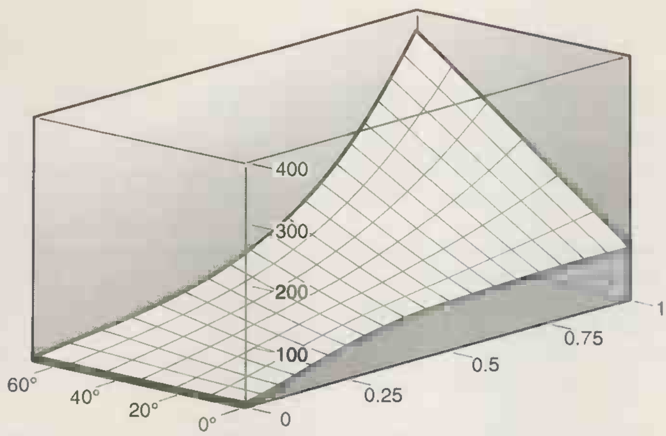


Fig 11. Peak power plotted as in Fig. 10. The vertical scale must accommodate much higher power levels than Fig. 10.

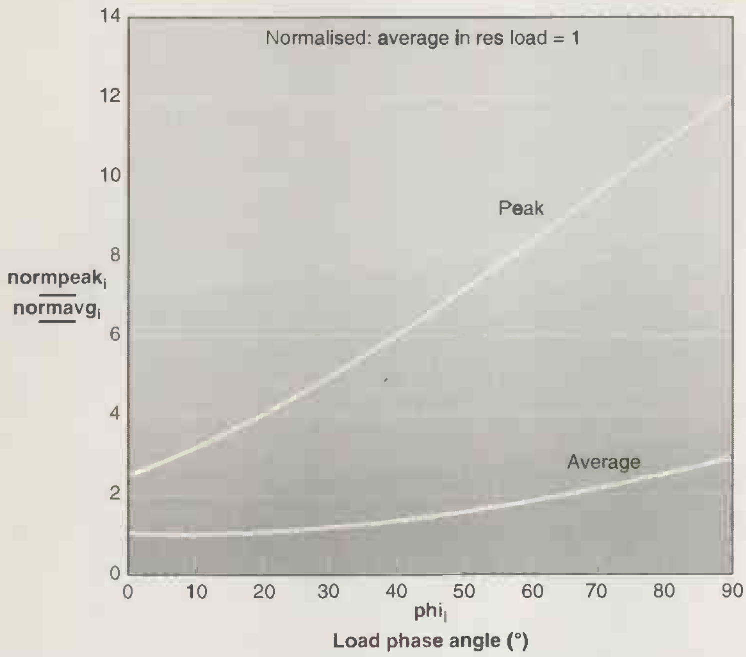


Fig 12. Peak power increases faster than worst-case average power as the load becomes more reactive and its phase angle increases. Class-B emitter follower as before.

Reactive loads observations

The following conclusions apply to reactive loads.

- Amplifier power consumption and average supply current drawn do not vary with load phase angle if the impedance modulus remains constant.
- Peak device current is not altered so long as the impedance modulus remains constant.
- Average current in the output devices is not altered so long as the impedance modulus remains constant. This follows from the first observation.
- Peak device power increases rapidly, as the load becomes more reactive. A 45° load increases power peaks by 2.7 times, and a 60° load by 3.4 times. See Fig. 12.
- Average device dissipation also increases, but more slowly, as the load angle increases. A 45° load increases average dissipation by 1.4 times and a 60° load by 1.8 times, Fig. 12.

much enhanced peak power.

Considering simple reactive loads like those listed in the panel 'Reactive load observations' gives an essential insight into the extra stresses they impose on semiconductors but is still some way removed from real signals and real loudspeaker loads, where the impedance modulus varies along with the phase, due to electromechanical resonances or crossover dips.

I looked at single and two-unit loudspeaker models in ref. 3 where the maximum phase angle found was 40°. In brief, the results were:

- Amplifier power consumption and average supply current drawn vary with frequency due to impedance modulus changes.
- The peak device current increased by a maximum of 1.3 times at the modulus minima.
- The average current in the output devices increased by a maximum of 1.3 times.
- Peak device power increased by a maximum factor of 2, mostly due to phase shift rather than impedance dips.
- Average device dissipation increased by a maximum of 1.4 times.

These numbers come from two specific models that attempted to represent 'average' speakers. Worse conditions could easily have been found.

Ultimately a comprehensive survey of the loudspeakers on the market would be required, but this would be very time-consuming. In ref. 4, which gives an excellent account of real speaker loading, 21 models were tested and the worst angle found was 67°. Eliminating the two most extreme cases reduced this to 60°.

The most severe effect of reactive loads is the increase in peak power, followed by the increase in average power.

Both are a strong function of load phase, and so the specification of the maximum angle to be driven has a big effect on the devices required, heatsink design, and hence on amplifier cost.

It is likely that a failure to appreciate just how quickly peak power increases with load angle is the root cause of many amplifier failures. ■

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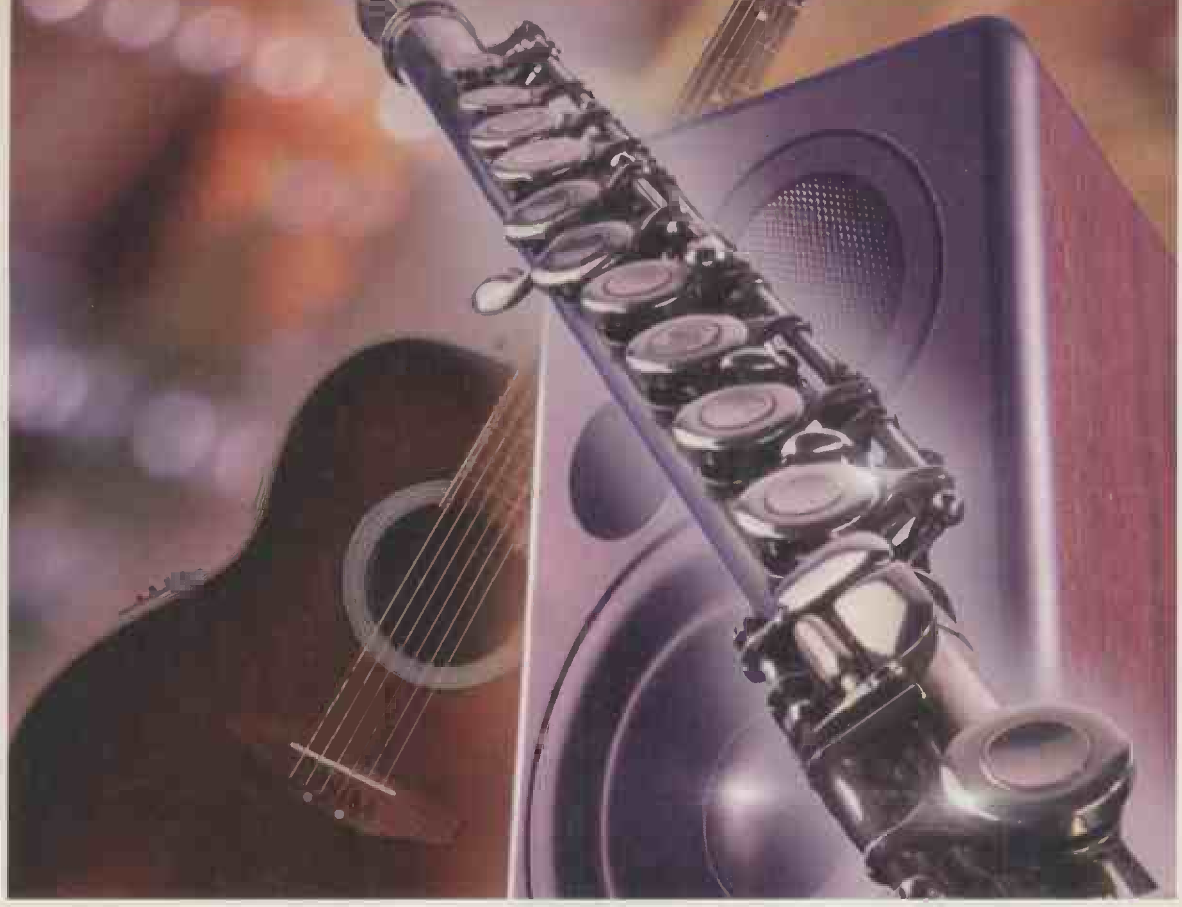


Stereo from all angles II

Obtaining consistently high quality in stereo sound reproduction is not a trivial problem. It requires an understanding of many aspects of sound and music. John Watkinson's second article begins by showing how the operation of the ear determines the structure of music and then moves on to consider how the ears determine direction.

Critical bandwidth has a large bearing on the way music has evolved. Beats are used extensively in music. When tuning a pair of instruments together, a small tuning error will result in beats when both play the same nominal note. In certain pipe organs, pairs of pipes are sounded together with a carefully adjusted pitch error, which results in a pleasing tremolo effect.

With certain exceptions, music is intended to be pleasing and so dissonance is avoided. Two notes that sound together in a pleasing manner are described as consonant. Two sine waves appear consonant if they are separated by a critical bandwidth because the roughness described in last month's article is avoided. However, real musical instruments produce a series of harmonics in addition to the fundamental and dissonance between the harmonics is then possible.

Figure 1 shows the spectrum of a harmonically-rich instrument. The fundamental and the first few harmonics are separated by more than a critical band. But from the seventh harmonic more than one harmonic will be in one critical band and it is possible for dissonance to occur.

Musical instruments have evolved to avoid the production of seventh and higher harmonics. Violins and pianos, for example, are played or designed to excite the strings at a node of the seventh harmonic to suppress this dissonance.

Harmonic distortion in audio equipment is easily detected – even in

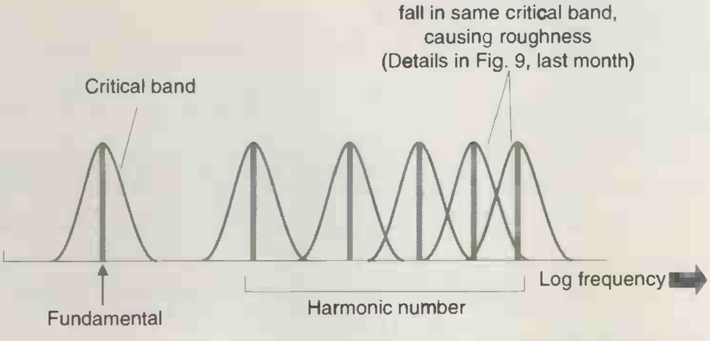


Fig. 1. Harmonics exist on a linear scale whereas critical bandwidth is constant on a logarithmic scale

minute quantities – because the first few harmonics fall in non-overlapping critical bands. Second-harmonic distortion appears to be much less offensive than third-harmonic distortion, to which the ear is very sensitive. Higher harmonics are generally lower in amplitude so it is often assumed that if the third harmonic performance is acceptable, the remaining harmonics will be inaudible.

Harmonic-rich notes together

When two harmonically-rich notes are sounded together, the harmonics will fall within the same critical band and this can cause dissonance unless the fundamentals have one of a limited number of simple relationships which makes the harmonics fuse. Clearly an octave relationship is perfect.

Figure 2 shows some examples. In **2a)** two notes with the ratio (interval) 3:2 are considered. The harmonics are either widely separated or fused and the combined result is highly consonant.

The interval of 3:2 is known to musicians as a perfect fifth. In **2b)** the ratio is 4:3. All harmonics are either at least a third of an octave apart or are fused. This relationship is known as a perfect fourth. The degree of dissonance over the range from 1:1 to 2:1 – unison to octave – was investigated by Helmholtz and is shown in Fig. **2c)**. Note that the dissonance rises at both ends where the fundamentals are within a critical bandwidth of one another.

Dissonances in the centre of the scale are where some harmonics lie within a critical bandwidth of one another. Troughs in the curve indicate areas of consonance. Many of the troughs are not very deep, indicating that the consonance is not perfect. This is because of the effect shown in Fig. 1, in which high harmonics get closer together with respect to critical bandwidth.

When the fundamentals are closer together, the harmonics will become dissonant at a lower frequency, reducing the consonance. Figure 2c) also shows the musical terms used to describe the consonant intervals.

The notes of the musical scale have been arrived at empirically to allow the maximum consonance with pairs of notes and chords. Early instruments were tuned to the just diatonic scale in exactly this way. Unfortunately the just diatonic scale does not allow changes of key because the notes are not evenly spaced.

A key change is where the frequency of every note in a piece of music is multiplied by a constant, perhaps to bring the accompaniment within the range of a singer. In continuously tuned instruments such as the violin and the trombone this is easy, but with fretted or keyboard instruments such as piano there is a problem.

The equal tempered scale is a compromise between consonance and key changing. The octave is divided into twelve equal intervals called tempered semitones. On a keyboard, seven of the keys are white and produce notes very close to those of the just diatonic scale, and five of the keys are black. Music can be transposed in semitone steps by using the black keys.

Frequency versus pitch

Frequency is an objective measure whereas pitch is the subjective near equivalent. Frequency and level are independent, whereas pitch and level are not. **Figure 3** shows the relationship between pitch and level.

Place theory indicates that the hearing mechanism can sense a single frequency quite accurately as a function of the place or position of maximum basilar vibration. However, most periodic sounds and real musical instruments produce a series of harmonics in addition to the fundamental.

When a harmonically rich sound is present the basilar membrane is excited at spaced locations. **Figure 4a)** shows all harmonics, **4b)** shows even harmonics predominating and **4c)** shows odd harmonics predominating.

It would appear that our hearing is accustomed to hearing harmonics in

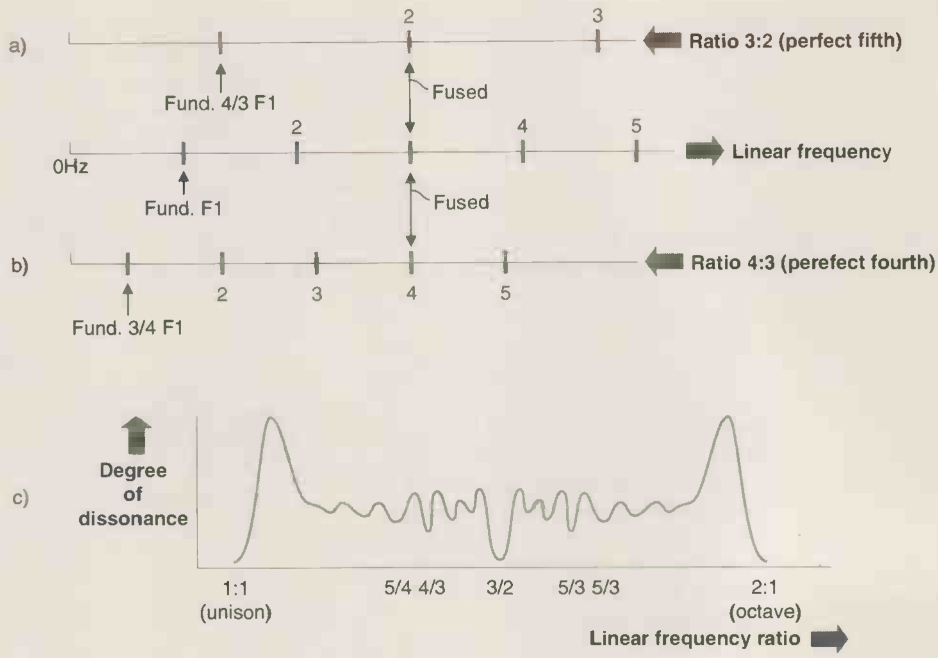


Fig. 2. Examples of consonance and dissonance. In a), two notes with a 3:2 ratio are considered. Their harmonics are either widely separated or fused, and the combined result is highly consonant. In b), the ratio is 4:3. Graph c) shows the degree of dissonance over the range 1:1 to 2:1.

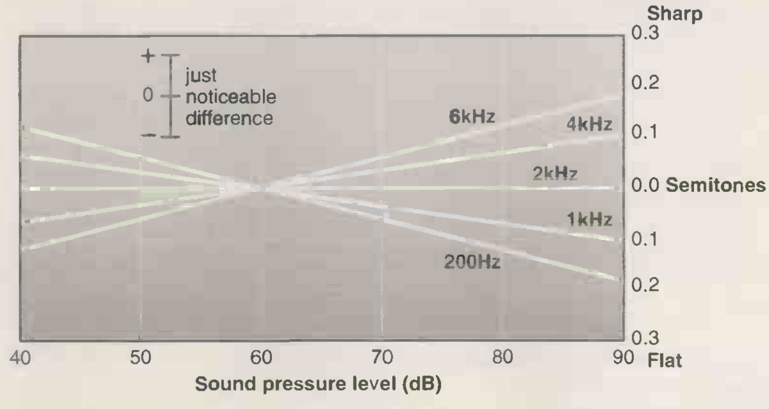


Fig. 3. Sensation of pitch is level dependent.

various amounts and the consequent regular pattern of excitation. It is the overall pattern which contributes to the sensation of pitch even if individual partials vary enormously in relative level.

As the fundamental frequency rises it is difficult to obtain a full pattern of harmonics as most of them fall outside the range of hearing. The pitch discrimination ability is impaired and needs longer to operate.

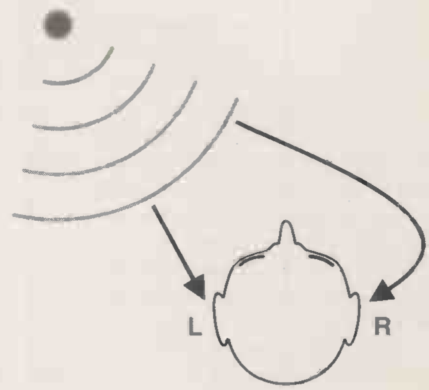
Figure 5 shows the number of cycles of excitation needed to discriminate pitch as a function of frequency. Clearly at around 5kHz, performance is failing because there are hardly any audible harmonics left. Phase locking also fails at about the same frequency. Musical instruments have evolved accordingly, with the highest notes of virtually all instruments found below 5kHz.

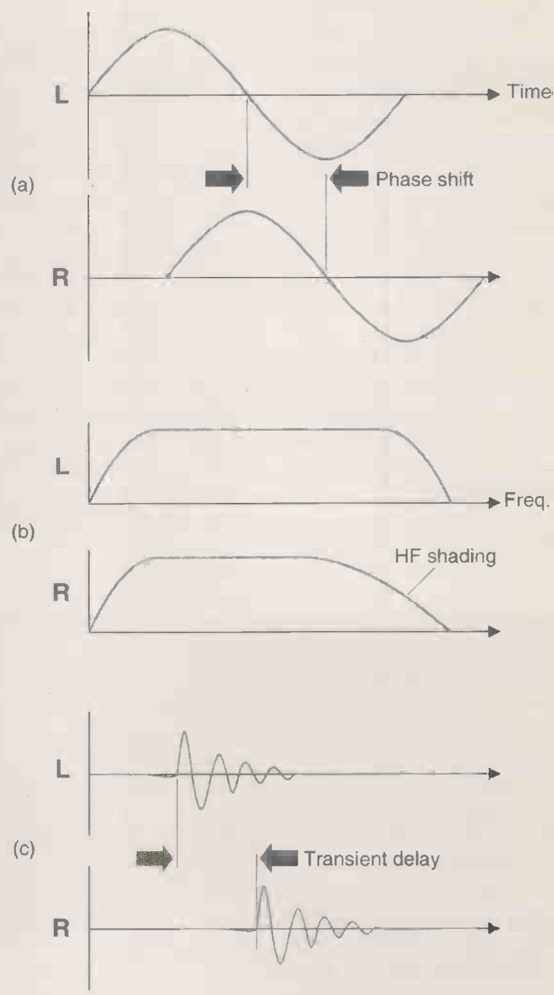
The hearing mechanism works in both the time and frequency domains. The frequency-domain response works relatively slowly, aiding the determination of pitch and timbre, and evolved later. The time-domain response works quickly, primarily aiding the direction sensing mechanism and is older in evolutionary terms. It evolved as a mechanism to assist flight from predators.

Direction sensing

Considering direction sensing, **Figure 6** shows that the brain can examine the differences between the signals from the two ears. At **6a)** a phase shift will be apparent between the two versions of a tone picked up at the two ears unless the source of the tone is dead ahead – or behind. At **6b)** the distant ear is shaded by the head resulting in reduced response com-

Fig. 6. Direction sensing mechanisms: a) phase shift b) shading c) inter-aural delay





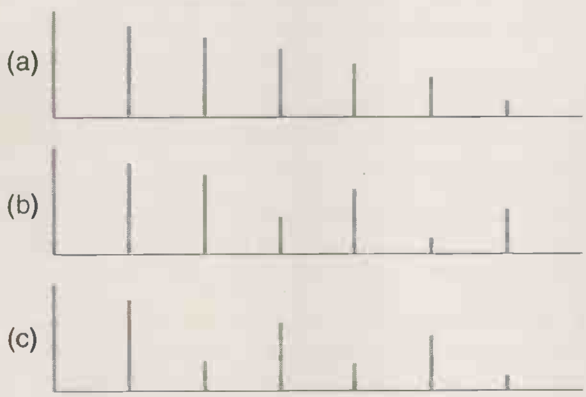


Fig. 4. Pitch recognition is based on the pattern of harmonics rather than their levels. Spectra a), b) and c) all have the same pitch.

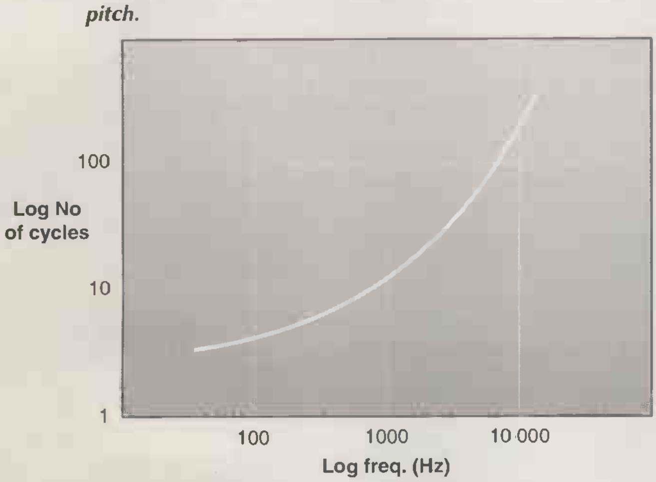


Fig. 5. Number of cycles required to discriminate pitch is frequency dependent

pared to the nearer ear. At **6c)** a transient sound arrives later at the more distant ear.

There will be considerable variation in phase shift with frequency. At a low frequency such as 30Hz, the wavelength is around 11.5 metres. Even if heard from the side, the ear spacing of about 0.2 metres will result in a phase shift of only 6° and so this mechanism must be quite weak at low frequencies.

At a high frequency such as 10kHz, the ear spacing is many wavelengths. Variations in the path length difference will produce a confusing and complex phase relationship.

The problem with tones or single frequencies is that they produce a sinusoidal waveform, one cycle of which seems much like another, leading to ambiguities in the time between two versions, **Fig. 7a)**. This suggests a frequency limit of around 1500Hz which has been confirmed by experiment.

The shading mechanism of Fig. 6b) will be a function of the important relationship between sound wavelength and the dimensions of the head. This suggests that at low and middle frequencies sound will diffract round the head sufficiently well that there will be no significant difference between the level at the two ears.

Only at high frequencies does sound become directional enough for the head to shade the distant ear causing what is called an inter-aural intensity difference.

At very high frequencies, the shape of the pinnae must have some effect on the sound that is a function of direction. It is thought that the pinnae allow some height discrimination.

Phase differences are only useful at low frequencies and shading only works at high frequencies. Fortunately real-world sounds are timbral or broadband and often contain transients.

Multiple frequencies

Timbral, broadband and transient sounds differ from tones in that they contain many different frequencies. A transient has a unique aperiodic waveform which **Fig. 7b)** shows has the advantage that there can be no ambiguity in the inter-aural delay between two versions.

Figure 8 shows the time difference for different angles of incidence for a typical head. Note that a one-degree change in sound location causes a inter-aural delay of around 10µs. The smallest detectable delay is a remarkable 6µs.

The actual transducer in the ear, which converts from vibration to nerve impulses is the basilar membrane and its associated hair cells. The basilar membrane is a frequency-analysis device that produces nerve impulses from different physical locations according to which frequencies are present in the incident sound.

Clearly when a transient sound arrives from one side, many frequencies will be excited simultaneously in the nearer ear, resulting in a pattern of nerve firings. This will closely be followed by a similar excitation pattern in the further ear. Shading may change the relative amplitudes of the higher frequencies, but it will not change the pattern of frequency components present.

Harmonics and nerve firings

The presence of harmonics means that a greater number of nerve firings can be compared between the two ears. As the statistical deviation of nerve firings with respect to the incoming waveform is about 100µs the only way an inter-aural delay of 6µs can be perceived is if the timing of many nerve firings is correlated in some way in the brain. Stereo sound reproduction systems ought to be able to reach this degree of accuracy for greatest realism.

Transient noises produce a one-off pressure step whose source is accu-

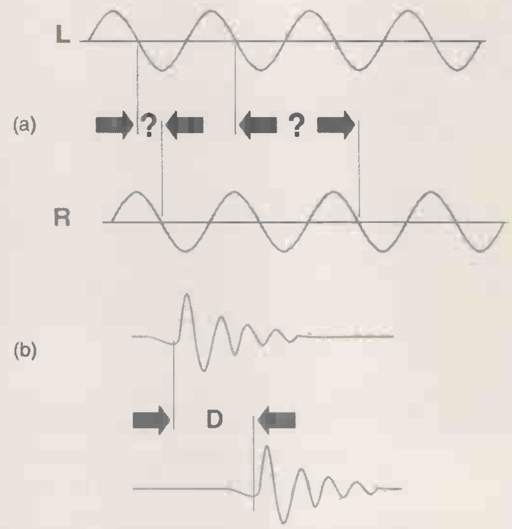


Fig. 7a) measuring delay between two sine-waves is ambiguous. As in b), with transients there is less ambiguity

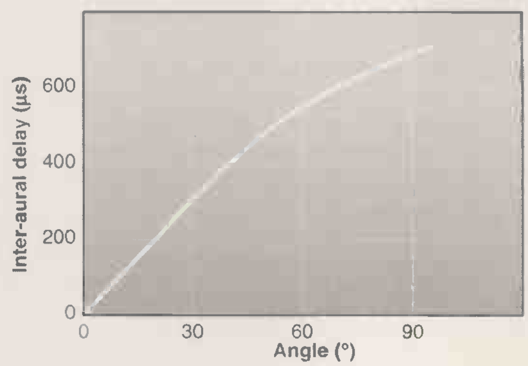


Fig. 8. Inter-aural delay is a function of the angle of the sound source from the median (straight ahead) position.

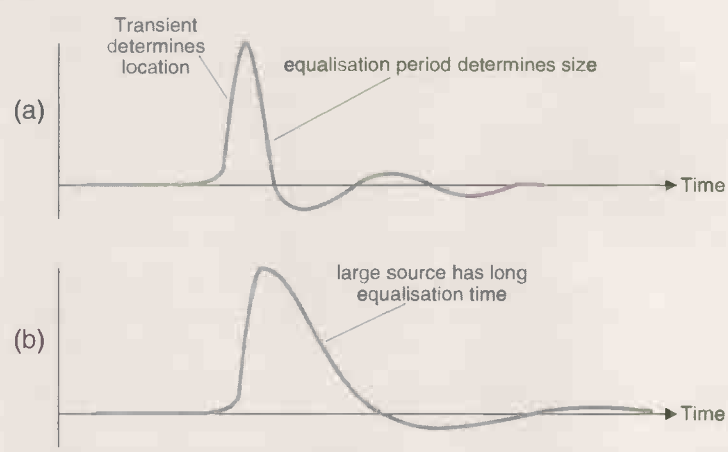


Fig. 9. A transient has a sharp leading edge and an equalisation period that is a function of the size of the object making the sound. The leading edge of the transient is used by the ear to determine the location of the object and this is known before the ear has analysed the timbre. A small source a) has a short equalisation time whereas it is longer in a large source, b). If this part of a transient is not reproduced properly there will be a lack of realism. Only a linear-phase speaker can do this; reflex speakers cannot.

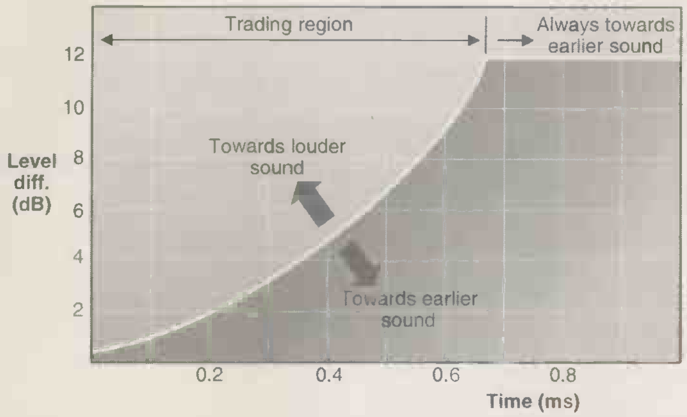


Fig. 10. Before about 700µs, the origin of a sound can be disturbed by a reflection. This error in location is a function of the relative level of the reflection. Sharp-edged rectangular speakers produce strong reflections in exactly this delay region and so are not capable of stereophonic fidelity.

rately and instinctively located. **Figure 9** shows an idealised transient pressure waveform following an acoustic event. Only the initial transient pressure change is required for location. The time of arrival of the transient at the two ears will be different and will locate the source laterally within a processing delay of around a millisecond, far before any perception of timbre has taken place.

Following the event that generated the transient, the air pressure equalises. The time taken for this equalisation varies and allows the listener to establish the likely size of the sound source. The larger the source, the longer the pressure equalisation time.

How important are transients?

The above results suggest that anything in a sound reproduction system that impairs the reproduction of a transient pressure change will damage both localisation and the assessment of the pressure equalisation time. Clearly in an audio system that claims to offer any degree of precision, every component must be able to reproduce transients accurately and should ideally be phase linear.

Human hearing can identify the position of a number of different sound sources simultaneously. The hearing mechanism must be constantly comparing excitation patterns from the two ears with different relative delays. Strong correlation will be found where the delay corresponds to the interaural delay for a given source. This is apparent in the binaural threshold of hearing which is 3 to 6dB better than monaural in the speech band.

Using a variable delay the hearing mechanism has an ability to concentrate on one sound source out of many. Sounds arriving from other directions are incoherent and are heard less well. This is known as attentional selectivity but is more usually referred to as the cocktail-party effect.

Tailored responses

The inter-aural phase, delay and level mechanisms vary in their effectiveness depending on the nature of the sound to be located.

A fixed response to each mechanism would be ineffective. For example on a low-frequency tone, the time-of-arrival mechanism is useless whereas on a transient it excels.

The different mechanisms are quite separate on one level, but at some point in the brain's perception a fuzzy logic or adaptive decision has to be made as to how the outcome of these mechanisms will be weighted to make the final judgement of direction.

In practice we hear very well in reverberant surroundings – far better than microphones can – because of the transform nature of the ear and the way in which the brain processes nerve signals.

When two or more versions of a sound arrive at the ear, provided they fall within a time span of about 30ms, they will not be treated as separate sounds, but will be fused into one sound which appears to have come from one place.

Only when the time separation reaches 50-60ms do the delayed sounds appear as echoes from different directions. As we have evolved to function in reverberant surroundings, reflections do not impair our ability to locate the source of a sound.

The first version of a transient sound to reach the ears must be the direct sound rather than a reflection. Consequently the ear has evolved to attribute source direction from the time of arrival difference at the two ears of the first version of a transient. Later versions, which may arrive from elsewhere, simply add to the perceived loudness but do not change the perceived location of the source.

The Dutch researcher Haas found that this precedence effect is so powerful that even when later arriving sounds are artificially amplified – a situation that does not occur in nature – the location still appears to be that from which the first version arrives.

Investigating weighting

Experiments have been conducted in which the delay and intensity clues are contradictory to investigate the way the weighting process works. The same sound is produced in two locations but with varying relative delay and shading-dependent level. The way that the listener perceives an apparent sound direction reveals how the directional clues are weighted.

Within the maximum inter-aural delay of about 700µs the precedence effect does not function and the perceived direction can be pulled away from that of the first arriving source by an increase in level.

Figure 10 shows that this area is known as the time-intensity trading region. Once the maximum inter-aural delay is exceeded, the hearing mechanism knows that the time difference must be due to reverberation and the trading ceases to change with level.

It is important to realise that in real life the hearing mechanism expects a familiar sound to have a familiar weighting of phase, time of arrival and shading clues. A high-quality sound reproduction system must do the same if a convincing spatial illusion is to be had.

Consequently a stereo system that attempts to rely on just one of these effects will not sound realistic. Worse still is a system that relies on one effect to be dominant, but where another is contradictory. Time-intensity trading is an interesting insight into the operation of the hearing mechanism, but it cannot be used in quality sound reproduction because although the ear is fooled, it is attempting to resolve conflicting stimuli and the result is inevitably listening fatigue. ■

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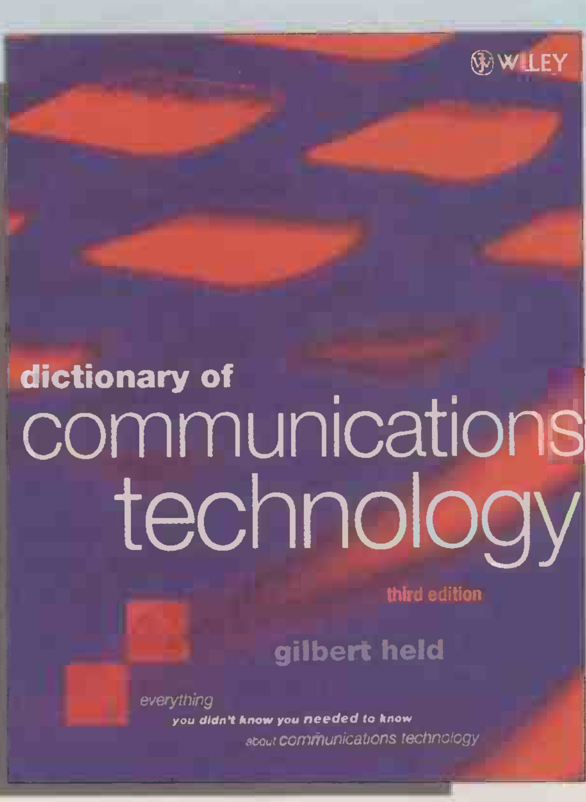

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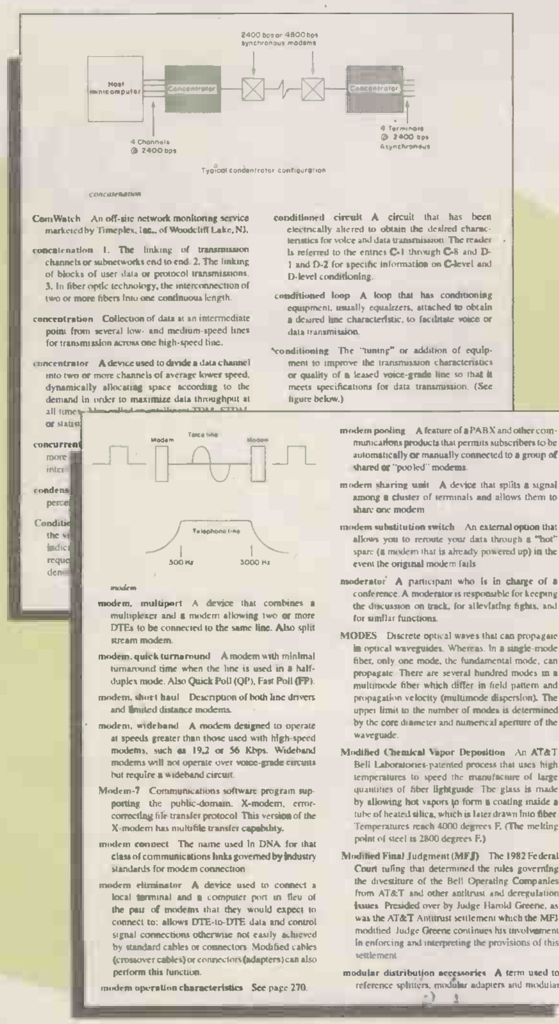

2400 bps or 4800 bps synchronous modems

Host minicomputer — Concentrator — Concentrator

4 Channels @ 2400 bps

4 Terminals @ 2400 bps Asynchronous

Typical concentrator configuration

concatenation

ComWatch An off-site network monitoring service marketed by Timeplex, Inc., of Woodcliff Lake, NJ.

concatenation 1. The linking of transmission channels or subnetworks end to end. 2. The linking of blocks of user data or protocol transmissions. 3. In fiber optic technology, the interconnection of two or more fibers into one continuous length.

concentration Collection of data at an intermediate point from several low- and medium-speed lines for transmission across one high-speed line.

concentrator A device used to divide a data channel into two or more channels of average lower speed, dynamically allocating space according to the demand in order to maximize data throughput at all times...

conditioned circuit A circuit that has been electrically altered to obtain the desired characteristics for voice and data transmission. The reader is referred to the entries C-1 through C-8 and D-1 and D-2 for specific information on C-level and D-level conditioning.

conditioned loop A loop that has conditioning equipment, usually equalizers, attached to obtain a desired line characteristic, to facilitate voice or data transmission.

conditioning The "tuning" or addition of equipment to improve the transmission characteristics or quality of a leased voice-grade line so that it meets specifications for data transmission. (See figure below.)

Modem — Telco line — Modem

Telephone line

500 Hz 3000 Hz

modem

modem, multiport A device that combines a multiplexer and a modem allowing two or more DTEs to be connected to the same line. Also split stream modem.

modem, quick turnaround A modem with minimal turnaround time when the line is used in a half-duplex mode. Also Quick Poll (QP), Fast Poll (FP).

modem, short haul Description of both line drivers and limited distance modems.

modem, wideband A modem designed to operate at speeds greater than those used with high-speed modems, such as 19.2 or 56 Kbps. Wideband modems will not operate over voice-grade circuits but require a wideband circuit.

Modem-7 Communications software program supporting the public-domain, X-modem, error-correcting file transfer protocol. This version of the X-modem has multifile transfer capability.

modem connect The name used in DNA for that class of communications links governed by industry standards for modem connection.

modem eliminator A device used to connect a local terminal and a computer port in lieu of the pair of modems that they would expect to connect to: allows DTE-to-DTE data and control signal connections otherwise not easily achieved by standard cables or connectors. Modified cables (crossover cables) or connectors (adapters) can also perform this function.

modem operation characteristics See page 270.

modem pooling A feature of a PABX and other communications products that permits subscribers to be automatically or manually connected to a group of shared or "pooled" modems.

modem sharing unit A device that splits a signal among a cluster of terminals and allows them to share one modem.

modem substitution switch An external option that allows you to reroute your data through a "hot" spare (a modem that is already powered up) in the event the original modem fails.

moderator A participant who is in charge of a conference. A moderator is responsible for keeping the discussion on track, for alleviating fights, and for similar functions.

MODES Discrete optical waves that can propagate in optical waveguides. Whereas, in a single-mode fiber, only one mode, the fundamental mode, can propagate. There are several hundred modes in a multimode fiber which differ in field pattern and propagation velocity (multimode dispersion). The upper limit to the number of modes is determined by the core diameter and numerical aperture of the waveguide.

Modified Chemical Vapor Deposition An AT&T Bell Laboratories-patented process that uses high temperatures to speed the manufacture of large quantities of fiber lightguide. The glass is made by allowing hot vapors to form a coating inside a tube of heated silica, which is later drawn into fiber. Temperatures reach 4000 degrees F. (The melting point of steel is 2800 degrees F.)

Modified Final Judgment (MFJ) The 1982 Federal Court ruling that determined the rules governing the divestiture of the Bell Operating Companies from AT&T and other antitrust and deregulation issues. Presided over by Judge Harold Greene, as was the AT&T Antitrust settlement which the MFJ modified. Judge Greene continues his involvement in enforcing and interpreting the provisions of this settlement.

modular distribution accessories A term used to reference splitters, modular adapters and modular

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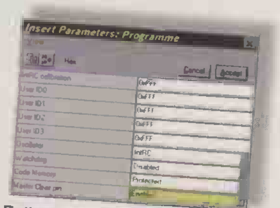

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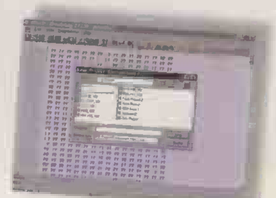

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Hands-on Internet

Designing filters used to be a bit of a slog, but now there's a number of time-saving filter CAD tools available free on the net, as Cyril Bateman has been finding out.

Designing time continuous filters – both active and passive versions – requires a number of mathematical calculations. These can be time consuming.

Traditional shortcut methods rely on use of standard filter design tables. While helpful, these tables often require awkward component values that are difficult to make up with resistors and impossible with capacitors.

When designing a band-pass filter to form part of an instrument, I decided to look for other more user friendly design methods, available on Internet. One good overview of filter design, called 'A Basic introduction to Filters – Active, Passive and Switched-Capacitor' can be found in application note 779 from National Semiconductor.[4] These 22 pages of advice can be downloaded as AN-779.PDF.

Before the introduction of integrated circuits, most filter applications were served by passive filter techniques. Today at high frequencies or when the lowest noise or widest dynamic range is needed, passive filters can still provide the best possible performance. Passive filters become almost inevitable when filtering a supply line carrying current.

At low frequency, the behaviour of both capacitors and inductors is predictable and can easily be simulated using Spice analysis. This is not so as frequency increases. Capacitance values reduce, while ESR increases rapidly.

Self-capacitance of inductors quickly dominates the inductor's behaviour. As a result conventional simulation methods and design tables become of little value and designs usually revert to cut-and-try methods.

Using a Spice based simulator it is difficult – maybe even impossible –

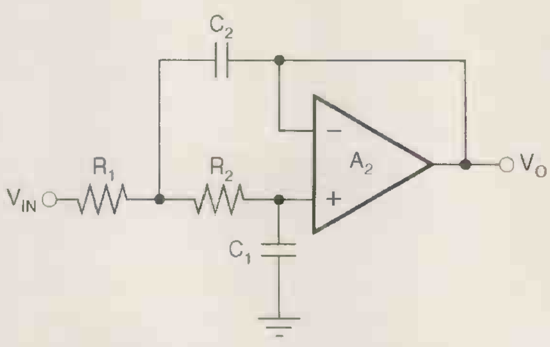


Fig. 1. Schematic of a second-order, unity-gain, Sallen and Key, low-pass filter. Addition of two feedback resistors achieves higher gains.

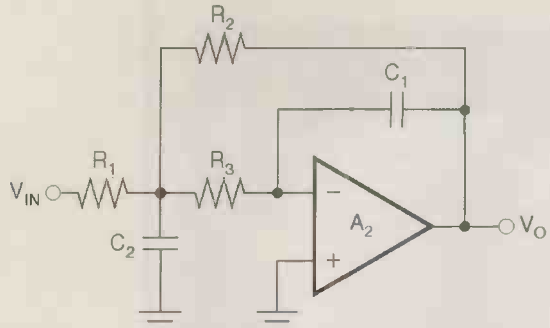


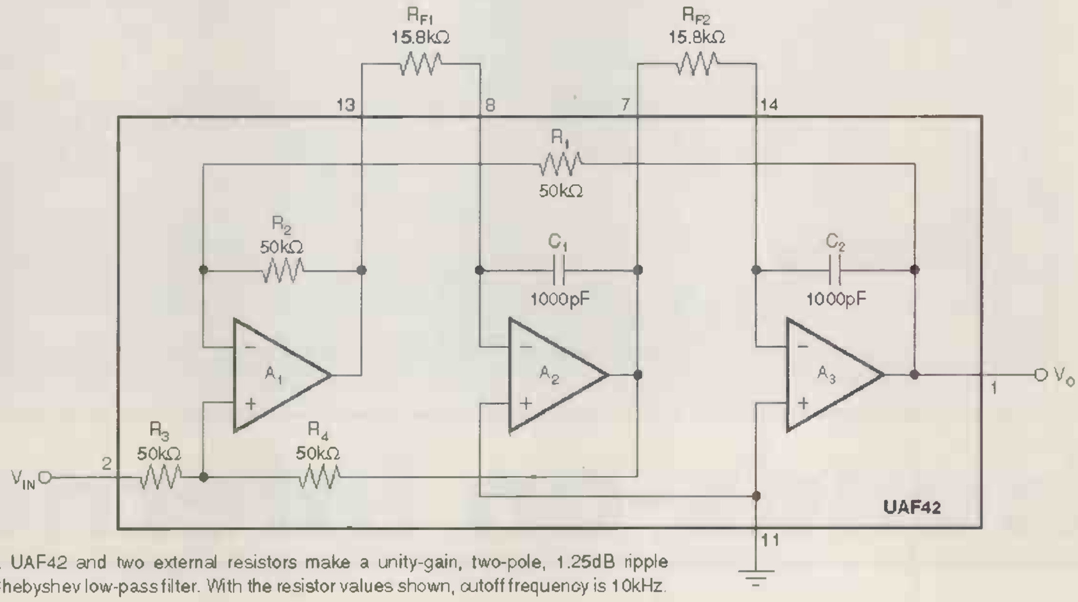
Fig. 2. Schematic of a second-order, multiple-feedback, low-pass filter having a gain that depends on the ratio of R_1 to R_2. This filter style is often called a 'Rauch' filter.

to compensate for these component variations, especially at high frequency. These difficulties resulted in my writing an easy to use frequency-domain simulator, dedicated to design of low-pass EMC filters.[5] My Internet searches failed to find any similar software for the design of general-purpose passive filters.

At low frequencies, the behaviour of both active-time-continuous and switched-capacitor filters is predictable. Active-time-continuous filter designs are usually based on the Sallen and Key **Fig. 1**, multiple-feedback **Fig. 2** or state-variable **Fig. 3** circuit styles. Equations to calculate component values for these styles, in low-pass, band-pass and high-pass configurations, are available in most good designers' reference books.

Unfortunately, while solving these equations is not difficult, repetitive calculations are needed to optimise performance and comply with standard component values.

Fig. 3. This unity gain second-order Chebyshev state-variable filter has been configured using a UAF42 integrated circuit and Burr-Brown's Filter42 *software. Its frequency is 10kHz and its ripple is 1.25dB. While the state-variable filter needs additional op-amps and components, it can provide high-pass, low-pass and band-pass simultaneous outputs.*



A UAF42 and two external resistors make a unity-gain, two-pole, 1.25dB ripple Chebyshev low-pass filter. With the resistor values shown, cutoff frequency is 10kHz.

Bugs

In January a rapidly spreading virus appeared that affects Windows 95, 98 and NT users, but not OS/2 or MacOS. The 'Win32/Ska.A' virus, is a 10Kbyte executable that attaches to e-mails or news messages. Infected machines then re-attach this file to outgoing messages, without the sender's knowledge. It has replicated itself extremely rapidly.[1] If you receive an e-mail with 'happy99.exe' attached and run this virus file, you will see a fireworks display as in **Fig. A**.

Your computer will continue to function apparently as normal but any e-mail or posting you send will now automatically include a copy of happy99, spreading this infection to your recipients. While running the fireworks display, happy99 copies your original WSOCK32.DLL as WSOCK32.SKA. It introduces two new files, SKA.EXE, SKA.DLL then modifies your working WSOCK32.DLL.

If your machine has been infected and you have passed on this virus to others, you will find a list of recipients in the file 'LISTE.SKA'. This file can be read using Notepad.

An infected computer is easily cleaned. Symantec[2] and DataFellows[3] provide full descriptions of the virus actions, together with simple clean-up instructions and advice.

This virus has appeared under many names besides 'Happy99.exe', for example, 'Trojan.exe', 'I-Worm.exe', 'Ska.exe'. Since a malicious sender could easily rename this file, it is best not to run any unexpected executable received via e-mail or a news message.

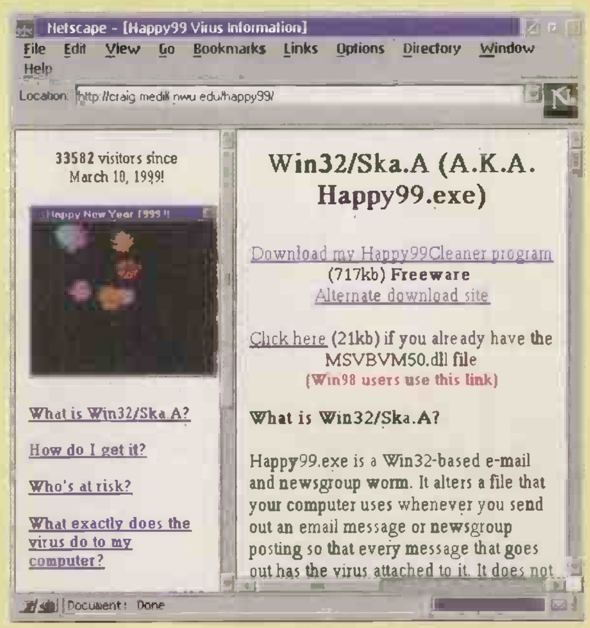


Fig. A. This site had over 30 000 visitors in just two weeks when I visited, reflecting just how quickly this virus has spread, both in Europe and the US. Many other sites also offer diagnostic information. To minimise further spread of this virus, e-mail and news-group users should check-out their computers for infection.

Sallen and Key filters

Dale Eager describes one simple tabular method for designing low-pass Sallen and Key filters on pages 22 to 24 of application note 67. This Linear Technology note includes several other useful filter topics.[6]

To simplify calculating these design equations, it is usual to accept equal values for the resistors used in the filter. As a result one is then left needing capacitors having impossible values. Dale Eager's tabular design method uses only standard E3 capacitor values together with E24, 5% tolerance resistors.

Dale's design tables list resistor and capacitor values needed for Bessel and Butterworth filters for each of 24 cut-off frequencies covering one decade. The only mathematics needed are for scaling these tabled values to your desired frequency decade. To confirm the results provided by these tabled values, Dale provides a PSpice simulation of a 1.6kHz third order Butterworth filter which he designed using table values, **Fig. 4**.

If you need other Sallen & Key filter styles, a series of application notes from National Semiconductor, written by Kumen Blake, might help. They are OA-27 for low-pass, OA-28 for band-pass and OA-29 for high-pass. Each details design methods for producing filters with low sensitivity to component and op-amp variations.[4]

To ensure the best production yields, these nominal filter designs must also compensate for component and printed-board parasitics. To this end, 'Component Pre-Distortion for Sallen and Key Filters' is discussed in application note OA-21, which can be downloaded from National Semiconductor.[4] Supported by Spice Monte-Carlo simulations, this method should ensure a reproducible design.

Multiple-feedback filters

Perhaps you prefer the multiple-feedback filter configuration. It provides a 'Q' value less sensitive to component variations than does the Sallen and Key circuit.

Burr-Brown provides the software suite *FilterPro*, which contains two DOS based programs.[7] The '*Filter2*' package facilitates the design of multiple-feedback low-pass filters. It caters for the three most commonly used filter types, Butterworth, Bessel and Chebyshev.

Particularly useful, this software also supports Sallen and Key designs. It can provide a sensitivity display for the chosen filter, re-calculating F_0 and 'Q' for a 1% change of component values.

Application note AB-034B[7] provides full details of this application software, which allows easy 'what-if' variations to be evaluated and the simulated responses plotted on screen, **Fig. 5**.

State-variable filters

Many designers favour the state-variable filter approach. While this

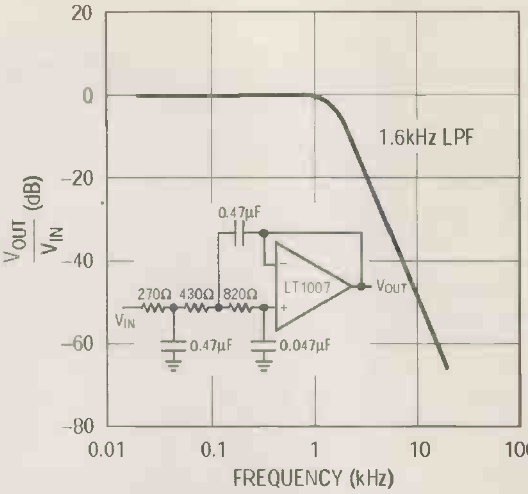


Fig. 4. This PSpice simulation shows the performance of a 1.6kHz third order, unity gain, Sallen-Key low-pass Butterworth filter designed using Dale Eager's filter design tables. Only standard component values and 5% tolerances are involved.

```
Press <enter> to enter capacitor menu—      Circuit.......... Sallen-Key
                                              Butterworth....
This option allows you to enter               Chebyshev......
actual capacitor values.  Since                  Ripple......
accurate capacitors can be difficult          Bessel.........  *
to obtain, you can measure capacitors       Number of poles..  8
and enter their actual values.              Cutoff frequency.  1.000 k
Resistor values are recomputed based         Response @ fx...  2.000 k
on your entries.                             Scale Resistors.  10.00 k
                                             Gain Entry......
DO THIS STEP LAST.  DESIGN ALL OTHER      ▶  Capacitor menu..        <enter>
ASPECTS OF THE FILTER FIRST.                 Resistors.......  Exact
                                             Display.........

          Sallen-Key, 8-Pole Low-Pass Bessel,  1.000 kHz Cutoff

Section      R1          R2          R3          R4          C1          C2

A          2.378 k     15.31 k                            10.00 nF    22.00 nF
B          2.479 k     20.35 k                            6.800 nF    22.00 nF
C          1.990 k     14.87 k                            6.800 nF    33.00 nF
D          2.154 k     15.82 k                            3.300 nF    47.00 nF

    <F1> Quit      <F2> Plot      <F3> Print
```

Fig. 5. The Filter2 program provides a simple design method for both Sallen and Key and multiple-feedback low-pass filters. Response to the questions in the upper screen, instantly shows component values needed on the lower screen. Press <F2> to view the frequency domain plot of your design on-screen.

```
                         UAF42 FILTER DESIGN GUIDE
                                 Rev 1.1
Desired Response :                                   Notch Parameters      :
▪Notch                                               ▪Order: n= 8
                                                     ▪fnotch  = 1.000k Hz
 Filter Type     :  <Enter> for more options => ▶▪Bandwidth= 100.0 Hz
▪Butterworth                                         Filter Response       :
▪Bessel *                                            ▪Freq = 2.000k Hz
▪Chebyshev                                            Gain =-31.99m dB
  ripple                                              Phase= 8.00 °
▪Inv Chebyshev                                        Tot UAFs : 4
  Amin                                               ▪Resistors: Standard 1%
                                                     ▪Topology : Auto
                                  Tutorial
  Notch Response      [Press <Enter> to toggle between Bandwidth and fl, fh
                       inputs].
Av    \  ◀ bw ▶  /     -Bandwidth: Notch band frequency range.
                       -fl, fh: Low and high notch band frequency end points.
                   f    Amplitude characteristics at fl and fh are dependent on
  fl   fnotch   fh      filter type selected (ie: -3dB points for Butterworth).

<F1> Exit     <F2> Plot     <F3> Component Values     <F4> Printout     <F5> Help
```

Fig. 6. Dedicated for use with the 'UAF42' integrated circuit, simply respond to the upper prompts to design your filter. Press <F2> for an instant performance plot, <F3> for a parts listing.

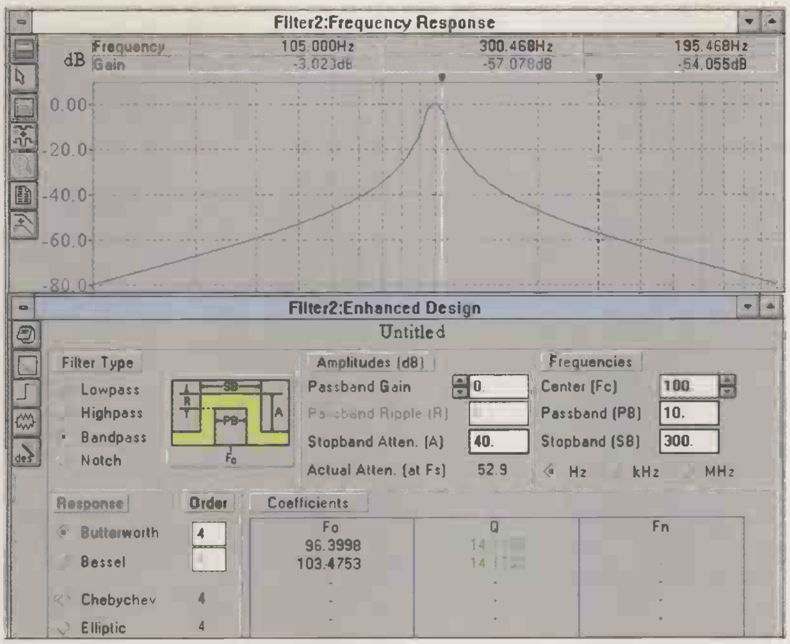


Fig. 7. Design of switched-capacitor filter and choice of integrated circuit needed for your desired performance is easy using FilterCAD. It provides both basic and enhanced level design screens. Simply enter your design goals into the design screen. The software then provides the frequency response shown.

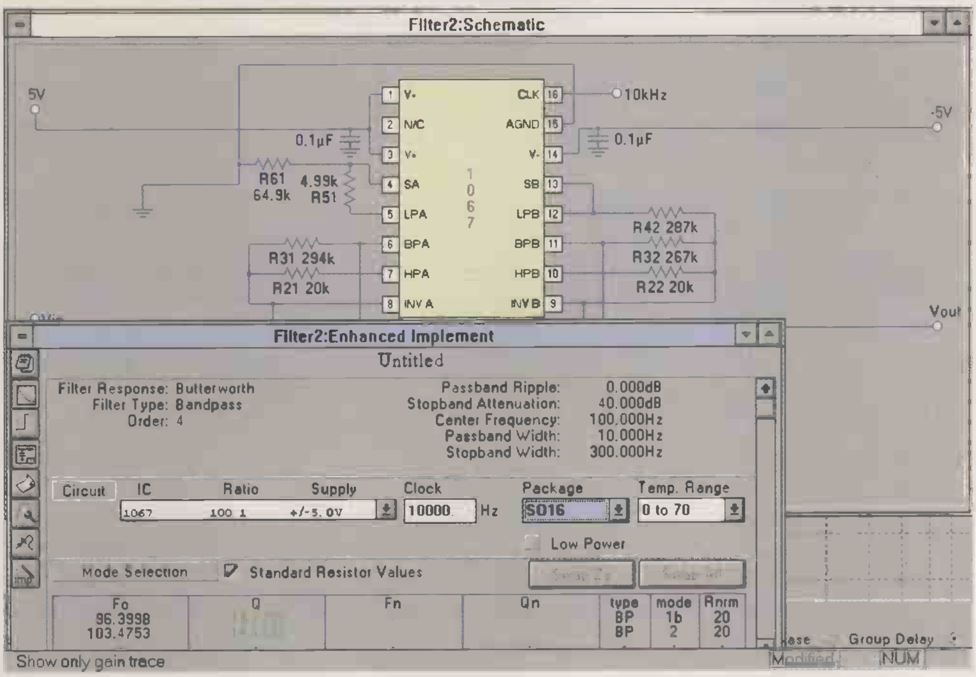


Fig. 8. Illustrating the final filter design stages. Having accepted the suggested integrated circuit or made your own choice, the software automatically calculates component values needed then draws your schematic circuit.

requires more op-amps than either the Sallen and Key or multiple-feedback filters designs, it provides high-pass, low-pass and band-pass outputs simultaneously from the one circuit. These outputs can be combined to produce any desired second-order transfer function.

Burr-Brown application note AB-035C describes the company's *Filter42* DOS based state-variable filter design program. This is part of the company's *FilterPro* package.[7]

Filter42 is aimed at designers using the *UAF42* integrated circuit – a state-variable filter on one chip. The device provides two precision 1nF capacitors together with four op-amps. Usually, it needs only external resistors to complete your design.

This software automates the design of the three most common filter styles – Butterworth, Bessel and Chebyshev – together with the less common inverse Chebyshev. Inverse Chebyshev provides a flat pass-band and step response similar to that of a Butterworth, but with a steep rate of roll-off, **Fig. 6**.

Filter42 also designs band-pass and band-stop filters with responses similar to parallel or series connected *LC* circuits respectively. As for the *Filter2* software, *Filter42* also facilitates 'what-if' evaluations, presenting simulated results on screen.

Switched-capacitor filters

Switched-capacitor filters tend to be easier to design, since each data sheet provides the necessary design equations. Switched-capacitor filters are capable of complex response curves, not easily produced using active continuous time filters or passive filters.

The disadvantage of switched-capacitor filters is reduced dynamic range and increased noise output. Their major advantage is change of filter frequency simply by changing clock frequency.

Designing a switched-capacitor fil-

Useful sites

1. Happy99.exe worm spreads through US and Europe	http://www.msnbc.com/news
2. Symantic Corporation	http://www.symantic.com/avcenter/vene/data/happy99.worm.html
3. Data Fellows' Virus Information	http://www.datafellows.com/v-descs/ska.htm
4. National Semiconductor	http://www.national.com
5. Understanding emi filters. Bateman. C.	Electronics World May 1996
6. Application Note 67 - AN67.PDF	http://www.linear-tech.com
7. Burr-Brown Corporation 'FilterPro' software	http://www.burr-brown.com
8. Linear Technology Corp.- 'FilterCAD 2.0'	http://www.linear-tech.com

ter first requires selection of the appropriate integrated filter circuit part. This choice of filter part number has now been automated for you. Linear Technology provides a CD rom entitled *FilterCAD 2.0* that runs under Windows 3.1 as well as Windows 95 or 98.[8] Installation is extremely easy: the installation program automatically detects which operating system is being used, then proceeds accordingly.

Filter design using this software is equally easy. It involves a two-stage process. Simply choose your required filter circuit response shape and low-pass, high-pass, band-pass or notch configuration, on the 'Design Screen'. Then input your target gain, ripple, stop-band attenuation, centre, pass-band and stop-band frequencies and select 'implement,' **Fig. 7**.

The software provides a selection of suitable integrated circuits, highlighting the most appropriate Linear Technology filter part. Having accepted this part or selected your preference, component values are automatically calculated and the schematic drawing of your circuit can be displayed.

Both frequency and time-domain simulation results can be quickly plotted on screen. All this is completely 'hands-off'. Simply request your desired action and the software completes the task, **Fig. 8**.

Which filter did I choose?

This article is the result of a search I carried for a filter to use as part of a product I am designing. So which of these design approaches solved my particular filter needs?

Apart from the state-variable filter software, I failed to find a single, simple, time-continuous band-pass filter design shortcut on Internet. In all probability such methods exist, but the search engines I used did not find them for me. My circuit board had insufficient space to accommodate a state-variable filter design.

Centre frequency and 'Q' being important, my initial design calculations had assumed using a dual op-amp giving two multiple-feedback, second-order band-pass filters in series. The prototype was designed using equal value capacitors, via the equations from 'Reference Data for Radio Engineers' and a pocket calculator. The prototype worked fine but the calculations were time consuming.

I needed to refine this design to accept a production spread of components. Rather than repeat the pocket calculator process, I decided to write a quick and dirty computer program. This prints out permutations of my options. I input the desired frequency and capacitor values, the program then instantly produces hard copy of all possible resistor combinations for the desired range of 'Q' values. ■

RF directional couplers

Passive directional couplers facilitate RF measurements and have uses in receivers. Joe Carr explains what they are and how they work.

Directional couplers are devices that will pass signal across one path, while passing a much smaller signal along another path. One of the most common uses of the directional coupler is to sample a radio-frequency power signal either for controlling transmitter output power level or for measurement.

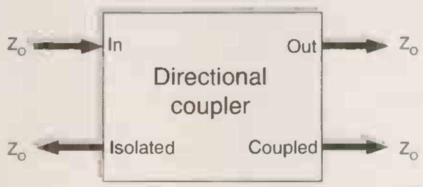


Fig. 1. Circuit symbol for a directional coupler.

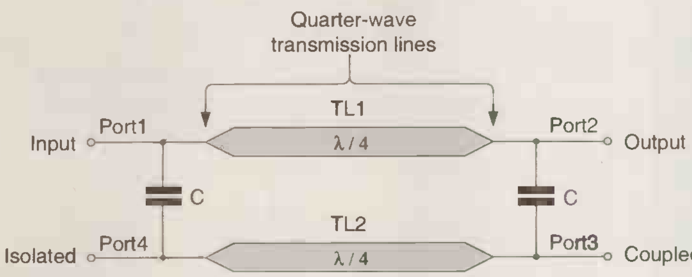


Fig. 2. Directional coupler made up of transmission line segments.

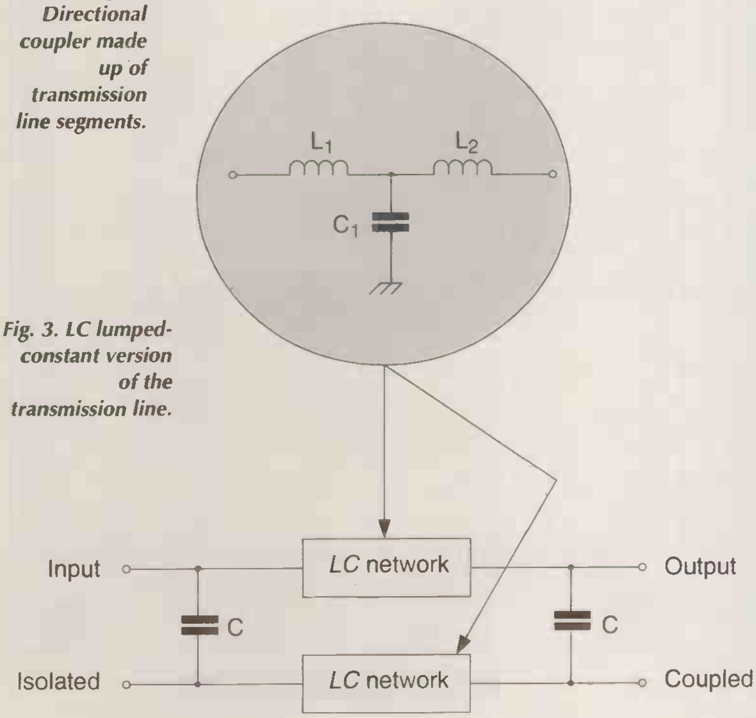


Fig. 3. LC lumped-constant version of the transmission line.

An example of how such a coupler can be used for measurement is a digital frequency counter connected to the low-level port, and the transmitter and antenna to the straight through, i.e. high power, ports.

The circuit symbol for a directional coupler is shown in **Fig. 1**. Note that there are three outputs and one input. The In-Out path is low-loss, and is the principal path between the signal source and the load. The coupled output is a sample of the forward path, while the isolated showed very low signal. If the In and Out are reversed, then the roles of the coupled and isolated ports also reverse.

An implementation of this circuit using transmission line segments is shown in **Fig. 2**. Each transmission line segment, TL1 and TL2, has characteristic impedance, Z_O, and is quarter wavelength long. The path from Port 1 to Port 2 is the low-loss signal direction. If power flows in this direction, then Port 3 is the coupled port and Port 4 is isolated. If the power flow direction reverses – Port 2 to Port 1 – then the respective roles of Port 3 and Port 4 reverse.

For a Port-3/Port-4 coupling ratio of –15dB or less, the value of coupling capacitance must be,

$$C_c < \frac{0.18}{\omega Z_o} \text{ farads}$$

The coupling ratio is $20\log(\omega C Z_o)$dB where ω is $2\pi f$. The bandwidth is about 12 percent.

The circuit shown in **Fig. 3** is an *LC* lumped-constant version of the transmission lines. This network can be used to replace TL_1 and TL_2 in Fig. 2. The values of the components are,

$$L_1 = L_2 = \frac{Z_o}{\omega_o}$$

$$C_1 = \frac{1}{\omega_o Z_o}$$

Figure 4 shows a directional coupler used in a lot of RF power meters and VSWR meters. The transmission lines are implemented as printed circuit board tracks. It consists of a main transmission line (TL_1) between Port 1 and Port 2 – the low-loss path – and a coupled line (TL_2) to form the coupled and isolated ports. The coupling capacitance in picofarads is approximated by $9.399X$, where X is in metres, when implemented on G-10 Epoxy glass-fibre printed circuit board.

Reflectometer directional coupler

A reflectometer directional coupler is shown in **Fig. 5a)**. This type of directional coupler is at the heart of many commercial VSWR meters and RF power meters used in the HF through low-VHF regions of the spectrum.

This circuit is conceptually similar to the previous transmission line, but is designed around a toroid transmission line transformer. It consists of a transformer in which the low-loss path is a single-turn primary winding, and a secondary wound of enamelled wire.

Details of the pick-up sensor are shown in Fig. **5b**). The secondary is wound around the rim of the toroid in the normal manner, occupying not more than 330° of circumference. A rubber or plastic grommet is fitted into the centre hole of the toroid core.

The single-turn primary is formed by a single conductor passed once through the hole in the centre of the grommet. It turns out the $^3/_{16}$in (4.76mm) outside-diameter brass tubing, of the kind sold in hobby shops that cater for model builders, will fit through several standard grommet sizes nicely, and will slip-fit over the centre conductor of SO-239 coaxial connectors.

Another transmission-line directional coupler is shown in **Fig. 6**. Two lengths of RG-58/U transmission line, each

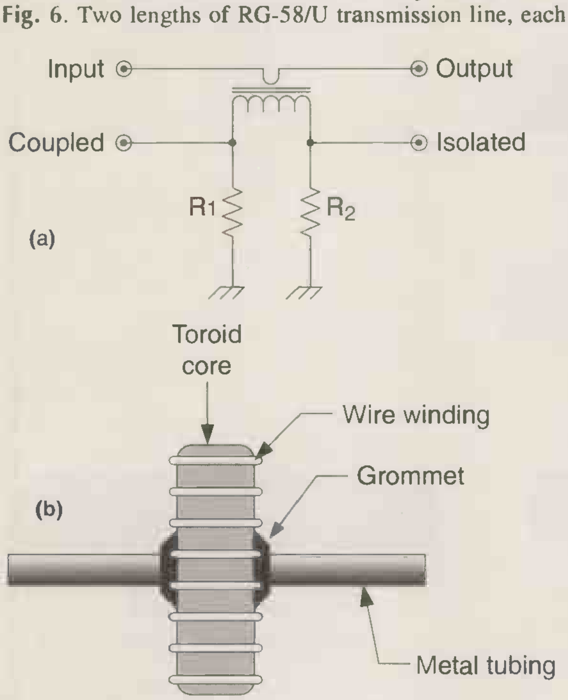


Fig. 5. Reflectometer directional coupler circuit, a), and details of the pick-up sensor, b).

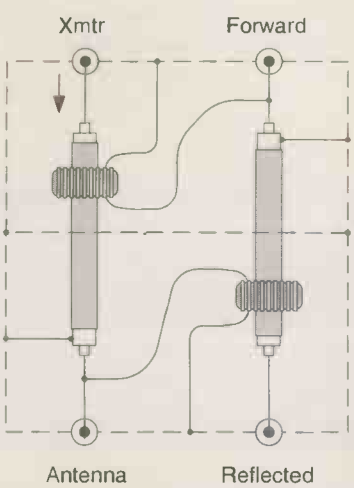


Fig. 6. Alternative directional coupler involving two short lengths of RG-58/U transmission-line.

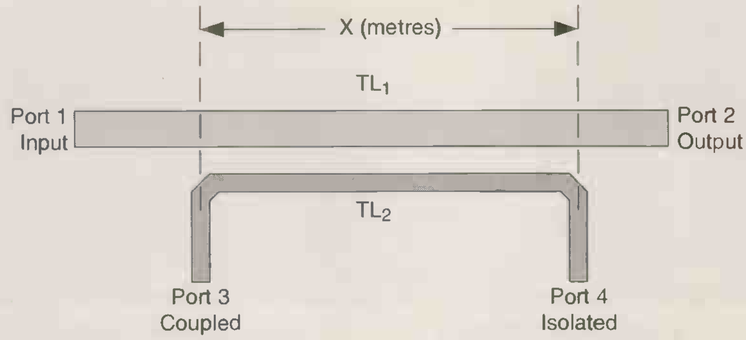


Fig. 4. Directional coupler of the type used in many RF power and VSWR meters.

around 6in long, are passed through a pair of toroid coils. Both coils are wound with 8 to 12 turns of wire. Note that the shields of the two transmission line segments are grounded only at one end.

Each combination of transmission line and toroid core forms a transformer similar to the previous case. These two transformers are cross-coupled to form the network shown.

The Xmtr-antenna path is the low loss path, while – with the signal flow direction shown – the other twom coupled ports are for forward and reflected power samples. These samples can be rectified and used to indicate the relative power levels flowing in the forward and reverse directions. Taken together these indications allow you to calculate VSWR.

Directional couplers are used for RF power sampling in measurement and transmitter control. They can also be used in receivers between the mixer or RF amplifier and the antenna input circuit. This arrangement can prevent the flow of local oscillator signal and mixer products back towards the antenna, where they could be radiated and cause electromagnetic interference to other devices. ■

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LETTERS

Letters to "Electronics World" Quadrant House, The Quadrant, Sutton, Surrey, SM2 5AS

E-MAIL jackie.lowe@rbi.co.uk

Engineering women

With reference to Richard Wilson's piece on page 480 of the June issue, it is good to see that these young women are providing positive role models in an industry dominated by men. Only 3% of the registered engineers – be they Chartered, Incorporated or Engineering Technicians – are women. But it should be pointed out that Richard's article is slightly misleading.

The competition for 'Young Woman Engineer of the Year' is open to Incorporated Engineer level, that is women who hold HNC or HND qualifications. As the competition is run by the Institute of Incorporated Engineers, it is no wonder that young women with degrees – be they from red-brick universities or not – do not feature. The so called 'stiff competition' is not as stiff as it could be!

This competition works on positive discrimination, which may give the message that women are not able to compete in a field dominated by men.

Surely it would be better to see women engineers competing and winning in competitions open to both sexes and to all levels? This would give a more powerful message.

John Phillips
via e-mail

A little thought: imagine a guy with an IQ of *n* who has played chess all his life. He's invited to play a game with a stranger with an equal IQ, but who only caught his first glimpse of a chess board a week ago. I'll probably be accused of being patronising here, but women will only be able to compete on equal terms when they've had equal opportunities for a while. Many thanks to John for raising the issue. Ed.

Heater debate

Bob Pearson's letter regarding the use of capacitors to supply heater current evokes memories of the ubiquitous 2μF paper capacitor – or 2mfd condenser, as we used to call it.

Small table radios from the US used a resistive line cord – wound on an asbestos core – to reduce the 110V line voltage: as the cord dissipated 12W, or more it was pleasantly warm to the touch. Cords allowing such radios to operate on 240V were hard to obtain and this led to the use of series paper condensers.

In rural districts in the 1930s, a new HT battery was a significant and often deferred expense. Wireless sets of that time had rather poor decoupling of the detector and output stages so that the high source impedance of a failing battery caused distortion and 'motor boating.' A 2mfd capacitor across the battery would restore normalcy – at least until the next payday.

Capacitors were also used to establish authority. Early in my service career, a voice called, 'catch!' Instead of the expected NAAFI bun, I caught a fully charged 2μF capacitor.

Harrold Shipton
St Louis

Inaccuracy logged

The news item regarding using logs to speed up calculations, May 1999 page 358, claims that the use of fixed-point numbers for logarithms avoids the rounding errors inherent in floating point calculations. This claim ignores the fact that precise logarithmic values typically require many more significant digits than the number they represent.

Printed log tables and computer calculations inherently conceal rounding errors. The explanation for the improved accuracy is that, for any total number of bits, fixed point calculations always provide a much greater number of significant bits.

R G Silson
Tring
Hertfordshire

Bandhogs

I've just read Andrew Emmerson's article 'Upwardly mobile' in the March issue, page 208, and I find myself aghast at the latest demands of the telecoms industry.

Analogue mobile phones currently occupy about 80MHz in the 900MHz region; the digital service has a chunk around 1900MHz. Now a further 155MHz is being rejected as 'woefully inadequate' for UMTS.

At the same time, our perfectly good analogue television services are being forced to squeeze themselves to compressed digital channels which may or may not provide better pictures. In my 25 years in the TV trade, I've never heard anyone ask for better pictures, wider screens or 'home banking'. Perhaps the telecom boys went to the same school as the Windows Software Writers' Guild, and simply expect their paths to be cleared in readiness for their bloated products. After all, it's only bandwidth isn't it?

Andrew Howlett
Dukinfield
Cheshire

The right to copy

I have just read Peter Marlow's editorial in the February issue. He elegantly presented the arguments heralded by both parties in this debate. Nonetheless it remains uncertain that this concern can be rapidly resolved.

I accept that the efforts of individuals need to be adequately recognised – monetarily or otherwise – in order to sustain any kind of competitive market. But it remains to be seen whether the public would indeed be willing to purchase software 'merchandise' under present conditions. It is difficult to justify spending a great deal of money in order to evaluate the suitability of any software in fulfiling personal requirements when the product explicitly excludes any liability as to accuracy, suitability, and other consequential damages.

It would also appear that under the pretence of 'essential' updates, one has to keep up with the evolution of software and *de facto* provide the suppliers with an unending stream of revenue – regardless of the fact that the previous version of the software 'suited us just fine'.

Many early PC applications were written with very compact code. They provided many of us – perhaps most of us – with the essential functions required at the time, and which are still fundamental today.

Here in Canada, legislation has been tabled to impose a levy on recording media to palliate plagiarism. In principle, one could consider this a reasonable measure. But don't forget those of us that use recordable media as a means of protecting our own intellectual property.

Unfortunately, I cannot propose an alternative at this time, but I am certain that this issue is definitively not going to go away.

Daniel Gratton
Nortel Networks
Montreal

Where did the weekend go?

31 December 1999 will be a Friday.
1 January 1999 will be a Saturday.
1 January 1900 was a Monday.
Will computers that have not been modified revert to the 1900 calendar and hence jump from Friday to Monday, losing two days?

R N Soar
Doncaster

Microstripline

The article in the May edition on microstripline, page 392, has a boxed note on drawing tracks to 0.001in. The following warning may be of interest.

The technique of drawing a thin track by juxtapositioning several hair-line' tracks should be used with caution. Although there may be no problem in Serif – of which I have no experience – with many graphics and dtp design packages a 'hair-line' has an indeterminate width, only calculated by the print driver when the output is produced.

Although a set of hair-lines may appear to abut each other on the screen, the printed output may show finer lines, with gaps between them.

An even worse potential problem is that although the software may detect your current printer's capabilities, and show the width correctly, it will be not be the same when an alternate printer is used.

In particular, if you eventually output to film using a bureau's image setter, working at 2400 dpi for example, the hair-lines will be almost invisible! This problem has caused many newbie graphics designers a lot of wasted expense.

John Walton
via e-mail

In the April issue of letters, Jim Cahner asked, "Why isn't this a perfect filter?" When the high and low-pass voltages of the filter above are summed, the result is a perfect straight line. Here's a few more of the answers to this very popular question, leading off below with a description of a filter that is almost as simple, and certainly nearer "perfect".

In the Letters section of the April edition of *Electronics World*, Jim Cahner asks for a simple explanation of why his third-order, two-way active filter is not perfect as a loudspeaker crossover network.

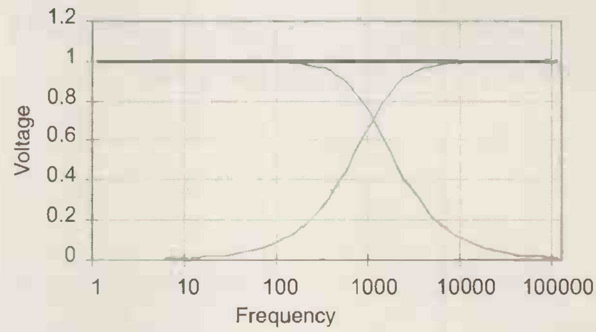
The simple explanation is that the total output from two speakers is not the sum of their individual pressures!

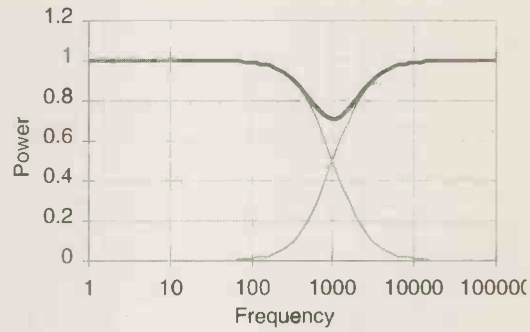
The mistake Jim – and many other crossover designers – makes is to assume that the the combined output from two loudspeakers is related to the sum of the voltage responses of the filters. It is true though that the pressure response of one speaker has the same shape as the voltage response of such a filter.

Consider this: let's say Jim's filter modules produce 1V in their respective pass-bands, and that the combined output is a flat 1V across the spectrum. Then at the crossover frequency, each must produce 0.5V output for his depicted flat response.

The power output at this frequency from each loudspeaker connected to these networks will therefore be 0.25 times the pass-band level ($W=V^2$I), giving a total power output of 0.5 times the pass-band level (0.25+0.25). But the power response will have a dip of 3dB at the crossover frequency.

To produce a flat combined response requires filters that are 3dB down at the crossover frequency (1/2 power), or 0.707 times the pass-band voltage, such as the Butterworth.





Adding Jim's simple crossover's voltage curves gives a straight line. It's only when you add the power curves that you see the problem.

Ian Taylor
Department of Psychology
University of Melbourne
Australia

There are a few questionable things with the plots in Jim Cahner's letter.

Judging from the two summing resistors, the amplitude plot looks like voltage. For example, take a simple first-order filter based on R=10kΩ and C=15.9nF giving a 1kHz cross-over. The first plot shows the voltages of both high and low-pass sections, taking into account the phase shifts.

The top line is the algebraic sum which is a flat line. The phasing

Too simple to be useful?

The third-order two-way crossover circuit presented by Jim Cahner in the April 1999 issue is not perfect because magnitude and phase of the vector sum of the two outputs are not constant.

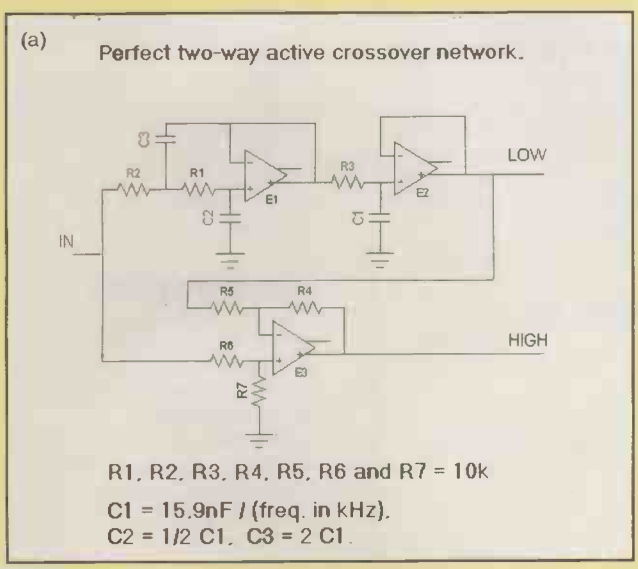
This requirement can easily be met by replacing one of the two sections by an elementary difference amplifier giving,

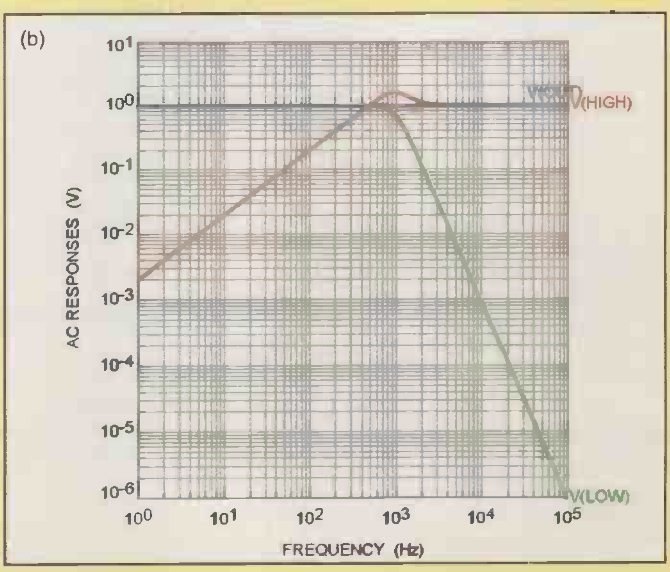
$V_{\text{high-pass}}=V_{\text{in}}-V_{\text{low-pass}}$

as in **Fig. 1a)**.

The combined frequency response, Fig **1b)**, and corresponding phase behaviour, Fig. **1c)**, are flat permitting a perfect transient response Fig. **1d)**.

Following the same principle, every crossover filter can be made perfect by having a 'correcting'





difference between these two signals is a constant 90°, whereby the high pass leads the low pass.

But what you hear is power. So low pass and high pass, each through its identical power amp, develops its own power $p=e^2/r$, whereby p=power, e=output voltage of power amp and r= speaker resistance.

The phasing of these is still high pass leading by 90°.

Imagine loudspeakers without dimensions such that the voice coils can be in the same location. No matter where your ears are, the sum of the powers is as shown in the second plot on the left.

Notice the dip to 70% of nominal power at the crossover frequency. If this is what a simple first order filter does, I'd hate to imagine what a third order will do, but I have not finished the mathematics behind it yet.

Michel Debuysscher
via e-mail

I read with some interest "Too simple to be useful?" in the April 1999 issue. His design is a well-known third order crossover, which can also be implemented with passive components.

The trouble with all odd-order filters is that the signals delivered from each speaker are in phase quadrature. But this results in constant level as Jim shows. There is a point slightly off axis that has a greater magnitude than on axis, because the signals are now in phase.

These problems are a direct result of the fact that the high and low frequency speakers are not at the same point. A truly coaxial speaker would not suffer from this problem but has problems of its own.

Even-order filters – second, fourth etc. – produce out-of-phase signals from the speakers, which results in a null at the crossover frequency. Any attempt to correct this problem by inverting the phase of one speaker results in a bump in the response at the crossover frequency.

There is one very clever second–order filter design that uses an additional speaker with a band-pass response at the crossover frequency. Here the result is a speaker system with constant level and linear phase.

A brilliant design that allows an even-order filter to produce a uniform response is the Linkwitz-Riley. It crosses over at –6B as opposed to the usual –3dB.

Finally, in a book entitled 'Advanced Speaker Systems', available from Radio Shack and written by Ray Alden, a reference is made to a design by D'Appolito where two vertical woofers with a centre tweeter are shown to produce excellent patterns with – of all things – third-order crossovers.

Jack Kouzoujian
Mill Neck
USA

Missing charge

Another view on the so-called 'missing-charge' paradox raised in Brian Cox's letter, January issue, can be obtained by postulating a variable capacitor rather than two capacitors connected by a switch, as this eliminates the need for the switch and interconnections.

As pointed out by Dr Shaw and Chris Ward in their replies, March issue, the idealised model used by Brian Cox results in the creation of an infinite current impulse at the instant of closing the switch, a situation which is avoided when a variable capacitor is used.

To illustrate, suppose we have a variable, parallel plate capacitor, the spacing between plates being x. By doubling the spacing to $2x$ the capacitance is halved and, following the standard argument for the paradox it will be seen that the energy is doubled in this case, i.e., energy is mysteriously created rather than destroyed – it's always better to have a positive approach.

What is missing here is the complete energy picture. Let A be the plate area, and for an air dielectric the permittivity is equal to that of free space, e_o and the initial capacitance is .

The initial stored energy is

$$E_1=Q^2/2C=Q^2x/(2e_oA)$$

where Q is the stored charge.

With Q constant, the force of attraction between the plates is

$$F=dE/dx=Q^2/(2e_oA),$$

also a constant. Doubling the distance i.e., increasing it by amount x requires work to be done equal to Fx which is equal to $Q^2x/(2e_oA)$.

This is the energy added to the system. Since capacitance is halved the new value of stored energy is $E_2=Q^2/(2(0.5C))=Q^2/C=2E_1$. The stored energy is doubled but the extra comes from the work done in separating the plates. Using the alternative formula for energy $E=0.5CV^2$ it is easily verified that voltage across the capacitor doubles when the separation is doubled. Incidentally, in Brian's original letter he seems to assume that voltage is always an *independent* variable which clearly is not the case.

The exercise may be repeated by halving the distance, showing that the stored energy in the capacitor is halved. The argument is just a little bit more complicated as it forces one to think about what holds the plates apart in the first place, but the "lost" energy is returned to the mechanical system holding the plates apart.

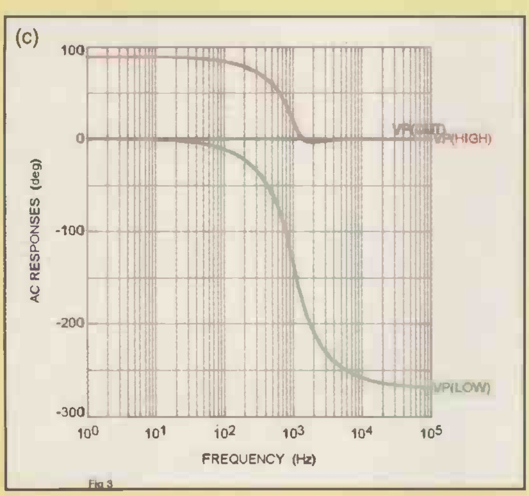
Dennis Roddy
Thunder Bay
Ontario

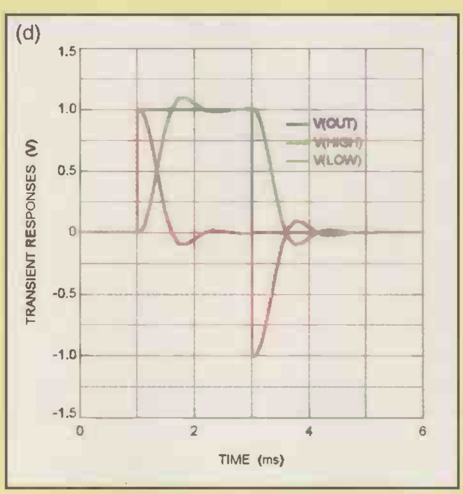
channel fed with the difference between the vector sum of the outputs of the other sections and the input. Of course, the loudspeakers must be able to faithfully reproduce the acoustical equivalent of the vector sum of all outputs of the filter.

Considering the results and the simplicity of this principle, I am somewhat surprised that I don't remember ever having seen it proposed or used before.

Jean-Pierre van Dormael
Wezembeek-Oppem
Belgium

Figs 1a-d). First is the enhanced crossover network, followed by its frequency response, phase behaviour and transient response.





Clearer curve tracer

With regard to Ian Hegglun's "Hotter Spice" article in the May 1999 issue of Electronics World, on page 395 is shown a curve-tracer set-up. The oscilloscope shown in the diagram has four terminal designations. One of those terminals is designated "common" and the others are designated 1, 2 and 3, respectively; 1 appears to be the vertical input and I assume 2 is the horizontal. I'm unsure as to what 3 is. Of course, it's possible I'm wrong about all of them, since nowhere in the article does it specify what inputs the terminals represent.

Bob Schoonmaker
Woodside
New York

Ian replies:
Sorry for any confusion. Since the curve tracer was not central to the modelling issue, I tried to keep the description as short as possible. Maybe it was too short.

The scope runs in 'Time Base' mode. All three channels are y-inputs. I used an HP54601A four-channel DSO.

The third channel is optional. If

Ch 3 is used, it allows the x-axis to be converted to volts/div later after a printout.

If you are using a two-channel scope then calibrate the time base to give meaningful graduations, eg 1, 2, 5V/div as follows:

Step 1. Apply the Ch 3 signal temporarily to Ch 1 and arrange the slope of the ramp to be 1:1, 1:2 or 2:1 div/div by varying either the timebase vernier or signal-generator frequency.

Step 2. Assuming Ch 1 has been zero referenced, use the x-offset to get the 0V-in over an x graduation of your choice. Note this reference point. Record the xV/div value used. Remove Ch3 and restore Ch1.

You could also use a single channel scope by alternating Ch3 then Ch 1 and Ch2.

Class-A comments

I read Colin Wonfor's article on Class-A with interest.

Having spent some time myself trying to make Class-A amplifiers both tractable and as linear as they should be, I was surprised he did not adopt any of the technology introduced in the Class-A (March 1994) and Trimodal (June, July 95) designs. No doubt there were good reasons, but it would be nice to know what they were. Likewise it is a great pity that no performance details were given, to confirm that the great cost and heat are really worthwhile.

My investigations showed that the use of power FETs in the Class-A output stages makes linearity worse. The Class-B problem of a horribly jagged crossover region is no longer relevant, but the low device transconductance still causes high distortion.

An automatic bias controller would have cost a few pence, and removed the need for a perilous quiescent-adjust preset that allows infinite current to flow at one end of its travel.

The instruction to trim the DC offset to less than 10mV with a preset seems rather strange when it is straightforward to make an amplifier that gives an offset of less than 25mV with no adjustments.

It concerns me that there is no effective DC-offset protection given the large amount of power available. I note the output fuse (no value shown) but I would be interested to know how the constructor is going to select its value so it can protect

Now Zen

Regarding Ian Hickman's article on the distortion free amplifier of March 1999, page 224, I think Nelson Pass has already designed this amplifier and the preamp to drive it. It is called Son of Zen with a balanced line stage.

As Nelson uses resistors to bias it, it is very efficient. Could Colin Wonfor's output stage be used in place of it, or my adaptation of Nelson Pass's Zen 10W amplifier? Maybe Colin or Ian would be interested in making it work?

All of Pass's designs can be found on the internet at passlabs.com. Cyril Bateman covered his designs in a recent 'Hands-on Internet' article.

Richard Woollcott
Canada

Ian replies:

A number of readers have contacted me, all like Mr Wollcott noting the advantages of a balanced bridge output stage.

Certainly the idea has been around a long time, the purpose of my article being simply to point out the advantages to the obscurantists who still advocate a single ended output stage with a big beefy triode.

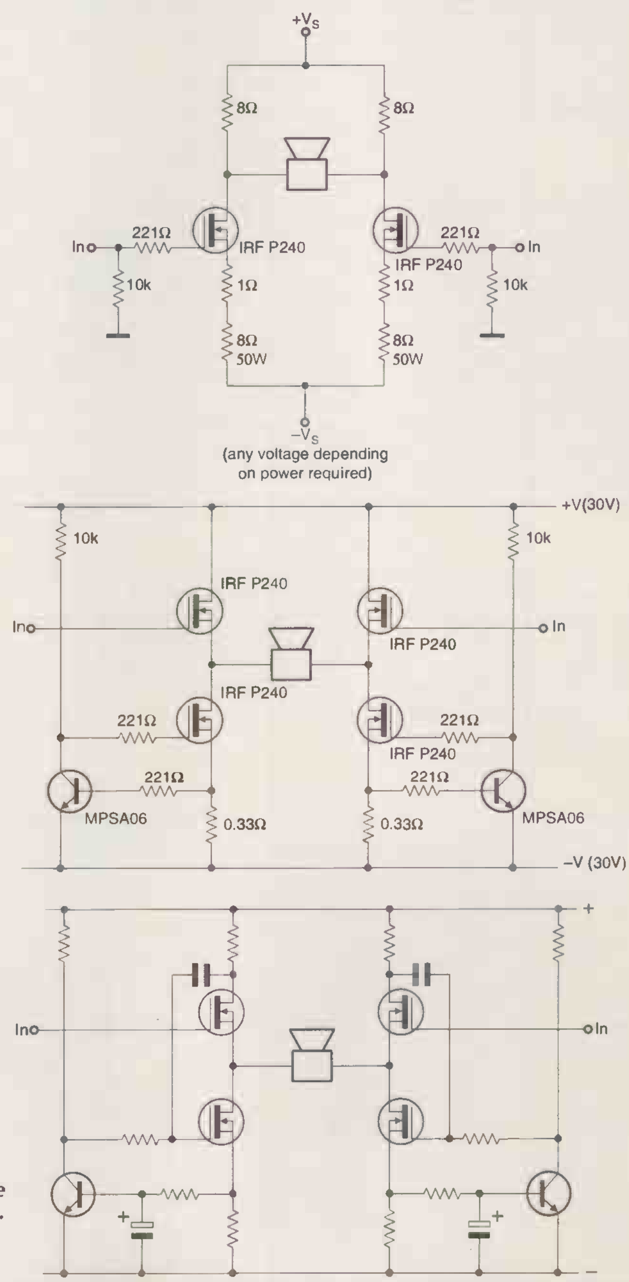
The two arrangements he shows, are, as he notes, rather inefficient. A useful improvement in this respect could be obtained by using the semiconductor equivalent of the White cathode follower.

This arrangement was popular in the days of valves. It offers increased drive and lower output impedance compared with a simple cathode follower with a resistor as the 'long tail'.

The stratagem was used, as thermionic complementary cathode follower:valves came in the 'n-p-n' flavour only.

Applied to Mr Woollcott's adaptation of Zen, it consists of modulating the drain current of the lower FET, in the direction of cutting off when the input goes positive, and *vice versa*. Thus on a positive output swing, the tail current reduces, the extra current being available for the load. On negative swings, when the source follower cuts off, the tail current increases.

The necessary gate drive voltage can be obtained from a small resistor in the drain lead of the source follower, AC coupled down to the erstwhile constant current tail FET.



'Son of Zen', top, with Richard's adaptation below it. At the bottom, is Ian Hickman's idea for increasing efficiency.

loudspeakers without nuisance-blowing or introducing distortion.

A 1A slow-blow output fuse carrying 20W/8Ω generates 0.01% THD (third harmonic) at low frequencies, so fuse distortion could easily make the finer nuances of Class A linearity somewhat irrelevant.

It seems astonishing to use the constant-current (single-ended) mode for a large Class-A power amplifier as this doubles the already enormous heat dissipation. The largest version described releases 3kW of heat, which will not be significantly reduced by playing music at full volume. Surely this will be an uneasy companion in summertime?

The quiescent current for the so-called 300W/4R version is prescribed as 10.2A, so negative peaks cannot go beyond 4 by 10.2 or 40.8V below ground. This corresponds to a maximum sine output of only 208W, so using supply rails as high as ±75V appears to do nothing but increase the power dissipation.

It allows no safety margin for loudspeakers that fall below a nominal 4Ω impedance. Perhaps there is a typographical error in the article for this huge emission of heat makes little sense to me.

At the risk of seeming discouraging, no information is given that makes me see this design as a significant advance in Class-A amplification

Douglas Self
London

Voice coil reaction

This comment refers to Speakers' Corner in the January 1999 issue. While I usually enjoy descriptive articles about odd facets of technology, I cannot let a piece that is so blatantly misleading go uncommented.

The discussion about the reaction of the voice-coil current on the magnetic field in the speaker magnet is way out. For a start, provided the voice coil physical excursions stay inside the linear region of the magnet field, there is no non linearity caused by the field reactions. Certainly the field is distorted. This is what provides the driving force, but this will never be non linear. If the field was not distorted, then the voice coil would see no force, and not move – not quite what you want in a speaker.

Even further out is the next discussion. If the reaction is indeed moving domains in the permanent magnet, then in the process the magnet will become progressively demagnetised. This is not a good state of affairs, and one which no good speaker designer would never allow.

The fields produced by the voice coil current are so low compared to the magnet coercive force that this is never a possibility. So the question of Barkhausen jumps inducing noise, as the domains shift is just silly.

Worse still is the disparaging discussion of ferrite magnets. It is obvious that the writer has never looked at the construction of a ferrite magnet speaker. If he had, he might have noticed that the pole pieces near the airgap are – nominally – steel, while the ferrite is far away, having its flux concentrated by the steel inner and outer poles.

Since these poles are electrically conductive, this would allow the eddy current effects to work as he describes, for what it is worth, with no effect on the ferrite magnet. I make no comment on the gobbledygook in the next-to-last paragraph about 'practical' tests on a tweeter using software?

Adrian Jansen BSc
Toowoomba
Queensland
Australia

John replies:

I must apologise to Mr Jansen for writing gobbledygook. Loudspeaker designers use the term software to mean the moving parts of a speaker, the surround, spider and so on and perhaps I didn't make that clear.

If Mr Jansen will bear that in mind, what I wrote is now clear. I was sceptical about the audibility of magnets myself, and so I built a drive unit with interchangeable magnets to find out. One was a traditional ferrite structure, the other using a neodymium magnet.

The signal source, amplifier, cone, coil, surround and spider remained the same. All that was changed was the magnet. When the neodymium magnet was fitted, the drive unit sounded cleaner.

Flux modulation is a matter of fact and I am by no means the first to describe it. Domain jumping occurs in all magnetic structures subject to variable conditions including electric guitar pickups, telephones, antilock brake sensors and cassette tape heads.

Flux modulation does not demagnetise a speaker magnet. If Mr. Jansen were correct, guitar pickups would demagnetise in use and tape heads would never need demagnetising – contrary to our experience.

The resistivity of steel is too great to resist flux modulation, which is why a great many manufacturers use copper plated pole pieces as I stated. A speaker magnet is a series circuit and all of the flux in the gap passes through the magnet.

If the gap flux is modulated, the magnet and pole flux is modulated. If this sounds silly, then wind a coil around the magnet and look at the flux modulation on an oscilloscope.

It was unkind of me to suggest that ferrite is only useful for fridge magnets. I also use ferrite magnets for picking up swarf when I'm machining prototype pole pieces for neodymium magnets.

Impeding measurement

Ian Hickman's article on inductors in the May issue raises some interesting points. However the section where he has performed a number of calculations based on his measured results in Table 1, raises more than a few doubts.

I note he used an *HP8753* Network Analyser, which used correctly and within its capabilities, is an excellent instrument. However when used to measure very low and reactive impedances, representing a high VSWR, its accuracy is much degraded. Ian's few nanohenries of inductance, certainly represent a very low impedance and high VSWR.

Regardless of frequency, a good short circuit must always present an inductive reactance. Any measurement that declares a short circuit to be capacitive is a bad measurement.

Secondly, while not detailed in his article, I suspect the 'measurement-plane calibration' to which he referred, was use of 'port extension' to compensate for the electrical lengths of his adaptor and SMA connector.

While valid when measuring small VSWR parts, port extension provides a mathematical correction only for phase. It assumes use of precision 50Ω components, so does not correct for resistive losses. Using less than perfect adaptors, other correction methods are needed – particularly when measuring very low or very high impedance, reactive components.

Hewlett Packard suggests alternative and much better measurement methods in its 'Impedance Measurement Handbook'.

Over the years I have measured many similarly difficult components using 8753, 8510 and 4195 analysers, having to declare the uncertainty of my measurements.

Assuming that no more appropriate measuring instrument was available, I personally would have used only the best precision adaptors. Having performed a 'full 1-port' calibration at the APC7 test port, I would not have used port extension. Rather I would have measured the best possible short circuit, open circuit and a known 'load'. Each of these would have been mounted precisely in place of the capacitors or inductors to be measured. These results would have been used to 'characterise' the behaviour of the test connector and adaptors.

Hewlett Packard provides two application notes, 302-1 for 'open/short' and 346-3 for 'open/short/load' correction methods. These could be applied to mathematically de-embed his test connector and adaptor errors, giving better accuracy.

Unfortunately Ian has only performed a short measurement. Consequently only vector diagram correction is possible. This will also result in errors, but probably less than accepting his measured values.

The results are interesting, adjusting his nanohenry values for Table 1 in sequence as:

88.8, 98.7, 360, 7.6, 100.9, 233 nH.

Obviously his 'capacitive' short circuit is not negligible compared to the inductances measured.

These correction methods and calculations, applied to capacitors, were covered in my articles of December 1977 and April 1988. These techniques also apply to inductor measurements.

Cyril Bateman
via e-mail

Ian replies:

Cyril Bateman rightly points out the limitations of network analysers when used to make measurements on components exhibiting a high VSWR.

As is obvious from the quoted results in the article, the VSWR presented by the inductors measured was not merely high, but very high. The inductive reactances could have been made nearer j50Ω by using a higher measurement frequency, but that only exacerbates other problems.

Naturally, a full one port calibration was carried out, at the shortened connector spill and solder tag. But the open and short conditions were obviously not as ideal as coaxial standards.

The 50Ω load 'standard' was two small 100Ω surface-mount resistors soldered in parallel – not the 50Ω disk resistor found in a quality coaxial standard. So the calibration was inevitably not ideal, with even some uncertainty as to exactly where the reference plane was.

The effect of this could have been reduced by using a lower measurement frequency – for this measurement there is no no happy optimum.

At the risk of introducing a little theory into what was an interesting article on practical experimentation, I was surprised that Ian Hickman (How Long is L?, May 1999, pp. 386-9) did not mention a couple of formulas for the inductance of a wire, which have a bearing on the results of his experiments.

Firstly, the self-inductance (or 'internal' inductance of a long straight wire is exactly 50nH/m. This arises from the definition of various SI units. However, as Ian Hickman pointed out, we cannot measure the inductance of an isolated short wire, because there needs to be a return current flow.

Well, it turns out that this current flow will interfere with the field enough to substantially change the inductance, which is why Hickman deduced a somewhat larger figure himself.

Secondly, it is possible to make use of an expression for the inductance of a wire loop that demonstrates Hickman's practical results quite nicely, and shows that the inductance is indeed neither proportional to length of wire, nor length of wire squared. The derivation of the formula is difficult, and involves considering the mutual inductance between the current flow on the inside and outside diameters of the wire loop.

The exact expression is complicated, but it approximates to

$$L=\mu_0 r(\ln(8r/w)-2)$$

where r is the radius of the loop and w is the thickness of the wire, which must be much less than r for the approximation to be valid. To this must be added the 'internal' inductance of 50nH/m as mentioned above.

Putting in Hickman's figures of w=0.375mm (28 swg) and r=32mm therefore gives L=192nH, which is quite close to his figure of 201nH.

For r=16mm, the formula gives 82nH as against Hickman's 86nH. If I read his article correctly, he implied that there was some unexplained behaviour going on. I think that the above formula demonstrates that his results are not without explanation after all.

Just as the inductance of a multi-turn coil is proportional to N^2 only if all the flux couples all of the turns; so, in similar vein, the inductance of a single turn depends on how much of the flux from one infinitesimal current 'tube' couples the other current tubes. As I have pointed out above, this introduces a logarithmic term and the simple concept of 'inductance per unit length' does not always apply.

On the practical side, if the wire is very thin, and the loop radius does not vary too much, then the log term varies very little from example to example, so the inductance is, for practical purposes, proportional to length – for a single turn loop. However, the constant of proportionality does depend broadly on r and w, which was a point that did not come over in Hickman's article.

David Gibson
Microsystem Solutions
Leeds

Ian replies:

I am grateful to Mr Gibson for adding some further useful information for readers. With regard to his comment about the dependence of the inductance on the ratio of r and w, the diameter of the coil and the thickness of the wire, that was precisely the point of my Table 2. ■

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SD 6054B Micro Counter 20HZ-18GHZ – N Socket – £700.
SD 6054D Micro Counter 800MC/S-18GHz – £600.
SD 6246A Micro Counter 20Hz-26GHz – £1.2K.
SD 6244A Micro Counter 20Hz-4.5GHz – £400.
HP5352B Micro Counter OPT 010-005-46GHz – new in box – £5k.
HP5340A Micro Counter 10HZ-18GHz – Nixey – £500.
HP5342A Micro Counter 10HZ-18-24GHz – £800-£1K – OPTS 001-002-003-005-011 available.
HP5342A + 5344S Source Synchronizer – £1.5K.
HP5345A 500MC/S 11 Digit LED Readout – £400.
HP5345A + 5354A Plugin – 4GHz – £700.
HP5345A + 5355A Plugin with 5356A 18GHz Head – £1K.
HP5385A 1GHz 5386A-5386A 3GHz Counter – £1K-£2K.
Racal/Dana Counter 1991-160MC/S – £200.
Racal/Dana Counter 1992-1.3GHz – £600.
Racal/Dana Counter 9921-3GHz – £350.

SIGNAL GENERATORS
HP8640A – AM-FM 0.5-512-1024MC/S – £200-£400.
HP8640B – Phase locked – AM-FM-0.5-512-1024MC/S – £500-£1.2K. Opts 1-2-3 available.
HP8654A – B AM-FM 10MC/S-520MC/S – £300.
HP8656A SYN AM-FM 0.1-990MC/S – £900.
HP8656B SYN AM-FM 0.1-990MC/S – £1.5K.
HP8657A SYN AM-FM 0.1-1040MC/S – £2K.
HP8657B SYN AM-FM 0.1-2060MC/S – £3K.
HP8660C SYN AM-FM-PM-0.01-1300MC/S-2600MC/S – £2K.
HP8660D SYN AM-FM-PM-0.01-1300MC/S-2600MC/S – £3K.
HP8673D SYN AM-FM-PM-0.01-26.5 GHz – £12K.
HP3312A Function Generator AM-FM 13MC/S-Dual – £300.
HP3314A Function Generator AM-FM-VCO-20MC/S – £600.
HP3325A SYN Function Generator 21MC/S – £800.
HP3325B SYN Function Generator 21MC/S – £2K.
HP8673-B SYN AM-FM-PH 2-26.5 GHz – £6.5K.
HP3326A SYN 2CH Function Generator 13MC/S-IEEE – £1.4K.
HP3336A-B-C SYN Func/Level Gen 21MC/S – £400-£300-£500.
Racal/Dana 9081 SYN S/G AM-FM-PH-5-520MC/S – £300.
Racal/Dana 9082 SYN S/G AM-FM-PH-1.5-520MC/S – £400.
Racal/Dana 9084 SYN S/G AM-FM-PH-.001-104MC/S – £300.
Racal/Dana 9087 SYN S/G AM-FM-PH-.001-1300MC/S – £1K.
Marconi TF2008 AM-FM-Sweep 10KC/S-510MC/S – £200 Fully Tested to £300, as new + book + probe kit in wooden box.
Marconi TF2015 AM-FM-10-520MC/S – £100.
Marconi TF2016A AM-FM 10KC/S-120MC/S – £100.
Marconi TF2171/3 Digital Synchronizer for 2015/2016A – £50.
Marconi TF2018A AM-FM SYN 80KC/S-520MC/S – £500.
Marconi TF2019A AM-FM SYN 80KC/S-1040MC/S – £650-£1K.
Marconi TF2022E AM-FM SYN 10KC/S-1.01GHz – £1K-£1.2K.
R & S SMPD AM-FM-PH 5KHz-2720MC/S – £3K.
Anritsu MG3601A SYN AM-FM 0.1-1040MC/S – £1.2K.

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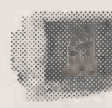

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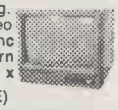

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Marconi 2030 opt 03 10KHz-1.3 GHz signal generator,New	£4995
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NEW PRODUCTS

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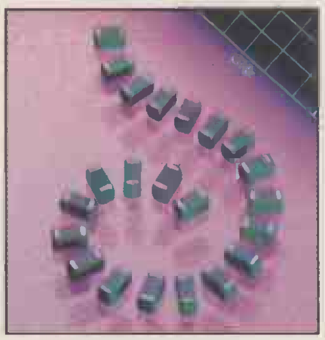
Jack and plug

Minitran is offering copper cabling with Giga-TX technology developed by Panduit. Giga-TX preserves the twist in paired cables within the jack and plug. The Giga-Channel system is claimed to exceed the proposed category six connector and channel specification, and supports transmission schemes with gigabit data rates, such as Gigabit Ethernet. Components include the Mini-Jack modular jack, patch cords, modular plugs and modular patch panels.
Minitran
Tel: 01279 757775
Enquiry No 502

Capacitors

DRC dual or triple resonance capacitors from AVX are suitable for RF bypassing in dual and triple mode and band cellular phones. The dual capacitor has series resonant frequencies at 880 and 1800MHz with tolerances of 3 per cent and a capability of attenuating main transmit and receive bands by more than 20dB. Rated at 50V with a frequency range of 500MHz to 3GHz, it comes in 0603 and 0805 chip sizes. The triple resonant version is for circuits requiring attenuation of three main bands.
AVX
Tel: 01252 770000
Enquiry No 503



Op-amp

National Semiconductor has launched the LMV921 op amp with rail-to-rail input and output. It comes in the SC-70 package and is suitable for portable consumer applications such as cellular phones and pagers. With operation from 1.8 to 5V, it has a quiescent current of 150μA. The device handles input voltage signals with amplitudes exceeding the supply voltages without distorting the output signal. Typically, the common mode voltage range can operate 300mV beyond the supply voltage. It is made on a sub-micron silicon-gate BiCMOS process.
National Semiconductor
Tel: 00 49 8141 351443
Enquiry No 504

Active GPS antenna uses 20mA at 5V DC

Matsushita has introduced an active GPS antenna. The VIC1LP has a 46 by 43mm footprint and is 12.5mm high. For automotive and related GPS applications, it is suitable for use under foliage, in a high rise urban environment, or with electromagnetic interference. Current consumption is typically 20mA at 5V DC. The standard, magnetic mounting, charcoal grey with shielded low noise amplifier, comes with 6m of coaxial cable and SMA, SMB, BNC, TNC, OSX (MSX) or GT5 connector. Operating temperature is –40 to +100°C.
Matsushita
Tel: 01908 231555
Enquiry No 505

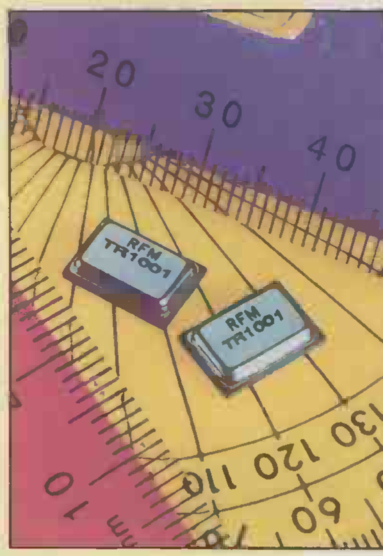
Rocker switches

The Otto K series illuminated rocker switches from Quiller Electronics are sealed and watertight to IP68, and can have through-panel holes for draining water from flat panels. They comply with standards for marine and offroad vehicles. The K3 has LED, incandescent or neon illumination. Customised legends can be stamped or laser etched on the lens. Available

in standard and low level contact ratings, its full rated load is 20A.
Quiller Electronics
Tel: 01202 436770
Enquiry No 506

DIP switch

Grayhill 97 half-pitch DIP switches are available from EAO. They come in four, six and eight position versions with lengths of 6.35, 8.89 and 11.43 mm, respectively. Contact rating is 100mA at 50V DC nonswitching, or 25mA at 24V DC and 10mA at 50V DC switching.
EAO
Tel: 01444 236000
Enquiry No 507



Mini transceivers for short-range wireless

Acal has introduced subminiature transceiver modules for short-range wireless data communications, initially at 433.92MHz (TR3000) and 916.5MHz (TR1000), but soon to be followed by 868.35MHz (TR1001) and 418MHz. Made by RFM, the modules use an amplifier-sequenced hybrid architecture. They support data transmission up to 115.2kbit/s and draw an average of 6mA from a 3V supply in transmit and between 1.8 and 7.5mA in receive, depending on setup. The TR1000's surface-mount package measures 10 by 7 by 2mm and the TR3000 11 by 9 by 2mm. Both have the same pin configuration.
Acal Electronics
Tel: 01344 727272
Enquiry No 501

CompactPCI connector

AMP has introduced a CompactPCI hybrid connector for transition board applications. Consisting of a two-piece connector on 2mm pitch, the Z-

Pack HM type AB comes in a 19 or 22 column format. The one-piece shroud is compatible with various backplane thicknesses to provide required pin mating levels. Used with select load feed-through pin-headers, the device will guide transition cards with connectors at positions three and five to avoid pin damage while mating. The hard metric receptacle comes in two styles – RJ3 and RJ5. The RJ3 is 37.9mm long with 95 signal contacts. RJ5 is 43.9mm long with 110 signal contacts.
AMP
Tel: 0181 420 8224
Enquiry No 508

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Diode array

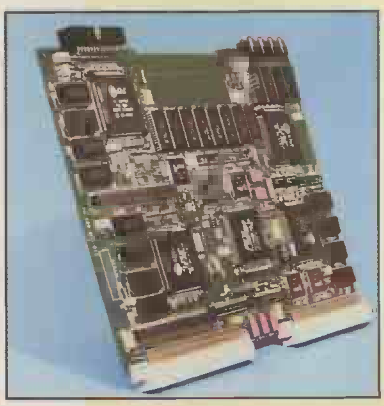
The SDC36 TVS diode array protects sensitive components in circuits used in industrial control applications and semiconductor devices connected to data and power lines. Protection is provided against overvoltages caused by electrostatic discharge, electrical fast transients and lightning. Available from Semtech, features include: transient protection for data lines to IEC1000-4-2 (ESD) 15kV (air), 8kV (contact); IEC1000-4-4 (EFT) 40A (tp +5/50ns); and IEC1000-4-S (lightning) 1kV, 2A (tp 1.2/50µs). It comes in a SOT-23 package.
Semtech
Tel: 01592 630350
Enquiry No 509

Pin isolators

Winslow has introduced modules for isolating individual pins. They can be supplied for various package styles, such as PGA, PLCC and dual-in-line. With dual-in-line switches, the user can activate or deactivate any of the IC pins, induce faults under controlled conditions or disable unwanted interrupt and resets when emulating. A module can be plugged into an IC socket on the motherboard or made available with a clip over option, enabling direct attachment to board mounted ICs.
Winslow Adaptics
Tel: 01874 625555
Enquiry No 510

Flash card

Samsung has launched a 32Mbyte Smartmedia memory card - the SMFV032 (32Mbyte plus 512kbyte) x 8-bit nand flash memory for nonvolatile storage applications, such as solid-state mass storage, digital voice recorders, digital cameras and other portable applications. Measuring 37 by 45 by 0.76mm and weighing 1.8g, it uses CMOS floating-gate technology.
Samsung Semiconductor
Tel: 0181 380 7200
Enquiry No 512



VME board

The VT-32C 6U VME board from Valley Technologies is based on four 80MHz DSP32C processors from AT&T. It provides each processor with 512kbyte SRAM, 512kbyte flash and 16kbyte DPRAM. Local buffered access connects to a configurable P2 and front panel interface. Global bus hosts a PMC site, 1Mbyte SRAM and A32:D32 VME interface.
Valley Technologies
Tel: 01506 885617
Enquiry No 511

Video LCD

Trident has released a 12.7cm video LCD module. The HC5031A is a 960 x 234 resolution TFT video LCD that will accept various input signals. Its analogue RGB format lets it reproduce 16 million colours and the panel has a brightness of 280cd/m².
Trident Displays
Tel: 01737 780790
Enquiry No 513

Computer board

The Maxi 6560 Eurocard-based CompactPCI board from Sight Systems is software compatible with desktop PCs and can operate without forced-air cooling. The 6U high single board computer comes with either a 166, 233 or 400MHz Pentium. User I/Os are routed directly to the front panel for access to the main module interface, which includes communications, VGA, PS/2 keyboard and mouse, printer, and LAN ports. Seven CPCI expansion slots are supported.
Sight Systems
Tel: 01903 242001
Enquiry No 514

PCI board

The PCICan from HM Computing is a single or dual Canbus PCI board that runs the Devicenet protocol, or other code, on its Siemens 167 processor. It provides dual ported memory to the PCIbus with bidirectional interrupts. The main engine is left to supervise the operating system and application software. The board provides remote, two-wire communication with standard I/O modules.
HM Computing
Tel: 01299 250997
Enquiry No 515

Power supplies

Acal has introduced the MVLT-100 medically certified single and multiple output 100W open-frame AC/DC power supplies. Made by EOS, the supplies measure 7.6 x 12.7 x 2.5cm. Typical efficiency at full load is 86 per cent. They weigh less than 280g and can be convection cooled up to full rated output at 50°C. Single output supplies are available with outputs of 12, 15, 24 or 48V, while the multiple output units offer selections of ±5, ±12, ±15 and ±24V. All accept inputs from 90 to 264V AC.
Acal Power Solutions
Tel: 01344 727272
Enquiry No 516

VMEbus module

A VMEbus module from BVM is a 33MHz 68040 CPU in a 3U form factor. With a typical power consumption of 10W and an operating temperature of –40 to +80°C, the BVME4500 is for use as a remote or embedded processor in industrial applications. Various memory options are dual ported to the VMEbus. Onboard are 2Mbyte of 32-bit wide non-volatile SRAM and 2Mbyte of 32-bit boot sector flash EPROM.
BVM
Tel: 01489 780144
Enquiry No 517

Analyser comes with 2GHz timing capability

The capability of HP's HP16600A and 16700A logic analysers, launched last year have been enhanced with the addition of two modules offering up to 333MHz state and 2GHz timing analysis. There is also an improved trigger capability. The upgrades have been in response to designers requirements to be able to work on systems operating with bus speeds above 100MHz. To help speed data capture, the analyser's triggering interface includes a scroll that allows users to select any of the most commonly used trigger functions quickly. When a trigger is selected, it is graphically depicted on the screen. This helps the design engineer to see exactly how trigger conditions will be defined.
Hewlett-Packard
Tel: 01344 366666
Enquiry No 526

Single-chip for motor control

Sunrise has announced the NEC µPD78098x single-chip microcontroller with on-chip features for controlling three-phase asynchronous or brushless DC motors such as in fans, pumps and home appliances. It has an 8/16-bit CPU with a 0.24µs minimum instruction execution time using an 8.38MHz clock. The architecture has four register banks, each with eight 8-bit registers. There are 63 instructions available, including full 16-bit operations. A 64kbyte linear address space is accessible using a 16-bit address pointer. Bit manipulation operations are supported on RAM and special function registers. On-chip memory combinations include mask ROM and flash. Flash memory can be written even with the device mounted in the target system. There are seven ports with a total of 39 input and output pins, most with pull-up resistors facilitating their use as inputs. Eight I/O pins can directly drive LEDs. An eight-channel 10-bit ADC is on chip. At 8MHz, average conversion time per channel is under 15µs.
Sunrise Electronics
Tel: 01908 263999
Enquiry No 518

Potentiometer

Xicor has announced a digitally controlled potentiometer, the X9317 100 tap with a typical standby current of 200nA and operation down to 2.7V in an eight lead TSSOP. It is for battery operated, mobile products such as laptop and palmtop computers, handheld barcode scanners and fibre optic communications. It uses an up and down three-wire interface. The resistance value is stored in a nonvolatile register and can be recalled on power up. The device can be used as a three-terminal potentiometer or as a two-terminal variable resistor.
Xicor
Tel: 01993 700544
Enquiry No 519

Temperature sensor

Telcom's TC74 digital temperature sensor is available from SEI Millennium. It enables temperature management in micro-based equipment such as printers, set-top boxes, fax machines and white goods. It comes in a SOT23A-5 package and occupies less than 10mm². It can be placed next to a high dissipation device such as a microprocessor. I²C and SMbus compatibility gives byte-based serial communication, and the device is addressable so several can be used in one system.
SEI Millennium
Tel: 01203 694555
Enquiry No 520

40MHz-1GHz RF amplifier

STMicroelectronics has introduced the TSH690 RF amplifier for use as a

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direct antenna driver in transmitters and for general purpose buffer applications. Using bipolar technology, it operates at 40MHz to 1GHz from a 1.5 to 5V power supply. A bias control pin allows adjustment of the output power and amplification mode. When the bias pin is connected directly to the supply rail, the device operates a class A linear amplifier, but when the bias voltage is reduced by inserting a series resistor, amplification efficiency is improved to class A-B behaviour. If the bias pin is grounded, the circuit is set in power-down mode without current consumption. The amplifier, 50Ω input and output internally matched from 300 to 1000MHz, exhibits 28dB gain at 450MHz, 20dB at 900MHz, and delivers an output power of +13.5dBm at 3V. The bias function controls the output power adjustment, achieving +18dBm at 4V and up to +20dBm at 5V. The device comes in an SO8 package.
STMicroelectronics
Tel: 01628 890800
Enquiry No 521

Limiting amplifier

Vitesse has introduced the VSC7958 2.5Gbit/s limiting amplifier for OC48 Sonet and SDH fibre optic links. Noise is 160μV RMS and it has a differential architecture and on-chip offset correction circuitry. The 5V IC is available in 16-pin glass-walled flat packs or in die form.
Vitesse Semiconductor
Tel: 011 510 226 2374
Enquiry No 522

HDTV chipset

Three chips make up the Giga GD145xx set for 1.485Gbit/s video transmission in HDTV systems. From DT Electronics, the set consists of the GD14510 equaliser, GD14515 serialiser and GD14516 deserialiser. The GD14526 retimer is available as an alternative to the GD14516. All meet the requirements of SMPTE 292M. The GD14510 has a loop-through connection for better than 15dB return loss at 1485MHz and cable driver or PECL output. It comes in a 40-pin MLC package. The serialiser has an on-chip clock multiplying phase locked loop, 20:1 multiplexer, scrambler, NRZI encode and cable driver. It has tuning capability in the range 1200 to 1500Mbits/s and a power dissipation of 1.2W. This device is housed in a 68-pin MLC package.
DT Electronics
Tel: 01203 466500
Enquiry No 524

Codecs

Cirrus Logic has introduced two AC-97 v2.1-compliant codecs for PCs and embedded systems. The Crystal audio chips provide Sony and Philips digital interface output for connection with consumer electronics products that support S/PDIF input. The CS4299 complements Intel's core logic chipsets due out soon. The CS4297A teamed with a PCI controller provides a two-chip audio product for embedded systems such as set-top boxes, hand-held computers and other devices that use the AC audio link.
Cirrus Logic
Tel: 011 510 226 2374
Enquiry No 523

Transient suppressor

The LCDA15C-6 transient voltage suppressor array from Semtech has been designed for multi-mode transceiver protection in telecommunications, networking systems, point-of-sale terminals and WAN. It protects up to six I/O lines or three line pairs operating between five and 15V covering all multi-mode levels. Featuring a low capacitance of less than 15pF per line, the device is suitable for use on high-speed interfaces. A low clamping voltage ensures minimum stress on the protected IC.
Thame Components
Tel: 01844 261189
Enquiry No 525

Multiplexers

National Instruments has announced two multiplexers for SCXI signal conditioning systems. The SCXI-1104 is for mid-level voltage analogue input

applications, and the SCXI-1127 for matrix and switching use in computer-based measurement and automation systems using the company's Labview or Labwindows/CVI software. The 1104 32-channel multiplexer can read voltage levels up to ±42V. It can monitor signals just outside the working input range of a data acquisition card, such as 12 or 24V signal sources. Each module multiplexes the 32 conditioned signals into one channel of the data acquisition module. Used together, several modules can multiplex hundreds of inputs into one channel. Each channel includes precision attenuation circuitry, instrumentation amplifier, and lowpass noise filter. Channels can be scanned at up to 333ksample/s. The 1127 is an armature relay device that acts as a multiplexer or matrix module for the SCXI platform. As a multiplexer, it can operate in one-wire mode for large channel-count systems, two-wire for differential pair systems, or three and four-wire for resistive measurements including RTDs and thermistors. With SCXI front-mounting terminal blocks, it operates as an 8 x 4, two-wire matrix switching module or as a 32 x 1, two-wire multiplexing module.
National Instruments
Tel: 01635 523545
Enquiry No 527

Hall-effect switch

The A3210 from Allegro combines pole-independent Hall-effect switching with a latched digital output. It has an operating voltage of 2.5 to 3.5V and a 0.1 per cent duty cycle. A chopper stabilisation technique eliminates inherent and environmental chip offsets to improve magnetic sensitivity and reduce switch-point shifts with temperature. It is fabricated in a BiCMOS process. Chip orientation allows for operation with a north or south pole magnet.
Allegro Microsystems
Tel: 01932 253355
Enquiry No 529

Limiting amplifier

Actel has available a 16 000-gate evaluation board for its SX and MX FPGAs. This board lets designers conduct real-time evaluation of functionality and performance of both devices. The MX is for 5V applications. The board's user prototype area lets designers evaluate their own designs to determine suitability for specific applications.
Actel
Tel: 01256 305600
Enquiry No 528

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Low power driver IC

From Ericsson, the PBL38085 driver IC lets designers replace LED display backlighting with an electroluminescent (EL) lamp in portable equipment such as mobile phones or palmtop computers. Supply voltage can be down to 2.7V. Output voltage is typically 170V, and it can drive an output current of up to 2mA. Power-down mode requires less than 0.1μA. The device comes in a ten-pin SOIC package, measuring 3 by 3mm. External components required are a choke, capacitor and resistor. A lamp dimming function is controlled by varying the duty cycle of one of the input signals. The device can drive a lamp area of up to 30cm².
Ericsson Components
Tel: 01793 488300
Enquiry No 530

RS485 drivers

Micro Call has introduced the Linear Technology LTC1688 and LTC1689 quad RS485 drivers that transmit at 100Mbit/s. This combination lets multipoint, high-speed backplanes be used in rugged environments where noisy grounds or common-mode variances have disqualified other technologies. They control the propagation delay window to 8ns ±4ns and use a standard pinout.
Micro Call
Tel: 01296 330061
Enquiry No 531

A-to-D converter

Cirrus Logic has introduced an integrated A/D converter chip for digital meters measuring residential electric power use. The Crystal CS5460 power-meter-on-a-chip is for helping utilities better access information needed to establish flexible customer rate plans, encourage off-peak energy consumption, and spur efficient energy resource management. It can also be configured for use in commercial and industrial meters.
Cirrus Logic
Tel: 001 510 249 4244
Enquiry No 532

Digital interface

Quintek has released the Sharcpac-FPGA10K programmable digital interface, integrating a 120Mflops 2106x Sharc processor and Altera Flex 10K-series FPGA on a standard Sharcpac module. The FPGA is memory mapped to the Sharc via a 1kword x 32-bit bidirectional Fifo providing an I/O bandwidth of up to 160Mbyte/s, depending on the interface functionality. All six onboard link ports are available at the Sharcpac connectors, as are the flag and IRQ signals.
Quintek
Tel: 01179 628196
Enquiry No 533

PCI development kit

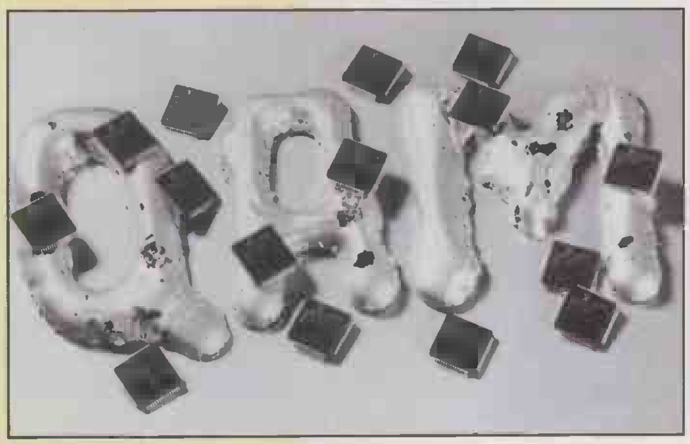
Quicklogic has available the QL5032 PCI reference development kit to support the first member of the QuickPCI ESP family. With version 8.0 of the company's Quick Works development tools, the development kit is a hardware-verified PCI product with a socketed 32-bit, 33MHz ESP device, compiled drivers for Windows platforms and source code for the software drivers and applications. The kit has been verified by the PCI SIG as PCI bus compliant, allowing for the maximum 32-bit PCI bus bandwidth (132Mbyte/s).
Quicklogic
Tel: 01932 579011
Enquiry No 534

CAN controller

Oki has introduced the MSM9225 CAN controller LSI for communications within internal automobile networks. The chip conforms to ISO CAN version 2.0B. As a multiple master bus, the chip accelerates the transmission at a data rate up to 1Mbit/s between microcontrollers operating in separate zones in a car, such as engine control units, ABS, airbags, power windows and suspension. It elevates the network bandwidth by transmitting up to 16 consecutive 8byte packets at once. This reduces the load on the controllers, and in turn speeds the transmission.
Oki Semiconductor
Tel: 01753 516577
Enquiry No 534

QAM demodulator uses triple-well technology

Fujitsu has announced the MB87L2070 DVB and DAVIC QAM demodulator. Digital processing for the QAM was developed by Libit Signal Processing. Fabricated in a CMOS process and using triple well mixed-signal technology, the chip includes analogue macros. It comes in a 64pin PQFP. The A/D and resampling algorithms are integrated. It complies with DVB-C and ITU-J.83 annex A and C standards, and supports direct IF sampling with variable frequency up to 60MHz for European, North American and Japanese IF frequencies. It includes a full QAM demodulation path and a 10-bit A/D function. The chip supports 256, 128, 64 and 32 QAM demodulation, with data rates up to 56Mbit/s in an 8MHz channel.
Fujitsu Microelectronics
Tel: 01959 562772
Enquiry No 534



Continued from page 655

Precise active x-over

Figure 7 compares the response of the low-pass filter with 4th, 6th and 8th order Linkwitz-Riley filters. The new filter has a slope between that of a 6th and 8th order Linkwitz-Riley filter.

Figure 8 compares the group delay of the new low pass filter (bottom) with Linkwitz-Riley 6th order and 8th order filters. In spite of the steep cut-off of our new filter it has the lowest group delay and the least variation making the application of group delay equalisation easier.

Figure 9 shows an implementation of the crossover. A Bainter filter provides two complex poles and two complex zeros. The following voltage-controlled voltage-source filter provides the remaining complex poles with zeros at infinity or dc, depending on whether it is in the low-pass or high-pass section.

A Bainter filter is used to produce the notch. Although it requires three operational amplifiers, it is tolerant of component variations. Ideally 1% components should be used throughout.

The op-amps could be *NE5532* types. The two paths do not have the same pass-band gain and they require some equalisation of level. Also, the internal operating levels are higher than the output, requiring the outputs to be run 10dB below the normal maximum.

For those of you who interested in experimenting with active crossovers, see the panel entitled 'Time delay equalisation for non coincident drivers'. For good stereo imaging, correction for misalignment of the acoustic centres of low and high-frequency drivers is absolutely essential.

As a practical example with the drivers used in this crossover design, the displacement between bass and treble drivers was 8mm. At normal room temperature the speed of sound is 345m/second, from this you can calcu-

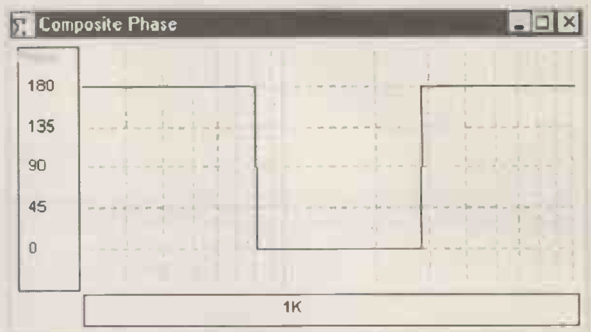


Fig. 6. Phase difference plot of the two filter sections shows that throughout the crossover region, there is no phase error.

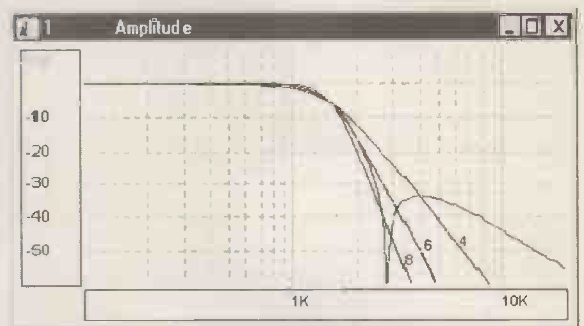


Fig. 7. The new low-pass filter response compared with 4th, 6th and 8th-order Linkwitz-Riley filters.

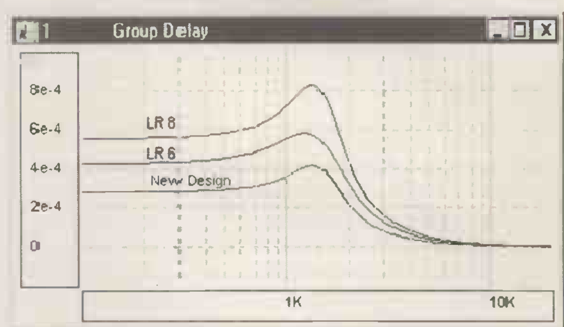


Fig. 8. Group delay of the new low-pass filter in comparison with those of 6th and 8th-order Linkwitz-Riley alternatives.

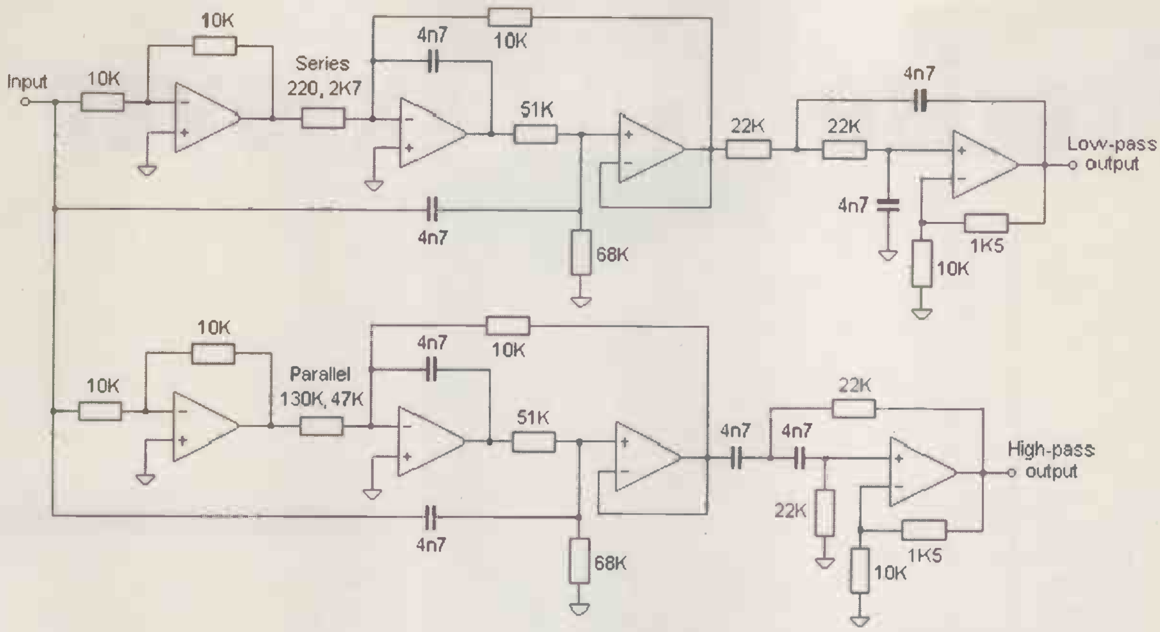


Fig. 9. Implementation of the new filter for 1.5kHz crossover. The design is reasonably tolerant of component variations. Op-amps like the NE5532 could be used here.

Time-delay equalisation for non coincident drivers

High-quality loudspeakers usually have the input signal divided into separate frequency bands, each with its own specialised driver. **Figure D** shows most common configuration – a cone bass driver and dome tweeter.

The larger the bass driver, the further back is the effective point of radiation. For both units individually, the optimum listening axis is straight ahead, but in the crossover region, where they are both radiating, the wave front will be bent downwards. This is because the centres of radiation are not on the same vertical axis and the output of the bass speaker is effectively time delayed compared with that from the treble speaker.

The effect of this misalignment is to prevent the composite speaker from effectively mimicking a single sound source. The source of the sound is diffused, not tied to any particular point at the front of the loudspeaker. Worse still, when listened to at a variety of angles the sound source may appear to move to some point either side of the speaker. These effects get worse as the bass driver gets larger and its effective point of radiation moves further back. Needless to say the effect of this on stereo image is disastrous.

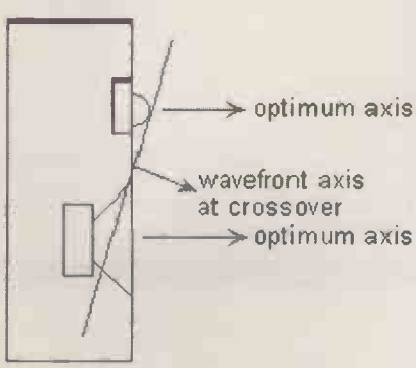


Fig. D. Because the voice coils in two-way systems are rarely aligned vertically, sound from the bass unit takes slightly longer to reach a listener directly in front of the speaker than does sound from the tweeter. At frequencies around the crossover, where both speakers are contributing to the same sound, this difference in distance causes phase errors – i.e. unwanted signal additions and cancellations.

The solution is either to move the tweeter back in line with the bass unit, or to electronically delay the signal to it by an equivalent amount. The effect of delaying the information to the treble speaker is to bring the perceived sound source to a point somewhere between the bass and treble unit; moreover it remains firmly rooted to that point irrespective of the angle of ones head to the speaker position. The effect of the treble delay correction is a dramatic improvement in stereo image.

The problem of acoustic misalignment becomes more serious when a large bass driver is used at a relatively high crossover – often as high as 3kHz. A 200mm bass driver and tweeter can have a acoustic centres misaligned by 40mm. The wavelength of sound in air at 3kHz is 115mm so 40mm represents a phase shift of 125° with 180° occurring at 4.3kHz. With these phase shifts, the polar response throughout the crossover transition will be seriously impaired.

With typical units the displacement is unlikely to be above 40mm. For some small low-frequency drivers with an integral dome intended for mid range operation, the error will be less than 10mm.

The simplest approach to correcting the problem is to generate a non linear delay that produces the time delay required only at the crossover frequency. A more thorough approach is to produce a linear-phase time delay, that is a time delay that is as near as possible constant over the audio range. This can be done using the poles of a Bessel filter to form an all-pass filter.

late that a delay of 23μs is required.

Since phase linearity is necessary, the desired time delay must be maintained up to high frequencies. A practical limit was set at 16kHz.

The pole and zero locations are derived from a second-order low-pass Bessel filter with the cut-off frequency set to give half the required group delay. Transforming from low-pass to all-pass doubles the group delay to the required value.

Figure 10a) shows how the high-frequency driver delay can be implemented using a second order all-pass filter. This provides the delay shown in Fig. **10b)**. ■

Crossover phase response

In general, the band-splitting filters used in a crossover can be described as in-phase, or linear phase or both. A set of filters that are in-phase may not be linear phase, and conversely, filters that are linear phase are not necessarily in-phase.

An in-phase set of filters are those which have the same phase throughout the transition or crossover region. The most notable of this type are the Linkwitz-Riley alignment. The characteristic is important because it produces a good polar response throughout the region where one driver is transferring sound output to the other.

A linear-phase crossover is one where each of the filters has a linear phase response – at least through its pass band and the transition region. However in some designs this is achieved at the expense of a constant phase difference throughout the transition region, of 90° or less, with the inevitable degradation of polar response.

There have been many attempts to circumvent these problems and achieve both phase linearity and in-phase transition by varying the overlapping of filters or utilisation of time delays in the summation of filter responses. These have varying degrees of success and are of sufficient complexity to make a digital implementation the only feasible solution. References 10 to 14 contain a considerable amount of practical information relating to these subjects.

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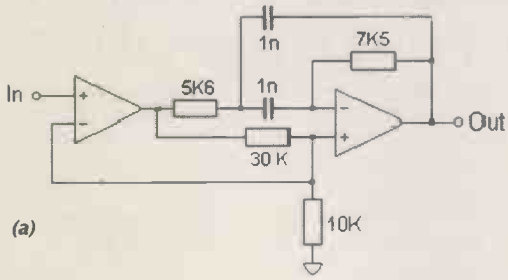
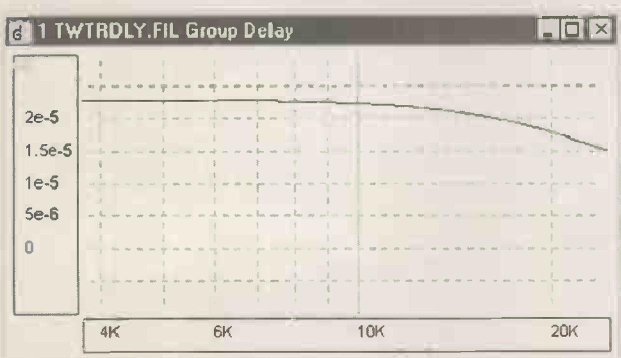


(a)



(b)

Fig. 10. High-frequency filter delay can be implemented using a second-order all-pass filter as in a). Delay performance of the circuit shown is plot b).

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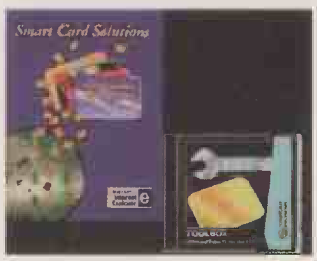

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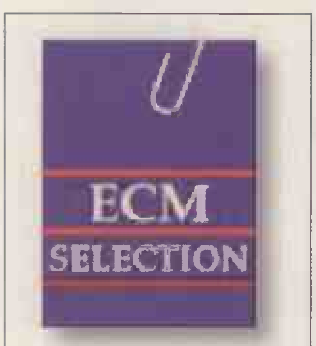

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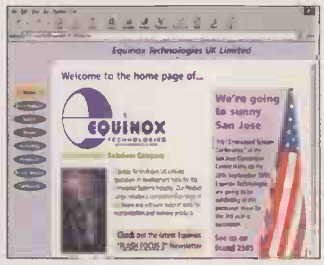

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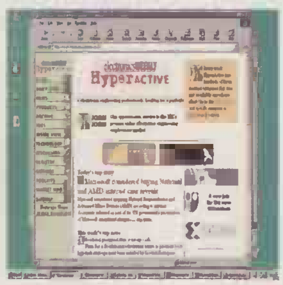

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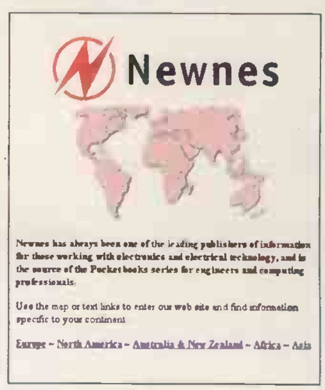

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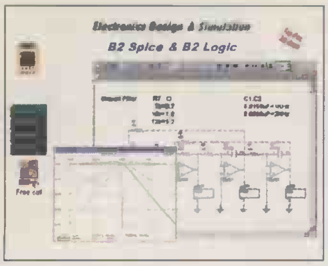

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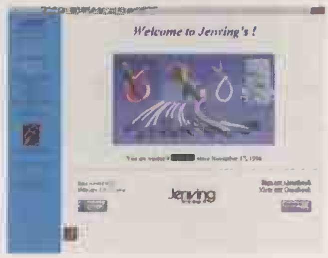

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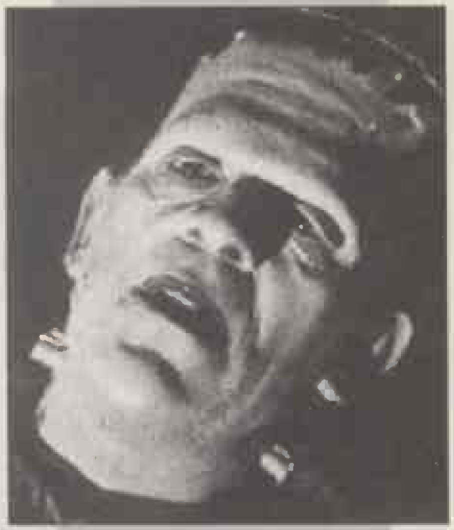

The route to simulation II

Continued from page 635

Another contentious issue is the provision of teaching aids. These are fine for teachers. But they are of little or no use to the electronics designer, who will nevertheless have to pay for them. Sometimes they can even get in the way. These aids should really be left out of the main program and offered as an add-on – with an appropriate price adjustment.

Many of the programs at this level do not have any operator control over which glitches are shown in digital analysis. This may not be important for a program aimed at the educational market, but a practical designer will want to see the glitches that are significant. The designer does not want to see a forest of unimportant glitches – for example, those that are too short in duration to affect the ICs being used. Manually sorting through an analysis on a program that shows all glitches regardless of duration can be extremely tedious.

On the other hand, if a design does not function correctly and the designer suspects a glitch, he or she does not want to see an idealised set of waveforms with some or all of the glitches removed.

Therefore some method of filtering out the glitches that the designer regards as unimportant is required, and this implies that it should preferably be operator-adjustable, to suit the varying requirements of the work in hand.

CircuitMaker Pro v6

CircuitMaker Pro version 6, referred to from here on as *CM6*, is an integrated schematic capture and mixed-mode simulator program. It has no pin limit. The simulation engine is based on Spice 3F5 and XSpice.

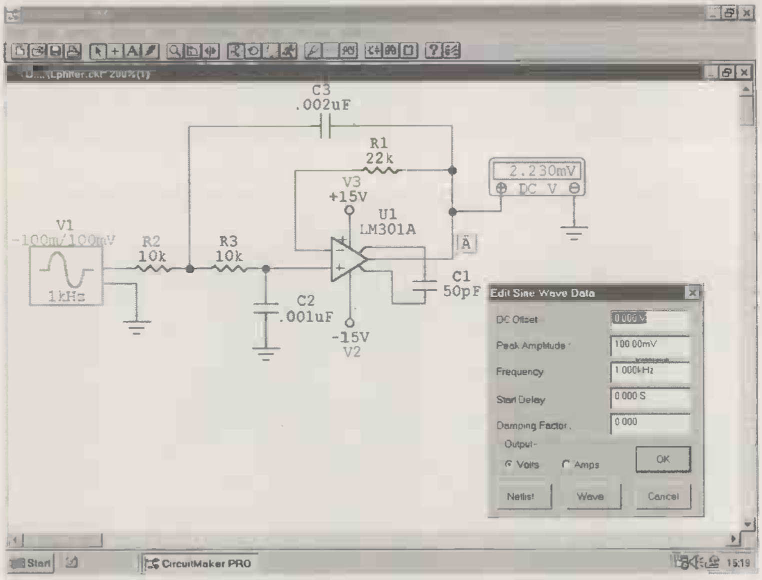

Fig. 1. View of a typical CircuitMaker *schematic, showing a signal generator being set up via a menu. Note the method of wiring the signal generator and dvm into the circuit under test.*

Schematic capture

You will find a detailed discussion of *CircuitMaker*'s schematic capture feature in the November 1996 issue. In brief, *CM6*'s schematic section is to be commended. It allows you to choose whether you wire up automatically or manually. Wiring errors are difficult to make, as they are automatically erased, and there is an excellent system for positive confirmation of connectivity.

There have been several changes since my last review. It is now possible to drop a schematic symbol into a circuit wire instead of having to delete the wire and then reconnect. It is also possible to gain access to the symbol editor by a single right click.

Borders can be drawn to split circuits into separate pages, and a fit-to-page function can now resize the circuit into the page. The drawing sheet 'auto-pans' when moving symbols around. All these, and a few more added features, make drawing quicker. However, one small change has been made that makes drawing slightly slower, and this concerns the 'hot keys'.

Symbols in *CircuitMaker* are nearly always selected from two permanent user-definable parts bins. These are

Requirements

CircuitMaker is provided on floppy disk or CD and security is by registration number. The requirements are WIN3.1 (installed with WIN32s as CM6 is 32-bit), WIN 95 or NT and 15Mbyte of disk space (20Mbyte to install), and a 486DX PC or better is preferred.

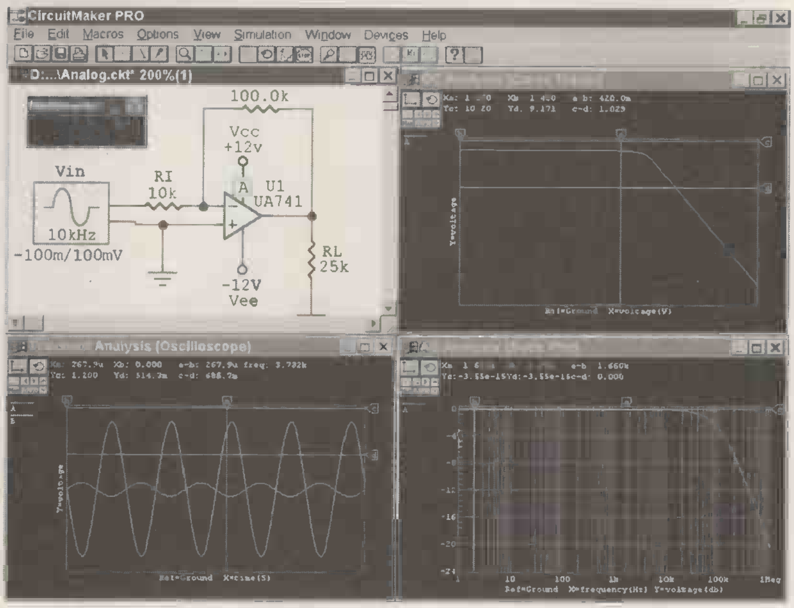

Fig. 2. A set of default analyses showing the presentation style in CircuitMaker*. Any of these can be resized to make accurate measurements.*

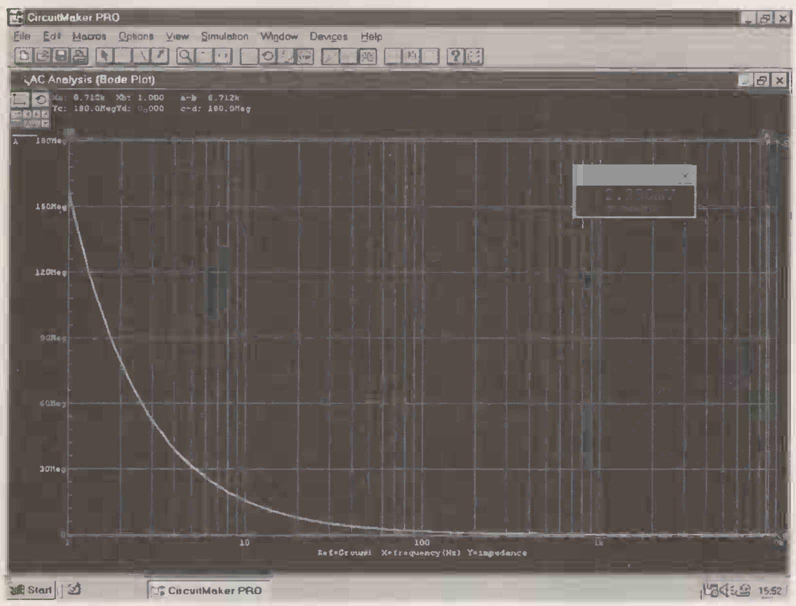

Fig. 3. With a little extra effort, an impedance plot like this one of a Sallen and Key low-pass filter can be made.

Those of you interested in CAD may like to know that MicroCode, the makers of CircuitMaker, has been acquired by Protel International. Protel's intention is eventually to incorporate MicroCode's simulation technology into Protel's EDA system. However, it seems that CircuitMaker itself will not only continue to be sold, but developed as well. It will be interesting to see what comes out of this Australian-American alliance.

referred to as hot keys 1 & 2, as all the symbols contained in them can be selected from the keyboard. This makes the schematic drawing section particularly well suited to experimenters who want to design on-screen.

In the previous version, the hot keys were directly accessible from the menu bar. In *CM6* the hot-key bins are hidden in the 'Devices' menu. The hot-key parts bins are opened and closed continually during circuit drawing, since it is impossible to memorise every hot key for every symbol, so the quickest possible access is desirable. This has now become a two-click process, whereas previously it was one. Just why this core aspect had to be changed is anyone's guess. It seems to be a promotion of style over utility.

Despite this small matter, *CircuitMaker* still one of the best examples of a schematic drawing system. Drawing is fast, intuitive, and almost foolproof, and the learning curve is gentle.

The notional drawing sheet area is large – 32in by 32in. The symbol library contains over 400 symbols, backed by a library of 6000 Spice device models, which is good coverage. It is possible to add your own symbols via a symbol editor, and attach Spice model data.

Net lists can be exported in *Spice3, Tango, Protel, OrCAD, PADS Cadnetix* and *Calay*, as well as in MicroCode's own *TraxMaker* pcb format.

CircuitMaker versus *Workbench*

It is perhaps inevitable that *CM6* is compared to *Electronics Workbench V.5* as both packages are the same price and appear to be similar at first glance. However, they do not cover the same ground, and they have very distinct and separate styles of operation.

Although *CM6* uses a type of virtual instrument concept, it is executed in a quite different way relative to *Workbench* and *Tina*. In *CM6*, virtual instruments are not set up via instrument-style front-panels that are reminiscent of the real thing. Instead, the instruments are configured from a Windows-style menu screen like that shown in **Fig. 1**. Nor are the results displayed as if on instrument front panels, but rather as graphs in a plain window. There is something to be said in favour of both systems, so I think which system is preferable is a matter of personal choice.

Although things like the signal generator and multimeter are wired into circuit in a similar way to *Electronics Workbench*, for some analyses, the method of connecting into a circuit to obtain a simulation in *CM6* is by using a mobile virtual probe. This has the benefit of providing several consecutive sets of results quickly just by moving the probe around – i.e. no re-wiring of instruments is needed. This is still a highly graphical and intuitive system. It is difficult to say which is the best method.

The presentation style of a typical set of results in *CM6* is shown in **Fig. 2**. Results from various analyses can be stacked for ready back-reference. Note the probe A in this illustration. You can insert several probes if required. Even if you do not choose probe points, *CircuitMaker* can still run simulations on a default basis, and are auto-scaled, which is good to have when you are learning and making frequent errors.

CircuitMaker differs from *Workbench* and *Tina* in another respect. With *Workbench* it is, to a large extent, possible to avoid getting involved in Spice, but in *CircuitMaker* you need some knowledge of Spice to get the best out of the program. This may give *CM6* more flexibility, but to take advantage of it means learning just a little more about Spice and its commands.

The virtual instrumentation consist of an oscilloscope, signal generator, pulse generator, data sequencer (word generator), and multimeter. Any number of oscilloscope probes can be attached to a circuit to give a multiple trace display. Any number of meters can be attached.

If you wish, virtual instruments that you use most frequently can be inserted in the hot-key bins like normal circuit components, to improve access speed.

There is an extensive range of analyses. These include transient analysis, DC analysis, AC analysis of amplitude and phase versus frequency, noise, Monte Carlo, worst case, Fourier, temperature sweep, transfer function, and input/output impedance.

The method of obtaining the input/output impedance graphs is indirect, but it is possible to produce good results. Some impedance results for a Sallen and Key filter are shown in **Fig. 3**. The only notable thing missing in the analogue section is distortion analysis.

As well as a mixed-mode simulator, *CM6* has a digital simulator. To distinguish the components that can be used for digital analysis, the library components are marked either for analogue, mixed mode or digital use. This approach differs from *Workbench*, where everything is treated in mixed-mode fashion. The digital simulator in *CM6* offers features such as variable simulator speed and the ability to show digital activity by wire colour. With this system, high is red, low is blue and tri-state is green, and it can be very helpful to watch the changes in wire colour to see what is happening. **Fig. 4** shows an example.

In this simulation mode, the probe tool doubles as a logic probe that can be freely moved around to show the state of nodes in the circuit. An unlimited number of waveforms can be observed in the window, whereas most other logic analysers have a fixed limit, usually 8 or 16. There is no glitch threshold control. And there is no logic converter device as there is in *Workbench*.

On the educational front, *CM6* has fault injection and a number of animations to add interest and introduce a practical note. Another interesting feature of *CM6* is that it gives access to *SimCode*. This is MicroCode's method of enabling digital models to be inserted into the analogue simulator to give mixed-mode operation. Some of the principles behind mixed-mode simulation were mentioned in this review's introduction last month. *SimCode* is a high-level language that takes some of the difficulty out of adapting any digital models that you want to add to the mixed-mode simulation.

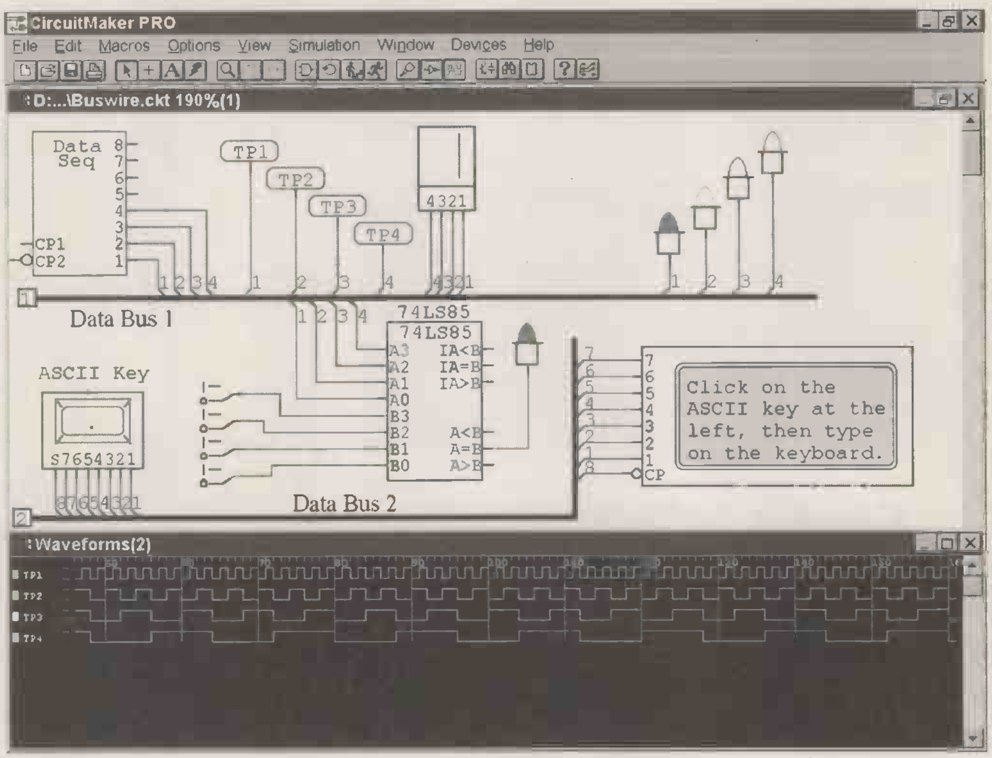

Fig. 4. A digital circuit showing the coloured wiring system which indicates digital state, a typical timing chart, and the good use of graphics – the leds in this example light up, as in the real world.

Documentation

Two soft-back books form the software's documentation. The larger of the two is a user manual describing everything to do with the program, from installation, through device modelling to troubleshooting. The second book lists the library contents and briefly describes the major components and how they are modelled. The main manual is well written and concise, includes a tutorial session, and is well complemented by the Help files.

The user manual does not assume much prior knowledge of this field and gives lucid explanations of many things a first-time buyer might find difficult. For example, the chapter headed 'Spice – beyond the basics' gives an insight into the practical problems that are often encountered when using Spice.

Taken overall, the documentation is good, clear and concise.

In summary

CircuitMaker 6 is a user-friendly program – especially in the schematic drawing section, which is notably easy to use. As a comparatively inexpensive experimental tool, the simulator section has many interesting features, and much to commend it to the first-time buyer, including a well-written manual.

The package comes close to meeting the requirements for a basic general-purpose simulator as suggested in the review's introduction. It only misses due to lack of a distortion analysis and glitch threshold control. But these are offset by the availability of input/output impedance graphs not available in *Workbench*.

CircuitMaker Pro v6
MicroCode Engineering, USA.
UK supplier Labvolt
tel. 01480 300695, price £199.

If you need neither distortion nor glitch control, then *CM Pro* is a very attractive proposition. The animations, fault injection and general ease of operation make it useful for educational use. Superficially, *CM6* may seem like *Workbench*, but there are many small differences, examples of which have been given above, that distance the two programs from one another.

At £199, *CM6* is good value for money considering the range and scope of what is provided. ■

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