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FEBRUARY ISSUE - ON SALE JANUARY 25.

January 1996 ELECTRONICS WORLD
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CIRCLE NO. 155 ON REPLY CARD
Chasing the dragon

A sia is an obsession of this Government. If only we could emulate Asian growth rates and success in the new high-tech industries, think ministers, the UK would be a 'Tiger' economy – a kind of Taiwan or South Korea. What is it about the Asian Tigers that our Government fears has created their success? Low tax, low interest rate, laissez-faire, deregulated capitalism red in tooth and claw, seems to be the Government's answer.

To anyone who follows the success of Asian high-tech industries this is a laughable misapprehension. The most successful of the Asian Tigers, Japan, have had governments which, very carefully and consistently, have followed a policy of acquiring – then disseminating as widely as possible – the key technologies for success in high-tech.

Take the most successful Asian economy – Japan. Back in the fifties, the Japanese government put import restrictions on US microelectronics products; forced US companies selling microelectronics in Japan to license their technology to Japanese companies; and forced Japanese holders of such licences to sub-license the technology to spread it to as many companies as possible.

Take Taiwan. In 1976 the government bought seven micron microelectronics technology from RCA and has subsequently refined it – in the same laboratory – through every subsequent technology generation down to a modern 0.25 micron process. From time to time the Taiwan government has spun-off its pre-production r&d process lines into the private sector forming such companies as UMC, TSMC, Winbond and Vanguard.

In Korea much the same thinking produced the microelectronics r&d organisation KIET (Korean Institute for Electronics Technology) and the 1982 Semiconductor Industry Promotion Plan which directly led to the moves of Samsung, Hyundai and Lucky Goldstar into the memory chip market with huge success – today microelectronics accounts for one tenth of Korea's total exports.

What is it about the Asian Tigers that the new head of one of the UK's largest electronics concerns – BT is to be a microelectronics man? Peter Bonfield, is ex ICL and, more importantly, ex of microelectronics giant Texas Instruments.

The two strengths of microelectronics men are: a belief in laissez-faire; and the user end and another at the exchange

disgustful to the citizen wanting better services, or to anyone who wants the UK to be an efficient industrial economy. But it makes perfect sense to a BT executive seeking maximisation of the profit earning potential of every technological change – which usually means delaying change.

To a microelectronics man, such a willful disregard to promote and implement available technology goes against the grain. If Bonfield can get the learning curve mentality of Microelectronics Man to replace the dead-head mentality of Telecoms Men and Government Man he will do the UK a massive favour.

David Manners
**GPS attacked over GPS chip claims**

EC Plessey Semiconductors, GEC, has clashed with Rockwell Semiconductor and Motorola of the US in claims over chip-sets for the global positioning system market.

“We are the only people offering a complete integrated chipset,” said Brian Hick at GPS. “All the others are built around dsp cores or use discreet front ends.”

“Absolute rubbish!”, said Nigel Williams, chairman of Manhattan Skyline, Rockwell’s UK distributor.

“We took the first large European order for Rockwell’s integrated gss chipset.

“We’ve had three to four design wins already and one production order. In the accounts we’ve been in we haven’t seen any competition."

Motorola also disputes GPS’s claim. The company has been competing in the gsp market at board level but is about to move to chipsets.

“We’ve not announced a product yet but we’re talking to selected customers about an integrated chipset,” one of the Motorola gsp team commented.

However, Brian Hick said that GPS had sold 150 of its development systems for its chipset, which consists of a bipolar silicon front end, a 35MHz surface-acoustic-wave filter and a 12-channel correlator. An ARM 60 processor is used with the chipset.

“We want to see if customers want further integration”, said Hick. “The obvious move is to integrate the ARM core and the correlator.”

Rockwell’s latest two-chip set is called Zodiac. Encapsulated together in a PQFP are a gallium arsenide front end and an analogue c-mos die.

The second chip contains a single die incorporating a 12-channel dsp, a microprocessor and co-processor in a TQFP. The set costs $70.

Others targeting the gsp chipset market are Philips and SGS-Thomson.

David Manners, Electronics weekly

---

**Pictures from radio to enhance your PC**

In an intriguing move, Philips Semiconductors has developed a module and a set of components that let pcs receive and decode RDS (radio data system) broadcasts, displaying the information on the screen.

Dubbed the Smart-Radio module, Philips claims it provides the first high-quality radio reception for pcs, superseding previous solutions that relied on chips developed primarily for car receivers.

“Now radio can be seen as well as heard,” said David Canha, sales manager for Philips Semiconductors.

“Smart-Radio expands the listening experience to computer users as well as providing on-screen text.”

Kaveh Kianush, project leader of Philips’ radio IC design team, adds that just as tv tuners have been added to the pc so some users have been demanding stereo-quality radio reception as well.

“In a multimedia pc it might be nice to have some background music playing while you perform another task,” said Kianush. “The RDS facility then adds the ability to receive data, which might be market information or advertisements.”

RDS broadcasts are widely used throughout Europe to transmit the station’s identification, traffic bulletins, weather and other information alongside the normal fm signal. In the US the Electronics Industries Association (EIA) is supporting the RDS standard and has launched this year a $1m campaign to install hardware encoders in the top 25 radio markets across the US. The EIA’s plan is to equip several radio stations with the encoders, allowing RDS signals to reach 85% of the US radio audience.

The heart of Philips’ strategy is the OMS604 module, which is the fm radio tuner and preamplifier. The module is carefully shielded to isolate it from the noisy pc electrical environment and prealigned.

It uses the TEA5757H tuner IC designed by Kianush’s team for this application. The tuner IC uses twin frequency locked loops to provide speedy scan tuning and tuning to up to 99 frequency presets. The module is completed with a preamplifier providing 900mV line audio outputs and an pc bus controller chip. The module is programmed and controlled via this bus.

PC and pc-card makers can then augment this module with the SAA6779 RDS demodulator and the CCC921 RDS decoder chips. Philips engineers have developed a prototype pc plug-in card based on the OMS604 module and written a Windows software utility to control the radio from the screen.

Kianush says the card and the software are for demonstration purposes and it is unlikely Philips will sell either on the pc user market. “The idea is to support pc and card makers with the architecture,” he said.

Simon Parry, EW
National Semiconductor has introduced an op-amp that has a slew rate of 4100V/μs which it claims is the world’s fastest voltage-feedback operational amplifier.

The LM7171 is indicative of trend in which innovative amplifier front-ends and high-performance processes allow voltage-feedback op-amps to encroach on current-feedback territory. This tempts engineers who are ignorant of current-feedback techniques to design high speed circuits.

However, National Semiconductor engineer Mark Holdaway thinks current-feedback op-amps will continue to reign at the top. “Current-feedback op-amps will allow voltage-feedback op-amps to drive heavy, capacitive loads like video cables with claimed low distortion, resulting in high-quality images. Supply current is typically 6.5mA and offset voltages down to 200μV are available.”

EMC awareness receives a boost

The DTI launched on 1 December 1995 its £100,000 extension to the EMC awareness campaign. The money will pay for five technical consultancy hotlines around the country.

The extension is to be managed by Salford University Technical Services which says the current 16 emc clubs, established as part of the original £450,000 awareness campaign, will be grouped into five consortia. These will each run a hotline providing full-time technical support for six months until May next year.

David Southerland of the DTI said the intention of the extension was to “deal with the immediate panic” before EC regulations are enforced on 1 January. “A lot of people still have questions but the emc clubs haven’t got any spare capacity,” said Southerland. “The extension will help.”

Southerland added that the extra money would be found within the DTI and said the Microelectronics in Business programme would not be affected.

UK X-ray litho work to continue

Oxford Instruments is continuing development of its Helios II X-ray synchrotron for IC lithography despite uncertainty over the viability of the technology. X-ray lithography is considered slow and too expensive for production purposes by some semiconductor manufacturers.

However, Oxford Instruments believes the synchrotron, due next year, is not an expensive machine considering its capabilities and is in discussion with a number of IC makers.

“I don’t think the synchrotron is expensive. Considering the investment needed, our machine doesn’t cost much. Optical steppers are slightly less expensive, but still comparable,” said Alistair Smith, head of the semiconductor processing division at Oxford Instruments.

Helios II is a twenty-stepper machine that can provide continuous operation from a high intensity X-ray source. Oxford Instruments expects X-ray lithography to become the norm before the end of the decade.

Apple pushes PowerPC against Intel

Apple Computer, worried about the progress of PowerPC microprocessor development, has pressured its partners IBM and Motorola to boost the microprocessor’s performance sooner to keep pace with Intel.

Although Apple says that PowerPC is still a better platform than Pentium Pro, it acknowledges that the PowerPC’s performance advantage has been eroded by Intel’s Pentium Pro family.

David Nagel, vice-president of

NiMH batteries could see doubling in capacity

Nickel-metal-hydride battery with a claimed energy capacity twice that of typical NiMH cells is being developed by Matsushita. The battery stores 300Whr/l compared to conventional NiMH cells which offer around 170Whr/l.

Nickel-metal-hydride cells offer higher energy densities than Nickel cadmiums and have a lower environmental impact. They are appearing in portable products like mobile phones and laptops, but there is some debate whether they will be overtaken by lithium technology secondary cells.

Lithium ion cells have higher energy densities, around 280Whr/l with 360Whr/l predicted. The Matsushita battery, if it proves to be mass-producible, brings NiMH strongly back into the fray.

AER Energy Resources of Columbus, Georgia, is claiming a 50% increase in the capacity with a 78% hike in output power for its rechargeable zinc-air batteries. AER makes batteries that fit under popular laptops increasing their run time to over 12 hours between charges.

Apple Research and Development, said: “We have been concerned about PowerPC development and we have communicated that concern to IBM and Motorola. They have promised to address the issue and we are confident they can get the PowerPC back on track.”

When the PowerPC was first unveiled, its supporters promised a 200% increase in performance compared to Intel microprocessors. The performance advantage has been reduced to between 10 and 50%.

“Fifty per cent is still an important advantage for a lot of computational tasks,” says Nagel. “I wish it wasn’t the case, but Intel has done a good job in narrowing the gap.”
UPDATE

Cabling makers dispute unshielded twisted pair emi

Claims that unshielded twisted pair cabling may not meet emissions rules under the EMC Directive have been attacked by three leading cable network suppliers.

Nortel, AMP and AT&T have demonstrated that category 5 UTP shielded data installed networking equipment outside the limits of the EN55022 emissions specification, provided the terminal equipment is fully EMC compliant.

"There is a great deal of misinformation rumour generated by the shielded cable manufacturers," said Andrew Green, marketing manager for Nortel Cable Networks. "It simply isn't true that shielded cabling is the only way to meet the EMC Directive and we have demonstrated it at 155Mbit/s."

Nortel tested a 155Mbit/s ATM broadband communications link running over 100m of UTP category 5 cable including patch panels. Measured emissions were at least 8dB better than the EN55022 Class B (domestic) limit from 30MHz to 1GHz.

Current legislation requires that all active LAN equipment should be certified to EMC Class A in commercial environments from the beginning of January 1996.

According to Green, as long as the cable terminations are suitably balanced category 5 UTP will support data rates up to 155Mbit/s and meet the EMC regulations.

Balancing the current flow in each arm of the twisted pair cable is necessary to ensure that noise currents are cancelled out at the receiver input. Nortel also suggests that inductive chokes are connected in series with the input transformer to keep the common-mode rejection ratio to the necessary 40dB.

Part of any common-mode noise signal is converted into a differential-mode noise signal which passes directly into the input of the receiver. The level of balance in the UTP link was measured by Nortel to be better than the necessary 40dB from 20 to 100MHz and better than 50dB from 1 to 20MHz.

According to Paul Cave, manager of AMP's networking group, manufacturers with only shielded products are using the EMC argument to boost sales. "There is confusion in the market caused by people with a vested interest in streaming the market down the shielded route," said Cave.

Richard Wilson, Electronics weekly

Surround sound for the pc embryo

Aimed at multimedia pcs, Dolby Laboratories has introduced Dolby Surround Multimedia -- a cut-down version of the Dolby Pro-Logic system used in home cinema.

The intention is to promote Dolby Surround-sound encoding among games companies who can use it to increase the sense of participation in their games.

The signals that Surround Multimedia decodes are the same as those in the Pro-Logic system, but the different listening requirements of a pc, allow the playback system to be simplified.

A pc user sits in a well defined position in front of the computer. The home cinema audience, on the other hand, is spread around a room. The '3D' sound field of a PC does not need to be so big, and the Multimedia Surround system exploits this.

Windows 95 sales misjudged

Despite being the most highly marketed product in history, Microsoft's Windows 95 is not reaching its sales targets and most home pc users say that they have no immediate plans to install the system.

Market research firm Dataquest says that it misjudged earlier forecasts of demand for the operating system and has reduced its forecast for 1995 shipments by 3.6m. It now says that 26.4m units will ship this year instead of 30m, and expects 10m units will remain unsold at the end of the year, gathering dust on shop shelves.

"The lower forecast is due to two factors. It is an artifact of forecasting a product at a time when it was not yet in the market, when there was no final shipment date and Dataquest is taking a more conservative view of the holiday season," said Dataquest analyst Paul Cubbage.

Also a survey of home pc users by market research firm Odyssey, found that just 6% of respondents had installed Windows 95. About 53% said that they would not install it and 30% said they didn't need it.

Quarter micron geometries for GPS

GEC Plessey Semiconductors (GPS) has unveiled its technology roadmap coinciding with potential fab capacity down to 0.25µm. Next year it anticipates having its first silicon germanium chips on 0.35µm technology.

By 1997, the company will prototype 0.35µm c-mos production, with a 3.3V 0.35µm process scheduled for 1998. A further 2.5V 0.25µm c-mos process is expected by the end of the decade.

GPS is to pursue a more aggressive track in bipolar with 0.5µm, 5V and 3.3V, triple metal layer processes in production next year.

The company will sample a 0.35µm SiGe process next year with production scheduled for 1998. By the end of the decade GPS could have 0.25µm SiGe process.

GPS's first SiGe products will be 2.4GHz rf front-ends, low-noise amplifiers and mixers.

Monitoring heartbeat

It has been brought to our attention that Baki Koyuncu's "Monitoring Heartbeat" feature published in our July 1995 issue is based on John Becker's "Biomet Pulse Monitor" design published in the February 1993 issue of Everyday with Practical Electronics and apologise to them and to their Technical Editor John Becker for this unintentional infringement.
You can simply plug the new TiePieSCOPE - HS508 into the parallel port of your portable or desktop PC. With the advanced software, you can use this two channel, 8 bits, 50 MHz measuring instrument as a fast digital storage oscilloscope, including a lot more features than a single oscilloscope! Moreover, the TiePieSCOPE - HS508 contains a multiple display voltmeter (up to 5 MHz true RMS), a spectrum analyzer with an harmonic distortion meter and a transient recorder for recording a variety of signals.

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Gravity reveals military vehicle secrets

Experiments to measure different aspects of gravity are unremarkable in the modern laboratory. But if three scientists at Mitre Corporation in the US are correct in their assumptions, we could one day see portable gravity gradiometers scanning dusty lorries as part of arms limitation monitoring. Or even being used by Police on the M25, checking if vehicles are overloaded.

According to the researchers, Steven Gray, John Parmentola and Richard LeShack, the basic principles which make gravity gradient measurements possible are relatively easy to understand - it is the hardware and software engineering principles that are really involved.

Suffice it to say that the team proposes to use gravity gradients to monitor the weight of military vehicles, to help assess what type of arms, if any, they are carrying. Other studies have suggested that gravity profiles could be used to distinguish between conventionally-armed and nuclear cruise missiles. Or used to count the number of warheads on board an intercontinental ballistic missile or submarine-launched ballistic missile.

The approach would have certain advantages not shared by other techniques. It is non-contact, there can be no argument that design secrets are being revealed - as is the case with x-rays - and background readings can cause problems with γ-ray detectors. Scale equipment can also suffer when being used for in-motion measurements.

The researchers point out that the idea that the mass of an object can be estimated directly from gravity gradiometer measurements is easy to understand. At sufficiently large distances, all objects can be approximated to point masses and various expressions exist to relate gravity to mass.

Of course things get a little trickier when measurements are taken closer in, since the objects are no longer point sources. Most of a recent paper (Estimating the weight of very heavy objects with a gravity gradiometer, "J. Phys. D: Appl. Phys, Vol 28 (1995) pp. 2378-2388) is taken up with explaining how to correct for the non-spherical nature of real world objects.

Obviously, much work has to be done to arrive at a reliable device giving reproducible results in a real situation. But in a world where scientists routinely talk in terms of the gravity exerted by a single grain of sand, such problems will surely not detain researchers for long.

Steven Gray, The Mitre Corporation, 202 Burlington Road, Bedford, MA 01730, USA.

Chemistry cracks open memory limits

Innovative but relatively minor changes in silicon-processing techniques could open up a three-fold reduction in the area required for capacitive components in microelectronic devices. Electronics experts are welcoming the advance as allowing further miniaturisation of memory devices which would otherwise soon hit capacitance limits using conventional methods.

Basis of the breakthrough, announced by R F Cava and colleagues at AT&T Bell Laboratories, is an increase in the dielectric constant of Ta2O5, tantalum oxide, by addition of 8% TiO2 titanium oxide ("Enhancement of the dielectric constant of Ta2O5 through substitution with TiO2", Nature, Vol 377, pp. 215-217).

Ta2O5 is known to form high-quality thin films, though its dielectric constant is low at around 35. Addition of TiO2 boosts that figure to 126. Ta2O5 and TiO2 are already well used and understood in electronics, unlike many of the more exotic compounds that have been suggested in the past to increase the dielectric constant. This allows the area of capacitors to be reduced.

The exact reason why the TiO2 boosts the dielectric constant of Ta2O5 is not yet wholly clear. So far TiO2 it seems to be unique in its effect. However the researchers say that at around the 8% TiO2 level, the new material is likely to be processable with very similar conditions to those currently employed to make pure Ta2O5 films.

As well as the area reduction, the workers say that the material could eliminate the need for complex three-dimensional capacitor geometries often resorted to, to yield acceptable capacitance in small-area components.

R F Cava, AT&T Bell Laboratories, 600 Mountain Avenue, PO Box 636, Murray Hill, New Jersey, 07974-0836, USA.
Putting electric cars under the microscope

Calls for ever-smaller electric vehicles have led to a breakthrough by Japanese workers at one of the largest suppliers of technology to the automotive industry worldwide. In shape it looks like an estate. It is light-weight, has a record breaking speed capability in its class, and you can even have it finished in pure gold. Only one drawback – the vehicle is only 7mm long.

In fact the microcar developed by researchers at Nippondenso Research Laboratories in Japan is currently the smallest wheel-driven mechanism in the world. Overall weight is 65mg and the car is reported to have recorded a top speed of 100mm/s.

Power for the car is through a miniaturised step motor built around an isotropic barium-ferrite magnet rotor that has been machined into a tube shape by a cylindrical grinder ("Performance of a 7mm microfabricated car, Akihiko Teshigahara et al, Journal of Micromechanical systems, Vol 4, No 2, pp. 76-80). The rotor's core is made into a four-pole magnet by placing it between four contact probes whose coils can generate a four-pole magnetic field. It is then joined directly to a zirconia front wheel shaft.

Chassis and wheels are made of stainless steel, and for the shell body, the main material used is a 30µm nickel film produced by plating onto an aluminium mould, etching away the aluminium, then protecting the nickel with gold. Microparts are so small that researchers had to resort to a mechanical manipulator usually found in bioresearch cell handling.

During testing, three running conditions were identified. In the low frequency range, alternation of the magnetic pole of the stator core is slow and the alternation interval long. So the rotor rotates step by step with each step being 90°. But at certain angles of rotation, motor torque increases too quickly and the wheels lose their grip, resulting in a stop-go movement. In the medium frequency range, movement is much smoother, with maximum speed being attained at 100Hz. But local variations in surface friction can cause some erratic movement – even backwards travel.

In the high frequency range, the rotor could not keep up with the changing field and simply rotated back and forth. However, surprisingly the effect produced the smoothest and most consistent forward movement. Researchers believe movement was due to a vibration of the stator caused by its asymmetry. Clearance between the wheel shaft and the chassis allowed this vibration to move the wheel into and out of contact with the surface, enabling a smooth and net movement in one direction.

A speed of 100mm/s might not seem sports performance, but it could be too fast for a 1/1000th scale car. Wear of the rotating parts is quite severe. Unfortunately, in tests, lubrication caused adhesion due to molecular force or surface tension. Plainly there is still work to do before a millimetre scale machine becomes useful in industrial and medical applications. And of course, the design does not really have any practical implications for real electric cars. Where on Earth are you going to be able to find atomic-scale furry dice?

Akihiko Teshigahara, Research Laboratories, Nippondenso Co Ltd, 500-1 Minamiyama Komenoki, Nisshin, Aichi 470-01, Japan.
Email: atesiga@rlab.denso.co.ip.

Plastic lasers breakthrough performance boundaries

Sealing laser modules in plastic can provide several operational advantages. But such modules have rarely demonstrated good operating characteristics and so far reliability has been uncertain. However, three researchers from NTT Opto-electronics Laboratories in Japan are claiming to have built a device that could make plastic moulding practicable, heralding the next stage in laser module fabrication technology for optical fibre transmission.

The lasers ("Pigtail type Laser Modules Entirely Moulded in Plastic," M Fukuda, Electronics Letters, Vol 31, No 20, pp. 1745-1747) are Fabry-Pérot type 1.3µm-band bulk or strained MQW BH devices. They have no facet coatings, are mounted on silicon heat sinks with a fibre guide. The heat sinks are then bonded on a TO18 stem usually.
Using infrared to see through the traffic jam

We can't - or don't want to - halt the unwavering progress of the automobile. But we can make a better job of keeping all that metal moving along the roads, if a simple system being tested in New York finds favour in the UK.

Good traffic control involves counting the number of vehicles, analysing their movements in real time and taking control measures, such as switching on green lights. Unfortunately magnetic-loop detectors are relatively inflexible and any modifications are going to mean digging up the carriageway - never the best way to improve traffic flow. They are also limited in their capabilities and can not be used to record speed and length data of individual vehicles, or for vehicle identification and tracking.

But infrared systems now being developed look as though they can do all that - and more. The first commercial system, produced by Eltec Instruments, has just been launched, and in an unrelated study recently, researcher Tarik Hussain and colleagues revealed the full potential of such systems to improve traffic flow ("Infrared pyroelectric sensor for detection of vehicular traffic using digital signal processing techniques", T M Hussain et al, IEEE Transactions on Vehicular Technology, Vol 44, No 3, pp. 683-689).

Pyroelectric systems rely on sensors mounted above a roadway to monitor traffic passing on individual lanes below. The detectors convert incident thermal infrared optical power into an electrical output signal.

Pyroelectric crystal sensors develop an electrical charge on their surfaces when incident radiation is absorbed by a coating on the crystal surface and converted to heat. The heat alters the lattice spacing of the crystal and causes a charge differential, measurable as an output voltage to sense the passing vehicles.

The signal, typically in the millivolt range, is then passed through a comparator block that uses mirror op amps to compare the received signal from the detector when a vehicle passes, to the steady-state condition. High cmrr op-amps - of the order of 60dB - allow the system to respond only to the difference signal.

For the 50-60m detection distances typically required, a gain of 40dB on the op-amp gives acceptable signal-to-noise ratios. The difference signal is fed to an LM2575 to generate a series of pulses which feed a counter.

In use, the pyroelectric detectors can be mounted on lamp standards or anywhere where they can have a line of sight to the vehicles. Counting output pulses is then a simple matter. But a second sensor along the road allows much more information to be obtained. The system can sense the duration of each signal in each sensor, and also the time between signals. So speed, acceleration and length of vehicle can all be calculated.

Different signal widths result from cars, vans and lorries in terms of signal widths and the field tests showed that weather conditions had little effect. The researchers say that the system could survive several hundred fold reduction in visibility to a level at which traffic probably couldn't circulate anyway.

Counting error was found to be less than one in 200 vehicles and was more accurate than magnetic loop sensors which overcount by detecting additional axles in the case of lorries.
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CIRCLE NO. 109 ON REPLY CARD
Ketil Parow details how to design a no-compromise, full-range tractrix horn loudspeaker featuring the combined benefits of high efficiency and crossover-free operation.

Why a horn? Much has been written about horn speakers over the years, and I am not going to repeat all of it here. I will however point out the obvious advantages, and my points of view. First, I will kill some of the myths about horn loudspeakers.

- Bass reproduction from a horn is not tinny. Some poorly-designed systems may have a tendency to produce 'empty' bass. In these cases, chances are that the designer compromised too much, foreshortened the horn in order to reduce the physical size. Poor wood quality may also have contributed to the poor quality.
- You do not need a vinyl record player and valve amplifiers to make a horn sound good. Advantages of a horn system are noticeable—regardless of equipment technology.
- Treble is not poor from a one-way horn system. There is a number of 'full-range' drivers available that are not truly range. This is not so with the Lowther drivers I used.
There is wide agreement on the following pros of horn loudspeakers.

- A horn system is far more efficient than an infinite baffle of bass reflex system. This offers not only more sound per input watt. For a given signal peak level, spl, the driver's excursion is accordingly smaller. As a result, the driver operates with more linear excursions so distortion is lower.

- Sound from a horn is more dynamic and more life-like than that from closed or vented boxes.

- In my experience, multi-way systems in vented or sealed enclosures can be seriously compromised by sloppy construction. A horn is less sensitive to materials and wood-working accuracy.

- Crossover networks are difficult to design, expensive, and the end results are unpredictable. A one-way horn removes these problems and produces a single sound source.

**Design goals**

Does the world need a new horn design? There is a number of well-designed horn systems in existence. However, on reviewing some of the existing designs, they did not appeal to me for various reasons. The modern designs seem to compromise too much in order to limit their physical dimensions. Older designs are no doubt good, but they look old fashioned. I want my speakers to look modern.

When I set out on this project, the only goal that was well-defined was that I wanted a back-loaded horn. Bass reproduction was to go well below 40Hz. Physical size was not an absolute restraint.

**Horn theory overview**

This section merely scratches the surface of horn theory. Those of you wanting to design your own horn are referred to the literature list. I will however provide all the theory relevant to the horn that I designed in enough detail to allow you to duplicate my work.

A horn is an acoustic transformer coupling the speaker cone movement to the air. It transforms the small volume of air moving with high energy into a low-energy, large movement of air. This is where the efficiency comes from.

The horn forms a funnel starting at the throat, closest to the driver and ending at the mouth. Cross-sectional area of the funnel expands along the length of the horn at a rate depending on the formula used. This is known as the contour.

Any given horn is only effective down to its cut-off frequency, and the chosen contour.

**Mouth dimensioning**

Mouth area is calculated from $f_c$. Circumference of a circular mouth should be

### List. Pseudo code for calculating the horn's tractrix contour.

:Sample code

Def var r as decimal . ;See the above formula
Def var x as decimal . ;See the above formula
Def var a as decimal . ;See the above formula
Def var r_throat as decimal . ;Radius at throat
Def var r_fe_mouth as decimal . ;Radius for a Fully Expanded mouth (free space loading)
Def var hl_fe_mouth as decimal . ;Horn Length when Full Expansion
Def var Size_factor as integer . ;See below
Def var step as integer initial 1 . ;See below
Def var oldx as decimal initial -1 .
Def var xfromthroat as decimal .
Def var prec as decimal initial 0.001 .

:End Sample code

### Performance

I don't have the tools to measure frequency response, but bass response seems reasonably flat in the deep end. I suspect, however, that there's a small dip in the response in the high bass region. It may be possible to reduce this with experimentation on cavity-damping/room-placement.

One of the most amazing things about these speakers is the sound output they are capable of. Out of interest, I tried powering them up using a Walkman, and they actually filled my basement - which is a big room. According to my measurements, 2-3W driving these speakers should suffice. At 1W, they are loud enough for most people to start holding their ears.

Lowther drivers are often criticised for having peaks in the high mids/low treble regions. I think the critics are, to some extent, right on this. There is a pronounced peak somewhere in that region, making some vocals/guitars sound a little harsh. There are some common tweaks to remedy this. Loosely stuffing some clean, long-hair wool between the whizzer-cone and the main cone is one of them.
equal to one wavelength of $f_c$, which gives, 

$$r_n = \frac{c}{2nf_c}$$

where, $c$ is the speed of sound, at 34290cm/s or 1125ft/s.

If you try inserting a sample value in the formula of, say, 30Hz, you will get an outrageous mouth radius of 182cm. This, however, is a mouth size calculated for free space loading. That means the dimensions are good if the horn is hanging from the roof, radiating into all eight quadrants of the space. For wall placement, the horn will be radiating into only two quadrants, so you can safely reduce the size by a factor of $\frac{1}{2}$, or four.

For corner placement, the situation is even better. The horn will be radiating into one quadrant only, and you can reduce the size by a factor of eight. In this example, that will bring the mouth down to $r_n/8$, which is 64cm. In a square horn, this corresponds to a square side of 114cm — which is still large.

It is possible to reduce the mouth size still further, even without compromising the overall design, as explained later.

This horn's $f_c$ was set at 32Hz, and calculated for corner placement.

**Throat dimensioning**

Theory here is somewhat more obscure. However, there are some valid thumb rules in calculating the throat area. In his 1974 article (see recommended reading) Dinsdale said, "For maximum bandwidth of a horn, one uses throat-to-driver ratios of 0.50 to 0.30; for maximum efficiency one uses ratios of 0.50 to 0.70."

Personally, I chose the throat area of my horn empirically. I browsed through 10-20 comparable horn designs, and decided on the basis of the designs most resembling mine. The driver I chose has an effective area of 211cm². This indicates that the throat area should be somewhere between 63cm² and 148cm². I decided on a throat area of 100cm². That gives a throat radius, $r$, of 5.65cm.

**Horn contouring**

Common contours for horn loudspeakers include conical, exponential, hyperbolic, and the tractrix. Of these, the conical is the easiest one to calculate and convert into a physical unit, but it is also the least efficient. Conical contours are never used for bass horns, because of the poor response and the impossibly long horns that result.

The exponential is the most common, and is easy to calculate. The hyperbolic contour is a variety of the exponential, and is the most efficient type. The trade-off is more distortion in the deep bass region. Hyperbolic horns are

### Table 1. Inside dimensions of the horn. Note that the mnemonics are referred to in Table 2.

<table>
<thead>
<tr>
<th>Dimension</th>
<th>Mnemonic</th>
<th>cm</th>
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<tr>
<td>Width</td>
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<td>Depth</td>
<td>ID</td>
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<tr>
<td>Height</td>
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### Table 2. Tractrix horn expansion from the throat

<table>
<thead>
<tr>
<th>Distance Radius</th>
<th>Square Area</th>
<th>Const. width</th>
<th>L. straight inch</th>
<th>Dist. w, inch</th>
<th>Const. w, inch</th>
<th>L. straight inch</th>
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</tbody>
</table>

Prototype enclosure design, cross section. Front and back panels are 37cm wide. Note these dimensions apply for 22mm mdf.
also somewhat longer than exponential horns. Combinations of all these types can also be found.

The tractrix is a curve well-known in the world of mechanics. In a stroke of genius, the late Paul Voigt applied the tractrix to horn speaker acoustics in 1926. The tractrix contour has characteristics similar to the exponential, but has the advantage of being shorter since the curve expands faster. The disadvantage is that it is somewhat awkward to calculate. This is because you cannot directly calculate the area \( A(x) \) at a distance \( x \) from the throat.

In this design, I decided to employ the tractrix contour to save space and wood.

Calculating the tractrix

The formula for calculating a tractrix contour is as follows,

\[
x = a \log(a + \sqrt{a^2 - r^2}) - \sqrt{a^2 - r^2}
\]

Where \( x \) is the distance from the mouth of the horn, \( a \) is the radius at the mouth and \( r \) is the radius at distance \( x \) from the mouth.

The tractrix curve can also be constructed using two straight rulers. However, if you have access to a computer and some form of computer algebra, the most convenient method is to write a small program to calculate the curve. Pseudo code outlining how this is done is presented in the List.

Coupling horn and driver

Before designing the enclosure, some attention needs to be directed to the volume of air - the cavity - between the driver's back side and the horn throat. It is best to keep higher frequencies out of the horn.

A good thumb rule is to let the horn handle 3-4 octaves or fewer. You should therefore dimension the cavity in such a way that it will act as a low-pass filter with an upper cut-off 3-4 times the horn's `true cut-off' frequency. Also, the upper cut-off should be set at a point where the horn's length equals an odd multiple of the wavelength. This is because the horn is loaded from the back of the driver, 180° out of phase. In this way, cancelling of the frequencies around the upper cut-off is avoided.

Calculate the volume of the cavity \( V_c \) using,

\[
V_c = \frac{C \times \text{throat area}}{(2 \pi \times \text{upper } f_c)^2}
\]

In my design, this was not considered. This is because I wanted to leave plenty of space available for experimenting with different upper cut-off frequencies and damping of the cavity. However, I advise would-be constructors to take advantage of this equation.

Folding a bass horn

This is by far the most difficult part of constructing a bass horn. Horn length should be measured along the middle of the duct. Through a bend, the length should be measured along the middle of the duct, all the way around the bend. That makes it a little difficult to construct the bend correctly, but by employing a pair of callipers and a ruler, it is possible. At first glance, the most convenient method appears to be to keep one wall of the horn straight, while expanding the horn with the other wall. This provides for a convenient way of making the cross-sectional areas match the distance. However, it also makes it harder to measure the distance, since the distance will be on an angled line.

The formula gives values for \( x \) measured along a straight centre line in the middle of the duct, while the radius - and calculated squares/rectangle heights - will expand in directions from the centre line.

I solved this trigonometrical problem with a spreadsheet. I inserted output from the tractrix contour program into the spreadsheet, and added a number of columns. One of the columns contains the length along the straight wall of the horn, corresponding to the length along the centre line, \( x \). This makes it a lot easier to draw the horn.

In this spreadsheet, the first four columns are created by the tractrix calculation program. These are, distance, radius, square and area. Height of the duct with a width of 37 is shown in the next column, entitled Const. width. In the L. straight column, length of the straight duct wall over the corresponding distance is shown. This has been found by applying simple trigonometrical functions to the angles given by the increase in height versus the length increase shown in the Distance column.

In the spreadsheet that I used, I also calculated the minimum space requirements for each row of the table. This gave a good indication of the final overall size, given the horn length.

And a time to compromise...

Comparing the spreadsheet to the drawing reveals that the mouth area of horn appears small. It has been foreshortened, i.e. terminated at a mouth smaller than the actually calculated mouth area.

An accepted method of foreshortening the horn is to base the mouth opening on a higher frequency than the \( f_c \) used to determine the flare rate. Normally, a horn's ‘true cut-off' will be at a frequency of around 1.25\( f_c \). This allows the horn to be terminated at an area of 7314cm². This corresponds to a duct height of 198cm in our 37cm constant-width horn.

Purists will tell you not to foreshorten the horn any more than this, and in some cases a lot more. The resulting design theoretically yields an uneven response versus frequency curve in the 1-2 octaves above the \( f_c \). However, so will any speaker located in any room. Depending on the acoustical characteristics of the speaker's surroundings, this natural reflection phenomenon may, or may not, be in excess of the irregularities of the horn.

After having done some thorough research on the subject, I have found a number of thumb rules for foreshortening horns. One of them is that the mouth area should be at least five times the area of the throat. This rule is probably best-employed for front-loaded than this - and in some cases a lot more. The resulting design theoretically yields an uneven response versus frequency curve in the 1-2 octaves above the \( f_c \). However, so will any speaker located in any room. Depending on the acoustical characteristics of the speaker's surroundings, this natural reflection phenomenon may, or may not, be in excess of the irregularities of the horn.

After having done some thorough research on the subject, I have found a number of thumb rules for foreshortening horns. One of them is that the mouth area should be at least five times the area of the throat. This rule is probably best-employed for front-loaded...
horns. Another is to cut the horn when the length is at a quarter of the theoretical $c$ wavelength.

My design’s foreshortening is well within these thumb rules. I cut the horn at an area of 91cm by 37cm, giving the horn a length of 288cm. Because of the mouth geometry of this design, the mouth will in fact be somewhat larger.

**Dimensioning**

Inner dimensions are listed in Table 1. Since there’s a number of options with regards to wood quality, thickness, etc., I have left final dimensioning to you. Using the original drawing, the outer board dimensions can be calculated like this, for two speakers.

I used 22mm medium-density fibre-board for the top/front and side panels, and 16mm MDF for the back/bottom and inner boards. My advice however is to use marine plywood, as thick as you can afford. Also, if I were to build these speakers again – which I am sure I will, incidentally – I would cut some boards for bracing the cabinet, between the top panel and the topmost inner board, between the front board and the middle (vertical) inner board, and between the side panels toward the mouth opening.

**Recommended reading**

- Dinsdale, J., Horn Loudspeaker design, Wireless World, Mar-June 1974. Note, there are errors in the tables of tractrix horn lengths.
- Edgar, BC, The Show Horn, Speaker Builder, 2/90.
- Edgar, BC, The Monolith Horn, Speaker Builder, 6/93.
- Edgar, BC, The Edgar midrange horn, Speaker Builder, 1/86.
- Edgar, BC, Solving the Klipschorn throat riddle, Speaker Builder 4/90.
- Edgar, BC, The Klipschorn throat revisited: Or, Ooops, Speaker Builder, 6/90.

Ketil Parow is a 30 year old programmer / software analyst at RADAR Software AS in Norway. His hobbies include loudspeaker design and building, wood-working and scuba diving. Ketil is also a part-time musician.
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CIRCLE NO. 112 ON REPLY CARD
Rod Green and Richard Hosking discuss how extending the polyphase direct conversion receiver concept produces a system with high performance, impressive selectivity and variable bandwidth.

There have been many direct conversion receiver designs published, many with acceptable performance despite their simplicity. Problems with simple designs are the ‘audio image’, and difficulty receiving modes other than cw and ssb.

The audio image – ie receiving signals on both sides of the local oscillator frequency – can be eliminated using phasing techniques. It is possible to achieve opposite side-band suppression with these circuits of 55-60dB. This level of suppression approaches that achievable with a crystal filter.

A block diagram of a polyphase dc receiver, Fig. 1, published in *EW* + *WW* March 1994, is shown in with a plot of the receiver selectivity in Fig. 2. You can see that as the receiver is tuned across the signal carrier frequency from one side-band to the other, the response drops very rapidly from 0 to –60dB.

'High-side' selectivity, illustrated in Fig. 2, depends on the low-pass filter following the audio phase network. In the case of the previous polyphase design a five-pole low-pass filter was used to give attenuation of approximately 30dB per octave above 3kHz. As a result the –60dB point occurs at about 12kHz from the carrier.

Obviously, better results could be obtained using higher order switched capacitor or digital filters, with response determined by the design of the filter and the number of poles. Alternatively, if the ‘crossover’ effect that occurs as the receiver is tuned across the carrier frequency could be utilised on the high side of the audio pass-band, the result would be a receiver of excellent selectivity.

By extending the polyphase direct conversion receiver concept it is possible to produce such a design – Fig. 3.

A different rf phasing network

The front end of the receiver, Fig. 4, is a direct conversion phasing design similar to the one described in *EW* + *WW* March 1994. The major difference in this part of the circuit is the rf phasing network. Instead of the bistable circuit, a two-pole rf polyphase network is used, see panel.

The first set of mixers translate from signal frequency down to an 'audio IF' at 0Hz and a wide-band polyphase network provides suppression of the opposite side-band, Fig. 5. Instead of combining the outputs of the polyphase network they are used as inputs to a second set of mixers at a second intermediate frequency of 461kHz, Fig. 6. The intermediate frequency is not critical and any convenient choice could be used.

As the second mixer local oscillators are in quadrature, this section is in effect an ssb generator at 461kHz. The 461kHz ssb signal is
amplified and passed through a band-pass filter before being applied to a third set of quadrature mixers, Fig. 7. The filter is not critical, but it is required to attenuate responses at harmonics of the intermediate frequency local oscillators.

For the third mixers, the local oscillator is offset from the second local oscillator by 1 to 10 kHz. Third mixer outputs are applied to a second audio polyphase network identical to the first, Fig. 5, to make this section in effect a second polyphase ssb receiver.

Phase relationships of all the signals are arranged so that the first section provides the selectivity on one side of the receiver passband while the third section provides the selectivity on the other side.

Variable offset between the second and third local oscillators allows a variable receiver bandwidth, i.e., frequency difference between the two oscillators. The AN4061 dual mixer is used for the second and third mixer pairs. This device is designed for use in colour tv cameras and is highly balanced and matched. In addition, it requires a minimum of external components.
Polyphase network principles

The principle of the polyphase network has been previously described. Briefly, if the network is fed with four signals in quadrature, they are cancelled in one direction – for example 0°, 90°, 180° and 270° – and passed in the other – 180°, 90°, 0° and 270°. More importantly, in this design, the signals are cancelled in the network. The outputs do not need to be combined to achieve cancellation and are available as quadrature pairs to feed into the second set of mixers. We used 9 pole audio networks to give an accurate quadrature response up to about 10kHz, Fig. 5 and the diagram below.

Network response was measured with a phase meter to be accurate to within 0.2° over the relevant pass-band. This accuracy is consistent with the measured system selectivity of approximately 55dB.

Note that networks have maximum attenuation within their quadrature pass-band. Above and below the quadrature pass-band, amplitude of the output products rises. A low pass RC section at about 15kHz was necessary after the first audio network to prevent overloading and cross-modulation in the second set of mixers from high frequency audio products. The network is relatively immune to component variation but it is still useful to choose matched capacitors and 1% resistors to achieve best performance.

And at rf?

The polyphase network can be made to work at rf up to about 30MHz. Above this, stray capacitance and phase errors in other parts of the circuit cause unacceptable phase error.

In the front end of this receiver, we used a two pole network to give good results over the 3.5 to 4MHz band,
Local oscillator quadrature signals for second and third mixers are generated using dual bistable devices in a ring configuration. Front end quadrature rf signals are generated using a broad-band rf polyphase network with values for the band of interest.

To minimise board space, surface-mount components can be used for polyphase networks. A notch filter is necessary in the audio output to eliminate a tone due to carrier feed-through in the second set of mixers. In this case we used a switched capacitor filter.

System blocks
Figure 8 gives a visual representation of the signals as they pass through various stages of the receiver.

Assume selectivity is set at 3kHz and that the incoming signal spectrum includes a dsb signal on 10MHz. Carriers C1-3, are on 9.990, 10.007 and 10.018 MHz, Fig. 8a). The first local oscillator, LO1, is set at 10.003MHz and the first set of mixers and phasing networks are arranged to receive the lower side-band.

Signals above 10.003MHz are attenuated by the phasing network. Note that this attenuation extends only to the limit of the phasing network response. In this case we used a polyphase network with poles extending to about 10kHz. As a result products from sig-
RF DESIGN

Fig. 8. Signal spectra at various points in the receiver.

Fig. 9. The receiver shows ‘brickwall’ selectivity on both sides of the passband.

and 15kHz respectively. Note that C2 has been eliminated. Recall that the polyphase network has output phases of 0°, 90°, 180° and 270°. A quadrature pair of these outputs is applied to the second set of mixers with phases arranged to produce lower side-band ssb with a local oscillator, LO2, at 461kHz. The IF spectrum includes the original dsb signal from 461 to 455kHz – inverted – and products from C1 and C3 at 448 and 446kHz respectively, Fig. 8c).

The third set of mixers is driven by local oscillator LO3, at 458kHz. This section is arranged to receive upper side-bands. Resultant output is demodulated audio from the upper side-band of the original dsb signal, Fig. 8d). The lower side-band and all the carriers have disappeared. Selectivity on both sides of the pass-band is sharp due to the crossover effect described above.

AM reception via direct conversion?

It is possible to receive amplitude modulation using the receiver. Assume that a 10MHz AM signal is being received and that selectivity is set to 10kHz. The first local oscillator would be set to 10.010MHz to give an audio output with the original AM upper side-band inverted from 0 to 10kHz. The AM carrier would be at 10kHz and the AM lower side-band would range from 10 to 20kHz.

Single side-band output from the second section would in effect be a reconstructed AM signal with the carrier at 451kHz. It is possible to detect this using envelope detection. In this case it is necessary to provide intermediate frequency selectivity at 451kHz as only signals between 10.010MHz and 10.020MHz would be attenuated by the first section.

In practice, heterodynes from other signals are a problem. Alternatively the AM signal could be limited and the third local oscillator phase locked to the carrier to give synchronous detection with good selectivity. In this case, audio output would represent the original amplitude modulation upper side-band from 0 to 10kHz.

The receiver’s performance

Front-end performance depends on the quality of the first local oscillator and first mixers. In this case, we used NE602 active mixers which have only average strong signal handling due to their rf gain and low power design. Maximum input level before limiting is about –25dBm. Dynamic range is about 90dB. This could be improved as in all receivers by using high-level mixers in the front end.

Selectivity is impressive, Fig. 9. These measurements were taken at 3.6MHz with the selectivity set at 3kHz. Reference point was 0dB at 1kHz from the carrier. ‘Lobes’ in the response within 300Hz of cutoff at each side of the pass-band were due to poor quadrature accuracy in the polyphase networks below 300Hz. This effect could be reduced by 300Hz high pass sections after the polyphase networks or poles at frequencies less than 300Hz in the polyphase networks.

There was a spurious response at –49dB at twice the selectivity frequency setting – in this case 6kHz. We assume that this is due to poor side-band suppression in the second set of mixers. It could almost certainly be improved with a second design.

Subjectively the filter improved intelligibility of signals over the original polyphase receiver alone.

Complex but advantageous

Though the receiver is somewhat complex, it offers several advantages over conventional superhet and dc receivers,

- Apart from responses within the audio pass-band – ie within about 10kHz of the signal frequency – any spurious responses are determined by first local oscillator quality and first mixer strong signal handling.
- ‘Brickwall’ selectivity can be achieved without crystal or mechanical filters and band-width is easily variable. This order of selectivity could be achieved using dsp techniques. However digital processing generally requires higher current drain which may not be ideal in portable equipment.
- Potentially, amplitude and frequency modulated signals can be received, making this a multimode design.
- Automatic gain control is easier to apply to the intermediate-frequency amplifier than to audio, as would be the case in a conventional direct conversion receiver.
- It would be interesting to measure group delay characteristics of the system though we have not done this. In theory the system should show a relatively linear characteristic over the whole pass-band.

Further reading

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CIRCLE NO. 115 ON REPLY CARD
In this extract from his book *Valve Amplifiers* Morgan Jones takes you through the steps of designing a valve power amplifier, and presents a prototype with a unique feature.

**A fresh look at valve power**

When designing your first valve amplifier, you need to be realistic. You are not going to design a world beating amplifier overnight, and by restricting your ambitions you stand a much better chance of making something that actually works.

The design example presented here is a 10W Class AB1 push-pull 'ultra-linear' amplifier using EL84 output valves. There are a number of reasons for this choice.

- It is cheap. If you have a 340V HT supply, this can be smoothed by 385V capacitors intended for switch-mode power supplies. In addition, the HT could be provided by a 240V isolating transformer with a silicon bridge rectifier. If any mistakes are going to be made, then it is best to make them with reasonably inexpensive components rather than expensive ones.

- There are many reasonably cheap second-hand amplifiers such as the Leak Stereo 20 or Leak TL12+, that can be cannibalised for their transformers.

- Powerful amplifiers require considerably more skill in layout and construction, and generate bigger bangs, so it is advisable for designers to start small.

**Bevois Valley amplifier**

This design acquired its name because the prototype was built from a pair of mono amplifiers bought for £15 – including pre-amplifier – in Bevois Valley. Once the output valves have been chosen, transformer configuration is limited, and therefore the entire output stage is fixed.

Transformer primary impedance needs to be around 8kΩ anode to anode, and with 43% taps for minimum distortion. This component might have been scavenged from a Leak, or it might even have been bought new. Either way, you will need an HT of 320V, and each valve will require 8V_{RMS} for full output. Our task is to design superior driving circuitry using the following specification.

- Low noise. With the low noise obtainable from CD or a good vinyl recording, noise in the power amplifier needs to be undetectable. A signal-to-noise ratio of 100dB, relative to full output power, is not an unreasonable figure to aim for. This rules out pentodes and high sensitivity.
No hum. This implies superb standards of construction, and/or dc heaters for the input stage.

Stability. To achieve good stability an absolute minimum of stages is needed.

Distortion. This is a tricky topic. If you want distortion measured in parts per million, then you had better buy a decent transistor amplifier. If you think that hearing is everything and measurement is nothing, then sell the house and buy a single-ended triode amplifier. We have to be honest about this. Valve amplifiers do not measure well, but they do sound good. Presumably, we listen to music to enjoy it, so this quality is important. However, I see no reason why we should tolerate obvious engineering faults. As a result, these will be removed — although this will not imply perfection.

Simplicity. Valve designs should be simple. Simple systems tend to have simple shortcomings. Additionally, they are repairable. Complex systems are built on silicon, have lots of legs, and are repeatable and disposable.

Together, these criteria demand that we use a concertina phase splitter direct coupled from the input stage without a driver stage, and we can instantly draw a circuit diagram. That this circuit is quite similar to the GEC912-Plus demonstrates that there is little new under the sun. The design rationale however is new, and to my knowledge, the cathode build-out resistor in the phase splitter is unique, Fig. 1.

Since the output valves are being driven directly from the phase splitter, linearity of the phase splitter is paramount. The chosen phase splitter only has a gain of 1, so the input stage will also need excellent-linearity.

Only three valves are really suitable for a concertina stage — the 6SN7, ECC82 and E88CC. We will use the E88CC.

Optimisation of dc conditions

Because the two stages are dc coupled, the design of the two stages will be interactive. As before, the way to deal with an awkward problem like this is to garner as many facts as possible, label the drawing, and see if anything useful appears.

Having chosen a concertina stage, we can start by labelling the anode and cathode loads as 22Ω. This traditional value is used because \( Z_{\text{out}} \) is approximately equal to \( R_L \), and while output resistance needs to be minimised, a significantly lower value would result in excessive power dissipation, Fig. 2.

Generally, with an anode voltage of 80 to 90V, linearity of the E88CC is best when the grid is at −2.5V. Although the concertina operates under heavy feedback, it would be preferable if it were linear before feedback. As a result, it is necessary to juggle conditions such that both valves are biased with \( V_{g} = −2.5\text{V} \).

Since the concertina has a gain of around unity, it might be possible to arrange component values such that the signal current drawn by the concertina is equal and opposite to the signal current drawn by the input stage. This would result in zero modulation of the HT supply, and would make the HT requirements less stringent.

After much drawing of loadlines, I found that all three requirements could be met simultaneously. Additionally, they met the previously unstated requirement of being achievable with the HT available.

Balancing signal currents is the easiest requirement to satisfy. If the concertina had an \( A_V \) of unity, then for equal and opposite currents we would use an anode load in the input stage equal to the sum of the anode and cathode loads of the concertina. Since the concertina has \( A_V \) of less than one, proportionately less signal current is swung into the input stage, which means a higher value of anode load.
For the concertina, \( R_t = R_c = 22 \text{k} \Omega \) obtains, and optimum biasing is achieved. Since the HT voltage is not known, gain has to be guessed. Fortunately, because the E88CC has such a low anode resistance, and the concertina has heavy feedback, it is possible to make a good guess at gain. Since the E88CC has a low anode resistance, gain for a given value of anode load does not change greatly with HT.

As a result, an HT of 300V can be guessed at, and a loadline can be plotted to determine the gain of an E88CC with a 44k\Omega anode load and -2.5V grid voltage. This results in an \( A_V \) of 28.75. The feedback equation can now be used to determine the gain when used as a concertina.

\[
A_{(\text{concertina})} = \frac{28.75}{1 + 28.75} = 0.966
\]

Even if had the approximation been based on \( \mu \), the error would only have been 0.3%, indicating that the guess should be quite accurate. From this result, the value of anode load for the input stage can be calculated from,

\[
R_{(\text{input})} = \frac{R_{\text{anode}} + R_{\text{cathode}}}{A_{(\text{concertina})}}
\]

Alternatively, \( R_t \) for the input stage can be set at 47k\Omega, and the concertina resistors reset to 22.7k\Omega. This is a more convenient choice since the 47k\Omega resistor will dissipate almost 1W, and so a 2W component is needed. This could be provided by 4x47k\Omega 0.6W devices in series/parallel, or by a single 2W component.

It is inconvenient to provide non-standard values in higher ratings, whereas the concertina resistors are only dissipating around 0.33W. This can be more easily met by standard resistors.

The closest approach to 22.7k\Omega is provided by 24k\Omega in parallel with 430k\Omega, but this means that the 24k\Omega resistor is dissipating almost all the heat, and a 0.6W component is marginal. You could use a 2W 24k\Ω resistor, but the tolerance of 2W resistors is usually marginal. A better solution is to use 36k\Ω in parallel with 62k\Ω, which is not such an accurate approach to 22.7k\Ω, but the resistors are closer tolerance. Also, the power is more evenly distributed between the components so that they are operated well within their ratings.

These choices of loads for the input/phase splitter stage will ensure equal and opposite signal currents, so we now need to arrange the correct biasing. The only way of doing this is by an iterative process.

Both stages will have an HT of less than or equal to 300V due to the voltage drop across the output stage. It is also known that each stage will have an anode voltage of 80 to 90V, for a -2.5V grid.

First draw the loadline for the concertina and find \( V_a \) for \( V_{ak} = -2.5V \). This value is then subtracted from the HT voltage to give the voltage across \( R_t \) and \( R_c \), and divided by 2 to give the voltage across \( R_t \). Voltage on the grid will be 2.5V lower than this, and will equal the anode voltage of \( V_1 \).

The next job is to draw a loadline for \( V_a \) to see if the optimum anode voltage corresponds with the voltage just derived. If it doesn’t, the only variable is HT voltage. Fortunately, a few iterations – by hand, not computer – found that a 285V HT voltage met all requirements, and this will be provided by a regulator.

I must say that the last determination was an incredibly tedious process. It was only carried out because in adjusting the biasing, it became obvious that it was also possible to fiddle both valves’ bias voltage into balance as well. A nice computer model using real valve characteristics would solve this problem in considerably less time.

Now that the HT voltage for the two stages is known, all the ac parameters can be calculated, and the value of the build-out resistor for the concertina determined.

Cathode bias and feedback resistance

This is easily the most complex calculation in the design of a power amplifier with negative feedback applied to the cathode of the input stage. These four factors are significant:

- Cathode bias voltage needs to be set correctly. This would normally be a trivial application of Ohm’s law, but in this case the bias current flows through the cathode resistor and the feedback resistor.
- The input valve generates a feedback current through the cathode resistor, in addition to any current sourced from the output of the amplifier.
- Ratio of the two resistors needs to be set so as to obtain the desired negative feedback.
- As far as ac is concerned, the cathode resistor is shunted by \( r_t \) of the valve.

Now, with the restrictions specified, it should be possible to label a diagram and derive some equations. Since 2.5V bias on the cathode is needed, and anode current is 190V/47k\Omega, the total resistance to ground from the cathode must be 618.4\Ω.

Anode signal swing for full output is 8.636V rms. This means that the anode signal current must be 8.636V/47k\Omega=0.1837mA rms. This current also flows in the cathode circuit and will develop a feedback voltage across any unbypassed cathode resistor.

If input sensitivity of the amplifier is to equal 2V rms, and we know that the unmodiﬁed sensitivity is 298mV rms, the feedback voltage required at the cathode will be 2-0.298V, which is 1.702V rms.

For full output of 10W, the signal at the output of the amplifier will be 8.944V rms. This means that there will be 7.242V rms across the cathode resistor.

Using this equation, you will find that \( r_t = 1.559 \text{m} \Ω \).

Assume that the output of the amplifier is a true Thévenin source driving the network through the feedback resistor ‘y’. The valve’s own feedback current is represented as a Norton current source, and the cathode resistor ‘x’ is shunted with \( r_t \). Note that Fig. 3 is an ac diagram.

Our first observation is that there is a resistor of known value \( r_t \) with a known voltage of 1.702V across it, so current through it is 1.091mA.
We can now see that node 1 has two known currents flowing through it, so we can find the third, using Kirchhoff. If there is 0.1837mA flowing into the node, but 1.091mA leaving it, then 0.9073mA must be supplied by the other node. Moving to node 2, you can see that any current coming into the node must be supplied by $I_x$ and that this splits through the resistor $x$, and to node 1. Formalising this:

$$I_x + 0.9073 = I_o \quad (Eq. 1)$$

You can use Ohm's law to make statements about the currents in resistors 'x' and 'y':

$$I_x = \frac{1.702}{x} \quad (Eq. 2) \quad I_y = \frac{7.242}{y} \quad (Eq. 3)$$

The final restriction is the dc restriction, which says that $x$ and $y$ in parallel must give 0.6184kΩ.

$$0.6184 = \frac{xy}{x + y} \quad (Eq. 4)$$

The way to solve the equations is to substitute the second and third equations into the first:

$$1.702 + 0.9073 = \frac{7.242}{x}$$

Rearranging, and simplifying:

$$7.982x - 1.876y = xy.$$

It is now possible to substitute this into the fourth equation, and solve it to give the ratio $y = 2.953x$. Substituting this ratio back into the equation yields $x = 82852$. Using the ratio you therefore need 1.2kΩ in parallel with 2.7kΩ for the cathode resistor, and 4.7kΩ in parallel with 5.1kΩ for the feedback resistor.

### Distortion consideration

Some of V1’s cathode current is now flowing through the output transformer, and it might be thought that this would cause distortion. Assuming that dc resistance of the transformer secondary winding is negligible, the current flow will be:

$$2.5V/2.44kΩ = 1mA.$$  

Now the current turns ratio of the transformer is 31.6:1 secondary-to-primary – so 1mA of dc flowing in the secondary is equivalent to 31µA out-of-balance dc flowing in the primary. Compared to 40mA each side, this is negligible, since output valve balance is highly unlikely to be as good as this.

All component values for the driving circuitry are now known, so the values for the output stage can be determined. The EL84 is allowed a maximum grid leak resistor of 300kΩ with grid bias. As cathode bias is being used however, this can be increased to 470kΩ. A 0.1µF coupling capacitor is necessary, which should be polycarbonate, or preferably polypropylene, with a rating of 400Vdc or more.

A value of 4.7kΩ is traditional for grid stopper resistors on the EL84. They may not be needed, but it seems sensible to fit them just in case. A resistance of 47Ω in series with $g_2$ is alleged to reduce distortion while reducing peak power. I have not tested this, so fitting them is a matter of personal choice. The Mullard circuits included them, but the Leaks didn’t.

From the data sheet, the cathode bias resistor should be 2700 and dissipate 0.45W. Resistors rated at 2W are commonly used here, but a 15W chassis mounting metal clad type with tabs is a much better choice. This is because an electrolytic capacitor is going to be placed very close to this resistor, so it needs to be kept cool. The resistor also provides convenient tags for anchoring the capacitor.

The cathode bypass capacitor should be 2200pF for a 1Hz cut-off. But as discussed earlier, this value would cause additional problems; a good compromise is 470µF at 63V. A rating of 63V may seem excessive, since it will only see around 11V, but the higher voltage component will have a lower effective
series resistance. This becomes significant when you are trying to bypass the 90Ω cathode resistance of the valve.

Because there is only one RC network plus the output transformer in the entire amplifier, low-frequency stability will not be a problem. High frequency stability is not assured, and so this should be investigated.

The input stage has its basic sensitivity reduced from 298mV to 2V, which corresponds to a gain reduction of 6.71. From this, you can calculate the new $r_a$ for the stage:

$$\frac{\mu R_L}{R_t + r_a} = 6.71 \frac{1}{R_t + r_a'}$$

Solving this, and using $r_2=5kΩ$, gives $r_a'=502kΩ$, in parallel with $R_L=47kΩ$, this gives $r_{out}=4kΩ$. You will find that applying global negative feedback invariably causes $r_{ext}=R_L$ for the input stage.

The concertina has 3.2pF of Miller capacitance. Allowing for strays, 5pF is a reasonable total value. In combination with 4kΩ, this gives a cut-off around 780kHz. The output stage will have an input capacitance that loads the 22kΩ output resistance of the concertina, so should be decreased. Although the EL84 is a pentode, it will still have Miller capacitance, albeit greatly reduced, so this should be included in the calculation.

You can find anode gain of the output stage by calculating voltage across the 8kΩ transformer primary for 11W. It is known that 16V RMS from grid to grid is needed to drive the stage. This gives a gain to the anode of 18.54. Since $C_a=0.5pF$, this would result in a Miller capacitance of 9.8pF. Unfortunately, this value of $C_a'$ is for the pure pentode connection. On the other hand, we will be using the 'ultra-linear' connection, where $g_2$ does not stay at a constant potential. This means that allowance must be made for the Miller effect from $C_g$. Unfortunately, the Mullard data sheet does not give a value for this, so it is probably wise to allow another 10pF.

Adding these to $C_a=10.8pF$ produces a total input capacitance of around 35pF, including strays. Driven by the concertina, this gives a cut-off of 200kHz, and is the dominant pole. To achieve high-frequency stability, slug the input capacitance of the output valves, and not the concertina, as is usually done. This will have the advantage that additional capacitance will swamp variations in the capacitance between valves, improving high-frequency balance. Shunt capacitors of 68pF across the EL84 grid leak resistors will slug this pole to 72kHz. It is now possible to draw a full circuit diagram of the amplifier, with component values, Fig. 4.

Further reading

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DIY

Circuit Analysis

Using brief Basic routines, John Hopkins demonstrates how you can analyse simple circuit networks on the PC - and in doing so, provides an insight into how Spice functions.

If you have used an analogue simulator, such as Spice, you may have wondered how all those intricate calculations are carried out.

If you have a version of Basic - such as the one included with MS-DOS - you can gain an insight into Spice by trying out the routines described here. What follows is not intended to be a replacement for a proper simulator, but rather an exercise in trying to understand how network analysis is carried out.

Consider solving a linear network consisting of resistors, capacitors, inductors and active devices driven from an appropriate signal source. Transistors - and other active devices - are non-linear, but by considering small signal operation at a suitable dc working point, you can linearise the problem.

Figure 1 shows a network involving only resistors and a dc source. The usual way of solving this without computer aid is to use Kirchhoff’s laws and set up a system of equations. For our purposes, it is best to use the nodal analysis method.

You begin by numbering the nodes, making the earth node zero and then allocating the numbers 1, 2, 3... as far as necessary, as in Fig. 1. Analysis is then conducted in terms of voltages at each node, by writing a statement of Kirchhoff’s current law at each node.

The sum of the currents flowing through the resistors, worked out from the node voltages and Ohm’s law, is equal to the current from the independent source connected to that node.

At node 1: \( V(1) + V(2)/10 + V(3)/100 = 1 \)

At node 2: \( -V(2)/10 + V(3)/10 = 0 \)

At node 3: \( -V(1)/100 - V(2)/10 - V(3)/1 = 0 \)

Equations can be tidied up so that terms involving a given node voltage are grouped together.

Node 1: \( 1 + 1/10 + 1/100 \cdot V(1) - V(2)/10 - V(3)/100 = 1 \)

Node 2: \( -V(2)/10 + 1/10 + 1/100 \cdot V(3)/10 = 0 \)

Node 3: \( -V(1)/100 - V(2)/10 + 1/10 + 1/100 \cdot V(3)/1 = 0 \)

There are some simple rules for constructing these equations.

- At each node, a term is formed by summing the reciprocals of the resistances attached to the node, multiplied by the voltage at the node.

- Remaining terms on the left-hand side of an equation are formed by considering the resistances connected to each of the other nodes in the circuit.

  These terms are all negative and consist of the other node voltages divided by the connecting resistances.

- The right-hand side is the independent current source value.

The equations can be written again with the numerical values worked out,

\[
\begin{align*}
1.11V(1) - 0.1V(2) - 0.01V(3) &= 1 \\
-0.1V(1) + 0.21V(2) - 0.1V(3) &= 0 \\
-0.01V(1) - 0.1V(2) + 1.11V(3) &= 0
\end{align*}
\]

An elegant way to solve the equations is by the use of Gaussian elimination.

If you take equation [1] and multiply each term by 0.1/1.11 and then add it to [2], the term involving \( V(1) \) in the new version of [2]...
Each node voltage can be calculated using back substitution. From [3]:

\[
V(3) = \frac{-0.101V(2) + 6.101 \times 5.12E-2}{0.201} = 0.474
\]

Putting this value into [2], V(2) is found to be:

\[
V(2) = \frac{9.01E-2 + 6.101 \times 5.12E-2}{0.201} = 0.474
\]

And using these values in [1],

\[
V(1) = \frac{0.01 \times 0.474 + 0.01 \times a(2,2) \times a(2,3) \times y(2) \times 1.11}{0.201} = 0.944
\]

This is the solution to the problem. Of course, this method can be employed in more complicated situations. Although Gaussian elimination is not the preferred method for solving the equations, it is simple and serves as an introduction to more advanced techniques.

**Gaussian elimination.**

When written in Qbasic with operations applied to the coefficients of each term in the equations, the algorithm is reasonably compact. You can therefore draw up an array (a matrix) of these coefficients, and then work out the procedure from that. A set of three equations, written with symbols for the coefficients, would look like,

\[
a(1,1) V(1) + a(1,2) V(2) + a(1,3) V(3) = a(1,4)
\]

\[
a(2,1) V(1) + a(2,2) V(2) + a(2,3) V(3) = a(2,4)
\]

\[
a(3,1) V(1) + a(3,2) V(2) + a(3,3) V(3) = a(3,4)
\]

So the array we are using is:

\[
\begin{align*}
a(1,1) & \quad a(1,2) & \quad a(1,3) & \quad a(1,4) \\
a(2,1) & \quad a(2,2) & \quad a(2,3) & \quad a(2,4) \\
a(3,1) & \quad a(3,2) & \quad a(3,3) & \quad a(3,4)
\end{align*}
\]

All coefficients are taken as positive at this stage, although for a real problem some numbers would be negative, as we have seen. To eliminate V(1) from [2] multiply [1] by a(2,1)/a(1,1) and then subtract the new first equation from [2]. To eliminate V(1) from [3] multiply [1] by a(3,1)/a(1,1) and then subtract the new first equation from [3]. At the second stage, to eliminate V(2) from [3], multiply the new second equation by the new a(3,2)/a(2,2) and then subtract it from the new equation [3].

Difficulties caused by differences between the old and new equations can be neatly avoided by simply storing new coefficient values in the array which held the old values. This is easy to do and has the advantage that it is economical on memory space. As a result, you do not need to discriminate between old and new values of coefficients, and the reduced array or matrix now looks like,

\[
\begin{align*}
a(1,1) & \quad a(1,2) & \quad a(1,3) & \quad a(1,4) \\
a(2,2) & \quad a(2,3) & \quad a(2,4) \\
a(3,3) & \quad a(3,4)
\end{align*}
\]

Referring to Fig. 2, you will see that it falls into four parts. Firstly, there is a straightforward section which defines a number of integer variables and dimensions arrays a(n,n+1) and V(N). The program then asks for values of the coefficients to be used in the calculation. Secondly, sections two and three headed ‘reduction of matrix’ and ‘back substitution’ do the calculations. These have been developed from a flow chart by Dorn and McCracken.1

If, like me, you need more explanation of computer programs, then the following comments may help. Think in definite terms about the procedure.

To start with, assume that the number of equations N=3. Integer K refers to the number of the equation which is to be multiplied and then subtracted from the others, so it will start at 1 and go to 2 to complete the procedure.

The general array element is a(I,J), with the first integer (I) being thought of as the row number and the second (J) as the column number.

Element a(K,K) is the one which divides the Kth equation, and is called the pivot. So the I loop runs from the next equation (starting at number 2). This is used to work out the multiplier M and set elements which are going to be eliminated equal to zero (there is no need to work them out). It is also used to multiply...
DECLARE SUB Equations ()

'Type a list of DATA statements at the beginning, as follows:

REM DC network of Fig.1
DATA 7.3
DATA CS,1,0,1
DATA R,1,0,1
DATA R,1,2,10
DATA R,2,0,100
DATA R,3,2,10
DATA R,3,0,1
DATA R,1,3,100

DECLARE SUB Equations ()

FOR I = 1 TO N: FOR J = 1 TO N
REM Set up matrix
G(J, J) = G(J, J) - 1/R: G(K, J) = G(K, J) - 1/R
G(J, K) = G(J, K) - 1/R: G(K, J) = G(K, J) - 1/R
NEXT J: NEXT I

REM Back substitution.
V(N) = a(N, N+1) / a(N, N)
FOR I = N-1 TO 1 STEP -1
V(I) = (a(I, N+1) - S) / a(I, I)
NEXT I
END SUB

Fig. 3a) QBASIC program to solve a dc network.

DEF I-N
READ Numele, N 'No. elements, nodes.
DIM A(N, N+1), V(N)
DIM G(N, N), CS(N)
FOR I = 1 TO Numele
READ Numele, N
DEFINT I -K, N
DEFINT I -K
CASE IS = "R"
DATA R,1,3,100
DATA R,2,0,100
DATA R,1,2,10
CASE IS = "S"
READ J, K, Csource: CS(J) = Csource: CS(K) = -Csource
READ J, K, R
G(J, J) = G(J, J) - 1/R: G(K, J) = G(K, J) - 1/R
G(J, K) = G(J, K) - 1/R: G(K, J) = G(K, J) - 1/R
CASE IS = "CS"
READ J, K, Csource: CS(J) = Csource: CS(K) = -Csource
END SELECT
NEXT I
REM set up matrix
FOR I = 1 TO N: FOR J = 1 TO N
a(I, J) = a(I, J) + a(J, I) / R: a(K, J) = a(K, J) + a(J, K) / R
NEXT J: NEXT I

PRINT "Solution"
PRINT USING "## "; a(I, J)
NEXT J: NEXT I
END

Fig. 3b) Sub program for solving equations.

REM Reduction of matrix, using Gaussian elimination.
FOR K = 1 TO N-1
FOR I = K+1 TO N
M = a(I, K) / a(K, K)
FOR J = K+1 TO N+1
a(I, J) = a(I, J) - M * a(K, J)
NEXT J
NEXT I

Next the program has a main section to print the matrix, the right-hand side and divided by the coefficient of the variable we need.

General comments
You should find that the program will work for any reasonable number of equations N. One of the problems that you can encounter if you simply invent equations without reference to a network is that some of the pivot values can be zero, in other words a(K,K)=0. This will not happen if you use a network as the source of the problem, provided that you always allocate an earth node and label it zero and number the other nodes in sequence 1,2,3... None of the values should be left out.

Creating a netlist
The usual way of getting circuit configurations into a package such as Spice is by means of a list of components and their node numbers. To deal with the problem in hand it is necessary to specify resistors and current sources. These methods are described fully in Vlach and Singhal2.

RESistors require their conductance values to be stored in an array called G(J,K), the J,K values being the node numbers to which each resistor is connected. Since the problem is going to involve N equations, the highest node number will be N, so the greatest values of J and K will be N. You therefore dimension the array: DIM G(N,N)

Assuming that J.K and the resistor value R have been read into the program, you can now create an entry into G as follows,

G(J)=G(J)+1/R: G(K)=G(K)+1/R
G(J,K)=G(J,K)+1/R
G(K,J)=G(K,J)+1/R

To understand this, consider the 100Ω resistor connected between the nodes 1 and 3 of Fig. 1. According to the assignments above, provided that all elements of G are zero at the outset, then the effect is to make,

G(1,1)=0: G(3,3)=0
G(1,3)=0: G(3,1)=0.

If these are interpreted as the elements of the matrix, i.e a(1,1)=G(1,1) etc, then you have the appropriate values for the coefficients on the left-hand side of the equations. When these instructions are used repeatedly, then the effect of the G(J,K) terms on the right-hand side is to add the conductances of any resistors which terminate on the given pair of nodes.

If a resistor is connected with one end to earth, say K=0, then all but G(J,K) will have no effect. This is because the corresponding matrix elements are not used in the Gaussian elimination.

Now for the current sources. Here we create an array CS(N), and assign values to it as follows,

CS(J) = Source : CS(K) = -Source

where J is the node connected to the positive end of the source, and K is the negative end. We enter these values into column N+1 of the matrix. Figure 3 reads in a netlist and then solves the resulting equations.

As you see, the program has a main section and a sub program to solve the linear equations. The only other thing you might need to do to convince yourself that the program is operating correctly is to add a short program segment to print the matrix, Fig. 4.

References

Further reading
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A.D. Ryder, in his letter 'Sallen & Key Distortion', November 1995, discusses a distortion mechanism often seen when op-amps are used with divergent inverting and non-inverting input source impedances and applied common-mode voltages. A Sallen & Key filter represents an extreme case of this, with medium-high impedance at the non-inverting input and - usually - zero impedance at the inverting input, as in his example. However, the distortion can also be seen in other non-inverting amplifier topologies when working from medium to high impedances such that the source/feedback impedances are not balanced.

The basic distortion culprit is a non-linear capacitance/voltage characteristic at the amplifier input(s), i.e., a varactor-like behaviour. This produces the distortion when appreciable common-mode voltage is applied. For a given signal input, the distortion generated is proportional to the source impedance, that is, higher impedances will produce more distortion from a given amplifier.

As Mr. Ryder notes, operating the op-amp in the alternate inverting mode avoids this distortion. This is because the non-linear C-V mechanism is not exercised for this condition.

Several means can be used to control this distortion in Sallen & Key filters. One is to simply scale the filter impedances to the lowest level possible. This reduces the distortion by simple brute force means. Use, for example, two 11kΩ resistors and 4.7nF/47nF capacitors, etc.

A more complete approach is to provide direct compensation for the distortion, by taking advantage of the intrinsic amplifier input characteristics to break the inherently differential-mode input devices; as such they produce this distortion at both inputs. It then follows that matching the impedances of the two inputs will provide a compensating delta-V/delta-C distortion at the feedback input. This conveniently causes the distortion at the output to be minimised by virtue of the amplifier's natural common-mode rejection.

A simple first-order RC pair in the feedback path provides some compensation, but full distortion reduction is seen when the input RC components seen at the non-inverting input are duplicated one for one in the feedback path. For Mr. Ryder's example, this would be accomplished by synthesising an equivalent two-terminal impedance network, composed of, at the top, a 470pF capacitor in parallel with two series 11kΩ resistors. The lower of these is shunted by a 4.7nF capacitor.

The bottom of this network is returned to the amplifier output, completing the filter. In this setup, the amplifier sees identical impedances at the two inputs for all frequencies, the desired condition for lowest distortion due to common-mode inputs.

This topic has been discussed previously in (1) and (2) for filter applications, and in (3) and (4) for straightforward amplifiers. Data contained in (2) shows an order of magnitude or more reduction in thd with the use of the compensating network in a Sallen & Key filter. Reference (4) also illustrates another method of beating the problem, by bootstrapping the input stage of a susceptible amplifier.

Walt Jung
Analog Devices Inc.
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(fax) 410 692-2152
Email: Walt.jung@analog.com

References
2. Walt Jung, 'Active Filter Circuit Subcircuits', Analog Devices OP176 data sheet.

Anyone remember this?

In 1914 the Marconi Wireless Telegraph Company, acting for the Admiralty, built a 100kW spark transmitting station on Ascension Island. This, by any standards, was a remarkable task. Apart from being one of the most remote in the world, this island had virtually no infrastructure, and the sole population comprised a garrison of naval staff and their servants. Every item required in the project had to be brought in. In addition to the transmitter/receiver site, with its six, 300ft tubular steel masts supporting a massive "T" antenna, the project also required the construction of complete power station, delivering 500V dc for the transmitter and a separate 220V dc supply for lighting and staff accommodations.

Incredibly, the entire job was completed within three months. My thanks, to the help and generosity of the Marconi Company, has a good deal of information concerning the description and specification of this installation. Sadly we have very little information concerning its operational life. My appeal to any readers who may have any such information, or even photographs, to contact me at the following address: Ascension Heritage Society, c/o BBC Atlantic Relay Station, Ascension Island, South Atlantic Ocean. Phil Brooks Secretary, AHS Ascension Island.

EMC clarified

Mr. Bore's argument, expressed in last December's Letters, is valid to some extent, but his description of the situation is inaccurate. He can be forgiven for being confused, because there is a positively scandalous amount of wrong information - even disinformation - rife in the industry. The authorities seem powerless to control matters.

Radiated emissions are by no means the only disturbances which are controlled by limits in standards. The EMC Directive does not require any testing at all to be undertaken: all it requires is that "apparatus" does not cause interference, and is not unduly sensitive to legitimate or permissible electromagnetic disturbances that can be expected to occur in the environment(s) in which the manufacturer intends or expects the apparatus to be used.

In most cases, the manufacturer is required to make a legally-binding declaration to this effect. This declaration can, for most types of apparatus, be supported in either of two ways: conformity to the appropriate standards which have been notified as acceptable to the Commission by publication in the EU Official Journal, or by the 'Technical File' route, known as TCF. With the TCF route, standards may be used in part, but essentially, it requires a technical report prepared or endorsed by a "Competent Body". This body may be an officially recognised test laboratory.

In spite of a concerted campaign to enable the UK to migrate to the TCF route (to support the high proportion of Competent Bodies in the UK - far higher than in other countries) the standards route is, for most products, simpler, cheaper and more certain. It is essential however, that the appropriate standard(s) are applied, and this is a particular source of confusion. For products that have no particular appropriate standards ('product standards' or 'product-familly standards'), there are Generic Standards EN50081-1 and -2 and EN50082-1 and -2.

These standards make provision for exemptions from testing where there is clearly no call for it - see for example, Clause 8 of EN50081-1. Some product standards are rather elderly, and do not contain such a provision explicitly.

However, provided the decision not to test is recorded with reasons, and preserved as required by the Directive, a statutory defence of 'due diligence' is provided in, for example, the UK Regulations implementing the Directive. Mr. Bore's example figures, however, are not realistic. For most types of product, radiated emissions are controlled above 30MHz. Below this frequency, the efficiency of most products as antennas is low because their dimensions are small compared with the wavelength. But at 30MHz and higher frequencies, quite a small amount of antenna power is required to produce, for example, the limit field strength of 30dB(uV/m) (31.6uV/m) at 10m specified in EN50081-1. If the equipment were as efficient a radiator as a half-wave dipole, the requisite input power to the dipole would be given approximately by: (31.6x10^-6) x (30/10) x (3.14x10^6) /1.8 = 1.84mW

KRS John Woodgate
(Chairman, British Standards Technical Committee EPL/100)
Improved hot audio

Having analysed Jeff Macaulay's valve power amplifier featured in the October 1995 issue of Electronics World, I have found an imbalance in the circuit that the following equations highlight.

In the equations,

- \( u \) is \( V_{i}\)
- \( i \) is change in \( Tr_1 \) emitter current from its quiescent value
- \( l \) is change in \( Tr_2 \) emitter current from its quiescent value
- \( m \) is the value of \( R_3 \)
- \( M \) is the value of \( R_4 \)
- \( p \) is potential at \( Tr_1 \) emitter
- \( P \) is potential at \( Tr_2 \) emitter
- \( E \) is potential of the negative rail
- \( f \) is the value of \( R_5 \)
- \( F \) is the value of \( R_8 \)
- \( E \) is the value of \( R_{18} \)
- \( h \) is the value of \( R_9 \)
- \( k \) is the value of \( R_17 \)
- \( m \) is the value of \( R_3 \)
- \( i \) is change in \( Tr_1 \) emitter current from its quiescent value
- \( u \) is \( V_{i}\)

Emitter current of \( Tr_1 \) is,

\[
\frac{p - 0 + p - u}{m + f}
\]

which in the quiescent state is,

\[
E \frac{1}{h} \left( \frac{1}{m + f} \right)
\]

and similarly the quiescent emitter current of \( Tr_2 \) is,

\[
E \frac{1}{B} \left( \frac{1}{M + F} \right)
\]

When \( V_{i} \) is instantaneously at some value \( u \) the emitter current of \( Tr_1 \) is,

\[
\left[ \frac{u + u - E - u}{f + b + b - h} \right] - \frac{1}{m} \left[ \frac{u + u - E - u}{f + b + b - h} \right] - \frac{1}{f}
\]

which simplifies to,

\[
\frac{Ef + u + f}{m + f + 1 + \frac{1}{b} + \frac{1}{h}} \frac{E}{m + f + 1 + \frac{1}{b} + \frac{1}{h}}
\]

and so the change in \( Tr_1 \) emitter current from its quiescent value is,

\[
\frac{u - f}{m + f + 1 + \frac{1}{b} + \frac{1}{h}}
\]

and so we have,

\[
i = \frac{u - f}{m + f + 1 + \frac{1}{b} + \frac{1}{h}}
\]

Similarly the instantaneous emitter current of \( Tr_2 \) is,

\[
\left[ \frac{E \frac{u - E}{B - h} - 0}{M} \right] \left[ E \frac{u - E}{B - h} - 0 \right] \frac{1}{F}
\]

which simplifies to,

\[
\frac{EF}{BM} + \frac{EF}{BM} + \frac{EF}{BM} + \frac{1}{h}
\]

and so the change in \( Tr_2 \) emitter current from its quiescent value is,

\[
i = \frac{EF}{BM} + \frac{1}{h}
\]

Applying the resistor values specified in the circuit (working in mA, volts and kΩ) the change in \( Tr_1 \) emitter current is,

\[
\frac{1}{1.8} \cdot \frac{1}{1.8} = 10.6u
\]

Similarly it can be shown that the change in \( Tr_2 \) emitter current is \(-10.6u\).

I have not yet built this circuit but I would expect it to perform better than the original. Also its component count is lower.

Finally, I am a little worried about the voltage rating of \( C_5 \) and \( C_6 \). Allowing a 2V drop across the bridge rectifier, the peak on these capacitors will be 394V. This is perilously close to the specified voltage rating of 400V. If the mains input voltage were to rise to 244V there might be an expensive bang.

DC Haigh
Winchester

The value of \( R_{18} \) has been chosen to give the same quiescent current as the original circuit.

There are two ways of looking at the operation of the phase splitting arrangement.

The potential at the junction of the equal-valued resistors \( R_3 \) and \( R_4 \) will be the average of the potentials at the transistor emitters so if phase splitting is perfectly balanced this potential will remain at 0V. By connecting this point to the inverting input of \( A_2 \), \( A_2 \) will control the potential at the emitter of \( Tr_2 \) so that the potential at the junction of \( R_3 \) and \( R_4 \) is as close to 0V as its raw gain will permit. Hence the phase splitting is as well-balanced as possible.

The other way of looking at the circuit is to consider \( R_{17}, R_5, A_2 \) and \( Tr_2 \) to be a conventional amplifier with feedback which takes its input from the emitter of \( Tr_1 \). The gain of the amplifier is -1 so it straightforwardly inverts the potential at its input and so provides the opposite phase.

\( A_1 \) has 100% negative feedback applied so the potential of the emitter of \( Tr_1 \) will equal \( V_{i} \). and that of \( Tr_2 \) will be \(-V_{i}\).

In this circuit, let \( r = \frac{V_i}{R_1}, r = \frac{V_i}{R_2}, r = \frac{V_i}{R_2}, r = \frac{V_i}{R_2} \) and \( s = \frac{V_i}{R_2} \). Quiescent emitter current of both \( Tr_1 \) and \( Tr_2 \) will be half that flowing through \( R_{18} \), i.e.

\[
1 = \frac{E}{2} \left( s + t \right)
\]

or

\[
\frac{E}{2s + t}
\]

Potential at the junction of \( R_3, R_4 \) and \( R_8 \) will be

\[
\frac{E}{2s + t}
\]

Hence when \( V_{i} = u \) and the emitter of \( Tr_1 \) is at potential \( u \) the emitter current of \( Tr_1 \) will be,

\[
u + \frac{E}{t} \left( 2s + t \right)
\]

which is,

\[
u + \frac{E}{t} \left( 2s + t \right) + \frac{u}{r}
\]

Applying the resistor values specified (working in mA, volts and kΩ) the change in \( Tr_1 \) emitter current is

\[
\frac{1}{1.8} \cdot \frac{1}{0.1} = 10.6u
\]

This circuit improves on the balance obtained by Jeff Macaulay's valve power amplifier.
Linearity confusion

Looking back at past debates of fet/bjt linearity, I noticed that the word linearity is being referred differently by some correspondents. A lot of the fet supporter comment on the linearity between $g_n$ and $V_{gs}$. Fair enough. Suppose $g_n$ does increase linearly with $V_{gs}$. Then we can write $g_n = kV_{gs}$, where $k$ is a constant.

Knowing that $i_d = g_m V_{gs}$ and substituting $g_m=V_{gs}/r_o$, gives $i_d=kV_{gs}^2$. What is important in audio signal amplification is linearity between $i_d$ and $V_{gs}$. The fet's square law does have an advantage though. Distortions produced by square law devices are mainly even order. Using such a device in push-pull results in cancelling of even-odd distortion. If precise matching can be done for output devices – which I doubt – fets may have less distortion in the end.

Bipolar devices are probably more linear on their own, but due to the doubling of odd-order distortion under push-pull operation, they may have higher distortion compared to a perfectly aligned fet stage. Using fets with reasonable $g_n$ also results in lower open-loop output impedance. This is due to the source follower output impedance of $1/g_m$. Output impedance below 1Ω would be enough to control most speakers. There is another point. Given low open-loop output impedance, any back emf and radio-frequency disturbance coming from the speaker will be earthed at the output terminals. Amplifiers relying on global negative feedback to lower output impedance behave differently. Any unwanted rf/emf coming in from the output terminals will be injected into the second base of the long tailed pair. As a result, the amplifier is forced to perform disturbance rejection on top of the amplification it has to do. This is not ideal.

Mr Self rejected the idea of

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- **Output waveforms**: sine, square, triangle, pulse and sawtooth

**Waveform characteristics**

- **Sinewave distortion**: <1% at ≤200kHz
- **Triangle non-linearity**: <1% at ≤200kHz
- **Square rise and fall**: <20ns

**Output**

- 50Ω o/p impedance: 50Ω ±5%
- 0dB amplitude: 2Vpp to 20Vpp no load
- −20dB amplitude: 200mV to 2V pk-pk, no load
- −40dB amplitude: 20mV to 200mV pk-pk, no load
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- **Offset**: ±10V no load, ±5V 50Ω load
- **Duty cycle**: 10% to 90%

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- **Size, overall**: 203mm x 195mm x 75mm
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- **Power consumption**: typically 10VA
- **Specification**: BS EN 61010-1/1993

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resistively loading the \( V_{	ext{in}} \) transistor due to distortion. However, it has its merits when bipolar devices are used at the output. Choosing \( R_I \) at 5kΩ, and assuming an output Darlington base current of 100kA, the open-loop impedance will be 0.5Ω. As a result, there is no need to resort to feedback to lower open-loop output impedance.

Varying the value of \( R_I \) results in a trade off between distortion and open-loop output impedance. Subjectivists may find it interesting to replace \( R_I \) with a 1000Ω potentiometer.

Lastly, why doesn't anybody use the Siliconix Vimos device which claim superior linearity over ordinary mosfets? I hardly ever see the Toshiba IGBT's used either.

Koji Kiyolewau,
Allstedre
Derby

Excellent grounds for more debate

I beg to disagree with sentiments expressed by John Watkinson in your letters column of EW, Nov '95. Contrary to John's point of view, I feel that the debate on audio amplifiers has been healthy and should be encouraged.

Whether for Aerospace, Marine, Defence or Industrial applications, amplifiers are fundamentally the same. As a result, whatever is learned by investigating audio amplifiers is certainly applicable to the other areas just mentioned.

The fact that audio-frequency amplifiers have been hotly debated over these many months, is a clear indication that a lot has been taken for granted over the years. Of course thanks to our new tools of trade, analyses are made so much easier. This book is not to take anything away from those who have burnt the midnight oil to bring it all to us.

For most of us in the "Third World" where we are more at home with wild life than wires, the closest we get to Spice is in our cup of morning tea. Hence we are very grateful to the Selfs, Ohlsson, Hoods, and Duncans for the privilege of having the chance to share their simulation and test results, and also their unquestionable technical expertise.

Clyve J. Caines
Technical Services Manager
Nairobi, Kenya, East Africa.

Duncan disputes

Prof. Cherry in Letters, Nov '95, has not read my words. On p. 20, Jan '95, I am glad that Prof. Cherry recognises that music signals are usually asymmetrical.

As a world-renowned advocate - along with top recordists - of a minimum record/replay path built from all-dc electronics, ideally free from unnatural high-pass capacitors, the discharge error he mentions does not much concern me; his caution should be directed as those like Self, whose power amplifier alone contains as many electrolytic capacitors as my entire, ideal dc audio chain.

Much as I love his approach to electronics, it appears that Prof. Cherry cannot see nor share the joke over the futility (on his own objectivist terms) of his scheme, viz: "there are no sharp edges nor sub-sonic signals surviving in multitrack recordings because of all the high-pass filters. So making the last and hundredth device in the chain have an ideal square wave response is a waste."

If Prof. Cherry chooses to ignore the role of capacitor constructions, impedances and microphony on sonics, and denies real electrolytic capacitor tolerances and temperature coefficients, that is not my problem. There are in-depth works on audio capacitor reality by Marsh and Jung and someone called Duncan. The five minutes of tweaking he suggests is ok by me but grossly unacceptable to most manufacturers, as rightly indicated by Self, and it ignores drift and temperature coefficient.

Since my colleagues and I design and produce audio systems that have helped enthrall many thousands of discerning customers, I have enough substantiation, thank you. At least we'll be open minded enough to give Prof. Cherry's low-frequency compensation scheme a listen some day, even if we think it is as deranged as his chassis earthing recommendations.

Ben Duncan
Lincoln

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Living with the Chip “Captures the excitement of the chip industry” – Gordon Moore.
A new direction in orientation

Sensitive down to 10nT, the FGM three terminal magnetic detector is suitable for a host of two and three-axis orientation applications. Richard Noble explains.

The most well known orientation detector is the common compass, the simplest form of which is a single axis device known as a steer-on-heading compass or "poor man's autopilot". It comprises a single magnetic sensor mounted on a rotatable disk, marked in degrees around the periphery and fitted with a stationary indicating pointer, Fig. 1.

Output of the sensor is connected to a frequency-to-voltage converter circuit feeding a centre-zero meter display as described in the November issue of EW.

In Fig. 1, if the boat veers to the left, the meter needle swings to the right, indicating the need to steer in that direction to correct the course. If the rotating disc is turned to a new heading the needle shows the best direction to steer in until the new heading is reached. At this point it returns to the centre position.

Magnitude of the deflection gives an indication of the amount of correction needed at any time. This type of steering system is said to be easier on the helmsman than having to remember and follow a degree bearing.

For this simple system to be useful, the sensor must be gimbaled and weighted so as to keep its axis level at all times. Since it is only a single axis device it only needs a single gimbal, provided that the gimbal rotates with the heading disc.

The next level of complexity is a two-axis compass. For this, it is best to replace the frequency-to-voltage converter with a microcontroller of some sort as a number of more complex operations need to be carried out - especially if a readout display is wanted.

Many varieties exist, all capable of dealing with the requirements of a compass. But because the sensors have their own analogue-to-digital feature, microcontrollers which have frequency or period determining features built in are the obvious choice in this instance.

Angular sensitivity

It is useful to look first at the angular response of an individual sensor. Because of its structure it 'sees' the full magnitude of a field, which is aligned along its long axis. For any field at right angles to this axis, it gives zero output in the sense that its period corresponds to that of a zero field condition.

For a field aligned at an angle between these two extremes the response is proportional to the projection of the field on to the long axis of the sensor, therefore to the cosine of the angle between field and sensor. This gives rise to the classic figure-of-eight polar diagram, comprising two contacting circles or, in the three dimensional case, two contacting spheres, Fig. 2.

If sensors are aligned along the axes of any two or three axis coordinate system the...
sensor outputs represent the direction cosines of the field vector with respect to that coordinate system. For convenience the chosen system is usually cartesian but this is not essential.

As described under the calibration and linearising techniques, it is convenient to normalise the sensor readings by dividing through by the zero-field period. In orientation type devices it is also convenient to then subtract one from these normalised values to yield equal positive and negative ranges about zero. These adjusted values are then proportional, but not yet equal, to the direction cosines of the field vector.

The reason is that no two sensors are exactly alike in absolute sensitivity, and must now be calibrated so as to achieve a standard sensitivity. This can be done by the calibration coil method described earlier, after which proportionality constants can be assigned as multipliers to equalise the sensitivities. Alternatively it can be done by aligning the individual sensors in turn along the local earth field vector in the two possible directions, 180° apart and determining the corresponding maxima and minima for each sensor. Proportionality constants are again assigned to equalise the sensitivities.

Two-axis orientation sensing

The two axis compass uses twin sensors superimposed at right angles to one another in the same location and both constrained to lie in the horizontal plane. The sensitivity equalising process in this case can be semi automated by rotating the sensors through a full 360° and allowing the software to determine the maxima and minima for both axes.

Assume that the two now standardised values are x and y components of the local field vector, \( h \), having a modulus equal to \( \sqrt{x^2+y^2} \). Now, the final normalisation can be realised by dividing each component by this modulus. This gives the true direction cosines of the field vector which together define the unit vector \( h \), having the same direction as the field vector, \( h \).

This process eliminates the effect of any variation of the absolute magnitude of the measured field, since the sum of the squares of the direction cosines always equals one. Earth field variations are insignificant in this context, but supply or ambient temperature changes are neutralised provided all sensors are equally affected.

The direction cosines can be readily converted to a more customary representation such as angular heading as follows.

Assume that the compass heading indication is aligned with the y-axis and label the components of the unit vector, \( i_x \) and \( i_y \). Then it can be seen from Fig. 3 that if \( \theta \) is the conventional heading angle,

\[ \tan \theta = -i_x/i_y \]

and

\[ \theta = \tan^{-1}(-i_x/i_y) \]

and the compass heading is simply the arctangent of the ratio of the x and y components of the unit vector in the Earth's field direction.

For a three-dimensional coordinate axis system with the z-axis at right angles to the other two there is no conflict with anything that has been said so far, provided that the z-axis remains vertical. In fact this becomes the necessary condition for the successful operation of this type of compass, which needs to be gimbaled in two directions and appropriately weighted.

It will be evident that some attention to signs and the possible divisions by zero will be required in considering the full circle of 360°. While this arctangent solution may be possible for a computer with trigonometric functions in a high-level language, it is not appropriate for a lower level of implementation such as a microcontroller. However, the underlying principle remains the same in alternative approaches.

The full circle in which the heading vector lies may be segmented into eight 45° octants and the octant occupied by the field vector can be identified by simple non-trigonometric tests, easily applied in software.

The rules which do this involve the signs of the \( i_x \) and \( i_y \) components and the comparative magnitudes of these components taken as an ordered set. For example if \( i_x < 0 \) and \( i_y > 0 \) the heading must lie in the first quadrant. If, in addition, \( i_x < |i_y| \) it must lie in the first octant between 0° and 45° in the previous diagram. Other combinations uniquely identify the remaining octants, Table 1.

<table>
<thead>
<tr>
<th>Sign of ( i_x )</th>
<th>Sign of ( i_y )</th>
<th>Octant No</th>
</tr>
</thead>
<tbody>
<tr>
<td>Negative</td>
<td>Positive</td>
<td>Less</td>
</tr>
<tr>
<td>Positive</td>
<td>Negative</td>
<td>Greater</td>
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<td>Negative</td>
<td>Negative</td>
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<td>Positive</td>
<td>Less</td>
<td>Greater</td>
</tr>
<tr>
<td>Less</td>
<td>Greater</td>
<td>Greater</td>
</tr>
</tbody>
</table>

Implementation of these rules on their own provide an eight point compass with a ±22.5° accuracy, which while not very precise may be adequate for some undemanding applications. There are other benefits in more sophisticated versions. The first advantage of this technique is the Gray-code like way in which the octant rules work.

At each octant boundary only one of the rule parameters changes. For example at 45° no sign changes occur but the inequality between \( i_x \) and \( i_y \) changes direction. At 90° no inequality changes occur and the sign of only \( i_y \) changes. This property prevents large scale jitter and confusion which might otherwise occur at the octant transitions if the changes were not totally synchronous.

A second advantage is that in each of the octants, a linear function of either \( i_x \) or \( i_y \) can be identified which is virtually equal to the desired heading angle, to within a small error.

In the 0° to 45° range if \( k_i \) is interpreted as a radian angle it is in fact very little different from the appropriate arctangent for that octant. If \( k = 1.08 \) the error in doing this is nowhere greater than about 1.25°. If \( k \) is ignored and the unit vector \( x \) component alone is interpreted as radians the error is never worse than 4.5°, permitting the implementation of a 5° precision compass very easily. Note that for this purpose the modulus of the \( x \) component is used, eliminating the need to consider signs.

In the next octant, between 45° and 90°, \( k_i \) interpreted as an angle and subtracted from 90° is very close to the correct heading. This pattern repeats around the full circle and leads to the following rule.

Whenever the direction cosines is the smaller is interpreted as an angle and in odd octants is added to the nearest quadrant boundary. However in even octants it is subtracted from the nearest quadrant boundary to obtain the heading.

If as is likely in a software implementation the octants are numbered 0 to 7 rather than 1 to 8, the odd and even should be reversed in the previous statement of the rule. In conjunction with using \( k = 1.08 \) this rule will provide almost ±1° precision in a software implementation requiring no trigonometric functions.

Alternatively, since the error is small, a very short lookup table of adjustments to be added to the heading obtained with \( k = 1 \) will improve the precision to a level of around ±0.5°.
This technique will not produce a compass of this accuracy. Rather, the contribution to the total error budget from this source will be minimised to the extent indicated. Other sources may contribute larger errors in a final design if they are not suitably addressed.

One important potential error is lack of orthogonality in the axes of the two sensors. This can cause a smoothly varying error around the whole compass circle which can be much larger than those discussed above. Fortunately there is a relatively simple correction technique for this as can be seen from the following analysis.

In Fig. 4, \( i \) is the unit vector in the field direction, \( \theta \) is the heading angle and \( \phi \) is the small angular error by which the \( x \)-axis sensor departs from the correct right angled position. Also \( i_x \) is the true \( x \) component of the unit vector, \( i_y \) is the true \( y \) component and \( i_x' \) is the apparent measured \( x \) component of the sensor in error.

You can see from the geometry of the figure that,

\[
\begin{align*}
  i_x & = \cos \theta \sin \phi \\
  i_y & = \sin \theta \\
  i_x' & = \sin(\theta + \phi)
\end{align*}
\]

Expanding the last relation,

\[ i_x' = \sin \theta \cos \phi - \cos \theta \sin \phi \]

Since \( \phi \) is small \( \cos \phi \) may be taken as one and \( \sin \phi \) as just equal to \( \phi \) giving

\[ i_x' = \sin \theta \cos \phi - \cos \theta \sin \phi \]

Hence,

\[ i_x' = i_y \cos \phi + i_x \sin \phi \]

It can be seen from this that the desired \( x \) component of the unit vector can be obtained from the apparent measured component, for all angles, by adding a small fixed portion of the \( y \) component. The proportion to be added is equal to the orthogonality error in radians.

The value of \( \phi \) can be found, for a standardised and normalised sensor set by rotating the configuration in the Earth's field and measuring the angle between the zero-field positions of each sensor.

Alternatively, the algorithm can be added retrospectively to an otherwise completed compass by checking the error during a full 360° rotation. The value of \( \phi \) can be taken to be the average of the errors at 90° and 270° shown by the digital display. Such a determination needs only to be made once.

If the microcontroller has no convenient way of memorising the correction it could alternatively be read from a trimmer value on power-up using \( RC \) timing or some other relatively crude analogue input method. The orthogonality then becomes one of the possible adjustments available to the user during the compass "boxing" exercise.

Detecting errors

Probably the largest of the final observed errors will arise from failure to constrain the axes of the sensors to the horizontal plane. The errors depend on the direction of tilt and the heading, and on some headings small angular tilts will multiply up to much larger heading errors.

For example on a north heading a 1° north south tilt will produce no error, but a 1° east west tilt will give rise to almost 2.5° of heading error. There is no simple cure for this other than effective double gimbaling, suitably weighted, though short term averaging of multiple readings can improve the stability of the displayed output.

Another more complex alternative is to use a gravity sensor to determine the direction of the gravity vector and use trigonometric calculation to correct for the effects of tilt.

A final aspect of overall accuracy concerns the required precision of sensor readings. Interestingly, this is surprisingly lower than might be thought. Using the type of algorithm described earlier, a full 360° of 5° precision requires only that the measured components be slotted into one of forty-five almost evenly spaced bands. A relatively low six-bit binary measure will cover this. For a 5° precision a miserly four bits is adequate.

In conclusion, for those who may design, build and use a compass, in anger, the illusion of three-dimensional systems but do not have any mechanism for rolling around the remaining axis, since it would be entirely pointless.

If only the human head was satisfied by the same mechanism. Virtual reality would be
much easier to implement.

The reason that this works is that as soon as the roll axis is constrained to remain horizontal, the rotational ambiguity around the field vector, mentioned previously, disappears. The trigonometry of the unit vector components is solvable and yields not only the azimuth angles, like a compass, but also the elevation angles.

Gun platforms fall into this category, as do steerable satellite type aerials, some robot mechanisms and any device which needs to point to a direction in space from a horizontal platform. Complex devices of this nature are probably well served by the expensive mechanical compass and any device which needs to know the direction in a polar diagram, what results is a time invariant 'signature' of the object. In some sense this signature contains information about the range, since for a close passage it will have a large angle polar diagram and for a remote passage a small angle diagram. This range is not absolute as it will also depend on the equivalent magnetic length of the magnetic moment being observed, which is roughly correlated with the size of the vehicle most of the time.

If a three dimensional sensor configuration were gimballed transversely and suitably weighted, it could perhaps maintain the pitch axis of the sensor set sufficiently horizontal to allow the strategy under discussion to generate a heading and additionally a bank angle of acceptable precision.

Whatever the precision, it would represent a vast improvement on the conventional fully gimballed compass and add half of an artificial horizon into the bargain. It would also weigh and cost less than any gyroscopic equivalent.

Three and two dimensional ferrous detectors

It is possible to elaborate the design of fixed single sensor vehicle detectors described earlier, with advantage, by using a two sensor version. Even when restricted to the horizontal plane, an orthogonal sensor set can provide more information, in the sense that it can provide both angular and magnitude signals for the anomaly caused by the vehicle passage.

An object with a magnetic moment possesses an external pattern of lines of force similar to that of a permanent magnet. This line of force pattern combines additively with the earth's field lines of force which consist locally of straight parallel lines. If the disturbing magnetic moment passes very close to the sensors it produces not only a variation in field magnitude but also large swings in the angular orientation of the detected field. If the passage is more remote from the sensors, not only is the magnitude of the signal reduced, but also total angular swing.

While the time variation of these parameters gives some indication of the speed of the passage, the magnitude of the signal is plotted against the angle in a polar diagram, what results is a time invariant 'signature' of the object. In some sense this signature contains information about the range, since for a close passage it will have a large angle polar diagram and for a remote passage a small angle diagram. This range is not absolute as it will also depend on the equivalent magnetic length of the magnetic moment being observed, which is roughly correlated with the size of the vehicle most of the time.

The fall off in field strength is proportional to the inverse cube of the ratio of the range to the magnetic length. As a result, the field from large objects falls off more slowly than that from small ones.

Actual magnitude and angle variations will be quite small but can be increased to usable size by the digital heterodyne method or, in

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**Output frequency v field**

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**The FGM-3 outputs a frequency between 50kHz and 120kHz whose period represents magnetic fields in the range ±50µT and is highly stable with temperature.**

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January 1996 ELECTRONICS WORLD
this case its software equivalent. The polar diagrams shown are simplified and guesswork, not based on any tests, Fig. 5.

It seems that this is an area worthy of more serious research on practical real life situations, since it may resolve the problems of lane separation and vehicle classification in multiple vehicle studies.

Orientation sensitivity elimination

The ferrous detection systems discussed so far have been static ones and the fixed large signals produced by the earth's field can be relatively easily eliminated from the desired indications. In situations where the sensor configurations will inevitably be subject to unpredictable movement, the high orientation sensitivity becomes a serious disadvantage in the search for very small signals.

However, consider a perfectly standardised and normalised, perfectly linearised and perfectly orthogonal sensor set. The problem is easy to deal with since the sum of the squares of the three outputs must always be equal to the field vector modulus squared - a scalar quantity without any orientation. Success of any real implementation will clearly be only a function of how close to perfection the above requirements come.

The mathematics are simple and readily implemented on computer or microcontroller. Basic sensitivity of the sensors is adequate, matching of the calibrations is more constructive than absolute accuracy, orthogonality correction can be carried out to a fairly high degree, but non-linearity may be a troublesome source of error.

A technique helpful in these circumstances is to use some sort of negative feedback to improve both linearity and stability. The method consists of overwinding the sensor with a solenoidal coil in which a controlled field can be produced. This field is automatically adjusted to cancel out to zero, the local field which the sensor would otherwise experience.

Solenoid current giving rise to this cancelling field must be proportional to the local field being cancelled. Since the sensor only ever sees a zero field, its own non-linearity is no longer of consequence and the cancelling current is a direct and linear measure of the local field magnitude.

This approach obviously calls for a d-to-a converter to control the current in the cancellation coil. With a microcontroller, this could be a pulse-width modulated, single-bit, output and low-pass filter arrangement, as used so successfully in many current low-cost digital audio devices. Software complexity increases but the hardware cost is still held low, probably calling only for a linear current generator of modest current capability.

In any case, total 360° orientation de-sensitising is not always needed. Reductions in the angular variation achieved by other means will often considerably improve performance. Examples are a detector carried in a normally floating vehicle, or a neutral buoyancy weighted float, trailed just submerged. An error may exist in the output but it remains passably constant.

This type of system, Fig. 6, could find uses as a detector of seabed wrecks in modest depths or as a search tool in archaeological studies. Constructed with sufficient care, it provides a low cost and compact alternative to nuclear magnetic resonance devices in some applications.

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**Fig. 6. In a three-dimensional sensor system, a microcontroller with PWM facilities can be used to supply the cancellation coils.**
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Modelling cable and cmr

As an aid to designers optimising balanced audio and instrumentation systems, Ben Duncan demonstrates how to model common-mode rejection and simulate cmr testing. In the process, he exposes some of the subtleties of connecting shielded cables.

In an earlier article, a number of IC makers’ Spice ‘macro’ models were tested for common-mode rejection. Some subsequent wider testing reinforces the impression that makers’ models have some way to go to simulate reality in this area.

For example, the Spice model of the SSM 2017 – a true differential receiver converting balanced lines to unbalanced – yields about −35dB common-mode rejection when tested. While the data sheet omits to graph cmr, real units sampled over five years have consistently measured around −65dB to −70dB. The comparison is across the audio band.

Rather than allow these failings to damn the use of today’s powerful circuit analysis tools, progress can be made by other means.

Ideal balanced receiver

For most investigations, a perfectly balanced differential receiver: or differential-to-single-ended converter with near infinite common-mode rejection is an ideal that will help designers isolate other causes of common-mode rejection degradation.

A linear ‘V-Of-V’ source that provides the ideal differential-to-single-ended conversion is available in MicroCAP-IV and -V. For other simulators based on Spice 2G, voltage-controlled voltage source ‘E’ is a similar element.

A major advantage of ideal source rather than an op-amp or other active device model, is that the device performance is completely free from physical limits. Slew limits, maximum voltage swing and dc offset are all initially irrelevant to cmr investigations. The linear ‘V-Of-V’ in MicroCAP has just one parameter – a setting for numeric V0/VVout ratio, for example 1.0 for 0dB.

Looking at the top left circuit in Fig. 1, the raw part appears as a circle – the output side – associated with two round input terminals – the isolated inputs – on the left. Here, we are interested in checking the cmr of the device, so the balanced input terminals are linked and driven by an ac common mode test voltage, Vcm.

In ac mode, the test level defaults to a nominal 1V and frequency is swept – irrespective of the generators’ amplitude and frequency defined in the .MODEL statement text beneath.

Macro capabilities

MicroCAP’s macro capability is easy to use and is a little more than just a means of grouping key strokes. In Fig. 1, the circuit on the right repeats the identical V-Of-V after being macro’d into a four-pin op-amp shape. This is called DTSEC, which represents a differential-to-single-ended converter. The tie ‘bb’ connects the Vcm test source. In analysis, the single ended output voltages are read between the named nodes V0, V02, etc, and ground.

In Fig. 1, the lower pair of circuits comprise a wholly independent differential gain test of the same V-Of-V, and again, the macro’d version of it, DTSEC, is on the right. This demonstrates MicroCAP’s ability to simultaneously test independent circuits, thereby greatly speeding up investigation and control investigative file branching.

Figure 2 is a plot of results. The recovered differential signal is 0dB. For this to be the case, the V-Of-V’s gain parameter needs changing from the default 1.0 to 0.5, i.e. −6dB. This is because the two voltage sources Voff need to model a perfect balanced source are both set to 1V and summed in ac analysis mode. This is an intrinsic feature which cannot be changed.

The DTSEC’s cmr is not quite infinite, but at −3000dB, the ratio comfortably exceeds physical limits assuming there are about 1x10^9 (1.6kHz) atoms in the local universe. Of course, the macros on the right-hand side of Fig. 1 perform exactly the same. To enable this to be seen, their expressions in Fig. 1 include ‘−1V’ and ‘−500dB’ in the upper and lower plots respectively, as an offset for clarity.
Virtual cmr workshop

In Fig. 3, the DTSEC is being tested in a circuit where the resistive connections and their errors are real enough, but where reactive effects are being ignored. This would be a reasonable comparison for an on-board interface, where transmitter and receiver connections are very short.

Resistors are defined in the statements on the right of the schematic. Values of \( R_{AB} \) or \( R_{source} \) and its twin, \( R_{c} \) or \( R_{source,c} \), are typical output stand-off resistors. Hot load resistor \( R_{L} \) and its cold twin \( R_{cL} \) are typical of a conventional audio bridging audio input, to the IEC 268 convention. These provide a bias and discharge path, as well as defining input impedances \( R_{diff} \) and \( R_{CM} \).

The DTSEC drives a 10kΩ resistor but this is in no way essential in MicroCAP. It just makes the single ended output visually stand out.

In the lower half of Fig. 3, a completely different pair of tests is going on. Here, the circuit above has been split and its output resistors, is split from the DTSEC, and its input resistors. The \( Z_{diff} \) of both halves is then read using individual 'I' sources, set up to force a frequency-swept ac current of 1A differentially up each. The current converts the ac Y axis from volts to an ohms scale.

In Fig. 5, the effects of shunt capacitative reactance are added and part isolated for study. This time the circuit source and load resistors are untoleranced, but offset with realistic high specification mismatches of 0.025% for 20Ω and 0.001% for 100Ω.

Shunt capacitance is also untoleranced, and perfectly balanced. As in Fig. 3, capacitors connect to ground. By practical inference, this is the local zero volt reference, seen beneath the DTSEC.

Common-mode test source \( V_{cm} \) shares the same ground by definition; don’t be deceived by its left side geographic location. In the upper circuit, the introduced shunt capacitance is just 22pF – believable if the interface is proximate on a hundred and fifty yard groundplane pcb. In the lower figure, capacitance is far higher, being typical of tens of yards or so of shielded cable, the total value possibly including shunt rf filter capacitors.

When looking at the outcome in Fig. 6, it is important to note that the added capacitors are not in any way unbalancing the line. Yet the higher capacitance, while not as a load, begins to significantly degrade the common-mode rejection at low rf of around 1MHz, and above. Of course, the model neglects other effects becoming significant at 1MHz but the trend is clear enough.

Introducing cable effects

Accurate balancing of analogue and audio signal feeds has a major role to play in emi immunity for eme. But attaining high cmrs in balanced transmitters and receivers alone is not good enough. Real cmr may depend as much, or more, on the
quality of cable used, and moreover, on the way that cable shielding is connected.

A recent AES paper has drawn attention to cmr degradation caused by core-to-shield capacitance imbalance in twisted pair shielded cables. It specifically highlights the effects of the pairs having naturally separate colours of insulation, for example red and black. Owing to the different additives used for colouring, each type has a different permittivity. This, and differences in the extruded insulation thickness, lead to a different capacitance between the circuit legs. Typically this is of the order of 1-5%. This may not seem much, but it has increasingly serious repercussions at rf in any system requiring high emi immunity.

After comparing the spot foil-shield capacitances of some balanced cable offcuts lying around the lab, Fig. 7 mimics the reality. In the upper circuit, the 1% typical imbalance between two cores of a foil-shielded pair, for example type 2401, is modelled by defining the core/shield capacitance caused by core-to-shield capacitance imbalance in twisted pair shielded cables. It specifically highlights the effects of the pairs having naturally separate colours of insulation, for example red and black. Owing to the different additives used for colouring, each type has a different permittivity. This, and differences in the extruded insulation thickness, lead to a different capacitance between the circuit legs. Typically this is of the order of 1-5%. This may not seem much, but it has increasingly serious repercussions at rf in any system requiring high emi immunity.

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As in Fig. 5, note that realistically imbalanced resistive values set a maximum cmr baseline, and that cable capacitances are tied to ground. This is the receiver output reference, alias local OV. This means the shield is connected to earth only at the receiver end. While disputable, this method often works in practice and is considered essential when a shielded multicore is unplugged from 90kW of front-of-house power amplification.

Fig. 7. Shunt capacitance definitions have been appended with core-to-shield capacitance tolerances that are typical of real shielded pair cables.

Fig. 9. Cable shield is strapped at the balanced transmitter’s ground end, alias V_{e},’s hot side. This scheme is ‘more’ correct in theory, than terminating at the receiver end.
But what are the hazards?

Figure 8 shows the effect of stepping the cables' capacitive imbalances onto the given tolerances. Clearly the random mismatching caused by tolerancing is wreaking havoc balance with cmr affected most. Using a cable with twice as much core/shield capacitance tolerance than the foil type is not helping – by 6 dB as you might expect.

In Fig. 9, the circuit has changed subtly. The cable capacitances' drain ends have been 'back-referred' effectively to the source ground. This ground is notionally the 'hot' side of the VCM sine wave. Some effect, connection of the cable shield has been changed from receiver to transmitter ground. Now, as shown in Fig. 10, high-frequency cmr degradation is far lower, and the effect of inter-pair capacitance mismatch is also diminished.

The fact remains, however, that in giant, touring multimedia audio systems, the most inobtrusive system noise quality has required a flexible approach as to which end of the shield should be grounded. Others have noted this.1 This means the Fig. 7 model, while a firm basis, and useful for predicting performance of specific systems, naturally requires some more parametric elements to be added to be useful in many real applications.

Output cmr test station

In Fig. 11, output cmr is measured by two external methods, where the source can be modelled as just an impedance. A third method, specified by the BBC, is outside the scope of this setup, as it requires the source to be active and driven.

In the upper figure, $R_{cmr}$ and $L_{cmr}$ represent the active source's impedance with high negative feedback, and the inductance modelling the effect of diminishing negative feedback at hf. Resistors $R_{cmr}$ and $R_{cmr}$ are the usual output resistors. Again, all parts have arbitrary but realistic fixed imbalances.

The test involves shifting the balanced output terminals and applying $V_{cmr}$, the test signal, to this point. A balanced receiver is then connected before, and across the output resistors. Common-mode rejection, i.e. balance, is proportional to the voltage here.

The ability of circuit modelling to mimic even the industry standard test gear you might use to verify computations, has not been widely published hitherto. To change this, on the right, the cmr of the analyser input of my Audio Precision System One test set has been modelled, by measuring the output of the reference feedback, $R_{cmr}$, then adjusting the values of the shunt and series $R_{cmr}$ and $C_{cmr}$ to closely match the curve.

In the lower circuitry of Fig. 11, output cmr is measured by another method, cited by SSM in their SSM-2142 data sheet. Here, $V_{cmr}$ is connected driven up the via a pair of quite tightly matched ±0.01% resistors, $R_{cmr}$, in series with a third resistor.

This method requires a fiddly fixture, but avoids any need to delve inside the box. On the bottom right, the AP input (above) has now been macro'd into the same shape as the dual-to-single-end converter.

Results, shown in Fig. 12, demonstrate there is not a lot between the two test methods – at least over 100 linearly stepped Monte Carlo runs shown. A few minutes spent running such tests can evaporate myths, thereby reducing the noise floor at audio conventions!
DESIGN BRIEF

High-performance thd meter

Taking advantage of modern components, Ian Hickman has developed a new distortion meter combining a measuring resolution of 0.001% with design simplicity.

Designed this thd meter for testing hi-fi amplifiers, in conjunction with a low distortion sinewave oscillator.

Figure 1a) is a schematic for a thd meter design dating from the early seventies. Figure 1b) shows its full circuit, which is interesting in that it highlights some of the problems in thd meter design.

The complete instrument includes a low-pass filter with a choice of switchable cut-off frequencies. This extends the lower limit of the measuring range by limiting the noise bandwidth. Provision for selecting a high-pass filter to reject hum is also included but not shown.

Designed as a distortion monitor, the circuit was intended for use with a separate external audio-frequency voltmeter – preferably one with true rms response.

Some quick sums show it to be still 7.5dB down at the second harmonic of the notch frequency. This is because of the low Q of the Wien network caused by the two capacitors and two resistors, not the complete bridge. It is well below unity – in fact just a third. Consequently, it is necessary to include the notch circuit in an overall negative feedback loop. This brings the response at the second harmonic to ideally much less than 1dB down.

The effective absence of negative feedback at the frequency of the test signal when the notch is correctly tuned to it, however, means that the fundamental is considerably accentuated in the stages within the loop. As a result it is necessary to keep the input amplitude well below the overload level of these stages. This helps prevent the distortion meter introducing distortion of its own.

A consequence is that the noise floor can limit attempts to measure very low thd levels. A further consequence of negative feedback is to narrow the notch, making tuning critical. The original article recommended a slow motion drive with at least a 100:1 reduction ratio. These and other considerations limited the measurement range of the instrument of reference 1 to the order of 0.01% at 1kHz.

Some years ago, I designed and built a thd meter with ranges down to 0.01% full scale, permitting readings down to around 0.002% or lower. It used a state-variable filter, svf, based circuit which, having a higher Q, does not

Notch circuitry

Using a Wien bridge arrangement, the circuit provides a notch or transmission zero at the fundamental of the sinewave test signal. However, the notch produced by this type of circuit is cusp-shaped. Not only is it very narrow at the null, but its return to full output at very low and very high frequencies is leisurely.

Fig. 1a) Above, schematic of a notch circuit with negative feedback to provide a (nearly) flat response by one octave either side of the notch.

b) On the right, part circuit of the distortion monitor described in reference 1.

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need so much overall negative feedback to obtain a reasonably flat response at the second harmonic. Measurements at the limit of the circuit's sensitivity were very tricky due to the narrowness of the notch, as detailed in the panel. Consequently, when making thd measurements on a low-distortion oscillator, the thd meter was preceded by an auxiliary test circuit. This consisted of a fixed-frequency twin-tee network followed by a two pole Chebychev high-pass filter. Tuning and peaking of the latter were adjusted to give a sensibly flat response, in conjunction with the notch, from the second harmonic upwards. Three such circuits permitted spot frequency testing at 20Hz, 600Hz and 10kHz. While this arrangement provided thd measurements which are reliably accurate – due to suppression of the fundamental in a passive network before it meets any active circuitry – the restriction to spot frequency testing is in practice a serious drawback. A new approach For many years, I have been planning to replace my original distortion meter with an improved design offering continuous tuning and a wider notch. The wider notch is needed for measurements down towards the 0.001% level since adequate suppression of the fundamental is not possible with a single two-pole notch circuit. As indicated in the panel, to measure even 0.01% thd, requiring suppression of the fundamental to at least 0.003%, implies an accuracy of tuning of 15ppm – the equivalent of 0.1Hz/0.1Hz at 1kHz. This required accuracy is not an absolute figure. It is relative to the frequency of the sinewave oscillator providing the test signal. Even if stability of the notch tuning were perfect, the oscillator may not exhibit the necessary long term stability to allow readings to be taken. Even if it did, the short term stability of an RC oscillator is likely to be inadequate. The inevitable close-in noise sidebands will be inadequately suppressed by the notch. If you prefer to think in the time domain, frequency of the oscillator will shuffle about by a minuscule amount. This results in the fundamental peeping out randomly on either side of the notch, preventing a steady reading representing the harmonics only. Notches in tandem The solution presented here is to use two notches in tandem, greatly reducing the suppression required of each. This four pole arrangement also permits a design that avoids the accentuation of the fundamental within the loop. This is necessary with a two-pole notch circuit to achieve a response at the second harmonic which is no more than, say, 1dB down on the 'flat', ie on the response far from the notch. The scheme is outlined in Fig. 2, where the first stage is an svf notch circuit with a Q of unity. As a result there is no accentuation of the fundamental, high pass, band-pass and low-pass responses. All are unity at the tuned notch frequency, Fig. 3. With the chosen Q of unity, there is a slight peak in the low-pass response of just over 1dB at 62% of the tuned frequency – and at 1.6 times the tuned frequency in the case of the high pass response. But in thd testing there is no signal present at this frequency. As low and high pass outputs are in antiphase, summing them produces a notch at the tuned frequency.

The first stage sums equal contributions from the low and high pass outputs, resulting in a symmetrical notch. With the chosen Q of unity, this is just 1dB down at twice the notch frequency. High pass output is summed with just 60% of the low-pass output by the second stage. This results in the notch occurring below the tuned frequency, with low-frequency response only 60% that of the high-frequency response, Fig. 2. Furthermore, the notch now occurs at a frequency below the svf stage resonant frequency. Resonant frequency is that at which low, high and band-pass responses are all equal, and is given by, $f = 1/(2\pi CR)$. Flat response By choosing a smaller value of CR for the second stage – in conjunction with the chosen ratio of low to high-pass contribution – its notch can be arranged to coincide with that of the first. By choosing a suitable value of Q for the second stage, a peak occurs at its resonant frequency, ie somewhere above the notch. Its amplitude can be made +1.6dB at twice the notch frequency. This compensates for the -1.6dB response of the first stage. In fact, by judicious adjustment of the second-stage resonant frequency, ratio of low to high-pass contribution and Q, overall response can be made flat at the second and all higher harmonics. The arrangement has certain similarities to a four-pole elliptic high-pass filter, but there is a significant difference. Instead of spacing the two frequencies of zero response apart so as to maintain a designed stop-band attenuation, A, all the way down to 0Hz, they are made coincident. This is because there is only one signal in the stop-band – namely the fundamental of a sinewave test frequency. The harmonics all lie in the pass-band, which in this design is flat to within less than 0.1dB.

Performance of the thd meter The two stages were made up in temporary form, as per Fig. 4a) and 4b) and tested. Note the tuning arrangement using a potentiometer to drive each integrator's input resistor R. This scheme using a fixed R provides linear tuning. As a result, poor resolution is avoided at the high-frequency end of the range which occurs if tuning is effected by varying R. Figure 5a) shows the notch output of the Fig. 4a) circuit, set to 1kHz, at 10dB/vertical division, the span being 0-5kHz. Over 60dB of rejection was observed, the residual due to the low and high pass outputs being not quite in antiphase: 60dB down corresponds to a depar-

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**Fig. 2. Schematic arrangement of an improved front-end of a thd meter. This provides reduced internal distortion and less critical tuning of the notch.**

**Fig. 3a) Frequency response of the low, band-pass and high-pass outputs of a state variable filter with a Q of unity (Bode plot).**

**b) As a) but shown as an Argand or vector diagram.**
Fig. 5a) First stage response; span 0 to 5kHz, notch trace 10dB/div, low-pass trace 1dB/div.

b) Second stage of the improved THD meter.

Fig. 5b) First stage response; span 0 to 5kHz, notch trace 10dB/div, low-pass trace 1dB/div.

C.1N tunes from 2kHz down to HP3580A spectrum analyser

Fig. 6a) First stage of the improved THD meter.

b) Second stage of the improved THD meter.

ture from 180° of just 0.053°.

Also shown is the low-pass output at 1dB/div, showing the expected peaking of about +1dB. Note that the traces are offset so the 0Hz levels of the two traces do not correspond. Figure 5b) shows the same results, this time with the notch trace at 1dB/div and the low-pass at 10dB/div. Compared to the response at 5kHz and higher, it can be seen that the response at 2kHz is indeed 1.6dB down as predicted by theory.

Figure 6a) shows the response of the second svf stage, with its tuning set so that the notch again occurs at 1kHz. The stage’s resonant frequency is actually 1.29kHz, with around 10dB of peaking at the low pass output. This results in a smaller peak in the notch output, the response still being +1.6dB at 2kHz relative to the far-out high frequency response. When this is combined with the first stage response of Fig. 5, the result is a response which is level at 2kHz and upwards, Fig. 6b).

The notch should be 120dB or more deep, but as displayed it is limited to the spectrum analyser’s noise floor at -90dB ref. To reliably achieve 60dB or more suppression in each stage, putting the fundamental of the test sinewave below noise, a phase trim should be provided for each stage, similar to Fig. 7.

To complete the picture

A complete THD meter front-end must include facilities to accept inputs of various amplitudes, so some kind of input attenuator is required as indicated in Fig. 7. The potentiometer used should be a wirewound type, to avoid introducing noise. The twin gang tuning potentiometers should also be wirewound, for the same reason. Also shown is a ‘set level’ switch which permits meter deflection to be set to full scale on the incoming signal, before notchfing out the fundamental.

Figure 7 shows a fine tuning trim on the second svf stage. This optional trimmer assists in obtaining the maximum possible fundamental rejection. It becomes necessary if the twin gang tuning potentiometers for the two stages are ganged together. In this case, an ‘initial tune’ position is needed which permits tuning of the first stage for maximum rejection.

The second stage is now also approximately tuned due to the ganging. Final adjustment of the second stage frequency and phase trims completes the tuning. Dual gang 2kΩ wirewound pots are available to special order with the shaft extended at the rear, permitting the tuning of the two svf stages to be ganged.

Due to the 20dB gain stage between the first and second svf stages, the lower limit of the measurement range is set only by first stage noise. Should THD measurements exceeding 10% be required, provision must be made to switch the 20dB gain stage to 0dB.

Results shown in Figs 5 and 6 were taken using TL084 quad op-amps in circuits prototyped on bread boards. In the final design, these op-amps would be unsuitable, due to their wirewound type.

Available from Spectrol Reliance Ltd. Tel, 01793 521351, fax, 01793 539255.
Classic state-variable filter

This circuit diagram is the classic four op-amp state variable filter. There is a three op-amp variant, but this involves taking the damping term from the band-pass output back to the non-inverting input of IC1, resulting in a common mode component at the filter input.

The tuned frequency or maximum gain frequency at the band-pass output is given by \( f = \frac{1}{2\pi RC} \). At this frequency the high, band and low-pass outputs are all equal in amplitude, with the band and low-pass outputs lagging the high-pass output by 90° and 180° respectively.

With resistor values shown, low-frequency gain at the low-pass output and high-frequency gain at the high-pass output are both unity. Gain at the tuned frequency is numerically equal to the circuit Q, where \( Q = \frac{RQ}{100k} \). The transfer function is given by

\[
\text{numerator} \left( s^2 + Ds + 1 \right),
\]

where the numerator equals 1 for the low-pass output, \( s \) for the band pass and \( s^2 \) for the high-pass output, and \( D = \frac{1}{Q} \).

Now \( s \) is the complex frequency variable \( \omega \sin \theta \), but for the purposes of determining the steady state response of the circuit to sinewaves, \( s \) can be ignored. This leaves just \( \omega \) as the variable, where \( \omega = 2\pi f \) radians per second.

Things can be simplified even further by normalising the frequency, that is, simply assuming that whatever tuned frequency you are interested in is unity. Thus gain at the low-pass output is given by \( \frac{1}{\left( \omega - \frac{1}{\omega} \right) + D\omega} \). At the tuned frequency, where \( \omega = 1 \), this amounts to \( \frac{1}{1/D - \frac{1}{1/D}} \) where \( -j \) indicates a phase lagging 90° on the input. At the tuned frequency, if \( D = \frac{1}{2} \), ie \( Q = 2 \), the gain is \( x^2 + \omega + 6dB \) at 90° relative to the input.

Notch output of Fig. 4a) is obtained by summing the high and low-pass outputs. As a result, gain is given by

\[
\left( \frac{-\omega^2 + 1}{-\omega^2 + D\omega + 1} \right).
\]

Clearly, the numerator is zero when \( \omega = 1 \). On substituting \( D = 1 \) and \( \omega = 2 \), the gain turns out to be \( x0.83 \), or \(-1.6dB \). This is the response of the circuit at the second harmonic of a sinewave test signal, when the notch is tuned to the fundamental. In Fig. 4b), \( D \) is set to 0.35 \( (Q = 2.86) \) and the contribution from the low-pass output is reduced to 60%. Now, gain is given by

\[
\left( \frac{-\omega^2 + 0.6}{-\omega^2 + 0.35j\omega + 1} \right),
\]

resulting in the response illustrated in Fig. 6a). Note that in this case, though, \( \omega = 1 \) corresponds to 1.29 times the \( \omega = 1 \) of the first svf section. In other words, this is 1.29kHz for a 1kHz test signal, making the notches of the two stages coincident.

Sharpness of the notch in the circuit of 4a) can be found by a little judicious approximation of the expression

\[
\left( \frac{-\omega^2 + 1}{-\omega^2 + j\omega + 1} \right).
\]

Remember that \( D = 1 \) for this circuit. At \( \omega = 1 \), the numerator is zero and the denominator is \( j\omega \), or just unity – amplitude-wise.

If frequency is changed by 0.1%, the denominator is virtually unaffected. However, with \( \omega \) now equal to 0.999, the numerator becomes 0.002, or only 54dB down on the response at \( \omega = 0 \) or infinity. Thus for a fractional detuning from the notch of 8, the output rises from zero to 28%.

For reasonably accurate thd measurements even down to a modest 0.01%, the fundamental must be suppressed to 0.003%. As a result, the accuracy of tuning must be at least 0.0015%, or 0.015Hz at 1kHz.

their thd of around 0.003% typical. A better choice is the Burr Brown OPA2604 dual fet-input audio op-amp, with its 0.0003% typical thd figure.

Clearly, careful construction and screening between stages is necessary to achieve 120dB or more of fundamental suppression. Given this, the limiting factor on readings is likely to be noise and hum. The former can be reduced by a low-pass filter immediately preceding the measuring circuit. It should have switchable cut-off frequencies of say 200, 80 and 20kHz, and include a switchable high-pass filter with heavy attenuation at 50Hz for the 20kHz selection.

References

Fig. 7. Suggested complete front-end for an improved thd meter.
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Now applying a train of pulses, for example 10kHz for a second, to the Data Terminal Ready, DTR, line charges C2, sets the flip-flop and extinguishes D4, so that power is removed from the pc.

Mount D4 and R4 close together in a light-tight enclosure, D4 being a high-efficiency diode.

Torsten Martinsen
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Missing-phase shut-down for three-phase motors

Three-phase motors are only switched on when all phases are present, but there is always the chance that one phase might fail during operation. This circuit shuts the motor down to avoid burning out the other two phases and indicates the missing one, efficiently and less expensively than is usual.

Power for the motor comes via the contactor through the triac driver circuit, the green led indicating the fact. The three-phase supply is converted to 12Vdc by the diodes D4,5,6 and the BZV85C12 zeners, the three lines being presented as inputs to the diode And gate formed by D1,2,3, the output of which drives the triac gate only in the presence of all three phases.

If a phase fails, the triac does not fire and the control circuit is inoperative, the relevant led in the And gate indicating which phase is missing.

In the original circuit, the contactor is a 440V, 50Hz type; for a 220V coil, connect between yellow phase and neutral.

Porus M Mehta
Bombay
India

Less expensive than usual, but still efficient, this simple circuit automatically shuts down a three-phase motor if a phase fails.

Low-voltage audio power amplifier

Power amplifiers are, perhaps, more vulnerable than other circuitry; in particular, those working on low voltages using bipolar transistors seem to suffer the most. This design has power mosfets in a circuit adapted to reduce non-linearity in a common-source configuration, in which low supply voltage is feasible. Gate drive in the usual source-follower circuit must exceed the supply to obtain a rail-to-rail output. In this circuit, the common-source output stage is bootstrapped, very simply, so that the common-source square-law transfer characteristic is greatly linearised. In this case, the two transistors appear not to be purely feedback devices, but behave as variable attenuators and make gate impedance roughly inversely proportional to transconductance. A high-impedance driver produces a linear output at the expense of two diode drops, matching the performance of the bipolar devices used in emitter-follower output stages.

A Ziemacki
Rotherham
West Yorkshire

Mosfet power amplifier for use with low-voltage supplies. Power mosfet output devices are in a linearised common-source arrangement. Amplifier provides about 7W into 8Ω on a 12V supply.
Marconi distortion meter type TF2331 - £150. TF2331A - £0.000.
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Marconi RCL Bridge type TF2700 - £150.
HP Frequency Counter type 5340A - 18GHz 11000 - rear output £800.
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Marconi TF2370 as above but late type Brown Case - £1000.
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Marconi TF2370 - 30Hz-110Mc/s 750HM Output (2 BNC Sockets+ Resistor for 500HM MOD with HP5370A Universal Time Interval Counter.
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HP8182A Data Analyser.
HP251A Bus System Analyser.
HP725A Optical Power Unit 0-100mV - £000.
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HP737A Pattern Generator.
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HP6057A Distribution Amplifier.
HP6131C Digital Voltage Source + 100V A Amp.
HP7378A Primary Multiplier Analyser.
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HP8150A Optical Signal Source.
HP8151A Bus System Analyser.
HP8136A Universal Counter A-B.
HP8150A Optical Signal Source.
HP8151A Bus System Analyser.
HP8136A Universal Counter A-B.
HP8150A Optical Signal Source.
HP8151A Bus System Analyser.
HP8136A Universal Counter A-B.
HP8150A Optical Signal Source.
HP8151A Bus System Analyser.
HP8136A Universal Counter A-B.
NiCd battery discharger/charger

NiCd cells should be discharged to 1V: this circuit does that and subsequently charges the cells at the 100mA rate.

Reference for the first TL082 comes from R3 and is set to 0.7V, V1 being adjusted to \( V_{\text{bat}}/n \), where \( n \) is the number of cells. While \( V_1 \) is above 1V, the BUZ10 is on and the cells are discharging, but when \( V_1 \) decreases to below 1V, the circuit is latched off via \( D_2R_9C_1 \) and discharge is complete.

Switching in R10 begins the 100mA charge, the value of this resistor being \( (8-V_{\text{ref}})/0.1 \).

Adjustment of R6 allows the discharge of up to 30 cells, assuming R6 is man enough for the job and can be adjusted with sufficient accuracy. Discharge resistor R11 is \( 5\Omega \).

The circuit will work with up to 20 lead-acid cells by setting \( V_{\text{ref}} \) to 1.75V, R6 as needed and selecting R11 appropriately.

Ken Hughes
Wokingham
Berkshire.

Fast, precise pulse generator

Newport Components’s 31A5500 tapped, active delay line (from RS), together with advanced c-mos logic (ACL), eases the design of a pulse generator to produce 3ns transients and pulse-width accuracy within 2ns.

An input pulse longer than the required output, rising at better than 10ns/V, goes to one input of an XOR and to the other by way of the selected delay in the 31A5500, so that the gate output is high during the delay; the AND prevents anything further happening during the delay, after the input pulse has gone low.

Propagation delay from the leading edge of the input is accounted for by two ACL delays, amounting to 8ns, and the transient times are determined by the AND gate – around 3ns for ACL. A possible problem is that the output of the line is not much more than the specified high for ACL.

Instead of the delay line, the output of the line is not much more than the specified high for ACL. A possible problem is that the output of the line is not much more than the specified high for ACL.

Unused gates could be used to make up the delay, the six available contributing about 3.8ns per gate at 25°C, although the delay will not be as precise as with the line.

The result shown in the oscillogram was obtained by the circuit built on Veroboard with the ics in sockets.

Nick Wheeler
Sutton
Surrey.

Notes:
1. 0.02pF capacitor on each device socket
2. Pinout of 31A5500 (and other 31A delay lines) as follows:

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</table>

Delay line pulse generator gives 3ns rise and fall times and pulse-accuracy to within 2ns.

Narrowest pulse available is 5ns.
Narrow-band, voltage-controlled oscillator

As a tone decoder to detect a single frequency in the presence of interference, a narrow-band VCO in a phase-locked loop works well. Unfortunately, combining a narrow band and a well-defined centre frequency is not easy, particularly if initial tuning is to be avoided. A crystal oscillator is one method, its frequency being "pulled" by a Varicap, but the frequency sweep is only a few hundred hertz at frequencies in the 1-10 MHz range and at maximum deviation stability is poor. Ceramic resonators offer the advantage of wider frequency adjustment. The main diagram on the left shows a low-power 6 MHz oscillator using a CMOS 4007UBE IC with an RS type 656-215 ceramic resonator. The BB102 varicap provides adjustment capacitance C, the centre frequency being set by C1 (about 33 pF). Sweep range is around 50 kHz for a change of 33-220 pF, the characteristic being logarithmic to give a falling frequency with increasing capacitance. Sensitivity therefore depends on capacitance: -4 kHz/pF at 10 pF and -80 Hz/pF at 470 pF.

A cheaper alternative is to replace the Varicap with a rectifier diode, having some junction capacitance. A 1N4001 – as shown in the diagram contained in the block – provides a 4 kHz sweep at 6 V.

R G Harrison
Charvill
Reading.

Electronic lock relies on magnetic poles

Giving an unauthorised user no indication of how to use it, this lock uses a magnetic key inserted into a field of reed relays. Out of nine (or more) possible positions, only three contain the magnets to operate the relays, the three operated being effectively connected as an And gate. Inserting the key energises the output transistor and, therefore, the lock, but only if the remaining relays are not operated; if any of the others operate, being connected as an Or gate, the lock stays locked and the alarm sounds.

Ronny Tegel
Arlov, Sweden

Reed relays and their magnets in a random pattern form an electronic lock which also provides an alarm if the wrong key is tried.
Evaluate DSP for

Allen Brown has been looking at an evaluation kit with a price tag designed to bring DSP development within the reach of every engineer.

Digital signal processing, DSP, techniques have become very popular for solving a range of problems in electronics. Whether it is filtering, spectral analysis, noise reduction or information coding, there is a very good chance that DSP can offer a solution.

Having decided that DSP is an appropriate direction to find a solution, the next question centres on what hardware should be used. There's a number of processor manufacturers to choose from, and to make the choice easier, a number of companies now provide low-cost evaluation modules hosting a signal processor.

One such product is the DSP56002EVM from Motorola which sells for £90, exclusive, and provides a very low-cost gateway into practical DSP. Shown in Fig. 1, the DSP56002EVM comprises a 40MHz DSP56L002, which is a low-power version of the DSP56002, and 128kbyte of SRAM. It also has a Crystal Semiconductor CS4215 which is a 16-bit multimedia audio codec, and is fitted with three 2.1mm jack sockets for analogue I/O.

The board can be controlled directly from a PC via a serial link. It is provided with Debug-EVM software, allowing the user to monitor the internal registers and memory of the DSP56002 processor. Facilities for emending assembly code at the processor level are also provided, as are technical data sheets on both the DSP56002 and the Codec.

DSP56002 architecture

The processor is a fixed point device with two 56-bit accumulators, designated A and B. It has a dual data architecture - X and Y memory with individual buses - feeding general purpose registers X and Y. There are also eight address registers r7, each with auxiliary registers, m0, m1, n0, n1, for implementing address modifications.

Instructions are fetched on the program bus and with its hard-wired multiplier and arithmetic and logic unit the DSP56002 is able to perform multiply, add and dual channel data move operations in a single clock cycle.

Harvard architecture on-board memory comprises 512-by-24 bit program ram and 256 by 24bit ram for both X-data and Y-data. There is also a boot rom in the program memory area for allowing external code to be downloaded from a slow eprom during the boot-up phase. Both X and Y data spaces have rom areas dedicated to A-law/u-law and sine coefficients respectively.

Fig. 1. DSP56002EVM featuring a DSP56L002 and a CS4215 multimedia audio codec from Crystal Semiconductor. Using the evaluation module serves as an ideal method to become acquainted with the signal processor.

Evaluation kit hardware

Analogue I/O operations are performed on the CMOS CS4215 16bit stereo codec. It supports CD, FM radio quality music, telephone quality speech and modems. The device's a-to-d converters are 64-times oversampling delta-sigma types with on-chip filters which adapt to the sample frequency. The CS4215 allows integration of microphone, line-level inputs and I/O gain settings, along with headphone and monitor speaker drivers, resulting in a very small footprint.

Codec sampling is between 4kHz to 50kHz and the device can perform 16bit or 8bit audio data coding compression in either A or u-law. This makes more efficient use of the 16bit dynamic range by applying a gain that is dependent upon the instantaneous signal amplitude.
The limitation of the codec lies in its output sample rate of less than 50kHz. Although adequate for stereo audio needs, the evaluation kit would not be suitable for investigating higher frequency applications. Its appeal would probably be limited to first time users of dsp technology.

In addition to the 128kbyte of sram, the DSP56002EVM also has a space for an AT29C256PC flash eromp which can serve as program memory or as X-data memory for read only purposes. Motorola only provides holes for the eromp - not even a proper dil socket. It would have been far more useful if an eromp in a dil had been supplied.

Without the eromp, once the power to the module is switched off, the memory loses its contents - thus limiting its usefulness. Also, Motorola leaves room for a second nine-pin D-connector to enable a dumb terminal to be interfaced to the DSP56002EVM. Again the company could have been a little more generous and supplied the D-connector as standard, together with some software routines for running an external led display or dumb terminal.

Software for dsp evaluation
Communication with the DSP56002EVM is effected via an RS-232 serial link which accesses the OnCE facility on the main processor. The OnCE permits all the register contents to be interrogated and dumped via the serial link to the pc. Registers can also be modified, and as a result allow the user to emend software errors. The accompanying software - Debug-EVM - is a development system package which greatly facilitates software testing. It only runs under dos, but it can be called from within windows.

The other attractive feature of using the OnCE is that it negates the need for a monitor program to be resident on the evaluation module. This feature is common on evaluation systems hosting conventional microprocessors and occupies valuable memory address space.

A typical screen display of Debug-EVM is shown in Fig. 2. The display can show several fields, for example register contents, disassembled code, graphical i/o and memory contents. However, these do not operate as smoothly as windows equivalent using the graphics user interface (gui) standard.

Provided with the package is a DSP56002 assembler. Once a sources code is run through the assembler an executable file is generated which can be downloaded into the evaluation module's memory. The whole operation is quite painless.

When the Debug-EVM software is running the user has the option of executing a host of features normally found within microprocessor monitor systems, for example implanting break points and single stepping the code. Although acceptable if you are only running one evaluation kit, if you wanted to run a second, the limitations of the dos version of the Debug-EVM would soon become apparent.

**Worked dsp examples**
To run a number of the worked examples, the user is asked to provide a stereo music source - such as a Walkman - and a set of headphones. The worked examples given in the Quick Start document - comprising 16 pages of A4 - include a program for removing a 60Hz signal using a notch filter and implementing the codec process on audio signals.

There is also a program implementing a low-pass digital filter. These are quite useful as demonstrations and serve as good introductions to the DSP56002. Routines are also provided for driving the codec. There should be more examples available code. This deficiency is not helped by the absence of a credible user manual containing information on the kit's hardware features. Although there is some information on a disk readme file - by no stretch of the imagination would this pass as a user manual.

There is however a printed manual for the Debug-EVM software. Probably a great deal could be done with the kit but without the relevant hardware details the scope for development will be somewhat limited. The general impression is that it is very much intended for the engineer who already has a reasonable knowledge of microprocessors.

**Summary**
The DSP56002EVM gives instant access to the possibilities of the DSP56002. Its stereo channel analogue i/o is a very attractive feature which allows the processor to used in audio signal processing applications or for any processing requirements under 50kHz.

However, in view of the fact that many potential users probably would not have a knowledge in dsp, I feel that the product would be better served if provided with a comprehensive user manual. Having performed the exercises laid out in the Quick Start document you could be left wondering 'what do I do with it now?'

On the whole, I feel that a stand-alone module for a dsp chip is a good idea. But unless more is provided for the non specialist interested in investigating the possibilities of the DSP56002, I feel its appeal will soon run out.

---

**Availability**
DSP56002EVM costs £90, excluding VAT, and is available from Arrow Jermyn, St Martin's Business Centre, Cambridge Road, Bedford MK42 OLF, tel 01234 270027, fax 01234 214674.
### TRANSISTORS

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January 1996
Claus Kühnel describes how combining the Basic Stamp microcontroller and a TSL230 forms an extremely simple light-to-RS232 interface.

For silicon photo diodes, the detectable spectrum of light extends from about 300nm to 1100nm.

Short circuit current of the silicon photo diode is proportional to incident irradiation and almost independent of temperature. The downside is that photo diodes generally produce only a very small signal current. As a result, the analogue circuits for processing the diode output can become expensive.

Design costs reduce however when the photo diode and its amplifier are integrated into the same chip or module. In the case of the photo diode used here, not only are the diode and its amplifier mounted in the same module, but there is also digital circuitry, allowing the device to communicate directly with a microcontroller. In addition, the chip can operate from supply rails down to 2.7V.

Light sensing with the TSL230

The TSL230, from Texas instruments, is a programmable light-to-frequency converter. Both silicon photo diode and current-to-frequency converter are housed in a clear plastic dual-in-line package with eight-pins.

When the device is set for maximum sensitivity, an irradiation of 450pW/cm² (4 of 660nm) produces an output frequency of typically 1MHz. The light sensitive area of this photo diode is typically 1mm².

This photo diode is configurable. Its sensitivity can be increased by a factor of 10 or 100. Further adjustments can be made via programmable on-chip frequency dividers allowing divide by 2, 10 or 100 of the output.

In addition to control inputs S3 to SO there is an output-enable input. When active-low, this pin switches the output of the device into a tri-state condition, allowing outputs of several devices to be connected to a common line.

Programming conditions are listed in the table. The first line in the table is hatched since these conditions invoke a special mode.

Listing – PBASIC-source

```
[Constants]
symbol RxD  = 7
Symbol TxD  = 6
symbol TSL230 = 4
symbol baud  = 115200

[Variables]
symbol key  = pin5
symbol ss    = 50
symbol period = 1

[Initialisation]
dirs  = $0b111111
pins  = $00000101

[Main Code]
start: if key = 0 then enter
loop: pullin TSL230.0,period
serout TxD, baud,(period,10,13)
dump period
pause 1000
period = 0
goto start
enter: serout TxD, baud, ("SS?",10,13)
serin RXD, baud, #ss
ss = ss - $30 & $Of
dump #ss
pins = ss
goto loop
```

Basic Stamp is designed for ease of use. It has eight i/o lines and is programmed using a high-level Basic dialect. As a result, forming an interface to read data from the light-to-frequency converter and translating the reading to RS232 for use on a pc is very simple.

Frequency division causes a symmetrical output, i.e. one with a 1:1 duty cycle. While S3 and S2 are low, duration of the output pulse extends from 125ns to 500ns and symmetry is not defined.

The TSL230 programmable light-to-frequency is linked to a host controller by a simple RS232 interface and one command line. After this command line is pulled low, the Stamp asks for a command. This command sets inputs S3 to SO of the TSL230. Following setup, the Stamp sends the measured value to the host periodically.
When S1 and S0 are low, the device switches to the power-down mode. This reduces maximum supply current from 2mA to 10µA.

**Basic stamp as an RS232 interface**

In order to determine light level, the TSL230’s output frequency must be measured. Interfacing the TSL230 to a pc or controller allows the conversion and display to be carried out in software.

The simplest solution results from building this interface using a microcontroller best suited for this kind of application. On the Basic Stamp, there are eight free configurable i/o lines and the device is programmable in a Basic dialect known as Pbasic, which is an abbreviation of Parallax Basic. This language was developed especially for microcontroller applications.

Interfacing to the host computer is carried out via a simple three-wire connection according to RS-232 standards and an additional command line.

**Light detection software**

At the label 'start' in the program, the command line CMD marked 'key' is queried. If the i/o line is pulled low by the host controller a jump to the label 'enter' results.

The Basic Stamp sends the characters 'SS?' to the host and waits to receive a value to setup the control lines S3 to S0. After masking the result can be transferred to the host. The command line. Results from the command 'pulsin' are saved in the debug window of the development system, for all eventualities, and a possible display in the debug window of the development system, the control inputs are set up. A jump to the label 'loop' leads to the endless loop, where the program runs normally.

The low period of TSL230 output pulses is measured using the command 'pulsin'. To calculate output frequency from this low time the pulse sequence must be symmetrical. As a result, the hatched condition in the table is unusable.

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<th>S3</th>
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<td>f</td>
<td>L</td>
<td>L</td>
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<tr>
<td>L</td>
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The TSL230 light to frequency converter is programmable for both sensitivity and output frequency scaling via four programming inputs.

**Basic stamp output frequency vs radiance**

**Photodiode spectral responsivity**

Typical performance of the TSL230 light-to-frequency converter. Both are for 25°C. Conditions for frequency versus irradiance were 5V supply, $\lambda_0$ of 670nm and S3=S2=L.

**Distortion of the Trimodal power amplifier in its class-A mode at 20W into 8Ω.**

Printed circuit boards for Douglas Self’s Trimodal audio power amplifier - detailed in the June and July issues of EW+WW are available exclusively via EW+WWW. This amplifier can be switched between Class A/AB and Class B to provide remarkable performance over a wide range of operating conditions. In Class A it delivers up to 27W with ultra-low distortion. But presented with a low impedance, the amplifier has recourse to an unusually linear AB configuration.

Designed by Gareth Connor and supplied with a 12 page manual, the silk-screened boards are supplied in pairs at £49.48 per pair, fully inclusive of VAT and UK or overseas postage. Send a postal order or cheque payable to Reed Business Publishing to Trimodal Power, EW+WWW, room L333, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS, together with your address. Alternatively e-mail your address, credit-card number, credit-card type (i.e. Access/Visa) and the card’s expiry date to jackie.lowe@rbp.co.uk or fax the same details on 0181 652 8956.
**Display Electronics 1995. E & O E 6/5**

**19" RACK CABINETS**

Superb quality 6 foot 40U

Virtually New, Ultra Smart

Less Than Half Price!!

Top quality 19" deep x 40U high cabinets from Philips Enclosures Ltd (EFT0690).

Reliable and solidly constructed, substantial Racks are designed to be sturdy enough to support heavy, full height adjustable lockable back door and tops. They are fitted with a fully adjustable internal fixing Stud, ready punched for any manufacturer's choice.

A unique feature is that all models are supplied with a new, plus ready mounted integral 12 way 13 pin switch mounted mains distribution strip made from high quality materials. Some of the models shipped so far are:

- **MB 6U** (30U flat rack, £125.00)
- **MB 6U 3WAY** (30U flat rack, £135.00)
- **MB 6U 6WAY** (30U flat rack, £145.00)
- **MB 6U 9WAY** (30U flat rack, £155.00)

All of these are in three sections, with lockable side panels. A 31 pin Female Euro Plug at the rear makes these racks the most versatile of all!

As usual, we have ever racks. May be stacked side by side and therefore any number of side panels can be supplied.

Overall dimensions are: **776 x 632 x 22 W**.

Order as: **OPT Rack 1**, Complete with 3 side panels, £225.00 (G)

**OPT Rack 2**, Less side panels...

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**Low Cost RAM & CPUs**

**INTEL "ABOVE" Memory Expansion Board.** Full length PC XT compatible card (EFT0691). Card is fully tested for Expandable or Expanded (266) processor. On board 64K / 128K VIA expansion memory. Fully tested and guaranteed. Windows Compatible.

Over 1000 racks in all sizes 19" 22" 24" and 26" wide available to order.

**Call with your requirements**.

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**Low Cost RAM & CPU's**

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**Video Screen Monitor**

The ultimate in micro screen technology made by MicroTouch, but sold at a price below cost. The system consists of a micro processor, memory, software, a specially made character generator, a display screen and an interface.

A parallel data connection is made to the computer via a standard 25 pin D connector. The display screen is made from a LCD panel with a 640 x 200 pixel resolution and capable of 80 colour displays. The display is controlled by the computer via a simple serial data containing positional X & Y co-ordinates as to where a finger is touching the screen as the "pen", the data moves in one second.

The character generator is a set board of characters or "icons" that can be displayed on the screen. The resultant image is a 24 by 47 pixel resolution.
**NEW PRODUCTS CLASSIFIED**

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**PASSIVE**

Ceramic capacitors. Kyocera's new ceramic component offers high capacitance (up to 10µF at 16V), low equivalent series resistance and good hf characteristics. Insulation resistance is better than 100GΩ or 500MΩ·µF, whichever is least, and the units operate from -25°C to 85°C. The capacitors are in Y5V with nickel barrier terminations. AVX Ltd. Tel., 01252 770000; fax, 01252 770001.

Cermet trimmers. The Spectral Reliability model 761 is a 0.25in square, single-turn cermet trimmer, available in three pin styles for top or side voltage and resistance adjustability. Voltage and resistance adjustability is ±0.05% and ±0.15%. The trimmers have barrier terminations. AVX Ltd. Tel., 01344 853313.

Connectors and cabling

Adaptable cigarette-lighter plugs. Since there appears to be no standard for the diameter of cigarette-lighter plugs, all those that bear the label "cigarette lighter" are 0.75mm in diameter. The leads may be gold or silver plated and may be inserted into the contact socket "probably won't. Accordingly, Pedoka has made all-purpose plugs with a sliding mechanism so that they will fit into any socket. Pedoka Ltd. Tel., 01462 224243; fax, 01462 422253.

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**ACTIVE**

A-to-D and D-to-A converters

3V, 16-bit a-to-d. Analog Devices’ AD7715 is a 16-bit analogue-to-digital converter suitable for battery-operated equipment. It has a programmable-gain input. A three-wire serial interface reduces the number of interconnection lines and couplers for isolated systems, the input taking differential inputs. Gain, signal polarity and update rate are all controlled by software. Polar Electronics. Tel., 01525 377093; fax, 01525 378367.

Discrete active devices

Reliable s-m discretes. Transistors and diodes in the Rohm SC-59 series allow more reliable automated manufacture than the SOT-23 devices. The range consists switching transistors and diodes, digital p-n-p and n-p-n types with built-in resistors and both transistor and diode arrays. Although the devices are direct replacements for the SOT-23 family, they have improved structures to reduce thermal stress to the bond wires and reduced vulnerability to humidity, with better power dissipation and improved alignment on the board. Also in stock is 1.6mm square EMS series. Polar Electronics. Tel., 01525 377093; fax, 01525 378367.

Logic

3.3V, low-eml clock drivers. SSL306/0/8/318/368 comprise AMCC's family of 3.3V clock drivers to meet the requirements of logic running at up to 100MHz. 306/0 provides 10 outputs at half fM and 10 at fM, 308 and 318 give 20 and 30 outputs respectively at fM, while 368 gives six at fM and eight at 0.5fM synchronously; all outputs at fM asynchronously. Output drivers provide 24mA, dropping to 4mA when the output reaches 0.8V, followed by a slower transition to 0V. This "virtually eliminates" ground bounce and any resulting emi. AMCC Inc. Tel., 001 619 450 9333; fax, 001 619 450 9885.

Microprocessors and controllers

Stepper drivers. Sanyo Densi offers the PMM8723/14 for driving two-phase and five-phase stepper motors. They are cmos lics and each, with a pulse oscillator and power switching transistors, forms a complete drive. Both have an excitation mode changeover terminals for different phase excitation. Power supply is 5V for the PMM8723 and 4.18V for the 8714. EAO-Highland Electronics Ltd. Tel., 01444 236000; fax, 01444 236641.
resonator for the 300-360MHz range of comms systems, ecc oscillators and rf modulators. Frequency tolerance is ±100ppm and series resistance 3Ω. The device measures 1.5 by 4.8 by 5.2mm and is suitable for automatic assembly. Advanced Crystal Technology. Tel., 01635 528520; fax, 01635 528443.

Displays
Super thin. Hitachi announces a new 'super-thin' lid technique that produces a 70° angle with no colour shift or change and is forecast to compete with 90° when several viewers are present. The technique used is in-plane switching (IPS), in which the liquid-crystal molecules switch the their longitudinal axis parallel to the substrate. First product to use this method will be a 262cm, 13.3in 1024 by 768 pixel display. Hitachi Europe Ltd. Tel., 01628 358163; fax, 01628 555160.

SVGA colour lcds. A new screen size of 11.5in is introduced by Sharp, with 200 by 200 by 6mm thickness and SVGA-compatible colour displays. Power consumption and weight are both less than in the current 10.4in types used in notebooks and they are more compact designs. Hero Electronics Ltd. Tel., 01525 405015; fax, 01525 402383.

Filters
Motor filters. Roxburgh offers the MF range of motor inverter filters with the aim of reducing noise from inverters using long runs of cable. Single-phase types handle 3-32A, the three-phase versions coping with 3x10-32A, which can take one or two 4-180A. They meet EN55022B requirements, with over 80dB of noise reduction at 1310nm is 31dB. Tektronix spectrum analysis. Marconi Instruments has a new range of optical-interferometric and microwave spectrum analysers. The 2390E consists of the 2393 for 9kHz-2.9GHz and the 2392 for 9kHz-2.5GHz. All three have built-in am/fm receivers and a 1Hz resolution frequency counter to allow the identification of interfering signal, and resolution bandwidth of 3Hz-30MHz enable signals from a large range of equipment to be examined. There is a 2.9kHz tracking generator for response measurements and an optional quasi-peak detector and filters are provided for emc testing. Marconi Instruments Ltd. Tel., 01436 742200; fax, 01436 726701.

Oscilloscope calibration. For both analogue and digital oscilloscopes, a Switchable filter. Kemo's VBF19 is an adjustable band-pass/band-stop filter frame which can take one or two channels, each of which is mode-selectable. Frequency is adjustable by front-panel rotary switches from 0.01Hz to 99.9kHz to a resolution of three digits. In band-stop mode, it gives a notch filter response with theoretical ripple of ±0.1dB in upper and lower pass-bands, response being 18dB down at 1/2 octave points and -50dB at 2% from centre frequency. Kemo Ltd. Tel., 0181 664 3288; fax, 0181 558 4084.

Test and measurement
Photometers. Tek's J86 range of hand-held digital photometers is now available. A range of pre-calibrated heads, automatic units selection, scaling and zeroing allow the measurement of chromaticity, luminance, illuminance, radiance, irradiance and led output. When used with the J810 chromaticity head, the instrument measures, in real time, colour for matching and balancing in television studios and monitor manufacture. Thurlby Thandar Instruments Ltd. Tel., 01480 412451; fax, 01480 450405.

Spectrum analysis. Marconi Instruments has a new range of optical-interferometric and microwave spectrum analysers. The 2390E consists of the 2393 for 9kHz-2.9GHz and the 2392 for 9kHz-2.5GHz. All three have built-in am/fm receivers and a 1Hz resolution frequency counter to allow the identification of interfering signal, and resolution bandwidth of 3Hz-30MHz enable signals from a large range of equipment to be examined. There is a 2.9kHz tracking generator for response measurements and an optional quasi-peak detector and filters are provided for emc testing. Marconi Instruments Ltd. Tel., 01436 742200; fax, 01436 726701.

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Fluke's 5500A-SC represents a new facility for use with the 5500A Multi-product calibrator, which itself handles thermometers, power meters, power harmonic analysers and other instruments. 5500A-SC is a levelled sine generator up to 1250kHz for bandwidth verification, a square wave for calibrating voltage gain, a low rms-time pulse generator for pulse response and a time-marker for time-base calibration. This is a single board that plugs into an internal slot in the 5500A. Fluke UK Ltd. Tel., 01923 240511; fax, 01923 225607.

Screened rooms. For emc testing, Seaward has modular screened rooms to provide fully rf shielded and controlled environments for compliance testing. The rooms can be stand-alone or linked to existing buildings. Features include minimal cavity resonance, supply filtering, the ability to take heavy equipment and a removable panel for rf penetration. As options, there are radio absorbent material to give a uniform field for radiated immunity test, are conditioning with waveguide protection, a revolting test table and cctv monitoring from a separate room. A larger version has a separate instrument lobby. Seaward Electronics Ltd. Tel., 0191 586 3511; fax, 0191 586 0227.

Otdr plug-ins for FiberMaster. Tektronix's FG Series plug-in modules for the TFP2A or FiberMaster optical time-domain reflectometer are meant for use in cabling, cable head-end facilities and cables. FG1300 works at 1310nm and the FG1315 at both 1315 and 1550nm, both being single-mode and having temperature characteristics for indoors or temperate outdoor weather. Dynamic range at 1310nm is 31dB. Tektronix UK Ltd. Tel., 01628 403300; fax, 01628 403301.

Interfaces
PCMCIA buffer. Elan has the B158 Buffer Card, which provides a simple interface between external circuits and the PCMCIA slot. It allows high-speed data transfer with no signal conditioning or control, providing an interface for portable applications, with eight bidirectional data lines, 15 address lines, memory read/write, i/o read/write signals and a single interrupt line. In-line resistive terminations reduce cross talk and reflection, with a view to reducing emc. A screened mini-iv connector is used. Elan Digital Systems Ltd. Tel., 01444 236000; fax, 01444 554171; fax, 01444 577516.

and the latest edition includes more than 1500 new products. Two of the new sections are Navigation, which includes GPS systems, and Education, including lasers, meteorology and solar power. Equipment from resistors to computer hardware is to be found in these pages. Mps Electronics. Tel., 01702 554171; fax, 01702 553305.

Keyswitches. Low-profile keyswitches are described in a new brochure by Cherry. The Mx switch is designed to give the full 3mm travel preferred by touch typists on standard keyboards. Cherry Electrical Products Ltd. Tel., 01582 783100; fax, 01582 768883.

Relays. Matsushita has a catalogue of relays, including power, signal, PhotoMOS, time delay, safety, automotive and surface-mounted types, with some ic modules. Highlights include the AOYD12 SOP, said to be the world's smallest semiconductor type at 2.1mm high, and the TX2, which is rated at 60W and 220V and 2A and which withstands a 2500V rms surge. Matsushita Automation Controls Ltd. Tel., 01908 231555; fax, 01908 231599.

Optical encoders. In 24 colour pages, Grayhill's new brochure describes the Series 61 family of rotary optical encoders, including a 128 cycle/rev range. The publication contains advice on applications, describes the quadrature decoding
Materials
Solders and adhesives. Multicore has a new range of low-residue solder pastes needing no cleaning and high-performance epoxy adhesives for surface-mounted components.
Multicore AGC3 paste leaves a little clear residue and provides a good print definition to 16mil pitch at printing speeds up to 100mm/s in air or nitrogen. Multicore SA-35 adhesive has a rapid cure for high-speed dispensing at over 16,000 dots/hour; it has high insulating resistance and low dielectric constant to make it electrically invisible after curing for 45-60s at 130°C. Flint Distribution. Tel., 01564 777169.

Production equipment
Fluid dispensers. For the accurate application of flux, masking agents, solvents and water-based materials, Intertronics supplies pens and refillable bottles in the Finair Flow-Seal range which are provided with spring-loaded nibs and a valve to control the amount of material and prevent evaporation. Nibs are available in chisel, bullet and pointed shapes in acrylic and in a polyester chisel shape. Intertronics Ltd. Tel., 01865 842842; fax, 01865 842172.

Power supplies
2V dc-to-dc. Ericsson’s PKG 4310 Pl is a new member of the PKG series of dc converter modules. This one offers up to 75% efficiency at 30W output, causing only 30°C case temperature rise at 1ms/air velocity, which reduces the need for heat sinking, although there is thermal protection built in. Case size is about 3in by 2in by 0.43in and the dual-in-line pin layout allows mounting on 1cm centres. Output is adjustable, there is a remote on/off function and with a current limitation function to avoid discharging batteries during mains failure. Ericsson Components AB. Tel., 01739 483300; fax, 01739 483801.

Crt supply. Cathode-ray tube supply M1S302 by Farrel Hivolt has been redesigned as the CRM180 to give better performance and efficiency. Output voltage is now 18kV at 300A, varying by less than 50V for 100μA load change at 180°C. An additional auxiliary output is now standard and tracks the 18kV supply. The unit is controllable by a signal from a d-to-a converter or op-amp, by an internal or external reference and potentiometer or by fixed resistors. Custom versions can be supplied. Farrel Hivolt Ltd. Tel., 01234 841888; fax, 01234 824698.

720W in a 3U rack. Melcher has the PKS family of switching regulators, from which is available up to 720W of output at 5V-36V, with inputs up to 144Vdc and with no additional heat sink or air cooling. Features include continuous short and open-circuit protection, sense line(s), true current sharing for parallel operation, inhibit and continuous output adjustment to 42.5V. The devices come in both chassis and 19in rack form. Melcher Ltd. Tel., 01425 474755; fax, 01425 474765.

Radio communications products
Tunnel-diode detectors. A family of tunnel-diode detectors in the ACT1500 range from Anglia exhibits output voltage stability of, typically, ±0.15dB over the −65°C to 100°C temperature range. The family operates in bands between 100MHz and 18GHz and versions are made for use in applications from broad or narrow-band ecm receivers to low-noise video amplifier inputs. Output impedances are between 75Ω and 125Ω and the square-law range is 34dBm, extended to 38dBm for high load selection. No bias is needed. Anglia Microwaves Ltd. Tel., 01277 630000; fax, 01277 631111.

Transducers and sensors
Optical encoder. Grayhill’s 610 series panel-mounted industrial optical encoder has a life span of over 105 cycles, 2.5us and 4.5us turn-on and turn-off times, position and direction of rotation outputs. Multiple, concentric versions, a type with one code change per revolution, extended temperature working, military specification, and various sealing and switching options are offered. Roxburgh Electronics Ltd. Tel., 01724 281770; fax, 01724 281650.

Switches and relays
Sil relays. From Astralux, the Series 160 SIL relays come in a transfer-moulded, e-packages occupying only 5 to 19.5mm of board space and 4-pin solder or socket mounting. Normally open and nc contacts are available, energised by 5V, 12V and 24V, handling 100Vdc, 0.5A and 10W switching capability. Astralux Dynamics Ltd. Tel., 01430 240255; fax, 01430 255657.

Microswitches. The D2F range of pcb-mounted microswitches for computer mice is now extended for other uses. Measuring 12.7 by 9.8 by 5.5mm, the switches handle up to 125Vac or 30Vdc at 3A and actuators are of pin plunger, lever or roller form as standard, other types being available. A quick reverse-action mechanism gives high speed. Citron Electronics Ltd. Tel., 0181 420 4646; fax, 0181 450 8087.

Sil changeover. A miniature, single in-line changeover switch for pcb mounting, the SECME M212 measures 10 by 2.5mm, standing 6.4mm off the board and can be mounted on a 2.54mm matrix. Contacts are gold-plated, the base is sealed and feet allow it to withstand wave soldering. EAO-Highland Electronics Ltd. Tel., 01444 236000; fax, 01444 236641.

Navigation systems
Pic, counters, a digital storage rotational angular-velocity sensor by Murata, intended for use in navigation, location systems and satellite antenna positioning. A triangular vibrating prism uses the Coriolis effect to provide output from two piezoelectric sensors mounted on the prism and driving a differential amplifier. In this way, output is much higher than that normally obtained from piezo vibratory instruments. The unit copies with seven direction changes per second at a maximum angular velocity of 100°/s at 25°C, producing a ±2Vdc swing. Supply is 5V at 15mA, Murata Electronics (UK) Ltd. Tel., 01252 811666; fax, 01252 811777.

Board-level products
Virtual laboratory. Multiple Instrument Station by ABI is a software/hardware package to emulate six measuring instruments on a pc, counters, a digital storage oscilloscope, direct-voltage probes, function generators, programmable analogue outputs and a power supply. It fits the 5.25in drive bay, so that connections are to the front panel; one expansion slot is needed. The software works under Windows and provides familiar-looking instrument panels and controls. ABI Electronics Ltd. Tel., 01226 350145; fax, 01226 350483.

Computer
January 1996 ELECTRONICS WORLD
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Computers

Pentium card. IMS introduces a new plug-in Pentium cpus card, the PCA-167V, and ISA/PCIbus unit to be used in a passive backplane or as an embedded controller. The board has temperature sensing and overheating alarm outputs. It is based on the Triton chipset, taking cpus working at frequencies up to 150MHz. Features include both ISA and PCI local bus, PCI SCSI-II interface, two PCI Enhanced IDE hard disk interfaces, two floppy interfaces, two RS-232 Interfaces with high-speed buffers and a bidirectional parallel port. There is also 256/512K of extra cache and the board can accommodate up to 128MB of memory. Integrated Measurement Systems Ltd. Tel., 01703 771143; fax, 01703 704301.

Pentium motherboard. From the Apricot subsidiary of Mitsubishi comes the Diamond PCI/ISA motherboard, based on the Triton chipset. It is intended for "cost-sensitive" pc systems, but can be configured as a Pentium motherboard. From AMC, the PC/lntel compatible AMC-490 is an all-in-one single-board 486 (up to DX4/100) computer with an on-board svga controller, PCI-bus and ISA-bus support. Display-intensive work is eased by the use of a Trident TGUI9440 chipset with 1 or 2MB of video memory, allowing 32-bit graphics at up to 33MHz, and by the standard Feature Connector. There is also a high-speed local bus IDE controller supporting modes 3 and 4 hard disks to enable data transfer at up to 11Mbs. Four IDE devices, including large hard disks, cd-rom drives, tape and other types may be connected. Features include two RS-232 serial ports and a bi-directional parallel port and a floppy controller. Smm sockets can take up to 256Mb of dram. Advanced Modular Computers Ltd. Tel., 01753 580660; fax, 01753 580653.

Data acquisition

Transducer control panel. Amplicon Liveline’s EX206 transducer excitation and signal-conditioning panel provides current or voltage for up to 16 channels and is intended to work with the PC220 expandable a-t-d board. This newest 200 series board offers a wider range of excitation and software programmable analogue threshold trigger. Each input is configurable for 2, 3 or 4-wire ndt, 4-20mA, solid-state temperature sensors, voltage-excited 2-wire sensors and strain gauge transducers. Amplicon Liveline Ltd. Tel., 0800 525 335 (free); fax, 01273 507215.

Data communications

PC serial interface. IMS offers the PCI-740, a high-speed serial port interface card, which can be switched between RS-232,422,485 or current loop. It has a 16C550 uart with an on-chip fifo buffer to reduce processing load, particularly in Windows applications. In RS-485 mode, a network of serial devices can be built over distances of 1200m with only two wires, the card sensing direction of incoming data and switching transmission to suit. The card comes with a suite of software for programming and debugging, together with high-level language drivers for most of the popular development languages. Integrated Measurement Systems Ltd. Tel., 01703 771143; fax, 01703 704301.

Software

Heart education. Guildsoft has two packages on cd-rom designed to teach the basics of the heart. SmartHeart Teacher, in addition to the information on the right, contains all the echocardiogram images and cardiac data and output of related information. Output can be faxed or sent by modem to a physician. Guildsoft Ltd. Tel., 01752 895100; fax, 01752 894833.

Faster pcb design. Computation’s Vutrax pcb design package is now in versions 1.6 and 1.7, in which double the previous speed and many extra features are offered for the same cost as earlier versions. Features include the provision of software in 32-bit or 16-bit form, as appropriate for the work in hand; Improved Windows presentation; more checking facilities; a new extra and options config. files to include support for controllers in the 86- by-400 modes for best trade-off between speed and resolution for dos users. Computation Systems Ltd. Tel., 01525 378939; fax, 01525 850489.

New catalogue

This last minute entry is SSI’s 1996 catalogue featuring new additions including wide range of computer cables, connectors and assemblies, loudspeakers, PA/disco equipment, instruments and hundreds of component-level products. SSI, Tel., 0181 6431126; fax 0181 6433937.
A sophisticated, linear stereo amplifier and audio front end. All circuits and breadboards are designed and manufactured in the Crow's Nest. Features: Hart Super Audiograde Silver Solder. Also includes: HC80 Replacement Stereo Cassette Head. Suitable for use with all models of Hart Audio Cassette Heads. Soldering: The modern technique of soldering components without test equipment. Ideal when fitting new heads. A professional quality, digitally masterd test tape at a price anyone can afford. Test Cassette TC1D. Our price only £9.99.

**SOLDERING**
The size of modern components makes the task of soldering essential for good results. Everything we offer is actually used in our own workshops. See our list for the full range. 85-3-93110 1009 Reel Superfine 24swg for ultra precise control and easy working. Not only does it give beautiful results but it is designed to melt at normal soldering temperatures (around 240°C) but it has the added bonus of the ability to apply the heat precisely where it is needed. For special high temperature irons. A very low residue flux makes possible the neatest joints easy but eliminates the need for board cleaning after soldering.

**HART SUPER AUDIOGRADE SILVER SOLDER.**
Solder is an integral part of any electronic device. Without it, the components would simply fall apart. Hart Silver Solder has been specifically formulated for use with all models of Hart Audio Cassette Heads. Not only does it give beautiful results but it is designed to melt at normal soldering temperatures (around 240°C) but it has the added bonus of the ability to apply the heat precisely where it is needed. For special high temperature irons. A very low residue flux makes possible the neatest joints easy but eliminates the need for board cleaning after soldering. 85-3-93110 1009 Reel Superfine 24swg for ultra precise control and easy working. Not only does it give beautiful results but it is designed to melt at normal soldering temperatures (around 240°C) but it has the added bonus of the ability to apply the heat precisely where it is needed. For special high temperature irons. A very low residue flux makes possible the neatest joints easy but eliminates the need for board cleaning after soldering.

**HART AUDIO KITS -- YOUR VALUE FOR MONEY ROUTE TO ULTIMATE HI-FI**

Hart Audio Kits and factory assembled units use the unique combination of circuit designs by the renowned John Linsley Hood, the very best audio components, and our own engineering expertise. To give you unobtainable performance and unbeatable value for money. We have been teaching the art of hi-fi for over 20 years and have built a body of professional standards, even in the fashions we were using easily assembled printed circuits when Heathkit in America were still using tagboards! Many years of experience and innovation, going back to the early Discos and Bakelite chassis makes us incomparable design back- ground in the needs of the home constructor. This simply means that building a Hart Kit is a real pleasure, resulting in a piece of equipment that not only gives you money but you will be proud to own. Why not buy the报复和导线图解手册 for the kit you are interested in and see how easy it is to build your own equipment in the Hart way. The FULL cost can be credited against your subsequent kit purchase.

**K1100 AUDIO DESIGN 80W POWER AMPLIFIER.**

The fantastic John Linsley Hood designed amplifier is the flagging of our range, and the ideal powerhouse for your ultimate hi-fi system. This kit in your way to get the performance at bargain basement prices. Unique design features such as fully FET stabilised power supplies give this amplifier World Class performance with crystal clarity and transparency of sound, allied to the famous Hart quality of components and ease of constitution. Useful components are a brand new LED power meter and a versatile passive phase input and switching inputs, with ALPS precision Blue Velvet tone-towel volume and balance controls. Construction is very simple and straightforward, even by begin- ner standards, even in the sixties we were using easily assembled printed circuits when Heathkit in America were still using tagboards! Many years of experience and innovation, going back to the early Discos and Bakelite chassis makes us incomparable design background in the needs of the home constructor. This simply means that building a Hart Kit is a real pleasure, resulting in a piece of equipment that not only gives you money but you will be proud to own. Why not buy the报复和导线图解手册 for the kit you are interested in and see how easy it is to build your own equipment in the Hart way. The FULL cost can be credited against your subsequent kit purchase.

**CHIARA** SINGLE ENDED "A" CLASS HEADPHONE AMPLIFIER.

This unit provides a high quality headphone output for 'stand alone' use or to supplement those many power amplifiers that do not have a headphone facility. Easily installed with special link-though feature the unit will have the power of our new Andante Ultra High Quality Is- land tone-mid speaker. However in the black finished, Hart minicase it flashes the wider frequency response, low distortion design, and overall HI-FI 'musicality' that one associates with design from the renowned John Linsley Hood. The termination interconnecting leads and PCB mounted sockets prevent supply polarity reversal and on-board diagnostics provide visual indication of failure by means of a visible voltmeter. Volume and balance controls are ALPS "Blue Velvet" components. Very easily built, even by begin- ner standards, even in the sixties we were using easily assembled printed circuits when Heathkit in America were still using tagboards! Many years of experience and innovation, going back to the early Discos and Bakelite chassis makes us incomparable design background in the needs of the home constructor. This simply means that building a Hart Kit is a real pleasure, resulting in a piece of equipment that not only gives you money but you will be proud to own. Why not buy the报复和导线图解手册 for the kit you are interested in and see how easy it is to build your own equipment in the Hart way. The FULL cost can be credited against your subsequent kit purchase.

**"Andante" SERIES 20VA AUDIOPHILE POWER SUPPLIES**

Specially designed for each audio unit using audio-grade mini- mum parts, high bias and total freedom from mechanical noise this unit is a logical development from our highly successful 150 series. Utilising linear technology throughout for smoothness and musically it matches the perfect power supply for any audio amplifier requiring fully stabilised 575v supplies. Two versions are available, K3565 has a 25v supplies and a single 15v for relays etc. and can be used with our K1450 preamplifiers and our K1400 preamp. A more useful mod.vous soon to be introduced. The K3565 is identical in appearance but only has the 25v-15v current supply unit for use with the K1450 RPS head and "Chiar" headphone amplifier. K3565. Full Supply with all outputs: £39.75. K3565 Supply for K1450 only: £50.42.

**ALPS "Blue Velvet" PRECISION AUDIO CONTROLS.**

Now you can throw out those nasty B individuals carded pots and switches. This fantastic John Linsley Hood designed amplifier is the flagship of our range and the ideal powerhouse for your ultimate hi-fi system. This kit in your way to get the performance at bargain basement prices. Unique design features such as fully FET stabilised power supplies give this amplifier World Class performance with crystal clarity and transparency of sound, allied to the famous Hart quality of components and ease of constitution. Useful components are a brand new LED power meter and a versatile passive phase input and switching inputs, with ALPS precision Blue Velvet tone-towel volume and balance controls. Construction is very simple and straightforward, even by begin- ner standards, even in the sixties we were using easily assembled printed circuits when Heathkit in America were still using tagboards! Many years of experience and innovation, going back to the early Discos and Bakelite chassis makes us incomparable design background in the needs of the home constructor. This simply means that building a Hart Kit is a real pleasure, resulting in a piece of equipment that not only gives you money but you will be proud to own. Why not buy the报复和导线图解手册 for the kit you are interested in and see how easy it is to build your own equipment in the Hart way. The FULL cost can be credited against your subsequent kit purchase.

**HART TECHNOLOGY LIBRARY**

NEW Another Classic by John Linsley Hood. "AUDIO ELECTRON- ICS." Following the enormous ongoing success of his "Art of Linear Electronics," now entirely re-written by the master himself. Underlying audio techniques and equipment is a world of electronics that determines the quality of sound. For anyone involved in design, adapting, auditing or using digital or analog audio equipment understand- ing of electronic theory is of greater concern than the reproduced sound. The subjects covered include tape recording, tuners, power output stages, digital test instruments and interconnect crossover systems. John's lifetime of experience and personal innovation in this field place him to apply the gift of being so familiar with the subjects he can write clearly about it and make it both interesting and comprehensible to the reader. Contains 240 pages and over 250 line illustrations this new book represents great value for money at only £18.99.


**SPECIAL OFFER:**

**PRECISION Triple Purpose TEST CASSETTE TC1D.**

Are you sure your tape recorder is set up to give its best? Our latest triple purpose test cassette heads make a perfect test tape at a price anyone can afford. Test Cassette TC1D. Our price only £9.99.

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WANT NEW Another Classic by John Linsley Hood. "AUDIO ELECTRON- ICS." Following the enormous ongoing success of his "Art of Linear Electronics," now entirely re-written by the master himself. Underlying audio techniques and equipment is a world of electronics that determines the quality of sound. For anyone involved in design, adapting, auditing or using digital or analog audio equipment understand- ing of electronic theory is of greater concern than the reproduced sound. The subjects covered include tape recording, tuners, power output stages, digital test instruments and interconnect crossover systems. John's lifetime of experience and personal innovation in this field place him to apply the gift of being so familiar with the subjects he can write clearly about it and make it both interesting and comprehensible to the reader. Contains 240 pages and over 250 line illustrations this new book represents great value for money at only £18.99.

APPLICATIONS

High permeability cores for emi filtering

Detailed in a technical bulletin from Allied Signal is a highly permeable core designed with a highly flat or 'sheared over' dc hysteresis loop. High permeability, typically greater than 90,000 at 1kHz, 2mA/cm, makes these cores particularly effective for noise suppression applications, such as electromagnetic interference filtering. Major benefits include high attenuation for excellent suppression of electromagnetic noise and low profile. Weight and volume reductions of up to 50% are possible and core loss is low. Other applications include high accuracy current and pulse transformers as well as ground fault protection devices.

Square-loop cores for magnetic amplifiers

Square loop cores detailed in a further Technical bulletin from Allied Signal claim to be able to operate at higher frequencies than previously possible. In addition, they are said to enable magnetic amplifiers to be made with unparalleled precision and output regulation efficiency. Magnetic amplifiers can be used for outputs with currents of 1A to several tens of amps but they are also used at lower currents, where tight regulation and efficiency are important. Conventional regulated outputs are limited at higher frequencies and output currents, and linear regulators are inefficient. Independent switch-mode sub-regulators avoid the inefficiency, but they also require more complex and expensive circuitry relative to a magnetic amplifier.

Power controller designed with emc in mind

A power controller designed specifically to comply with the soon to be introduced emc regulations is described in an editorial feature from Sutronics. The BFM-TH burst firing trigger circuit requires only the inclusion of a 100kΩ potentiometer and suitable triac to give a compact versatile power supply. Unlike phase-angle controllers and electromechanical contactors, the device generates virtually no radio-frequency interference. This is a result of switching occurring when the supply waveform is within ±5V of the zero-crossing point. Also, because the switching element is a triac, there are no moving contacts and as a result no arcing. Due to the controller operating on a burst-fire principle manual control of resistive loads such as hot plates, ovens etc can be achieved. Sutronics, 62 Park Road, Swanage, Dorset BH19 2AE, tel/fax 01929 426400, email: sutronics@tcp.co.uk.
90% smps efficiency using flyback topology

Data sheet PWR-TOP200-4114 from Power Integrations details the operation, device characteristics and application issues of the company’s three-terminal off-line pulse-width modulation switch family.

TOPSwitch is a self biased and protected current-to-duty cycle converter with linear control and an open drain output. High efficiency is achieved through the use of mosfets combined with high integration. Integration eliminates the external power resistors normally needed for current sensing and/or supplying initial start-up bias current.

This three pin device implements buck, boost, flyback or forward topology and will easily interface with both opto and primary feedback. Supporting continuous or discontinuous operating modes, it is intended for 100, 110 or 230V ac off-line power supply applications in the 0-100W range. The device can also be used for 230/277V ac off-line power factor correction up to 150W.

Primary feedback regulation

Figure 1 is a simple 5V, 5W bias supply based on the PWR-TOP200. This universal-input flyback power supply employs primary-side regulation from a transformer bias winding. Line and load regulation of ±5% or better can be achieved from 10 to 100% of rated load.

Voltage feedback is obtained from T1’s bias winding, eliminating the need for an opto-coupler and secondary-referenced error amplifier. High-voltage dc is applied to T1’s primary winding. The other side of the transformer primary is driven by the integrated high-voltage mosfet transistor within the PWR-TOP200.

The clamp circuit implemented by VR1 and D1 limits leading-edge voltage spikes caused by transformer leakage inductance to a safe value. The 5V power secondary winding is rectified and filtered by C2, C3 and L1 to create a 5V output voltage.

Output of the T1 bias winding is rectified and filtered by D2, R1 and C4. Voltage across C4 is regulated by U1, and is determined by the 5.7V internal shunt regulator at the control pin of U1.

When rectified bias voltage on C5 begins to exceed the shunt regulator voltage, current flows into the control pin. Increasing control pin current decreases the duty cycle until a stable operating point is reached.

Output voltage is proportional to bias voltage by the turns ratio of the output to bias windings. Capacitor C5 is used to bypass the control pin. It also provides loop compensation for the power supply by shunting ac currents around the ‘control’ pin.

Dynamic impedance. In addition, it determines the auto-restart conditions.

Boost PFC pre-regulator

As a fixed frequency, discontinuous mode boost pre-regulator, the TOPswitch can be used to improve power factor and thd in applications such as power supplies and electronic ballasts. Figure 2 operates from 230V ac, delivering 70W at 430V dc with typical power factor of over 0.98 and a thd figure of 7%.

Bridge rectifier BR1, full wave rectifies ac input voltage. A boost power stage comprises L1, D1, C4 and the TOPswitch. Diode D2 prevents reverse current through the device’s body diode due to ringing voltages generated by the boost inductance and parasitic capacitance.

Resistor R1 generates a pre-compensation current from the large filter, C3. This prevents an averaging effect which would increase thd. Capacitor C1 filters high frequency noise currents to prevent errors in the pre-compensation current.

When power is first applied, C3 charges to typically 5.7V before the TOPSwitch starts. It provides bias current until the output voltage becomes regulated. When this occurs, series connected zener diodes VR1,2 begin to conduct. They drive current into the control pin, and directly control the duty cycle.

Capacitor C2 together with R2 performs low-pass filtering on the feedback signal to prevent output line frequency ripple voltage from varying the duty cycle.

Circuit performance:

- Load Regulation ±4.5%
- Line Regulation ±1.25%
- 95 to 370 V DC
- Ripple Voltage ±25 mV

Resistor R2 decouples pre-compensation current from the large filter, C3. This prevents an averaging effect which would increase thd. Capacitor C1 filters high frequency noise currents to prevent errors in the pre-compensation current.

When power is first applied, C3 charges to typically 5.7V before the TOPSwitch starts. It provides bias current until the output voltage becomes regulated. When this occurs, series connected zener diodes VR1,2 begin to conduct. They drive current into the control pin, and directly control the duty cycle.

Capacitor C2 together with R2 performs low-pass filtering on the feedback signal to prevent output line frequency ripple voltage from varying the duty cycle.

Power Integrations Inc, 411 Clyde Avenue, Mountain View, California, 94043. Applications Hotline, (800) 552-3155, fax, 468-0809

Fig. 1. Minimum parts count 5V, 5W bias supply using the PWR-TOP200.

Fig. 2. 70W, 230V ac input boost input power factor correction circuit using the PWR-TOP202.
20mA vco operates to over 1GHz

Described in Motorola's *ECLinPS LITE* technical data book is a low-power voltage controlled oscillator designated the MC12148. This LC-tank-based device operates at up to 1100MHz. Housed in 8-pin SOIC packaging, it needs just 20mA from a 5V supply, and it features a phase noise of typically -90dBc/Hz at 25kHz.

The MC12148 requires an external parallel tank circuit comprising an inductor and capacitor. A varactor diode may be incorporated into the tank circuit to provide a voltage-variable input for the oscillator. Alternatively, the 12148 is suitable for many fixed-frequency applications, but it will not operate in conjunction with a quartz crystal.

Based on the vco circuit topology of the MC1648, the device uses advanced bipolar process technology which results in a design capable of operating at a much higher frequency, but needing only half the current.

Typical frequency stability of the 12148 is 3.6kHz/mV with supply fluctuations, but only 0.1kHz/°C with temperature change. Second harmonic from the carrier is -25dBc, while signal-to-noise ratio is 40dB. Emitter-coupled logic output circuitry of the 12148 is not a traditional open emitter output structure. Instead it has on-chip termination with a nominal value of 500Ω. This facilitates direct ac-coupling of the output signal into a transmission line.

Because of this output configuration, an external pull-down resistor is not required to provide the output with a dc current path.

Output is intended to drive one ECL load. If you need to fan the signal out, an ECL buffer such as the MC10EL16 line receiver/driver is useful.

Motorola Semiconductor Products
Sector, Buckingham Street, Aylesbury, AY1 1XX, tel 01296 395252, fax 01296 21999.

Tank component suppliers
Below are suppliers who manufacture tuning varactors and inductors which can be used to build an external tank circuit.

Coilcraft Inductors AO1T thru AO5T
Coilcraft-Coilcraft, Inc
1102 Silver Lake Rd
Gary, Illinois 60013
tel, 708-639-6400

Loral Tuning Varactors GC1500 Series
Loral
16 Maple Road
Chelmsford, Massachusetts 01824
tel, 508-256-8101 or 508-256-4113

Alpha Tuning Diodes DVH6730
Alpha Semiconductor Devices Division
20 Sylvan Road
Woburn, MA 01801
tel, 617-935-5150

Typical test circuit for the gigahertz low-power vco.
Operating from a nominal supply of 5V, the device is also useful in applications requiring a fixed frequency.

Internals of the MC1248 vco. The device exhibits a phase noise of -90dBc at 25kHz.

Typical evaluation results with the low power oscillator, 930MHz cw.

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January 1996 ELECTRONICS WORLD
Distortion

Building on his previous articles discussing ac analysis with Spice, Owen Bishop details how distortion can be assessed on a pc.

More advanced versions of Spice include the command .DISTO for analysing distortion. Here, I use the routine included in IsSpice. This does not adhere exactly to Spice syntax, but it does cover much the same aspects of circuit behaviour.

Before looking at the routine in operation, it is instructive to examine the test circuit using the standard ac frequency sweep. The circuit is a wide-band amplifier based on a single rf transistor, Fig. 1. Its frequency range is extended by including an inductor in the collector circuit. Assuming that other component values are suitable, it is necessary to settle on a suitable value for the inductor which will give an extended but reasonably flat-topped frequency response.

Draw the schematic and enter component values, including the voltage source, V2 1 0 AC 1. Use the Edit control box to add the .AC command statements to the netlist. The lines required here are, .AC DEC 10 1K 100MEG, and .PRINT AC V(2). These call for the voltage at node 2 to be displayed for a frequency sweep from 1kHz to 100MHz, with ten steps per decade.

Overall response
Select Simulate from the Actions menu. When the analysis is complete, the Simulation Status screen displays a small graph of V(2) in decibels against frequency. These plots are intended only to give an overall picture of the response, not for reading off values. In fact, there are no values displayed on the x-axis and only the lowest and highest level marked on the y-axis.

Often the plot does not completely fill the frame in the vertical direction. As a result, its shape is somewhat obscured and the curve is flatter than it would be with a better choice of scale on the y-axis.

This can be remedied. Click twice on the plot; when the Rescale Plots dialogue box appears, click on the Auto button, then on the OK button. The plot is re-displayed with a more sensible scale. Of course, the associated Intuscope program provides much clearer plots with fully graduated axes, but the small plot is good enough for the moment and allows the overall effects of sweeping component values to be observed.

Sweeping inductances
Sweeping the inductor value is most easily done by using the Simulation Control dialogue box. I normally keep this minimised as an icon, because it obscures the plots when there are more than two on the screen. Clicking on the icon restores the box, Fig. 2.

Now click on the Persistence panel and enter '10'. This allows up to ten curves to be displayed before the plot is cleared and renewed. Click on the Stimulus button to bring up the Stimulus Picker dialogue box. Click on II to list the sweepable parameters associated with the inductor. Double-click on Inductance.

At this point, a small control panel headed 'II:inductance' appears and displays the present value of the inductance, 10µH. Frequency response that is already displayed shows a sharp peak which indicates that this inductance is too large. Clicking on one of the buttons in the row below increases or decreases inductance, the buttons further to the left or right effecting the biggest change.

Check the Always option, then place the mouse on the next-to-leftmost button and hold down the mouse button. Inductance is decreased in steps of 1µH and at each step the simulation is re-run and its output curve displayed. In Fig. 2, the top curve with the prominent peak is for 10µH and the bottom one for 1µH.

The shape of the lower curve shows that 2pF is too small a value, insufficiently extending the amplifier’s bandwidth.

The best response, with no appreciable
Defining distortion

When a pure sinewave passes through a circuit, its shape is almost always altered. The distorted non-sinusoidal shape of the output signal can be described as the sum of the original fundamental sine wave, frequency \( f \), plus sine waves of frequencies \( 2f, 3f, 4f, \ldots, nf \) the harmonics. Amplitudes of the harmonics usually differ from each other and from the fundamental. Some harmonics may have zero amplitude. The fundamental and harmonics may also differ in phase.

Usually, the fundamental has the greatest amplitude. Amplitudes of the harmonics decrease with increasing order. For this reason the biggest contributor to harmonic distortion is the second-order harmonic, followed closely by the third-order. In a distortion analysis, Spice calculates the amplitudes of both second and third-order harmonics.

This can be done for a single value of the fundamental or, more often, input frequency is swept over a given frequency range and distortion is calculated at a number of points within this range. In Spice2 distortion is expressed with reference to the fundamental’s amplitude. Second-order harmonic distortion, HD2, at any given frequency of the fundamental is calculated as the amplitude of the second-order harmonic, divided by the fundamental’s amplitude. Third-order distortion, HD3, is calculated similarly.

Harmonics of the fourth and higher orders can usually be ignored and are not calculated. Note that values of the distortion terms are ratios, so they have no units.

Another type of distortion is produced if the input signal consists of at least two simultaneous sine waves. If two signals have frequencies \( f_1 \) and \( f_2 \), the output signal contains ‘sum and difference’ signals, with frequencies \( (f_1+f_2) \) and \( (f_1-f_2) \). This is known as intermodulation distortion.

Amplitudes of these signals relative to the fundamental are calculated by Spice as SIM2 and DIM2. Not only is there intermodulation between the fundamentals but it also occurs between the fundamental and the harmonics.

An enormous number of intermodulation signals is obtained, even with relatively few harmonics, but usually most of these are insignificant. Spice calculates only one of these signals, \( (2f_1-f_2) \). The command line refers to this as DIM3. The two harmonic distortion values and the three intermodulation values constitute complete distortion analysis.

The range of a Spice distortion analysis is specified in the accompanying AC command, for example,

```plaintext
.AC DEC 10 100 10MEG
```

This specifies 100Hz to 10MHz, with ten steps per decade. Distortion is calculated for a load resistor, described by,

```plaintext
.DISTO ROUT 5 0.9 1M 0.25
```

ROUT is the netlist name of the output resistor. Distortion is to be calculated for every fifth step, that is twice per decade. Analysis is for two frequencies, with \( f_2/f_1=0.9 \) – the ratio must always be between 0 and 1. Power developed by the fundamental in the output resistor is set at a reference level of 1mW – this is the default, so this parameter can be omitted. Amplitude of \( f_2 \) is 0.25 times that of \( f_1 \). Output is specified by a line of the form,

```plaintext
.PRINT DISTO HD2 HD3 SIM2 DIM2 DIM3
```

This calls for values of all five distortion terms, based by default on the magnitudes of the amplitudes. Output can also be calculated in terms of real (R) or imaginary (I) components, or for phase (P), by keying these letters after the appropriate terms. Values on a decibel scale are obtained by including DB, for example,

```plaintext
.PRINT DISTO HD2 (R) HD2 (I) HD3 (DB)
```

Fig. 3. A plot of the distortion analysis shows how second and third order distortion vary with frequency.

Fig. 4. Transient output of the amplifier (L1=4µH) when the input is a 1kHz sine wave with an amplitude of 0.1V.

Fig. 5. Increasing input amplitude to 0.5V shows the beginnings of clipping and general asymmetry of the curve. Although the curve looks like a sine wave, thd is nearly 10%.

peak and with maximum bandwidth, occurs when inductance is 4µH. Working in this way makes selection of component values easy, and is one of the more novel and useful facilities of this simulator.

Circuit analysis

Having found a value which provides the widest bandwidth and a level response, analyse the circuit for distortion. IsSpice syntax differs slightly from Spice2 – see panel 1. Command lines are,

```plaintext
.DISTO DEC 10 1K 100MEG
.PRINT DISTO V(2)
```

The .DISTO line has the same format and syntax as the usual AC line. The .PRINT line specifies the output node. There is no need to state which distortion term is required, as amplitudes of both second and third-order harmonics are calculated automatically. Also make additions to the voltage source statement,

```plaintext
V2 I 0 AC 1 1 DISTOF1 DISTOF2 0.1
```

The DISTOF additions specify magnitudes of the signals \( f_1 \) and \( f_2 \) during the frequency sweep. Only DISTOF1 is used for calculating harmonic distortion. DISTOF2 is used if intermodulation distortion is being calculated.

By default, as with DISTOF1 above, magnitude of the \( f_1 \) signal is 1 and 0.1 for the \( f_2 \) signal. A second value after each of the two keywords can be used to specify their phases, the default of which is 0°. DISTOF may be used with more than one independent voltage source in the same circuit.

Note that IsSpice calculates amplitudes (in volts), not the ratio between the harmonic and fundamental amplitudes, as in Spice2. HD2 and HD3, if required, are obtained by using a calculator.

Figure 3 shows initial results of the analysis with a 4µH inductance. To begin with, the screen shows only second-order distortion. This is indicated by the legend DISTO1 in the Plots panel of the Simulation Control box. To
plot both terms on the same grid, click on the Start button. This plots both second and third-order amplitudes as shown in the figure. The curve with the pronounced peak is the third-order curve.

Distortion curves can not be plotted in Intuscope. For precise details, look in the Output file – Exit the simulation, select Edit Text Files, then click on the OUT button. This displays the output file which lists amplitudes of both of the frequencies as they are swept over the prescribed range. Both curves have a similar shape.

Over a wide range at the lower frequencies, amplitude is almost constant, increasing by no more than 1% of its 1kHz value. It then rises fairly sharply to a peak at the high-frequency end, falling steeply beyond the peak. Table 1 summarises these features for this and subsequent analyses. Comparison with the ordinary ac analysis-results shown in Table 2 shows that at 1kHz, second-order distortion is 0.42/8.05, i.e. 0.05 of the fundamental, ie HD2 is 2.7%. It rises to about 0.15 at a frequency where the curves peak, between 12 and 15MHz.

Intermodulation values are obtained by running the same simulation after adding the f2/f1 value to the .DISTO line. These too are summarised in Table 1, which shows that they are relatively unimportant. For example, the sum of the first two terms is 0.01 of the fundamental at 1kHz. f2/2f1 is only 0.01 of 1kHz. Repeating analyses with L1=10pH presents a different picture. Behaviour of the amplifier is unaffected at low frequencies, starting at 1kHz, but the frequency at which distortion begins to exceed 1% is much lower than when L1 is 4pH. The distortion peaks to higher values and the intermodulation distortions also peak at lower frequencies. These results confirm that 4pH is a preferable value for L1.

Further investigation of distortion in more detail can be undertaken using a full print-out of the output files and a calculator, but this short discussion illustrates some of the ways in which distortion may be examined.

THD analysis – a different angle

Transient analyses include a routine that is also of value in examining distortion. If a pure sine wave of given frequency is fed into a circuit, the composition of the set of harmonics emerging at the output indicates the degree of distortion.

Ideally, only the fundamental should appear at the output, but circuits are nearly always far from ideal. The Fourier analysis, FOUR, examines the output signal and computes amplitudes of the fundamental and its harmonics.

It sounds as if FOUR does the same things as DISTO. There are some points of similarity, but the way in which they are carried out is totally different. A DISTO analysis is preceded by an AC analysis, which applies small-signal inputs to the circuit when at its dc operating point.

Non-linear components are linearised during the analysis. A FOUR analysis, by contrast, simulates the action of the circuit in the time domain. All voltage and current sources are fully operational, and with intrinsically non-linear devices behaving in their non-linear ways. Results are saved in a file that lists one or more output values at each instant in time.

Fourier and transient analysis

Next follows the .FOUR analysis. This takes the results of the Transient analysis and mathematically analyses them to characterise the set of sine waves that constitutes the waveform. Since Fourier analysis depends on data-processing, it is a more efficient and powerful tool for the fundamental and the first nine harmonics. It is 2.7% harmonic. It is a useful single-value measure of the overall distortion.

For comparison with the .DISTO analysis, take the netlist for Fig 1 and edit it for transient analysis followed by .FOUR. Delete lines containing AC and .DISTO, including the statement defining V2. Then add these lines:

\[ .TRAN 10U 10M \]
\[ .FOUR 1V 2(2) \]
\[ V2 = 0.1 \text{ SIN}(0.1 \text{ U1}) \]
\[ .PRINT TRAN V2(2) \]

Since Fourier analysis depends on data produced by a transient analysis, it is essential to include the .TRAN line, even if the precise shape of the transient curve is of no particular interest. The line quoted above specifies time steps of 10µs lasting for a total of 10ms – i.e. “milli” in Spice. The .FOUR line sets the fundamental frequency to 1kHz, and the voltage to be analysed is the voltage at node 2.

The voltage generator is set up between nodes 1 and 0, excited by a sine wave, delay time zero, amplitude 0.1V, and frequency 1kHz. Total time of the analysis allows for ten complete cycles of the waveform, although only the final cycle is analysed by .FOUR. This gives time for the circuit to get into a steady operating state.

In a transient analysis capacitors begin uncharged, unless initial values are set. Similarly, inductors begin with zero current through them and with zero induced emf. In a simple circuit such as this, values reach steady levels after fewer than ten cycles, but it is a safe habit always to run for at least ten cycles.

Select Simulate from the Actions menu; the Simulation Status screen displays a plot of ten sinusoidal cycles, or rather, ten inverted sinusoidal cycles, since this is an inverting amplifier. Select Scope, then select VOUT=V(2) on the Waveform menu.

Transient analysis is displayed as in Fig. 4, and is a typical inverted sine wave with no visible distortion. Exit Intuscope, then Exit Simulation Status, select Edit Text Files and click on the Out button. The output file has the usual netlist, followed by a table of the Initial Transient Solution (dc operating point values), Fourier analysis, and data generated by the Transient analysis. Running the analysis for ten more-or-less repetitive cycles creates a lot of data – 20 pages of it, if printed out.

Fourier analysis table Table 3 begins with a statement of the thd. Harmonic zero is the dc component of the waveform. This is calculated from:

\[ \text{THD} = \frac{\sqrt{a_1^2 + a_2^2 + \ldots + a_n^2}}{a_1} \times 100 \]

where \( a_1 \) is the amplitude of the fundamental, and \( a_n \) is the amplitude of the \( n^{th} \) harmonic. It is a useful single-value measure of the overall distortion.

| Table 1. Key values of distortion in the amplifier circuit of Fig 1. |
|-----------------|-----------------|-----------------|-----------------|-----------------|-----------------|
| Inductance of L1 | Distortion term | Initial amplitude (V) | up to (MHz) | Peak amplitude (V) | at (MHz) |
| 4pH             | 2nd-order       | 0.42             | 0.63         | 1.26             | 15.8            |
|                 | 3rd-order       | 0.35             | 0.32         | 1.26             | 7.9             |
|                 | f1+f2           | 0.08             | 1.26         | 0.18             | 25.1            |
|                 | f1-f2           | 0.08             | 1.26         | 0.18             | 25.1            |
|                 | 2x f1-f2        | 0.11             | 0.5          | 0.31             | 12.6            |
| 10pH            | 2nd-order       | 0.42             | 0.40         | 4.03             | 15.8            |
|                 | 3rd-order       | 0.35             | 0.32         | 2.24             | 7.9             |
|                 | f1+f2           | 0.08             | 0.79         | 0.47             | 15.9            |
|                 | f1-f2           | 0.08             | 0.79         | 0.47             | 15.9            |
|                 | 2x f1-f2        | 0.11             | 0.39         | 0.66             | 10.0            |

| Table 2. Amplitude of the fundamental at the amplifier output (Node 2). |
|-----------------|-----------------|-----------------|-----------------|-----------------|
| Inductance of L1 | Initial amplitude (V) | up to (MHz) | Peak amplitude (V) | at (MHz) |
| 4pH             | 8.05             | 6.3           | 8.25           | 12.6           |
| 10pH            | 8.05             | 1.6           | 13.57          | 15.8           |

Total harmonic distortion

This is calculated from:

\[ \text{THD} = \frac{\sqrt{a_1^2 + a_2^2 + \ldots + a_n^2}}{a_1} \times 100 \]

where \( a_1 \) is the amplitude of the fundamental, and \( a_n \) is the amplitude of the \( n^{th} \) harmonic. It is a useful single-value measure of the overall distortion.
angles for harmonics three to nine lead the input by between 125° and 172° respectively. Absolute values are sometimes of interest, but it makes possible comparisons between tests under different sets of conditions if the values are normalised.

The last two columns of the table show differences only in the last or next-to-last of the six significant figures. Overall there is a slight increase in amplitudes but they are negligible.

Whether inductor \(L_1\) is 4\(\mu\)H or 10\(\mu\)H is immaterial when frequency is low. Fourier analysis operates at only one frequency; it does not sweep the frequencies as does .DISTO. As a consequence, edit these lines of the netlist to:

\[ \text{.TRAN 800P 800N} \]
\[ \text{.DISTO} \]

Using frequency spectra

Many simulators lack .DISTO but Fourier analysis can be used for assessing harmonic distortion. In SpiceAge, enter a netlist to describe the circuit of Fig. 1, with \(L_1\) equal to 4\(\mu\)H, and with \(V_2\) amplitude 100mV, sine excitation at 1kHz. Set probes to measure the input signal (node 2) and the output signal (node 7).

The procedure is to run a transient analysis and use the Fourier option to analyse and display the result. SpiceAge performs a Fast Fourier Transform on the whole wave train not just the last cycle. For most accurate results, sample several cycles and make the sampling period cover an integral number of cycles. In the Sweep Times dialogue box, set start time to zero, stop time to 5ms and step time to 10ps. This gives time for 5 cycles of the waveform. Run a Transient analysis which shows a sine wave, amplitude 0.725V, Fig. 6.

Follow with a Fourier analysis but first use the Probe Control box to de-select the input probe, as we do not need to analyse the pure sinewave input. Also in the Graph Scaling box, set the x-axis maximum to be held at 10k. Frequencies higher than this have such a low amplitude that they may be ignored.

Under Y-display mode, select Lines to Origin – to give a spectrum – and Phase Plot. Only the fundamental (1kHz) and the first two harmonics appear, Fig. 7 – the rest are too small to show on this scale. One way to bring more into view is to select dB scale under Y-display mode. This will also bring into view lines for many other frequencies, some of which may be cross modulations but many of which will be 'mathematical noise'. To cut these out, set the y-axis minimum to be held at -85dB. Frequencies at amplitudes less than this may certainly be ignored.

It is clear from these results, Fig. 8, that the distortion products of this circuit are all so small that they can be ignored.

Clipping

One important application of .FOUR is to investigate clipping and similar forms of distortion. This is quite outside the scope of .DISTO.

Using the same circuit as above, re-edit to the lowest distortion settings – with a frequency of 1kHz and \(L_1\) at 4\(\mu\)H but increase amplitude of \(V_2\) to 0.5V. The simulation shows a waveform that is clipped but only just, Fig. 5. Fourier reveals a generally unacceptable THD of 9.67%. Second and third harmonics contribute most to this (normalised amplitudes 0.0874 and 0.0375) but the fourth and subsequent harmonics are very small – there is a change of pattern. Alter \(L_1\) to 10\(\mu\)H and repeat. THD increases to 2.8247% and harmonics increase roughly in proportion.

The analysis after editing the netlist to make \(L_1\)=10\(\mu\)H. Figures obtained show differences only in the last or next-to-last of the six significant figures. Overall there is a slight increase in amplitudes but they are negligible.

Whether inductor \(L_1\) is 4\(\mu\)H or 10\(\mu\)H is immaterial when frequency is low. Fourier analysis operates at only one frequency; it does not sweep the frequencies as does .DISTO. As a consequence, edit these lines of the netlist to:

\[ \text{.TRAN 800P 800N} \]
\[ \text{.DISTO} \]

Repeat the analysis after editing the netlist to make \(L_1\)=10\(\mu\)H. Figures obtained show differences only in the last or next-to-last of the six significant figures. Overall there is a slight increase in amplitudes but they are negligible.

Whether inductor \(L_1\) is 4\(\mu\)H or 10\(\mu\)H is immaterial when frequency is low. Fourier analysis operates at only one frequency; it does not sweep the frequencies as does .DISTO. As a consequence, edit these lines of the netlist to:

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