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ELECTRONICS WORLD

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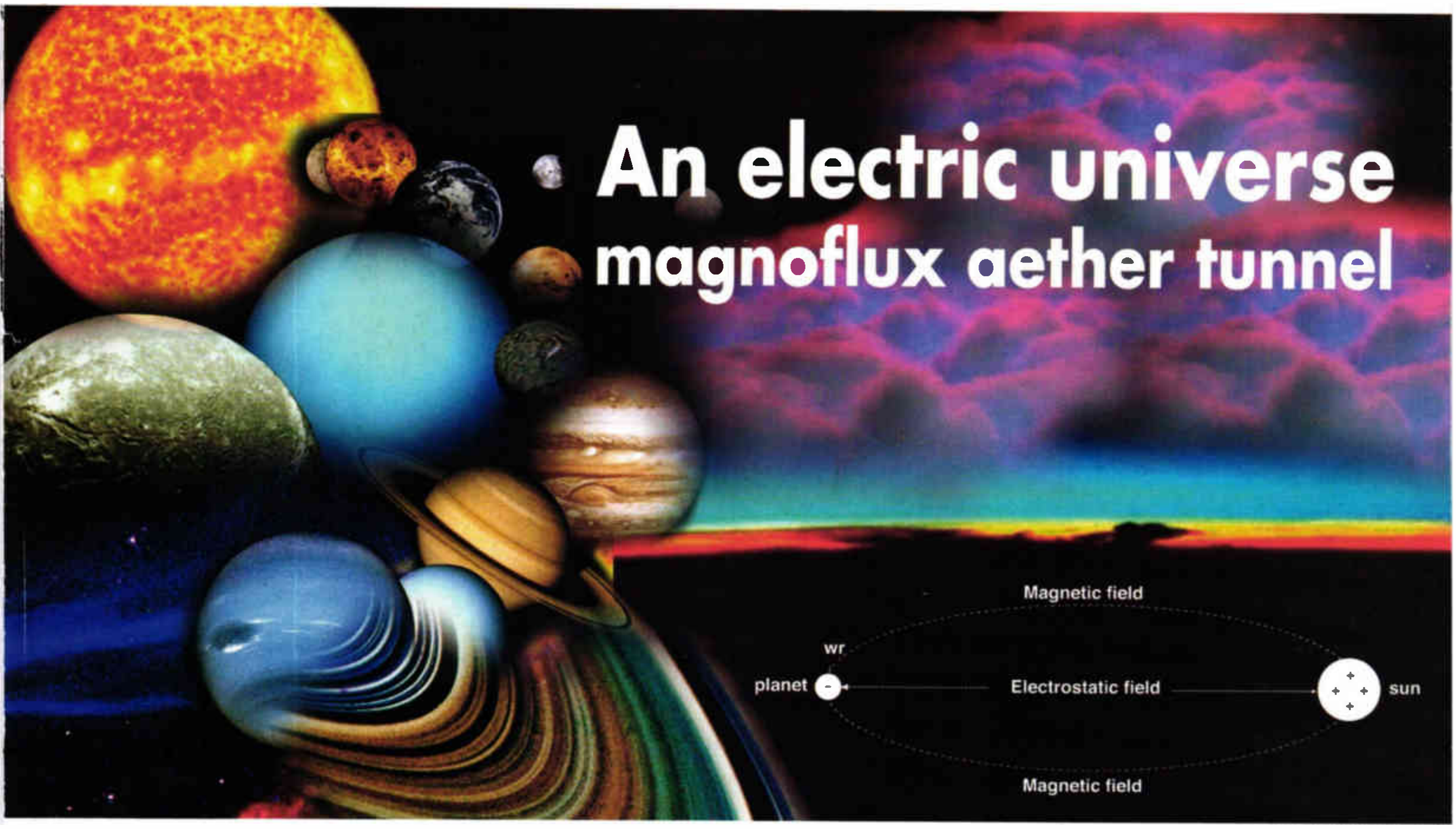

Class-A imagineering: PART III

SDI vision mixer

Automatic heterodyne filter

Single-frequency element oscillators

Circuit Ideas

- Frequency divider
- Accurate one second clock
- Low range AM transmitter
- Ceramic cartridge preamp
- *and much more ...*

CONTENTS

AUGUST 2004 VOLUME 110 NUMBER 1820

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Speak up!

As I said last month, we have finally got around to doing another reader survey. A lucky winner (pulled from the hat) will receive a bottle of Champagne, but if you'd rather get a free subscription or one of our archive CDs, do let me know. This survey is very important to us, as it will steer the direction of *Electronics World* for the foreseeable future. So please participate if you want your voice to be heard.

In this issue we have yet another bumper crop of circuit ideas in an effort to reduce the time taken to publish ideas. And I can report that for new ideas, the wait time is down to just under a year – but it was 18 months not long ago! My aim is to get this down to three months, so as long as no one objects, and no one has so far, I'll continue to do bumper *Circuit Ideas* sections until my goal has been reached.

Also this month we have a slightly off topic contribution from Clive Stevens on the subject of magnoflux aether tunnels, which I'm sure will stimulate debate in the *Letters* section. And talking about letters, Mr. Catt has decided to put his money where his mouth is. To find out how – *Letters* begin on page 54.

Elsewhere, we've got Graham Maynard's excellent continuing saga of things audio, an automatic heterodyne filter for the RF aficionados, a different slant on oscillators and another thought provoking piece from Leslie Green.

I read on the internet this week that in April this year Tim Berners-Lee (the inventor of the World Wide Web), had won the Finnish Technology Award Foundation's first Millennium Technology Prize. Highly commendable all round. But how come the first award was in 2004 not 2000? Perhaps they ought to rename the award.

I'm sure a lot of you are getting bored with my monthly ramblings so, if anybody else has some comments or observations on our industry and would like to air them – feel free to send in a leader column. This page normally runs around 4-500 words – so any volunteers gratefully received. And unlike the *Letters* section – you'll actually get paid!

Phil Reed

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UPDATE

Lead-free legislation in free-fall

Europe's forthcoming RoHS Directive is causing chaos in the industry as firms, trade groups and governments argue over definitions within the Directive.

Much of the industry confusion stems from the term 'homogeneous materials' which is used to define the basic component by which concentrations of lead, chromium and other dangerous materials are measured.

Three possible alternatives exist for a definition of homogeneous material: Whole assemblies like modules, PCBs or even a whole TV or computer; individual components such as ICs; or the raw materials themselves.

The latter definition is the only one that would see lead eliminated completely from components. This is the definition to which chip firms such as National Semiconductor, Intel and Zetex are working.

Using the component as a basic definition would allow lead to be used on the terminations of components. This definition is supported by trade bodies including EICTA, the pan-European group.

However, such a classification could result in major problems for manufacturers. Components marked as 'RoHS Compliant' might, or might not, contain lead and the manufacturer would have no immediate way of knowing which components contain lead, and which do not.

Components that use lead on their terminations could reduce reliability if used with lead-free solders and could ruin solder baths.

Manufacturers would have little option but to test every batch of components to see whether they contain lead.

The problems extends to other materials banned under RoHS. Hexavalent chromium, for example, is carcinogenic – far more so than lead - and is often used to plate mechanical components. The thin layers used means that the compound could be used to plate a screw and still meet the 0.01 per cent limit.

At a meeting of the EC's technical adaptation committee (TAC) in May, the final definition should have been made, but EC procuedures and voting rules meant it had to wait.

Following this TAC meeting, a clear definition of the term was called for by the DTI in a report it commissioned from ERA Technology.

This report was presented at the most recent TAC meeting. It said "There is a requirement for a clear understanding of how the definition of the maximum concentration values will be interpreted".

However, at this latest TAC meeting the issue was still not cleared up, despite a vote on the matter that should have seen the relevant paragraphs inserted into the Directive.

A source told *EW* that the number of votes needed to make a decision was not reached. Uncertainty over the new voting rules meant not enough members were available at the meeting.

All European member states must implement their version of the RoHS Directive by August 13, but the failure of the TAC to come to a conclusion about the term 'homogeneous materials' means that Governments will have to provide their own definition, which might not be common to all states.

LEDs close to 100lm/W

LED firm Osram Opto Semiconductors is claiming an efficiency of 96lumen/Watt for an LED.

The 300x300μm device emits more than 12lumens at 70mA, claimed the firm, although the highest efficiency is achieved at a more modest 20mA of drive current.

"The efficiency of 96lm/W is the highest reported efficiency at this wavelength (618nm)," said a spokeswoman at the German firm. "However, even more important might be the fact that it is achieved with a purely surface emitting LED."

Surface emitting LEDs are more useful, and are easier to couple light in and out of fibres.

"Most other high-brightness LEDs use all sides of the chip for light extraction. Even if these chips achieve high efficiency values, the luminance is much lower," said the spokeswoman.

"For operation at higher currents, we just increase the emitting area," she added. The firm has thin-film LEDs available for nominal operating currents of 50mA (chip-size 300μm), 200mA (500μm) and 400mA (1000μm).

The devices are made using AlGaInP materials.

High powered LED

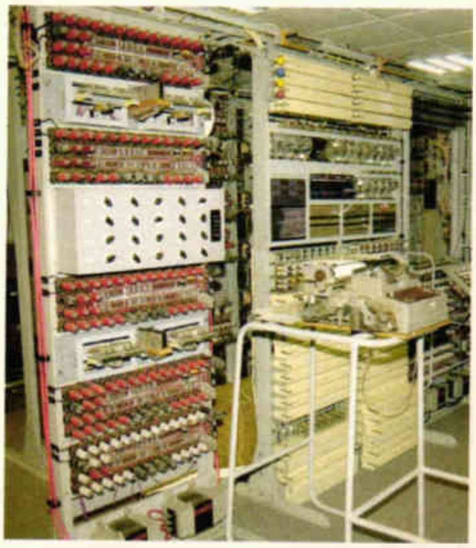
LED chip maker Cree recently demonstrated 57lm/W of white light from an LED die operating at 350mA, and 142lm from the same device operating at 1A.

"The company believes that these are the highest reported results for a high power packaged LED chip to date," said a spokeswoman.

North Carolaina-based Cree uses a combination of silicon carbide and gallium nitride materials to make its LEDs. It produces blue, green and near-UV diodes as well as near-UV lasers.

www.cree.com

Colossus 60 years old comes out of retirement



The Science Museum in London hosted an important anniversary on June 1st. It was 60 years ago to the day that the first Colossus Mark II was commissioned at Bletchley Park. Colossus was the world's first electronic digital computer and it played an important role in enabling the deciphering of high level German tactical messages during World War II.

Over one hundred and twenty people were present including Harry Fensom, an original Colossus engineer, now 83 years old.Wrens played an important role in the operation of Colossus and 20 attended the event offering useful stories about operating the machine.

The event also marked the near-completion of a rebuild of Colossus MkII. Tony Sale, who is running the project gave a presentation showing for the first time, on the rebuild, how the machine was actually used to break a cipher in 1944.

Getting to grips with 90nm

Each shift to a smaller manufacturing process brings a raft of new problems for chip makers. The move from 0.13μm to 90nanometre has proven exceptionally tricky.

Now Intel is claiming to have gained ground on one of the major problems - that of current leakage.

As the process and hence transistor channel length shrinks, the oxide layer separating the gate from the drain-source channel must also shrink. However, with gate oxide thickness approaching 1nm, or 10Angstrom, leakage through the oxide becomes a major problem.

Using exotic materials such as hafnium oxides is a solution, but not one that can be integrated into today's processes.

Intel in its 90nm Dothan processor - sold as a mobile Pentium - has managed to double transistor count, while raising clock speed and lowering power compared to its 0.13μm chip (codenamed Banias).

Dothan has 140 million transistors, Banias has 77 million, but both have a die size of 83mm^2

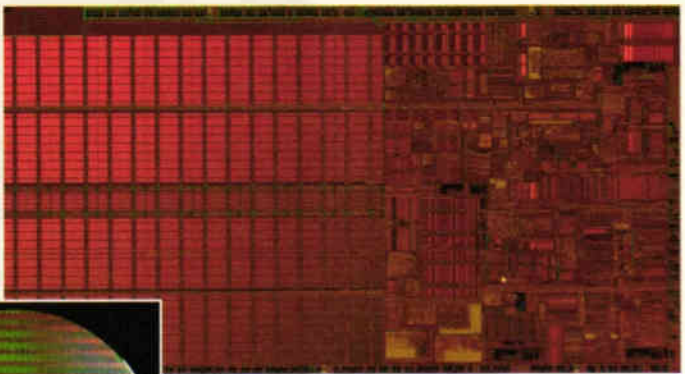
Intel has used multiple threshold transistors in the core. Slow transistors not on the critical path can have thicker gate oxides to reduce current leakage.

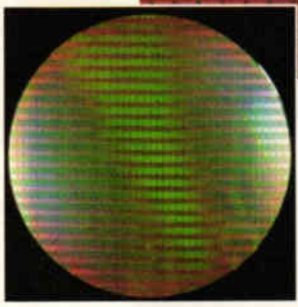
To improve speed the substrate uses strained silicon. This stretches the silicon lattice by growing a thin layer of germanium. The larger lattice spacing leads to increase carrier mobility.

Various sections of the design industry are also working to help with the problems faced at 90nm, not least the EDA tool suppliers and process library developers.

EDA firms are releasing tools that can automatically design with 'voltage islands', where different areas of the chip have different supply voltages. They can also cope with several types of transistors with differing threshold voltages.

Library vendors are also offering cells that cope with voltage islands and multiple thresholds.





Above: Intel's 90nm Dothan processor

Left: A 300mm wafer of 90nm Dothan processors

Lasers make microstructures on a hair

Researchers at Boston College in Massachusetts adapted a micro-fabrication technique to mould polymer features onto hair.

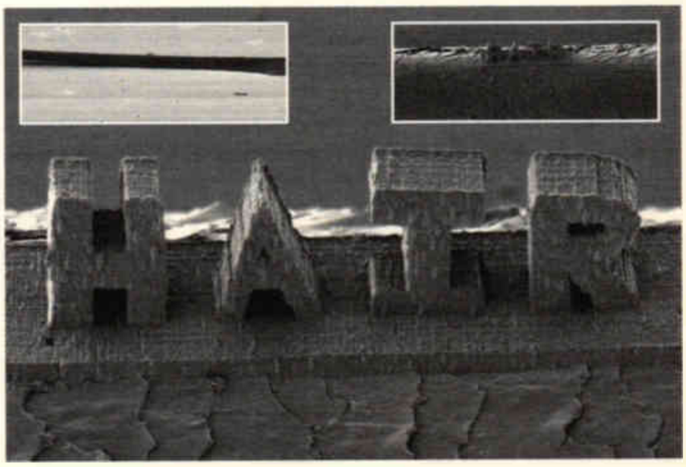
Called multi-photon-absorption photo-polymerisation (MAP), the technique involves depositing polymer at the focal point of a laser. Scanning and modulating the beam "allows for the formation of intricate, three-dimensional patterns," said the college, which claims that it "makes it possible to create features that are 1,000 times smaller than the diameter of a human hair."

The discovery of bio-compatibility was serendipity, following on from the original purpose of the study which was to demonstrate intricate and resilient structures could be created with MAP using cheap, common materials.

In order to demonstrate the size of the features that could be created, the researchers fabricated structures near a human hair and in the course of these experiments discovered it was also possible to fabricate structures on the hair itself.

"We built the structure on top of the hair with a material that is akin to Plexiglas [Perspex]," said Professor John Fourkas. "One of the really exciting and unexpected things about this is that we found that we could make this structure on the hair without harming it in any way."

This has prompted the researchers to look for other bio-applications. "It suggests that we could accomplish the same with other biological materials. One could imagine, for instance, building devices directly on skin, blood vessels, and eventually even a living cell," said Fourkas, who cautioned: "While this idea is currently in the realm of science fiction, our results represent an important step in that direction."



Back at the original reason for the project, Fourkas sees MAP as a way to make optical components for opto-electronics. "While there are applications of the technique we used in the optical communications area that are being pursued by us and by others, writing a structure on a hair does not have direct bearing on optical communications," he said. "On the other hand, we can and have done exactly the same sort of thing on optical fibres that are of comparable size, and this does have direct bearing."

Students dream up novel sensor delivery

University of Florida students have turned to a paint ball gun to position wireless sensors in hazardous and difficult-to-reach places.

The project is part of a military-sponsored programme to find ways of safely sniffing out hidden explosives, led by Florida-based Lockheed Martin Missiles and Fire Control.

Numerous delivery mechanisms, including a gun made of plastic tubing, were tried before settling on an off-the-shelf paintball gun.

The projectile, currently a prototype, includes a transmitter, a sensor and a wire antenna, all powered by a watch battery.

As miniature chemical and explosive detectors are not yet available, the team opted to install an accelerometer which allowed force measurements to be made during launch and landing.

"I think the most important thing for the proof of concept was to see if the electronics could survive the impact," said electrical engineering student Felipe Sutantri.

Getting the projectile to stick proved problematic. Eventually an "industrial polymer" was discovered that provided instant grab. The weight of the polymer, together with the arrangement of the components, causes the projectile to be heaviest at the front, which helps it fly straight and strike the target with its sticky end. "What we did was we made its tip heavy so it's like a dart – it doesn't tumble over," said Sutantri.

Tests show the transmitter could report data from over 70m, while the paintball gun can shoot the projectile 20m. "Both distances could be extended in the production version," said Florida.

Top: A sticky-tipped wireless sensor, the green lead is an aerial.

Inset: engineer Syad Sohab with the sensor and the paintball gun which delivers it.

Games console is a PC in miniature

The Eve console

OS	Windows XP Embedded
Processor	533MHz Eden-N
RAM	128MB DDR266
Hard disc	20Gbyte 1.8in
Display	10cm 640x480 LCD
Graphics	S3 UniChrome Pro (VIA)
Battery	Liion and external power

San Diego-based Ministry of Mobile Affairs (MoMA) has revealed the tiny Eve mobile gaming console, based around an x86 processor and running Microsoft's XP Embedded operating system.

"With its traditionally high levels of power consumption and heat dissipation, the x86 architecture has never been able to break into the mobile space," said MoMA CEO Andrew Huang.

To get around the power problem, Huang adopted VIA technologies' 'Grace platform' which includes a low-consumption processor and an MPEG2/4 accelerating chipset for 2D and 3D graphics.

"The single largest challenge to enter the mobile gaming console market has always been signing up developers to deliver games for unique [non-x86] hardware architectures," said VIA spokesman Richard Brown.

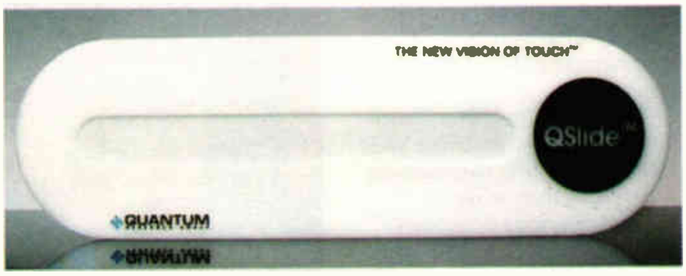
Ministry of Mobile Affairs is "dedicated to providing affordable yet unsubsidised [by games sales] mobile gaming based upon the x86 architecture", said Huang, who is the author of 'Hacking the Xbox: An Introduction to Reverse Engineering'.

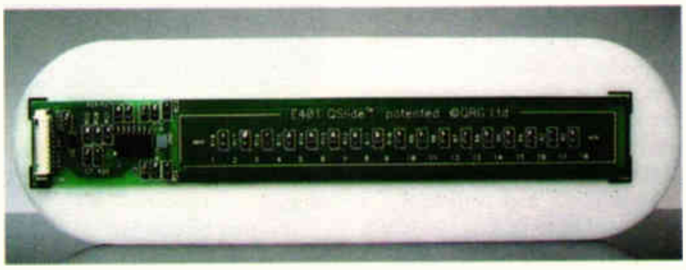
www.ministrymobile.com

Flexible LCDs

Scientists at Philips have created flexible LCDs by coating a plastic foil with a fluid mixture, followed by a single step ultra-violet (UV) light exposure process. The firm said it is an improvement over the company's previous process, which required two separate UV exposure steps to produce the array of liquid crystal cells. "Using a simpler and less time-consuming offset printing stage is a major step towards the realization of reel-to-reel LCD manufacturing," said Philips.

Sliders, but not as you know them





Fabless chip touch sensor chip firm Quantum Research Group of Southampton has developed a replacement for electromechanical slider controls.

QT401 is a 'slider chip' that senses along a line or arc and provides a linear to a resolution of one part in 128.

A resistive element, which can be made from transparent indium-tin oxide, acts as the capacitive sensor and can be located behind up to 3mm of insulator including glass and plastic.

"This chip is for control panels in appliances, or in just about any consumer product where you would expect to see a knob or linear control, for example for brightness, volume, temperature, power and other functions," said Quantum.

The output is a SPI serial interface. Data output can be interpreted as either a linear control, or as a series of buttons or a mixture of the two, depending on the needs of the designer, said the firm.

To create up to eight buttons, the output range is interpreted as a number of sub-ranges.

"It can be used with clear sensing elements to create LED-backlit touch zones for a very dramatic effect, or to show position after the touch is released," said Quantum.

See Quantum's Concept Cooktop at www.qprox.com

Low-cost chip make fibre affordable

A Scottish spin-off firm from Heriot-Watt University has developed a single chip that can drive multiple fibre optic standards.

The device could dramatically lower the cost of fibre links and beat copper is many applications.

"So far it has been prohibitively expensive to add optical interconnects to most devices," said Dr Keith Symington, founder of Conjunct. "It has only been telecoms switches and routers costing thousands of dollars."

Conjunct's device could be applied to a PCB and would offer short range optical links.

"What we're creating is a system package that incorporates the integration of optical transceiver, passives, serialiser/deserialiser and a field programmable gate array," says Symington.

The FPGA allows the device to implement multiple comms standards across the fibre, including USB, FireWire, Gigabit Ethernet, Fibre Channel, SCSI and PCI Express.

"The advantage of a field configurable standard product is that an engineer can just pick it off the shelf and choose to implement the protocol required," Symington said.

Scotland gets opto design centre

Scottish Enterprise has opened a design centre aimed at the commercialisation of optoelectronic research.

Optocap is based in Livingston and hopes to attract university spin-offs, researchers, start-ups and established companies. They will be able to use the centre's facilities and service to develop and test their ideas.

"It's the kind of practical support for innovative products that will give them the competitive edge in the market place," said David Ruxton, chief executive of Optocap. The £4m centre is jointly funded by Scotiish Enterprise and the European Regional Development Fund.

One of the main focuses of the centre will be the packaging of optoelectronic devices.

"Packaging of these very fragile and complex devices is often the main barrier to producing a commercially viable product," said Ruxton.

"With as much as seventy five per cent of the product cost lying in developing a packaging solution it is critical that its design is integrated into the device development process."

The first two firms to make use of Optocap are Edinburgh's Micro Emissive Displays and Conjunct, a spin-off from Heriot-Watt University.

Tiny RF tag reader opens door to new applications

Not much bigger than an RF identification (RFID) tag, Wokingham-based Innovision claims this is the world's smallest and lowest-cost 13.56MHz RFID reader.

"The reader has been designed to enable the next generation of RFID applications. For example, by passing your MP3 player with a built-in reader over an NFC-enabled music poster, you could download a sample track from the advertised album instantly or even purchase the entire CD," said Rob Kitchen, head of the firm's consumer business.

Known as 'io', it is aimed at the fledgling Near Field Communications (NFC) standard and includes a risc microcontroller to allow users to keep up with changes in the standard. "The reader's future-proofing for the emerging NFC standards opens up the possibility of RFID applications in completely new areas," said Kitchen.

io runs off 2.8V and reads and writes to industry-standard 13.56MHz RFID tags and smart labels.

Innovision specialises in RF tag products and is probably best know for developing the technology behind talking Star Wars toys a few years ago.

www.nfc-rfid.com
www.innovision-group.com

RF amplifier gets efficient

A firm from Cambridge claims to have improved the efficiency of power amplifiers used in mobile phone basestations, in some cases doubling or tripling the efficiency to 35 per cent.

Nujira said its technology is suitable for wideband-CDMA and OFDM modulation techniques and could be applied to other air-interface schemes.

The firm is using a tracking technique to increase efficiency, which involves modulating the voltage to the drain of the RF power transistors in sync with the signal envelope.

This technique of envelope tracking is well known, but can create significant distortion in the output while making the power supply design rather difficult. "This in itself can cause some other distasteful effects," said Nujira.

The firm has managed to reduce distortion by ensuring the drain voltage supply very accurately tracks the RF envelope and claims to have produced the required power supply.

Power supply modulation has been achieved with an efficiency of greater than 85 per cent and bandwidths over 60MHz, it said.

"We are demonstrating a 20 Watt multi-carrier WCDMA amplifier using high accuracy envelope tracking," said Tim Haynes, Nujira's CEO. He said the firm is "demonstrating that the poor efficiencies associated with WCDMA and OFDM systems need no longer be the case".

Other techniques to improve performance of RF amplifiers include pre-distortion and feed-forward. However, the firm said these "are complex and require complicated hardware circuits for implementation and rarely achieve efficiencies of over 20 per cent in cellular systems".

Nujira has received almost $1m in funding to develop its products.

If the technology can be made successfully, a mobile operator could save several hundred million pounds over the ten year life of a 10,000 basestation network in electricity charges alone.

Novel technology shows its TV credentials

Claiming it to be the largest of its kind, US firm iFire has unveiled a 34in. colour inorganic electroluminescent (EL) flat panel display.

"Producing a 34-inch prototype display of this quality in our R&D facility proves both the scalability and the simplicity of our electroluminescent technology," said company president Anthony Johnston. "Proving the scalability of our technology is an important milestone as we move from R&D into pilot production."

Based on iFire's proprietary thick-film dielectric electroluminescent (TDEL) technology, the prototype is also the first native high definition display of its kind. The 1280x768 display can handle 720p video and XGA computer data.

"iFire is well on track to produce high-definition, mid-30in. sized flat panel television modules," said Barry Heck, President & CEO of The Westaim Corporation which owns iFire.

iFire claims a 30in. TDEL display will cost 30 to 50 per cent less than a similar LCD or plasma display.

This may be true, but techniques for making these other technologies are well established and iFire will have to work hard to get market acceptance.

"We see a major market opportunity for an affordable high-quality flat panel display technology," said Johnston. "Our analysis indicates that iFire's TDEL will have a sustainable manufacturing cost advantage and will require a smaller capital investment for manufacturing. This presents a very attractive alternative to other flat panel display technologies for potential manufacturing partners."

The firm plans to begin pilot production of 34in. high-definition television display modules in 2005.

Record claim for OLEDs

Universal Display Corporation (UDC) is claiming a record for power efficiency using white OLEDs (organic light emitting diodes).

The firm has used phosphorescent OLED technology to make a device with luminous efficiency of 38candela/Amp and a power efficiency of 18.4lumen/Watt at 1,000cd/m^2.

The device has a white emission with a colour rendering index (CRI) of 79.

UDC is aiming its OLEDs at both displays and at white backlighting applications. In the latter case it is working with Toyota.

"This demonstration is crucial to establishing white OLED technology as a suitable, and more efficient, alternative to traditional displays and LCD backlighting," said Dr Julie Brown , chief technical officer at UDC.

The firm said the use of phosphorescent materials with OLED allows quantum efficiency to approach 100 per cent, four times that of a standard OLED.

Phosphorescent materials are commonly used to convert higher energy photons from blue or violet LEDs into lower energy green and red photons, thereby creating a white light source.

UDC has also developed flexible OLEDs built on thin metallic substrates.

Making flexible OLEDs is tricky, as the organic layer must be kept sealed from the outside environment as moisture and oxygen are poisonous to the materials. UDC used alternating layers of polymer and ceramic film applied in a vacuum to seal the display.

The 150x150mm display was built on a 100μm substrate.

Professor Joe McGeehan CBE

Professor Joe McGeehan, managing director of Toshiba Research Europe's Bristol-based Telecommunications Research Laboratory, has been awarded the CBE for pioneering innovation in mobile communications.

Professor McGeehan has published more than 200 papers in the field of mobile communication systems, RF engineering, microwave theory and techniques, radio-wave propagation, signal processing and plasma physics.

His research into Wideband Code Division Multiple Access (W-CDMA) during the early 1990's formed the basis of today's 3G cellular standard. He also led the development of speech scrambling technology which is used today by all the UK's police forces for secure mobile radio transmissions. This technology won the Prince of Wales Award for Innovation (1992).

Professor McGeehan is a Fellow of the Institute of Electrical Engineers (FIEE) and a Fellow of the Royal Academy of Engineering (FREng).

SDI vision mixer

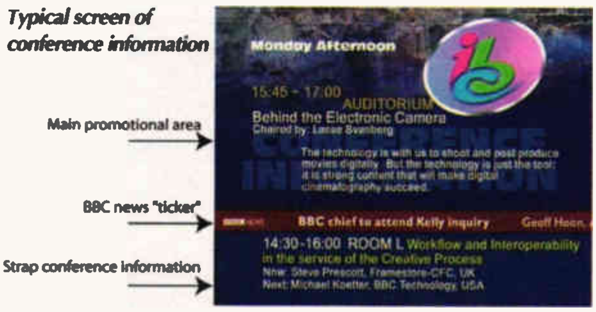
Mike Cox, video design guru and analogue dinosaur designs his first digital vision mixer. This is the story of how he came to love digits and is the account of an 18 month project to build the unit

The IBC show in Amsterdam (one of the world's largest and most prestigious broadcast hardware events) has long had an information channel, which trails and promotes Conference Sessions and other events in the show. This has been displayed on around 70 – 80 screens around the RAI Exhibition and Conference Centre. I have been involved with this for the last 10 shows as technical director. The images are mainly derived from Aston Character Generators (CGs), which Aston kindly lend the IBC, augmented by several computers with digital video recording facilities and associated graphics composing. Some years ago, the RAI-ACS decided that it wanted an SDI (Serial Digital Interface, 270Mb/s) feed to the PAL modulators that feed the RAI cable distribution system. The Info Channel at that time was arranged as an analogue component system. To meet the SDI requirement, we fitted a Vistek YPbPr to SDI converter at the end of the local chain.

All the video switching and mixing was done in YPbPr, using a device called a Title Assembler. This is an 8 input, 2 bank YPbPr device, originally designed for broadcast facility graphics compose areas, with simple dissolve between the banks. Following this was a simple Split Screen unit, which allowed insertion of a 'strap' along the bottom of the screen, carrying information about the current conference session, and who is currently speaking. This is vital information for those who are combining conference attendance and visits to the exhibition.

SDI mixer in action at IBC

Typical screen of conference information



A final fade to black facility completed the video path. These analogue products had been designed over the years, drawing heavily on work I had done with Cox Associates and Vistek. In between IBCs, they live in my garage, along with all the power and video cables for the complete control system: and some other analogue component mixers that are used for the IBC Conference A/V set-up.

After IBC2001, it was felt that the Info Channel should move into the digital age and perform all switching and mixing on SDI signals, particularly as the Aston CGs all had SDI outputs. The target was to build an SDI mixer incorporating all the switching, mixing, split screen and fade facilities that we needed by IBC2002. The mixer has now been used in anger at IBC2002 and IBC2003. The following is an account of the design and development process.

Early work

It might be worth setting out my resources to undertake this task. I used to operate a company that made video mixers, as well as PAL/NTSC/SECAM coders, switchers and a variety of broadcast items. Some readers may have come across the COX 841 series of mixers, or even the COX T8/T16/T24 series. I had a lot to do with the 841s, but very little to do with the T series as they used processors such as the Z80 in the control system, and I am an analogue dinosaur! However, after the company was bought by Carlton in 1985, Cox Associates (CAL) was set up to produce test generators, and also the 2036 Title Assembler mentioned earlier. CAL designed numbers of small spine cards using Surface Mount (SM) components for such things as video crosspoints, 75 ohm driving output amplifiers and blanking regenerators, which were used in the Title Assemblers and various routing switchers.

In parallel with this, I helped Vistek Electronics financially with a management buy-out from GEC-McMichael, and after Vistek bought back my original company from Carlton in 1990, I threw Cox Associates into the pot, and semi-retired. Vistek continued to make some CAL designs, and I continued my responsibility for these, mostly working from home.

To do this, I have a reasonably well-equipped lab at home, although it only measures 8 feet by 5 feet. It contains an SPG, a component (YPbPr) test generator, a caption camera (CCD) looking at a card, oscilloscope, spectrum analyser, EPROM programmer etc. In addition there are several picture monitors, some proprietary analogue to SDI converters and a PC and printer. ORCAD and Boardmaker software (DOS versions) cover the CAD side of things.

At a leisurely pace over the previous years, I had done some simple digital video experiments, building up on a large peg board a 27MHz clock generator, video ADC and DAC, simple digital $\sin^2$ pulse and bar/ramp generator, frame store, 3 x1 switcher, and a number generator driven by a shaft encoder. This led to some thoughts on interconnection.

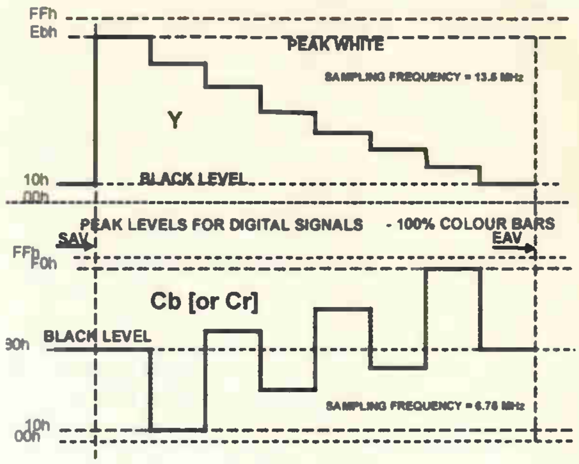
All the parts that are easily available are 8 bit – EPROMs, 8 wide latches and buffers such as 74HC574/541. Hence the decision was taken that all work would be 8 bits wide, although the CCIR and SMPTE specifications do allow for 10-bit working. The standard 8-bit interconnection uses a 10-pin IDC connector on 0.05" ribbon cable. By extension, the '601' interconnect uses a 14- pin IDC connector and ribbon cable. The extra pins cater for the clock signal, and some extra ground. This has proved remarkably simple and economic.

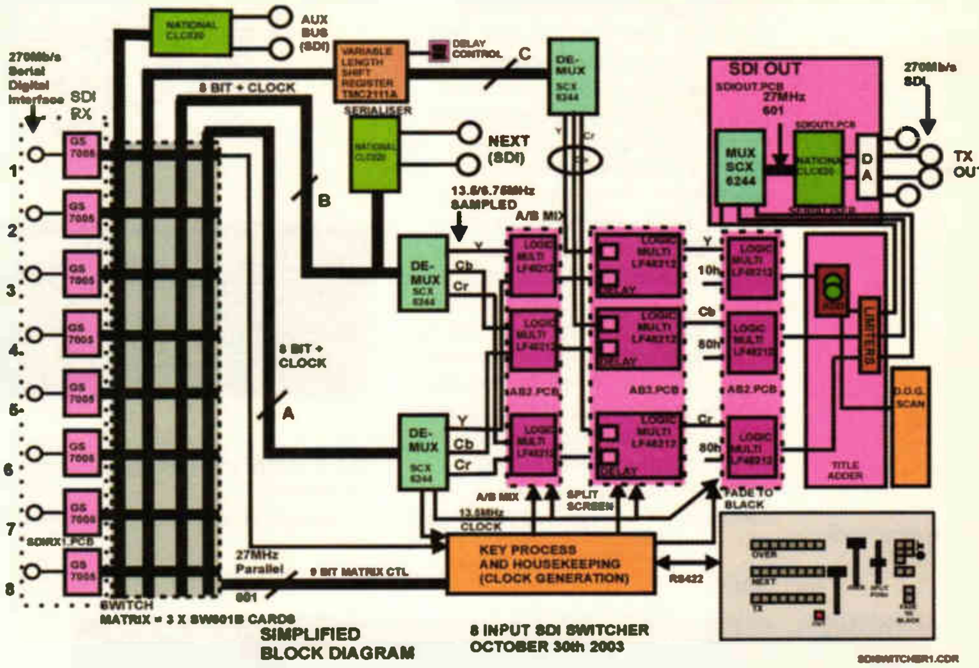
A pair of SDI equalisers were built before the mixer project started. These proved useful at IBC some years ago when a long length of indifferent quality coaxial cable was used for a particular run, and the signal was beginning to go over the digital 'cliff'. These use Gennum GS9004/GS9007 parts, each giving three outputs.

Brief tutorial on SDI

Standard definition digital television involves sampling the analogue-luminance signal (Y) at 13.5MHz (864 x H frequency), and the two colour difference (Pb, Pr) signals at 6.75MHz. Due attention has to be paid to levels at the input to the ADCs. Black level for digital signals is set at 16 (10h), and peak level at 235 (EBh) for Y and 240 (F0h) for Cb and Cr. Note that Pb/Pr notation is used for analogue colour difference signals, and Cb/Cr for their digital version. Care must be taken in any digital processing to ensure that levels do not stray into the 00h or FFh region as these levels are used to indicate End of Active Video (EAV) and Start of Active Video (SAV).

The three streams are multiplexed together with an effective speed of 27MHz. At this point the streams are parallel 8 or 10-bits wide. The object of SDI is to serialise the parallel signals into a single stream at 270Mbits/sec. The stream is sent at around 800mV pk-pk on a 75-ohm coaxial distribution system. Thus, standard good quality video cable and jackfields can be used. Because of the high frequencies involved, there is a limit of around 300 metres for signal transmission using the best quality 75-ohm cable. The failure is very sudden, and is called the 'digital cliff'. Good installation practice would introduce equalisers at around 250 metres in the run. Devices called 'cable clones' simulate lengths of cable and allow testing of equipment and cable runs by adding 'cloned' cable until the cliff is reached. This will indicate the margin of error in the particular run.





***Top:* Levels**

***Bottom:* Block Diagram**

Start of design phase

Early in 2002, the project was started. It was a bit like trying to assemble flat-pack furniture – little or no instructions; and have I got all the bits? The first action was to produce a block diagram of the mixer as a guide for the detailed design.

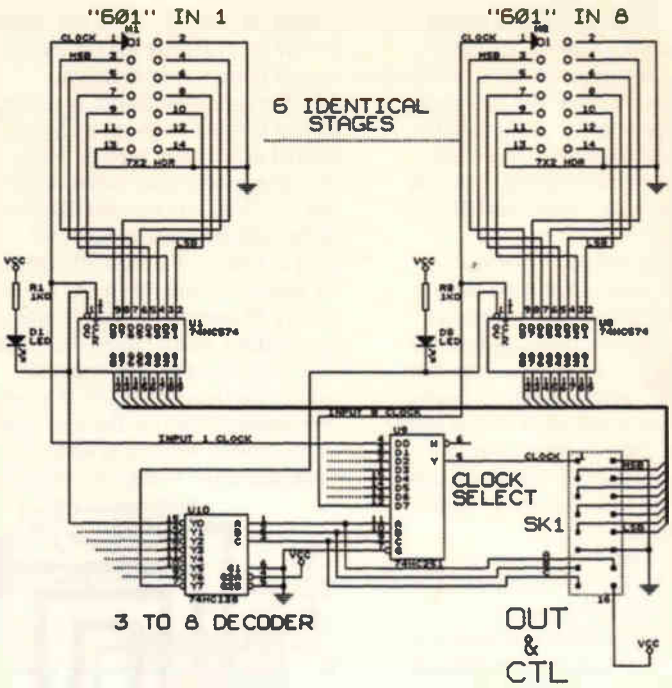
Before this, decisions had to be taken about the internal switching arrangement. There were two options. Either to switch the SDI signal (along the lines of the article by Emil Vladkov in the August 2003[3] issue of *EW*), deal with de-multiplexing the SDI bus signals to CCIR601 and then to the components YCbCr for mixing and split screen; or to have an SDI receiver for each input, and switch the resulting '601' signals. This latter makes for an easier switch matrix design in that 74VHCxxx parts can be used as they are only handling 27MHz clock frequency, and the layout becomes much simpler. After the '601' signals are de-multiplexed to YCbCr, the clock rate drops to 13.5MHz, which is even simpler.

Input section

Most SDI receiver systems use at least two chips, but Gennum offer a single chip receiver, GS7005, which has built-in equalisation for up to 75 metres of cable. For a video mixer, where the sources are relatively close, this was not a problem. Also by using PCB mounting upright BNCs, the SDI signal path length was about 1 inch, thus minimising any return loss issues. The GS7005 has a 52-pin QFP package, and one of the early jobs was to create a Boardmaker library shape for it. Deep joy was in the house when the chip was married up to a print of the shape, and it fitted.

The eight GS7005s were laid out on a strip PCB 270mm by 38mm, and was designed to fit directly to the rear panel of the mixer. Each receiver chip has a pin, which can light an LED to indicate a valid input. These pins were also remoted to a panel at the front of the mixer to give a clear indication of which inputs are being fed.

The 601 outputs were taken away on 14 pin plugs, mating with the 14 pin IDC sockets on ribbon cables. Note that I refer to '601'. This is an unbalanced version, perfectly valid to use within a unit. True CCIR601 uses balanced signals, specially twisted ribbon cables, and D25 connectors.



Switch Matrix Card

Several problems showed up

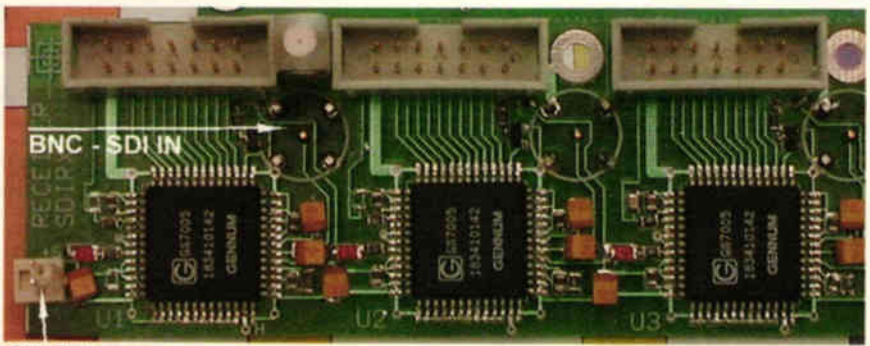
One of the inputs did not work, and on investigation, it seemed as if the chip was suspect. Gennum's agent very kindly sent a replacement, and the original chip was carefully removed. This was achieved by grinding down the blade of a cheap Stanley knife so that it could slide under the legs of the chip once they had been de-soldered. Unfortunately, I do not possess an infra-red reflow tunnel so all SM components have to be soldered using solder cream and the finest solder bit that Weller offer.

When the chip was replaced, it still did not work. A close examination of the rear of the board showed that a 'via' hole that was supposed to connect to the rear ground plane did not. A scrape and some solder cured the problem.

Another problem was the +5 volts distribution. Power is fed in near input 1, and a long track takes it down the board. The +5 volts at input 1 fell to around 4.5 volts at input 8. A piece of 30 SWG wrap wire in parallel cured any problem due to volts drop. The proper cure is of course to use wider track.

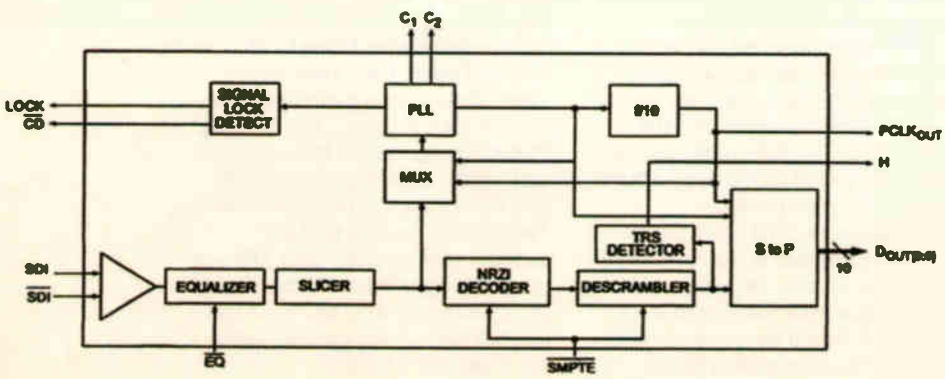
SDI receiver card

"601" OUT



POWER +5 V

PART RECEIVER CARD



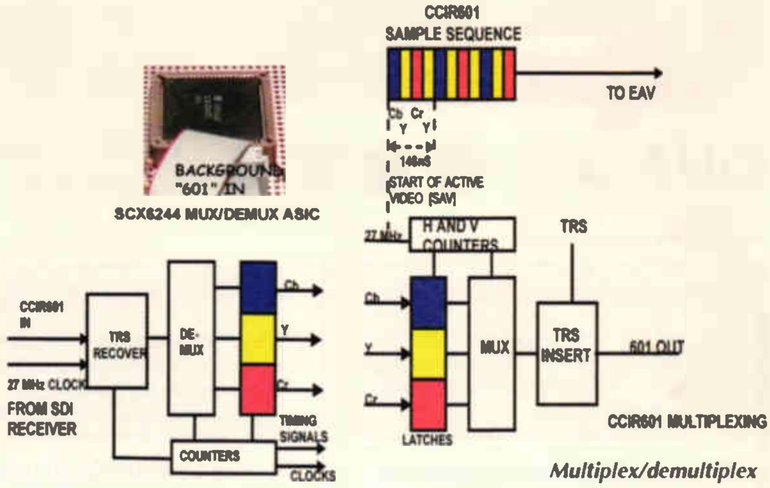
Switch matrix

Attention now turned to the switch matrix. The choice of switch element was the 74VHC574 8-bit latch. A card was laid out with 8 of these crosspoints with their outputs in parallel, and the output enable pin (/OE) used for crosspoint control. Each latch uses the clock input from the '601' input. The eight clock feeds are then selected by a 74HC251 8 to 1 multiplexer. A three line to eight-line decoder is used to translate the 3-line control input to each of the output enable pins. To indicate matrix status, SM LEDs are fitted to each crosspoint. The output and control connections to each card are via a 16-pin socket, designed to mate with a plug on the mixer motherboard. Each card therefore makes up one bank of the matrix.

The inputs are on 14-pin headers, which mate with IDC connectors on ribbon cable. Note that each SDI receiver drives a 4-drop cable,

looping through the 4 matrix cards. Four PCBs for these cards were ordered, although originally only three banks were to be fitted. At a late stage, it was felt that it would be useful to populate and fit the fourth card as an Auxiliary bank (AUX). To return the '601' signal to SDI, a device such as a National CLC020 serialiser (or its Gennum equivalent) is used.

One down side to switching the 601 stream showed up at IBC. We were using the AUX bank to route a source to a recorder. The source had audio embedded in its SDI output. When it reached the recorder, it was mute. By de-multiplexing to 601 and then re-multiplexing in turn to both 601 and SDI, the embedded audio had been stripped from the stream. Direct switching of the SDI signal would have obviated this problem. However, the layout of the matrix would have been more complex, and would have involved multi-layer boards. For the A, B and C banks, the SDI receiver chips would be placed at the output of the matrix, followed by the de-multiplexers to deliver YCbCr to the Mix/Effects (M/E) system.



Multiplex/demultiplex

De-multiplexing

Because of the way in which the components are multiplexed in the CCIR601 parallel stream, it is not practical to carry out any operations on it other than switching. Therefore the signal has to be de-multiplexed back to the original components.

Some years ago, I was given as samples a pair of multiplexer/de-multiplexer ASIC devices (SCX 6244) that had been designed by a consortium of UK broadcast companies such as Aston, Acron, Avitel and Cintel. The device can operate as multiplexer to take in Y, Cb and Cr 8/10 bit parallel data and give out 8/10 bit parallel '601' signals. By taking a pin (39) low, it operates as a de-multiplexer, giving three parallel streams (Y, Cb, Cr) from a '601' input. Unfortunately there was a small snag in the masking of the device, resulting in occasional upset of the order of outputs, with Cb coming out of the Y ports etc. To correct this, a PAL was supplied in a 16-pin package. This corrected the counter reset problem.

Through friends in the industry, I was able to acquire several more of these devices and had enough for each of the three banks of the mixer as de-multiplexers, and one for the output multiplexer, as well as one spare.

Clock and housekeeping

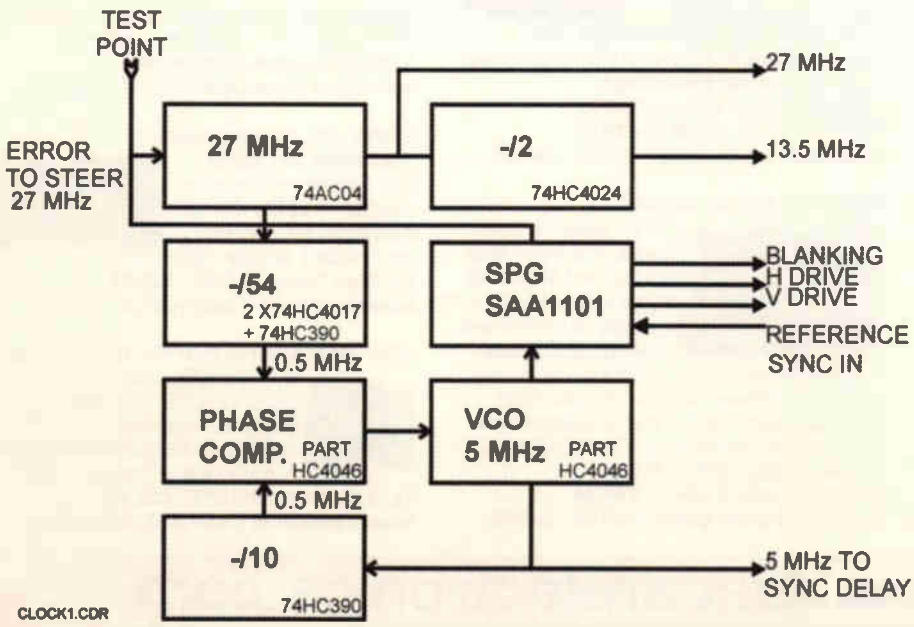
We have now recovered the digital component signals YCbCr from each of the three banks of the switch matrix, and are ready to mix, split the screen and finally fade to black if needed. However clock and housekeeping signals have to be considered.

The assumption was made at the start that all signals into the mixer would be synchronous. This is a very practical point as all the sources used at IBC are referenced to a master black and burst signal. Accordingly, a local Sync Pulse Generator (SPG) was designed into the mixer such that it could be genlocked to the external house reference feed. The master oscillator for the SPG is a 27MHz crystal; so clocks at 27MHz and 13.5MHz are readily available. The output multiplexer requires 27MHz clock, and special H and V drive pulses that have to be carefully timed. These are derived from the SPG outputs.

The SPG is based on a Philips SAA1101 device. A genlocking colour SPG using this device was described in an article[1] in *EW* for February 1996. In the SPG for the mixer, colour subcarrier lock is not required. For sections such as the split screen unit, H and V drive pulses from the SAA1101 are used as counter resets, and 13.5MHz and 6.75MHz feeds are used for clock and for the H counter. Mixed blanking is also used to ensure that no key signal strays into prohibited time zones.

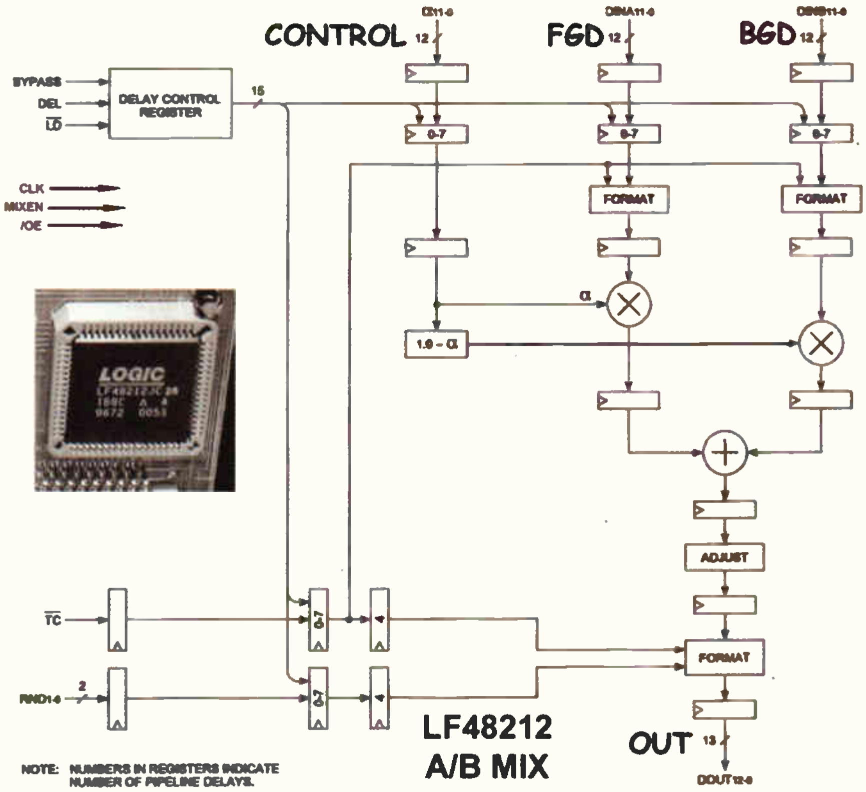
In the early experiments, the SPG used a 5MHz crystal oscillator, using the internal phase comparator to lock to input reference mixed sync. In a second loop, the 5MHz frequency was divided down to 0.5MHz and compared with the output of the 27MHz clock oscillator divided by 54. The phase comparator section of a 74HC4046 was used for this, and the error signal used to steer the 27MHz oscillator into lock. After the

Clock and housekeeping



mixer had been used in anger, it was found that this was not the optimum arrangement for jitter performance. Subsequently, the 5MHz crystal was removed from the SPG and the 74HC4046 reconfigured to act as a 5MHz VCO. The VCO output fed the SPG oscillator input. The dividers remained as before, and the SPG error steered the 27MHz clock oscillator directly. This resulted in greatly improved jitter performance.

It was also found that some sources exhibited a long time delay between their reference inputs and the SDI output. Consequently a large passive delay was necessary in the reference input feed to the mixer to avoid pictures shifted over to the right. Accordingly an internal delay has now been fitted. The sync signal from the reference input sync separator is fed into a 15-stage shift register, clocked by the 5MHz signal from the SAA1101. An 8 x 1 and a 2 x 1 multiplexer are used to select the outputs, giving a sync delay adjustable in 16 x 200nS steps. A hex push button switch on the front panel completes the modification.



Mix-effects (M/E) system

A number of companies offer digital multiplier devices, but Logic offer one of the most interesting. Their LF48212 has two 12-bit inputs, and a 12-bit alpha or key input. The output is such that it equals a x A + (1 – a) x B, just what we want for dissolves, fades or split screens. Accordingly, 11 of these were purchased, and a (ME) card was designed to carry three of them. (For the three channels Y, Pb, Pr). Three cards were needed, for dissolve between TX and Next Event banks, for split screen and for fade to black.

Experience has shown up another negative. The Split Screen unit has one of the LF48212 devices as a Key fade stage. This is of no great consequence in a split screen mode, but if a luminance key signal is taken from the C bank input to the split screen M/E card, the key will be about 7 clock cycles later than the fill. This is due to the latency in the LF48212. A second glance at the device block diagram shows why. There are seven latches from input to output. A PhD in maths is not needed to work out that this gives a latency of 518nS. As this is 1/100th of the screen width, it is fairly noticeable. Consequently a new version of the M/E card has been made which carries a programmable 8-bit wide 16-section register (TMC2111AR3C) for each of the six inputs. The key signal is taken off before the registers, which are set to correct the delay due to the key processing stage. Satisfactory luminance and very simple 'blue screen' keying using the Cb signal has been achieved.

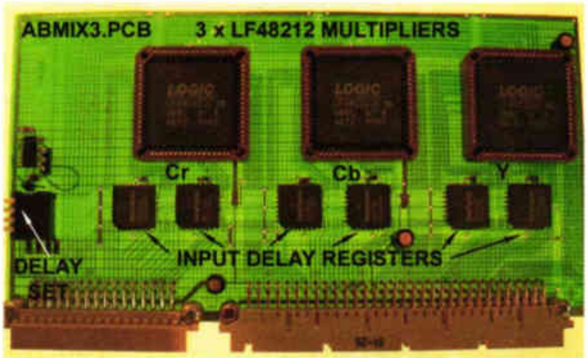
Split screen

The split screen card produces ramps at H and V rate, appropriate selection system, a full adder to add the ramp signal (8-bit) to the fader signal, followed by a clipper to tailor the signal to give the required split screen. Ramps are generated by EPROMs as look-up tables, and the clipper is another EPROM, programmed with a series of slopes to allow for changes of softness on edges, as shown in Fig.10. In Key mode the ADC, normally used with the split fader, is used with a potentiometer to give a KEY CLIPPER control.

The card is designed as a split screen rather than a wipe generator; hence the 8-bit ramps are not a problem. Obviously to go completely from one side of screen to the other would require 288 levels for the horizontal, and 351 levels for the vertical. This requires 9-bit EPROMs, or using some other device.

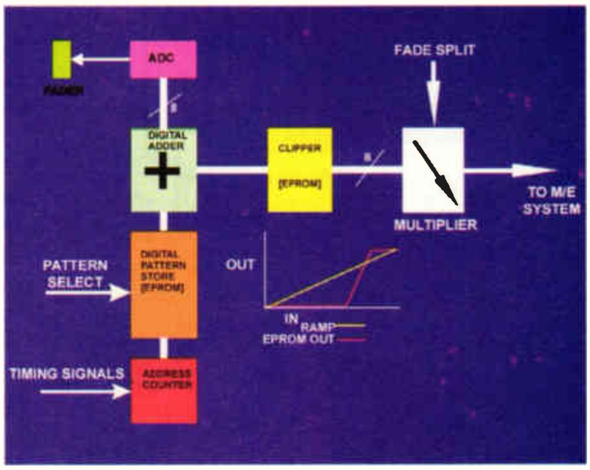
Fade to black

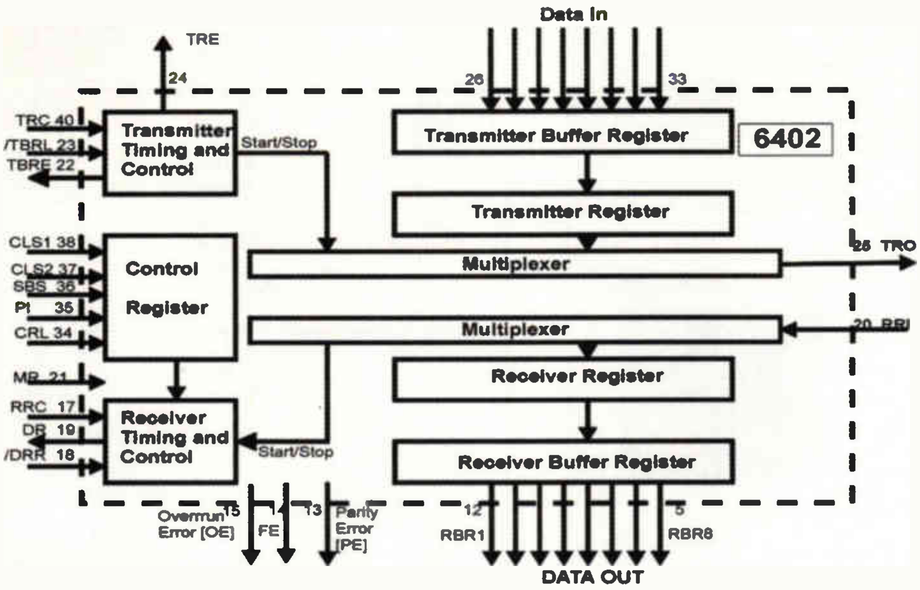
This uses another M/E card, with the output from the split screen card as one input, and the other input hard wired to 10h (00010000) for Y, and 80h (10000000) for Cb and Cr.



Above: M/E system and M/E card.

Split screen system



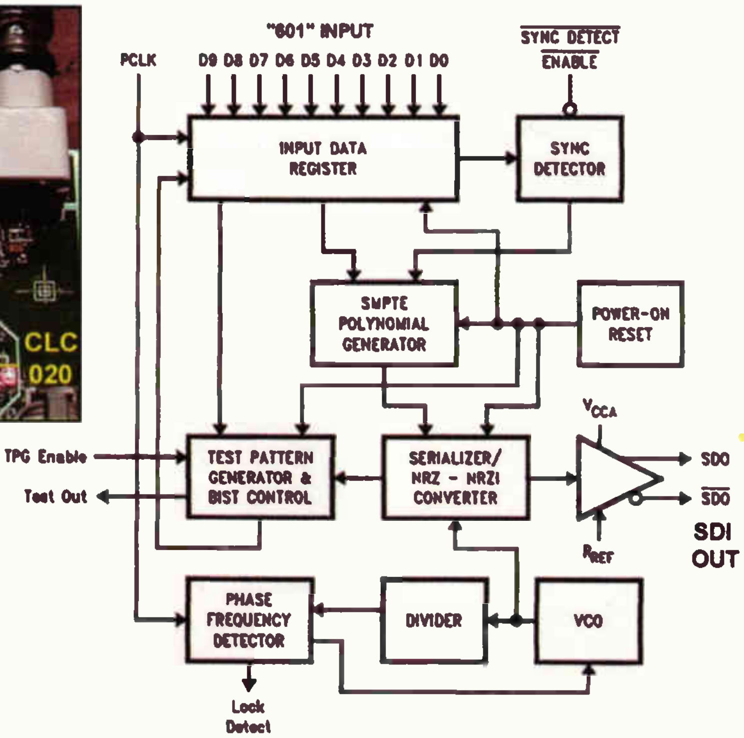


Block Diagram of UART

Control could be either from a fader, or from an up/down counter initiated by push buttons on the control panel. For simplicity the counter approach was chosen. A pre-set clock frequency control on the card is set to give around ½ second fade time.

Limiters and title addition

In the tutorial section above, the need for the digital component streams to avoid the levels 00h and FFh was stressed. To ensure this does not happen, three EPROMs arranged as limiters are fitted in the path from the fade to black M/E card to the output multiplexer and serialiser. The opportunity was taken to provide a simple title or channel ident injection facility into the Y channel only. This is achieved by using a full adder (74HC283 x 2) stage in the Y channel, together with a switch to control the insertion; and to equalise time delay, dummy full adder stages are fitted in the Cb and Cr channels. Any title video can be used, but a card has been produced to generate an appropriate logo/message. This uses an EPROM, which is programmed with the title and appropriate scan counters to read it. The EPROM is programmed so that the highest level is at 80h, i.e. half peak amplitude. The title is added to the video rather than keyed, and hence the limiter prevents peaks exceeding EBh. To give confidence to the operator, a digital to analogue converter is provided on the card so that a simple monitor can show the title in the EPROM.



Block diagram of the serialiser, and its place on the output card.

Output section

Following the limiter and titler section, the three streams Y, Cb, Cr are fed to the output card. This carries a multiplexer (SCX6244) and a serialiser (CLC020) and output driver (GS9007). This card is mounted on the rack unit rear panel, and provides a very short path for the four SDI outputs. An identical card, without the multiplexer or output driver, is used to serialise the '601' stream from the next event bank. Two next event SDI outputs are provided.

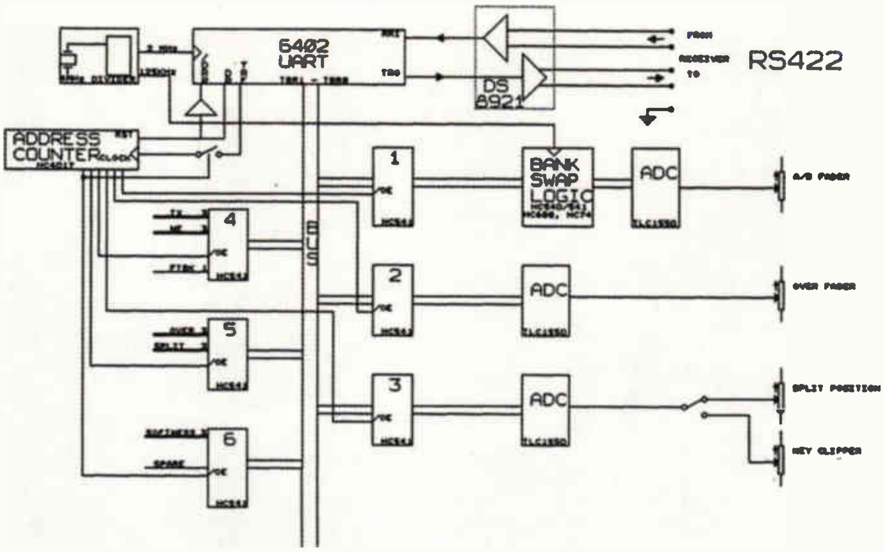
Control

In these sections, the video path has been described, but only passing mention of controls. For the mixer's first outing, the control panel from the analogue units was used, with the ADCs mounted at the rack end. It worked, but it was a bit small and not very user friendly. Furthermore it used two ribbon cables, one 25 way; the other 15 way.

Accordingly, thought was given to a panel that was specifically designed for the job, and which would carry two integrated LCD monitors for TX (Transmission) and Next Event. For the 21st century, it seemed ludicrous to use wire-a-function, and a multiplexed system was chosen. The T series mixers that my old company made used serial communication using RS422 protocol and a two pair cable as the interface. However it used microprocessors, and you will recall that I am an analogue dinosaur. So a compromise was chosen – a UART. This is a fancy name for a bi-directional serialiser.

I had done some work on an analogue processing amplifier, which was to be controlled using RS422 protocol and using a 6402 UART. The project did not go ahead, but there were some prototype cards lying around. These were set to work – after about 10 years shelf life, and a design began to emerge.

The way the panel operates is that a source is previewed by selecting it on the Next Event bank either by pressing the CUT button, or by moving the A/B Fader over, that source is now sent to transmission. The source that was previously on transmission is now on the Next Event bank. In this mode of operation, the fader end position is



Block diagram of control panel

irrelevant.

Latching of buttons is achieved using a counter (74HC161) and a 4 to 16-line decoder (74HC4514). When a button is pressed, the counter is enabled and counts round until its decoded output puts a 'high' on the LED (which illuminates) and simultaneously de-enables the counter. The Qa to Qc outputs of the A and B counters are stored in a latch (74HC574) and the output of this store taken to the load inputs of the counter for the other bank. When the CUT button is pressed, or the fader ADC reaches FFh, the bank swap process is initiated, and the stored information exchanged between the banks. The C bank uses a similar latch arrangement but without the store.

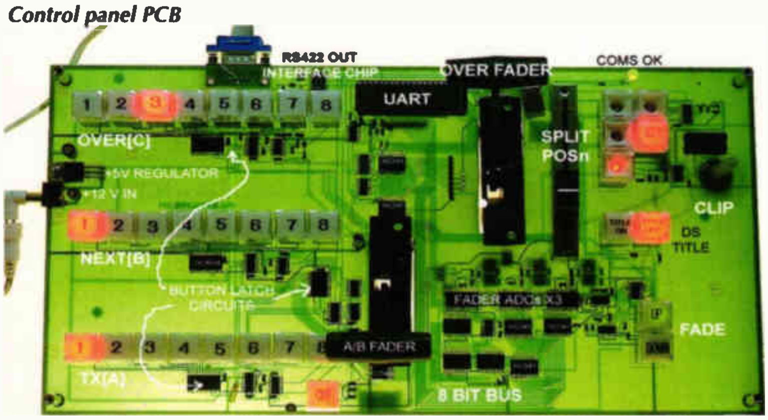
Three ADCs are used with the faders. To generate the bank swap flag, the A/B fader ADC is followed by a digital comparator (74HC688). When the fader ADC output reaches FFh, the bank swap results and the flag also toggles a bistable stage to select either the ADC output or its complement. The other ADCs are used with the over (C bank) fader and the split screen position fader, without any complication except for a NAND gate to signal when the over fader is off the end stop. This is used to gate an oscillator to make the LED in the selected C bank button flash brighter to indicate that the over system is active.

The UART system comprises a transmitter at the control panel, and a receiver at the rack unit. Both UARTs are clocked at 2MHz, giving a baud rate of 128Kb/second. The receive UART sends a word to the panel initiated by a field pulse. The transmitter UART receives this and this initialises the data transfer process. Each UART has 8-bit inputs, and these are fed via six buffers (74HC541) with their outputs in parallel. These are fed with the 8-bit words corresponding to their function. Words 1 to 3 are from the fader ADCs, while Words 4 and 5 are for source selection, fade to black, pattern selection, title insertion etc. Word 6 is spare for future use. Thus 48 bits of information are sent during every field-blanking interval.

As soon as the first word has been sent, an end of word flag from the UART is used to clock a counter that moves selection on to next word. After the UART has had the last word, the counter is reset for the next field. The receiver card at the rack unit drives a series of latches (74HC574) in parallel. The same counter arrangement as used at the transmitter generates a series of clocks at the correct time to clock the words into their respective latches. The latch outputs are then taken to their respective destinations. Note that all changes occur during the vertical interval, so there is no unexpected picture disturbance.

To give comfort to an operator, a re-triggerable monostable circuit is fitted to each UART's End of Word pin. When all is well, the monostable Q output will be high. Should the RS422 link fail, there will be no End of Word signal, and the monostable output will go low. A bicolour LED between Q and /Q of each monostable will be green when all is well and red when communication fails. Apart from the UART and the line receiver/driver, all the circuitry is surface mount, on a double-sided circuit board.

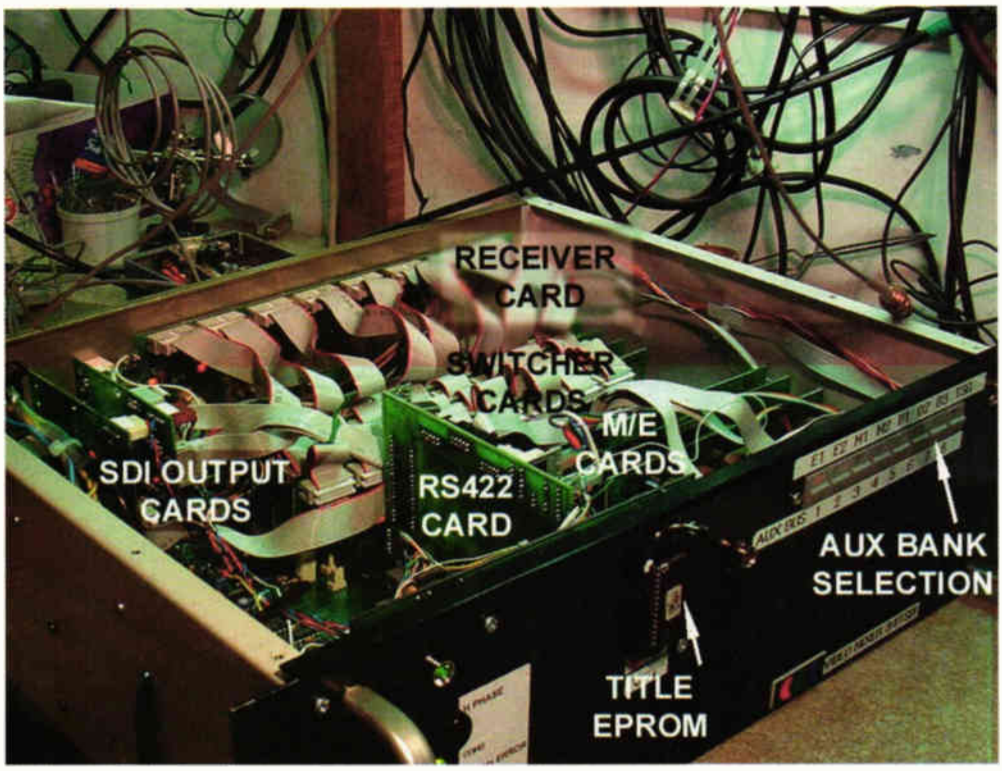
Control panel PCB



In the first trials of the mixer using the old analogue control panel, tally outputs were integrated into the panel. For the serial control panel, the necessary information is available at the rack unit, and that is where the tally card is. Every input has its own relay to provide a set of closing contacts when that input is 'on air'.

LCD monitors

The two monitors are based on 7-inch LCD panels, switchable between 16:9 and 4:3 operation. These have PAL inputs, so an SDI to PAL converter is necessary. The converter chosen is made by ShootView (DAC271) and has analogue component outputs as well as PAL. A suitable case was found in the Maplin range, which accommodates the LCD panel, the SDI decoder and an auxiliary card for power and video distribution. The size of this case defined the width of the mixer control panel. The LCD panels require 12 volts at around 700mA, whereas the SDI decoder normally takes in 7.5 volts which is sub-regulated to +5 volts on board. A more efficient process is to use a switch mode buck regulator from 12 volts to 5 volts. This was done using a National device (LM2575), mounted on the auxiliary card. Thus a single 12 volts supply drives the monitor and decoder. Switching is provided on the front panel for aspect ratio and for SDI/PAL selection. Switching is provided between the PAL output of the SDI decoder and an external PAL input. The switch out is also fed to a BNC at the rear so that an extra PAL



Rack unit

monitor can be used for mixer outputs. The YPbPr outputs are also brought out to the rear for similar reasons.

Power

All the rack unit power rails are +5 volts; these are obtained from a switch mode block supply giving a maximum of 4 Amps. Individual LC filters split the supply up to the various sections and cards. The control panel requires +5 volts at about 100mA. As each monitor requires 12 volts at around 1.1 Amps, a 12-volt, 2.5 Amps switch mode block supply is used, with a linear regulator providing 5 volts for the panel electronics. Monitor power connections are made using 2.1mm plugs and sockets. This allows the monitors to be used elsewhere if necessary using Sony type BP90 batteries, or NP1 batteries with an adapter.

Metalwork

The rack unit uses a proprietary tray, standard 3U panel and a purpose built rear panel. As previously mentioned, the receiver card is mounted directly on the rear panel, as are the TX and Next Event output serialisers. The rest of the electronics sit on a motherboard mounted in the base of the tray. The three M/E cards are plug-in, using DIN41612 connectors, the switcher cards are also plug-in, and other cards are mounted on the motherboard and connected using Molex or IDC ribbon interconnects. This has made for a remarkable degree of flexibility, very desirable in what is an experimental unit.

One of my former colleagues, now in the metalwork business, very kindly offered to make the control panel. He also replicated the Maplin's monitor boxes, saying that it was as easy to build from scratch as to pierce an existing box.

Conclusions

It is very gratifying when a project succeeds in all its objectives. Perhaps only someone who has had to struggle with designing, aligning and setting up analogue component video mixers can appreciate the complete absence of 'tweaking' in the digital equivalent. If it is assembled (and designed!) correctly, it works!

Every piece of kit has to have a name, and as the architecture of this mixer is very similar to the 841 of blessed (?) memory, it is called the 841SDI.

This has been a challenging 18 months spent designing, proving and laying out circuit boards. The most costly parts of the process have been circuit boards. A total of 12 different types of board have been made, all double-sided, with colander ground plane where at all possible. The motherboard was the trickiest, with the interconnection to and between the M/E cards the most complex. I must confess to cheating, in that some of the inputs are done in wire-wrap wire. Tracking from three 84-pin PLCC sockets to the M/E connectors was too difficult on a two-layer board. A handbook of sorts has been produced, highly essential if one is to keep track of modifications along the way. Major vision mixer manufacturers such as Grass Valley or Ross have nothing to worry about, as this is definitely not a commercial project.

Finally, I must thank a few people such as Tony Frere and Andy Eden of DT Electronics, who helped with the supply of Gennum and Logic ICs and Veetronix push buttons; Graham Barker of Fastform for his panel making skills; PC Designs for their efforts in the PCB making area; and Phil White, IBC's Technical Resources and Events Chairman, who suggested the project in the first place. I must also thank Sheila, my wife, for her patience, particularly at tricky moments when the air got a bit blue and the cats were scared.

Finished Control Panel

References:

1 *Locking onto Video Sync*, Michael Cox, *Electronics World*, February 1996, Vol. 102, No. 1719

2 *Analyse Your SDI*, Emil Vladkov, *Electronics World*, February 2002, Vol. 108, No. 1790

3 *Professional SDI Router*, Emil Vladkov, *Electronics World*, August 2003, Vol. 109, No 1808

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Class-A imagineering: Part 3

Graham Maynard's investigation into loudspeaker induced amplifier distortion continues; here his simulations match some previously reported subjective findings

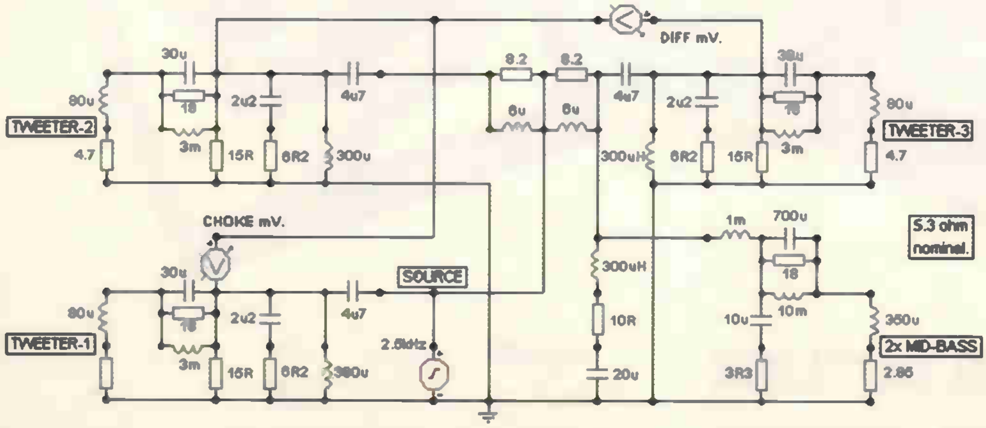
I could understand some readers still doubting whether I have conclusively illustrated how microsecond delays and the small quadrature potentials which reactively develop, go on to behave as I suggest and integrate to cause what I describe as a smearing of amplifier output in both amplitude and time, which then subsequently affects the sonic character, rightful position and stability of a reproduced sound stage image. However, we must remember that the transient errors caused by crossover filter section components and lower frequency drivers are due to entirely natural characteristics which act most significantly over tens of microseconds – long enough time frames to develop higher audio frequency potential changes – which thereby become capable of modifying tweeter drive due to their parallel action upon any interposed series impedance between the loudspeaker system itself and a low output resistance NFB loop controlled amplifier. Nor should we forget that some of these reactively generated back EMFs can be due to loudspeakers having become excited by waveforms that have gone before, but which may no longer be observable, or, worse from an amplifier's output powering viewpoint, have become shifted completely out of phase with driven waveform at its NFB loop controlled output node.

Series output chokes

For this reason, I simulate using the circuit of **Figure 7**, effects that the commonly used series output choke can have upon tweeter drive due to none other than its insertion between a low impedance amplifier output node and a loudspeaker system, as exampled here both with and without the additional connection of paralleled twin mid-bass drivers and their associated crossover plus driver compensation and impedance correction components which are essential for properly balanced full range reproduction.

Here the most significant back EMF induced errors shown in **Figure 8**, arise during the first 90 degrees of 2.5kHz sinewave amplifier drive. The combination of loudspeaker system components acting throughout their initial 'transitory' inductive delay and settlement periods have reactively generated the completely separate, different slew rate, parasitic and tweeter modulating, error waveforms illustrated.

Figure 7: the circuit for examining tweeter drive error at 2.5kHz.



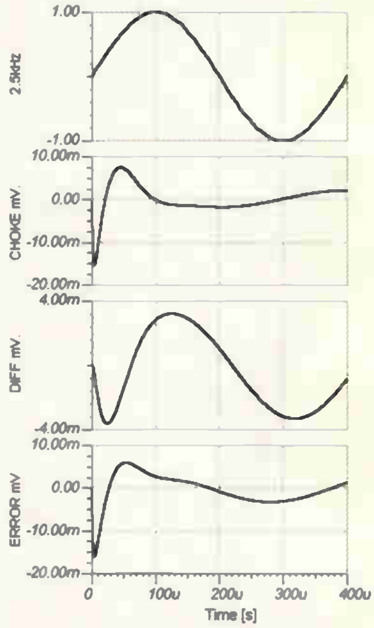


Figure 8: The higher energy tweeter driving error components caused by 2.5kHz excitation of parallel connected crossover section circuitry.

The simulated 'CHOKE mV' trace of Figure 8 illustrates the effect upon Tweeter-2 of the treble-section reactance interacting with current flow through a 6μH//8.2Ω series output choke. This causes an initial negative error (minute delay wrt direct connection), with a natural (catch-up) settlement of crossover section voltage development, plus an overshoot that is determined by the treble section components themselves and which hopefully is properly electrically damped, as here and reasonably so by the series connected 2μ2F+6.2Ω. After 100μs, there remains a minute but phase shifted steady state 'undistorted' residual. Note that this is not the actual voice coil waveform apparent at Tweeter-2 itself, but what is equivalent to a series output choke insertion error development due to the crossover-section loading and back EMF generation, when compared with the directly connected Tweeter-1 option.

The waveform appearing at Tweeter-1 is already crossover section modified, and although less accurate and more poorly damped than it could be with a line-level signal crossover and direct connection between amplifier and driver, it remains useful as a directly connected reference. The waveform at Tweeter-2 has been series output choke modified during the first 90 degrees of 2.5kHz drive due to approximately 4μS of treble-section crossover to output choke lead-dividing wrt the amplifier, followed by an entirely natural settlement between the series output choke, section components and the driver itself. The induced error has been reactively modified by the treble crossover section, and this clearly indicates a complex inter-reaction; the less reactive a treble driver and its section components, then the less image distorting the resulting electrical delay and overshoot error will be.

The Figure 7 test circuit also illustrates the way in which a parallel connected mid-bass driver section with additional but necessary impedance correction circuitry can further affect tweeter output when an amplifier incorporates a series output choke; see the 'DIFF mV' trace simulated in Figure 8. Here capacitive loading of the additional mid-bass inductors, has separately divided a fraction of amplifier output potential across the series output choke; there is a lower level, but longer acting delay, circa 20μS, plus an entirely separate natural recovery into the undistorted steady state 2.5kHz sinewave at Tweeter-3, wrt Tweeter-2. The now compositely utilised Tweeter-3's transient response has been additionally inter-reactively modulated by no more than the parallel connection of mid-bass driver and compensation components, such that the tweeter potential now carries a quite separately generated component from the mid-bass section, which looks like almost 4mV of a more highly energetic third harmonic wavelet, before it too runs into its own steady sinewave residual, again from 100μs onwards.

The 'ERROR mV' trace in Figure 8 shows the combined effect of waveform distortion at Tweeter-3 when compared to the directly connected drive of Tweeter-1. Both crossover section error contributions are identifiable; though they result from different crossover section delays with differently timed overshoot errors that increase in amplitude with frequency until each runs straight into phase shift. Both sections contribute to the generation of a characteristic leading edge 'output choke' induced sound that is discernible on sibilants, transients, etc. Here, where the crossover driven tweeter has its own -12dB (steady state) first half cycle 2.5kHz contribution at approximately 200mV (dynamic), there is a more clear illustration of dynamically induced transient distortion than the steady state and common ground sensing illustration of Part 2, Figure 4 suggests; this being followed by a reactive 'catch-up' plus overshoot that is akin to those illustrated in Part 2, Figure 6, though generated via more realistic crossover-driver inductances with their longer self delay periods.

We are very unlikely to hear these errors on a single sinewave start-up, but when their relatively frequency-constant crossover-section induced effects are repeatedly superimposed upon wide ranging music waveforms, they are likely to engender an audibly recognisable sonic affectation, and I suggest that readers might listen for these sorts of differences for themselves by inserting a small inductance in series with their loudspeakers. Also, it is not as if the mid-bass driver can itself generate a matching contribution that will correct tweeter output, because the short time delay which creates the induced tweeter error arises more quickly than the mid-bass driver is capable of responding unless directly connected in the first place; waveform changes arise more speedily than the mid-bass-section can itself react.

I have no doubt that the reproduction from Tweeter-1 could have been deemed reasonable in its own right, but the reproduction via Tweeter-3 will be notably different due to additional leading edge modulation by what is not just a normal-forward output choke delay induced error, but almost 100μS of mixed and reactively induced-reverse, and no longer phase linearly damped, circuit inter-reaction.

I note that;-

- the timing and frequency of the differently delayed 'transient' waveform settlements into steady state are due to individual crossover-section characteristics, not the series output choke;
- the crossover-section inductor-capacitor delay induced error increases with frequency until it runs straight into phase shift;
- the natural frequency at which each individual crossover-section becomes momentarily excited does not change with start-up frequency, and thus repeated dynamically induced affectations become a noticeable superimposition;
- also, the cross-coupling between excited crossover-sections, and the loss of phase linear amplifier controlled reactance damping, increase with the value of any amplifier's series amplifier+choke+lead inductance.

Reproduction fidelity

We could go round and round in ever increasing circles discussing the fundamentals of transients - sinewaves - group delays - driver response capabilities, etc., but one most important point stands out here - this being that it is the composite loudspeaker's individual crossover-section delays and back EMF induced reactions arising at the output choke itself that cause the development of these frequency constant and thus characterisingly recognisable errors. Should there be any argument about what exactly the choke generated error potentials from my Figures 7 and 8 comparison illustrate, then imagine the relative frequency and voltage errors that will arise when a good NFB amplifier is not fitted with an integral series output choke – there won't be any undesirable leading edge errors – not ever – that is as long as the loudspeaker leads are both thick and short, and the amplifier's internal delay, stabilisation (C.dom) and NFB arrangements do not make the amplifier itself behave auto-inductively in the presence of delayed loudspeaker system generated back EMFs. Often the high audio frequency delays due to composite loading do not become fully established until after the first half cycle, with initial lag turning into an additional steady state leading load such that first cycle output and loading is much reduced compared with the second cycle onwards; this first cycle weakness cannot be observed and measured via steady state testing procedures, this leaving us with a loudspeaker which might 'sound' smooth and be easy to live with, but which is dynamically challenged! This is covered in Part 6.

The composite loudspeaker model I am using for these simulations induces lower levels of series output choke induced waveform distortion than do some other more complex loudspeaker types; I do however want to stick with this circuitry because it is representative of modern design, even though some older loudspeakers with sharper crossovers and a lack of impedance compensation will induce worse and much more displeasing responses that will still not show up via swept or steady state investigation.

My simulation has demonstrated how, because it is the loudspeaker itself that reactively induces series output choke throughput distortion, changing from one type of loudspeaker to another, can simultaneously change the nature of the choke generated transient distortion, and thus alter the amplifier's reproduced 'sound' much more significantly than can be simplistically explained by examining conventional measurement results. I redouble my emphasis here by stating that the resulting reproduction distortion is due to this amplifier's initially 'perfect', but series inductive (or interconnecting cable inductance) first cycle response to the loudspeaker itself. We are indeed humanly capable of describing dynamically induced amplifier (impedance) 'characteristics' that conventional test bench or SPL measuring equipment still cannot reveal, and it never has been the listeners who have been the problem. All good amplifiers that utilise series output chokes are likely to make a given type of loudspeaker 'perform' very similarly, but that does not mean that any of them is capable of making a loudspeaker reproduce realistically, no matter how impressive its specification!

This could relate to some of the treble differences I noted decades ago in the tweeters of three way loudspeaker systems, when I compared the sound of 'linearly running' but choke coupled solid-state class-B power amplifiers with my 50W class-A amplifier, and another superb home-built reference - a hybrid 4x KT88 ultralinear monoblock. The kind of waveform aberrations shown in Figure 8 make the human voice lose high frequency clarity, whilst imparting an additional throatily catchy, over sibilant, overly loud and as if prematurely distorted excitation response upon crescendos. The reproduction of wind instruments like saxophone and trumpet is affected too; whilst other similar loudspeaker induced amplifier (inductance) distortions cause an inconstant harshening confusion to the percussive attack of composite audio, thereby degrading the leading edge reproduction that the loudspeakers themselves should more naturally be capable of.

String instruments too are well known for producing sound that is rich in overtones, so how realistically can a two or three way composite loudspeaker system reproduce their music when an amplifier's series output choke permits an alteration of their leading edge harmonic signature in a manner that is directly related to the loudspeaker's internal circuitry? Output choke induced energy transformations make a tweeter reproduce incorrectly, when it isn't the tweeter itself that is the problem! When fitted with a higher value output capacitor the original JLH-69 has a resistive output stage which is incapable of allowing this kind of inductively generated first cycle harmonic intermodulation, but they were disadvantaged by a lack of realistic driving power, and that is why I have pressed on with this work.

The loudspeaker that I have based this computer model upon is the widely acclaimed *Ariel* design due to Lynn Olssen. To read more about its developmental history and DIY construction details, visit;- http://www.ariel.club.tip.nl.

There have been comments that the Ariel loudspeaker performs more satisfactory with most valve amplifiers than it can with most solid state designs, and from long experience I can understand such a statement. Now however, I believe that my imaged examples of the inseparably related time and amplitude and therefore stereo image changes that can be introduced by the output chokes already fitted to many existing solid state 'hi-fi' amplifier chassis, show that this wholly subjective statement does indeed have a provable theoretical basis. Note that the Ariel does make correct use of impedance compensation circuitry.

Output choke induced audio image smearing and peaking can be minimised by not implementing crossovers which operate at frequencies where our ears are so sensitive, as is possible with high quality wide frequency range drivers mounted against well damped or non-resonant enclosures. It is though, far more sensible to **not** use an output choke within an *audio* amplifier in the first place, and to minimise amplifier inductance at audio frequencies by **not** using a Miller connected C.dom at the push-pull voltage amplifier stage either.

No series inductance - no initial delay and no higher audio frequency back EMF induced smearing, thus a loudspeaker will have a much better chance of reproducing sound in the manner its designer intended.

This comment also applies to compensation circuits which might be inserted in series with a driver to counter peakiness or to provide baffle step compensation with a full range unit, as well as to all 'normal' passive loudspeaker crossovers themselves. Such frequency selective level adjustments are better implemented at signal level ahead of the power amplifier, so that amplifier-loudspeaker damping is properly maintained and the driver waveform cannot be reactively degraded by any driver-cabinet induced back EMF impingement upon that series impedance. A milli-Henry series choke with an ohmic parallel resistor which is used for baffle step 'compensation' might well level out the SPL frequency response for less strident reproduction via a particular driver and cabinet combination, but at what cost to its dynamic waveform integrity and the overall reproduction fidelity?

Such reasoning has also reinforced conclusions I developed from earlier empirical work; i.e. that crossovers and impedance correction circuitry must be sited as near to a chokeless amplifier's output terminals as is possible, with bi- or tri- wiring then separately completing each driver circuit. There is then no shared cable inductance to interfere with reproduction clarity due to a source amplifier having an extremely low driving impedance, which can quite literally turn a choke or a loudspeaker cable into a back-EMF energised high audio frequency spring. A sensibly chosen short plain lead in series with a single driver on the output side of a crossover has negligible effect upon voice coil current flow. Besides, an amplifier sited crossover arrangement facilitates the fine tuning of a composite loudspeaker's response by allowing crossover adjustment from the listening position, though the sound would then change significantly if, for reasons of completeness, the parts are subsequently reassembled at cable end inside the enclosure, even with bi- or tri- wiring. This is because parallel-connected and reactive crossover, impedance correction and driver circuitry still generates differentially delayed return currents that cannot fail to reactively modify (distort) transient throughput over some range of interacting frequencies.

No two loudspeaker systems sound exactly the same either, and even if they could one sited farther away from a low impedance solid state amplifier at the end of a longer cable will sound worse due to additional cable impedance audibly altering reproduction in a manner that cannot be explained by the fractional dB losses observable via swept steady-state measurement with sine waves. It is not exotically woven, deoxygenated or rare metal loudspeaker cables that are needed, just amplifier sited crossovers (input or output, but preferably input) followed by short, separate and sensible cable pairs out to each individual driver. From this, you should be able to imagine what I think when I see an impressively large showroom at a hi-fi emporium, which is full of expensive loudspeakers - each capable of being 'professionally' compared via switch or plug-box selection and trunked cabling! Hi-fi?

Series output capacitors

Unfortunately series output chokes are not the only components concealed away from user modification within an amplifier chassis which might subsequently affect reproduction. At this juncture I am referring to the output capacitors that are necessary when single ended or single power rail amplifier designs are used. By believing that high quality output capacitors have no more an effect than setting a reasonable subsonic -3dB point, some designer-writers continue to present such designs as if *audio* competent. I observed different capacitor induced reproduction changes with different loudspeakers before I built my solid state class-A 50W monoblock over thirty years ago, and descriptions like 'boomy', 'bouncy', 'hollow', 'soft' and 'woolly' immediately come to mind. I also observed that the nature of the sound changed according to the bass driver-auxiliary-baffle-enclosure combination being driven, and this is why I offer one more demonstration with my first simulation circuit.

Using the Part 2, Figure 3 layout with the top choke/resistor replaced by a 4,700μF (= 4m7F) capacitor to drive the simulated loudspeaker, and the bottom choke/resistor shorted out to give a directly coupled resistor measured reference potential; driving with a 55Hz sine wave causes the quadrature capacitor current flow 'back EMF' error that is simulated in Figure 9. This is due to no more than the capacitor's reactive insertion between a NFB loop controlled output node, and a real-world-like dynamic loudspeaker.

Yet again the worst waveform distortion arises during the first cycle, before the ongoing error becomes a quadrature component that will mix with the original to create a phase shifted but harmonically undistorted resultant.

Most valve and directly coupled solid state designs do not permit this degree of low frequency loudspeaker induced first cycle 'transient' waveform distortion to arise, and so if an output capacitor is essential to prevent direct current flow, then its value should be at least 22mF. A 4m7F capacitor might give a static -3dB point of about 5Hz with a passive 8Ω load, and on the surface this might measure as if it is superior to the performance any valve chassis can achieve; but just look at how it affects the dynamic presentation of a 55Hz first cycle. The waveform arriving at the loudspeaker terminals has become indisputably distorted by the capacitor's reaction to coincidental impingements of amplifier-forward output, plus loudspeaker-reverse back-EMFs.

When a capacitor is used in series with a low impedance output stage it is not just the amplitude and phase responses we need to be thinking about. As with the series output choke there is always an additional quadrature induced and spring-like first cycle storage and release of transient energy as the leading amplifier output interacts with driver, crossover and cabinet induced reflections and resonances. We must thus remain constantly aware of the way in which any energy exchange can affect ongoing audio waveform throughput. This capacitor induced first cycle bass distortion will not especially affect treble, but its effects can extend into and degrade the clarity of mid range audio as it alters the relationships of those bass tones that had previously carried dynamically combined harmonics which do not become equally phase shifted between amplifier and loudspeaker, with the result that the reproduced ambient image loses its coherent clarity, and bass notes their former firmness.

Owners might be able to improve the reproduction of older amplifiers that are fitted with a series output capacitor and/or inductor; - by replacing an ageing output capacitor with a modern more compact and higher value electrolytic, say 22mF; also by connecting a 1μH choke in series with a 0.22Ω resistor directly across the existing output inductor. Of course power rail electrolytics may need increasing too, which might suggest a further need for higher peak current rectifiers; also, observation would be necessary in case crossover distortion suppression and high frequency stability become challenged. Amplifier replacement is not always essential in order to achieve genuine loudspeaker driving improvements.

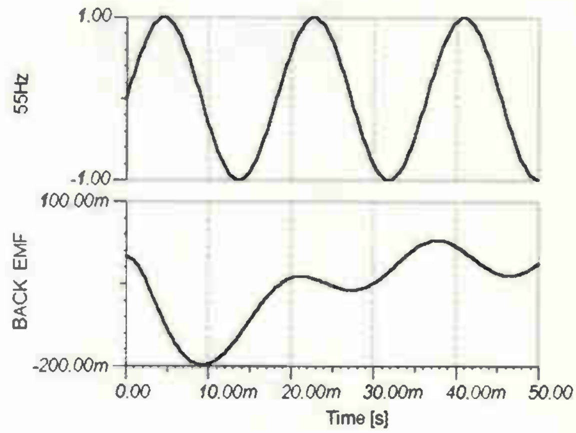


Figure 9: The first cycle error potential that develops at an amplifier's output terminals when a real-world loudspeaker loads an internal 4,700μF series output capacitor.

All of the effects I simulate here were audible long before home computers became available, so even though many good valve amplifiers, which take their NFB error sensing potential directly from the loudspeaker-output terminal, were incapable of generating such reactively erroneous waveforms, this first cycle distortion analysis cannot be considered as being some wizard new approach. What it does start to show however, is that enjoying music is not necessarily directly related to reducing steady state THD to an n'th zero, or ensuring low radio frequency pass-bandwidths with 100'sV/μS slew rates and micro-ohm output impedances, but to providing an amplifier which has a genuinely non-reactive plus low resistance output characteristic in order to minimise propagation and NFB loop delay, and thereby directly dampen all dynamically generated loudspeaker back EMFs! Actually I have come to understand that designers have distracted us by so fluently and frequently emphasising the needs for their narrowly focussed ideals, which of course simple low gain circuits cannot hope to match - though still without their 'basic sound' being as off-putting as is often the case with supposedly 'technically competent' architectures!

In part four, I will introduce a simple, very-low 'first cycle distortion', 25W-8ohm, JLH based class-A audio amplifier circuit without an input filter, C.dom, output capacitor or choke, though with low drift dc coupling and twin power rails.

An electric universe – magnoflux aether tunnel

Throughout the universe there is time. For if there was no relative time then nothing has space to move in, a reason for moving or time to even exist. At creation, there was an almighty flash of electromagnetic (EM) plasma. However, to prevent infinite dispersion, alternating frequency energy was restricted from escaping into outer space at the speed of light, thus allowing the concentrated plasma to condense and develop the universe permitting our planet (the Earth) to revolve around a star (the Sun), that God made part of it. Clive Stevens, C. Eng. M.I.E.E., explains

The probability of the sun and planets having the same surface voltage is remote in a dynamic system, thus consideration must be given to what physical effects a direct current (DC) voltage difference will have. To support light in space, it must be assumed that the in between volume of space contains electrostatic force fields and magnoflux magnetic aether fields, that once tunnelled, will allow the transmission of alternating current (AC) frequency Electromagnetic (EM) waves.

In classic physics it is assumed that mass attraction and tensors are the attractive forces between stars and planets. But we can obtain the same basic balance by considering it as mass attraction plus electrostatic charge attraction, because both forces are reducing in proportion to the square of the distance between sender and receiver. This is mathematically sound, and physically true because the e/m ratio of all matter is a constant. So overall mass attraction calculations need only a small correction to include the electric force set up by the unbalance of the positioning of the charges, meaning that the proton/electron pair do not quite weigh the same as the balancing neutron. This will result in planets having a negative environment and DC charge and the stars having a positive overall balancing DC charge and environment. A helium atom on a planet comprises a neutron plus proton pair surrounded or immersed in an electron enclosure. However, a helium atom on a star is envisaged as a neutron plus negatron pair surrounded by a positron enclosure which is in contact with the proton's shell creating a positive environment.

However, this change in concept to include electrostatics, will undoubtedly lead to an open universe, which is restricted in size only by the lack of magnoflux tunnels in the outer universe space to support EM transmissions. This restriction will effectively redefine the solar constants, as the only reason for EM energy to move across space is electrical.

The solar wind

The most obvious evidence of electricity in space is the solar wind. This material comprises mostly of positive hydrogen ions. From the positive ion's point of view, once it has been ejected it is being repelled by its star or the Sun's positive surface. The ion becomes aware that there are several negatively charged planets in the far distance attracting it outwards. This attractive electrostatic force accelerates the ion across space to, say, planet Earth. We now know that the solar wind takes up to 11 days to reach Earth. From this time we can calculate the average speed of the wind to about 155km/sec resulting with a final velocity of 310,000 metres per second.

From this, using e/m constant for a proton and formula: $\frac{1}{2} mv^2 = e.V$ (1/2x1/9.59x10^-7x310^2x10^6) we obtain the voltage per metre of about 500 volts. As the distance between the Earth and the Sun is 150 million kilometres, then the DC voltage will be around 75 million, million Volts.

Figure 1

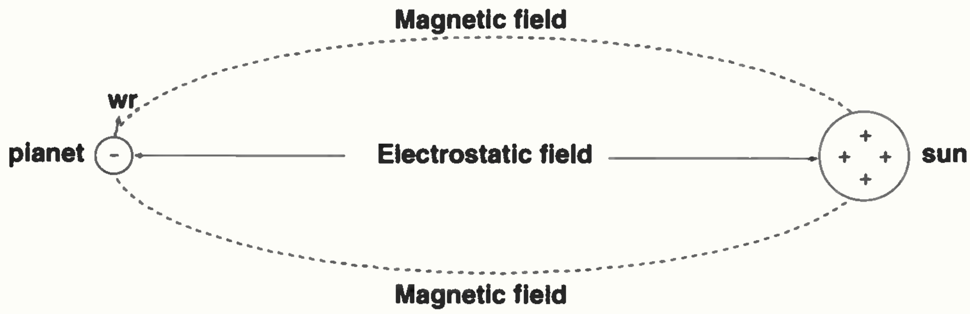
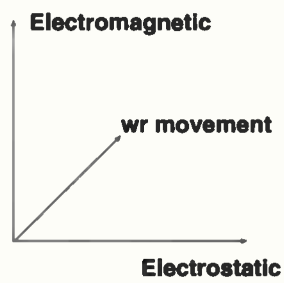


Figure 2



Solar system planet to sun charge attraction table

Classical							*Electrical*
T^4/D^2					SxP(weight)/D^2		
=1.4/0.318							
Solar kW/m² Constant	Planet (P)	Radius Asrt. Unit	Weight Astr. Unit	Distance to sun Astr. Unit	Attraction Astr. force	In earth Astr.units	Solar kW/m² Constant
9.3477	MERCURY	0.3794	0.056	0.387	373,909.15	0.83	3.6391
2.6783	VENUS	0.9541	0.817	0.723	1,562,951.36	0.55	2.4060
1.4000	EARTH	1	1	1	1,000,000.00	0.318	1.4000
0.6028	MARS	0.5292	0.108	1.524	46,500.09	0.05	0.2327
0.0517	JUPITER	11.195	318.354	5.203	11,759,873.75	0.03	0.1315
0.0154	SATURN	9.4701	95.222	9.539	1,046,481.61	0.00	0.0164
0.0038	URANUS	3.7002	14.58	19.192	39,583.76	0.00	0.0041
0.0015	NEPTUNE	3.4964	17.264	30.058	19,108.27	0.00	0.0022
0.0009	PLUTO	0.4704	0.93	39.518	595.52	0.00	0.0038
	Sun (S)	109.13	1,000,000				
	Galaxy centre		?????????			????	

To balance the hydrogen ion's positive materials arrival at a planetary environment, an equivalent amount of anti-positive neutron energy must be released from the sending star. We recognise this transmission as EM light.

The table uses mass to mass gravity as equivalent to +charge to -charge attraction for sun-planets but shows the differences between the classical and new electrical approach. The new DC magnoflux-aether tunnel approach conserves magnitudes of energy, as the EM transmissions are only released to balance a known electrical target and not everywhere equally. The calculated solar constant for the planets shows this divergence in philosophy.

The Sun generates and attracts the planet electrostatically using the DC magnetic aether field (magnoflux) at right angles to it as per Maxwell's law as shown in **Figure 1**. The planet then moves in the third right angle dimension around the Sun as per Fleming's electro-magnetic three dimensional rule shown below in **Figure 2**

The amount of Energy to be delivered is directly proportional to the electrical attractiveness of planet to the star/Sun. The electrostatic voltage reflects the mass to mass ratio through the fixed e/m unbalance ratio.

From the table it can be seen that the solar constant of Mars is 0.232kW/m²only a third of the amount of 0.603kW/m² predicted from the classical gravity approach, which could result in a space probes' battery failure. At the other end of the scale is Jupiter with a solar constant of 0.131kW/m² as opposed to the 0.052kW/m² predicted by gravitational physics.

Stars will repel each other DC electrically, this will cause a bowshock block or barrier in their magnoflux tunnel. Stars also repel any positively charged ions from another stars solar wind, making their solar wind warp around the intervening stars. The return electrons form a electro-magnetic magnoflux tunnel at right angles to the solar wind line as it moves towards an attractive planet .

Thus the space between stars and planets will be crisscrossed with mostly straight line DC magnoflux tunnels connecting individually each of the stars to each of the planets.

The law of conservation of AC plasma energy to move only within DC magnetised space or aether, was necessary at Gods creation to confine the Almighty plasma flash AC light energy and stop it moving away at the speed of light which would have lead to an infinite dispersion. This restriction still applies today, for if 99.99% of the stars/Sun energy is just blasted off into space which is 99.999% empty, then 99.99%of the energy will arrive at the edge of the universe with nowhere to go except outwards leading to a catastrophic dilution. As the edge of the universe is defined by a known amount of red-shift and the edge is not shining brilliantly, we can deduce that the random T^4 energy is not being emitted in all directions, as believed by Stefan but only through the DC magnoflux tunnelled aether filled space. Thus the Sun beams most of

Figure 3

Over simplified view of Magnoflux tunnels between stars and planets

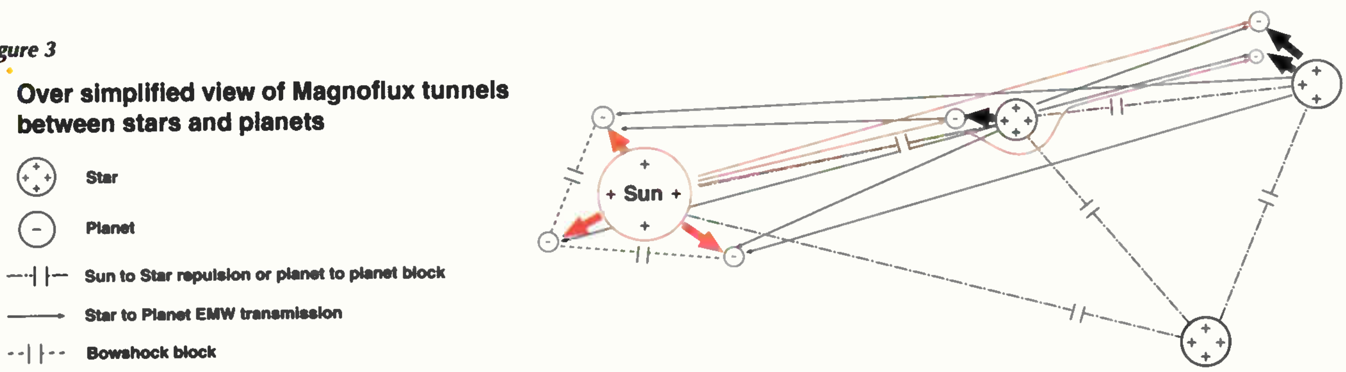
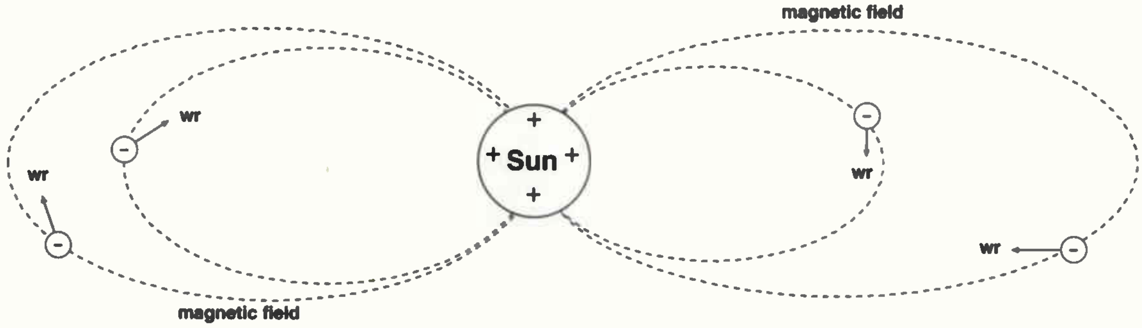


Figure 4



its output directly to the planets and the intervening space is mostly dark. Stars appear as lighthouses, beaming light directly to the planets and to the super black hole planet at the galaxy centre which they rotate.

Consider a complete system (solar system)

All the major planets in the solar system revolve the same way around the sun, and at approximately the same level.

The reason for this is that it is not desirable to cross magnetic flux lines or magnoflux tunnels at any given time, because this will create a voltage and energy disturbance on the Sun's surface. To obtain an overall minimum energy condition the sun must organise its planets into a flat trajectory. If magnoflux paths from various planets do sometimes cross then the excess energy will probably cause sun spots and radio interference.

This same logic can be applied to all other systems in this galaxy and in all other galaxies in the universe.

The dynamic energy balance of sun-earth connection

Historically, in Newtons day when the dimensions of Length, Mass and Time were adopted the scientists believed that matter could not be created or destroyed, but now this has changed to accommodate electricity, light and nuclear energy. Any attempt to try and explain the universe using existing dimensions must be inadequate, as only energy must be conserved and not mass. Plasma energy was the only item present at Gods creation when the Flash plasma temperatures were too high for matter to form and is the essence of all matter but it must be balanced electrically i.e. the positive charged proton matter must be balanced by negatively charged neutron matter or some plasma-energy must flow AND at the speed of light.

Physicists consider the neutron as neutral but in reality to balance mass/charge ratios it must be negative. However, as we live in a negative environment everything we touch is negative neutron matter and is measured from the neutrons negative outside potential. So when we touch the outside of a neutron it is negative but appears neutral. The proton is charged positive but only half of the charge is positive. The negative electron appears to enclose the pair and is connected electrically directly by touching the neutrons outer shell.The E/M ratios for protons and neutrons (negatrons) are well known quantum constants thus MASS and unit volumes of balanced electrical CHARGE are equivalents; they are one and the same thing. NOTE The electrical packing fraction energies appear as quantums of energy to physicists who oddly visualise them as particles instead of packets of electrical binding energy.

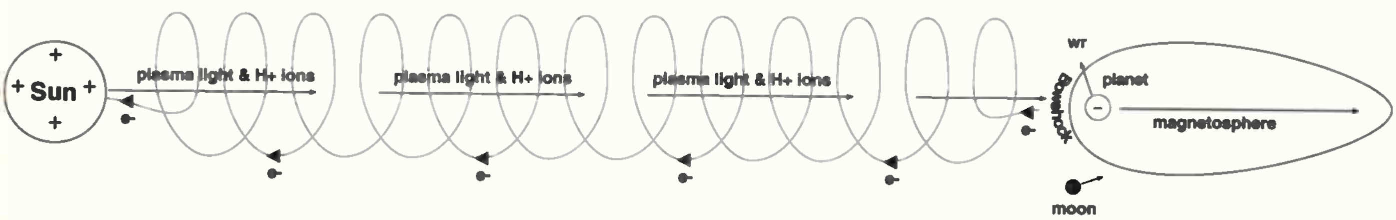
The small difference in weight between the proton/electron and negatron/positron pair is identified by physicists as QUANTUMS of binding energy and electrically produces a positive environment in stars, instead of the negative environment found on planets. This concept requires an EM transmissions of light from stars to planets to balance the charges every time a + ion moves into a planets environment.

This new electrical concept of the universe introduced above will unfortunately require the revision of some of the existing laws of physics. In particular this means that the Stephan Boltzmann fourth power black body temperature law appears to be correct only for incoming absorption of energy but not for full light radiation. Also revision to Plancks constant which dimensions energy to wavelength only; needs to be broadened to include the effects of the electromagnetic B/H volume media curve. The permeability of free space needs to refer to magnetised space only.

The model of the all electric universe sun-earth connection

See **Figure 5**. Planets and our Earth are negatively charged attracters of the + H ions and the balancing plasma energy, which they may combine or convert by nuclear fusion into matter. On arrival the DC hydrogen ions in the solar wind will be swept around to the back of the planet by the planet's magnetic field possibly for water and nucleonic acid formation later. . At the bow shock most of the electrons will be stripped off and returned to the star helixing backwards around the outside of the incoming magnoflux tunnel on their way back to the star/Sun. The AC light plasma EM light will then be released at the stars (Suns) surface to travel inside the sun-planet magnoflux tunnel

Figure 5: The sun-earth connection



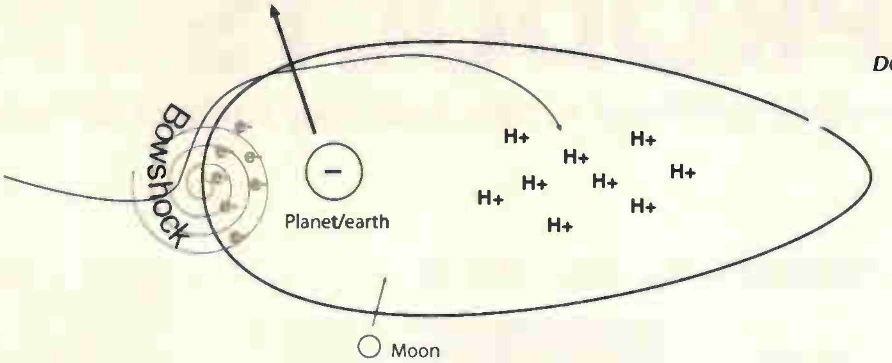


Figure 6: DC view of area near planet

connection to balance the equation. Any movement of the bow-shock will increase or decrease the feed back electrons thus stabilising the Sun Earth connection

DC view of area near the planet

See **Figure 6**. **Moons** appear to be non nuclear and neutral although they will probably be magnetically induced by the planet. They are locked to the planets relative time and are within the planets magnetic field space. They appear to have no active volcanoes so must totally reflect or loose all energy that hits their surfaces. Therefore the mass attraction theory is correct for moons and no additional electrical force would appear to be necessary.

Black holes? Smaller scale black-holes might just be a free planet that have been released from their star when the electrical attraction force fields collapsed after their star exploded. Newtons first law will then take over and the planet will continue in a straight line at its original velocity. However, the planet will of course be attracted towards any and all stars thus zig-zagging its way across space. If it targets one only star then it could be tragic.

Conclusion

When God breathed out the Almighty Flash of Plasma Electricity, He created energy, time, space, and movement but He restricted expansion to volumes of space that containing DC aether or magnoflux tunnels. These are used to stabilize, balance and conserve the whole of the creation and applies to all AC electro-magnetic light transmissions.

Universal space is not empty but full of magnetic aether or magnoflux tunnels, and the more we explore space the more we become aware that everything basic in the universe is moved by electrical attraction through these magnetic fields. The magnoflux tunnel concept opens the possibility, that if a catastrophic event suddenly interrupts the transmission of AC light, then the DC +ion can connects up with the feedback electron and light thus releasing the solar wind contents to combine into any convenient form of matter particle or helium atom that its new relative time reference will permit, such as comic dust or debris.

Similarly, the deeper we delve into the smallest atom, the more we discover that all energy movements are associated with electrical charge attraction immersed within the local time tied within balanced electric fields.

By adopting the electrical concept the following advantages over the mass attraction theory of modern physics are evident.

- It explains why the universe is not collapsing inwards due to all the matter being attracted together. (Like charges repel) We do not have to look for more missing matter or dark energy any more nor invent quint-essential theories.
- It explains why the solar system and galaxies rotate around a central point within the same plane. (To avoid crossing magnetic magnoflux lines)
- It explains how and why the solar wind works (DC Hydrogen +ions in solar wind are emitted from a positive star surface to a negatively attractive target like a planet and a DC magnoflux tunnel aether connection is formed by the released return electrons)
- It explains how the nuclear fissioned AC neutron plasma can pulsate through the magnetic DC aether in magnoflux tunnels to allow the balancing negative light to travel from stars surface to planets (Plus stabilising the star-planet connection with an electrical feedback.)
- It explains why suns and stars loose weight over time and conversely why the planets increase in size and weight (Geological evidence indicates the planet earth increases in girth.)
- It explains why some huge planets emit radio signals. (Due to intense nuclear fusion active within their cores.)
- It explains small black holes, which are only free wheeling planets that have lost their centrifugal attraction force when their star collapsed. (Super massive black-hole negatives have been found at the centre of galaxies surrounded by their bonded rotating stars).

However, if God suddenly stopped time, then the universe would collapse as the electrical charges would cancel each other out and in a flash there would be nothing at all left. We live in the God granted time gap between when there was nothing and when there will be nothing again.

NEW PRODUCTS

Please quote *Electronics World* when seeking further information

Singlemode fibre optic connectors

A singlemode version of the LightCrimp Plus range of fibre-optic connectors has been designed to complement the established multimode connectors.

Featuring a new fibre splice engine technology that allows for precision alignment of cabled fibre to the factory-polished stubbed ferrule, they enable installers to obtain low insertion loss and low reflectance.

The performance of the connectors exceeds the requirements of the TIA/EIA 568 B.3 and ISO/IEC 11801 insertion-loss and return-loss specifications.

The LightCrimp Plus family of connectors is ideal for direct termination applications at patch panels and wall outlets, and for either backbone or horizontal cabling.

Like the multimode LightCrimp Plus connectors, the singlemode design uses simple, reliable, and proven Tyco Electronics splicing technology.

The connectors feature a metal body which delivers robust and stable performance over time and can withstand temperature fluctuations over the range -10°C to +60°C, meeting the relevant TIA/EIA, IEC, CECC and EN standards.

The same toolkit and procedures are used for the singlemode product as for the multimode LightCrimp Plus connectors.

A complete installation Guide on CD-ROM is provided with every toolkit.

Amp NetConnect Group
www.ampnetconnect.com

DMOS full-bridge motor driver IC

A new device, the A3949 DMOS full-bridge motor driver integrated circuit designed for the PWM (pulse-width-modulated) control of DC motors is capable of peak output currents of up to ±2.8A and operating voltages of up to 36V, with low 'on' resistance outputs: typically 0.68 ohm for the source driver and 0.576 ohm for the sink driver.

'Phase' and 'enable' input terminals are provided for controlling the speed and direction of the motor with externally applied PWM control signals.

A 'sleep' mode is provided to minimise power consumption when the A3949 is not in use and a braking function is implemented by driving the device in slow decay mode via the 'mode' pin and applying an 'enable chop' command. Because it is possible to drive current in both directions through the DMOS switches, this configuration effectively shorts out the motor-generated back EMF.

Internal circuit protection includes thermal shutdown with hysteresis, undervoltage monitoring of load supply and charge-pump voltages, and crossover current protection.

Supplied in a choice of two power packages, a 16-lead plastic SOIC with a copper batwing tab (suffix LB), and a low-profile (1.1mm) 16-lead TSSOP (suffix LP) with exposed power tab. Operating temperature range is -20°C to +85°C.

Allegro MicroSystems Europe Ltd
www.allegromicro.com

COG, the component obsolescence group has published the fourth booklet in the 'Minefield' series. The other titles are The Obsolescence Minefield; The Date Coding Minefield; The Supply Chain Minefield. Costing £5 each they are available from COG telephone 01582 762934 or online at www.cog.org.uk. COG also run workshops.

Made to measure enclosures

PS Analytical produce instruments for automating the laboratory. Recently they extended their applications to detecting mercury emissions from coal fired utility stacks and mercury in the petrochemical industry.

Using a combination of Rittal enclosures, the analysers were built into three Rittal Vario Module enclosures that could be linked together and used on a desk or built into a Rittal floor standing PC enclosure.

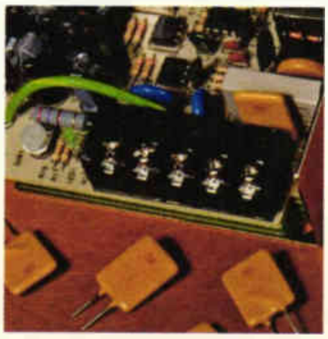
A Top Therm cooling unit was installed because the units could be used in countries where the ambient temperature is higher that the maximum operating temperature of the equipment therefore refrigerated air must be used. Analysers built into a PC enclosure make a portable unit.

Despite the enclosure being a special build, Rittal was able to deliver the equipment ready to be installed and on schedule.

Rittal Ltd
www.rittal.co.uk

LVR line voltage protection device

Raychem Circuit Protection have introduced a new device that increases the maximum hold current rating from 400mA to 550mA and helps protect devices from damage caused by overcurrent and overtemperature fault conditions. The LVR series now covers applications up to 120 Watts at 220VAC and 60 Watts at 120VAC. Designed for in line voltage applications the polymeric positive temperature coefficient devices are rated at 240VAC, permitting maximum voltages of up to 265VAC. The thermally active devices help protect against faults on the primary side of power supplies and transformers. When installed



on the primary side of the circuit, in proximity to potential heat-generating components such as magnets, FETs or power resistors, the LVR device can help provide undercurrent and overtemperature protection with a single installed component. Available in straight-lead or kinked-lead configurations the devices can be supplied in tape-and-reel packaging for compatibility with high-volume assembly.

Raychem Circuit Protection
www.circuitprotection.com/lvr

Please quote *Electronics World* when seeking further information

Comprehensive UMTS R5 protocol test package

Tektronix, provider of test, measurement and monitoring solutions for mobile equipment manufacturers and operators, has announced the addition of Universal Mobile Telecommunications System Release 5 (UMTS R5).

UMTS R5 is the latest step in the migration towards full implementation of third-generation (3G) UMTS networks. In order to develop network elements that comply with the standard, mobile equipment manufacturers are challenged to test the compliance of their existing products against the newest specifications. The Tektronix K1297-G20 protocol tester Version 2.60 (V2.60) delivers compliance testing of UMTS R5 in the industry's leading protocol test platform with the ability to address more than 1000 protocols.

The K1297-G20 is a full-featured, compact protocol tester that can be configured for a wide range of protocols. It supports multi-channel and multi-port protocol testing and incorporates a broad selection of conformance test suites. The K1297-G20 is set apart by its ability to simultaneously monitor and enable interoperability of GSM, GPRS and UMTS network elements. The existing GPRS-related features of the K1297-20 have made Tektronix a leading supplier of GPRS signalling protocol test equipment.

Tektronix
www.tektronix.com

Pocket-sized PC-based test instruments

A range of low-cost pocket-sized PC-based test instruments which combine high-speed sampling with deep data buffers are now available from Computer Solutions.

The range includes the DSO-2102 digital oscilloscope, the DSO-2102M which also includes a spectrum analyser, and the LA-2124 logic analyser. The instruments connect to a desktop or laptop PC via the parallel port (with an additional USB option for the DSO-2102) and are supplied with software compatible with most versions of Windows including NT and XP.

High-impedance (100 kilohm) probes minimise interference with the circuit under test, and both state and timing data can be captured simultaneously via a single probe. Easy-to-use software allows data to be displayed as timing waveforms or in the form of state-list displays. A search function is included as standard, and I2C monitor software is available as an option. There is an optional battery pack.

These handy devices cost from £420.00 (¤630.00) and are available for next day delivery.

Computer Solutions Limited
www.computer-solutions.co.uk

Appointment

Nigel Power has been appointed managing director of Cooper B-Line Ltd, a leading manufacturer of enclosures, support systems and spring steel fasteners for the electronic, electrical, mechanical and telecom/ networking industries.

Power has set about developing the business in the UK, France, Germany, Switzerland, Russia and the Middle East expanding Cooper B-Line's business with OEMs, end users, installers and distributors. Cooper Industries is a worldwide organisation trading as CBE on the New York Stock Exchange.

"My aim is to enhance profitability which will be achieved through intelligent assessment of market information and a combination of team work with employees and partners." Power said. "We will further enhance our business profile by offering our customers a comprehensive source of supply through the Cooper Industries Group of Companies".

Hall-effect current sensor offers higher bandwidth and enhanced accuracy

Allegro has introduced ACS752a new member of the ACS75x family of current sensors.

The new ACS752 boasts a bandwidth of 50kHz, compared with 13kHz for the original ACS750, and an accuracy of 7.5% compared with 13%. It is available in 50 and 100 A versions. The device runs off a single +5V supply, provides 3kV isolation, and has a resistance of only 130 micro-ohms.

The package consists of two primary leads that pass the current through a concentrator core and past the sensing element of a Hall-effect integrated circuit. It requires minimal effort to implement on a PC board in a high-volume production environment and may be soldered or welded. The volume of the package is less than one-eighth that of comparable competing current sensors.

Suited to applications where isolation, low power dissipation, and low voltage drop are an advantage, typical uses are in motor control, power-supply load management, power distribution or overcurrent fault protection.

Allegro MicroSystems Europe Ltd
www.allegromicro.com

D-Sub connector is sealed to IP 67

Harting has extended its D-Subminiature connector range with a new series designed for use in applications requiring water resistance to the rigorous requirements of the IP 67 standard.

The main usage is in equipment situated in harsh environments, where connectors need to withstand water, other fluids or dust. Typical applications can be found in industrial machinery, process control, mobile equipment and instrumentation.

The D-Sub range provides complete solutions for both PCB and cable mount applications. For cable applications, Harting provides connectors and hoods, while for PCB applications connectors with or without rear frames for water-resistant panel mounting purposes are available. Most of the connectors in the range are available with from nine to 50 poles.

Harting
www.harting.co.uk

Please quote *Electronics World* when seeking further information

Catalogue of safety & health products

To help improve working environments, Nederman has just produced a new 116 page guide to its comprehensive range of safety and health products.

A range of high vacuum systems is offered for the removal of fume and dust problems, while exhaust extraction units are used for the removal of exhaust fumes from stationery and moving vehicles.

For effective workplace screening, Nederman panels are supplied in modular form and will absorb noise and are both flexible and load bearing. For personal protection, welding helmets incorporate auto darkening and fresh air supply.

Nederman
www.nederman.com

Small components coated in drum machine

The coating of small electronic components, up to 150mm in length, has until now proved to be difficult to achieve without major investment.

Now at a fraction of the cost of conventional coating methods it can be done with the ISPC Walther Trowal Rotamat Coating Drum.

The Rotamat range comprises three models from the entry level R-55 with a capacity of eight litres to the 50 litre capacity R-90. Each is available with a range of options and can be up and running in minutes with minimal operator training.

International Surface Preparation
www.surface preparation.com

PCB test points – the easy way to get hooked up

If you are a design or development engineer working on prototype or small quantity printed circuit board production, you will know how difficult it is to make quick and reliable connections to test points on a printed circuit board. But now, the innovative PCB Test Points from William Hughes provide a neat and inexpensive way to completely overcome this problem.

Featuring a large loop for easy probe attachment and bowed legs for quick insertion into the board without damage to a through plated hole, the phosphor bronze spring strip is solder tinned to ensure good electrical contact. A special glass bead gives a high resistance to thermal shock and a choice of eight standard colours enables individual test points to be quickly identified when installed on the board.

Popular as terminals too, the test points are also ideal as mounts for components that may need scheduled replacement, or for elevating heat-generating elements above the PCB surface.

William Hughes PCB Test Points are available in three popular sizes. The Large Type 100 (for 1/16 inch board thickness and 1.32 mm hole size) has an internal loop diameter of 2.1 mm and leg length below the glass bead of 2.8 mm. The Small Type 200 is also for 1/16 inch boards but is designed to suit a 1.0 mm dia. hole. For customers using the thicker 3/32 inch boards, the long legged Type 300 with its 3.6 mm legs and 2.1 mm loop provides the perfect solution.

Test points can be shipped in large or small quantities, bagged as required, for both prototype and volume production. For applications that require a test point style outside the standard range, there is a choice of special colours and a custom design service is also available.

William Hughes Ltd
www.wmhughes.co.uk

Low cost USB analyser

USB is becoming more and more popular as a way of interfacing embedded systems to PCs. However, its protocols are complex, and an analyser greatly simplifies debugging both the hardware and the application. To date, USB analysers have cost around £4000, but the new USB Tracker 110 sells for little more than £500: a price level that makes it feasible to purchase one for every engineer working on the project.

It is the ideal companion for anyone developing USB peripherals, embedded software or drivers. Its analysis and display software allows the user to quickly learn all about USB. It can be connected between a personal computer or a laptop and any USB peripheral to provide an instant record of the traffic.

The USB Tracker 110 can carry out USB 'low-speed' (1.5Mbit/s), 'full-speed' (12Mbit/s) or mixed-speed capture, automatically detecting the speed. Unlimited capture time is provided through a high-speed USB 2.0 connection to the host PC, and power is taken from the USB.

Computer Solutions
www.computer-solutions.co.uk

Energy saving anti-pollution IC for digital systems

The new STR-T4000, developed by Sanken Electric and available in Europe from Allegro MicroSystems Europe, is a partial resonant convertor integrated circuit optimised for use in power supplies for IT-related equipment and flat panel displays.

The new device meets the latest energy-saving and anti-pollution requirements of microprocessor controlled digital systems operating in the 1.5V supply voltage range. It uses a new circuit topology which allows it to handle anything from zero load condition in standby mode to 300W of output power as encountered in IT-related equipment and flat-panel TV applications.

The power control IC automatically adjusts its operating mode to the load conditions, and features automatic dead-time setting.

Together, the STR-T4000 and the multi-path transformer form the basis of a 300W switch-mode power supply that provides the benefits of low standby power, system downsizing, a minimal component count, high efficiency, low noise and a multi-output capability. Typical improvements include a cost reduction of more than 30%, a 30% smaller 'footprint', and a 20% reduction in power loss.

Allegro MicroSystems Europe Ltd
www.allegromicro.com

Automatic heterodyne (notch) filter using a switched capacitor filter

Short-wave listeners and radio amateurs are only too accustomed to the many birdies, spurii etc to be found on the short wave bands. Modern high-grade equipment now comes with DSP notch filters that can eliminate up to six separate heterodyne whistles simultaneously and automatically. Michael Slifkin and Michael Rotenfeld discuss a practical example

In the early days of radio, the only recourse to remove such heterodynes was the manually tuned notch filter placed in the audio line and tuning from about 30 to 3000Hz. In the early 1970s an automatic notch filter was described in the literature using of course analogue techniques[1]. This worked using the principle that passing a signal through a filter will cause a phase shift unless the signal frequency is the same as that of the filter. These early circuits monitored the phase of the incoming and outgoing signals through the filter using a phase sensitive detector (see the article *in Electronics World* April 1999 for a description of the phase sensitive detector) and if the phases were not equal they would send an error signal to a variable resistor (i.e. an FET transistor configured to act as a variable resistance and placed in the tuned circuit of the notch filter) to bring the filter in tune with the signal.

Nowadays with the introduction of the switched capacity filter which comes as a integrated circuit, it is possible to build an automatic notch filter with few parts and no need for adjustments to circuit parameters. We give a simple explanation of how this filter works.

Shown in **Figure 1** is a simple active integrator or low pass filter. This can be used as the basic building block of any type of filter by a suitable combination of several of these units. By varying the resistor we can alter the characteristic slope of the filter. However using a variable resistance is not convenient as often, unrealistic values are required. We can simulate the resistor by using a switched capacitor as shown in **Figure 2**. The resistance of the switched capacitance will vary with the frequency, the higher the frequency the apparent lower the resistance. It can be shown that the simulated resistance is given by:

$R = 1/C_1 f_{clk}$, where f_{clk} is the clock frequency

This gives a very versatile and simple way of tuning the filters. Commercial switched capacitors are available in a number of forms containing two to eight subsections that can be cascaded for sharper characteristics. Switched capacitor filters are available in which the clock frequency bears a fixed ratio to

Figure 1: active RD inverting integrator.

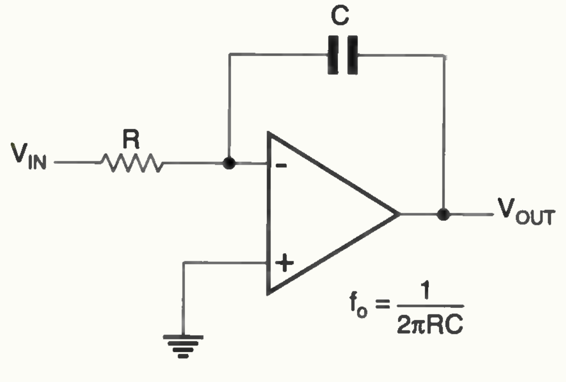
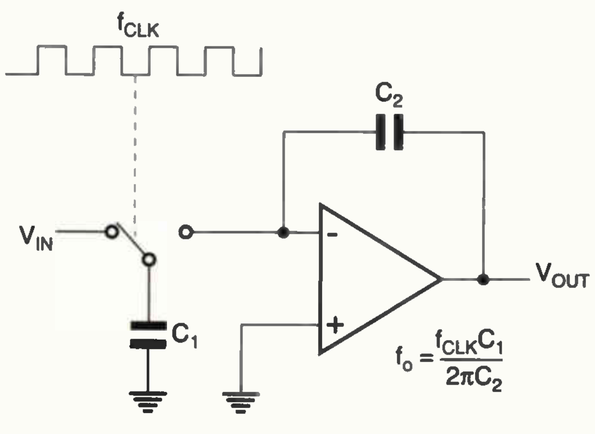


Figure 2: Inverting switched capacitor integrator.



the filter centre frequency usually 50 or 100 times the frequency. This means that we can build a versatile notch or peak filter using only four chips. The schematic is shown in **Figure 3**.

The signal plus the heterodyne enters the comparator that is set at such a level that the audio signal is rejected but the much stronger heterodyne is passed through. This enters the frequency synthesiser that consists of a phase lock loop chip (PLL) together with a divide by 100 counter in the feedback loop. Lock is obtained when the error signal is exactly 100 times the frequency of the input heterodyne. This signal is now used as the clock frequency of the switched capacitance filter which is configured as a notch filter and to have a peak frequency 1/100th that of the clock frequency. The signal plus the heterodyne is now fed to the filter and the heterodyne is rejected.

We have looked at two different switched capacitor filters; a simple device which has become an industry standard the LMF100 and a more sophisticated device the LTC 1068. These can be configured as low pass, band pass and high pass filters as well as notch filters. They can also be operated with a centre frequency at either 1/50th or 1/100th the clock frequency.

The LMF100 is a dual second order switched capacitor filter. The two sections can be cascaded to give fourth order i.e. sharper characteristics. The manufacturers schematic diagram of one section is given in **Figure 4**. The filter is particularly easy to use as only three resistors, R_1, R_2 and R_3, set three parameters of the filter. The ratio R_3/R_2 gives the Q of the filter. The ratio R_3/R_1 gives the gain at the maximum and the ratio R_2 to R_1 give the gain as the frequency goes to zero.

Our test circuit is shown schematically in **Figure 5** and the circuit in **Figure 6**. We used two signal generators to produce the signal and heterodyne. The first operational amplifier (the 741) on the left hand side of the diagram is a summing amplifier. If you wish to use these circuits in an audio line then this first stage should be removed. The second stage the IC LM311 is the comparator. As we are using signal generators with fixed signal voltages we have fixed the comparator level. If used as a stand-alone filter one would need to find the best level by trial and error. In which case one should replace the 5.1kΩ resistor with a 10KΩ potentiometer. It is important to use good wiring practice, as our

Figure 3: schematic of the automatic filter.

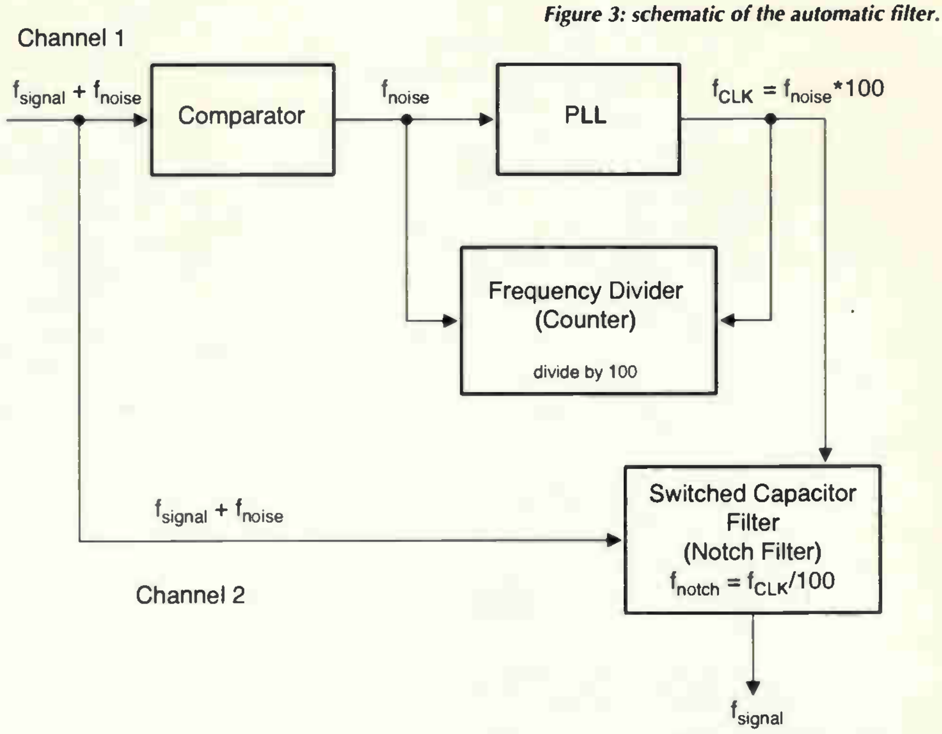


Figure 4: LMF100 schematic.

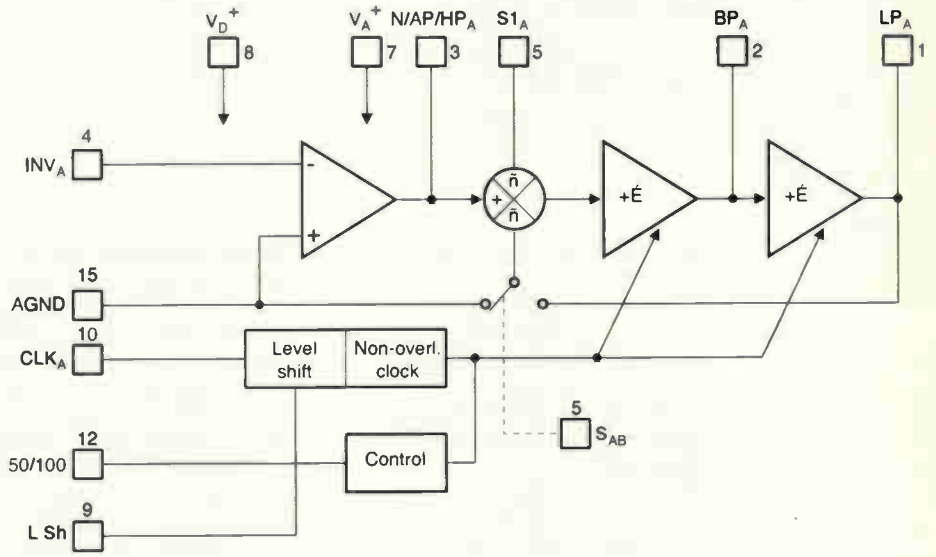
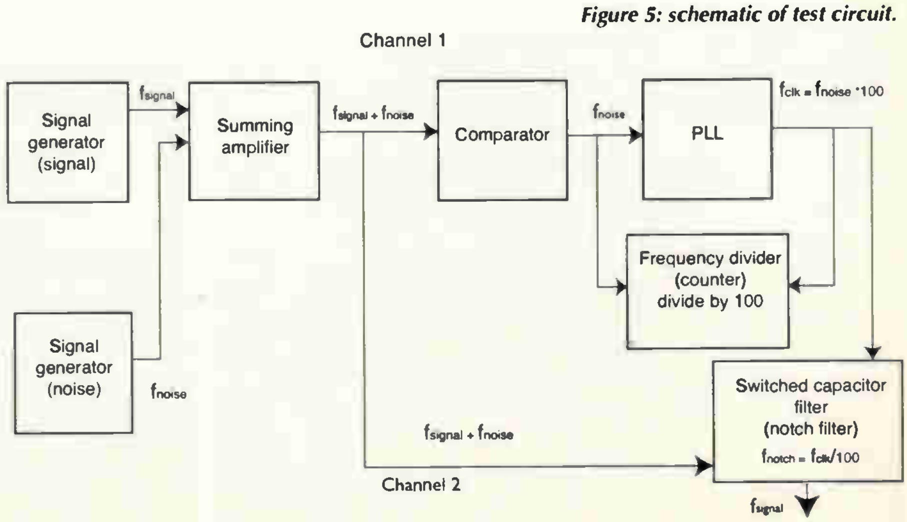


Figure 5: schematic of test circuit.



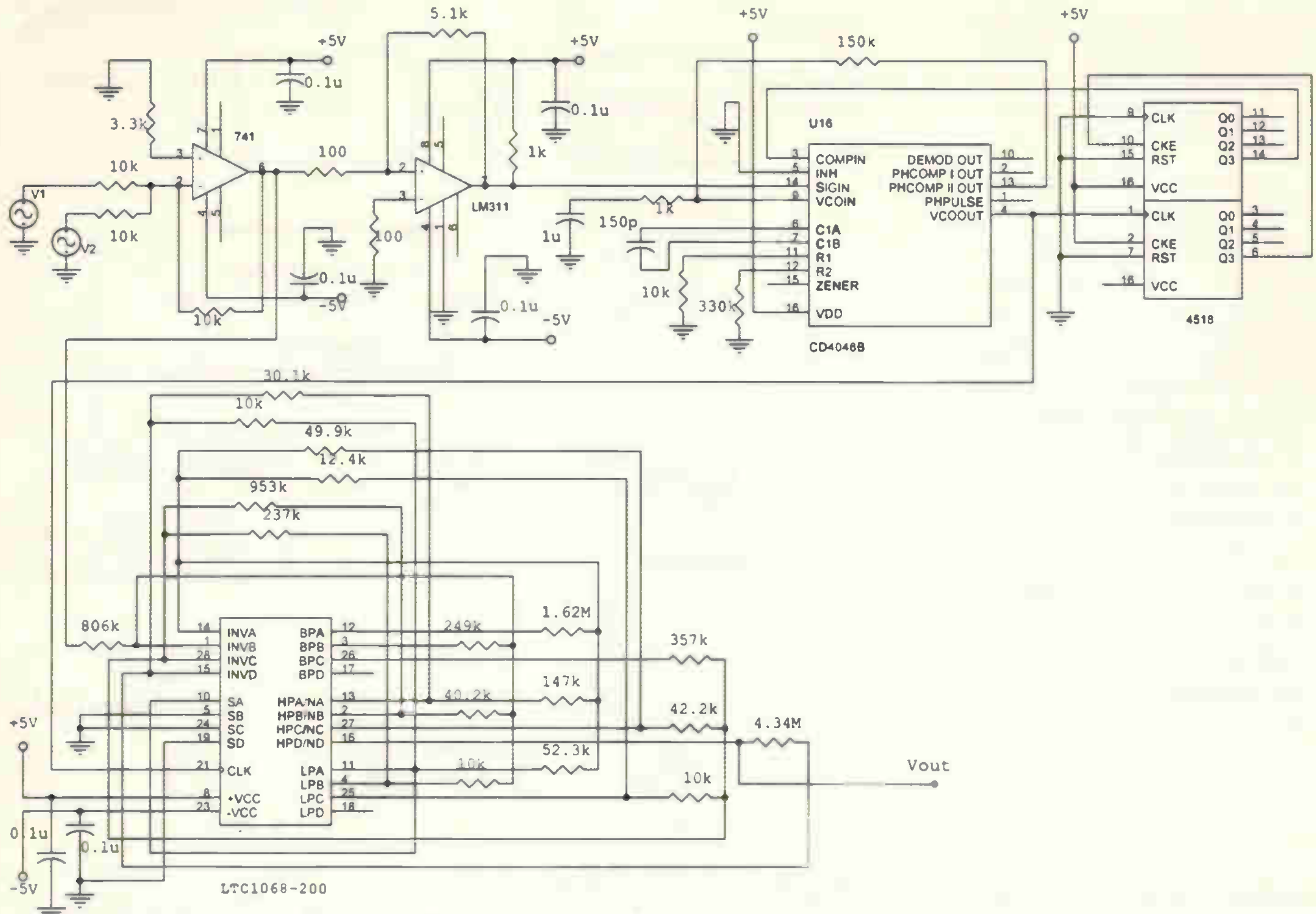


Figure 6

highest frequency is a 4 megahertz square wave.

We built our first filter using the LMF100 on a 'Superstrip' breadboard. This has two 2nd order filter sections on board. The datasheet can be downloaded from

http://www.national.com/ds/LM/LMF100.pdf.

The results were not quite as good as we expected. The best notch we could obtain using a second order section, which yielded a good stable noise free performance over the range 300 to 3kHz gave a notch depth of about -22 dB with R1 = 10kΩ, R2 = 10kΩ and R3 = 68kΩ Almost certainly one would do better by hard wiring the circuit rather than bread boarding it. In addition, one could cascade the second section to give a fourth order filter and even cascade more chips for higher orders.

However, as we were more interested in the better specified LTC1068 from Linear Technology our main effort was put into using this as the heart of our automatic notch filter. This device has four identical 2nd order sections in the one packet. Technical details for this filter are available at

http://www.linear.com/prod/datasheet.html?datasheet=104

This filter also comes with a design program called Filter CAD, so that we could do the theoretical design and compare it with the actual results. This program can be

Figure 7: screen display for input to Filter Cad.

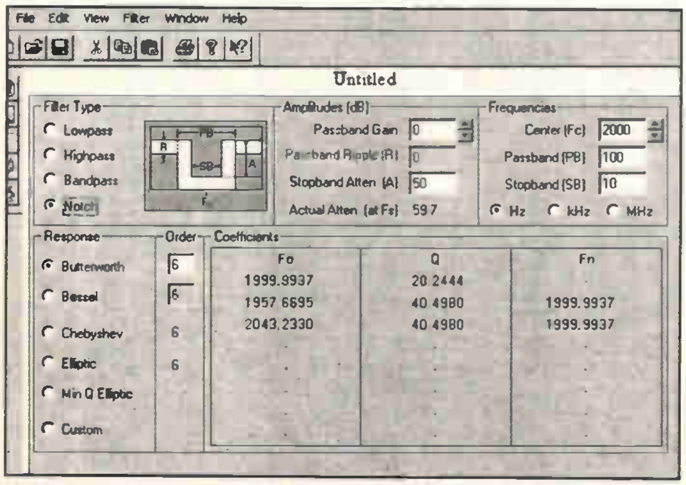
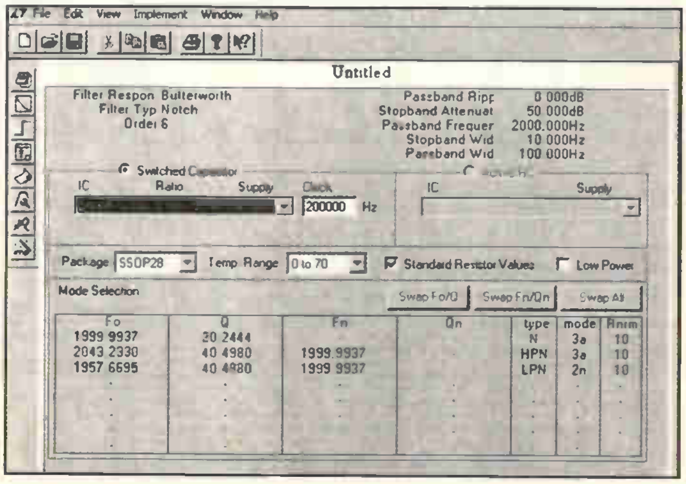


Figure 8: results screen from Filter Cad.



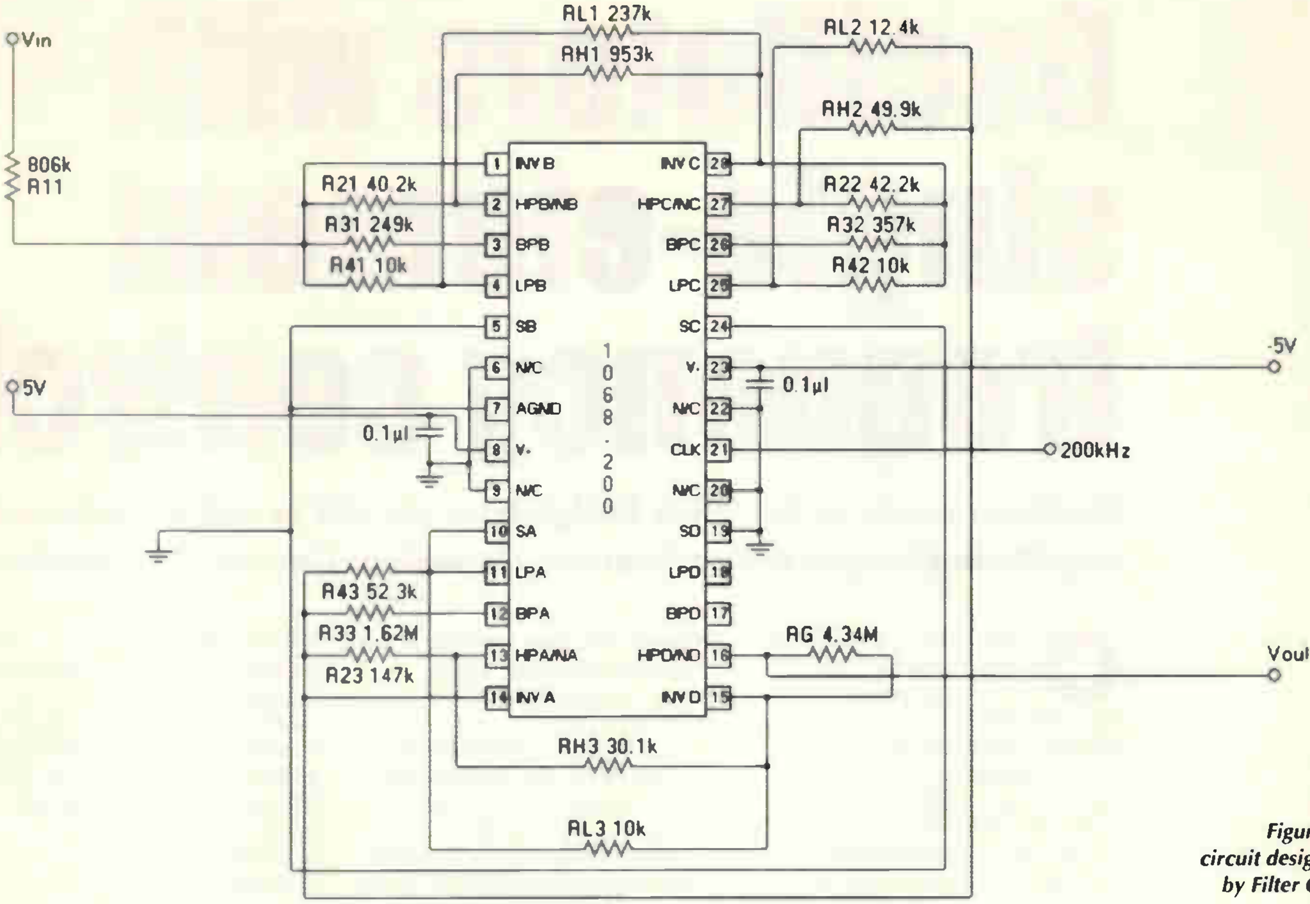


Figure 9:* *circuit designed by Filter Cad.

downloaded from http://www.linear.com/prod/prod_home.html?product_family=filter

Because of the high accuracy of this design program we have used 1% tolerance resistors as shown in the circuit diagram. **Figure 7** shows the computer screen display of this filter program.

The display is self-evident. One writes in the specification on the **Figure 8** shows the results screen and is also self-explanatory.

The program gives the design of he circuit as in **Figure 9**. Note that the values are given to a very high tolerance. The program then proceeds to give the predicted performance of the filter as in **Figure 10**.

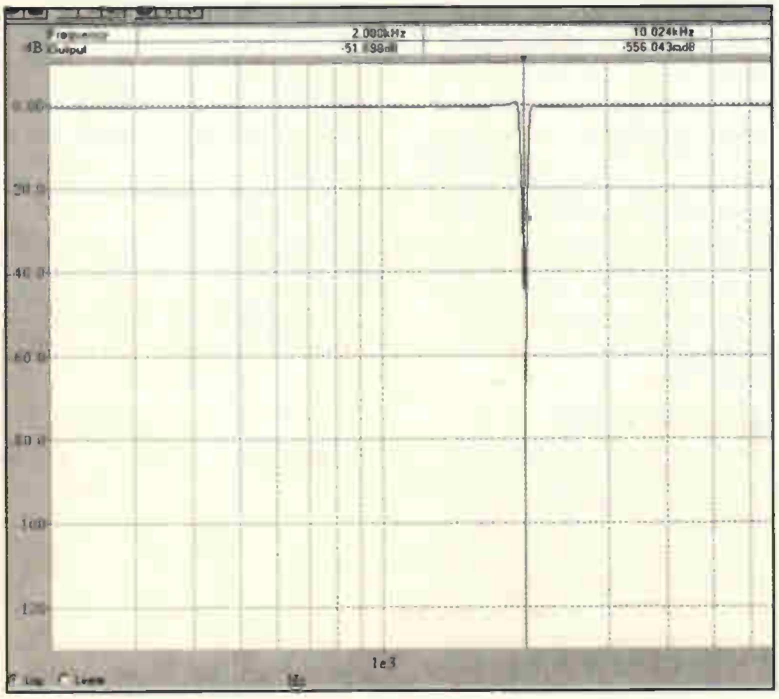
The circuit was built on a prototyping board. The actual circuit **Figure 6** differs very slightly from the designed circuit, as we needed to replace the resistors by the nearest ones in the 1% tolerance range. The measured gain at 2kHz was -48dB compared to the design figure of -50dB. The bandwidth was 107Hz as compared to the design figure of 100Hz. All in all, it was a very satisfactory agreement.

One could of course make a more versatile filter. One can switch to an external clock to give a manually controlled frequency. One can switch the filter to give a peak for CW reception.

In conclusion the use of the switched capacitor filter enables us to build a high Q notch filter using only four ICs and a minimal amount of wiring which tunes automatically to the interfering heterodyne.

Reference:

1: G.J. Deboo and R.C. Hedlund EDN/EE, 38-41 (1972)



(Left) Figure 10:* *theoretical performance of filter as given by Filter Cad.

Oscillators with single-element frequency control

Oscillators such as the Wien bridge type are apt to suffer from erratic amplitude changes as the frequency is varied. George Short explains

One potent cause of amplitude bounce is ganging errors in the two-gang variable resistance, which controls the frequency. (**Figure 1a**, R_1 R_2). The effect is often made worse by the addition of an amplitude-stabilisation circuit (thermistor, lamp, etc.), which is itself liable to provoke amplitude hunting. When the network is used to tune a selective amplifier the ganging errors translate into uneven performance over the tuning band.

The use of a single variable element for tuning avoids the ganging error but introduces a new one. Changing any one element in a Wien or comparable network changes the attenuation at the tuned frequency. Thus one source of amplitude variation is exchanged for another.

Wigan's modification

If some means could be devised of effecting tuning by a single element while keeping the attenuation constant the problem would be solved. An ingenious solution was proposed some forty years ago in the pages of this magazine's long-defunct sister journal, Electronic Technology (1).

Ernest Wigan's modified Wien network (**Figure 1b**) maintains the attenuation constant (at 3 times when $C_1 = C_2$, $R_1 = R_2 + R_3$). Here the resistance in the parallel arm is split to form a potential divider, R_2, R_3. A variable resistance P adjusts the frequency. If the ratio R_2/R_3 is correct the attenuation remains constant as P is varied.

Wigan's circuit is very effective. It has, however, two drawbacks. One is that the tapping point of R_2, R_3, which is critical, is affected by the tolerances of the fixed resistors and capacitors. Hence, in a range-switched circuit provision must be made for timing R_2/R_3 on each range. This drawback is shored by other similar circuts. The other, which is not, is that the amount of frequency sweep, which can be realised is too restricted. In the usual case of equal Cs changing P from zero to infinity gives a ratio fmax/fmin of only 2.6. In practical multi-range oscillator one wants a frequency ratio of at least $\sqrt{10}$, and preferably 10 to keep low the requirement for numbers of scales, switch contacts and capacitors.

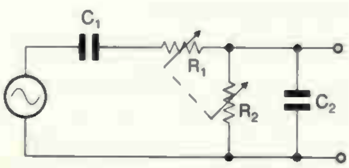


Figure1 (a) The attenuation of standard Wien arms is affected by ganging errors in R_1, R_2.

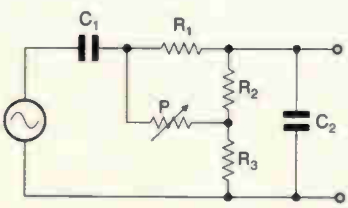


Figure1 (b) Wigan's modified Wien has a single variable, P. Attenuation is kept constant by correct proportioning of R_2/R_3.

Alternative strategy

The network shown in **Figure 2a** makes practicable ratios of 10 while preserving constant attenuation. Inspection reveals that when P = 0 the circuit turns into **Figure 2b**. This consists of the standard Wien reactive arms plus A and B in parallel in series with the output. Provided that the load is infinite A and B then have no effect. The zero-phase-shift frequency is the usual $1/(2\pi RC)$ and the attenuation factor is 3. As P increases from zero A and B come into play and introduce additional attenuation. This can exactly offset a reduction in attenuation, which results from the parallel arm increasing in impedance as P is increased.

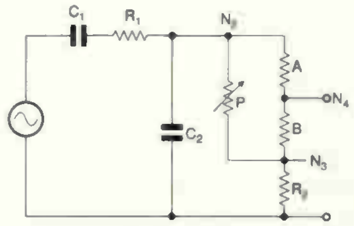


Figure 2a: alternative modified Wien, giving a wider tuning range.

Consideration of the circuit shows that as the frequency is reduced towards zero the resistance R1 becomes negligible compared with the reactance of C_1. In the parallel arm, if P is infinite and A and B very high only the reactance of C_2 is significant. Hence at very low frequencies the circuit approximates to 2(c). The attenuation factor due to $C_1 = C_2$ is 2. To restore the original value of 3 the resistive arm must

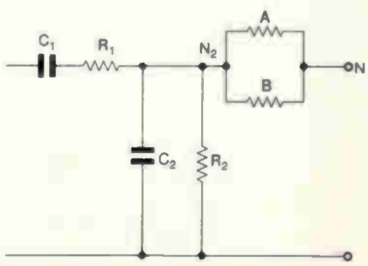


Figure 2b: effective network when P = 0

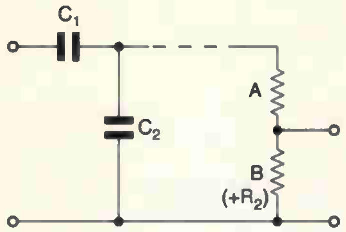


Figure 2c: effective network at very low frequency.

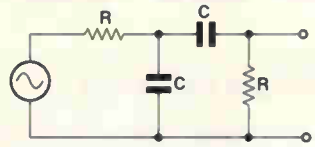


Figure 3a: Cascade RC,CR network has a response like the Wien.

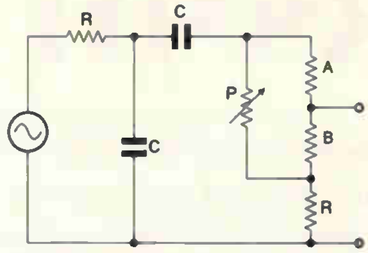


Figure 3b: modification for single-resistance tuning.

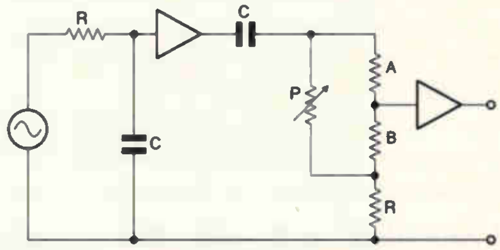


Figure 3c: split network with buffers.

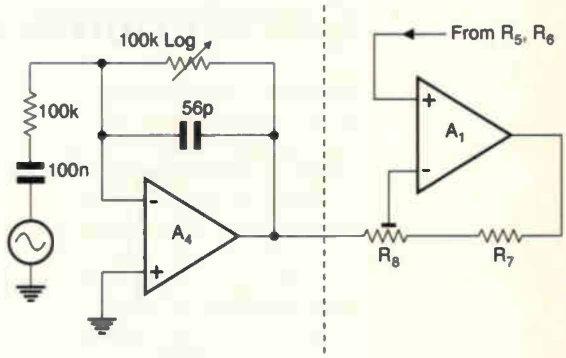


Figure 5: adding an input buffer/attenuator turns the circuit into a selective amplifier. For maximum Q, R8 is set for 'just not oscillating'. In the oscillating state application of a near-synchronous signal can lock the frequency

introduce an additional attenuation by 3/2. Since R_2 is negligible compared with A and B the requirement is that B = 2A. Analysis shows that when A, B and P are finite the attenuation remains constant so long as B = 2A.

In this extreme form the circuit has no frequency-selective properties. It is merely a capacitive potential divider followed by a resistive divider. In the practical case with finite resistances there is always some selectivity though this diminishes as P becomes large with respect to R_2. Experiment reveals that with careful setting up good sine waves can be obtained when the effective resistance of P, A and B combined is as high as $100R_2$. The frequency ratio is then $\sqrt{101}$.

Other zero-phase-shift networks

The same technique can be used to modify the cascade z.p.f. network of **Figure 3a** for single-element tuning. The result is **Figure 3b**, where B = 2A as with the Wien network. The cascade z.p.f. network can be separated into an RC arm followed by a CR arm, buffered from one another (**Figure 2c**). The constant-attenuation condition is now A = B.

Eliminating the A, B divider

In **Figure 2a** the resistive network P, A, B, R_2 is merely a special form of potential divider. It operates on V_2 in such a way that its output V_4 is always V1/3, whatever the value of P. It follows that any circuit arrangement which does the same thing can be substituted. To sustain oscillation V4 needs to be amplified so that $3V_4 = V_1$.

By inspection, the A, B divider reduces the voltage across P by a factor of (A + B)/B. The output V_4 is this reduced voltage plus V_3. That is:

$V_4 = (V_2 - V_3)\,[B/(A + B)] + V_3$

When B = 2A this yields:

$3V_4 = 2V_2 + V_3$

Evidently $3V_4$ can be computed from V_2 and V_3 by an operational circuit. One realisation of this is shown in **Figure 4**. Here A2 has an output of $2V_2$ and A_3 of V_3. The combining network R_5, R_6 reduces the output to 3V4/2 so to sustain oscillation A1 must have a gain of 2. Note that the input capacitance of A_2 can now be absorbed in C_2.

As P becomes much larger than R_2, V3 sinks towards zero leaving A_3 with a vanishing input. A_2 then provides the required gain of 2. When P = 0 the normal Wien network is restored and its loss is made up for by the summed gains of $A_2 + A_3 = 3$. As P is increased from zero to maximum the normal Wien frequency is reduced by a factor of v(1 + m) where m = Pmax /R_2.

The circuit can be set up initially with P = 0 and R_4, R_8 halfway. Adjust R_8 for onset of oscillation. Set P to maximum and set R_4 for the same level as before. Repeat the procedure to get constant output at low distortion.

In this form the amplitude is stabilised by a tiny amount of peak clipping. For reduced distortion a bead thermistor could be substituted for R_7 plus the top section of R_8 with appropriate scaling of the lower section. For selective amplification or synchronisation a signal can be injected via an input buffer A_4 (**Figure 5**). With components of the nominal values shown the measured frequency range was 68-693Hz.

Design considerations

The amplifiers should be f.e.t. - input types. The gain-setting potentiometers should be capable of fine adjustment. If range switching is required the capacitors should be very accurately matched or a separate equalising adjuster (for A/B or R4 in **Figure 4**) provided on each range.

Reference: (1) E.R.Wigan, Single Control Element Wien Bridge, Electronic Technology, June,1960, vol. 37, no.6, pp223-231

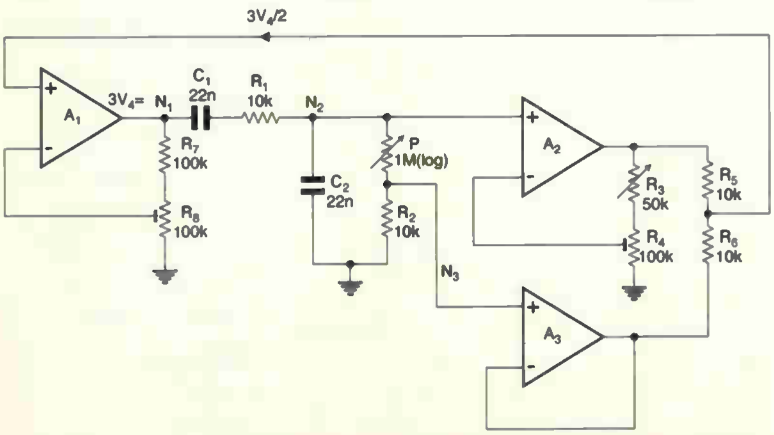


Fig.4 Here A2 and A3 perform the same function as the A, 8 divider of Fig.2 (a)

CIRCUIT IDEAS

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

The best circuit ideas are ones that save time or money, or stimulate the thought process.

This includes the odd solution looking for a problem – provided it has a degree of ingenuity.

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Clear hand-written notes on paper are a minimum requirement: but you will stand a much better chance of early publication if you supply your idea electronically, preferably by email. Any diagrams need to be in a graphic format, not a CAD/CAM or any other obscure file.

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Where software or files are available from us, please email Caroline Fisher with the circuit idea name as the subject.

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Circuit mimics battery discharge

In a battery-powered system, two identical removable lithium rechargeable batteries (11V) are used to meet required operating hours. At any time, only one battery will be used. Once the voltage of that battery drops to below 8.3V, a low voltage protection circuit inside of the battery pack will cut off the battery connection to prevent the lithium battery from over discharging. Meanwhile, the other battery will quickly connect to the system such that the operation will not be interrupted. The battery switching process takes around 200μs.

Two computer controlled bench power suppliers are used on the production line to simulate the lithium batteries. Unfortunately, turning off the power of a power supply does not necessary generate a sharp voltage drop on its DC output due to a fairly large output filter capacitance. Depending on the size of the output capacitor in the power supply and the load conditions, the output voltage fading time can easily exceed several hundred ms. With this 'soft' turn off characteristic, the system under test does not work properly since it requires a sharp power cut off to switch to another battery.

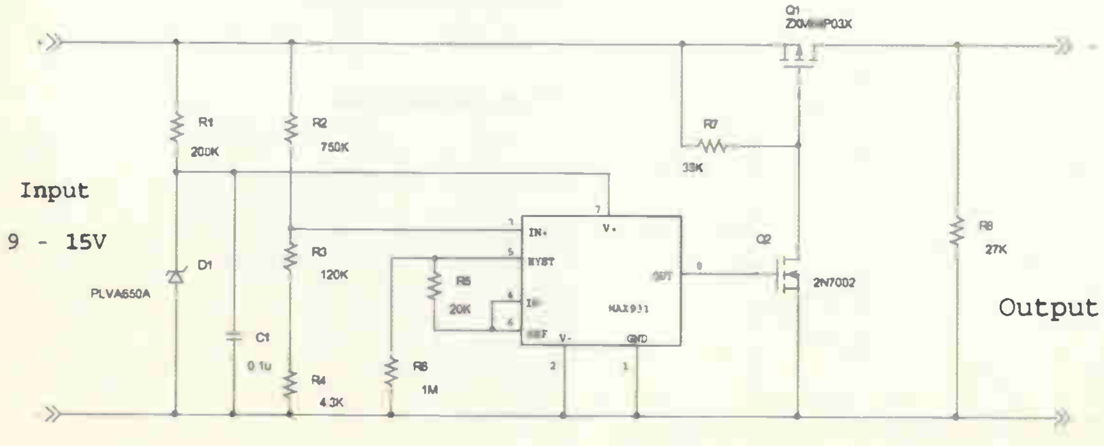
The circuit in **Figure 1** solves the problem by adding a voltage controlled MOSFET switch in series with the DC power line. IC1 (MAX931) is an ultra low power comparator with a 1.182V built-in reference. If the power supply voltage is higher than a predetermined cut-off threshold, the IC1's output is high which turns Q2 (2N7002) on. Since Q1 is a P-channel MOSFET, it will be turned on if Q2 is on. If the power supply voltage is below the predetermined cut-off voltage, both Q1 and Q2 will turn off. The cut-off voltage is set by R2, R3 and R4:

V_{cutoff} = 1.182 x (R2+R3+R4) / (R2+R3)

With the values shown in figure 1, V_{cutoff} is around 8.3V.

When the load current is heavy, a fast response power switch may generate oscillation at the output during the transition. This is because when the power-supply voltage drops just below the cut-off threshold, the switch cuts off the load. Without load, the power supply output voltage will bounce back due to the no load condition. This voltage may exceed the cut-off threshold such that the switch will turn on again. To avoid the oscillation, a hysteresis is added to the MAX931. With R5=20K and R6=1M, the hysteresis band is about 0.33V.

Q1 has an extremely low on resistance. The voltage drop on Q2 is less than 100mV when the load current is 1.2A. The MAX391 only needs 4μA supply current. Its power supply is regulated by D1 at 6.5V. The whole circuitry requires no additional power supply and consumes less than 1mA total current when the power supply is 12V. The turn off time is around 100μS, which is several hundred times faster than that of the original DC power supply.



Yongping Xia
Torrance, California U.S.A

Electronically tuneable CM oscillator using FTFN and OTAs

The oscillator employs a single FTFN, two each of OTAs and capacitors one of which is grounded. The oscillator is free from the preset condition of oscillation. The active and passive sensitivities are low.

Sine wave oscillators find applications in communication, measurements, control systems etc. Employment of FTFN in designing continuous-time circuits is receiving more attention among circuit designers as it is a more flexible and versatile active element compared to OA and CCII. Moreover the use of OTAs in any signal processing circuit provides much needed highly linear tunability and wide tunable range of its transconductance gain which is adjustable through its bias current and as such lending electronic tunability features to the circuit. The current-mode oscillators are advantageous compared to voltage-mode counterparts, as they do not require buffers while driving the loads in the analog signal processing systems. A number of oscillators built around FTFNs operating in CM or VM have been reported in the literature [1 – 7]. However, these oscillators lack the electronic tunability feature which is highly desirable in contemporary IC technology.

Here we are proposing an oscillator circuit that uses two OTAs, a single FTFN, and two capacitors (one grounded). It offers the following salient features:

- operates in current-mode,
- electronically tunable,
- free from preset condition of oscillation,
- resistor-less, and
- low sensitivity figures

A routine analysis of the proposed oscillator circuit shown in the figure yields the following characteristic equation:

$$s^2 + \frac{g_1 g_2}{C_1 C_2} = 0$$

It follows from Eq. (1) that the oscillator is free from the preset condition of oscillation. The oscillation frequency is given by

$$f = \frac{1}{2\pi}\sqrt{\frac{g_1 g_2}{C_1 C_2}}$$

For g1 =g2 = g, Eq. (2) becomes

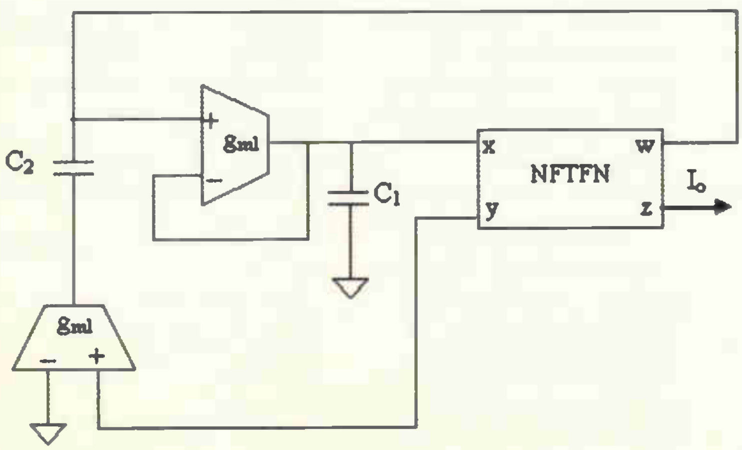
$$f = \frac{g}{2\pi}\sqrt{\frac{1}{C_1 C_2}}$$

Using g = IB/2VT in Eq. (3) one obtains

$$f = \frac{I_B}{4V_T\pi}\sqrt{\frac{1}{C_1 C_2}}$$

From Eq. (4) it is clear that the frequency of oscillation is function of bias current of OTAs. The bias current can be obtained from digital-to-analog converter (DAC), lending the circuit digitally programmable electronically tunable oscillator feature that is important in contemporary circuit construction [8].

N A Shah, S Z Iqbal & B. Parveen
Kashmir
Srinagar



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Shen-Iuan; Yu-Hung Liao: **Current-mode qudrature sinusoidal oscillator using single FTFN.** *Int. J. Electronics, vol. 81, No. 2, (1996) 171-175.*

Abuelmaatti, M. T.; Alzaher, H. A.: **Comment on Current-mode qudrature sinusoidal oscillator using single FTFN.** *Int. J. Electronics, vol. 85, No. 2, (1998) 177-180,.*

Abuelmaatti, M. T.; Alzaher, H. A.: **Current-mode sinusoidal oscillator using two FTFNs.** *Proc. Natl. sci. Counc. ROC(A), vol. 22, No. 6, (1998) 758-764,.*

Abuelmaatti, M. T.; Alzaher, H. A.: **Current-mode sinusoidal oscillator using single FTFN.** *IEEE Trans. on CAS-II, vol. 46, No. 1, (1999) 69-74.*

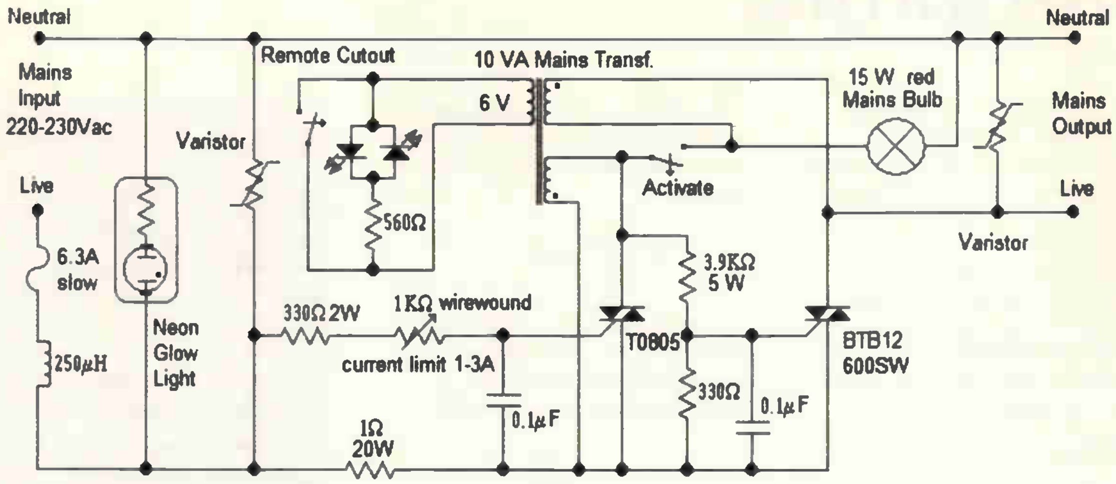
Cam, U.; Toker A.; Cicekoglu, O.; Kuntman, H.: **Current-mode high output impedance sinusoidal oscillator configuration employing single FTFN.** *Analogue Integrated Circuits and Signal Processing, vol. 24, (2000) 231-238.*

Shen-Iuan Liu: **Single-resistance-controlled sinusoidal oscillator using two FTFNs.** *Electron. Lett., vol. 33, No. 14, (1997) 1185-1186.*

Bhaskar D. R.: **Single-resistance-controlled sinusoidal oscillator using single FTFN.** *Electron. Lett., vol. 35, No. 3, (1999) 190.*

Abuelmaatti, M. T.: **New minimum component electronically tunable OTA-C sinusoidal oscillators.** *Electron. Lett., vol. 25, No. 17, 1114-1115.*

Electronic fuse with remote cut-out

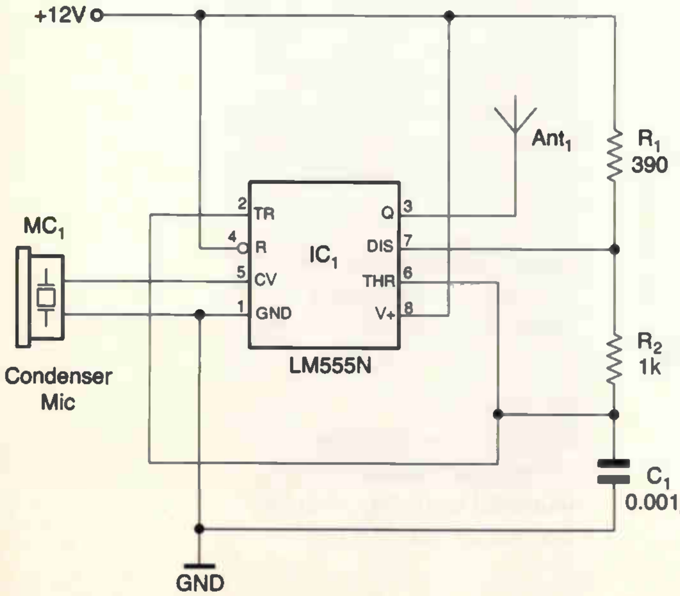


Full short – circuit and overcurrent protection is given by this circuit that suitable for workbench applications in technical schools and laboratories where there is a need to work directly with the mains. Additional features are a clearly visible red lamp indicating that the voltage is present, good isolation of the output circuit when the unit is off, only a few millivolts were measured with no load, current threshold adjustable over a limited range and the possibility of remote cut-out: the 6V from the secondary can be taken anywhere, normally where you are working, even far away from the protection circuit.

Pressing the push button will short-circuit the winding thus deactivating the circuit and removing the mains voltage. A suitable led is placed together with the push button to show whether the circuit is in operation or not. Additional remote cut-out circuits can be wired in parallel if so required. The circuit will switch off if a short is applied at the output without blowing the fuse but it will blow if you try to activate the circuit if a short is already present. If in doubt, first activate the circuit and then apply the load. The BTB 12-600SW is a snubberless triac while the T0805 is a standard triac: you may use other equivalent types but because of the way triacs are driven you cannot use, in this circuit, a snubberless triac instead of a standard triac and vice-versa. The 250μH inductor is a coreless inductor made with 100 turns of 1mm, enamelled wire over a form 27mm diameter and 12mm wide. The mains transformer is a standard transformer with split primary wired in such a way that the circuit will self-sustain once it is activated. The same circuit was implemented with a current limit between 0.1 and 0.3A. In this case you have to change the fuse from 6.3 to 1.5A and the sensing resistor from 1Ω to 10Ω.

D. Di Mario
Milan
ITALY

Low range AM transmitter



Here is a simple radio transmitter for transmission up to 25 metres. It is basically an AM modulator whose signal can be received on a normal AM radio. It can also be used as an AM radio tester.

IC 555 (IC_1) is used as a free running multivibrator whose frequency is set above 540kHz. Here the circuit is designed for a frequency of around 600kHz. The frequency of the multivibrator can be calculated as follows:

$$F=1.44(Rl+2R2)C1.$$

Where resistors R_1 and R_2 are in ohms, C_1 is in hertz. This frequency can be changed by simply replacing R_2 with a variable resistor or C_1 with ganged capacitors, but this may increase the complexity of the circuit. A condenser microphone is used for speech. The IC 555 multivibrator is used as a voltage to frequency converter. The output of the condenser microphone is given to pin 5 of IC_1, which converts the input voltage or voice signal into its appropriate frequency at output pin 3 This frequency produces an electromagnetic wave, which can be detected by a nearby radio receiver and you can hear the mic output in that radio.

Note that the receiver should be AM type.

The circuit operates off a 9V battery. For antenna, connect 2-3m long wire at pin-3.

Raj Gorkali
Kathmandu
Nepal

Frequency divider

The frequency divider of **Figure I** comprises an astable of a 555 timer whose output after inversion with an inverter is given as clock to a MOD-16 counter. The resets of astable and counter are R1 and R2 respectively. The outputs of counter are given to a magnitude comparator as 'A' inputs. The 'B' inputs of magnitude comparator accept binary preset value with which division is to be effected. The range of preset values are 1-15 in binary. The output A=B of magnitude comparator goes high when the counter output is equal to the set value at 'B' inputs. Normally A=B output is held low. The A=B point goes to trigger point of the 555 timer monostable. The A=B point also goes to an inverter whose output is the desired frequency divided output. The monostable resets the counter at high to low transition as shown in Fig.2, Z point of the circuit and the sequence repeats. Fig.2 shows the various waveforms at points X, Y, Z, W and S of the circuit.

For illustration referring **Figure 2** when the 'B' inputs of magnitude comparator are set to binary 0110 (6) and the period of astable is adjusted to 100ms.

I.e.: T = 100ms, thus its frequency $f_1 = 1/100x10^{-3} = 10Hz$
Frequency divided output time period $= 6T = 6x100x10^{-3} = 600ms$
Its frequency $f_2 = 1/600x10^{-3} = 1.67Hz$
Thus divided output $= f^1 / 6 = 1.67Hz$

Thus a range from 1 –15 in binary is obtained by this method.

V. Gopalakrishnan
Bangalore
India.

Figure I

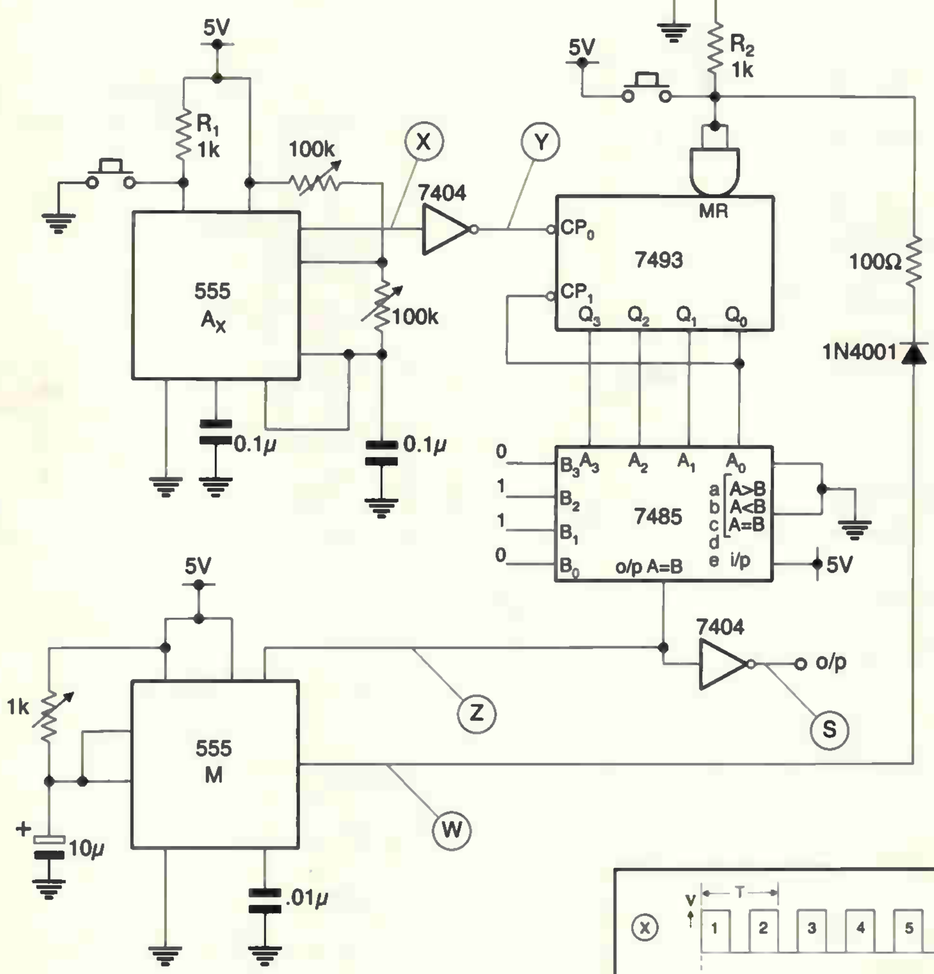
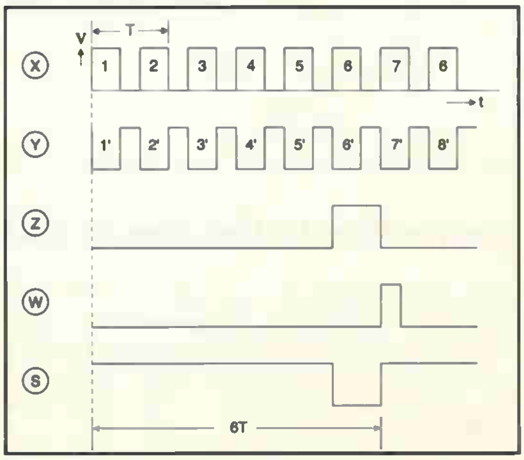


Figure 2



Versatile flasher revisited

In Ref 1 I've stated that it is not possible to create a versatile flasher circuit using an NE555. The reason is the reference voltage divider made up of 5k resistors. As these resistor values are larger in a TLC555, it will perform this function without problem.

As the circuit uses an IC, an internal power supply is created with Cl and D1. R2 limits the charge current of C1. The presence of C1 may cause a light flickering in connected LEDs when the circuit is switched on. Diode D2 will protect the circuit against reverse connection. The timing circuit is built with R1 and C2. By connecting R1 between the output 3 and the capacitor, it is possible to obtain an output duty cycle of 50%. This configuration avoids the need to use diodes in the timing circuit.

As the discharge pin isn't used for the timing circuit, it is possible to use it to control the load. When the timing capacitor is charging, output pin 3 is high and the discharge pin switched off. Under these conditions the complete circuit is powered by the power supply. With the capacitor discharging, the discharge pin is switched on and the circuit terminals are connected together switching on the load. In this state the circuit is powered from capacitor C1.

It is not clear to me how much current this pin can carry as I've haven't found any information concerning this. Measuring the voltage on a working example makes it clear that the current through the discharge pin is best limited to 100mA. At this current the voltage is about 900mV. For larger currents it remains possible to use an external transistor or FET as shown in circuit 2. Here a MOSFET IRF540 is used, switching currents up to 14A.

Ref 1: Versatile flasher circuit, *EW* April 2001

Bernard van den Abeele
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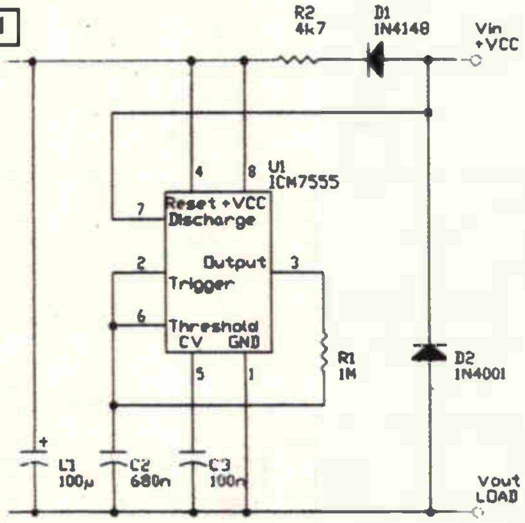
Belgium

1



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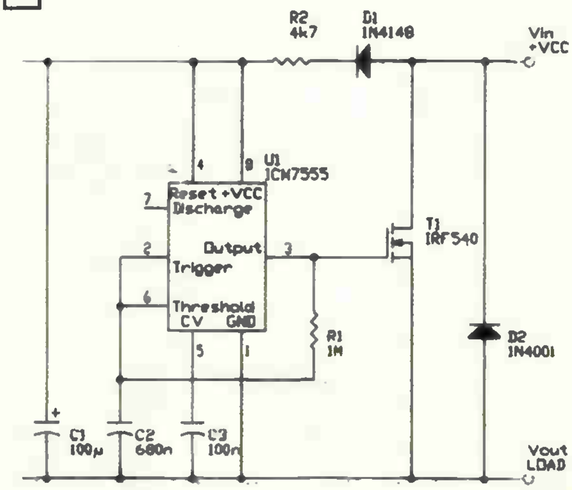
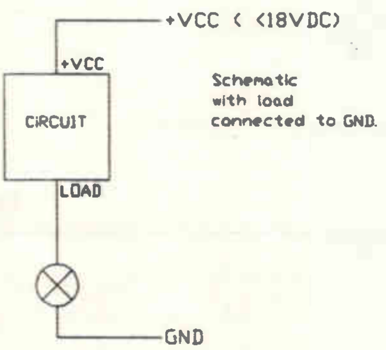


3 CONNECTION EXAMPLES



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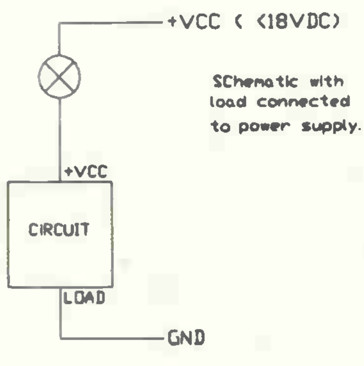


Correction to the simple voltage monitor/alarm, July 04

Unfortunately there is an error on the circuit. Pin 2 of U1a should go to pins 7&6. The cathode of D3 should be connected to pin 5 of U1b, in order for U1b to buffer the zener voltage.

It must have lost something in translation either from paper to PC or vice-versa. Sorry about that.

Incidentally, the unit did prove useful. The battery voltage did remain fine. On a hot day, the alarm system appeared to fail, and the solenoid supply was reduced as the isolation relay turned out to be faulty!

Michael Fallon-Williams

A simplified method for calculating reactance

For capacitor reactance, the basic capacitance reactance formula:
X = 1/(+j2 πfC)

X = reactance Ω ● f = frequency Hz ● C = capacitance μF

Using practical units. C = capacitance μF and drop the imaginary symbol 'j'.

Let C(μF) = 1000000*C(F)

X(Ω) = {1000000/(2π) } / f(Hz) / C(μF)

X(Ω) = 159155 / f(Hz) / C(μF)

If we increase the constant 159155 by 0.0025% 159159
A very small increase.
Now 159159 is twice up the main diagonal of the calculator key board. This allows for quick entry.

X(Ω) = 159159 / f (Hz) / C (μF)

f (Hz) = 159159 / C(μF) / X(Ω)

C(μF) = 159159 / f (Hz) / X(Ω)

These are all of the similar form.
Now using f(kHz) and C(nF); or f(MHz) and C(πF).
The multiples of 1000 cancel and the formulae remain the same.

X(Ω) =159159 / f (kHz) / C (nF)

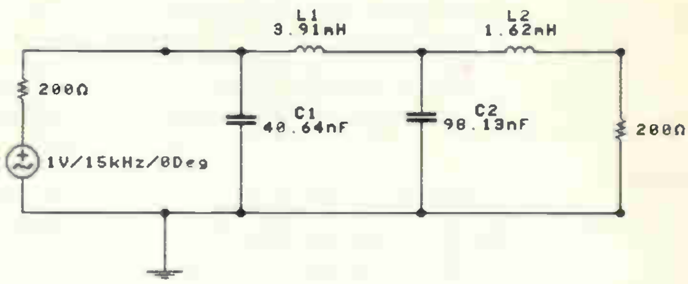
X(W) =159159 / f (MHz) / C(pF)

Combining the results

X = 159159 / f / C) (Ω Hz μF)
F = 159159 / C / X}———{Ω kHz nF)
C = 159159 / f / X) (Ω MHz πF)

For inductor reactance: X = +j2π*f*L
Converting the reactance X to micro Siemens B(μS).

B(μS) = -j 1000000 / (X(Ω))
Again dropping -j and rearranging the formula,
Reactance formula



B(μS) = {1000000 / 2π}/ f / L

but 1000000 / 2π = 159159 (as above)

B(μS) = 159159 / f / L

This is in the same form as the capacitance formula.
Again using f(kHz), L(μH); or f (MHz) , L (μH) .

B (μS) = 159159 / f (Hz) / L (H)
B(μS) = 159159 / f(kHz) / L(μH)
B(μS) = 159159 / f(MHz) / L(μH)

Combining the results

B = 159159 / f / L) (μS Hz H)
f = 159159 / L / B}———{μS kHz μH)
L = 159159 / B / f) (μS MHz μH)

Example in the use of the new formulae.
To design a four element Butterworth low pass filter.
Butterworth Parameters 0.7654. 1.848. 1.848. 0.7654.

Cut-off frequency	15 kHz		
Impedance	200Ω	(=5000μS)	
Co(nF) = 159159 / 15(kHz)/	200(Ω)	=53.1nF	
Lo(μH) = 159159 / 15(kHz)/	5000(μS)	=2.12μH	
Cl = Co * Bl	0.7654	* 53.1nF	= 40.64nF
Ll = Lo * B2	1.848	* 2.12μH	= 3.91μH
C2 = Co * B3	1.848	* 53.1nF	= 98.13nF
L2 = Lo * B4	0.7654	* 2.12μH	= 1.62μH

R.Gooding
Cheltenham
UK

Accurate one second clock

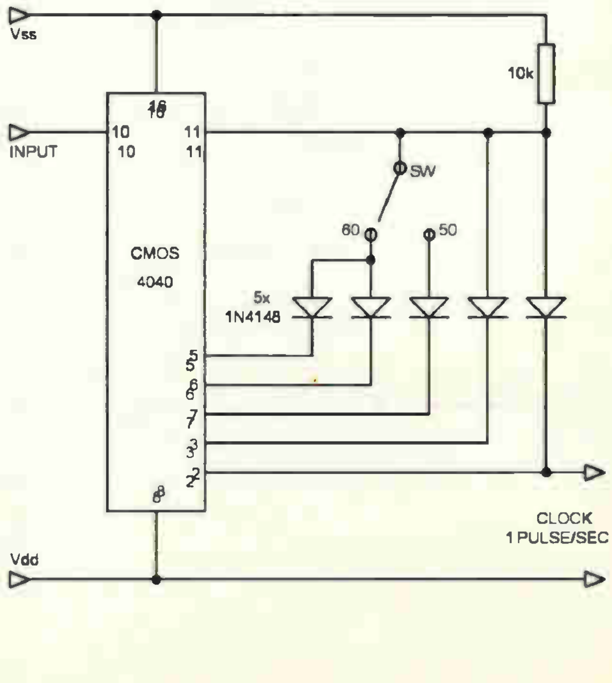
A handful of common components will make a very accurate one-pulse-per-second clock, driven from a 50 or 60 Hertz mains supply but with no direct connection to it. The circuit diagram shows a CMOS 4040 counter/divider chip set up with 5 diodes to divide the frequency of the input signal at pin 10 by 50 or 60, depending on the position of the SPDT switch.

The impedance at pin 10 is so high that simply touching the pin is probably enough to provide the necessary input signal. An absolutely certain input is provided by a piece of wire wrapped a couple of times around any convenient mains cable or transformer. No other connection is necessary.

The output signal has very fast rise and fall times and is not far from being 50% symmetrical.

Obviously if only one mains frequency is being considered, the circuit can be further simplified by the omission of the switch and the unwanted diode(s).

David Ponting
Bristol
UK



Optical powering and isolation of a loop alarm

Low-power, dielectrically-isolated supplies are useful for industrial sensors and, as here, for simple loop alarms. While 50/60Hz transformers are available in powers down to about 0.25VA, even this is overkill when we only need a few milliwatts. High frequency ferrite or capacitive coupling transformers are possible, but need an oscillator drive. Optical powering is an attractive option.

The International Rectifier PVI5100 5V 10µA isolated power source is designed just for this, but you can generate far higher currents if you build your own. The circuit fragment of **Figure 1** uses a string of three silicon photodiodes illuminated by infrared (IR) LEDs to generate a couple of milliamps photocurrent, sufficient for robust drive of a long circuit loop, including a standard optocoupler LED for the isolated return path.

With 880nm SFH487 LEDs driven by about 22mA from the 5V supply, the three BPW34 photodiodes together generate an open-circuit voltage V_{OC} = 1.45V and short-circuit current I_{SC} = 1.85mA (**Figure 2**). Connected to the TLP627 optocoupler I obtain about 1.55mA operating loop current, quite adequate to drive the phototransistor to saturation with a 10k or even 1k load resistor. Opening the external circuit produces a 5V output from the dark phototransistor. The resistor labelled 0R is not strictly necessary, but can be added to stabilise the external current with temperature changes. Two photodiodes don't quite have the terminal voltage to drive the optocoupler LED. Note that the total photocurrent is equal to the least of the individual photodiode currents, so try to equalise the illumination powers.

A solar cell might be used instead of the three photodiodes, but they are optimised for high intensity broadband ambient light, typically perform poorly with IR LED illumination, and are unnecessarily large and expensive. Even a small silicon photodiode is big enough to capture most of the LED's light and it achieves about 0.6mA/mW responsivity near to 900nm wavelength. To optimise coupling I polished off the LED lenses to within about 500µm of the wire-bond. Lensless types like the LD242 series avoid polishing. If you need high voltage isolation then leave the lenses on, maintain long 'creepage' distances, and improve coupling with a glass tube waveguide.

If you really want to push your luck, the chain of three GaAs LEDs and silicon photodiodes can be replaced by a single pair of visible-wavelength LEDs, one used as source, one as detector. The advantage of a visible-light LED as detector is its much higher open circuit photovoltage, which scales with the material band-gap. Red and green LEDs can generate more than 1V and 2V respectively. Using a pair of 660nm LEDs placed face to face I obtained V_{OC} = 1.4V, enough to drive a GaAs optocoupler LED, but only with ISC = 30µA. Nevertheless, this was sufficient to obtain a few microamps in a 1MΩ load, as with the PVI5100. Source modulation and a proper transimpedance circuit behind the optocoupler can make even this a robust approach.

Mark Johnson
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Figure 1

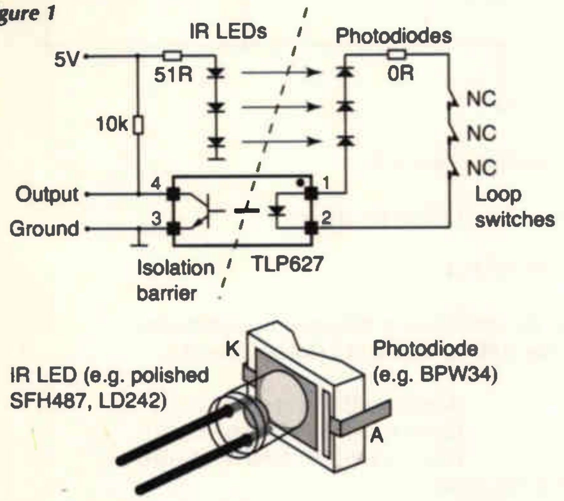
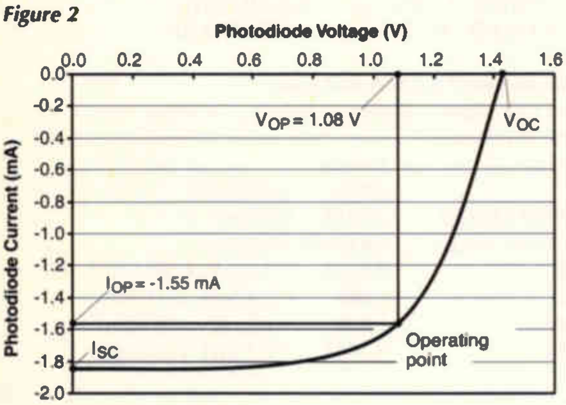
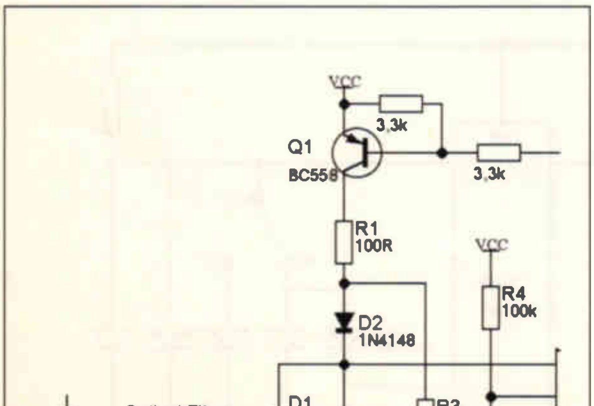


Figure 2



Two-way serial communication over a single optical fibre



Optical fibre is the best transmission medium when reliable communication has to be performed in electrically noisy environments; especially when galvanic isolation is required. However, a drawback of fibre optics is that usual transmitters and receivers are unidirectional, since they must be optically coupled to the fibre. Two-way communication usually requires two fibres running in parallel, and also two pairs of optical transmitters + receivers.

This circuit, on the other hand, provides bidirectional, half-duplex optical communication using a single fibre. It works due to a not widely known characteristic of infrared LEDs, in that they can also work as photodiodes, converting optical input power into electric current. Thus the same device, optically coupled to a fibre, works as a transmitter and a receiver.

Transistor Q1 drives infrared emitter D1 during transmission. When receiving (signal TXD at high logic level), Q1 is cut off and D1 develops around 10-20mV peak over R2. U1 is a FET input op-amp, operating as a voltage comparator (negative rail input range is required). R4 and R5 set the slicing level. Diode D2 helps to isolate D1 from Q1' collector capacitance.

This circuit has been tested with more than 5 meters of 1mm plastic fibre, and was found useful at more than 19.2kbps. Transmit and receive levels are TTL / CMOS compatible.

Higher speeds can be obtained reducing R2; but received voltage will be lower, requiring a high speed, lower input offset comparator to adequately recover the data.

Guido Stolfi
Sao Paulo
Brazil

Current-controlled sawtooth generator

Sawtooth signals whose amplitude and frequency can be controlled by varying current can be generated using this circuit. It operates as follows.

When the voltage at pin 3 is high, at V_{cc} that is, the output current of the first operational transconductance amplifier, namely OTA_1, equals the input auxiliary bias current I_{b1}. This is attributed to the fact that when the differential input voltage of the OTA is greater than few millivolts, then all the bias current will be pushed into the output of the OTA.

Meanwhile, a high voltage at pin 3 will switch off the switch *S*. Thus, the capacitor C will charge at a constant rate equal to I_{b1}/C.

The charging cycle terminates when the voltage across the capacitor reaches V_s, the value of the voltage at pin 5. Voltage at pin 3 drops to zero, and switch *S* closes. The capacitor discharges immediately through the very low resistance of the switch.

When the voltage across the capacitor drops below $V_s/2$, the voltage at pin 3 goes high again, switch *S* opens and another charging cycle starts. During the charging cycle, the amplitude of the sawtooth increases from $V_s/2$ to V_s.

Using a standard 555 timer, with three equal resistances each of 5kΩ, and assuming that,

$$\frac{1}{g_{m2}} << \frac{10}{3}\text{k}\Omega$$

the voltage at pin 5 can be approximated by,

$$V_s \cong \frac{V_{cc}}{5g_{m2}}$$

where,

$$g_{m2} = \frac{I_{b2}}{2V_T}$$

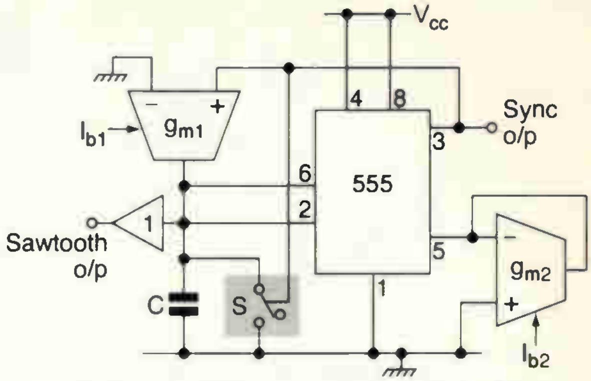
is the transconductance of OTA_2 in mA/V and V_T is the thermal voltage of around 25mV at room temperature. Thus the amplitude of the output sawtooth can be expressed as,

$$A \cong V_{cc}\frac{V_T}{5I_{b2}}$$

Discharging time is very small, especially for very low values of switch on-resistance. Ignoring it, the frequency of the output sawtooth can be approximated by,

$$f \cong 5I_{b1}\frac{I_{b2}}{C}V_{cc}V_T$$

The frequency of the output sawtooth can be controlled by adjusting the



S is an analogue switch, e.g. type HI-201HS
Amplifiers g_{m1} & g_{m2} are OTA3080s

It is possible to change both amplitude and frequency by varying the current signals to the operational transconductance amplifiers g_{m1} and g_{m2}.

auxiliary bias currents I_{b1} and I_{b2} and the amplitude can be controlled by adjusting the auxiliary bias current I_{b1}.

The sawtooth output is taken from the voltage across the capacitor via a buffer to isolate the load. Output at pin 3 is a pulse with very narrow width, which may be useful for synchronisation.

Muhammad Taher Abuelma'atti
King Fahd University of Petroleum and Minerals
Dhahran
Saudi Arabia

Controller for high torque servo

(diagram on page 46)

It is sometimes convenient to make use of a commercial high torque servo such as the Futaba S3301 rather than build one from scratch. This servo is intended for radio-control applications, and as such expects a pulse 1.5ms wide in order to obtain the central position. It would have been possible to strip out the existing circuit and replace it with a simple linear voltage control; however, it was decided 'not to wreck it' but instead try to make an external pulse generator. This turned out to be more tricky than anticipated.

A feature of these servos is that they consume a large current unless in the 'idle' mode, i.e. with the input pulse gated low. In our particular application, which is to activate a sequential gearbox, the 'non-idle' current of 1A would drain the 12V battery needlessly. A better strategy is to wake-up the servo only when a gear-change is required.

Unfortunately, simply gating-off the pulses does not work very well, as the internal circuit of the servo tends to make the shaft wander off the central position immediately after the final pulse - probably a low duty-cycle pulse-train is necessary to keep the servo half-awake. Even then, there is a risk of noise causing the occasional twitch. Since our gear-box requires its lever to return to the central position after each up or down shift, a sure-fire method is essential, and the best solution was found to be to disable the servo totally after each shift-and-return by removing its supply via the p-channel mosfet Q1 shown in Fig. 1, red lead connected to drain. This has the added benefit of keeping the quiescent current below 10mA.

A pair of micro-switches actuated by the shift-paddles on the steering wheel have their signals filtered and de-bounced by means of R1 C1 and R2C2, and the nand-gates N1, followed by the monostables N2. These each produce a pulse some 625ms wide, which is enough for the servo to move at least 600 under load. The next monostable (N3) produces a longer pulse, about 1.25s wide. The servo is activated for this total period, and as a result has 625ms to flick the gear-lever up (or down) and then a further 625ms to return to centre. Note that the pulses as well as the supply to the servo are gated 'on' and then 'off'.

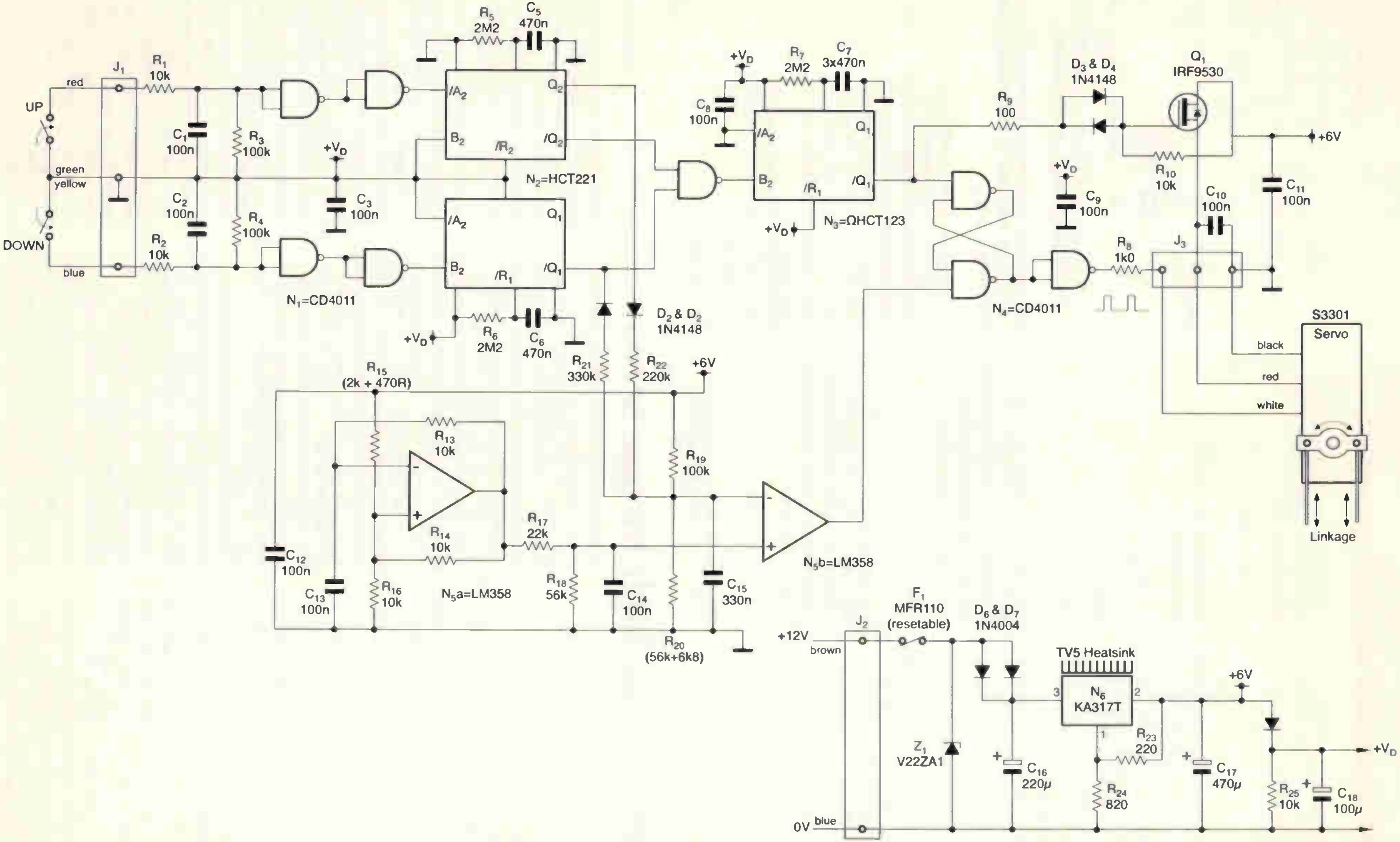
The pulses themselves originate from the generator-modulator constructed around the dual op-amp N5. The left section (N5a) is a square-wave generator running at 200Hz, and is guaranteed to start-up when the power is first applied. Its output is converted to a reasonably good triangular wave by R17,R18,C14, and the right section (N5b) behaves as a comparator whose reference level is set at the junction of R19,R20. The capacitor C15 acts as a filter, to prevent noise caused by the current-hungry servo-motor upsetting the circuit. For the central position, the 1.5ms pulse-width is determined by R19,R20, but for the 'down' pulses 2.2ms wide and the 'up' pulses 0.8ms wide the resistors R21 R22 come into play: moving the reference level down or up causes pulse-width modulation, as desired.

Since the servo-motor requires a +6V supply capable of providing current surges of at least 2A peak, a hefty regulator is called for, and the KA317T was found adequate at N6, as long as a low-impedance capacitor C17 is included. The digital section employs some HCT devices, and so the Vd rail is reduced to approximately 5.4V by D5. It is advisable to fit the protective components FIZ1, as well as the diodes D6D7 which prevents mishap in case the battery is reverse-connected.

This circuit may find use in industrial and scale-model applications also. If a smaller servo is employed, and idle current is not a concern, the scheme can be simplified, even to the extent of using the generator-modulator only. Simple 'down' and 'up' commands can be introduced direct from the microswitches to R21 R22, omitting the logic and monostables. Another possibility is to connect a joystick-potentiometer between the OV/6V rails, and take its wiper to C15 via a limiting resistor >200K. The pulse-width can be trimmed at R20.

C J D Catto
Cambridge

Controller for high torque servo



A collection of non-inverting logic translators

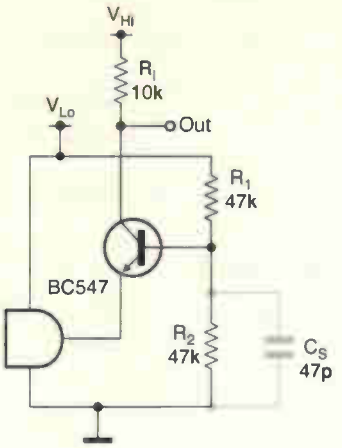


Figure 1a

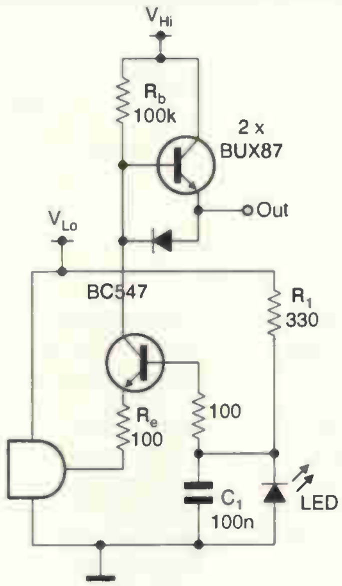


Figure 1b

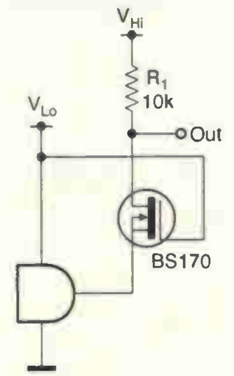


Figure 1c

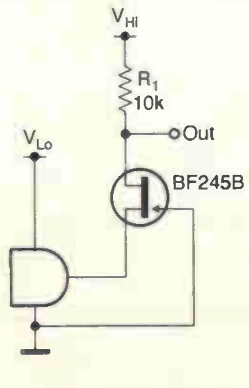


Figure 1d

Many designers seem to be unaware of the possibility to build voltage translators using only a single-stage. You only have to look at the schematic of a VCR for instance, to realise that whenever such a function is needed, as in the Vacuum Fluorescent Display drive section for example, two common-emitter stages are generally cascaded. Although this may be justified in some cases, when a high source-current is required, it is most of the time totally superfluous.

The key to avoiding this wasteful double-inversion is to use a common-base (or gate) configuration. Of course, the current gain is reduced to unity, but in most cases, low-voltage logic families, such as TTL, tend to have a healthy low-level output capability and do not need current boosting.

How is it done in practice? **Figure 1a** shows the basic configuration: the signal to be translated drives the emitter of a general-purpose transistor whose base is held at mid-supply by means of the divider formed by R_1 and R_2; on the collector side, a resistor pulls up the potential to the higher voltage supply. The component values shown are typical and are suitable for general-purpose applications. If VHi exceeds 50V, other types of transistor can be used: the 2N5551 withstands 160V and the BF422, 250V.

In its basic form, this circuit is relatively slow; adding a small capacitor between base and ground can reduce the switching times, and if maximum speed is needed, substitute a HF type for the transistor: a BF494 for example. Note that even with these modifications, the rise time will depend essentially on the load capacitance; if it cannot be reduced, the only way to increase the speed is to decrease R_1, at the expense of the current consumption.

The logic low output voltage of this level shifter will be that of the low-voltage gate plus the saturation voltage of the transistor; in the case of a TTL IC, this amounts to about 600mV. The logic high output voltage is essentially equal to VHi (unloaded of course).

In practice, R_2 can often be omitted; it is certainly the case with CMOS families, thanks to their rail-to-rail swing, but despite their theoretical high level output voltage of only 3V, TTL families also seem to operate satisfactorily under those conditions.

When operation at really high voltages is required, it is preferable to use the circuit of **Figure 1 (b)** rather than simply choose a transistor having a higher breakdown voltage. This circuit has rigidly controlled operating conditions, which provide increased protection against breakdowns; in addition an active load is used instead of a simple resistor: this increases both the speed and the output current with no power penalty.

For applications requiring the lowest low-level output, a Schottky diode must be used between the base and emitter of the upper BUX87; in other cases, a 1N4148 is perfectly adequate. This circuit operates at up to 450V and down to DC and 1000V under pulsed conditions; it is therefore suitable for off-line converters, frequency variators etc.

If a small MOS transistor is substituted for the bipolar type, the base resistors can be eliminated **Figure 1(c)**. The main drawback of this circuit is the relatively large drain capacitance of the MOS leading to slow rise-times as compared with the bipolar version. You also have to make sure that the threshold voltage of the transistor is lower than the supply voltage VLo. This is not likely to be a problem in 5V, but at 3.3V and below, this may be a concern.

The highest performance version is that of **Figure 1d:** not only are the bias resistors unnecessary, but the circuit also benefits from the low capacitance and high speed of the jFET. Interestingly, this circuit is not referenced to the logic supply, but relies on the depletion mode of action of the transistor.

For proper operation, you simply have to ensure that the transistor has a Vp smaller than the supply VLo (at lower supply voltages, you may substitute a BF245A).

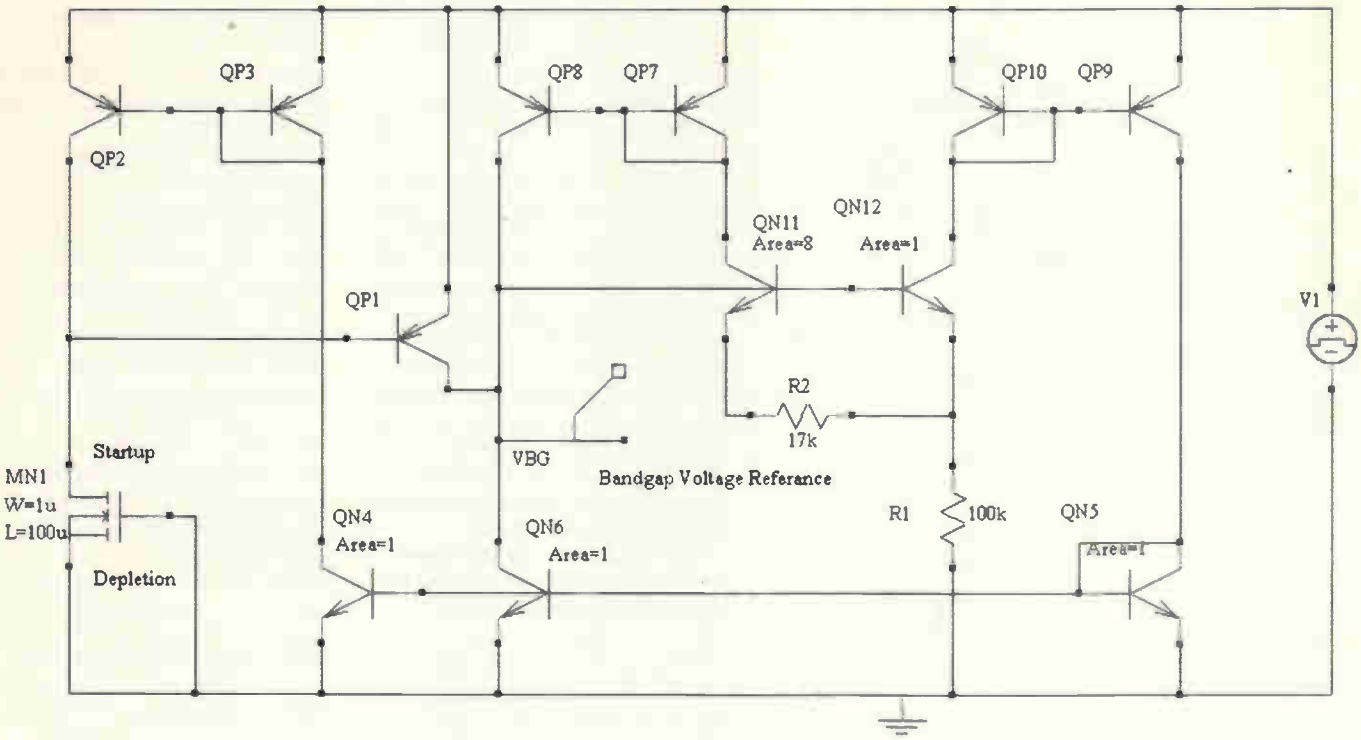
Finally, why would you bother to reinvent the wheel? After all, logic translators exist in integrated form and come in many variants. Here are some reasons:

Integrated translators come in packs of 4 or 6; very often, you only need one or two. Even when you can use up the whole package, it may not be practical because the four operators will be needed at the four corners of the board, making the routing of the PCB terribly awkward.

With home-brew translators, you can tailor the circuit to your application: if you change the resistance values from kΩ to MΩ in **Figure 1a**, the circuit operates as before, but at micropower levels. And finally when you have to operate at hundreds of Volts, discretes are the only option.

Louis Vlemincq
Auderghem
Belgium

Low area, low power startup circuit



The circuit design described here was developed to solve some specific issues in a design of a Band Gap voltage reference in a BiCmos process. However, the technique described here can be applied to many alternative situations.

The principle issues involved were:

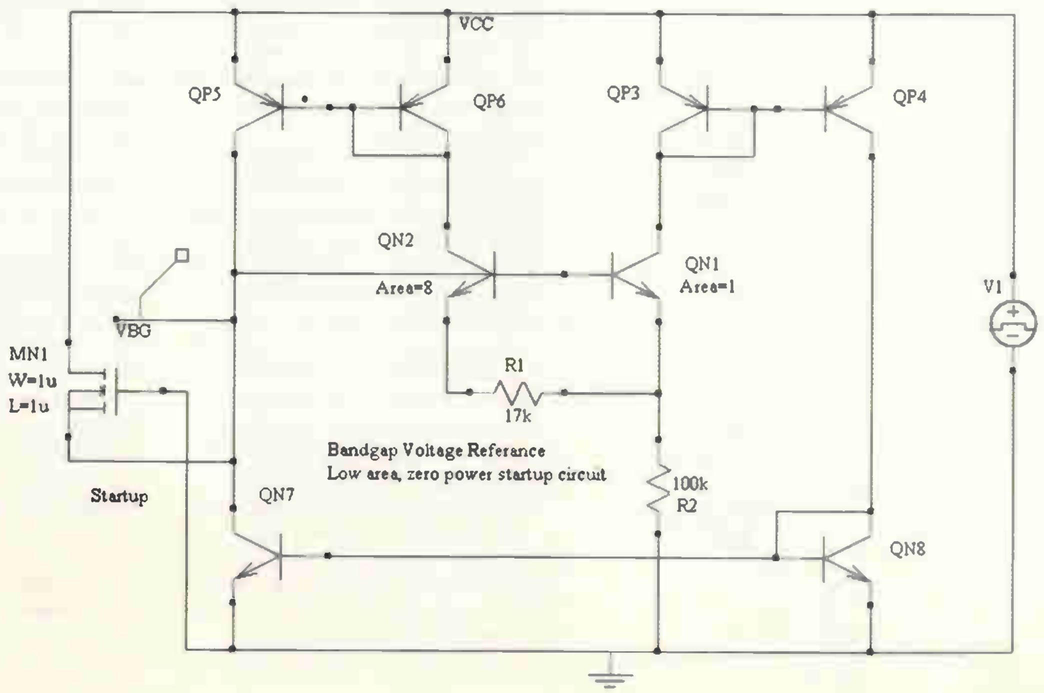
1 Die Area - The product was to be packaged in a pico package.

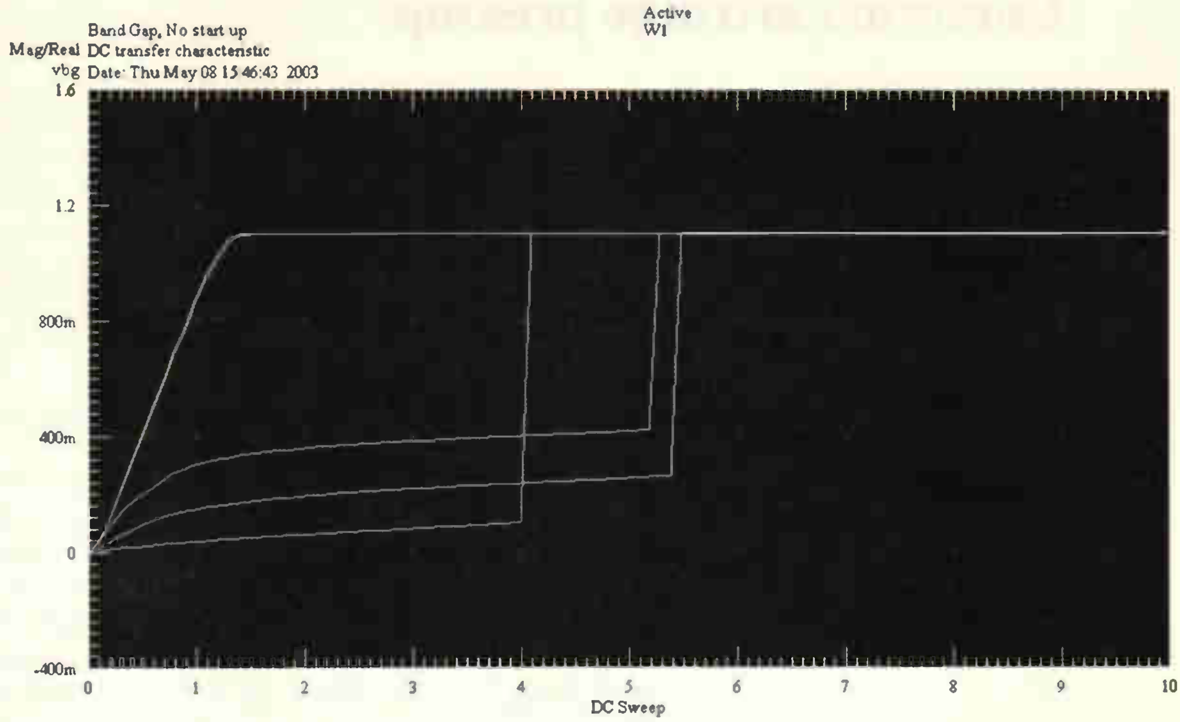
2 Low power - The total supply current was to be at the 1ua level.

The fundamental consequence of these specifications is the difficulty in achieving such low current without using high, mega ohm resistors, or long channel mosfets, which take up significant die area.

Standard PTAT (proportional to absolute temperature) bias circuits automatically solve one aspect of generating a low current with low value resister by applying a small delta Vbe of the order of 50mV across the resistor. However, the classic problem to these circuits is that they need a startup current. This is because they can start in a stable mode where all currents are zero. To avoid this, a current must be applied to ensure startup, but is then switched of once startup has occurred. The issue is that this startup current usually has to run all the time, it is only diverted from the bias circuit. If this startup current is to be small, very large resistors or alternatively, long depletion mode mosfets must be used. **Figure 1** shows such an approach

Fig. 1 is a basic band gap reference, with main currents being set by delta Vbe/R2. In





this case delta Vbe is a standard 52mV from 25mV.Ln(8), giving a current of 3µa. If QN6 and QP8 are non conducting, indicating that the main circuit is off, MN1 will pull current through QP1 forcing QN11 and QN12 to turn on, which will force Q6 to turn on. Once Q6 is on, it will cause the startup current of MN1 to be diverted away from QP1. .

New Circuit

The modified circuit also uses a depletion mosfet, but in this case, the mosfet is minim sized, rather than a long channel device, and the startup current is disabled completely once started.

If VBG is lower than the pinch off of MN1, MN1 will feed current into that node. This will startup up the circuit via QN2 and QN1. Once the circuit starts up, VBG will set itself to the bandgap voltage say 1.2V. This voltage will causeMN1 to turn off, since it results in a pinch off voltage being applied to MN's gate and source.

Figure 3 shows simulation results from **Figure 2,** using SuperSpice, with temperature stepped from -50 to 150 in increments of 50 degrees, without MN1 in circuit.

Figure 4 shows **Figure 2** simulation results using SuperSpice, with temperature stepped from -50 to 150 in increments of 50 deg., with MN1 in circuit.

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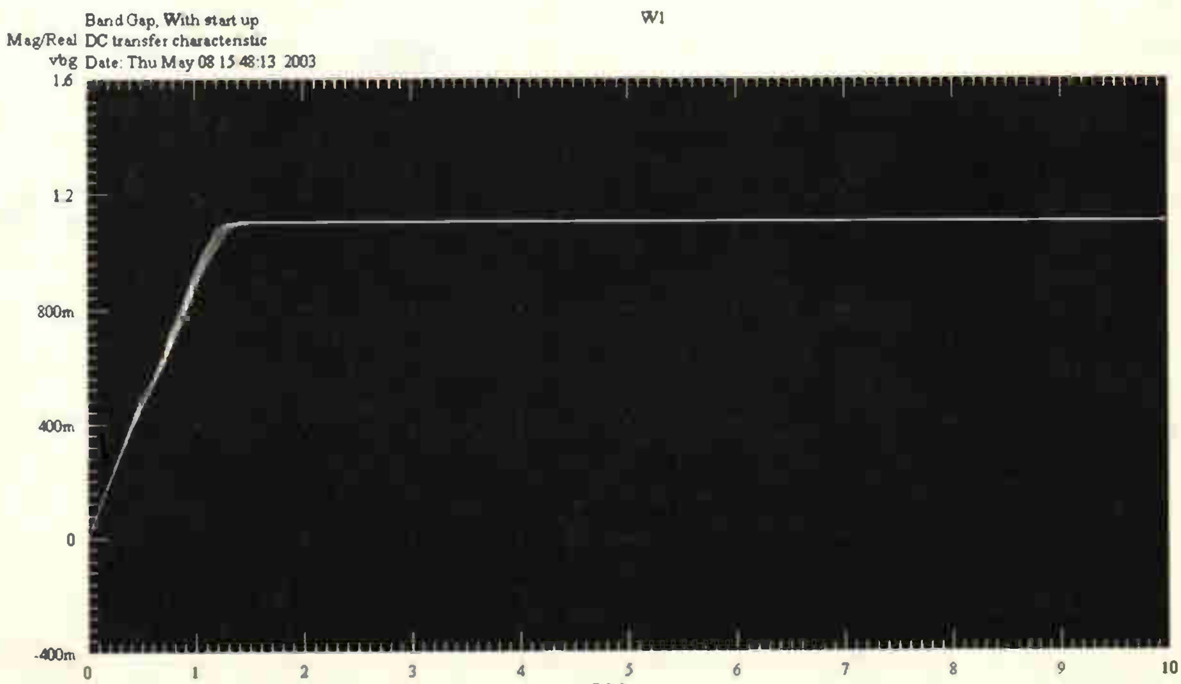
U.K.



Ceramic cartridge preamp

This circuit would appeal to anyone restoring a Garrard SP25 turntable who would prefer to retain the Sonotone cartridge for authenticity but would prefer an improved sound quality.

It is often stated that ceramic cartridges produce an output voltage proportional to the force applied, which is only true if no load resistor is connected. More accurately, the *charge* is proportional to the applied force. With no load

$$V = \frac{Q}{C}$$

where C is the self-capacitance of the cartridge. So voltage is proportional to charge, and to applied force.

Applying a sinusoidally varying force: $F_0 \sin(\omega t)$ and differentiating gives:

$$\frac{dQ}{dt} = F0\,\omega\cos(\omega t)$$

obviously $\frac{dQ}{dt}$ is the output current, so the output current varies with frequency, in exactly the same was as output voltage varies with frequency in a magnetic cartridge.

The ceramic cartridge can be modelled as a Norton source which varies with frequency loaded with a capacitor (the cartridge's self capacitance)

To achieve the 3183µs time constant all that is required is to load the current generator with a resistor and a capacitor in parallel, with the correct time constant. The capacitor is already in place (the cartridge's self capacitance) so the resistor is chosen as

$$R = \frac{3180\mu s}{C}$$

C was measured at 560pF, which gives R as 5.6MΩ. It should be noted that this is much greater than the 1MΩ normally used as a load resistor for a ceramic cartridge, and could explain the poor bass response often observed. A 1MΩ resistor in this position would produce a roll-off in the overall frequency response starting at 280Hz falling off below this figure at 6dB per octave. Having achieved the 6dB/octave slope from 50Hz, the next task is to obtain the flat section between 500Hz and 2122Hz and the continued 6dB/octave slope above 2122Hz. This is achieved with the op-amp circuit which gives a gain that rises from unity at 500Hz to 12dB above 2122Hz, R2/C1 being the 75µs time constant. This part of the circuit is normally missing altogether from a ceramic cartridge preamp, usually accompanied by the claim that the cartridge is 'mechanically equalized' to achieve this, and if you believe that, you'll believe anything.

A FET op-amp is mandatory here, as the input source impedance is 5.6MΩ at low frequency. R4 buffers the output of the op-amp against capacitive loads which could make it unstable.

Overall results were impressive. it was never intended to compete with the best of the magnetic cartridges. but it would certainly put some of the cheaper ones to shame. One thing that is immediately apparent is the lack of noise, but with an overall gain of only 12dB against over 50dB for the average magnetic cartridge it is hardly surprising. The circuit does have a tendency to pick up 50Hz hum, so mounting it underneath the chassis of the turntable as close as possible to the tone arm is a good idea. It is a matter of choice as to whether it is worth spending the extra money for a AD712 over a good old TL072.

Ian Benton
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R1 = 5.6M ¼W

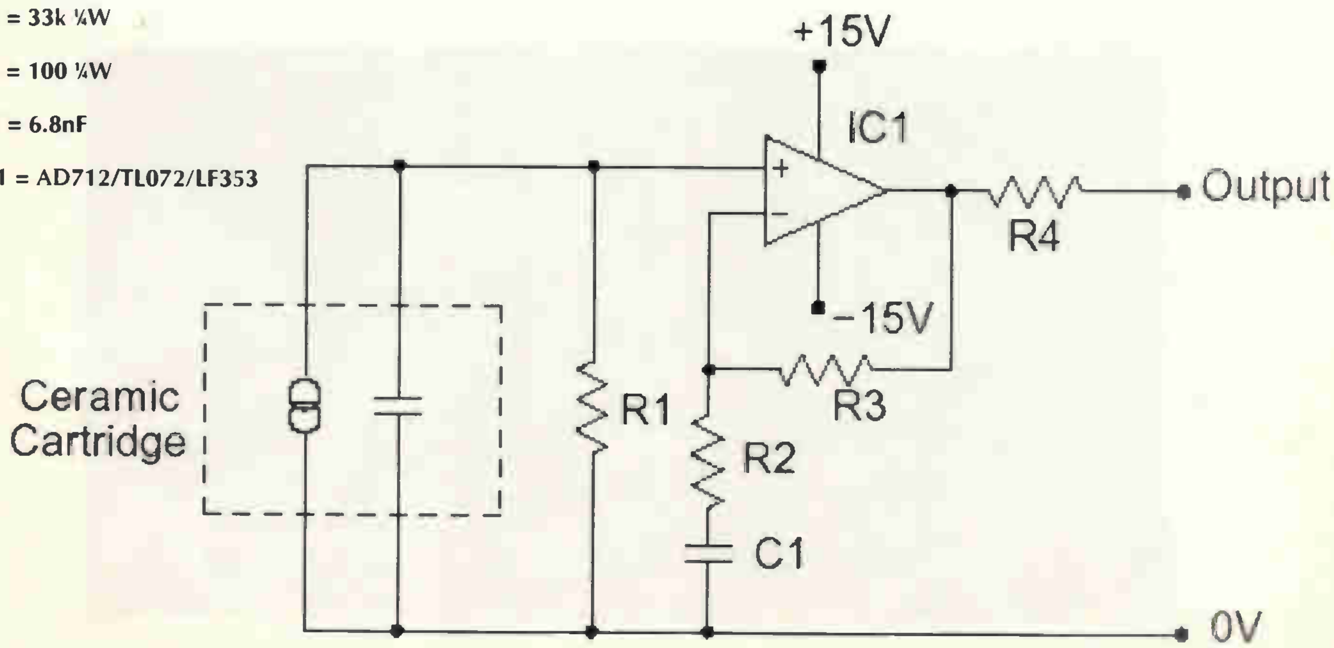
R2 = 11k ¼W

R3 = 33k ¼W

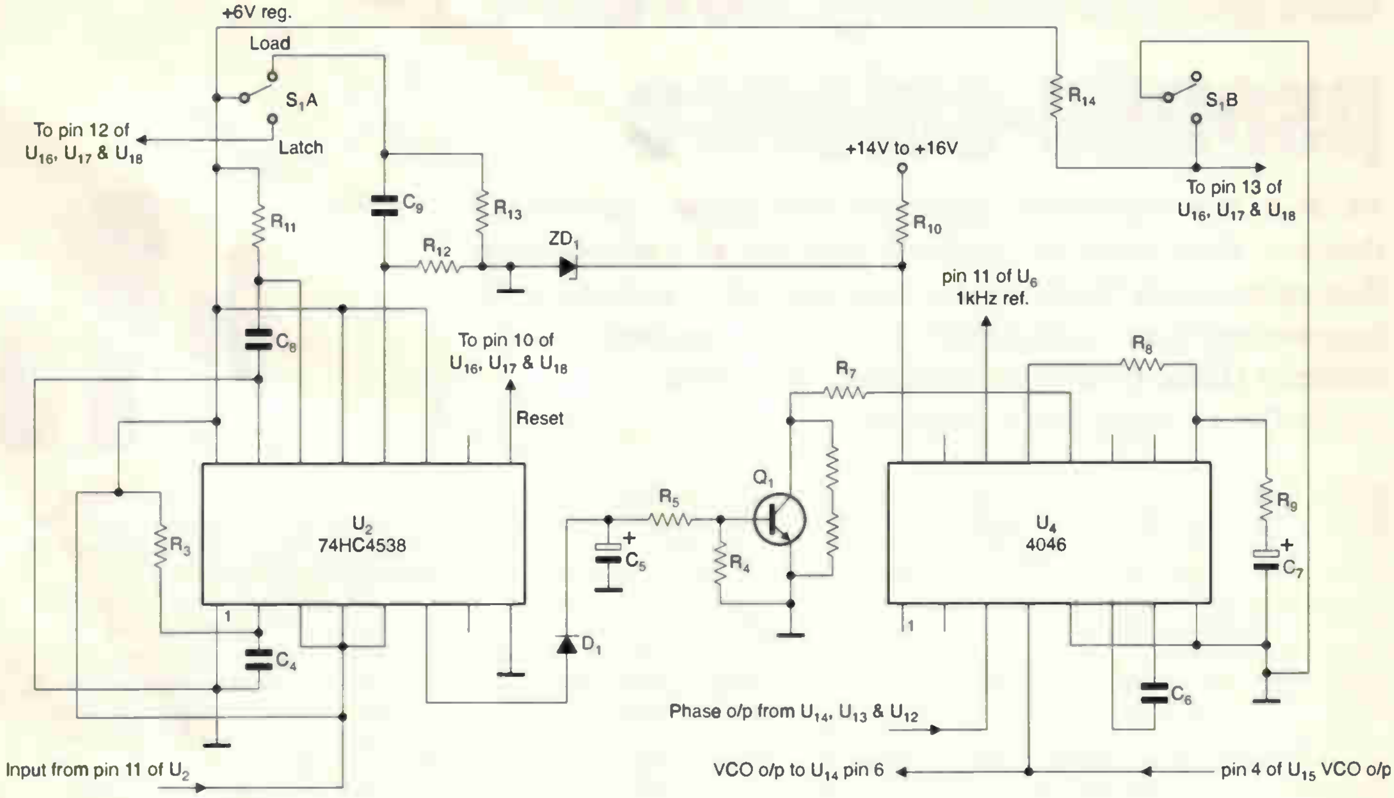
R4 = 100 ¼W

C1 = 6.8nF

IC1 = AD712/TL072/LF353



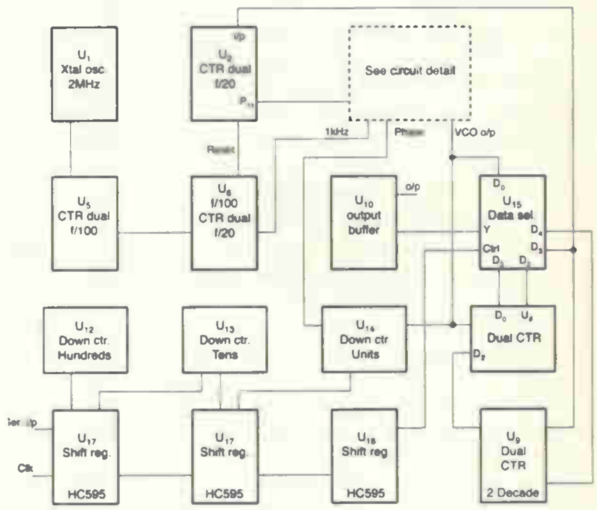
Frequency synthesiser with auto ranging and reset on loading



The detailed schematic shows the two ICs which give the ideas shown in the title. Refering to the block diagram, U_9 divides the o/p of the VCO, U_4, by 1000 which is further divided by 20 in U_2 and fed to U_3. U_2 is reset at a 200Hz rate by division of the 2MHz crystal oscillator by U_5 and U_6. U_6 also outputs the 1kHz reference to U_4 the PLL. When the VCO o/p reaches or exceeds 228kHz, a pulse output from U_2 inputs U_3 before U_2 is reset. This triggers the lower half of the dual MV13 for the time set by R_3 and C_4. Capacitor C_5 is charged through D_1 switching Q_1 on through R_5, in effect shorting out the VCO timing resistor R_6, leaving only resistor R_7 with C_6 in the tuning circuit. This allows operation from 1kHz to 1.5MHz with a post VCO division of 0, 10, 100, 1,000 and 10,000 by programmed data selection in U_{15}, of the outputs of U_8 and U_9. Practical limitations make selection by the units counter alone not suitable since modulation of the pulse width by the phase detector of the PLL may be visible. To obtain three digit frequency resolution the minimum selection would be 111kHz (1.1.1) and the maximum 1.221MHz (11, 11, 11) from the usual keypad entry of zero to 11. The keypad clock uses six pulses to shift each data nibble into the shift registers involved, U_{16}, U_{17} and U_{18}.

The upper section of U_3 along with switch S1A and B is used to reset U_{16}, U_{17} and U_{18} when set in the load position, this triggers U_3 as C_9 charges through R_{12} giving a negative reset pulse of ~300µS which may extend to ~ 600µS because of switch contact bounce. The enable common line to the shift registers holds them in the hiZ condition through R_{14} during loading. R_{13} discharges C_9 when set to enable, preparing for reset when required. Loading is done LSB first, division selection 0 for no division (or no entry) 1 for 10, 2 for 100, 3 for 1,000, or 4 for 10,000. Units next 0 to 11, Tens 0 to 11 and hundreds last 0 to 11. A single entry of 1 to 11 will set the hundreds counter only, from 100kHz to 1.1MHz. Putting the switch in the down, enable position, will latch and activate the registers and down counters U_{12}, U_{13} and U_{14} resulting in the selected frequency going to U_{10} the buffer output amplifier from data selector U_{15}. U_{10} is resistance matched to a 50 or 75Ω line and will give 3V into 50Ω or 6V open circuit.

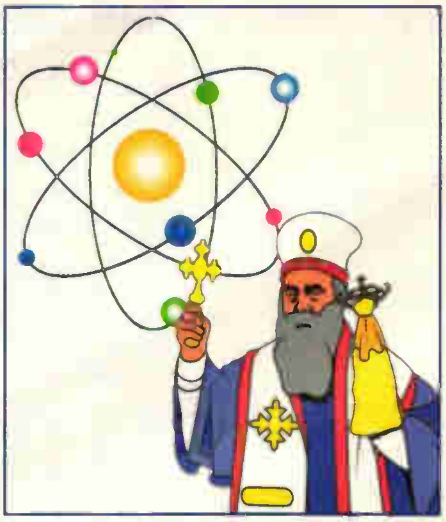
The choice of a 6 volt supply was due to the use of mixed CMOS and



HCMOS available when this project was built as a switch controlled version. The zener control of U_4 was to get the best performance from the 4046 PLL. The use of all HC devices with a 5V supply would probably extend operation several MHz by the use of the range switching idea described here, with component selection and the range switching frequency chosen.

C.V. Carlson
Calgary
Canada

Engineering versus pseudo-science



“ Men are deplorably ignorant with respect to natural things... they must be made to quit the sort of learning that comes only from books, and that rests only on vain arguments from probability and upon conjectures. ” William Gilbert from De Magnete, AD 1600.

Leslie Green CEng MIEE, explains

In the olden days, science did not exist as we know it today. Ideas were propagated based on opinions, unsubstantiated by any facts. Today we find it amusing that the Catholic Church insisted that the Earth was the centre of the universe. But in those days, anyone who tried to go against the Church's opinion was in trouble. Galileo, for example, was convicted of heresy in 1633 for this very offence.

Consider what is known today as Newton's second law: F=ma. This says that if you apply a constant force to a mass, the mass will accelerate indefinitely: even school-kids know this. But every adult knows you have to pull a cart harder to make it go faster. This 'common sense' view was both known and taught by Aristotle. In fact modern car manufacturers agree with Aristotle, using bigger engines in cars when they want higher top speeds.

Of course Newton was eventually accepted as being correct in this matter, the problem being that friction had not been taken into consideration in the cart and car situations. Nevertheless, advancing new ideas can be a brave thing to do. The history of science and technology is littered with examples of those explorers of the natural world who merely reported their findings or theories, and were vehemently attacked for it.

If you can show somebody a simple experiment that clearly demonstrates your idea then there is never really a problem. The problems occur when the result of the experiment is subtle, or where the cause/effect relationship is not clear. Worse still is where the only proof of your theory comes from mathematics, particularly when this mathematics is of a very advanced nature. And worst yet is where you have to convince somebody to fund your research to develop the theory, in order to devise an experiment, for which even more money will be required!

How not to proceed/succeed

Consider the case of Oliver Heaviside and long distance telegraphy circa 1887. Heaviside argued that inductance in the wires was not only a good thing, but that more should be added. William Preece, Electrician to the Post Office, had the opposite view. Now Preece was a very important man and was knighted for his services. Heaviside, on the other hand, whilst respected by a select group of scientists, was poor all of his life. Guess who prevailed at the time! It was not until 1899 that Heaviside's additional inductance idea was used, although the patent was taken out by Pupin; a patent which incidentally earned Pupin some $1,000,000!

Heaviside's problems were that he was right, but unable to convince anyone of his rightness, and equally unable to profit from his ideas. Nowadays we would say that he lacked people skills and financial acumen! One possible reason why he failed to convince people is that he had developed his own technical notation. This notation superseded the ghastly notation used by Maxwell, and is in fact the notation still used today. However, in order to understand what he was trying to explain, one first had to understand the terminology being used. This is still a large hurdle in many cases.

Another possible reason for his failure was his personal attacks on Preece, a well respected and important person. Heaviside's public condemnation of Preece's theories and experimental work was so public, and so frequent, that it made it difficult for Heaviside's papers to get published. Thus this anti-Preece campaign was actually counter-productive in the long run.

Modern Views

In modern science, no physical theory would be considered as validated without measurements. These measurements should preferably be reproducible by other labs, and in order to do so, the experimental requirements have to be given in sufficient detail that duplication is possible.

One difficulty is that the complexity of modern experiments can be so great that the results are merely 'consistent with the theory', rather than proving it beyond any reasonable doubt.

There was an experiment, reported in 1977, of an anisotropy in the cosmic microwave background radiation, measured by flying a U2 plane around in the upper atmosphere. The anisotropy meant the radiation was not the same in all directions. Anisotropy of the 4°K cosmic radiation suggests what might pictorially be described as "aether drift". But when you then read that the amount of anisotropy amounted to less than 0.004°K you have to wonder if the experimental conditions were sufficiently perfect for this to be a genuine result. One thing is for sure: very few scientists would have the interest to repeat this experiment, and even fewer would have the budget!

Any new theory has to have sufficient financial potential that somebody is willing to spend the time, and therefore the money, to reproduce the verification experiment. Remember that funding for just reproducing somebody else's work is not an easy thing to obtain.

Consider this simple question: Is the alternator in your car working properly? To test this question experimentally you might set the car idling with the headlights on, measuring the battery voltage with a hand-held DVM. If the voltage were 13.0V then you would know that the alternator or charging circuit was faulty. If the voltage were 14.0V then the alternator would be declared satisfactory. Of course in such an experiment one would first check the calibration of the DVM being used, avoiding unnecessary costly repairs on the one hand, or possible breakdown on the other. Nevertheless the experiment is simple enough to be convincing on its own.

Compare this with the question: Does a pointed lightning conductor attract lightning to a building? Experimental work in this field was (historically) so difficult that huge and heated controversies on the size, shape and deployment of lightning conductors existed for not just decades, but centuries.

Such controversy is easily defused by a good experiment. When such unpleasantness is encountered, both warring factions should seek a resolution in terms of definitive experiments, rather than continued personal mudslinging. This is the difference between scientific subjects, such as engineering, and non-scientific subjects such as art. Nobody will ever be able to devise an uglyometer to quantify the artistic merits of a painting, for example.

Compare this view with the contents of one particular website, for which further advertising is not warranted. This site has, amongst its simple introductory material, a description of a theory about the phase relationships between the electric and magnetic fields within a photon. Now understand that any convincing portrayal of an internal structure of the photon would undoubtedly attract a Nobel Prize in Physics for its author.

On the site in question, the photon's structure is presented with no experimental evidence, but just some simple, and from my point of view unconvincing, arguments. The author then asserts that he must therefore be right, and challenges all comers to "prove him wrong". Then, having been ignored for some time, the lack of refutation somehow means that the theory must be right as nobody has 'successfully' challenged it — according to rules laid down by the site's author.

This self-publishing of non-reviewed material is a recent 'benefit' of the internet for which there is no historic precedent. Ordinarily one would at least have to convince an editor that what was being said had some sort of merit!

Pseudo-experimentation

Pseudo-scientists conduct experiments, some of which may be cheap and simple, but for which controls are not used, and for which the results are not measured in any objective sense. Consider "pyramid power". People have proposed that placing used razor blade at the centre of a pyramid sharpens the blade. No doubt many people could be convinced that the blade is sharper when left over-night in the pyramid.

The trouble is that the sharpness of the blade is measured subjectively: the experimenter shaves with it and agrees that it is sharper. The first step is a control, use two blades: one is left in the pyramid and one is left somewhere outside the pyramid. The next step is the double blind. The person conducting the tests, interacting with the subjects, and recording the results, must not know which is the 'pyramid-enhanced' blade and which is the control. This process is necessary to prevent personal prejudice (bias) affecting the results.

Furthermore any statistical analysis of results should ideally be agreed upon before the results are obtained. It is all too easy to inadvertently draw a "best fit" line through noisy data points and get the slope or curvature desired. Thus data analysis should ideally be done only by computer, using a regression algorithm decided beforehand. Picking the 'correct' regression algorithm can itself bias the results of an experiment.

All this sort of experimental technique is necessary when the experiments do not produce hard evidence. Results of a statistical nature are much more open to interpretation and personal bias.

It is important to point out that just declaring a theory foolish because it violates known scientific principles is not necessarily good science. If one is only allowed to check for actions that agree with known scientific principles, then how can any new scientific principles be discovered? In this respect, Einstein's popularisation of the Gedankenexperiment (thought-experiment) is potentially a backward step.

Failures of experimental technique are however a valid and important target of criticism. If a piece of equipment is said to fail a two-tone intermodulation test, for example, the first thing I would want to know is what test equipment was used, and how was it established that the intermodulation was not being produced just by the action of summing the two signal generator outputs. It is actually very easy for intermodulation to be produced when two signal generator outputs are summed.

Of course being a stickler for accuracy, even if the equipment passed the two-tone test, I would still want to know what effect summing the generators together produced by way of intermodulation products. The reason is obvious: it is quite possible for the intermodulation tones created by summing the generator outputs to null against the tones created in the equipment under test, thereby giving a false pass!

Pseudo-science?

One has to wonder whether present day hi-fi systems constitute engineering or pseudo-science. Certainly there is heated debate on the subject, some of which gets very personal. It is therefore a brave engineer who tries to advance any ideas in this field, facing the wrath of those outspoken 'golden eared' individuals who know better, or at least who claim to. One might hope that with the raft of measurements available to the test engineer, golden ears would no longer be either needed or wanted, and yet this does not seem to be the case.

It is demonstrably true that harmonic distortion tests alone are inadequate to quantify the performance of an audio amplifier, but you would think that a specification involving THD+N, SFDR, frequency response flatness, 1dB bandwidth, noise, hum, pulse fidelity, output level for 0.1dB compression, input and output impedance (versus frequency), input CMRR (versus frequency), and two-tone intermodulation distortion, (and any I may have missed) would quantify an amplifier sufficiently to make testing an objective rather than a subjective process.

Now if hi-fi generates so much passionate debate, imagine the problem when dealing with atomic phenomena. The measurements made are inherently difficult and expensive. Even the most brilliant scientist could not 'knock up' a superconducting supercollider in their garden shed to experiment with. Thus important but expensive scientific areas can only be dealt with by institutions using funding extracted by excellent PR work.

Consider the case of beta particle radiation. We have a nucleus sitting there, minding its own business, and then, for no explicable reason, the nucleus decides to rearrange itself, ejecting a beta particle in the process. Now a beta particle is an electron, so one has to wonder what an electron was doing in the nucleus in the first place. But in the second place, this electron can be ejected at a speed of something like 90% of the speed of light. The more you think about this, the more you realise that it is a mind boggling concept.

Whilst an atomic radius is of the order of 200pm (2Å), a nuclear radius is of the order of 10fm, some twenty thousand times smaller. It is then left to the imagination of the reader as to how a supposedly stationary electron, which probably didn't exist in the nucleus beforehand, suddenly coalesces out of the quark plasma (forming the nucleus) and accelerates from 0 to 0.9c in the space of less than 20fm.

In fact if you look too deeply into modern physics, with zero-point fluctuations, spontaneous pair production, superconductivity based on Cooper pairs of electrons many atomic radii distant from each other, Hawking radiation from black holes, superstrings occupying eleven dimensional space, quark binding force increasing with separation distance, and so forth, you may be glad that you took up electronics rather than this 'occult subject' known as modern physics!

In this modern world, big experiments are seldom done because of 'interest' in the sense of scientific curiosity; the people paying for the research want immediate material benefits from their investments. This is unfortunate; since the history of technology shows that scientific curiosity opens up the major advances in technology.

As a modern professional engineer you should want to know certain things about any new scientific phenomenon. How can you measure it? How can you improve or refine it? What can you make using it? How much money is it worth? This may sound cynical, but it is really a question of whether engineering is a profession or just a hobby for you.

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LETTERS to the editor

Letters to "Electronics World" Highbury Business, Media House,
Azalea Drive, Swanley, Kent, BR8 8HU
e-mail EWletters@highburybiz.com using subject heading 'Letters'.

No amps for sale

In the May 2004 issue of *EW*, D Lucas stated that I "build and sell nil distortion amplifiers". This is not the case. I design products for one of the more significant companies in the audio business, and they would be less than pleased if they thought that I was running a cottage industry making amplifiers in my spare time.

I should therefore like to make it very clear that I do not sell amplifiers, PCBs, or anything else pertaining to my designs. However, PCBs approved by me (many years ago) and kits of parts can be obtained from Gareth Connor at www.signaltransfer.freeuk.com

I also I read with great interest Ian Hickman's article on spam.

Interestingly, I have received virtually no spam since I moved to a broadband service provider, though why that should be I can't imagine, as I am certainly no less active on the net.

I can offer a solution to the problem of putting your email address on your website.(some ISPs insist you do this, presumably so outraged surfers can complain directly, but I have a feeling this is not really enforced) The answer is- don't. Instead show a picture of the email address, as a small GIF or whatever. Don't make it a clickable link to your address, as that puts the address right back in the HTML, where the web-crawlers will nab it.

This technique requires potential correspondents to write the address down and type it in, but that should not discourage anyone with real motivation to contact you. For an example see my own modest offering at www.dself.dsl.pipex.com

Until the spammers come up with character-recognition software, this method appears to be pretty safe. When they do, well, there's always the Fraktur Gothic font...

Douglas Self
By email

Intelligent electric fence

Living in rural New South Wales electric fences are common for confining stock and fencer repair is one of my constants. These units look like a capacitor discharge Ignition system with 2-300Vdc inverter and SCR dump into an ignition-like transformer.

I can't quote the figure, but the safe energy output of a fence is defined in Joules by an Australian Standard. (0.1μF x 300V typical). Most will throw a fat pink spark over 10mm.

I was gobsmacked when I realised that this unit is simply the active side of the mains. This is totally illegal in Australia and stupid and very dangerous everywhere, even Kerala, India.

In fact this arrangement is defined as an illegal 'mantrap' here if used for 'homes and security' and I know of several cases where people who have used this method are now cooling their heels in gaol.

In one memorable incident a policeman suffered burns to 'the groin area' after he grabbed an electrified guttering while climbing over a tin fence during a drug raid. We may laugh but the occupier copped a double-whammy in court.

In another case a child was killed after contacting such an arrangement put up by a paranoid neighbour to prevent housebreaking - not a lawful excuse. If I were you, I wouldn't have published it.

By an odd co-incidence I have followed a very similar trajectory to Mike Arnold (April letters) having finally completed a counter using TTL and 'numitrons' (filament 7-segment displays) started around 1972 and finished in 2002.

Roly

April fools

Messrs Carri and Aylward's letters about my hybrid amp article show the truth of the adage, "a little knowledge is a dangerous thing". What seems to have upset them both is that I have the temerity to use a valve to improve the performance of a transistor amplifying stage instead of vice versa. Well I make no apologies for it. It certainly performs better than an ECC83 stage although I find those valves microphonic and overrated. A guitar amp is the best place for them!

If Mr Hawkins thinks that the valve performs no function in the circuit I can assure him the circuit stops working when it's removed. The argument is as vacuous as suggesting that the collector resistor or constant current source is not required in a common emitter amplifying stage.

Turning to Mr Aylward's comments, the design originally used a current source in the collector load of Q1. However this is a single ended amplifier. Simply substituting the current source for the valve results in 10A+ current surges through the output capacitor when the amp is switched on. By the time a slow start circuit has been added the circuit becomes very complex. Worse, slow start circuits don't appear to be very robust. The use of a valve as a constant current source for this part of the circuit gives built in slow start every time without compromising the high impedance seen by Q1's collector. Actually I put more credence for the good performance of the circuit on the fact that I used shunt feedback. I agree with JLH this configuration sounds a lot better than the series connected circuit. Probably because the common mode voltage of the circuit is zero.

On the subject of valve sound I submit that it is possible to get the same sound from any solid state amp by the simple expedient of adding a resistor, 3.9 Ohms will do, in series with your loudspeaker. This will reduce damping on the speakers providing bass lift around the speaker's resonant frequency. More importantly it will also produce a small lift at the crossover frequency of the speaker where its impedance peaks. This will impart a pleasing although inaccurate presence to vocals.

Hi output impedance is a property of valve amplifiers especially the much vaunted single ended power amp. Of course if you replace the speaker system with a load resistor you will measure a reassuringly flat frequency response! The best sounding amplifiers don't have a sound which is characteristic either of valves or transistors, they are simply transparent. Although it is a conceit, I like to think that this can only be achieved by good design. Using appropriate devices and circuit topologies. Often getting the sound right seems to be a matter of perspiration and luck.

Jeff Mcaulay
Chichester
Sussex
UK

Master and slave

With reference to the article 'Archiving' in the letters section of the may 2004 issue of *Electronics World.*

Perhaps Master and Minion might be more appropriate when say describing functions such as main and mere defendant, of parts of apparatus or devices.

Best wishes to *Electronics World.*

R.C.T. Stead
Hampton
Middlesex
UK

Colossus

Having just received my latest copy of *Electronics World* I was very interested to read the article on the Colossus Computer. I was particularly interested in the bit about leaving the valve heaters on and switching the HT off.

Back in 1959 I was working in the R&D department of Powers-Samas Accounting Machines at Whyteleafe in Surrey. Powers-Samas was a well-known company in the Croydon area for the manufacture of punched card accounting machines.

I was part of a team given the task of improving the reliability of a computer known as the PCC then being manufactured at the Crayford factory of Vickers(Eng). One of the problems was the short life of some of the valves. This machine was also being operated with heaters on but no HT.

Samples of the failed valves were sent to Mullard who discovered that the coating of the cathodes was being destroyed and put this down to cathode poisoning by the electrons that were not being attracted to the anodes.

I would be interested to know if the team at Bletchley Park know of this effect and have had similar problems.

D.F.W Jones IEng MIIE
Westerham
Kent,
UK

Colossus... computer

Dear friends, Colossus was not a "computer" according to the definition of a computer = a computing device governed by a stored program (very important restriction!).

Colossus (1943) was indeed the first electronic calculator, built for one specific application only, but not a computer. After Colossus, there came (Howard Aiken, Harvard 1944) Mark I, also an electronic calculator, but not totally electronical, as it used relays, too; then IBM ASCC (Automatic Sequence Controlled Calculator, 1944). The first computer (with stored program) was actually Z4 (Kondrad Zuse, Germany, Neukirchen 1945), which appeared before ENIAC (Electronic Numerical Integrator And Calculator, 1946), which is usually accepted as the first real computer. Zuse made his first calculators as early as 1938/1939 (Z1) with relays!

E.P. Mänd
Helsinki
Finland.

Blown fuse indicator

I write about the circuit idea 'Blown Fuse Indicator' in the May issue. I cannot see how the circuit can work as described.

I must assume, due to no markings on the circuit, that the lower horizontal line is the negative supply line. If so, when the fuse is in tact the green LED will not glow as it is reverse biased. The red LED will glow when the fuse blows.

If the diagram is wrongly drawn and the green LED should be the other way around, both LEDs will be grossly overloaded as they are in series across the supply with no current limiting resistance.

Also, it will not work with shunt diodes across the LEDs with an AC supply, but will work with series diodes to protect the LEDs.

Marcus Wilson
Wellington
New Zealand

Colossus IS a computer

Ivor Catt's article (Boolean castles in the air, July 2004) says: "I have checked back to find that circuitry was so expensive and small in number that machines like Colossus had virtually no logic design content" (middle column, page 29). This is not so, Colossus has a very large amount of logic, mainly equivalence/non-equivalence in the form of XNOR gates, implemented using pentodes. The other logic used, also in large amounts is NOT, OR, NOR, AND, NAND. The machine does up to 100 logical comparisons on data from the paper tape every 200µS, mechanical devices could not do that. Triodes are only used as cathode following buffers. The whole purpose of Colossus was to break the German Lorenz code, no expense was spared. Even if logic (including XOR) had been ten times the price it was in 1943, they would still have used it. In fact Colossus was never intended to be a "Computer", it was a tool to help some very clever mathematicians implement some very clever techniques. In 1943 the term Computer meant a person sitting at a desk calculating and performing logic. It so happens that Colossus now fits the name Computer, a name that was devised after it was. I invite Ivor to come and look at the circuit diagrams for Colossus and admire the rebuild.

Charles Coultas
charles@dataskil.com

Reflections

The statement by J. S. Linfoot (*EW* Letters, June 2004) that reflected power may be absorbed by the source needs to be treated with care. Certainly, as a National Service trainee radar technician in the 1950s, I was taught that power reflected from the load is absorbed in the source, but this needs qualification. Before deciding what happens to the power, it is necessary to consider what load the source sees, and what are the resultant currents and voltages. For example, in the case of a 1Ohm Thevenin source delivering 1W to a matched 1ohm load, there is additionally 1W internal dissipation in the source, whose open circuit terminal EMF would be 2V. If all of the power is reflected due to a total mismatch, viz. a short or open circuit seen at the source terminals, the internal dissipation in the source will be 4W or zero respectively, in the case of the usual type of generator (Thevenin), or zero and 4W respectively in the case of a Norton generator. In either case, the internal dissipation is not equal to the 1W internal dissipation of the matched case plus the reflected 1W.

There is however, a class of generators where the total dissipation is independent of the degree of mismatch of the load. A 1W Hickman Type I generator always results in 2W total dissipation, whether 1W in the load and 1W internal, or 0W in the load and 2W internal, or anywhere in between. Thus it may truly be said to absorb any reflected power completely. It consists of the combination of a Thevenin and a Norton generator, and full details can be found in my Design Brief article 'RF reflections', *Electronics World & Wireless World*, October 1993, pp 872 – 876. It also describes the Hickman type II generator, which behaves in the same way.

Ian Hickman
By email

The mystery of magnetic lines of force

I have been involved in several discussions with astronomers about their perceptions of magnetic lines of force.

These magnetic lines of force are alleged to have physical properties that are used to explain many astronomical phenomena. The Earth's Dynamo theory is based on the actions of these lines and sunspots are alleged to be created when these lines get "tangled". Some astronomical sources claim that ionised particles can travel along these lines while others claim that ionised particles spiral around them.

In the 40-odd years of reading *WW* and its successors I cannot remember reading anything about any device that can detect magnetic lines of force. I do not believe that magnetic lines of force exist any more than contour lines or isobars exist. However, I am prepared to admit that there is a gap in my knowledge if a magnetic field line detector exists.

Is there such a device? I know that magnetic field strengths can be measured with a field strength meter as easily as atmospheric pressure can be measured with a barometer. Yet these measuring devices give no indication of where magnetic lines or isobars are physically located - if they exist.

The classic demonstration of the shape of a magnetic field with iron filings on a card placed over a magnet will show apparent lines. If this demonstration is repeated many times with the same magnet and fresh filings, it will be noted that the apparent lines never occur in the same places twice. In other words, iron filings do not indicate the presence of stationary magnetic field lines.

Can any *EW* reader offer some enlightenment on this subject?

Wilf James
By email

JLH info

I was very sad to read in the May 2004 issue of *Electronics World* that JLH has passed away. I have been enjoying reading his professional but always very practical articles over the years in various magazines.

In the same issue there was an interesting schematic of JLH's distortion monitor that I was not previously aware of. Instead, I have built his portable distortion meter (with a test signal) published in the October 1982 issue of HFN/RR, please see a photo attached. This meter utilises opamps and has a very good performance; I used it in the 80s to compare the distortion levels produced by my Teac X-1000 3-head open-reel tape recorder with various tapes, biases, etc.

I also attach a useful web address which has lots of information on JLH designs including a valuable list of his published articles: http://www.tcaas.btinternet.co.uk.

Sauli Palo
Helsinki
Finland

Not me guv

Whilst Ivor Catt's thesis, "Not me, guv" (*EW* May 2004) deserves credit for drawing attention to a potential EMC hazard, there are practical reasons why, in the majority of cases, the problem described is of little consequence.

In the third paragraph of the article, measurements of the inrush current of a transformer due to magnetic remanence of the core are quoted as 'some 100A'. The article does not give the context of the measurement (i.e., characteristics of the transformer measured). The majority of equipment using the circuit of figure 2 in the article will have significant transformer primary winding resistance. For example, a 100VA transformer will have a primary resistance of the order of 40 ohms so that, on a 230Vac line, the inrush current would be limited to 8A if the supply were turned on at the peak. It would seem therefore that the hazard described would only arise for transformers of above, say, 1kVA.

Finally, good practice in power supply design can include a voltage dependant resistor in series with the transformer primary where inrush current is likely to be a hazard. A 4 ohm resistor might not be appropriate in all cases.

Quite apart from any "higher mathematics", another reason why the problem does not appear in EMC tests is because the testing methodology imposes a LISN network (having its own impedance) between the power source and the EUT (equipment under test). This is done in order to isolate the EUT from supply-borne interference so that interference generators can be applied in a controlled manner. Like Mr Catt, I am only too aware that EMC testing is overly concerned with the steady state. Furthermore, the statutory tests are restricted to a common range of interference modes which have been successful in improving equipment quality. However, for a particular equipment design there may be other interference modes that statutory testing does not reveal.

In my experience, too little design attention is given to 'real world' malfunctions arising from turn-off or turn-on transients. Before EMC testing became a discipline of its own - with exotic and expensive test gear, I used to wrap the mains lead of an (un-suppressed) electric drill around the EUT. If it survived that, it rarely went wrong in the field.

Ray Penny
Arundel
West Sussex
UK

Ivor question answered

Ivor must be referring to XOR.

My boss, Jon, used to pose the problem at interview to prospective employees of how to swap the contents of two registers without any other change of state.

Solution:
XOR A,B; Destination on the right BTW
XOR B,A
XOR A,B
Cute.

Paul Bartlett
By email

Less Catt please

I felt that I should write to express my disappointment at seeing yet another article in your magazine by Ivor Catt. Like many I feel that his work is insulting, ill informed, badly written rubbish by someone who clearly hasn't worked for the past fifty years and has the writing style of someone stuck in the mid 50's. For example I read that Ivor Catt has moved on from insulting anyone who has dealt with EMC issues and now has problems with Boolean algebra and what he calls "Oxford logic" and he doesn't believe that you need OR gates in logic design or some such nonsense. I would also add that if you are going to quote references to support some argument its best not to quote your own work, as he tends to do otherwise it looks as if you are agreeing with yourself.

I see by your response to such comments, labeling them as a flap, that you feel such views do not matter. I for one will not be buying another copy of Electronics World as it has been reduced from a magazine for working engineers to Ivor Catts mouthpiece for his ill-informed views.

Unsigned
By email

Well over half of the letters I get on the subject of Ivor Catt, either support him or at least like his style. As editor, I have to go with the majority vote. – Ed.

More Catt 'history'

I am always interested and often amused by Ivor Catt's articles and letters. This time (Boolean Castles in the air) he approaches rather closer than normal to some of my specific interests and so I feel I ought to comment on a couple of his assertions.

The Ferranti Atlas and the IBM 7030 (Stretch) were indeed two of the earliest supercomputers. Both were foundations of much of the technology that appeared in later computers. Stretch is normally considered a failure in spite of being the world's fastest computer from 1961-1965 because only nine systems were built and they were mostly sold at a loss. Atlas is normally considered a success yet only three systems were built (+ three more "cut down" and slower Atlas II systems). In both cases counts of systems sold are the official figures and there might have been one or two more sold to the "spooks" (NSA/GCHQ or their predecessors). Fans of the Atlas had, perhaps, the last laugh when in 1970 the IBM system 370 was launched with the slogan "IBM announces tomorrow" when the main new feature was the virtual memory pioneered in Atlas.

I am inclined to leave the defence of mathematical logic from Catt's serious misunderstandings of its nature to others with greater skill. What he thinks of as "logic" is "Boolean Logic" – a very small part of a much larger subject.

However, I will come back in on the Exclusive Or. When I read his original letter I wondered what he was on about - it couldn't, surely, be our old friend the XOR, could it? Now I see from his article that it is precisely that to which he refers. The reason the XOR isn't given the prominence he feels it deserves is quite simple - it is not a fundamental operation but can be built-up from AND/OR/NOT (or in fact from either NOR or NAND) whereas it cannot be itself used as a building-block to produce one of these more fundamental operations. In the logic design of older computers XOR is most often found as part of the "half-adder" (the XOR produces the "sum" bit; an AND gate produces the "carry").

Andy Holt
Rayleigh
Essex
UK

Capacitance meter

As the author of the article on the construction of a Capacitance Meter (EW March, 2004). I read with interest Paolo Palazzi's letter published in EW for July.

I think his theoretical comments on excessive current flow are correct and it would appear that the CMOS 40106 IC is being required to supply current well beyond its means.

However it would seem that Mr. Palazzo may have forgotten that the first gate of the 40106 is behaving much more like a linear amplifier with the unknown capacitor minimally charging and discharging about 6 volts. However, even with the highest value capacitors sufficient current is available for gate N1 to provide a waveform able to reach the trigger points of the paralleled buffer gates N2 and N3. They produce an exact copy of the frequency of gate N1, but now rail-to-rail: nothing is under electrical stress and nothing heats up. The slight drift in component value that he refers to is not due to heating but as I said in my article, to the capacitor reforming.

David Ponting
Bristol
UK

Buzz Lightyear

Infinity has been defined by Leitz (the famous optical company) and here I quote from 'Leica M The Advanced School of Photography'

"Leitz applies the largest number that can be utilized in the computer, namely: (infinity) =
1,701,411,834,605 x 10^35 m"

Huw Finney
Maidenhead
UK

The Catt question – *EW* challenge

Two letters in the July issue prompt me to take a major initiative, and to relegate my replies to the www in the hope that our editor will be able to publish them later. For the present, they are at the bottom of www.ivorcatt.com/44.htm

Regarding "The Catt Question", *EW* reader Alan Robinson favours the 'Westerner' answer of Dr Neil McEwan, Reader in Electromagnetism, while the second letter writer Ian Darney favours the 'Southerner' answer of Pepper FRS. They reinforce our concern that the question is unresolved.

For many years, for instance in the pages of *Electronics World*, I have offered a large sum of money to any student who succeeds in getting his accredited lecturer or text book writer to write a signed, dated document discussing "The Catt Question", to no avail. Now the offer is put on a firmer footing.

I have opened a bank account containing £2,000 which our Editor can access. This is the prize behind the *EW* challenge.

I suggest to the editor that 'accredited' should mean Reader in Electromagnetism, author of published text book from mainstream publisher, or similar. Professor of Electronic Engineering will do, but not Professor of Electrical Engineering.

My advice to the Editor is that the first student who succeeds in getting his accredited lecturer or text book writer to write, sign and date meaningful, intelligible comment on 'The Catt Question' should receive the first prize of £500. The next ten successful students should receive £100 each.

A student who provides documentary proof that his accredited lecturer or text book writer refused to make meaningful comment should receive £50.

Prizes should be paid out until the money runs out. Any third party contributions will be added to the fund.

In due course of time, all written, signed, dated comment (and all proof of refusal to comment) will be reached via my website at www.ivorcatt.com/44.htm.

Only written comment the whole of which our editor understands can receive a prize. The challenge is open to the whole world. All comments must be in English.

The decision of the Editor is final.

Ivor Catt
St. Albans
Hertfordshire
UK

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John Crisp

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Feb 2000 ▲ 302 pages ▲ Glossary ▲ Index
PB ▲ Published in UK
Code 0-7506-4583-0 **£18.99**

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S W Amos; R S Amos

Aimed at engineers, technicians and students working in the field of electronics, this dictionary provides clear and concise definitions, including TV, radio and computing terms, with illustrations and circuit diagrams.

4th edition ▲ Mar 2002 ▲ 394 pages
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Eugene Trundle

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3rd edition ▲ Feb 2001 ▲ 432 pages ▲ Index
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Richard Brice

Covering all aspects of digital television, this text encompasses the electronics of the equipment, data compression, television production, servicing and the different transition methods - terrestrial, satellite and cable. The text has been updated with developments since the 2000 edition.

2nd edition ▲ Oct 2002 ▲ 304 pages ▲ Index
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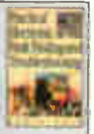

Robin Pain (Design Engineer, Cotag International Ltd)

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Mac E Van Valkenburg; Edited by Wendy Middleton

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9th edition ▲ Aug 2001
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Farid Dowla (Lawrence Livermore National Laboratory, University of California-Davis, USA)

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November 2003 ▲ 720 pages ▲ HB

Code 0-7506-7695-7 **£59.99**

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John Watkinson, Francis Rumsey

A detailed manual for those who need to get to grips with digital audio and video systems. It sheds light on the differences between audio interfaces and shows how to make devices "talk to each other" in the digital domain. Also includes detailed coverage of all the regularly used digital video interfaces. New information included in this third edition: dedicated audio interfaces, audio over computer network interfaces and revised material on practical audio interfacing and synchronisation.

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Edited by Peter Marlow

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January 2004 ▲ PB & CD-Rom ▲ 352 pages

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Steve Beeching

A comprehensive guide to domestic VCR technology and repair techniques. This edition brings the information fully-up-to-date, with expanded coverage of camcorders, sections on DVD equipment and the latest VCR technology.

5th edition ▲ Apr 2001 ▲ 323 pages Illustrations ▲ PB ▲ Published in UK

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Keith Jack

This edition has been updated to include information on digital television, datacasting, interactive video, digital camcorders and VCRs, and video interfacing. Coverage is international, including European, Asian and North/South American video standards, methods and techniques.

3rd edition ▲ Jul 2001 ▲ 784 pages & CD-Rom References ▲ Glossary ▲ Index ▲ PB Published in UK

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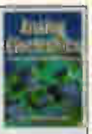

D Crecraft (formerly Open University); S Gergely (formerly University of Coventry)

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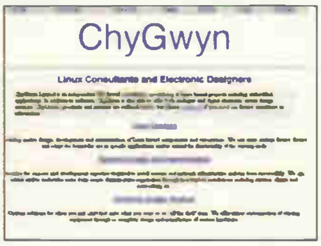

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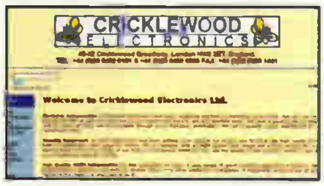

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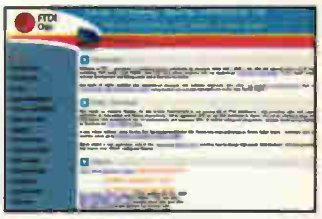

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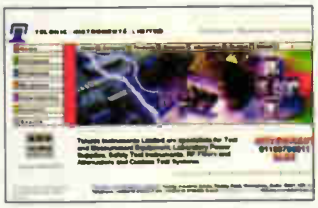

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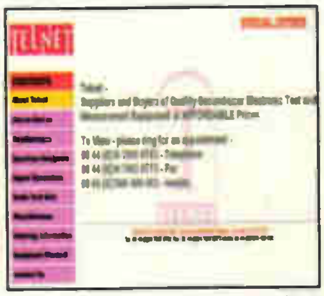

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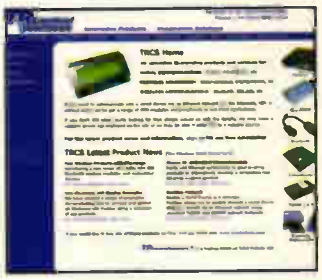

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Amplifier Research 10W1000B 1GHz 10W RF Amplifier	2500	76
Amplifier Research 1W1000 1GHz 1W RF Amplifier	950	40
Kalmus KMS737LC 25W 10kHz-1GHz Amplifier	4500	136
CALIBRATORS		
AT/HP 58503B/001 GPS Time & Frequency Ref Receiver	1950	78
Fluke 5500A/SC300 Calibrator With 300MHz Scope Option	11750	466
Fluke 720A Kelvin-Varley Divider	975	49
Keithley 220 Programmable Current Source	2450	103
Keithley 224 Programmable Current Source	2250	93
COMPONENT ANALYSERS		
AT/HP 16097A Accessory Kit For HP4192A	950	31
AT/HP 4192A 13MHz Impedance Analyser	4350	218
AT/HP 4193A 110MHz Impedance Analyser	3550	165
AT/HP 4194A 40MHz Impedance Analyser	15950	635
AT/HP 4285A 30MHz Precision LCR Meter	7200	309
AT/HP 4294A 110MHz Precision Impedance Analyser	22595	927
AT/HP 4339A High Resistance Meter	2450	125
AT/HP 4815A Vector Impedance Meter 500kHz-108MHz	1150	57
ELECTRICAL POWER		
BMI A-116 1-600A Current Clamp For BMI 4800/100G	650	33
Dranetz PP4300 Power Quality Analyser	4950	179
HT Italia SPEEDTEST RCD Test Set	400	24
Kikusui TOS8850A 5KV Insulation Tester	2550	100
FREQUENCY COUNTERS		
AT/HP 53131A/030 3GHz Universal Counter	1350	72
AT/HP 53132A/030 3GHz 12 Digit Frequency Counter	1650	50
AT/HP 5335A 200MHz Frequency Counter	950	40
AT/HP 5350B 20GHz Frequency Counter	1650	75
AT/HP 5351B 26.5GHz Frequency Counter	2200	95
AT/HP 5372A 500MHz Frequency/Time Interval Analyser	2650	100
AT/HP 5386A/030 3GHz Frequency Counter	1450	68
EIP 548A 26.5GHz Counter	1575	68
Racal 1992 1.3GHz Frequency Counter	1150	35
FUNCTION GENERATORS		
AT/HP 3245/001 DC-1MHz Function Generator	3650	163
AT/HP 3314A/001 20MHz Function Generator	1335	56
AT/HP 3325B 21MHz Function Generator	2250	68
AT/HP 8111A 20MHz Function Generator	1250	50
AT/HP 8116A 50MHz Function Generator	1895	70
AT/HP 8904A/001/002/003/004 600kHz Function Generator	2950	91
MULTIMETERS		
AT/HP 34401A 6.5 Digit Digital Multimeter	650	33
Keithley 2001 7.5 Digit Digital Multimeter	2150	99
Keithley 2400 200V Digital Sourcemeter	2995	127
NETWORK ANALYSERS		
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AT/HP 35689A 150MHz 50 Ohm S-parameter Test Set	1500	45
AT/HP 3577A 5Hz-200MHz Vector Network Analyser	4750	142
AT/HP 3589A 150MHz Network/Spectrum Analyser	5450	164
AT/HP 8714ET 3GHz Vector Network Analyser c/w TR	8950	351
AT/HP 8714ET/1E1 3GHz Vector Network Analyser c/w TR	9950	307
AT/HP 8722C 40GHz Vector Network Analyser	33500	1232
AT/HP 8753D 3GHz Vector Network Analyser c/w S Param	9750	351
AT/HP 8753D/1D5 3GHz Vector Network Ana c/w S Param	10250	369
AT/HP 8753ES/006 6GHz Vector Network Ana c/w S Param	16250	506
AT/HP 89441A-Various option sets avail - Call - prices from	11950	406
Anritsu 37369C/10A 40MHz-40GHz Vector Network Analyser	34950	1358
Anritsu MS4624B 9GHz Vector Network Analyser	18750	755
Anritsu MS4630B/010 10Hz-300MHz Network Analyser	5250	189
OSCILLOSCOPES		
AT/HP 54540A 4 Channel 500MHz 2GS/s Digitising Scope	5250	211
AT/HP 54645D/101 2 Ch 100MHz 200MS/s + 16 Ch LA	2350	71
AT/HP 54825A 4 Channel 500MHz 2GS/s Digitising Scope	6450	199
AT/HP 54846A 4 Channel 2.25GHz 8GS/s Infinium Scope	12600	511
Lecroy LC584AL 4 Channel 1GHz 1GS/s Digitising Scope	10500	417
Tek TAS485 4 Channel 200MHz Analogue Scope	1350	53
Tek TDS320/14 2 Channel 100MHz 500MS/s Digitising Scope	1500	45
Tek TDS350 2 Channel 200MHz 1GS/s Digitising Scope	1950	07
Tek TDS420A 4 Channel 200MHz 100MS/s Digitising Scope	2250	91
Tek TDS460A 4 Channel 400MHz 100MS/s Digitising Scope	3950	155
Tek TDS540/1M/13/2F 4 Ch 500MHz 1GS/s Digitising Scope	5100	199
Tek TDS640A 4 Channel 500MHz 2GS/s Digitising Scope	3950	166
Tek TDS644B/24/4D 4 Ch 500MHz 2GS/s Digitising Scope	5950	180
Tek TDS694C 4 Channel 3GHz 10GS/s Digitising Scope	14550	571
POWER SUPPLIES		
AT/HP 6030A 200V 17A Power Supply	2300	95
AT/HP 6031A 20V 120A DC GPIB Power Supply	1600	72
AT/HP 6034L 60V 10A Power Supply	1100	47
AT/HP 6038A 60V 10A 200W Power Supply	1250	53
AT/HP 6060B 300W Electronic Load (0-60A 3-60V)	1600	71
AT/HP 6236B +-20V 0.5A and 6V 2.5A Triple Output PSU	350	20
AT/HP 6612C 20V 2A DC Power Supply	600	30
AT/HP 6624A 50V 5A 40W Quad DC PSU	1400	62
AT/HP 6627A Quad 20V 2A Power Supply	1500	70
AT/HP 66312A 20V 2A DC Source	950	40
AT/HP 6632A 20V 5A DC Power Supply	695	30
AT/HP 66332A 20V 5A DC Source	950	40
AT/HP 6634A 100V 1A DC Power Supply	11[illegible]	56
AT/HP 6642A 20V 10A GPIB Power Supply	1250	57
AT/HP 6652A 20V 25A DC GPIB Power Supply	1250	39
AT/HP 6673A 35V 60A DC Power Supply	2250	96
AT/HP 6675A 120V 18A Power Supply	2100	87
AT/HP E3634A 200W Power Supply 25V 7A / 50V 5A	1250	45
Farnell AP70/30 70V 30A Power Supply	950	34
SIGNAL & SPECTRUM ANALYSERS		
Advantest R3361C 9kHz-2.6GHz Spectrum Analyser	6250	268
AT/HP 3585A 40MHz Spectrum Analyser	3950	119
AT/HP 3588A 150MHz Spectrum Analyser	5950	253
AT/HP 53310A 200MHz Modulation Domain Analyser	3350	102
AT/HP 85024A 3GHz Active Probe	1350	41
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AT/HP 8593E/021/027 26.5GHz Spectrum Analyser	14950	539
AT/HP 8594E/041 2.9GHz Spectrum Analyser	4750	143
AT/HP 8594E/140 2.9GHz Spectrum Analyser	4950	149
AT/HP 8594L/041 2.9GHz Spectrum Analyser	3950	11[illegible]
AT/HP 8595E/004/041/105/151/163 6.5GHz Spectrum Ana	9350	201
AT/HP 8901B 1.3GHz Modulation Analyser	1950	02
Anritsu MS2711B 3GHz Handheld Spectrum Analyser	3750	135
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AT/HP E4431B/1E5 2GHz Digital Signal Generator	6150	257
AT/HP E4432A 3GHz Synthesised Signal Generator	6600	256
AT/HP E4432B 250kHz-3GHz Synthesised Signal Generator	7750	308
AT/HP E4432B/1E5/UN5/UN8/UN9/UND 3GHz RF Signal Gen	11250	330
AT/HP E4433A/1E5 250kHz-4GHz Synth Signal Generator	7950	239
AT/HP E4433B/UN5/UN8/UND 4GHz RF Signal Generator	12500	490
AT/HP E4436B 250kHz-3GHz Signal Generator	13950	565
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Marconi 2023 10kHz-1.2GHz Signal Generator	2250	01
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Marconi 2030/001 1.35GHz Synthesised Signal Generator	3300	133
Marconi 2031/002 2.7GHz Synthesised Signal Generator	4500	135
R&S SMH 2GHz Synthesised Signal Generator	4250	193
R&S SMHU58/B6 100kHz-4.32GHz Signal Generator	10350	421
TELECOMS		
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AT/HP 3708A/001 Noise And Interference Test Set	5750	222
AT/HP 8920B/1/4/7/13/14 1GHz Radio Comms Test Set	3950	119
AT/HP 8920B/1/4/7/13/14/51 1GHz Radio Comms Test Set	3950	11[illegible]
IFR 2967 GSM 900/1800/1900 Radio Comms Test Set	5950	245
IFR 54421-003J RF Directional Power Head	250	[illegible]
Marconi 2945/05 Radio Comms Test Set	5950	179
Marconi 2955A/2957A 1GHz Radio Comms Tester c/w AMPS	2500	90
Marconi 2955B 1GHz Radio Comms Test Set	3500	126
Marconi 2955R 1GHz Radio Comms Test Set	2950	107
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