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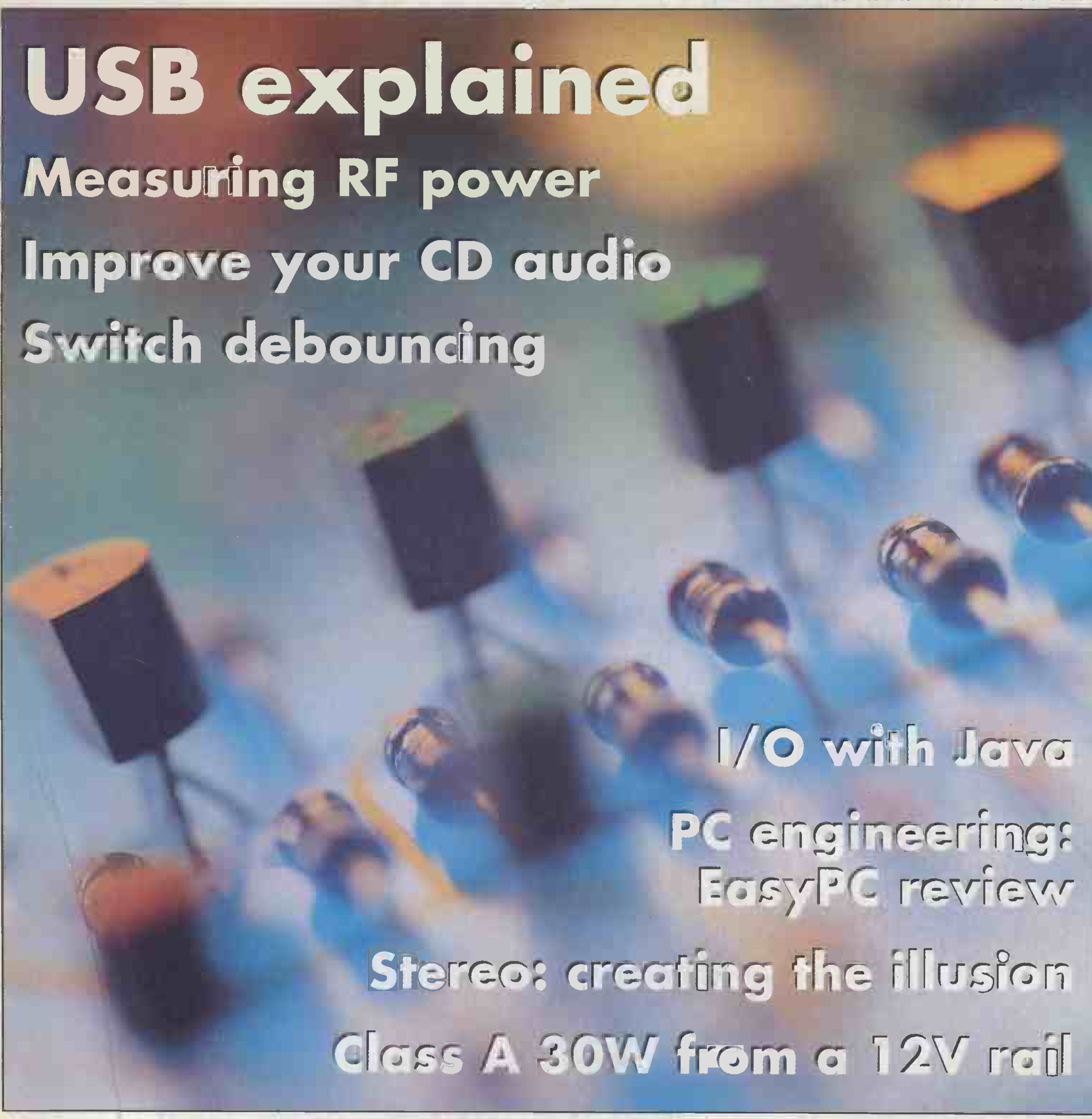

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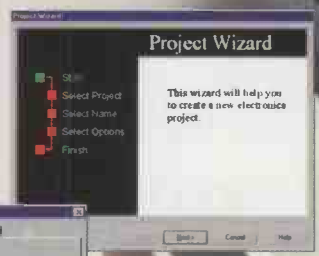

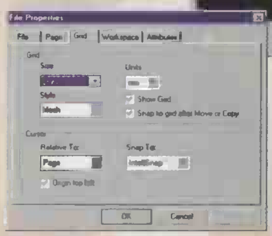

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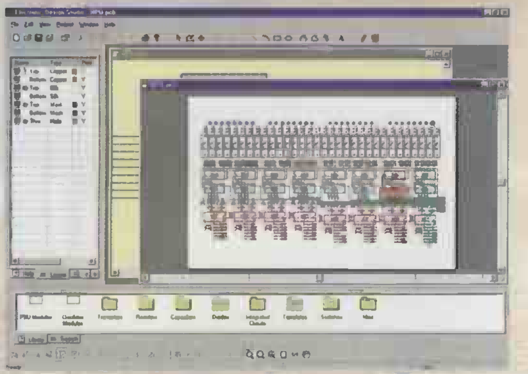

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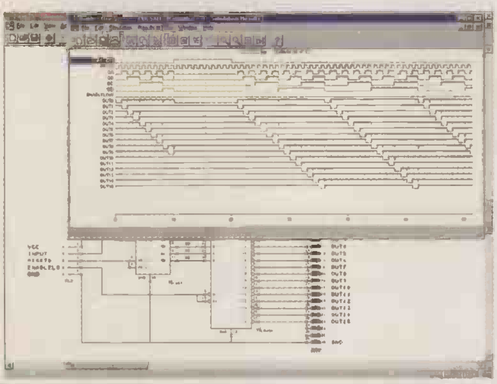
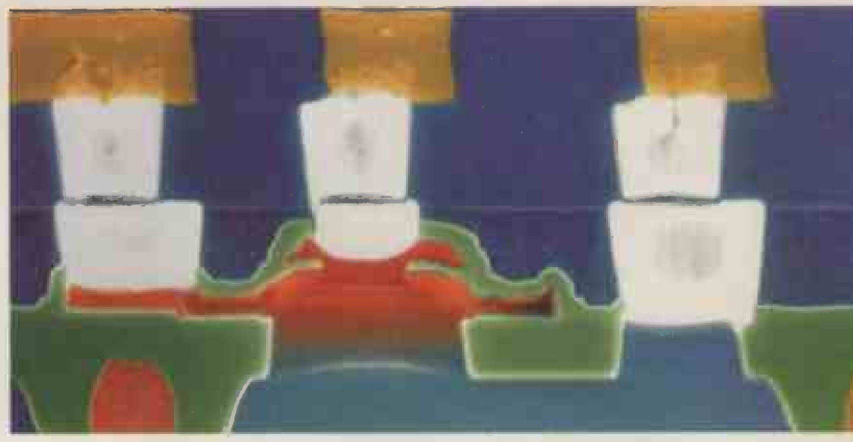
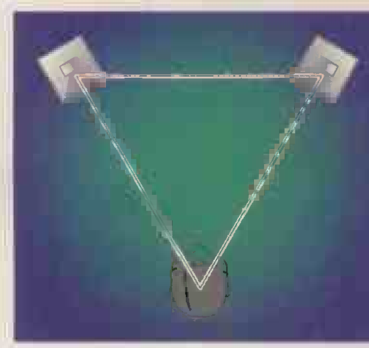


Easy-PC's digital analyser shows not only timing relationships, but also the glitches. Find out what else it can do on page 902.



Silicon-germanium technology is beginning to catch on. But why? It's already been shown that existing technologies are adequate for RF applications? Find out on page 888.



Higher quality hi-fi equipment allows the speakers to be further apart. This, and many other aspects of creating the stereo illusion, are discussed on page 920.

Encapsulated RF coils – one of 25 new products outlined, starting on page 927.

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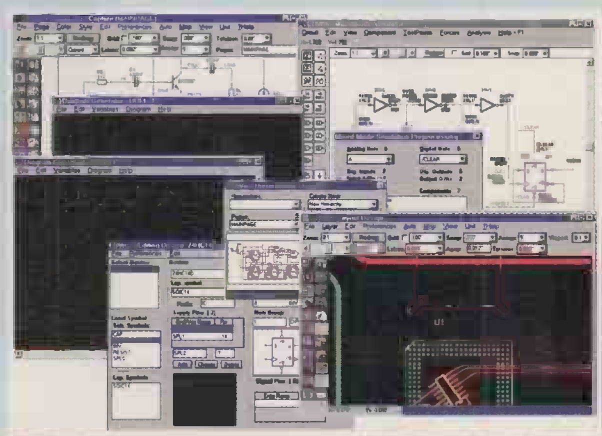

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Society gets the scientists it deserves

A few months ago Dr Peter Cochrane, the outspoken director of technology at BT, hit out at the media for its sloppy reporting of scientific issues like mad cow disease, genetically modified foods and mobile phone health scares.

His view is that inaccurate reporting of half-understood scientific facts has unnecessarily stoked the fears of the public.

"There is nothing quite like widespread ignorance fuelled by a good advertising campaign to trigger panic," said Dr Cochrane. "Judging from the media and popular press you might draw the view that science and technology are some kind of curse inflicted upon us."

Good on yer Peter! Men and woman researchers cheered in their laboratories from Bristol to Edinburgh. The scientists who could not defend themselves against the sensationalist tabloids and broadsheet newspapers had found someone prepared to fight for their corner.

Dr Cochrane, who is himself a regular newspaper columnist, seems to have made some perfectly valid claims. However, there was one important point that he failed to hit upon. That is the crucial fact that no one in society truly believes or trusts what scientists and technologists say any more.

A healthy scepticism about all things scientific or technical is to be welcomed – even cherished. After all wasn't it scientists who once argued that the Earth was flat and claimed that there was life on Mars?

But it is possible that this scepticism has spread into something altogether more serious. Science has an image problem and its practitioners are in danger of being bracketed along with priests and politicians as people we cannot wholly trust.

Society no longer trusts what our white-coated scientists tell it. This is unfortunate because science is particularly big news at the moment.

Concern over the impact of the Millennium Bug, fears over the consequences of genetically modified foods and, "should I or shouldn't I have this mobile phone so close to my head?" are big three issues making the headlines.

You can forget clever inventions like optical fibre or the microprocessor. It is on the reporting and eventual outcome of these three issues that most people in the street judge our scientists. And many unfortunately are seen as laughing stocks.

Don't take my word for it. Consider the British university being funded to carry a scientific investigation into the origin of the crop circles of Wiltshire.

Scientists are in danger of being put in glass boxes and having purple gunge poured over them on prime-time TV. For those with a passion for engineering innovation I must remind you that, Heinz Wolf's Egg Race has been replaced with Craig Charles and Robot Wars.

Things have got more powerful and much more violent, but the science behind the TV programmes is being dumbed down and with it falls the credibility of the scientists.

Dr Cochrane puts part of the blame with the scientists themselves. Seeking kudos for findings before they have been rigorously proved. "The scientific community, for reasons that escape me, has seen fit to break ranks, abandon the scientific principal and go public on the leanest of evidence," said Cochrane. "If an experiment cannot be repeated by at least two independent groups across the planet, then going public is plain irresponsible."

But that is just part of the cultural problem. Everyone wants to be a media star for five minutes and perhaps scientists are no different from anybody else.

Whether it is revelations about government's genetically modified Soya crop research or observations about the heating effects of mobile phones, it is the scientific information that is the valuable commodity. If a scientist can get to the market first there is no saying what rewards may be waiting for them.

The very fact that a renowned scientist like Dr Cochrane felt strongly enough about sensationalism and misrepresentation in the media to

"...the current trend for scientific sensationalism is seriously undermining the credibility of many of our scientists, who now find themselves as mistrusted as your average TV game show host."

say what he did is evidence of a scientific community under siege. It must be difficult for the scientific establishment to come to terms with the realisation that society not only no longer respects its status, but frankly does not always believe what it says anymore.

In the 1930s the cold-hearted scientists of Cambridge, Heidelberg and Los Alamos had a near god-like aura around them.

In contrast, today's scientists and technologists are seen as fallible human beings with a string of nonsensical letters after their names. Nothing wrong with that. Hero-worship is rarely beneficial to the worshiper.

However, the danger is that the current trend for scientific sensationalism is seriously undermining the credibility of many of our scientists who now find themselves as mistrusted as your average TV game show host. ■

Richard Wilson

Electronics World is published monthly. By post, current issue £2.45, back issues (if available £3.00). Orders, payments and general correspondence to **L333, Electronics World, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS.** Tlx:892984 REED BP G. Cheques should be made payable to Reed Business Information Ltd
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Subscriptions: Quadrant Subscription Services, Oakfield House Perrymount Road, Haywards Heath, Sussex RH16 3DH. Telephone 01444 445566. Please notify change of address.
Subscription rates 1 year UK £36.00 2 years £58.00 3 years £72.00. Europe/Eu 1 year £51.00 2 years £82.00 3 years £103.00
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Overseas advertising agents: France and Belgium: Pierre Mussard, 18-20 Place de la Madeleine, Paris 75008. United States of America: Ray Barnes, Reed Business Publishing Ltd, 475 Park Avenue South, 2nd Fl New York, NY 10016 Tel; (212) 679 8888 Fax; (212) 679 9455
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REED BUSINESS INFORMATION

Euro chip R&D – UK could miss out

The UK could lose out on an opportunity to play a role in the next phase of collaborative European microelectronics R&D unless the government gets its act together quickly.

This is the stark warning from a leading industry spokesman.

"If a country is not acknowledging that IT is the driving force of their economies they will condemn themselves to a low ranking in the economic league," said Dr Jürgen Knorr, chairman of the European Microelectronics Development for European Applications programme, known as MEDEA. "If you're not participating, you're not going to participate in wealth and job creation."

The programme is gearing up for MEDEA Phase-2, with Austria, Belgium, Finland, France, Germany, Ireland, Italy, Holland, Sweden and Switzerland taking part. The UK's financial commitment is zero.

For Tony Blair's government to pay the lip service it does to the benefits of high-tech, and then take no part in pan-European R&D is a surprise to many Europeans.

The value of high-tech consortia has been well proven with Europe's collaborations on GSM and Jessi (MEDEA's predecessor programme). "Jessi closed the technology gap and MEDEA realised that in products," said Knorr. "Phase-2 keeps Europe involved with the revolution that started with the invention of the transistor and became the IT industry."

Phase-2 will concentrate on process technologies and design, standardising software tools and interfaces, and specific applications such as advanced mobile voice and advanced digital audio and video systems.

David Manners
Electronics Weekly

New wireless LAN standard supports 54Mbit/s

A wireless LAN standard has been launched this week that will support data rates up to 54Mbit/s.

The HiperLan 2 Global Forum – comprising Nokia, Ericsson, Telia, Bosch, Texas Instruments and Dell – is promoting the technology as a global standard for in-building and metropolitan applications.

HiperLAN 2 works at 5GHz, a frequency dedicated to wireless LANs globally, and features a radio interface with a guaranteed 'quality of service' for given data types such as video.

It will also work with the UMTS third-generation mobile phone standard. "That really gives you ultimate mobility," said Vesa Walldén, v-p of marketing for Nokia's wireless business communications. Offering data rates from 9 to 54Mbit/s, HiperLAN 2 will adapt its data rate depending on channel conditions.

Walldén expects HiperLAN 2 to eventually be used in homes. "All wireless LANs and Bluetooth work on the 2.4GHz frequency," said Walldén. "Once we see the massive deployment of Bluetooth enabled devices we will run out of bandwidth – not today but soon."

Already specification work is under way to extend HomeRF, the wireless home networking standard. HomeRF offers 1.6Mbit/s links. According to Benno Ritter, wireless connectivity product marketing manager at Philips Semiconductors, HomeRF multimedia extensions promise data rates up to 60Mbit/s. The underlying technology for HomeRF multimedia is still to be decided and Nokia's Walldén believes it could use HiperLAN 2.

The specification will be completed by November with products expected in late 2001.

Roy Rubenstein
Electronics Weekly

NV memory smaller than DRAM, fast as SRAM

An innovative memory replacing both DRAM and flash could be in production within five years, claim its inventors at Hitachi's Cambridge Labs.

The PLEDM, or phase-state low electron-number drive memory is being developed by Hitachi's central research labs in Japan.

"We expect, or hope for, commercial success around the year 2005," said Dr Hiroshi Mizuta, head of the Cambridge research centre.

PLEDM seems to be the perfect technology – smaller than DRAM, non-volatile like flash, and can be made as fast as SRAM.

"Its area is half that of DRAM and is smaller vertically and simpler to manufacture," said Dr David Williams, senior researcher at the Hitachi Cambridge Labs.

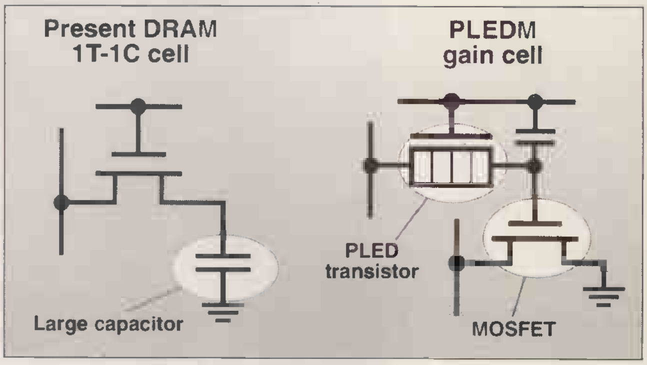
The devices are made using standard CMOS processes and could easily be integrated with logic.

Since handing the PLEDM over to Japan, the researchers in Cambridge have been looking further at single electron memory and logic.

Initial research with micrometre-scale devices worked at close to absolute zero. Now at tens of nanometres, test circuits are working in liquid nitrogen and some even at room temperature.

However, these device are not expected to become commercially viable until at least the year 2015.

Richard Ball



DRAM beater? Hitachi's PLED memory is based on a standard MOSFET. On top of the gate, a second vertical transistor is fabricated which writes, erases and stores the state of the cell. Unlike DRAM it needs no large storage capacitor.

Internet via mains venture turned off

Nortel Networks and United Utilities have decided to disband NOR.WEB, their jointly-owned Internet-over-the-mains company.

"The decision has been made to close the joint venture even though the technology is robust and well proven," said Kate Thomson, NOR.WEB's director of marketing programmes.

Launched two years ago, the company developed its digital powerline technology which used electricity substations to modulate data onto the mains. This provided homes and businesses connected to the substation with 1Mbit/s links for Internet access.

When the company first launched the technology it announced services would start a year ago. However, deployment of the technology suffered a year's delay.

The result of the delay has been increased competition from broadband technologies such as digital subscriber line (DSL) and cable modems, and ultimately the venture's closure.

"The projected volumes for digital powerline, with the roll out of xDSL and cable, are not significant enough," explained Thomson.

This is despite the successful completion in May of a technology trial in Manchester involving 70 homes, a school and small businesses. In turn the technology has also been tested by power utilities in Germany and Sweden.

No redundancies are expected – the 50 NOR.WEB staff will be redeployed by the two parent companies.

There are also no plans to use the technology in other applications. "The two companies still own the patents but there are no current plans at present," said Thomson.

Roy Rubenstein

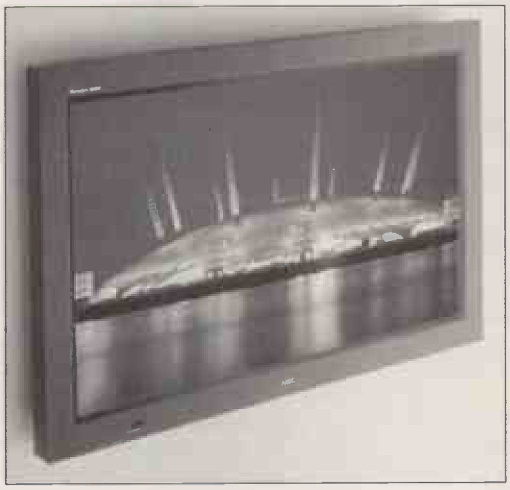
New figures say Asian crisis is over

Further indication that the Asia crisis is over is provided by the latest figures from the Semiconductor Industry Association (SIA). Worldwide semiconductor sales in July were $11.55bn, up 19.3 per cent from the same period last year, with the upturn being led by the Asia Pacific market – up nearly 30 per cent.

"July's global sales continued the robust growth that began in mid-1998," said George Scalise, SIA's president.

The Americas market grew 18 per cent over the last year whereas European sales were up a modest 6.3 per cent.

The SIA cited strong PC demand which is driving microprocessor sales as well as the double-digit growth of DSP and flash memory as a result of the burgeoning communications market.



Welcome to the plasma dome... NEC has won a contract to supply the Millennium Dome with plasma TV screens and LCD projectors. The New Millennium Experience Company (NMEC) is expected to need up to 300 42in. plasma displays and 100 LCD projectors. Plasma screens were chosen because of their thickness – under 100mm – and large viewing angle. NMEC's order is the largest yet received in Europe by NEC. Screens will be delivered during the next two months.

54 000 years of computing time fails to find ET

Screen-saver software designed to search for extra-terrestrial intelligence has become the world's biggest supercomputer.

Over a million users around the world have downloaded the SETI@home software which searches radio telescope data while the user's computer is idle.

Launched on May 17 this year, SETI@home has become the largest computation in history involving users in 224 countries. On August 14, Ed Bradburn from the UK became the millionth user of the screen-saver software.

Since May, over 54,000 years of computer time has been logged on the project, a number rising by 600 years per day.

Unfortunately, despite the massive resources being thrown at the project, not a shred of evidence has yet appeared to show that ET exists.

● The Arecibo Observatory in Puerto Rico, pictured left, is the world's largest radio telescope. It provides data for SETI@home – the world's biggest computer project.

The Arecibo Observatory in Puerto Rico

Picture courtesy of David Parker and Arecibo Observatory

TiePieScope HS801 PORTABLE MOST

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- The versatile software has a user-defined toolbar with which over 50 instrument settings quick and easy can be accessed. An intelligent auto setup allows the inexperienced user to perform measurements immediately. Through the use of a setting file, the user has the possibility to save an instrument setup and recall it at a later moment. The setup time of the instrument is hereby reduced to a minimum.

- When a quick indication of the input signal is required, a simple click on the auto setup button will immediately give a good overview of the signal. The auto setup function ensures a proper setup of the time base, the trigger levels and the input sensitivities.

- The sophisticated cursor read outs have 21 possible read outs. Besides the usual read outs, like voltage and time, also quantities like rise time and frequency are displayed.

- Measured signals and instrument settings can be saved on disk.This enables the creation of a library of measured signals. Text balloons can be added to a signal, for special comments. The (colour) print outs can be supplied with three common text lines (e.g. company info) en three lines with measurement specific information.

- The HS801 has an 8 bit resolution and a maximum sampling speed of 100 MHz. The input range is 0.1 volt full scale to 80 volt full scale. The record length is 32K/64K samples. The AWG has a 10 bit resolution and a sample speed of 25 MHz.The HS801 is connected to the parallel printer port of a computer.

- The minimum system requirement is a PC with a 486 processor and 8 Mbyte RAM available. The software runs in Windows 3.xx / 95 / 98 or Windows NT and DOS 3.3 or higher.

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CIRCLE NO. 106 ON REPLY CARD

Silicon-germanium starts to make its mark

Until recently, IBM was the only volume semiconductor manufacturer to take silicon germanium (SiGe) seriously. Now, Motorola, Lucent Technologies and Infineon Technologies are among those companies following IBM's lead on SiGe.

In recent weeks, two more alliances are a clear indication that SiGe technology is not just about producing transistors with transition frequencies of 75GHz. The technology is set to become the basis of a whole new generation of low-power communications chip sets for both wireless and broadband networking applications.

Atmel's Temic subsidiary, itself one of the first companies to produce commercial SiGe parts, has struck development and manufacturing alliances with RF component specialists M/A-Com, part of the AMP group, and Anadigics.

The companies say that the alliances will produce new SiGe devices which will target both wired and wireless infrastructure especially LANs and the local loop.

What is significant about these deals is that it brings together established RF components specialists with Temic's proven SiGe process technology. As Dr Charles Huang, chief technical officer for Anadigics describes the move: "Having access to Temic Semiconductors' SiGe facility and technology provides us with an opportunity to complement our existing gallium arsenide (GaAs) and silicon programmes."

Temic has already introduced RF components for the DECT cordless telephone and GSM mobile telephone standards. Like IBM, it has been in volume production of SiGe parts since the start of the year.

But these alliances will now give Temic the specialist RF IC design capability it will need to capitalise on the emerging market for low-power RF SiGe components.

SiGe transistors can be integrated into devices with standard CMOS components but bring faster speed and lower power consumption. And it is the lower power consumption rather than high-frequency performance which is the key to this latest interest in SiGe technology.

Ask Dr Neil Morris, director of advanced technology development at Philips Semiconductors Albuquerque if SiGe is needed for IC design in the two to 4GHz range. "The answer is no,"says Morris, "we've already shown that."

He is referring to the company's QUBiC3 0.5μm silicon BiCMOS process, which is being used for a range of RF communications devices including a digital cordless phone (DECT) transceiver and a family of frequency synthesisers operating at up to 3.7GHz.

Philips also has considerable SiGe expertise, but it still believes that it is not cost competitive with its 0.5μm BiCMOS process in the two to 4GHz band.

It is fine if you have the silicon process capability of a company like Philips to, "push the silicon envelope" to the limit, as Morris describes it. But there are sufficient big name semiconductor leaders going after SiGe to make one wonder whether there is something else in this newer technology. After all neither IBM or Motorola are exactly novices when it comes to pushing silicon technology to the practical limits.

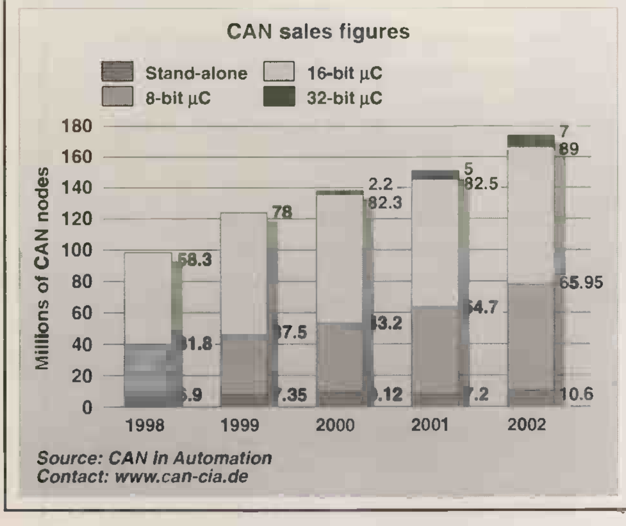
It seems that it is SiGe's inherently lower power consumption, rather than high-frequency silicon process technologies, which is providing the trigger for this latest push for the technology.

Like most big producers, Infineon is using 0.25μm CMOS process technology for its mixed-signal parts. That produces transistors with 25GHz transition frequencies, says Danny Thomas, a marketing manager at Infineon. "With the silicon-germanium process, the f_T is 75GHz."

But if these silicon processes are more than adequate for today's 900MHz and 1.8GHz mobile phones and inherently cheaper, why use SiGe? "We use SiGe for reduced current consumption," answered Thomas, but not necessarily for its faster operation.

So it is the latest moves to make all electronic products from PCs to digital TVs battery-power and portable that probably lies behind SiGe's new trendy status. Perhaps, unlike poor old GaAs, SiGe is an alternative process technology that *is* about to make its mark.

DATAUPDATE



New digital camera sees only ultraviolet light

A team of scientists from North Carolina State University, the Night Vision Laboratory and the Honeywell Technology Center have demonstrated a digital camera that senses ultraviolet light.

The camera is destined for use in military night vision systems and environmental monitoring, the developers believe. Objects that emit UV include rockets, soldering and welding equipment and astronomical objects such as stars.

In order to sense only UV light, the team fabricated an array of p-i-n photodiodes using aluminium gallium nitride (AlGaN). So far the researchers have managed to fabricate a 32 by 32 array.

The array starts life as a sapphire wafer – transparent to UV. Metal organic vapour phase epitaxy is used to deposit the n-layer of AlGaN, followed by the undoped GaN and finally the p-type GaN.

The array is bonded to a standard CMOS chip containing the control and interface circuitry.

The array is sensitive to light between 320 to 365nm wavelengths, making it blind to visible light an infra-red. ■

Understanding USB

Devised to simplify the job of connecting peripherals to a PC, the increasingly popular universal serial bus communicates data at up to 12Mbit/s over two of its four wires. Simple connection, yes, but the protocol needed to make the bus efficient and transparent to the PC user is complex, as Tony Wong explains.

Universal serial bus, or USB, is recommended for the new generation of IBM-compatible PCs in the 'PC 98 System Design Guide'.[1] It is also supported by Windows 98.

This bus provides an easier way of connecting a PC to a variety of peripherals via a serial bus. The universal serial bus is a four-wire cable of which there are two wires for power and ground – namely V_{bus} and Gnd – and two for data transfer, D+ and D–.

Up to 126 devices can be simultaneously connected to a PC via USB without the fear of running out of PC i/o addresses or having conflicts on IRQs and DMA channels used. The bus can also reduce cost and PCB space by removing the need for traditional attachment ports such as keyboard connector and serial ports.

Other benefits of USB are its low cost and that it supports data transfer at up to 12Mbit/s in 'full-speed mode' This is described in USB specification 1.0/1.1. At this speed, it is possible transfer data such as voice and compressed video signals in real time.

At the end of 1999, the USB 2.0 specification will be officially released and will move the maximum transfer rate to between 120 and 240Mbit/s. However, behind of these features, there is a need for sophisticated USB embedded controllers to handle the unique protocol and algorithm for the data communication activities on the bus.

In this article, I present a summary of the USB protocol format based on USB 1.1.

As an example of how the bus is implemented, I describe the Infineon Technologies C541U embedded USB microcontroller. Infineon Technologies was formerly Siemens Microelectronics by the way.

What is an 'end point'?

An end-point, or device end-point, as used in USB terminology, is not the easiest of concepts to understand. In the specifications, an 'end-point', or EP, is described as, "A uniquely identifiable portion of a USB device that is the source or sink of information in a communication flow between the host and device." Physically, an end-point can be considered as a memory area for data flow.

Take, for example, a CD-ROM drive with USB interface. The drive can be accessed by the file manager tool to read a data file from the CD, but it can also be used as to play audio CDs. As a result, you need to use two end-points to handle these two functions. One end-point is configured as 'bulk transfer', for transferring data files, while the other end-point is configured as 'ISO transfer' for real-time audio data transfer.

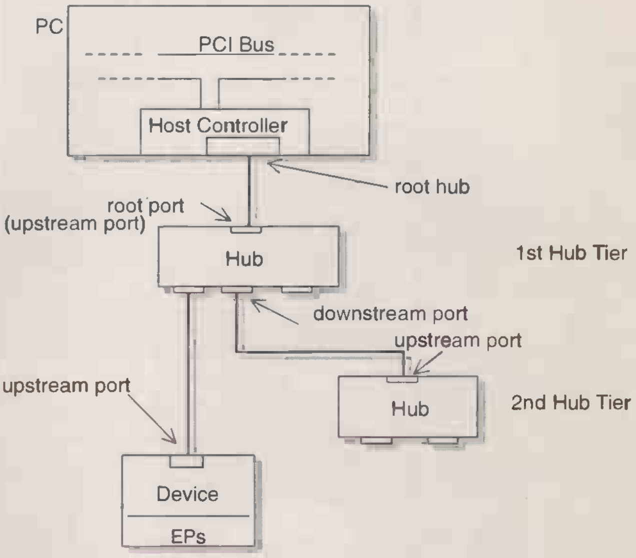


Fig. 1. With universal serial bus, hubs can be used to expand the system, allowing connection of up to 126 devices to one PC.

USB terminology

Descriptor
A Descriptor carries the information used for identify the device, for example, number of end-points, or EPs, in the device and the end-point type for each. This information is usually accessed by the host from the device.

Function
This is a USB device that provides the host with a capability, such as an ISDN connection.

IN
A packet-identification, or PID, type used to perform the 'in' transaction. Data packets flow from device to host.

HID
Human Interface Devices class. Devices are used by humans to control the operation of computer systems. Examples are keyboards, pointing devices and bar-code readers.

Host
A computer system with the installed host controller driver is a host. This includes the hardware USB root port, which may be a PC system running under the Windows 98 operating system for example.

Host controller
The host's USB interface.

Hub
A full-speed USB device that allows additional connections to the USB bus is know as a hub.

OUT
A packet-identification, or PID, type used to perform the 'out' transaction. Data packets flow from host to device.

PID
Packet identification – a field in a USB packet that indicates the types of packet. Examples are SOF, IN and ACK.

SETUP
A packet identification, or PID, command type used to perform the 'set-up' transaction. Data packets flow from host to device. One use of SETUP is to allocate an address to a device.

SIE
Serial Interface Engine is a hardware circuit and a part of USB module in silicon. It performs USB data processing tasks.

SOF
Start-of-Frame. SOF is a packet-identification type and also the first transaction of a frame. It allows end points to synchronise their internal clock to the host.

USB peripheral/USB device
A device that performs an USB function such as a hub, a USB keyboard, or USB speakers.

USB peripheral silicon
A USB embedded silicon chip.

Table 1. Some dedicated USB chips and microcontrollers with USB support.

	Samsung	**Zilog**	**Cypress**	**Infineon Technologies**	**NetChip Technology**	**Motorola**
Part number	KS86C6004	Z8E520	CY7C64113	C541U	NET2888	MC68HC08KH12
Product type	8-bit controller with USB	8-bit controller with USB	8-bit controller with USB	8-bit controller with USB	USB peripheral controller	USB keybd 8-bit controller
with hub						
USB speed	Low	Low	Full	Full/low	Full/low	Full/low
ROM (bytes)	4K	6K OTP	8K EPROM/OTP	8K OTP	—	12K OTP/Flash
RAM (bytes)	208	160	256	256	—	384
Clock	6MHz	12MHz	6MHz	12MHz	—	6MHz
USB buffer	8 bytes for transmit 8 bytes for receive	8 bytes	Up to four 8-byte data EPs Up to two 32-byte data EPs	Configurable USB buffer 8 bytes for control EP Up to 64 bytes for data EPs	64 bytes each of transmit and receive for bulk/ISO transfers	8 bytes each of transmit and receive for control EP 8 bytes of transmit for data
EP				Total buffer size 256 bytes	8 bytes for bulk EP 8 bytes for interrupt EP	
End points	1 control EP 2 data EPs	1 control EP 2 data EPs	1 control EP 4 data EPs	Full-speed mode: 1 control EP 4 data EPs Low-speed mode: 1 control EP 2 data EPs	1 control EP 4 data EPs	1 control EP for hub 1 data EP for hub 1 control EP for device 2 data EP for device
Op. voltage	4.5V~5.5V	4.0V~6V	4.0V~5.5V	5V	3.3V	5V
Int. transceiver	Yes	Yes	Yes	Yes	Yes	Yes

Notes: With the C541U, each data EP can be configured to support any one of four USB transfer types. The NET2888 acts as a bridge between processor Independent local bus and USB bus. The Motorola chip has 1 full-speed upstream port, 4 full/low-speed downstream ports and it supports interrupt transfer on the data EP.

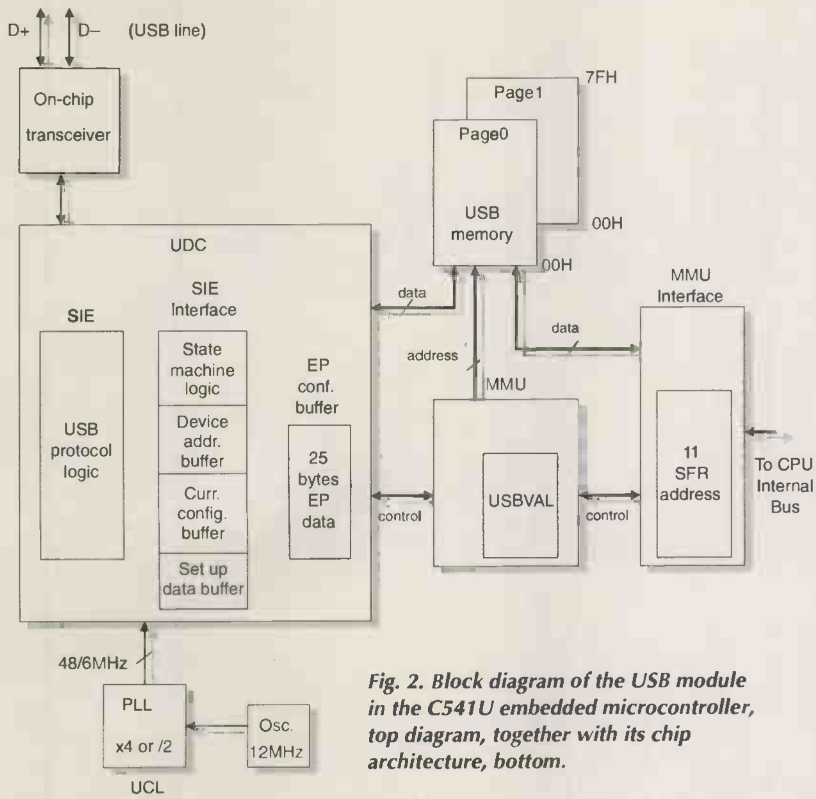
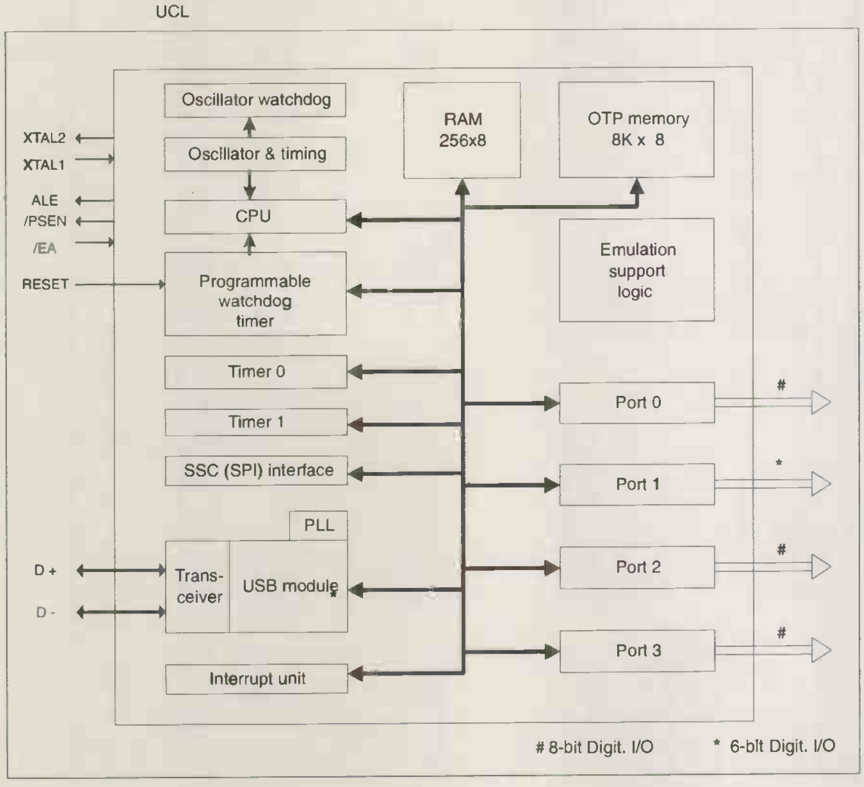


Fig. 2. Block diagram of the USB module in the C541U embedded microcontroller, top diagram, together with its chip architecture, bottom.



This device provides a simple solution to the USB peripheral implementation. Its hardware can handle the USB protocol transmissions automatically.[2]

USB system architecture

There are three basic hardware elements in the USB system architecture. They are host, hubs and devices, **Fig. 1**.

The connection uses the 'tiered-star' topology and can be connected up to five levels – i.e. have five hub tiers. Normally, the host controller and root hub are implemented via a chip set on the PC motherboard.

The host controller controls transactions over the USB system. There are two types of host controller. They are the 'open-host controller interface', or OHCI, and the 'universal-host controller interface, which is shortened to UHCI.

From an application point of view, the OHCI can support multiple transactions for a particular device end point, or EP, within a 1ms 'frame'. There's more on this later. On the other hand, the universal host controller supports one transaction for a particular end point in each frame. Software for USB devices should be able to handle transactions with either of these controllers.

A root hub acts as a port for attaching the USB device, Fig. 1. A USB hub allows multiple connections to the USB system and detects when devices are attached to or detached from the system. It also forwards the bus traffic between its upstream port and downstream ports.

Each USB device is allocated end-point numbers. End point number EP0 is reserved for the device's configuration by the host. It provides a point of communication to the host by means of EP descriptors.

End-point descriptors communicate device attributes and characteristics to the host. According to this information, the host configures the device and locates the USB client software driver.

Other device end points can be considered as a function of the device and can be separately configured for one of the different transfer types to communicate with the host.

For example, a keyboard application, which comes under the USB standard's 'human interface device, or HID, class uses EP0 for the device configuration and may use EP1 as an interrupt transfer to send the key-scanned data to the host. More details on the EP descriptors are discussed in references 3 and 4.

USB supports four types of data transfer:

- Control transfer – transfer request commands from host to device.
- Interrupt transfer – data transfer from an interrupt driven device to host.
- Bulk transfer – transfer for a large amount of data.
- Isochronous transfer – for applications requiring constant data transfer rate.

Implementing USB

Commonly, a USB embedded microcontroller is used to implement the USB functions. There are also other types of USB interface chip to suit different applications.

Table 1 shows some of the examples of USB chip. Only the USB related features are highlighted here; for details of other on-chip features and updated technical specification, refer to the corresponding company's product home page on the web site.

In the Table, the first two columns show two low-speed USB controllers with different clock frequencies and buffer sizes. The third column shows a full-speed chip, and the next one is a full/low-speed chip. The fifth column shows a USB interface chip without an integral controller and the last one is a controller with hub function.

Generally, a USB chip consists of a USB module in which the serial interface engine, or SIE, plays an important role in the USB activities. It performs all the front-end data processing functions such as NRZI and NRZ conversion, token packet decoding, bit stripping and bit stuffing, and cyclic

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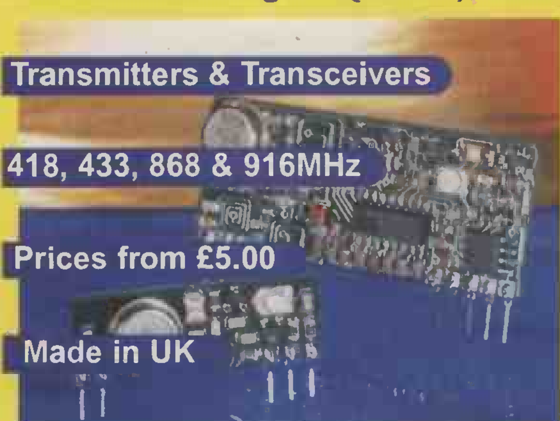

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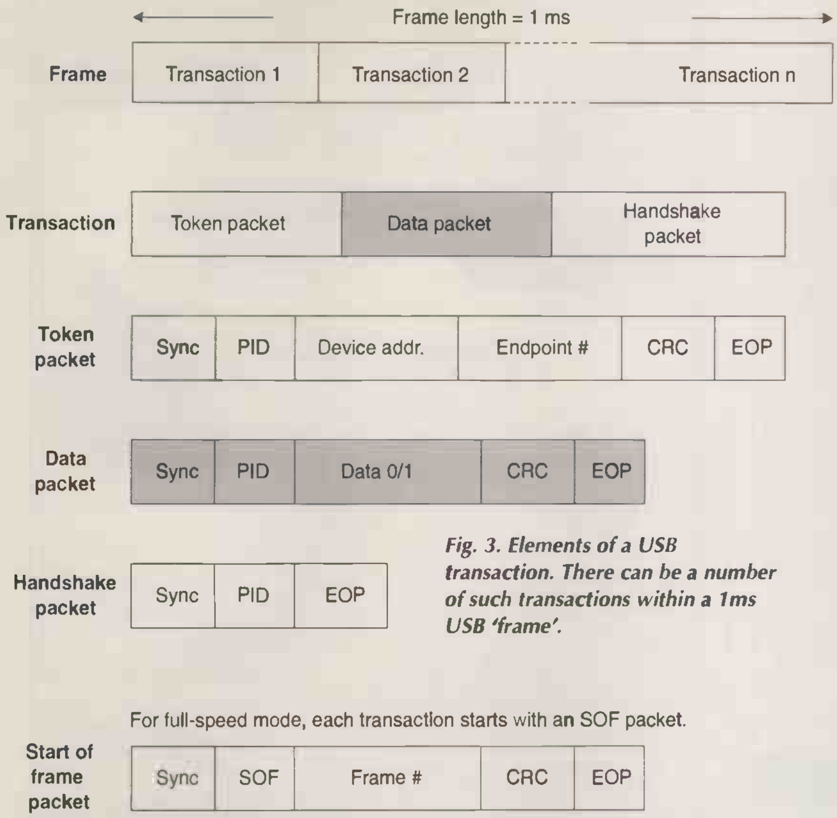

Fig. 3. Elements of a USB transaction. There can be a number of such transactions within a 1ms USB 'frame'.

Table 2: Packet transmission between host and device in full-speed mode.

File ENURATE.USB, recorded using CATC Inspector USB Analyzer S/W Version 2.10, packets 0-377 .

Packet#

0 | RESET(109.46 milliseconds)

109 | Sync(00000001) SOF(0xA5) Frame #(0x079) CRC5(0x15) Idle(11964)

110 | Sync(00000001) SOF(0xA5) Frame #(0x0E3) CRC5(0x0C) Idle(5)

111 | Sync(00000001) SETUP(0xB4) ADDR(0x00) ENDP(0x0) CRC5(0x08) Idle(3)

112 | Sync(00000001) DATA0(0xC3) DATA(80 06 00 01 00 00 40 00)
| CRC16(0xBB29) Idle(5)

113 | Sync(00000001) ACK(0x4B) Idle(11801)

114 | Sync(00000001) SOF(0xA5) Frame #(0x0E4) CRC5(0x12) Idle(5)

115 | Sync(00000001) IN(0x96) ADDR(0x00) ENDP(0x0) CRC5(0x08) Idle(4)

116 | Sync(00000001) DATA1(0xD2) DATA(12 01 00 01 00 00 00 08)
| CRC16(0xC8E7) Idle(6)

117 | Sync(00000001) ACK(0x4B) Idle(11799)

118 | Sync(00000001) SOF(0xA5) Frame #(0x0E5) CRC5(0x0D) Idle(11964)

119 | Sync(00000001) SOF(0xA5) Frame #(0x0E6) CRC5(0x0F) Idle(5)

120 | Sync(00000001) OUT(0x87) ADDR(0x00) ENDP(0x0) CRC5(0x08) Idle(3)

121 | Sync(00000001) DATA1(0xD2) DATA() CRC16(0x0000) Idle(4)

122 | Sync(00000001) ACK(0x4B) Idle(11866)

123 | Sync(00000001) SOF(0xA5) Frame #(0x0E7) CRC5(0x10) Idle(1697)

124 | RESET(Start of Reset)

redundancy generation and checking.

Figure 2 shows the block diagram of the USB module in the C541U controller chip from Infineon, and also the functional block diagram of the chip. More details of the operation of the USB module can be found in reference 5.

USB communication structure

Communication over the universal serial bus is performed with a series of frames. Within a frame, which is 1ms long, there can be a number of transactions.

The number of transactions depends on the number of attached USB devices and how often the host needs to communicate with these devices. A transaction can be viewed as a transfer of data. It consists of three phases.

Figure 3 shows the elements forming a packet phase. A token-packet phase comprises commands sent from host to a device and has four possible packet identifiers, known as PIDs. They are SOF, IN, OUT and SETUP.

Data is transferred during the data-packet phase. Two PID types are available for this, namely DATA0 and DATA1. This allows a data toggling mechanism, which is used to synchronise the transmitter and receiver of a transfer.

Acknowledgement of the data packet transfer is carried out during the handshake packet phase. It carries one of these PID codes, ACK, NACK or STALL, representing the current data receiving status.

Under the USB communication protocol, two kinds of control transfer can be performed. **Figure 4** shows the sequence of the communication for a three-stage control transfer involving a 'get descriptor' command transaction and a 'setup' token. Three-stage control transfer consists of a setup stage, a data stage and a status stage. It is mainly performed by the host to get information from the device.

Figure 5 shows a two-stage control transfer. Such transfers are used by the host to assign data to the device. For example, the host sets an address number to the device using the 'set descriptor' command as shown. Note that there is no data stage for the device to send back data information. The device only performs the 'status' stage to acknowledge the host by sending a zero-length data packet.

In addition to the two control-transfer formats, USB provides another data transfer format that is used to perform interrupt, ISO and bulk transfer types.

Figure 6 shows the data transfer procedure with two examples, namely 'interrupt' and 'ISO' transfers. For the interrupt transfer, the host keeps polling the bus by sending out 'in' tokens to the particular device.

The time interval for polling end-point for data transfer is user defined between 1 and 255ms. If the device has data to be sent, it will transmit the data packets to the host following the 'in' token. If not, it will send out an NAK response, which represents 'negative acknowledge'.

For an ISO transfer, which involves real-time data transmission, the 'in' token should be sent to the device every 1ms.

In order to illustrate the protocol involved in a real application more clearly, we used a monitor system called a CATC USB Inspector. It was set up to capture data transfers between a UHCI host and C541U device controller.

Figure 7 shows the test set-ups involved. The host was running Windows 98 and the device chip contained source code with USB keyboard function.

Test set-up 'A' captures the USB traffic between host and device directly. A full-speed hub and low-speed device are included in the test set-up 'B'. This second set-up can monitor the combined transmissions of full and low-speed packets on the bus at the same time.

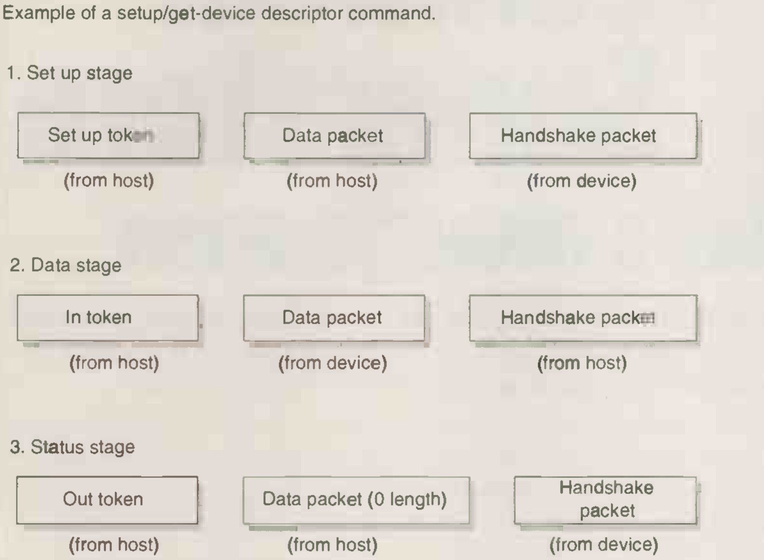

Fig. 4. In the USB protocol, two types of control transfer are possible. This diagram represents a three-stage control transfer, mainly used by the host to get information from a connected device.

Fig. 5. Two-stage control transfers are mainly used by the host to assign data to the device. In this case, the device is being assigned an address.

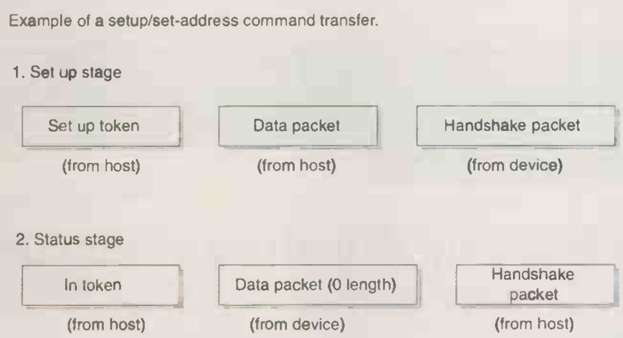

Table 3: Packet transmissions in low-speed mode.

```
File LS-1.USB, recorded using CATC Inspector USB Analyzer S/W Version 2.2,
packets 0-394.
Packet#

18      | EOP( ) Idle(1497)
19      | EOP( ) Idle(114)
20      | Sync(__000001) SETUP(0xB4) ADDR(0x00) ENDP(0x0) CRC5(0x08) Idle(4)
21      | Sync(00000001) DATA0(0xC3) DATA(80 06 00 01 00 00 40 00 )
        | CRC16(0xBB29) Idle(6)
22      | Sync(00000001) ACK(0x4B) Idle(13)
23      | Sync(00000001) IN(0x96) ADDR(0x00) ENDP(0x0) CRC5(0x08) Idle(5)
24      | Sync(00000001) NAK(0x5A) Idle(11)
25      | Sync(00000001) IN(0x96) ADDR(0x00) ENDP(0x0) CRC5(0x08) Idle(5)
26      | Sync(00000001) DATA1(0xD2) DATA(12 01 00 01 00 00 00 08 )
        | CRC16(0xC8E7) Idle(6)
27      | Sync(00000001) ACK(0x4B) Idle(4)
28      | Sync(00000001) OUT(0x87) ADDR(0x00) ENDP(0x0) CRC5(0x08) Idle(4)
29      | Sync(00000001) DATA1(0xD2) DATA() CRC16(0x0000) Idle(6)
30      | Sync(00000001) ACK(0x4B) Idle(887)
31      | EOP( ) Idle(238)
32      | RESET(Start  of  Reset)
```

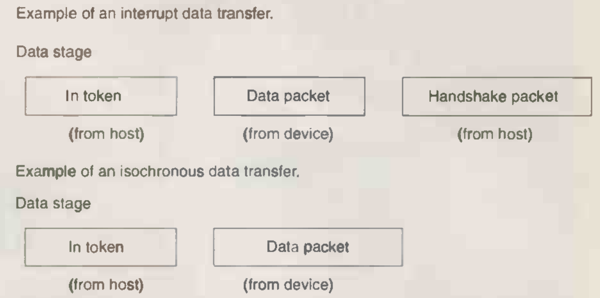

Fig. 6. Two examples of USB data transfer structures. The main difference is that interrupt transfer is polled and the isochronous transfer is real time.

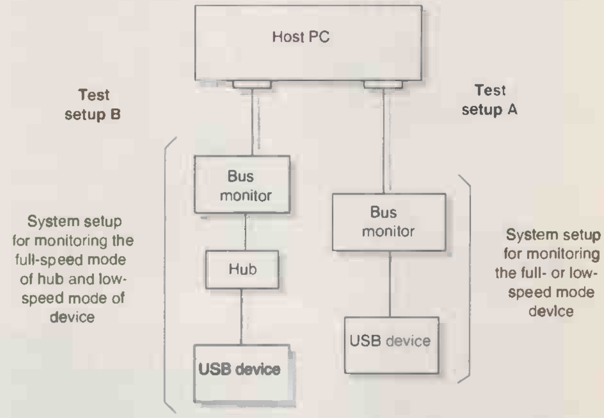

Fig. 7. Set-ups used to analyse USB communications. In this case the host is a UHCI and the device a C541U.

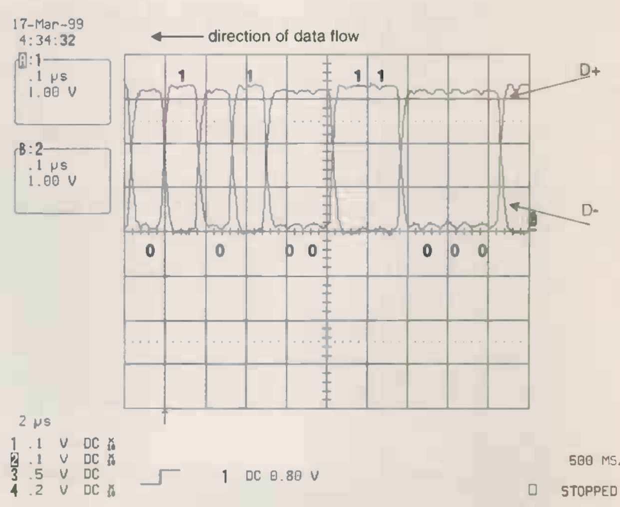

Fig. 8. USB signal, encoded in inverted non-return to zero form during sending of a negative-acknowledge NAK response to a host at full speed.

Snap-Shot of the USB Protocol

Table 2 shows part of the enumeration process of the device in full-speed mode. Enumeration is a procedure that allows a device to be recognised by the host and for setting up a communication pipe between them.

Packets 110 to 122 show a three-stage control transfer involving a 'get-descriptor' command. Packets 110 to 113 form the set-up stage, packets 114 to 117 form a data stage and packets 119 to 122 form a status stage.

Note that packets 110 to 113 are within one frame time, i.e. 1ms. Since there is only one device connected to the host,

Fig. 7 test set-up 'A', the transactions can not use up the whole time frame, resulting in an idle time of 11801 bit times.

For full-speed mode, the data transfer rate is 12Mbit/s. By counting the number of bits in each packet as shown in **Table 2**, it can be worked out that there's around 12K bits within a 1ms time frame. The table below shows the number of bits used for different tokens in the packet.

Field name	No. of bits
Sync	8
ADDR	7
SOF	8
ENDP	4
Frame #	8
DATA0	8
CRC5 / CRC16	5/16
DATA contents	64
SETUP	8
ACK	8

Low-speed mode

The captured packets transferred on the bus in low-speed mode are shown in **Table 3**. You can see that there is no start-of-frame SOF token. The time frame in this case is defined as being between 'end-of-packet' EOP tokens. Packets #20 to #31 are located within one frame.

Packets #23 and #24 show a negative-acknowledge NAK response issued by device to indicate that it is not available to respond to the 'in' packet from the host at that time. The descriptions on the packets are as follows:

- Three-stage control transfer with 8-byte data length:
  Packets #20 ~ #22, setup stage
  get-descriptor command
  Packets #25 ~ #27, data stage.
  Packets #28 ~ #30, status stage.

Full-speed and low-speed signals on the bus

This section looks at how full-speed and low-speed modes can operate simultaneously. Assume test set-up 'B' in Fig. 7. The USB bus between host and hub is in full-speed mode but

Table 4. The Low-speed packet transmission on Full-speed transactions.

File PRE-1.USB, recorded using CATC Inspector USB Analyzer S/W Version 2.2, packets 0-840.

Packet#	Idle(10654)
777	Sync(00000001) IN(0x96) ADDR(0x02) ENDP(0x0) CRC5(0x15) Idle(5)
778	Sync(__000001) DATA1(0xD2) DATA(03 03 00 00) CRC16(0xF0F9) Idle(7)
779	Sync(00000001) ACK(0x4B) Idle(42)
780	Sync(00000001) OUT(0x87) ADDR(0x02) ENDP(0x0) CRC5(0x15) Idle(3)
781	Sync(_0000001) DATA1(0xD2) DATA() CRC16(0x0000) Idle(5)
782	Sync(__000001) ACK(0x4B) Idle(765)
783	Sync(00000001) SOF(0xA5) Frame #(0x270) CRC5(0x0E) Idle(5078)
784	Sync(00000001) PRE(0x3C) Sync(00000001) SETUP(0xB4) ADDR(0x00) ENDP(0x0) CRC5(0x08) Idle(10)
785	Sync(00000001) PRE(0x3C) Sync(00000001) DATA0(0xC3) DATA(80 06 00 01 00 00 40 00) CRC16(0xBB29) Idle(48)
786	Sync(00000001) ACK(0x4B) Idle(87)
787	Sync(00000001) PRE(0x3C) Sync(00000001) IN(0x96) ADDR(0x00) ENDP(0x0) CRC5(0x08) Idle(40)
788	Sync(00000001) NAK(0x5A) Idle(73)
789	Sync(00000001) PRE(0x3C) Sync(00000001) IN(0x96) ADDR(0x00) ENDP(0x0) CRC5(0x08) Idle(38)
790	Sync(00000001) DATA1(0xD2) DATA(12 01 00 01 00 00 00 08) CRC16(0xC8E7) Idle(26)
791	Sync(00000001) PRE(0x3C) Sync(00000001) ACK(0x4B) Idle(10)
792	Sync(00000001) PRE(0x3C) Sync(00000001) OUT(0x87) ADDR(0x00) ENDP(0x0) CRC5(0x08) Idle(10)
793	Sync(00000001) PRE(0x3C) Sync(00000001) DATA1(0xD2) DATA() CRC16(0x0000) Idle(48)
794	Sync(00000001) ACK(0x4B) Idle(2865)
795	Sync(00000001) SOF(0xA5) Frame #(0x271) CRC5(0x11) Idle(1619)

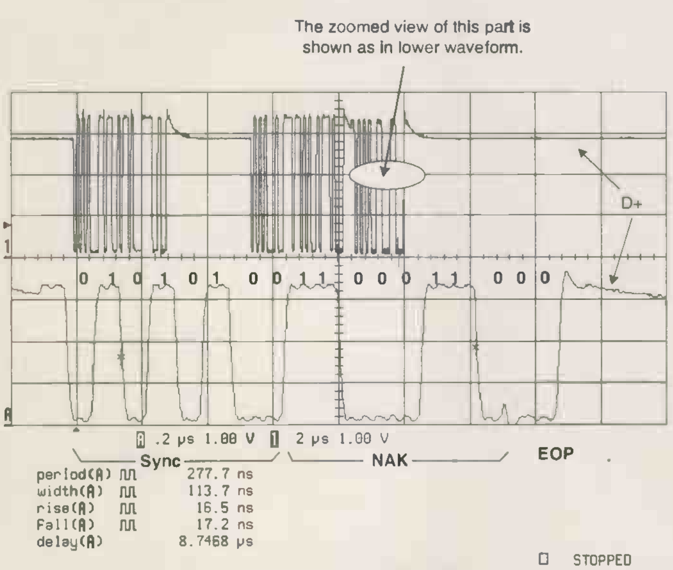

Fig. 9. Signal on USB D+ data line showing the sync, NAK and EOP data formats.

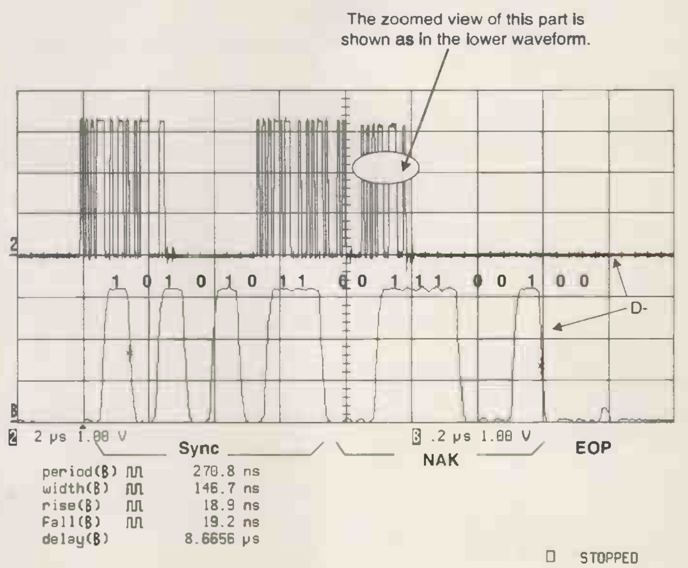

Fig. 10. Signal on USB D– data line showing the sync, NAK and EOP data formats.

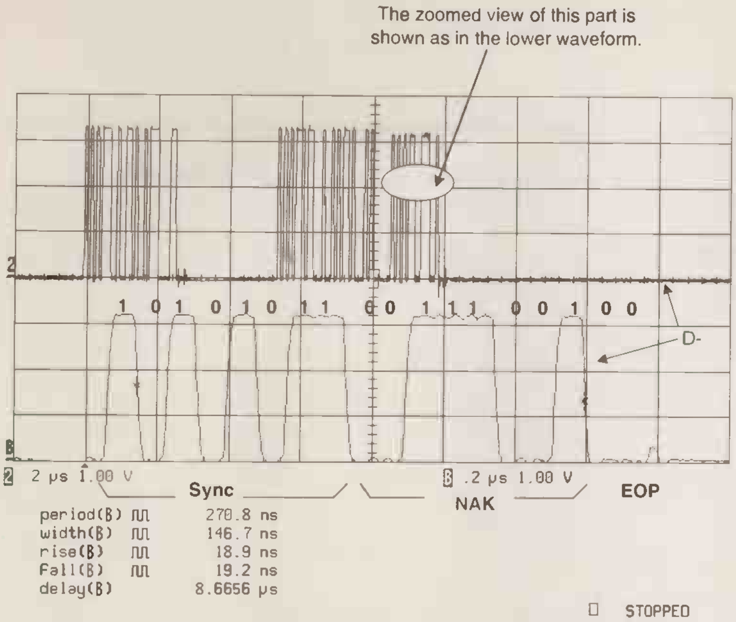

Fig. 10. Signal on USB D– data line showing the sync, NAK and EOP data formats.

(a)

Actual data		0							
Transmitted data (NRZI)	1	0	1	0	1	0	1	0	0

last bit of previous packet

(b) "Sync" data ID

Actual data		0	0	0	0	0	0	0	1
Transmitted data (NRZI)	1	0	1	0	1	0	1	0	0

(c) "NAK" data ID

Actual data		0	1	0	1	1	0	1	0
Transmitted data (NRZI)	0	1	1	0	0	0	1	1	0

Fig. 11. Decoding the inverted non-return to zero signal from the bus.

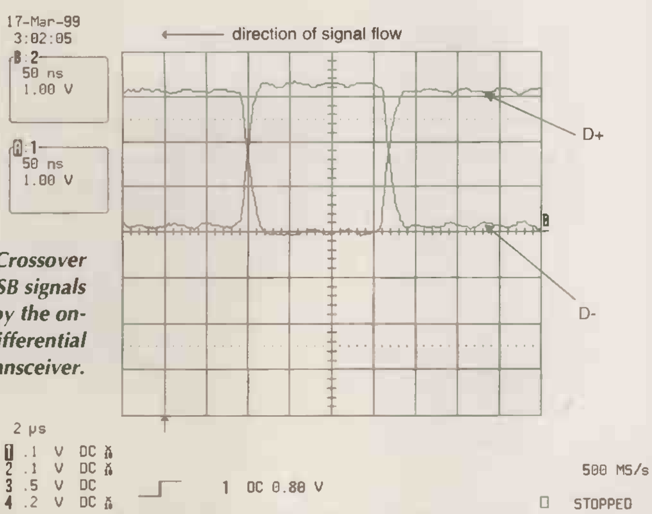

Fig. 12. Crossover points of USB signals generated by the on-chip differential transceiver.

the bus between hub and device is in low-speed mode since the device is low-speed.

We used the bus monitor to observe how the device packets are transmitted on a full-speed bus. **Table 4** shows part of the packet sequences. Packets #777 to #782 are the data and status stages of communication between host and hub using the normal full-speed transfer format.

Packets #784 to #794 perform three-stage control transfer between host and the low-speed device through the full-speed hub. In order to differentiate the low-speed signals from the high-speed for the hub to broadcast them to the downstream ports, a preamble (PRE) packet is required as shown.

Packets following the PRE (0x3C) are the low-speed data. Packets #784 and #785 are the get-descriptor command from host to the device, and packet #786 is the ACK from device.

How USB signals are transmitted

USB protocol involves non-return to zero, inverted, or NRZI, encoding to encode the data before transmitting onto the bus. This encoding method does not need a separate clock signal. In NRZI encoding, a transition between two consecutive data bits represents logic '0' while no transition represents logic '1'.

Figure 8 shows a serial data stream transmitted on the USB bus, encoded in NRZI format. The waveform was captured at the host side by probing on the data lines, D+ and D– while the device was sending an NAK response to the host in full-speed mode.

Logic bits for the D+ signal are shown in bold. For clarity, the D+ and D– signals are separated as in **Figs 9** and **10** respectively.

The upper waveform in Fig. 9 shows two packet transactions and the bottom waveform shows the magnified view of the last part of data transfer. It is easy to work out the actual data value from the waveform. **Figure 11** shows the transmitted NRZI data from Fig. 9.

The left-most bit '1' in the table of Fig. 11a) represents the last bit transferred from the previous packet. This bit is used to decode the following 8 data bits. The second NRZI bit is '0'. The change from '1' to '0' is a transition, so the first bit of actual data is logic '0'.

In Fig. 11b), logic '1' in the actual data row indicates that there is no data transition between the previous NRZI bit, in this case '0'. Following the same procedure for the rest of NRZI bits, the 8-bit actual data decodes as 00000001, which is a specific synchronisation data pattern.

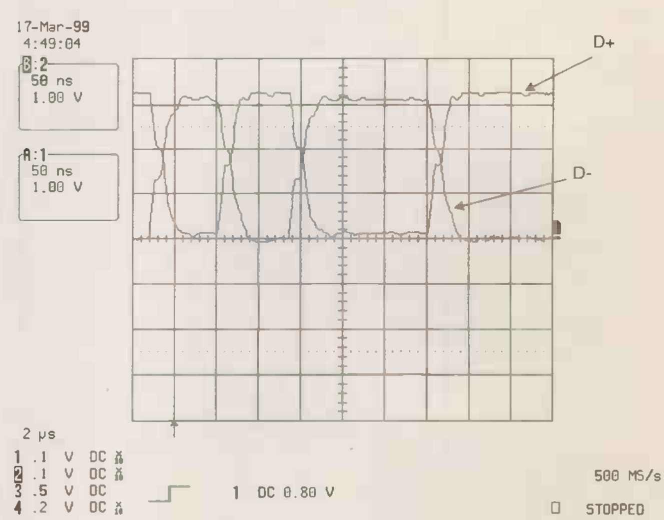

Fig. 13. Full-speed USB signals captured at the device side of the bus during sending of a negative-acknowledge NAK signal to the host.

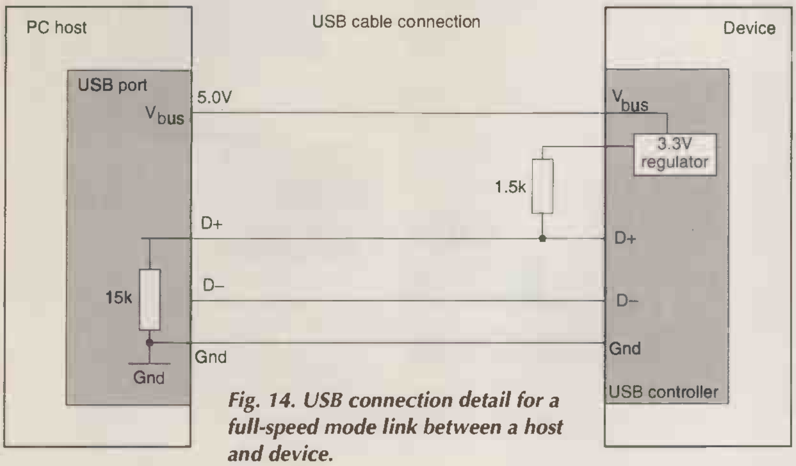

Fig. 14. USB connection detail for a full-speed mode link between a host and device.

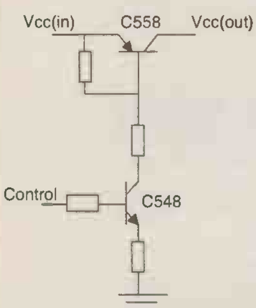

Fig. 15. Power switch control circuit used to switch off high-current circuitry during standby mode.

In Fig. **11c)**, the next 8-bit NRZI data stream is 11000110_2, so the decoded actual data will be 01011010 which is an negative-acknowledge NAK signal.

The three bit times at the end of the waveform in Fig. 9 indicate the end-of-packet (EOP). Both data lines D+/D– are driven low for two bit times and back to high again at the third bit time for D+. The D– line stays low after the EOP as shown in Fig. 10. Some more NRZI examples are shown below:

Name (packet ID)	Actual data (hex)	NRZI code
Data1	D2	0 0 1 1 0 1 1 0
SOF	A5	0 1 1 0 1 1 0 0
ACK	4B	1 1 0 1 1 0 0 0
IN	96	0 1 0 0 1 1 1 0
Setup	B4	0 1 1 1 0 0 1 0

The two data lines should be overlapped in order to show the cross-over point. **Figure 12** shows a close-up view of the waveform of Fig. 8. The cross-over voltage point is about 1.7V and the maximum signal level is about 3.3V.

Signal quality issues

When measuring signals at the device side, a different signal quality can be obtained. The signal shown in **Fig. 13** was captured at the device side while the device was driving NAK to the host.

The shape of waveform is not very smooth at the cross-over points; this is due to impedance mismatch. The signal sent out from the device is partially reflected at the host side and back to the device. This makes it confusing for engineers measuring the signal quality as required in the USB Compliance checklist.

The signals should be captured at the host side when measuring the device transmission signal quality, and captured at the device side when the receiving signal quality is required. According to the USB specification, the rise/fall times for the transmitted signal should be both less than 20ns and 30ns for the received signals.

Power management

A pull-up resistor connected at the downstream end of the cable is used to determine the operating speed of the device. D+ is pulled high for full-speed mode, as shown in **Fig. 14**, while D– is pulled high for low-speed mode.

When a device is attached to the host or hub, there is a pull-down terminator with resistance of 15kΩ at the host/hub side to form a complete loop.

It follows that there is a constant current of about 200μA flowing from the pull-up resistor, through the USB cable and to the pull-down resistor. This current needs to be taken into account when designing USB equipment.

Current consumed by a USB device is an important issue for designers trying to meet the USB specification. This is especially true where devices are involved that obtain power from the host/hub through the V_{bus} power line of USB cable.

To avoid overloading the host or hub, a current of less than 100mA should be drawn by each attached device during normal operation. It should be less than 500μA in suspend mode.

Suspend mode and power down mode in the device and USB controller chip are interpreted differently. During suspend mode, the device should be able to receive the wake-up signal from the host so that it can resume normal operation.

This can be done by turning on the signal receiving circuit of the transceiver, which is normally an embedded on-chip module. The rest of the on-chip modules can then be set to power-down mode in order to reduce power consumption.

However, the constant current through the pull-up, and the current consumed by the transceiver in suspend mode result in less of the 300μA current being available to support other hardware in the device. For this reason, it is recommended to power-down the external hardware circuits to meet the specification. A simple power-control switch circuit is introduced, **Fig. 14,** to switch off the high-current components in the device while in suspend mode.

This circuit can be controlled by an on/off signal from the controller. It also provides a 'soft-turn-on' function that prevents excessive surge current drawn from the V_{bus} by all components at the same time during power on. There is a commercial chip, Infineon's *BCR 48 PN*, that can provide a similar function.

In summary

More details on the USB HID class device,[6] such as a keyboard interface, can be found in the application note mentioned in reference 7.

The introduction of the USB 1.1 standard provides an easy way for the connection of PC peripherals and more and more peripherals are being embedded with USB. Dedicated USB controllers play an important role in USB products.

The imminent USB 2.0 specification will enhance the capability of USB for transmitting multi-media signals. Its higher bandwidth provides a wider range of applications to the next-generation peripherals. ■

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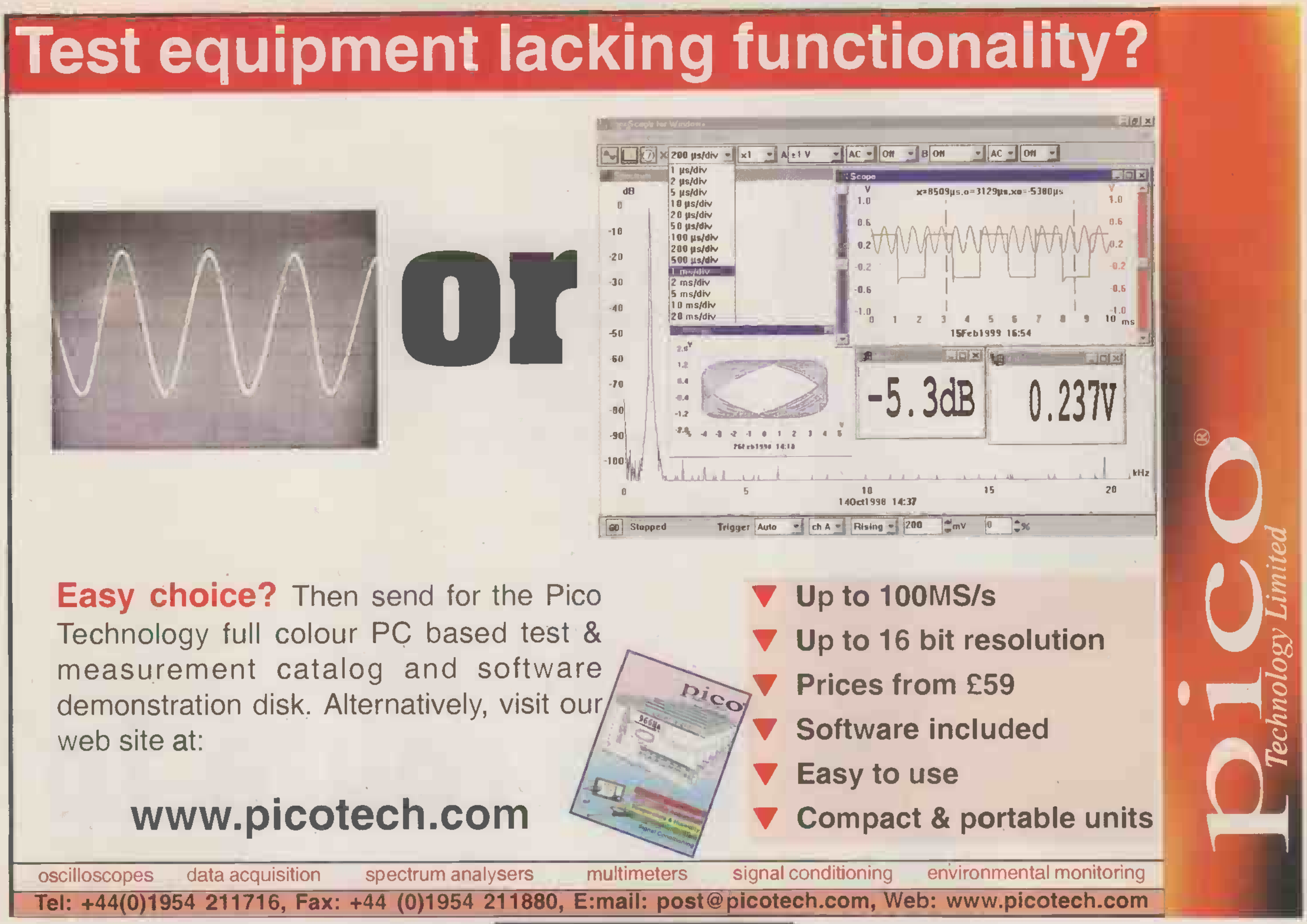


CIRCLE NO.116 ON REPLY CARD

CIRCLE NO.115 ON REPLY CARD

CIRCLE NO.117 ON REPLY CARD

Rod Cooper investigates Easy-PC – a PCB design package with separate analogue and digital analysis add-ons. Virtual instruments are not available, but this suite has modules that extend its usefulness to the microwave region.

The route to simulation V

Review subjects

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

Rod looked at *CircuitMaker* in the August issue. This £199 package is made by MicroCode in the US and supplied by Labvolt in the UK, tel. 01480 300695.

Tina Pro, priced at £299, was featured in the September issue. Quickroute supplies Tina in the UK, tel 0161 4760202.

Proteus, distributed by Labcenter Electronics, appeared in the October issue and is priced at £295 to £624 excluding VAT, tel. 01756 753440.

Now owned by Sightmagic Ltd, *Easy-PC* for Windows is the latest version of the well-established Number One Systems product.

The program is an example of a system using separate analogue and digital analysers; this aspect was commented on in the introduction in the August issue. Strictly speaking, '*Easy-PC*' is a PCB design package. It is the add-ons *Analyser* and *Pulsar* that allow the package to analyse analogue and digital circuits respectively.

Unlike *Workbench*, *CircuitMaker* and *Tina*, where the simulator is an inseparable part of the program, the separate analogue and digital analysers of *Easy-PC* can be purchased on their own, and can operate independently.

Reading matters

Three separate loose-leaf A5 binders cover *Easy-PC*, *Analyser* and *Pulsar* respectively. They are well-written and contain sufficient information to enable the first-time buyer to get started, as well as covering more advanced aspects.

The style of each book is similar, starting with a tutorial-style introduction. A reference section explains all the program commands and controls, and a library section lists the contents.

Links to other programs and net-lists are also discussed. All three books have an index, but no glossary, and are well complemented by the program's Help files.

Capturing schematics

The latest version of Easy-PC reviewed – 2.1 – comes on a CD and security is via a registration number. Pin-limited versions are available, at various prices.

If you are interested in the schematic drawing section of *Easy-PC*, take a look at its review in the August '98 edition of *Electronics World*. In brief, the schematic capture program is a conventional system with symbols being loaded to the screen from the library. It is mainly menu-driven, but the most common functions are included in a single tool bar, which can be turned on or off. There is no parts bin as yet.

Figure 1 shows the graphics quality of a typical schematic and the economical

Requirements

A minimum of a 486 with 16Mbyte of RAM are required needed to run the package. *Easy-PC*, *Analyser* and *Pulsar* occupy 20Mbyte, 2Mbyte and 3Mbyte of hard-disk space respectively. The programs are designed for Windows 95/98 and NT.

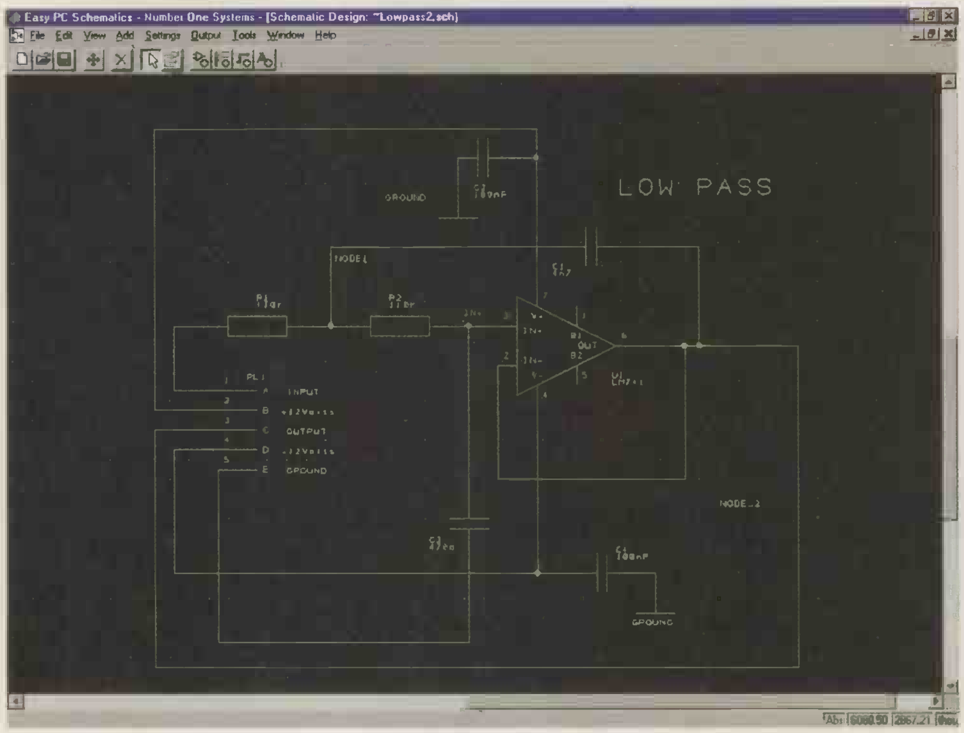


Fig. 1. The circuit used for the results shown in Fig. 2, illustrating the labels on the left of the circuit to be used for setting up the simulation. The colours are the default set. They are readily changed, as indeed they have been in the schematic in the background of Fig. 3.

style of the Windows set-up. When I wrote the first *Easy-PC* review, there were no connections to third party programs, but is now possible to import a net-list in *OrCAD* and *Workbench* formats.

Other new features include multi-level undo and redo, cut, copy and paste, auto-pan, a graphic preview of symbols from the library before use, and a 'best-fit' zoom option.

Analysing analogue circuits

The first thing to note about this product is that it is not Spice-based. This precludes using the Spice data handed out by semiconductor makers with their products. However, it is possible to model new devices, as described in the handbook.

The program comes on two floppy disks. *Analyser*'s main theme is plotting graphs of frequency versus gain, phase, input and output impedance, group delay, input and output voltage standing wave ratio and Y and S parameters. It also plots maximum available gain and stability according to either Linvill, Stern or Rollett criteria.

Where appropriate, real and imaginary parts can be shown directly. This is quite a different range of simulations from the other programs in this review.

There is no transient simulation, or any of the simulations associated with transient analysis. Noise and distortion cannot be simulated either, and there is no dc analysis. So, the range of simulations is limited, and those available seem to be directed towards high-frequency work.

This does not prevent the simulator from being used on, say, audio-frequency circuits. Indeed, graphs from *Analyser* can be seen in published work on audio. But it clearly has less utility outside its intended area than a program offering a broad spectrum of analyses.

Analyser comes into its own though when combined with its sister programs, the electromagnetic simulator '*Layan*', and the rf Smith-chart designer 'Z-Match'. These are outside the scope and the budget of this review, but should be borne in mind when assessing the package.

Analyser can be started from *Easy-PC* schematic capture, or operated on a stand-alone basis with a typed net-list. If you use the schematic-capture route, simulation is started automatically by simply choosing *Analyser* in the 'tools' menu. The graphs that appear are initially auto-scaled to give an acceptable default set of results. They can then be re-scaled if need be.

Alternatively, you can set the parameters up in advance from the menu system in *Analyser* if you know what they are. One graph can be overlaid on another, to compare the before-and-after results. You might want to do this if a component has been changed for example.

The simulations are conventional, i.e. no virtual instruments and the like are used. Input from the circuit under test is done not by a probe tool, but via a menu by naming specific points in the schematic. This assumes that you are using the schematic capture method.

Normally, graphs use both the left and right Y-axis to display two parameters simultaneously. For example the left axis displays gain and the right axis displays phase.

Graphs can be displayed in their own window over the schematic, as shown in **Fig. 2**, and can be resized and moved about in the normal Windows style to suit whatever is on the screen, and several can be displayed together if required.

Alternatively, graphs can be expanded full-screen to make measurement easier. The clarity and presentation style is uniformly good. The library for *Analyser* stands at about 650 device models.

Digital analyses

Pulsar is supplied on CD. Like *Analyser*, it runs automatically from *Easy-PC* schematic capture. Alternatively, it accepts a typed net-list from the net-list editor.

As you would expect, the style of operation is very similar to *Analyser*. There are no virtual instruments, and a menu system is used as before to set up the input, output, etc.

Two digital signal generators are provided. One gives a simple constant-frequency pulse chain; the other is more sophisticated and has user-defined parameters.

Both generators can be attached to any signal in the circuit, automatically disconnecting the existing signal. Removing the generator returns the circuit to its original condition. This procedure considerably speeds up investigation of the circuit under test. Step-by-step simulation is not possible.

Results are displayed in the familiar timing chart, and includes all glitches, down to a picosecond. There is no user-defined control over glitches, so they all appear regardless of duration down to the 1ps limit.

From a practical point of view, this is of more value in trouble-shooting than an idealised timing chart. A typical result, **Fig. 3**, shows the glitches, which are highlighted in red to make them easier to identify. When analysing glitches, a zoom feature allows you see it in more detail.

Like other simulators in this review, such as Labcenter's *Lisa*, *Pulsar* recognises more than just the two usual strong high and low logic states. It can also deal with weak high and low states, high impedance and open circuit. This can give a better correlation with circuits where the drive is via pull up or down resistors, or where non-ideal events occur. Each state can be allocated a colour in the timing chart.

The library of digital models consists of over 120 devices in the 74LS series, the same in the 74HC series, over 110 in the 74HCT series and over 90 in the 4000 CMOS series. As already mentioned, these can be added to manually. There is a library of about 50 logic primitive elements.

Summary

Analyser addresses a somewhat different field from the other simulators reviewed. Although it can be used as a general-purpose simulator, the range of simulations is shorter, but includes areas not covered by the other review products.

If you are thinking of buying this package, check that the scope of the simulations, focussed as it is on a specific area, is suited to your fields of work.

For those of you interested in audio circuits and those who want more than the basic analyses of gain, phase and impedance, then *Analyser* may not have much appeal. But if you design microwave circuits, for example, the availability of the sister programs mentioned earlier makes *Analyser* an attractive proposition.

I found *Analyser* straightforward and pleasant to use. The fact that it can produce a default set of auto-scaled graphs helps when you are in the early stages of learning how to use it. This feature also enables *Analyser* to produce quick 'snap-shot' results when you are working rapidly.

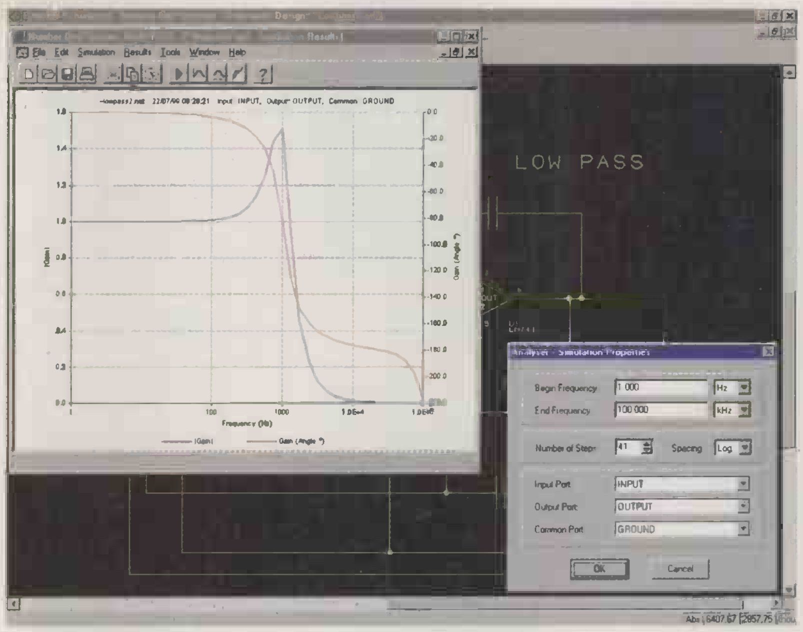
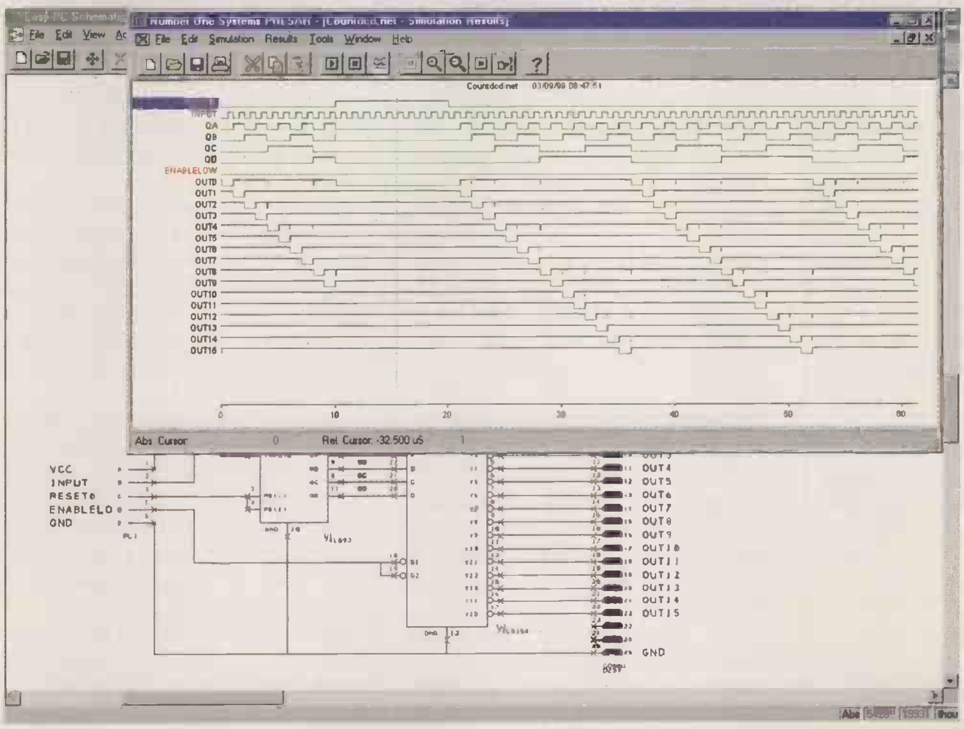


Fig. 2. Analyser simulation of gain and phase, showing the menu system for setting up the simulation. Note especially the method of calibrating the X axis – the same as that in Electronics Workbench. The graphs can be displayed full-screen if required.



I also found *Pulsar* pleasant to use. It performs well as a basic no-frills digital simulator. I would describe the overall learning curve as moderately steep.

Despite the commercial ups-and-downs of this program in recent months, Sightmagic has said that the *Easy-PC* suite is to continue being developed by them in the UK. It will be interesting to see what the company does with it in the future. ■

Fig. 3. Pulsar timing display of a counter-decoder. Note that the all the glitches are shown, picked out in red. To show the colour capabilities of the chart, in this diagram I have arbitrarily chosen strong high as black, strong low green.

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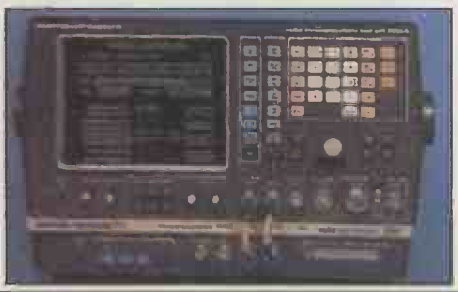

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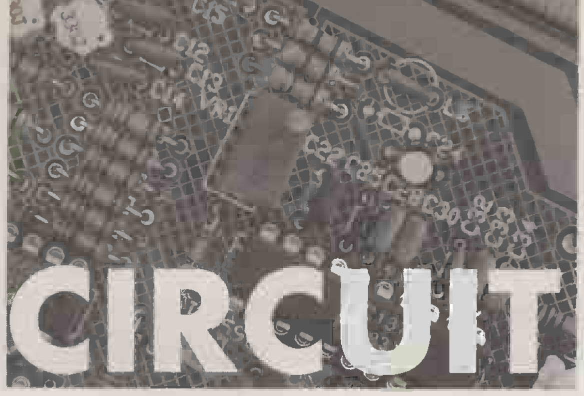


CIRCUIT IDEAS

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

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

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

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

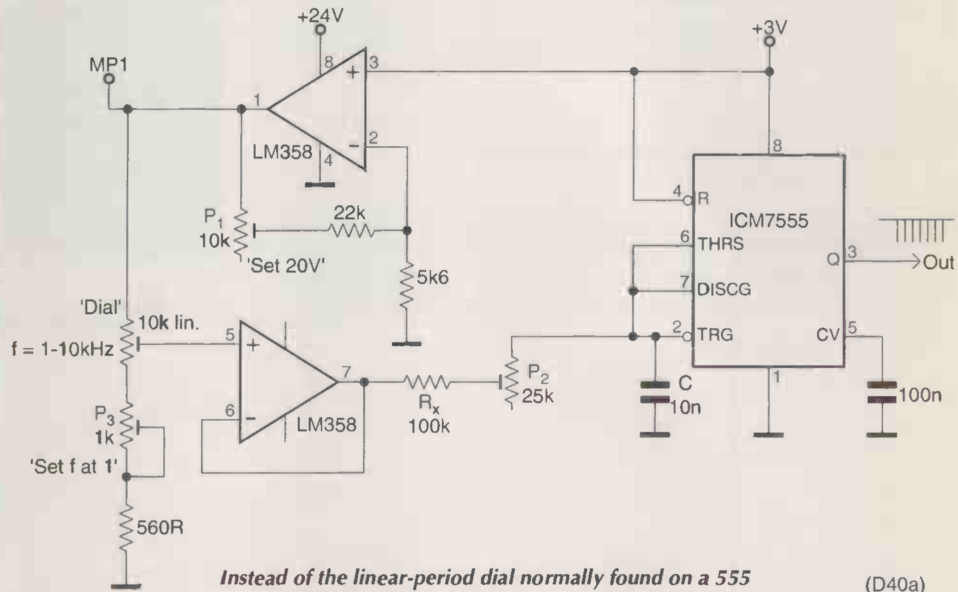
Clear hand-written notes on paper are a minimum requirement: disks with separate drawing and text files in a popular form are best – but please label the disk clearly.

555 oscillator with a linear frequency scale

Normally, the tuning dial of a 555 oscillator is linear with period, not frequency. For a linear frequency scale, it is necessary to vary the charging voltage of *C*, although it must be referred to the supply voltage of the 3V *ICM7555* used here to obtain stability.

To adjust the circuit, set the voltage on MP1 to 20V±0.1V by P_1. Dial adjustment consists of turning it to read 10 and setting maximum frequency by means of P_2; then turn it to 1 and set minimum frequency by P_3. If you find that the maximum frequency is not obtainable, vary R_x. Output is a narrow, negative-going pulse.

Ernst Schmid
München
Germany
D40a



Instead of the linear-period dial normally found on a 555 oscillator, this provides a linear frequency scale.

(D40a)

Push-button analogue voltage generator

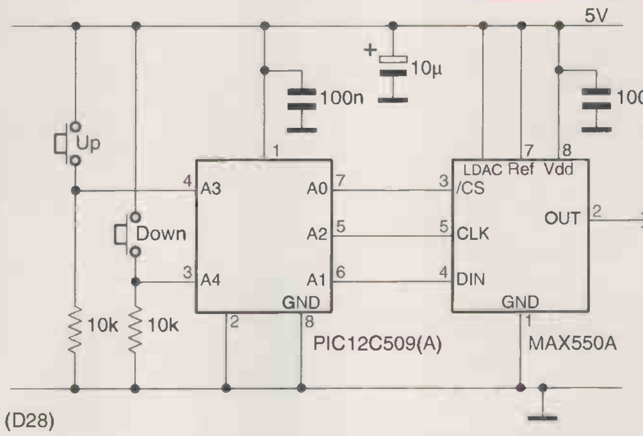
£50 WINNER

A *MAX550A* serial controlled d-to-a converter is driven by the *PIC12C509* microcontroller, which in turn is controlled by two switches, the result being a precisely set voltage output.

One switch increases the output and the other decreases it, but on reaching the limit, there are no sudden leaps from rail to rail. Pressing either button for more than two seconds increases the rate of voltage change, so that there are both coarse and fine settings. Zero is obtained by operating both switches simultaneously. The PIC code is exportable to larger systems or to other types of controller families.

Copies of the source code may be obtained by e-mail from simon-bramble@ccmail.mxim.com.

Kevin Bilke
Maxim Integrated Products
D28



(D28)

Operating two push-buttons causes the d-to-a converter to produce an accurate analogue voltage, which may be set to change rapidly or more slowly for precise setting.

Programmable Sallen and Key filter

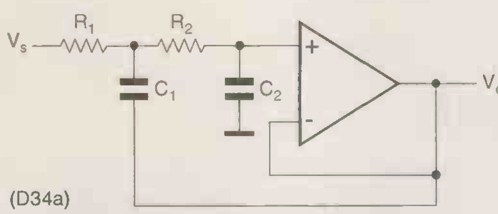
A digitally controlled potentiometer controls the Q and f_0 of a Sallen and Key low-pass filter, avoiding the critical specification of circuit components.

Figure 1 shows the usual circuit, in which the *R*s and *C*s determine frequency and Q. Since the two resistances are in series, they can be replaced by a digitally controlled Xicor *X9418* potentiometer. A quantity k may be said to represent the position of the 'wiper', zero at one end and 1 at the other.

Resolution depends on the number of programmed wiper positions, *R* representing the total resistance. The transfer function of the circuit is:

$$\frac{V_0}{V_s}=\frac{1}{s^2+s\left[1/k(1-k)RC_1\right]+1/k(1-k)R^2C_1C_2}$$

$$\frac{V_0}{V_s}=\frac{A_0\omega_0^2}{s^2+s(\omega_0/Q)+\omega_0^2},$$



(D34a)

Fig. 1. Classical Sallen and Key low-pass filter which requires close control of component values for precise performance.

which is the expression for a second-order low-pass filter, where A_0=1.

$$\omega_0=\sqrt{\frac{1}{k(1-k)R^2C_1C_2}}$$

and

£75 WINNER

$$Q=\sqrt{\frac{k(1-k)C_1}{C_2}}.$$

Using the values shown in **Fig. 2**, a theoretical Butterworth response is obtained when k=13/63, giving a Q of 0.704 and a cut-off of 6.85kHz. With the potentiometer hard over in either direction, the circuit becomes a first-order low-pass filter.

Chuck Wojslaw
Xicor Inc.
Milpitas
California
USA
D34

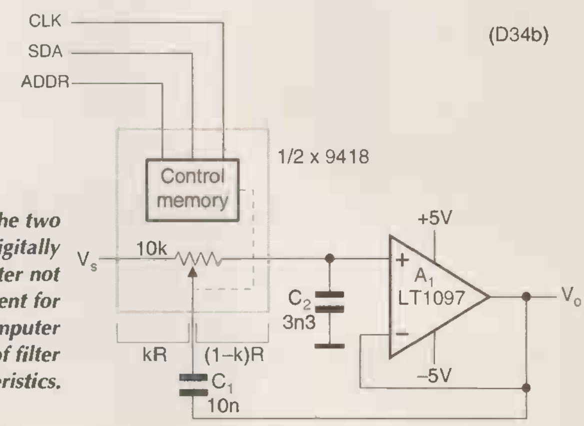
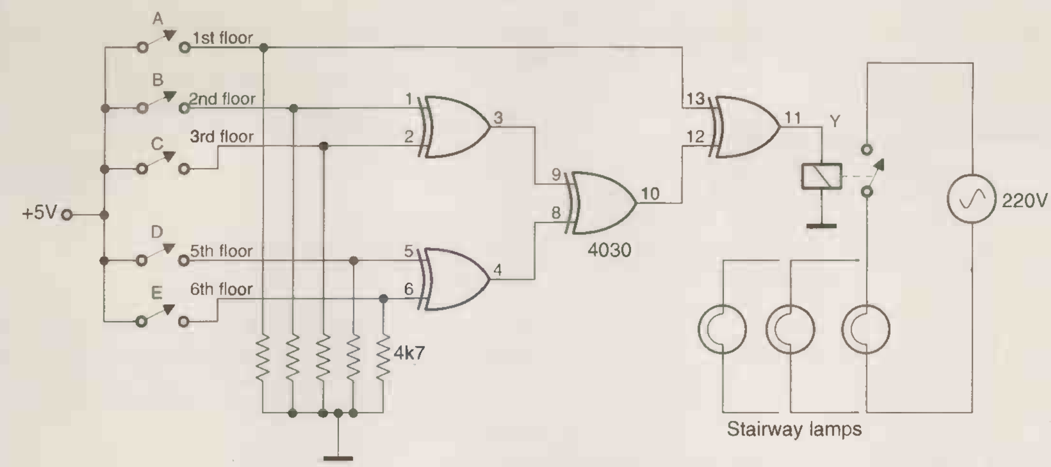


Fig. 2. Replacing the two resistances by a digitally controlled potentiometer not only allows adjustment for tolerances but also computer control of filter characteristics.

A	B	C	D	E	Y
0	0	0	0	0	0
1	0	0	0	0	1
0	1	0	0	0	1
1	1	0	0	0	0
0	0	1	0	0	1
1	0	1	0	0	0
0	1	1	0	0	0
1	1	1	0	0	1
0	0	0	1	0	1
1	0	0	1	0	0
0	1	0	1	0	0
1	1	0	1	0	1
0	0	1	1	0	0
1	0	1	1	0	1
0	1	1	1	0	1
1	1	1	1	0	0
0	0	0	0	1	1
1	0	0	0	1	0
0	1	0	0	1	0
1	1	0	0	1	1
0	0	1	0	1	0
1	0	1	0	1	1
0	1	1	0	1	1
1	1	1	0	1	0
0	0	0	1	1	0
1	0	0	1	1	1
0	1	0	1	1	1
1	1	0	1	1	0
0	0	1	1	1	1
1	0	1	1	1	0
0	1	1	1	1	0
1	1	1	1	1	1

(D45)



Y = lamp status '0' = off '1' = on
A,B,C,D,E = switch status '0' = normal off '1' = normal on

Simplifies stairway switches avoid power wastage and provide a longer lamp life.

Multi-switch stairway light

When a single light must serve several flights of stairs it is common to have a fairly complicated wiring to a number of switches or to have a delay circuit to keep the light on for a couple of minutes. This circuit avoids wiring problems and allows any switch to control the light, thereby avoiding the waste of power. in a delay device.

The circuit is simple, consisting only of a quad Xor *CD4030*, which allows up to five switches.

Ashraf Saad Awad Ebrahim
Farwaniyah
Kuwait
D45

Power-saving, switched-mode power supply

An oscillator in this smps acts a voltage output detector to control the series switch; no inductors are needed and power consumption is extremely low.

A *7413* nand, with a capacitor and diode, form the oscillator. Initially, Tr_1 has insufficient base current to open and C_1 charges until the schmitt nand triggers. Pin 6 of the 7413 goes low and the diode discharges C_1 quickly to the point at which the nand again toggles and the capacitor starts to charge again. The second half of the nand is simply a waveform shaper to drive the series-pass pair. In the condition in which Tr_1 receives base current from the potentiometer, the oscillator stops.

In this way, the setting of the potentiometer determines whether the oscillator opens and closes the series switch or blocks it, using very little power in the process, assuming that Tr_1 is a high-beta type. The red leds blink at a rate depending on the current demand and therefore function as a kind of current meter.

As an example of performance, varying output current between zero and 2A causes an output variation of no more than 20-40mV when a small C_1 is in use. This capacitor may be between 8.2nF and 10μF to provide either high speed and lower current or the reverse.

Roland Vanthomme
Sambreville
Belgium
D44

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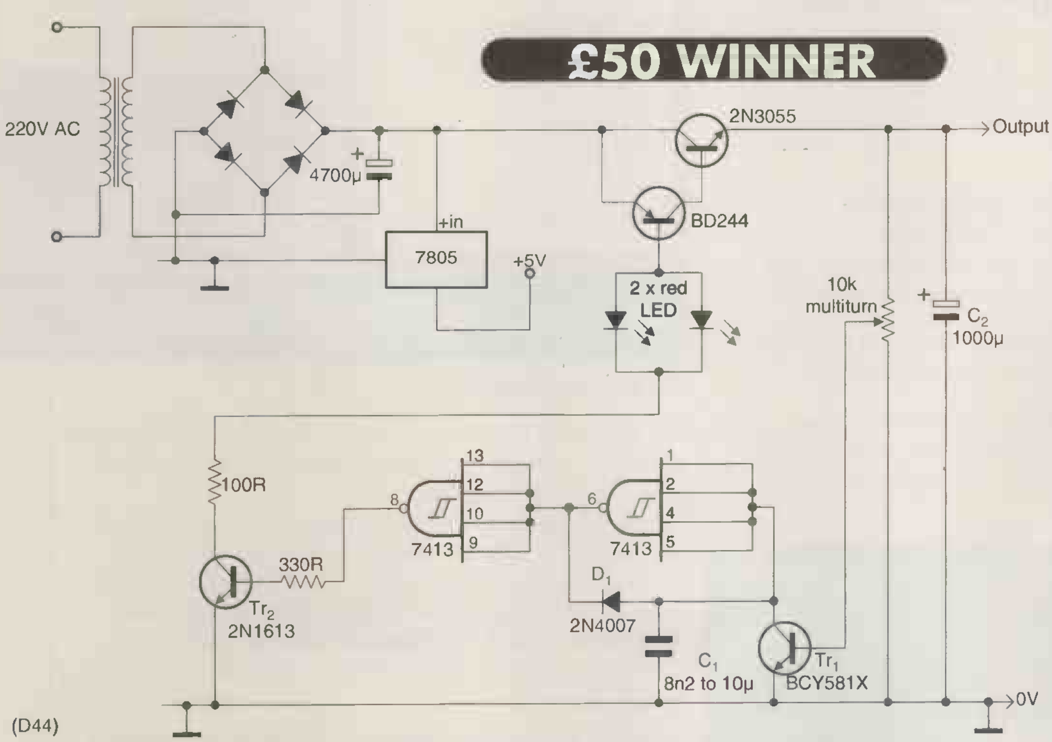


(D44)

Switched-mode power supply consumes little power and reacts to

Low-current battery monitor

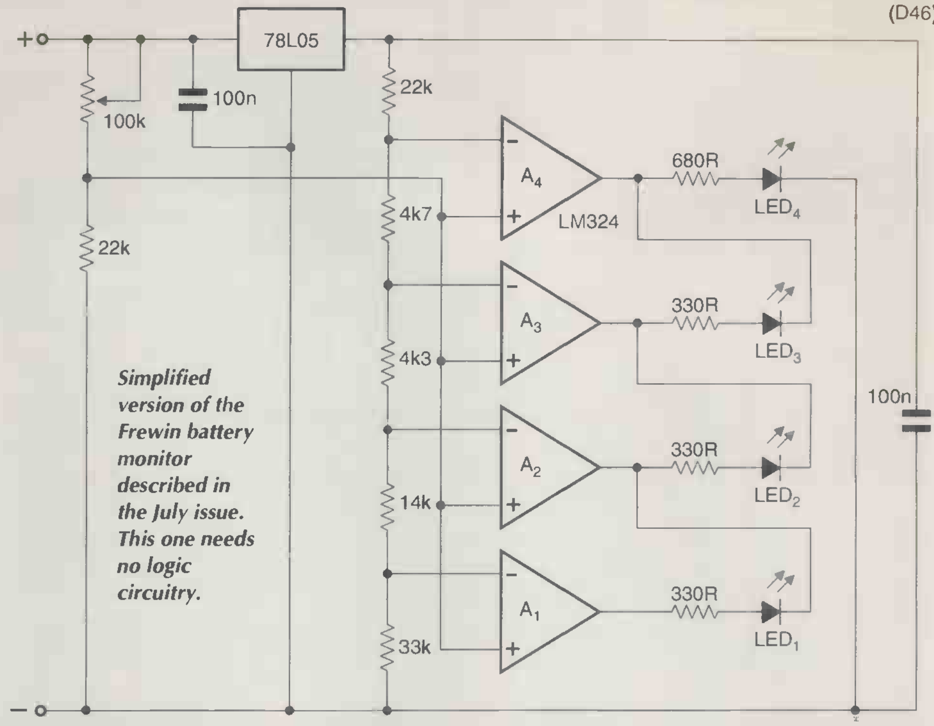
Frewin's circuit in the July issue uses Xnor gates to ensure that only one led is illuminated at a time to save battery current. In this circuit, the leds are connected between op-amp outputs and no gates are needed.

LM324 outputs supply enough current to drive the leds, using reduced value series resistors; at the 2.5mA led current used in the Frewin design, there is about 2.85V to drive leds 1-3 and 3.7V for led 4, since the op-amps source 3.7V from a 5V supply and sink 0.85V.

With the values shown, led 1 lights at 8-11.4V, led 2 at 11.4-12.4V, led 3 at 12.4-13.5V and led 4 above 13.5V

Paul Whiteley
Chester-le-Street
County Durham
D46

(D46)



Simplified version of the Frewin battery monitor described in the July issue. This one needs no logic circuitry.

Delay-length-locked loop

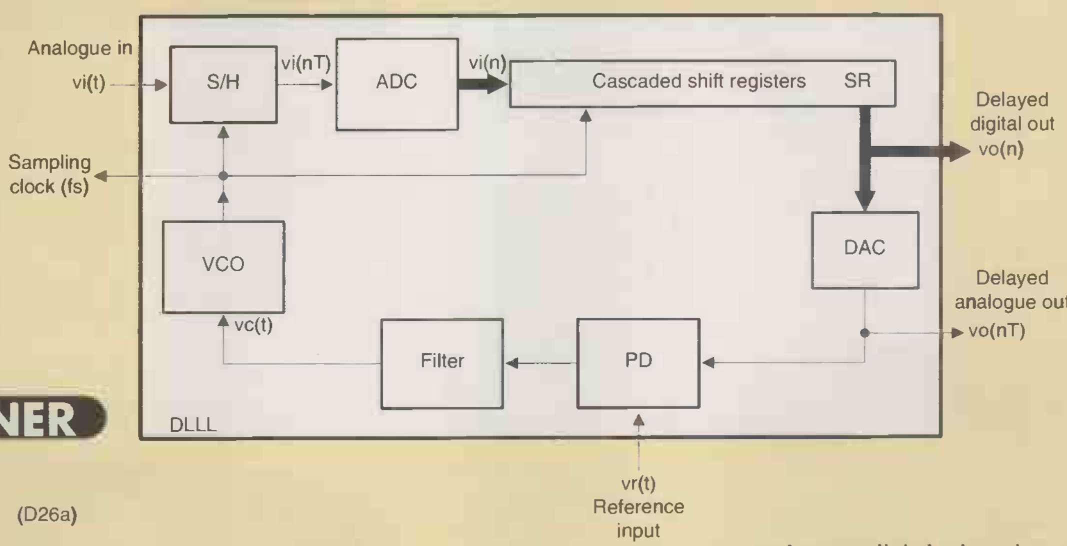
It is often necessary to be able to adjust the delay in a delay line to equal the time lapse between analogue and reference signal inputs. This note describes, by analogy with a phase-locked loop, a delay-length-locked loop or dlll, using the principle described above, in which a digital delay line is controlled in delay by a voltage-controlled oscillator. Here, there are 12 shift registers in cascade, so that the delay is v_i (n–12). A d-to-a converter produces $v_o(nT)$, the analogue of the digital signal $v_o(n)$.

A reference input becomes one input to a phase detector, the other being the delayed analogue output; for any delay between the two inputs, the pd produces a voltage proportional to the delay. This is low-pass filtered and used as the control input of the vco, the result being that the clock frequency is automatically adjusted to bring the analogue output and the reference input into coincidence, the delay line being now 'locked'.

The phase detector could use pulse-width modulation, as in **Fig. 2**, or analogue or digital subtraction, as seen in **Fig. 3**.

As regards uses, the arrangement may be used in applications where the received signal arrives after a delay from the transmitted one, as in radar, sonar and flaw detection, in which amplitude and quality of the received signal is lowish. The

Fig. 1. Delay-length-locked loop, where the analogue output is the delayed version of the analogue input, the delay being automatically adjusted to the delay between analogue input and reference input.



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(D26a)

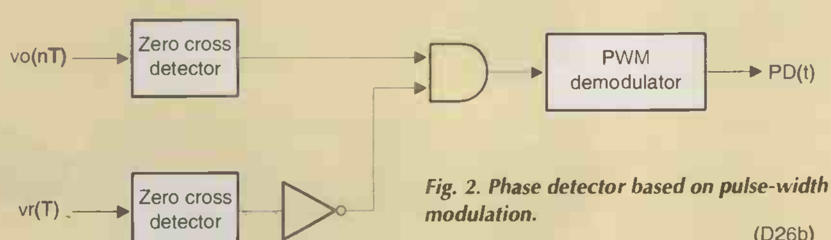


Fig. 2. Phase detector based on pulse-width modulation.

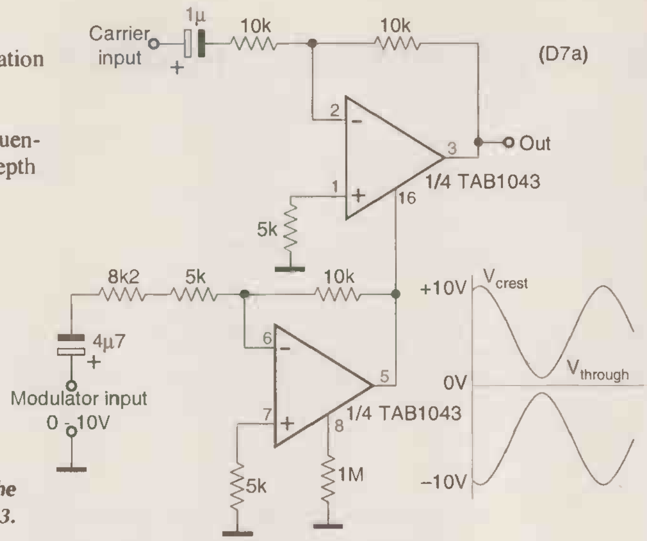
(D26b)

Audio modulator

Again using the *TAB1043* programmable op-amp, the modulation level and phase of this amplitude modulator are determined by the modulating op-amp, the mod. frequency being 10-600Hz. Modulation depth is

$$A = 1 + \frac{R_1}{R_2}$$

Kamil Kraus
Rokycany
Czech Republic
D7a

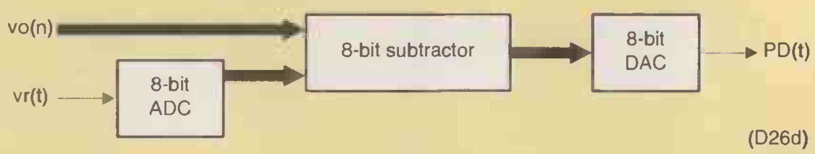


Audio modulator using the programmable op-amp TAB1043.

Details of significant new prizes for the best circuit ideas, sponsored by National Instruments, will be appearing next month – Ed.

principle is shown in **Fig. 4**, where the transmitted signal becomes the signal input of the dlll and the received signal the reference. Since the dlll locks to the delay between the two, the analogue output is the delayed version of the input.

K Balasubramanian
K B Gayathri
H Camur
European University of Lefke
Turkish Republic of Northern Cyprus
D26



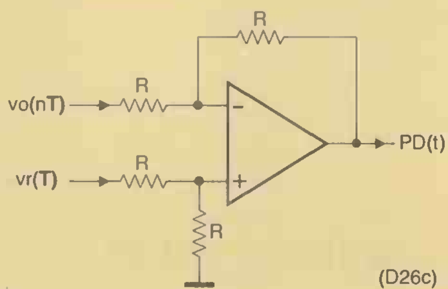
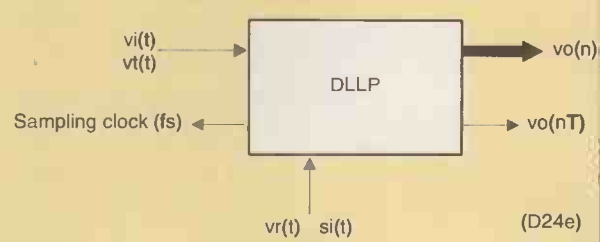


Fig. 3. Phase detection using (a) analogue subtraction and (b) digital subtraction.

Fig. 4. Using the dlll for signal detection where the received input is delayed on the transmitted one.



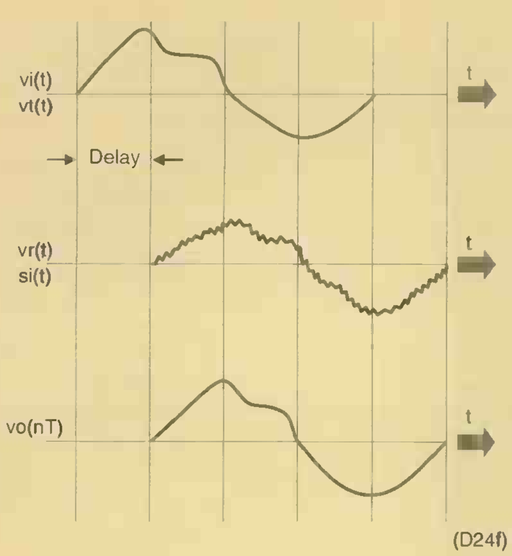
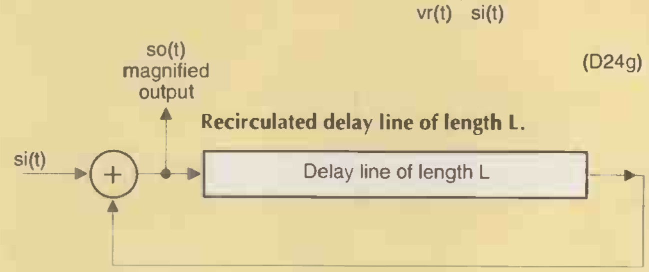
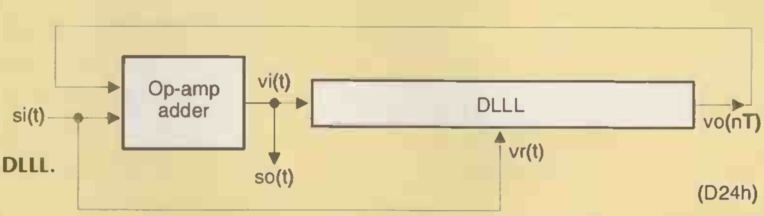


Fig. 5. Waveforms in the arrangement of Fig. 4.



Recirculated delay line of length L.



Recirculated delay line with DLLL.

Wideband photodiode amplifier

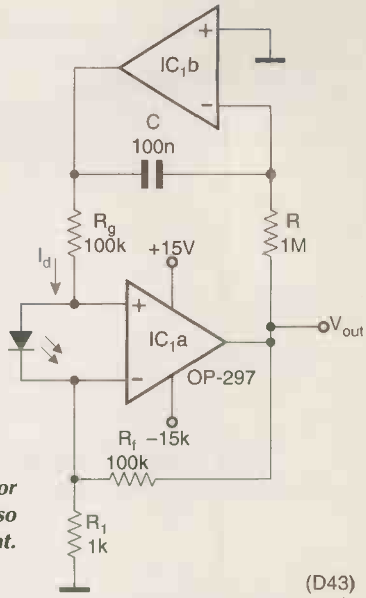
In this dc-restored photodiode amplifier, the conflict between wide bandwidth and gain is resolved, the dc restoration reducing the effect of ambient light below a time-varying signal.

Current from the photodiode flows through R_g to drop a voltage at the non-inverting input of the op-amp, where it is subject to a gain of $1+R_f/R_1$. Voltage output at mid-band is therefore,

$$V_o=I_dR_g(1+R_f/R_1).$$

Using the values shown, equivalent resistance is 10MΩ. For the dc restoration, the inverting integrator drives the restoration current through R_g, cancelling diode current at frequencies below the lf cut-off.

Michele Frantisek
Brno
Czech Republic
D43



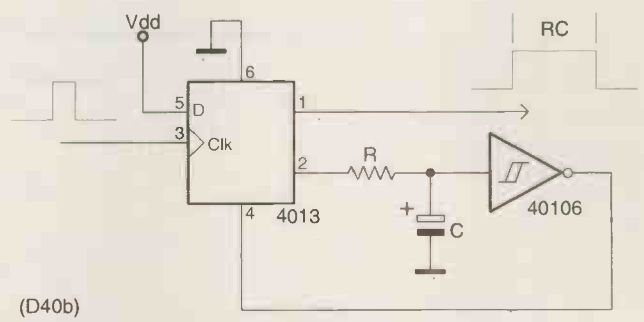
In this wideband photodiode amplifier, one resistor determines equivalent transimpedance and also provides the path for the dc restoration current.

£75 WINNER

Thrifty one-shot

For a simple and component-saving one-shot, use a spare D-type flip-flop and an inverter to give an output pulse width of *CR*, using a cmos inverter.

Ernst Schmid
München
Germany
D40b



Save components with this one-shot.

Two op-amp oscillator MkIII

An earlier design of two-integrator oscillator[1] used an inverting integrator and a non-inverting one, gain setting being by allowing the first integrator just to clip at the ±5V rails.

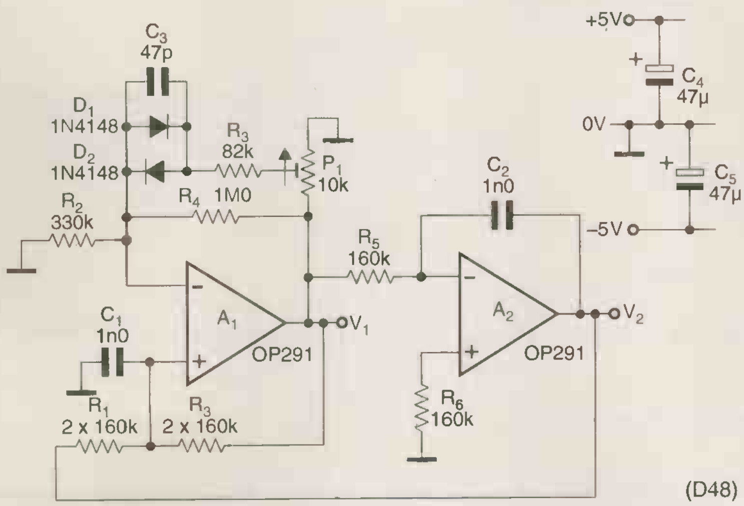
The method of gain setting used in this new version is by the use of diodes, described by Hickman and others[2]. This does cause more distortion unless measures are taken to reduce it, but the advantage is that of easier starting, as the diodes are not in conduction until oscillation is under way.

Op-amp A_1 in the diagram is the non-inverting integrator and A_2 the inverting one, A_1 being provided with increased gain by the ratio of R_4:R_2 to offset the reduction in gain caused by the diodes.

If the pot. is adjusted to give ±4V at the output, the output at V_1 is also around ±4V, together with distortion due to the diodes in the feedback path. Integrator A_2 reduces the distortion, but it is still present and is reduced by the presence of C_3 across the diodes to smooth their turn on/off.

With the values shown, frequency is 1kHz and, for a temperature change of +2°C, changes by about 0.1%; the amplitude falls by about 1%.

C J D Catto
Cambridge
D48



Latest version of the two-op-amp oscillator using diodes instead of clipping to set the gain.

References

1. Catto, C J D, 'Two op-amp sine generator', MkII. *Electronics World*, July 1999, p. 574.
2. Hickman, I, 'A perfect variable oscillator', *Electronics World*, June 1998, p. 485.

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Serial and parallel ports require no additional operating system support and offer flexibility of use when it comes to interfacing. Les Hughes examines Java's support for this type of connectivity.

Ins and outs of Java

In previous articles, I have peeked at the nature of the Java technology and investigated custom interfacing at a low level. This time, I examine how Sun's Java Communications API – also known as `javax.comm` after the package in which the code is located – provides for a simple interface to RS232 and IEEE-1248 ports on Windows PCs and Solaris workstations.

Platforms

Both Win32 and Solaris platforms support `javax.comm`. Version 2 of the package can be downloaded from `java.sun.com` free of charge. Its documentation is somewhat basic although the code examples provide help in getting started. Included with the download is a number of applications; a simple reader and writer, a black box tester, a serial port chat program, etc.

Installation is relatively simple. Some files need to be copied manually, but the instructions are clear and helpful.

Linux users are not yet supported by Sun but a third-party product in the form of the Java Communications Library is available from `http://www.interstice.com/kevinh/linuxcomm.html`. This package functions best with a newer 2.2 kernel, although certain features can be disabled during compilation to allow use with a 2.0.xx system. You will also need to obtain and install the Solaris version of `javax.comm`.

Architecture

The Java Communications API is extensible, meaning that developers are able provide support for other platforms (Linux, MAC, etc.) and interfaces (USB, ISDN, etc.) without waiting for Sun.

Within `javax.comm` is a basic communications framework around which extra classes can be built to provide support for specific hardware. As previously mentioned, this currently extends to RS232 and IEEE-1248 ports.

Central to the whole package is the `CommPortIdentifier` class. `CommPortIdentifier` is a manager class used to determine available ports, negotiate access and to open ports. Actual hardware ports are represented by classes that extend `CommPort`. `CommPort` provides high level port methods, leaving specific things such as reading and writing to a subclass; examples of such subclasses are `SerialComPort` and `ParallelComPort`.

These classes, and their associated `CommDriver`, form the actual read/write interface to the hardware.

An example

Code examples are worth a thousand words and in the case of `javax.comm`, an example shows how simple the framework is to use.

The program below is a simple application to list all of the serial ports available to you on your particular platform. This program doesn't actually check that you have the correct hardware or that it is configured. It merely provides a list of ports that *could* be managed by Java. In list 1, the first three lines import various library classes. Note the way in which `javax.comm` is imported.

Two variables, known as fields in Java speak, are declared; a `CommPortIdentifier` and an `Enumeration` – an object that allows you to traverse through a collection or list of objects. The line:

```
ports =
CommPortIdentifier.getPortIdentifiers();
```

invokes a static method in the `CommPortIdentifier` class to obtain a list – or `Enumeration` – of manageable ports. Next, spin through this enumeration, checking each entry to see if it is a serial port. If the entry is of type `PORT_SERIAL` then print out its name.

At this point, you could also check for `PORT_PARALLEL` if we were trying to identify manageable parallel ports.

Reading and writing

Reading and writing to files or devices in Java is achieved using the 'Streams' input/output model. `InputStreams` act as a source of data to your program and `OutputStreams` are a sink for data to flow into from your program. In order to read and write to `javax.comm` ports, you need to obtain an `InputStream` and an `OutputStream`.

The abstract `CommPort` class provides methods called `getInputStream()` and `getOutputStream()` that enable us to obtain the required i/o objects. Once you have these objects, you can simply call `read()` and `write()` on them to send data to your hardware.

Figure 1 shows an interaction diagram of the steps involved in obtaining input and output objects. A Java code example shows this in action, **List 2**.

The main method in this case first creates a `ReadWrite` object and then tells that object to '`doSomething()`'. This method writes three bytes (0x01,0x02,0x03) and then reads a byte from COM1 in this case.

Note Java's error catching mechanisms at work in this example with the `try{…}catch(){}` blocks.

The streams-based i/o model allows us to add extra functionality to the i/o pipe by wrapping the basic `InputStream` or `OutputStream` with other stream objects.

For example, you could connect a stream capable of handling strings, decimal numbers or even serialised objects to our i/o channel. This would allow you to send different types of data to our hardware almost transparently. The full use of streams in this manner is beyond the scope of this article but extra information can be found in the references below.

List 1. A class that displays the ports that could be managed by Java.

```
import java.io.*;
import java.util.*;
import javax.comm.*;
  public class portlister {
    public static void main(String[] args) {
    CommPortIdentifier ID;
    Enumeration ports;
ports = CommPortIdentifier.getPortIdentifiers();
  while (ports.hasMoreElements()) {
     ID = (CommPortIdentifier) ports.nextElement();
     if
(ID.getPortType()==CommPortIdentifier.PORT_SERIAL){
        System.out.println(ID.getName());
     }
   }
}
}
```

List 2. Java code example for the interaction diagram, Fig. 1, which shows the steps involved in obtaining input and output objects. It writes three bytes, then reads a byte from COM1.

```
import javax.comm.*;
import java.io.*;
public class ReadWrite {
   SerialPort port;
   CommPortIdentifier ID;
   InputStream inp;
   OutputStream out;
   public ReadWrite() {
   try{
      ID = CommPortIdentifier.getPortIdentifier("COM1");
      port = (SerialPort) ID.open("ReadWrite",1000);
      out = port.getOutputStream();
      inp = port.getInputStream();
   }catch(Exception e) {
      System.out.println("Oops..something went wrong:");
      e.printStackTrace();
      System.exit(-1);
   }
   }
   public void doSomething() {
   try{
      out.write(1);out.write(2);out.write(3);
      System.out.println("Read: "+inp.read());
   }catch(IOException io) {
      System.out.println("Oops....");
      io.printStackTrace();
   }
   }
public static void main(String args[]){
   ReadWrite theApp = new ReadWrite();
   theApp.doSomething();
}
}
```

Port specifics

Serial and parallel ports both have specific behaviours and control signals. Each specific class – `SerialCommPort` and `ParallelCommPort` provides extra methods that enable the programmer to have full control over the device.

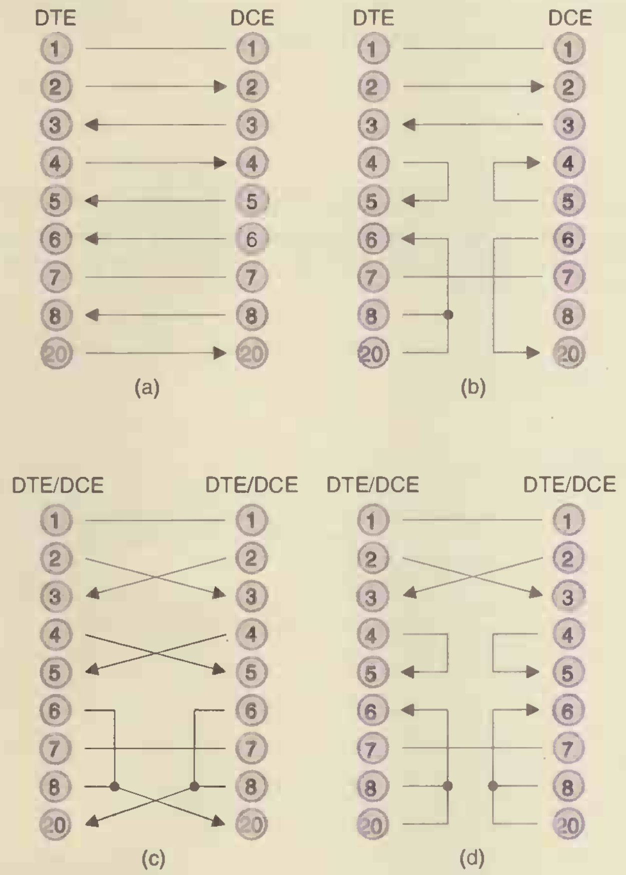
For example, the `SerialCommPort` class lets you set or check such lines as RTS/CTS, DCD, DSR, DTR, etc., while the `ParallelCommPort` class has methods for checking 'out of paper conditions', setting SPP/EPP/ECP mode, etc.

For a complete list of all available methods, refer to the API documentation included with the software download.

Following on from the serial port example above, in order to use a serial line correctly, certain parameters need to be set. To change the line speed, parity, data and stop bits and flow control discipline, use the `setSerialPortParams()` and `setFlowControlMode()` methods.

These two methods allow you to set line speed, data, stop and parity bits and to enable hardware (RTS/CTS) or software (XON/XOFF) flow control if required.

Some serial devices do not actually use the handshaking lines as the standard intends. For example, some packet radio modems use these lines to provide power. To set or to check to status of these lines, you can use such methods as setRTS(), isCTS() etc. Again, a full list is provided in the API documentation.



Zen and the Art of RS232

Serial lines can often cause problems. Is the device a DTE or a DCE ? Neither? Does your program expect and provide the correct flow control? The list is endless.

25way		9way
1	Protective GND	
2	TD	3
3	RD	2
4	RTS	7
5	CTS	8
6	DSR	6
7	signal GND	5
8	DCD	1
20	DTR	4
22	RI	9

A break out box comes in handy but failing that, these diagrams show the most common connections, from a straight-through line to a full null-modem using 25 way connectors. Also 9 to 25way translations are shown.

Configuration (a) is a straight-through cable while (c) is a null modem. In (b) and (d), each side provides its own handshaking. Pin 8 is omitted for DCE/DCE. This diagram is based on the one in Horowitz and Hill's 'The Art of Electronics,' page 725 in my edition.

Listen here

The package javax.comm also supports Java 1.1 lightweight event subscription model, as used by the AWT, Swing, Beans, etc. Although it may sound somewhat complex, event subscription is really a simple concept to understand.

When you buy *Electronics World*, you could visit the newsagent every day until the latest copy arrives, or you could take out a subscription and let the postman deliver the magazine when it becomes ready.

Subscribers to events act in this latter manner. Your program expresses an interest in certain events and then gets on with more interesting things. When an event occurs, for example data arrives on a serial line, your program is notified and can deal with the event. If, at a later time your program is no longer interested in certain events, it can cancel its subscription.

The object that registers its interests in events is called a listener and you will find many `addXXXListener` type methods all through the core API. For serial ports, you would subscribe using `addSerialEventListener()`.

Unfortunately, javax.comm only supports one listener per serial port at the present. However, when modelling the actual device on the end of the serial line, you could include a method that receives serial events and then 'multicasts' these events onwards.

Events can be triggered by a number of changes on the port. Data arriving is probably the most common, but the API allows you to receive serial events for changes to CTS, CD DSR, RI, etc.

The parallel port driver also offers this type of functionality but is somewhat restricted to error and buffer empty conditions.

Putting it all together

One of my current pet projects is an MP3 player for my car. For those of you not familiar with MP3 compression, it is sufficient to say that one CDR of MP3 compressed audio can hold about 15 'normal' audio CDs! For more information on this technology, see www.mp3.com.

The system is based on a single board PC, which runs RedHat Linux, in the boot of the car. This PC has no keyboard or monitor – but it does have an ethernet adpator – so an interface was required for mounting on the dash.

Matrix Orbital manufactures a range of LCD and VFD modules, providing an RS232 and I^2C interface, keypad driver, programmable output lines, simple graphics capabilities and much more. Details of these modules can be found at `www.matrixorbital.com` or `www.linuxcentral.com`. For the MP3 player, I chose a 20 character by 4 lines LCD module with integrated 5V regulator. Its model number is *LKD204V*.

Of course, the system software was written in Java – what else? That is where javax.comm comes in. The *LKD204V* module has an RS232 interface and javax.comm is used to provide low level communications. Higher level processing of data is achieved through the use of various classes and abstractions.

Programming in the abstract

The code mentioned in this section can be obtained from my website, given later.

One of the many tenets of object-oriented programming is to isolate things that can change from things that stay the same. Isolating areas of code allows you to unplug old code and plug in new code without breaking the system.

The serial interface is one area that could change in future. Perhaps a USB or TCP/IP aware version of the module may become available. In fact, you may well want to replace the entire user interface module, so you should allow for some

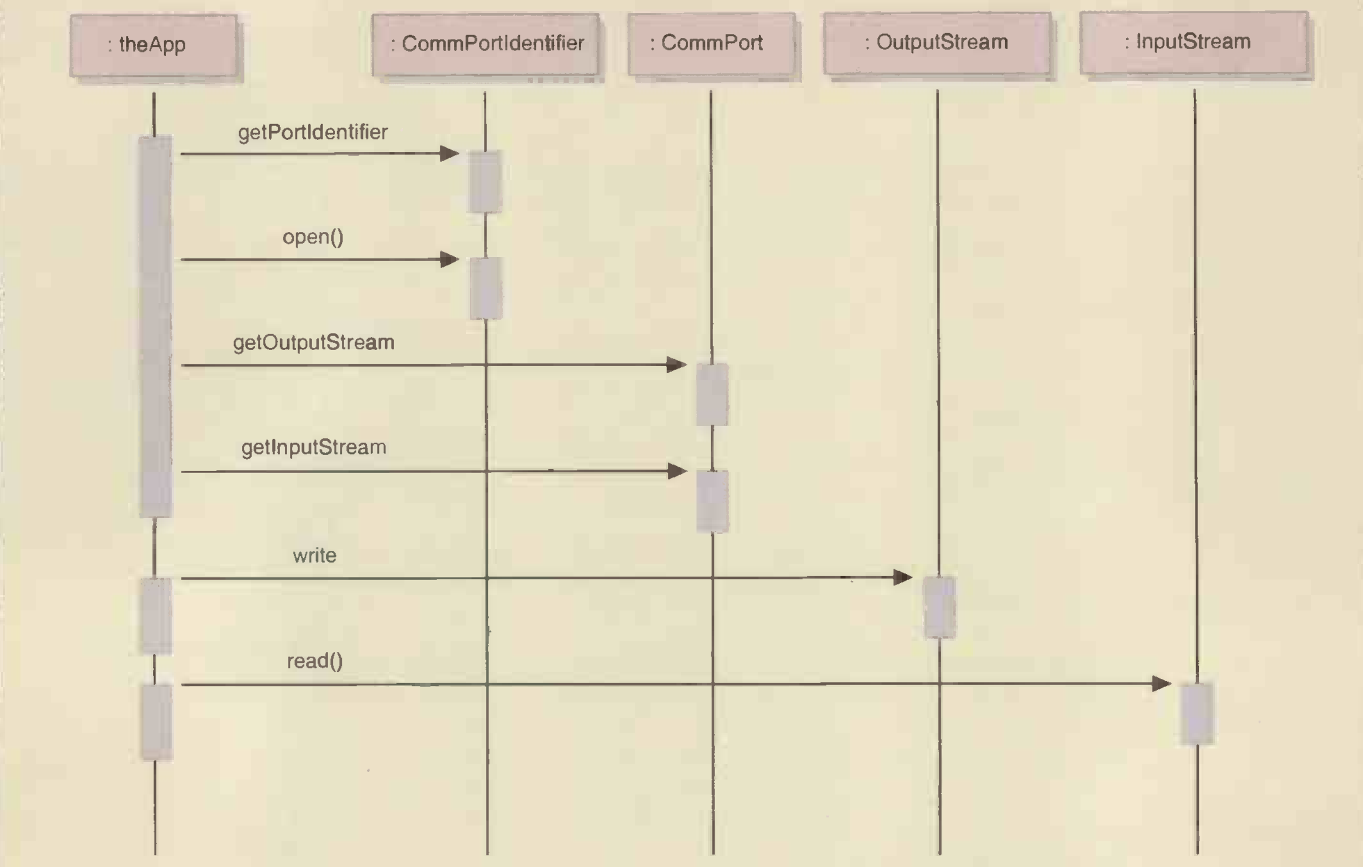


Fig. 1. Diagram illustrating interaction of the steps involved in obtaining input and output objects.

'future proofing' during your design. However, you also have to be careful not to be caught in the 'analysis paralysis' trap.

Firstly, an abstraction of the actual module was required. The class `LKD204.java` holds methods that duplicate the functionality of the module; writing text, creating custom characters, bargraphs, etc. *LKD204* is also responsible for loading a properties file from disk and creating the correct flavour of communications driver.

Each of the *LKD204* methods generates a command sequence and sends these commands to the module via an `OutputStream` obtained from the communications driver. Keypad input arrives either via an `InputStream` or through events – again obtained from the communications driver.

The `LKDCommDriver` is an abstract class, used as a 'placeholder'. In this system, the actual driver is using serial lines. By plugging in a different driver though, for I^2C for example, other communications channels become available and no other code would need changing. See the April 1999 issue of *Electronics World* for a discussion of abstract classes.

So far, I haven't mentioned the `javax.comm` serial communications side of things. In fact, all of the serial communications are factored out and encapsulated inside a concrete subclass of `LKDCommDriver`, `LKDSerialCommDriver`.

The concrete driver class for this application is called `LKDSerialCommDriver`. This class is responsible for managing serial communications with the *LKD204* module.

No other classes are aware of the fact that the module is hanging off of a serial line – it could be using the I^2C bus or even be connected to a remote serial port out on the network somewhere. `LKDSerialCommDriver` hides all of these gory details.

The complete listing for `LKDSerialCommDriver` is shown in **List 3**. The key operations are

- Obtaining a `CommPortIdentifier`
- Opening the port and obtaining a `SerialCommPort` reference
- Setting the serial port parameters
- Obtaining input and output streams
- Registering as an event listener for the port.

Again, isolating things that change, this class has no references to actual COM ports (`/dev/ttyS1` on my Linux

More information

The Java Communications API pages:
  java.sun.com/products/XXXX
JCL for Linux 2.0+
  www.interstice.com/kevinh/linuxcomm.html
Matrix Orbital
  www.matrix-orbital.com
MP3 resources
  www.mp3.com
MP3Mobile, inspiration for the author's MP3 player – written in C not Java :-(
utter.chaos.org.uk/~altman/mp3mobile
Full code examples are available from the author's web site:
www.parallax.co.uk/~leslieh

Formerly a Senior Lecturer in Software Engineering at the University of Greenwich, Les is now with Parallax, the Keane emerging technology practice (www.parallax.co.uk). He is a Sun Certified Java Programmer.

List 3. Complete listing for LKDSerialCommDriver. Key operations are, obtaining a CommPortIdentifier, opening the port and obtaining a SerialCommPort reference, setting the serial port parameters, obtaining input and output streams, and registering as an event listener for the port.

```
public class LKDSerialCommDriver extends LKDCommDriver implements SerialPortEventListener{
  SerialPort serialPort;
  CommPortIdentifier portID;
  InputStream inp;
  OutputStream out;
  public LKDSerialCommDriver(){
   try{
     portID = CommPortIdentifier.getPortIdentifier(System.getProperty("LKD204.CommPort"));
     serialPort = (SerialPort) portID.open("LKD204",1000); //1 secs timeout on open request
     int baudrate = 19200;
     try { baudrate = Integer.parseInt(System.getProperty("LKD204.BaudRate"));
     }catch(NumberFormatException n){}
     serialPort.setSerialPortParams(baudrate,
                   SerialPort.DATABITS_8,
                   SerialPort.STOPBITS_1,
                   SerialPort.PARITY_NONE);
     serialPort.setFlowControlMode(SerialPort.FLOWCONTROL_NONE);
     out = serialPort.getOutputStream();
     inp = serialPort.getInputStream();
     serialPort.addEventListener(this);
     serialPort.notifyOnDataAvailable(true);
   }catch(TooManyListenersException a){
     a.printStackTrace();
   }catch (PortInUseException piue) {
     piue.printStackTrace();
     System.exit(-1);
   }catch(NoSuchPortException nspe) {
     nspe.printStackTrace();
     System.exit(-1);
   }catch(UnsupportedCommOperationException ucoe) {
     ucoe.printStackTrace();
     System.exit(-1);
   }catch(IOException e){
     e.printStackTrace();
     System.exit(-1);
   }
  }
  public InputStream getInputStream() { return inp;}
  public OutputStream getOutputStream() {return out;}

  public void serialEvent(SerialPortEvent e){
   int data;
   Vector local = (Vector)listeners.clone();
   switch(e.getEventType()){
   case SerialPortEvent.DATA_AVAILABLE:
     try{
      data = inp.read();
     }catch(IOException ioe){
      return;
     }
     break;
    default:
     return;
   }
   for(Enumeration en = local.elements();en.hasMoreElements(); )
     {
      LKDEventListener lis = (LKDEventListener)en.nextElement();
      LKDEvent ev = new LKDEvent(LKDEvent.KEY_EVENT,data-64, this);
      lis.LKDEvent(ev);
     }

  }
}
```

List 4

```
 import LKD204.*;
 public class HelloLKD implements LKDListener {
   LKD204 lkd;

   public HelloLKD(){
   lkd = new LKD204();
   lkd.addListener(this);
      lkd.clear();
      lkd.setText("Hello World");
   }
   public void LKDEvent(LKDEvent ev) {
      System.out.println("Got an LKDEvent!");
      System.out.println("You pressed key" + ev.getData());
   }
 public static void main(String args[]) {
      HelloLKD theApp = new HelloLKD();
 }
 }
```

PC or COM2 on Win32). Instead, it relies upon entries in a properties file. By using this configuration file approach, code can be moved between platforms without the need for re-compilation: the installation program need only provide a correct properties file.

The final thing that the `LKDSerialCommDriver` does is register as a listener for events on the serial port. On receiving an event from the actual RS232 driver, it reads the port and the fires another event to any objects interested in `LKDEvents`. In this way, the single listener limitation imposed by the serial port driver is circumvented. This all happens in the `serialEvent()` method.

Using the package

As ever, a 'Hello World' program demonstrates the use of the five LKD classes:

- `LKD204`,
- `LKDCommDriver`
- `LKDSerialCommDriver`
- `LKDEvent`
- `LKDEventListener`.

The program is shown in **List 4**. You can see that by encapsulating all of the gory details for each sub-section, gives the module a clean interface. You could now extend this idea further to encapsulate the LKD204 module into a user-interface framework. Again, this would allow you to unplug our present UI and to drop in for example, a touch screen LCD flat panel.

In summary

Communications provided by the Java Communications Framework is still in its infancy. Support is limited to Solaris and 32-bit Windows systems, although third party developers have ported to other platforms. In order to protect you programs from future changes and enhancements it is still necessary to provide an isolation layer by wrapping the `javax.comm` classes in your own code as demonstrated.

However, the `javax.comm` package provides a simple interface to two of the most popular types of port and affords the developer some of the platform independence and other benefits of the Java technology. ■

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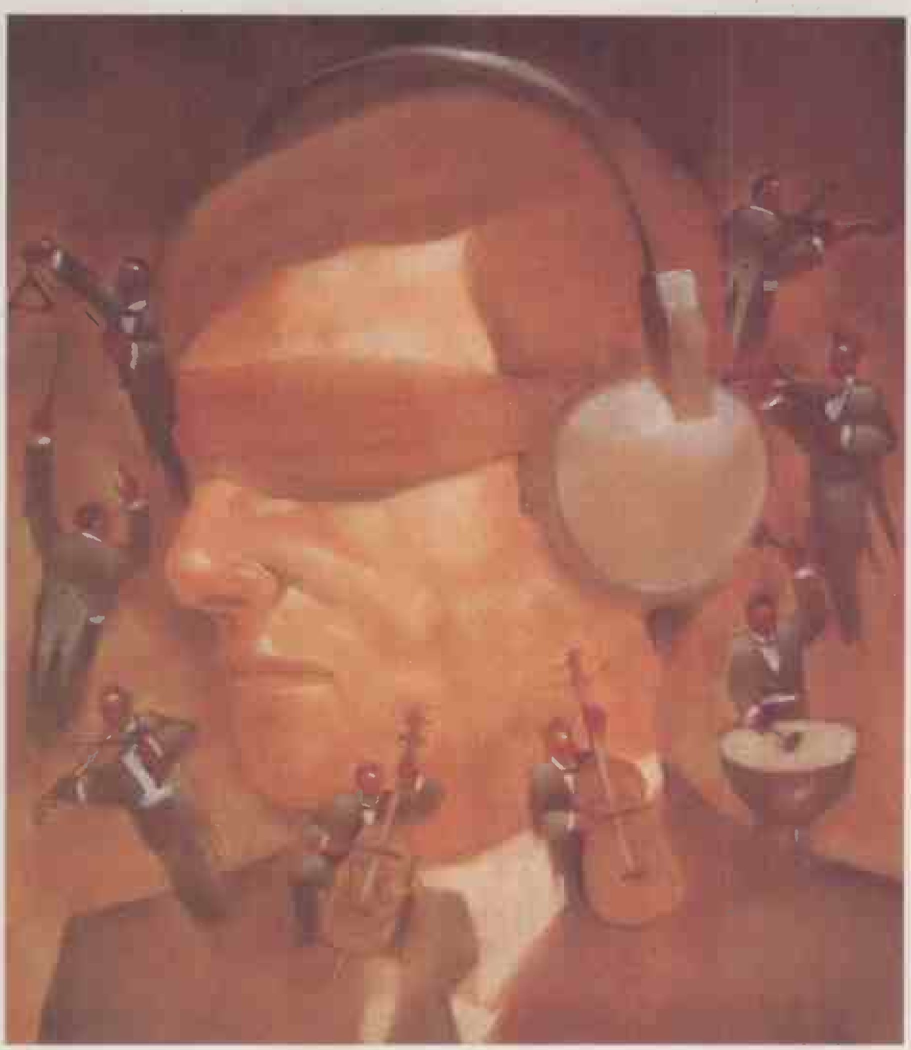


In the third article analysing stereophonic reproduction and the way that we receive and perceive sound, John Watkinson considers how the stereo illusion is created.

Stereo from all angles III

The term stereophony is derived from the Greek for 'solid sound' and is today invariably abbreviated to stereo. Stereo is based on two simultaneous audio channels feeding two spaced loudspeakers. The best listening arrangement for stereo is shown in **Fig. 1** and is where the speakers and the listener are at different points of a triangle, which is almost equilateral.

Stereophony works by creating differences of phase and time of arrival of sound at the listener's ears. My last article showed that these are the most powerful hearing mechanisms for determining direction.

Figure 2a) shows that this time of arrival difference is achieved by producing the same waveform at each speaker simultaneously, but with a difference in the relative level, rather than phase. Each ear picks up sound from both loudspeakers and sums the waveforms.

The sound picked up by the ear on the same side as the speaker is in advance of the same sound picked up by the opposite ear. When the level emitted by the left loudspeaker is greater than that emitted by the right, you will see from Fig. **2b)** that the sum of the signals received at the left ear is a waveform which is phase advanced with respect to the sum of the waveforms received at the right ear.

If the waveforms concerned are transient, the result will be a time of arrival difference. These differences are interpreted as being due to a sound source left of centre.

Figure 3 shows that the apparent position of the virtual sound source is a function of the level difference between the channels. This defines the characteristics that a microphone must have.

As I will explain in more detail in a subsequent article, a stereo microphone must produce a pair of in-phase signals whose relative amplitude varies with direction according to Fig. 3. If this is achieved, the spatial disposition of sound sources with respect to the microphone will be the same as the spatial disposition of virtual sources between the speakers.

Note that with only two channels, virtual sound sources can only be created between the speakers. At the recording venue one hears sound from all around due to reverberation, but two speakers cannot recreate this. Stereo relies on some reverberation in the listening room to make up for that.

If you don't consider two channels to be good enough, then more channels will have to be used. The number of channels that can be used is anything from three to infinity.

Two channels is the best compromise

The fact remains though that two channels done properly gives by far the best realism to complexity and cost ratio. For a given budget it is entirely possible that a four channel surround sound system could sound a lot worse than a stereo because each loudspeaker can now only cost half as much.

The stereophonic illusion only works properly if the two loudspeakers are producing in-phase signals. In the case of an accidental phase reversal, the spatial characteristic will be ill-defined and lack images. At low frequencies the two loudspeakers are in one another's near field and so antiphase connection results in bass cancellation.

As the apparent position of a sound source between the two speakers can be controlled solely by the relative level of the sound emitted by each one, this format is called intensity stereo.

Moving sources by pan-potting

It is possible to 'steer' a monophonic signal from a single microphone into a particular position in a stereo image using a form of differential gain control. **Figure 4** shows that this device, known originally as a panoramic potentiometer, or today as a pan-pot, will produce equal outputs when the control is set to the centre.

If the pan-pot is moved left or right, one output will increase and the other will reduce, moving or panning the stereo image to one side.

Pan-potted audio can never be as realistic as the results of using a stereo microphone because the pan-pot causes all of the sound to appear at one place in the stereo image. In the real world the direct sound should come from that location but reflections and reverberation should come from elsewhere. As a result, artificial reverberation has to be used on pan-potted mixes.

The mechanism of Fig. 2 only works as described at low to medium frequencies. As frequency rises the sound from the left speaker to the right ear and *vice-versa* will be attenuated in level due to shading by the head.

This phenomenon is shown in **Fig. 5a)**. Note that as the wavelength of sound falls the head becomes a more significant obstacle. The result is that high-frequency sounds appear to have come from further apart in the stereophonic image than the rest of

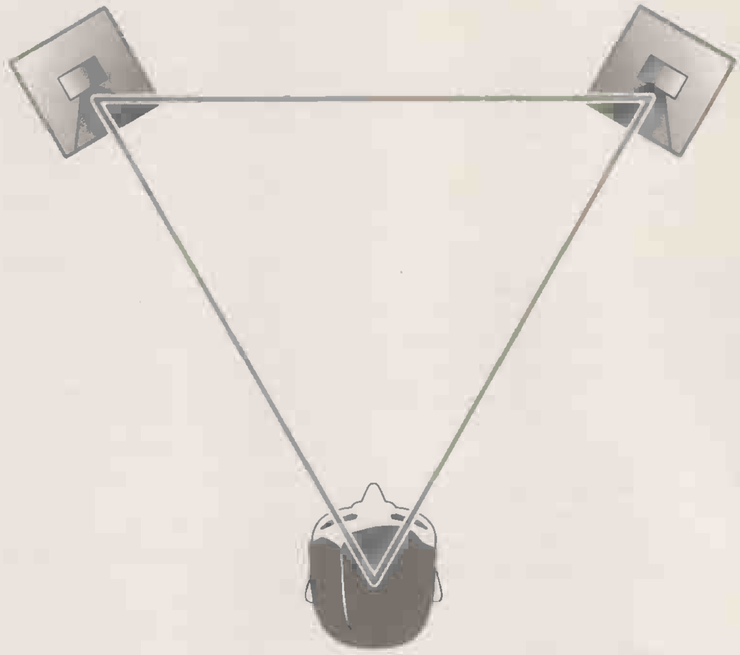


Fig. 1. Stereo listening arrangement requires a near equilateral triangle. Higher quality equipment allows the speakers to be further apart. With poor equipment, this results in a 'hole in the middle' effect.

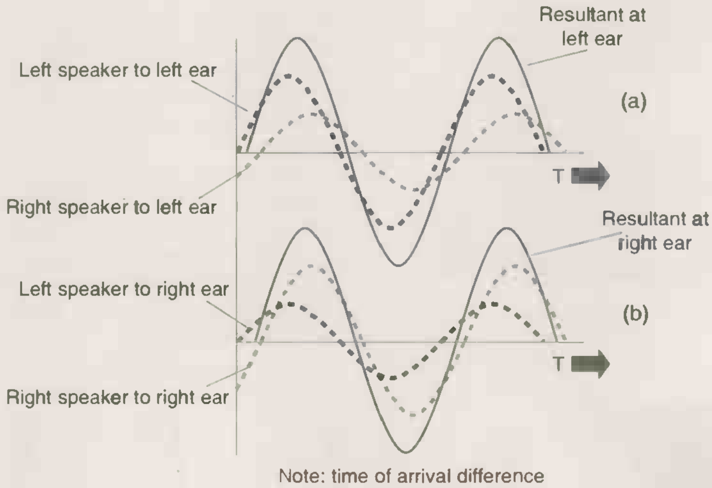


Fig. 2a). A virtual source partially offset to the left produces a larger amplitude signal in the left speaker than the right. As both ears hear both speakers, but with time delays on the 'opposite' ear, the result is a time-of-arrival difference which the ear uses as a directional clue.

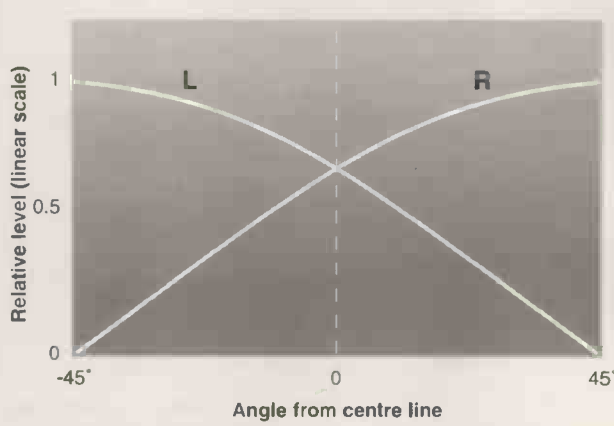


Fig. 3. Apparent position of the virtual sound source is a function of the level difference between the channels. This defines the characteristics that a microphone must have.

the spectrum. This is a form of image smear which is shown in Fig. 5b).

The effect is that wideband sound sources grow wider as they are panned to the extremes of the sound stage. This shading smear can easily be distinguished from loudspeaker smear due to poor speakers because that is independent of pan position as Fig. 5c) shows.

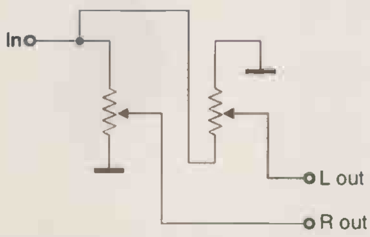
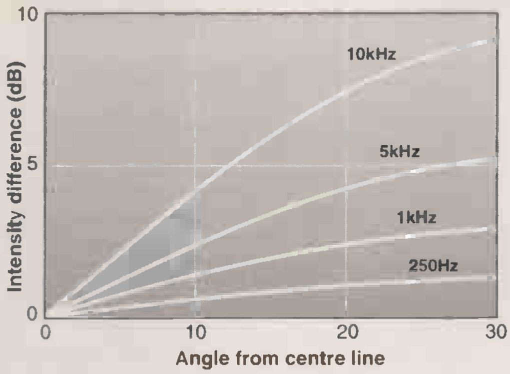


Fig. 4. It is possible to 'steer' a monophonic signal from a single microphone into a particular position in a stereo image using a form of differential gain control called a pan-pot, producing the function of Fig. 3.

Fig. 5. At a), you can see that shading cases a greater inter-aural intensity difference as frequency rises. This causes smear on off-axis images, as shown in b). This is distinct from the smear due to poor speakers, which is shown in c).



(a)

Blumlein was aware of the smear

Shading smear was known to Alan Blumlein and the stereosonic sound system that he designed contained a compensator for it. This is basically a frequency-sensitive stereo width control which electrically narrows the high-frequency image by the same amount as the shading widens it.

Shading smear is unavoidable and occurs in all stereo speaker systems. The need for compensation is universal, as is the audibility of the improvement that results. Although Blumlein knew about shading smear and published the solution in papers, the number of commercially available systems which incorporate it is tiny.

An analogue shading compensator can be implemented by introducing controlled frequency-dependent HF crosstalk between channels.

It is sometimes said that people prefer vinyl discs to CDs because they have crosstalk. Plausible as the argument seems, unfortunately it's a myth. While vinyl discs do indeed have crosstalk, the amounts are far too low to have any effect on imaging and the crosstalk does not vary with frequency in the appropriate way. As a result when properly set for the subtended angle of a particular pair of speakers, a shading compensator gives a similar degree of improvement on CD, vinyl and tape sources.

If this old wives' tale were correct, you would expect a shading compensator to make vinyl reproduction worse, but it doesn't.

Headphone listening and shuffling

For practical reasons, audio engineers sometimes have to use headphones for monitoring. Domestic listening on headphones minimises disturbance to others.

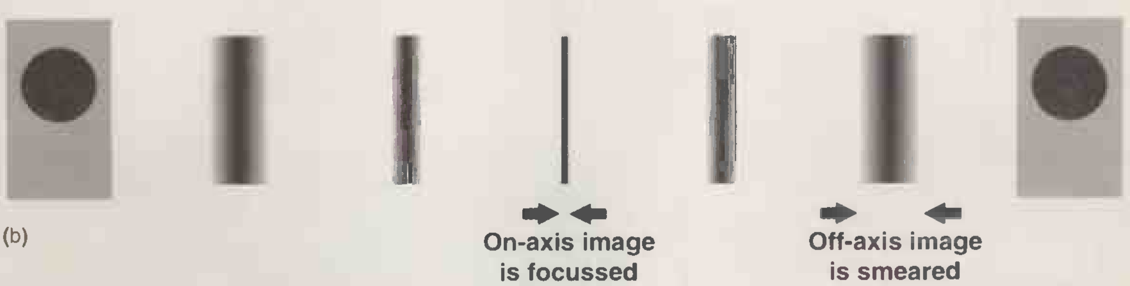
Unfortunately conventional headphones are of no use for assessing the spatial characteristics of a stereo signal. Conventional headphones prevent both ears receiving both channels and so it should be clear that the result of Fig. 2 cannot be obtained.

Anyone who has worn conventional headphones will know that there is no similarity to the sound stage produced by speakers. Instead the sound appears, quite unrealistically, to be inside the listener's head.

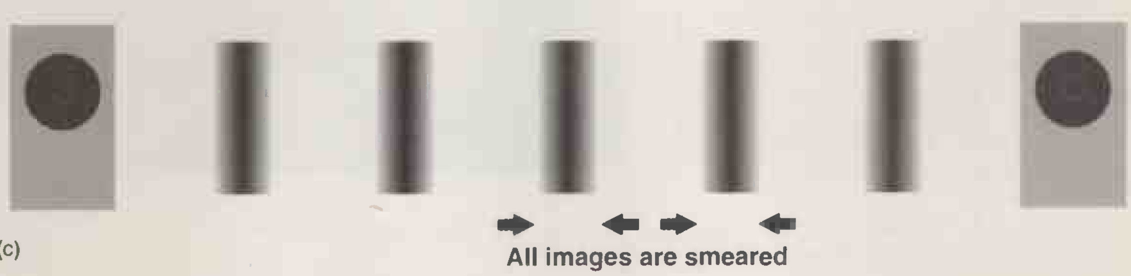
Headphones can readily be made compatible with intensity stereo signals intended for loudspeakers using a signal processor known as a shuffler. This device, again devised by Blumlein, simulates the cross-coupling of loudspeaker listening so that both ears receive both signals once more.

Figure 6 shows that this is done by feeding each channel to the other ear via a delay and a filter. The delay simulates the additional path length to the distant ear and the filter attenuates high frequencies to simulate the effect of head shading.

The result is a sound image that appears in front of the listener so that



(b)



(c)

decisions regarding the spatial position of sources can be made. Although the advantages of the shuffler have been known for decades, the information appears to have eluded most equipment manufacturers. To my knowledge, the only commercially available headphone shuffler is made by Sennheiser.

You would particularly expect that a shuffler would be fitted in audio devices specifically designed for headphone use, such as personal cassette and CD players, but this is not the case. Over time one learns that the audio industry generally prefers tradition and empiricism to theoretical knowledge.

As an aside, highly realistic results can be obtained, on conventional headphones only, using the so-called dummy head microphone which is a more or less accurate replica of the human head with a microphone at each side. These will be considered in a subsequent article.

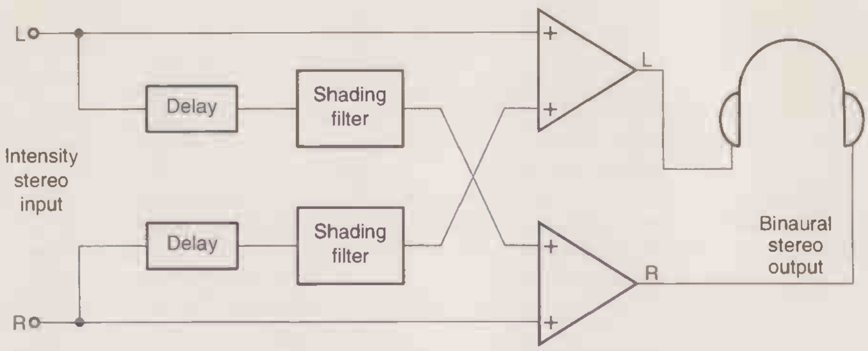


Fig. 6. Headphone shuffler results in a forward image with intensity stereo inputs.

Analysing stereo via vectorscope

When trying to obtain the best results in any endeavour, the process is always easier when objective measurement tools are available. This is particularly true in audio where the impressions gained by listening can only be subjective.

The audio vectorscope is a useful tool which gives a lot of spatial information in stereo systems. If an oscilloscope is connected in X, Y mode so that the L signal causes vertical beam deflection and the R signal causes lateral deflection, **Fig. 7a)** shows that the result will be a trace which literally points to the dominant sound sources in the stereo image.

Unfortunately in this display, the straight ahead sound source, which produces identical L and R signals, results in a line inclined at 45 degrees to the horizontal and this makes interpretation difficult.

Figure 7**b)** shows the solution. The coincident stereo signals L and R are passed through a sum and difference unit which produces two signals, M and S. The M, or Mid, signal is the sum of L and R whereas the S, or Side, signal is the difference between L and R. The sums and differences are divided by two to keep the levels correct.

When signals in the M, S format are supplied to the X and Y inputs of an oscilloscope the straight ahead condition results in a display pointing straight up which is much better ergonomically.

Self-contained audio vectorscopes are available that perform the sum and difference processing internally. It is also possible to employ a unit that synthesises a video signal containing the vectorscope picture. This can then be keyed into the video signal of a convenient picture monitor.

With an audio vectorscope, visual estimation of the width of the stereo image and the disposition of sources within it is possible. An out-of-phase condition causes the trace to become horizontal. Intensity stereo signals should ideally not contain phase shifts between the channels. Any phase shifts would result in the vectorscope displaying a Lissajous figure of some kind, instead of straight lines.

Other uses for M & S

The mid and side signal format has many further uses in stereo systems. In audio production, the apparent width of the stereo image may need to be adjusted, especially in television to obtain a good audio transition where there has been a change of shot or to match the sound stage to the picture. This can be done using M, S stereo and manipulating the S signal gain. Following this a second sum and difference unit is used to return to L, R format for monitoring.

Figure 7**c)** shows why this works. Considering a vectorscope display, the M signal produces a forward vector and the S signal produces a sideways vector. If the S gain is reduced, an off centre source will move inwards in the reproduced sound stage. The converse is not true. Increasing the S gain above unity causes sources on the edge of the sound stage to become anti-phase.

The use of M, S techniques is especially useful in microphones and this will be detailed in my next article.

And for mono listeners?

While almost all fixed audio equipment is now stereo, the portable radio or television set may well remain monophonic for some time to come. As a result, it will be necessary to consider the mono listener when making stereo material.

There is a certain amount of compatibility between intensity stereo and mono systems. If the S gain of a stereo signal is set to zero, only the M signal will pass. This is

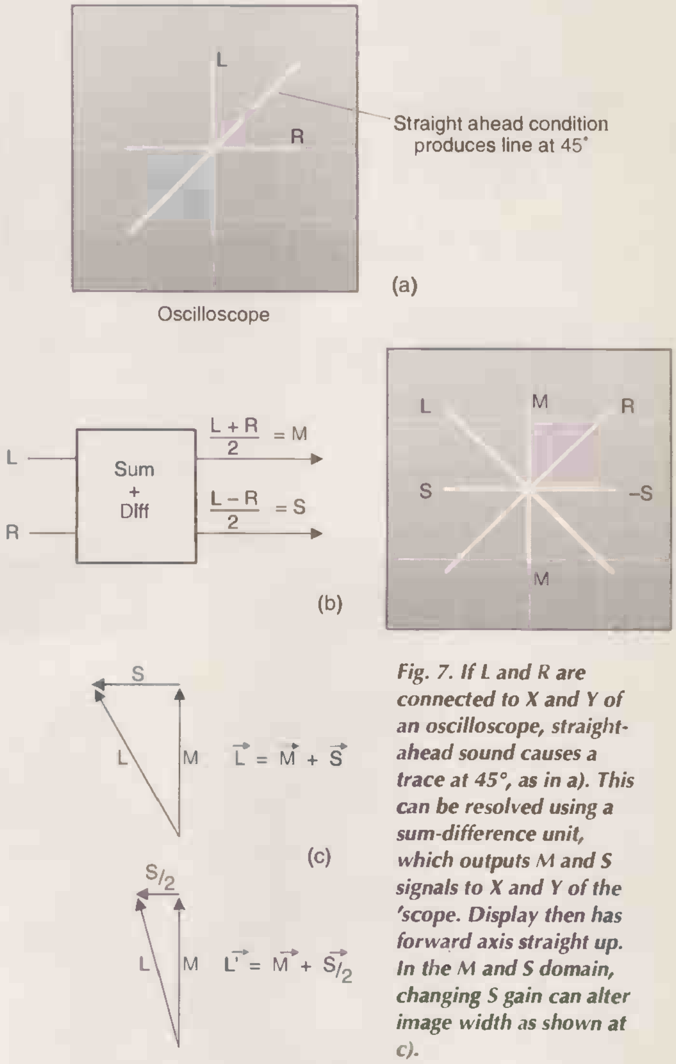


Fig. 7. If L and R are connected to X and Y of an oscilloscope, straight-ahead sound causes a trace at 45°, as in a). This can be resolved using a sum-difference unit, which outputs M and S signals to X and Y of the 'scope. Display then has forward axis straight up. In the M and S domain, changing S gain can alter image width as shown at c).

the component of the stereo image due to sounds from straight ahead and is the signal used when monophonic audio has to be produced from stereo.

Sources positioned on the extreme edges of the sound stage will not appear as loud in mono as those in the centre and any antiphase ambience will cancel out, but in most cases the result is adequate. Clearly an accidental situation in which one channel is phase reversed is catastrophic in mono as the centre of the image will be cancelled out. Stereo signals from spaced microphones generally have poor mono compatibility because of comb filtering.

One characteristic of stereo is that the viewer is able to concentrate on a sound coming from a particular direction using attentional selectivity (the cocktail party effect). Thus it will be possible to understand dialog, which is quite low in level even in the presence of other sounds in a stereo mix.

In mono the listener will not be able to use spatial discrimination and the result may be reduced intelligibility, which is particularly difficult for those with hearing impairments. Consequently it is good practice to monitor stereo material in mono to check for acceptable dialog.

A mono signal can also be reproduced on a stereo system by creating identical L and R signals, producing a central image only. While there can be no real spatial information most people prefer mono on two speakers to mono on a single speaker, probably because more complex reverberation is created in the listening room.

Know left from right

In stereo systems is it important that the left and right channels display the same gain after line up. It is also important that the left and right channels are not inadvertently exchanged, and that that both channels have the same polarity.

In some stereo equipment a twin PPM is fitted, having two needles operating coaxially. One is painted red (L) and the other green (R). In stereo line up tone, the left channel may be interrupted briefly so that it can be distinguished from the right channel. The interruptions are so brief that the PPM reading is unaffected.

Unfortunately the twin PPM gives no indication that the unacceptable out of phase condition exists. A better solution is the 'twin-twin' PPM which is two coaxial PPMs, one showing L, R and one showing M, S. When lining up for identical channel gain, obtaining an S null is more accurate. Some meters incorporate an S gain boost switch so that a deeper null can be displayed.

When there is little stereo width, the M reading will exceed the S reading. Equal M and S readings indicate a strong source at one side of the sound stage. When an antiphase condition is met, the S level will exceed the M level.

The M needle is usually white, and the S needle is yellow. This is not very helpful under dim incandescent lighting, which makes both appear yellow. Exasperated users sometimes dismantle the meter and put black stripes on the S needle.

In modern equipment the moving coil meter is thankfully giving way to the bargraph meter which is easier to read. ■

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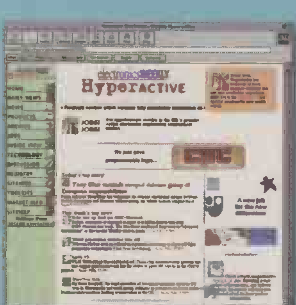

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Papst
Tel: 01264 333388
Enquiry No 502

Bubble chokes

From Schaffner EMC are two PCB-mounting buckle chokes. The EV vertical and EH horizontal are common-mode suppression chokes with current ratings up to 5A. Typical

dimensions are 32 x 22 x 35mm for a 5A-rated EV28 device or 29.5 x 30 x 22.5mm for a similarly rated EH28.
Schaffner EMC
Tel: 01189 770070
Enquiry No 503

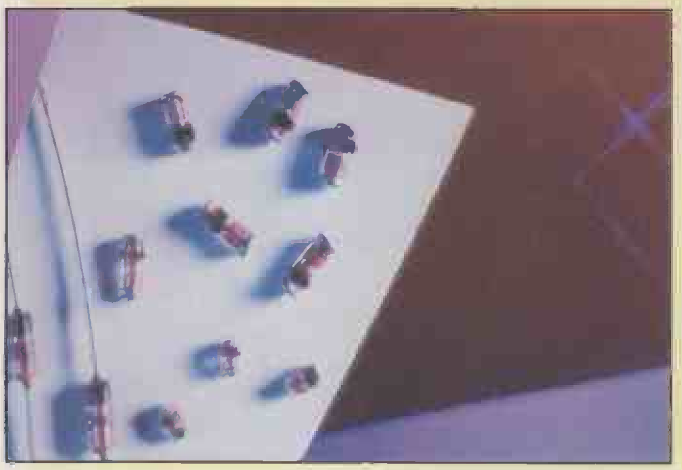
SIM connector

A connector for use with SIM cards has been introduced by AMP. Applications include mobile phones and epos equipment. It is a stripped-down version of current connectors for the same application, and lets SIM cards be inserted and removed by hand.
AMP
Tel: 01189 770070
Enquiry No 504

Twenty Schottkys in glass

Vishay has introduced 20 Schottky glass diodes, including the first devices in the firm's Micromelf package. Applications include computers, cell phones and data storage systems. They combine a choice of 30, 40, 50 and 60V maximum reverse voltages with four package options – DO35, SOD80 Minimelf, SOD80 Quadromelf and Micromelf.

Typical forward voltage is from 320mV for the 30V devices to 410mV for the 40 and 60V devices.
Vishay Semiconductor
Tel: 001 610 644 1300
Enquiry No 501



RF coils

TFC encapsulates its moulded RF coils to provide mechanical stability. They can be supplied as fixed value or adjustable with a shielding can. The moulding material is polypropylene and the shielding cans are plated brass. Core material for the adjustable versions can be ferrite, carbonyl, brass or aluminium.
Total Frequency Control
Tel: 01903 740000
Enquiry No 505

Modem chip set

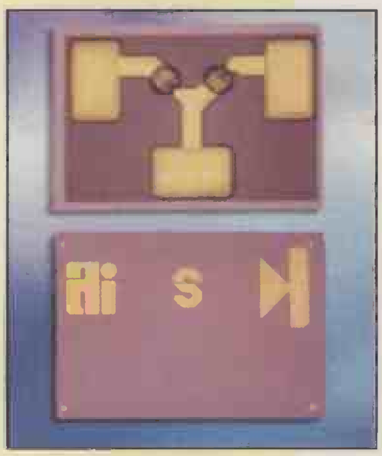
Semtech has released a modem chipset which delivers full duplex operation over a single fibre cable and allows sharing of up to 8.44Mbit/s bandwidth between four fully independent transmission channels. With the firm's ACS406 chipset, users can achieve 8.44Mbit/s on a single E2 channel or divide the bandwidth into four separate E1 channels at 2.04Mbit/s each with independent clock domains for each channel. The modem chipset takes advantage a proprietary Ping-Pong time division duplexing architecture, which achieves full-duplex operation over a single fibre channel. The ACS406 is comprised of the ACS9010 analogue chip, providing the laser and LED driver, and other analogue circuitry and the digital counterpart, ACS4060 which provides the logic necessary for time compression and decompression of data.
Thame Components
Tel: 01844 261188
Enquiry No 507

Miniature relay

Measuring 15.4mm high by 10.2mm wide, Finder's 43 miniature relay is for PCB applications such as alarm systems and medical equipment. Rated at 10A 250V, the relay is a single-pole double-throw device. Coil supply voltage is from 3 to 48V, and

the coil dissipates 250mW, making it suitable for low-powered control signals, such as those sent from PLCs, and battery operated circuits.

The standard unit is sealed to IP40, although protection to IP67 is available. Operating range is –40 to +85°C. It can be flow soldered, but a PCB mounting socket is available as an option.
Finder Components
Tel: 01785 818100
Enquiry No 508



GaAs diodes as a series pair

Alpha Industries has announced a GaAs diode in a series pair configuration for balanced and double balanced mixers. The DMK8001 is a single chip with a matched series pair suitable for flip chip mounting or wire bonding. For millimetre wave operating frequencies for broad and narrow band use, the diode has applications in LMDS, point-to-point, automotive collision awareness, VSAT, PRD, millimetre wave frequency conversion products and industrial sensors.
Alpha Industries
Tel: 001 781 935 5150
Enquiry No 506

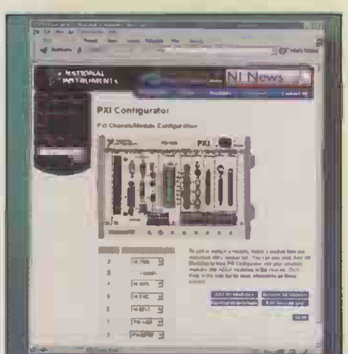
Please quote *Electronics World* when seeking further information

Z80-compatible 8-bit micro

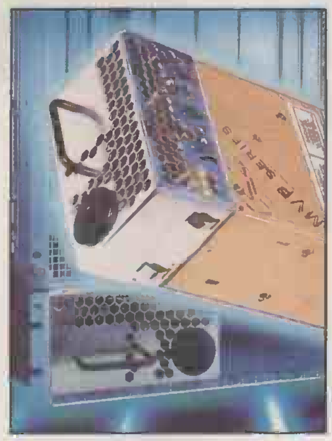
AB Semicon has introduced the AB181E-20 Z80-compatible 8-bit microprocessor using a one cycle architecture. Applications include copiers, digital cameras, digital signal processing, mobile phones, robot controllers and network connected devices such as printers. It incorporates 8080, Z180 and Z80 compatible code sets. In operations where the existing processor is already fully used, such as printer controllers, the AB181E-20 can take over the handling of network protocol stacks and provide the raw data to the printer controller. It will also handle SNMP data to and from the printer controller or NPMP information, without taking up any of the main printer controller's time.
AB Semicon
Tel: 01444 870408
Enquiry No 510

PXI configurator

National Instruments has announced an on-line utility for selecting chassis and modules needed to build a PXI/CompactPCI measurement and automation system. System developers can use the configurator to print a graphical image of a system, export an image for use in other documents, or generate an on-line order for the system. System developers are presented with the option of building a new system or retrieving



and viewing an existing system. It steps users through questions, including chassis type, processor speed, memory, controller type, software, power supply, power cord type and modules, then graphically places them in a chassis. The part numbers representing the PXI system configuration are automatically added to a shopping cart. Every PXI system configuration is assigned an ID so users can retrieve their configurations later without starting over. System developers can also use the ID to create a new configuration from a saved one by modifying the chosen modules.
National Instruments
Tel: 01635 523545
Enquiry No 511



Rack-mounting custom PSU

The PSE2 from Astec is a customisable rack-mounting PSU supporting redundant n+1 configurations with in-service swap-out. Applications include central-exchange telecoms racks, mobile-phone base stations, internet switches and servers, and industrial control systems. Based on the modular design of the firm's MVP series, the PSE has one to ten outputs factory preset to any value between ±2 and ±60V, and can deliver up to 600W. A single-wire current share facility on all outputs above 10A lets n+1 redundant systems be constructed. A built-in safety interlock system ensures that AC voltage cannot be applied until the module is engaged in the system backplane, so any unit can be isolated for field replacement.
Astec
Tel: 01384 842211
Enquiry No 512

16-bit micro with on-chip flash

Hitachi has available a 16-bit microcontroller with third-generation on-chip flash memory. The

H8S/2328F provides almost ten Dhrystone Mips at 25MHz and 3V with a power consumption of 50mA at that clock speed. It contains 256kbyte of flash memory and is made using a 0.35μm process.
Hitachi
Tel: 01628 585163
Enquiry No 513

Development kit

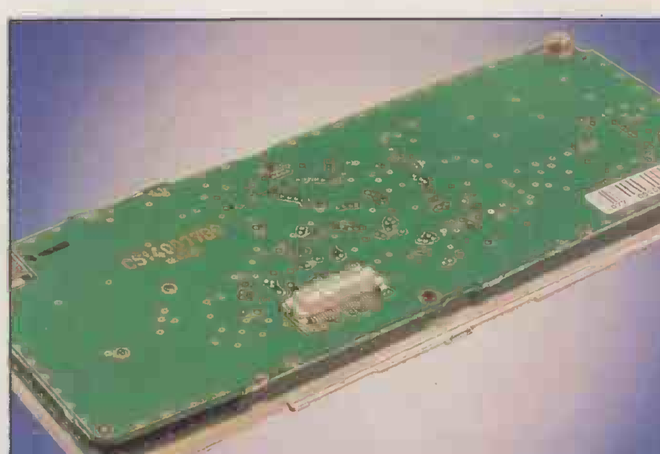
A Bluetooth development kit is available from Ericsson. Designed with Symbionics, the kit consists of a toolbox of equipment to let engineers integrate Bluetooth technology into information appliances. There are two development boards each with a functional Bluetooth radio for testing application software or demonstrating prototypes. Documentation ranges from a novice's introduction to wireless connectivity, to application examples and a description of the Bluetooth packet structure, link types and the protocols used. Also available are two Bluetooth components – a precertified module and a radio transceiver for embedded applications.
Ericsson Microelectronics
Tel: +46 8 757 4700
Enquiry No 509

Limiting amp

The SY88913 1.25Gbit/s limiting post amplifier has been added to Micrel's three-chip fibre-optic set. To work with the firm's SY88902 laser-diode driver and SY88905 laser-diode controller, it is the sister device of the SY88903 limiting post amplifier. The chip set is for use with Gigabit Ethernet, 532 and 1062MHz fibre channel, and 622Mbit/s Sonet. The post amplifier has chatter-free loss-of-signal (LOS) generation and a PECL LOS output. The SY88903 has an open collector TTL LOS output.
Micrel Semiconductor
Tel: 01635 524455
Enquiry No 514

GSM modem

Telital Automotive has released a GSM modem for embedded applications. Available from REP Design, the GM360 transceiver has a serial interface, enabling control via standard AT commands and GSM-specific commands. The serial connection is via a Molex 50-pin connector on one face, letting the unit be mounted directly on a system motherboard, with no need for a cable. The serial connector operates at TTL voltage levels. Weighing 85g, it measures 192 (L) by 51.4 (W) by 13.6mm (D), and has an integral slot for the mini-SIM and SMB connector for an external antenna connection. It is suitable for industrial, railway and automotive applications. Communication rate is 300 to 9600bit/s, supporting data, fax, SMS and voice. Pins on the serial connector are allocated for external microphone and 150Ω earphone connection for voice use.
REP Design
Tel: 01462 670770
Enquiry No 515



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ISDN adapter IC

Rohm has launched an ARM7 IC containing the microprocessor and peripheral functions needed for an ISDN terminal adapter. The BU6611KS processor doesn't need an external CPU, so it can replace the multiple ICs and discrete components normally used to construct ISDN terminal adapters. It is supplied in a

160-pin, SQFP with a PCB footprint of 31.2 by 31.2mm, and is based on an embedded, 24MHz 32-bit Risc.
Rohm Electronics
Tel: 01908 282666
Enquiry No 516

Micro debug modules

The latest addition to the Lauterbach BDM background debug mode embedded systems tool kit is a set of intelligent modules to speed the development and debugging of applications across various processors. Available from Noral Micrologics, the Power-Debug modules let users of Lauterbach' s Trace32 BDM debugger download

pure code to the target via an Ethernet network at speeds up to 600kbyte/s. Each module has a built-in 32-bit Risc controller operating at up to 40Mips.
Noral Micrologies
Tel: 01254 295800
Enquiry No 517

CPLD platform

DWA Technology is offering a CPLD development and prototyping platform to design, simulate, debug and prototype a logic design. Targeted at Altera's Flex 10K CPLDs, the Digilab 10K10 designed by El Camino is for logic designs up to 10 000 gates. It has a built in Byteblaster download cable for device configuration and is supported by general purpose switches, LEDs and seven segment displays. The board contains a socket for an optional EPC2 configuration EEPROM to allow for nonvolatile designs. The platform supports the free Altera baseline version of Max+Plus II.
DWA Technology
Tel: 01234 241818
Enquiry No 518

CAN micro development

Toshiba has expanded its family of microcontroller development platforms with a starter kit that provides programming, testing and implementation of CAN-based embedded systems. For automotive and other applications using the CAN bus, the Topas 900 kit incorporates the hardware and software developers need to evaluate and prototype applications based on Toshiba's TMP95PS54F 16-bit CAN microcontroller. The kit combines two TMP95PS54F evaluation and programming boards, (CAN-I and CAN-II) for CAN communication verification, C compiler, TMPro debugger and the IAR tool chain. A software library and CD-ROM

containing supporting documentation are included. Both CAN boards are equipped with the TMP95PS54F CAN MCU.
Toshiba Electronics
Tel: 00 49 211 52960
Enquiry No 520

Flash for digital machines

Ambar Components has added 32 and 48Mbyte densities to Its Silicon Storage Technology family of memory cards. Using ATA controller technology and flash memory design, the CompactFlash cards have a sustained write performance up to 1.4Mbyte/s. Applications include digital audio players and digital video recorders. Features include dual port SRAM buffer, direct memory access and flash file system.
Ambar Components
Tel: 01296 397396
Enquiry No 519

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Digital multimeter

The DL97 handheld digital multimeter from Kenwood has an accuracy of typically 0.05 per cent and a count rate of 4000 to 40 000. The measurement ranges include AC and DC voltage and current, resistance, diode and continuity, capacitance, frequency, temperature, square-wave frequency generation, and timer mode. It has a dual read-out backlit display with bar graph, enabling simultaneous read-out of combination measurements. Mathematical calculations are standard, as is peak data hold for voltage and current. It comes in an ABS case with rubber safety holster. Features include auto switch off, auto and manual measurement, maximum and minimum memory with recording time display, and a current input connection alarm. It comes with probes and there is an optional RS232 interface.
Kenwood Electronics
Tel: 01923 655291
Enquiry No 521

RF push-pull MOSFETs

Semelab has announced RF push-pull transistors in surface-mount eight-pin ceramic packages. The 28 and 12V parts are available in a choice of 10 or 5W at 1GHz. They are threshold-voltage matched to provide easier biasing and reduce distortion. The package has straight unformed leads, which are soldered to tracks on the top of the PCB in the usual way, while the body of the device occupies a hole cut in the PCB.
Semelab
Enquiry No 524

Shielded connectors

Thomas & Betts has announced Triad 01 shielded connectors with a maximum power rating of 3A and 360° shielding. They are for medical, instrumentation and other applications where airwave pollution can cause failure. The range includes crimp contact connectors, PCB-mounted connectors, retractile connectors and gender maters. They are available in three, four, five, seven or eight pole versions and in D-sub, PCB-mounted or standard style. They can be attached to any style of cable and use twisted contacts attached to the cable by a manual retraction clip with four points of retraction. The panel transfer connector permits female-to-female or male-to-male connection through a gender changer. Custom cable assemblies can be supplied that are shielded through a copper cable, moulded into the circular connector, providing a seamless seal to guard against noise pollution, magnetic fields and radio waves. They are rated to IP65 and DIN40050 for liquid and dust ingress.
Thomas & Betts
Tel: 00 32 2 359 8300
Enquiry No 523

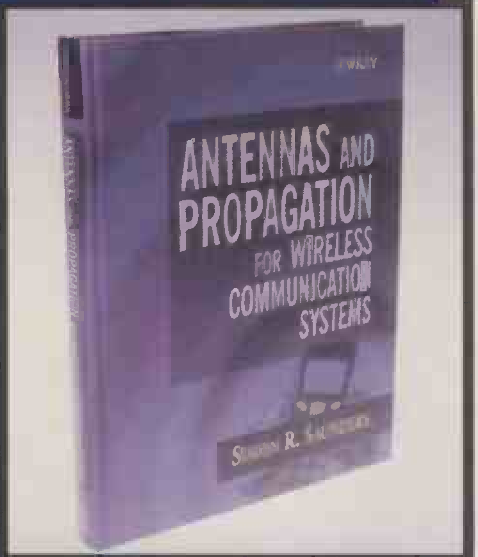
Coax needs no solder

Siemens Electromechanical Componets has introduced an IDC coaxial connector for transmitting high-frequency signals in telecoms and datacoms systems. The three-piece unit can be connected using one tool. There is no need for soldering or the removal of an intermediate cable layer, making it suitable for on site and field operations. Contact resistance is more than 0.5MΩ, even after mechanical and climatic stressing. Screening effectiveness is more than100dB. It is available in various runner types and packaging, including designs for cables with a solid inner conductor.
Siemens EC
Tel: 01344 396000
Enquiry No 522

ARM7TDMI platform

LSI Logic has announced its AMCU silicon development platform, an integrated microcontroller for developing embedded ARM7TDMI Asic designs. It has peripherals, the ARM7TDMI core, 4kbyte of on-chip memory and power-saving modes. The chip can be used as a prototyping vehicle for system-level hardware and software development, and as a reusable modular microcontroller design database, which can be customised and used as part of an Asic design. The platform was designed using the firm's Coreware Asic design methodology and is verifiable, testable and portable. It comes in a 176-ball MBGA. The peripherals that surround the core include Uarts, timer-counters and external memory interface. There is external interrupt capability on all 32 GPIO pins. Debug access on the development board is via a serial port and diagnostic software routines.
LSI Logic
Tel: 01344 413209
Enquiry No 525

BOOK TO BUY



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32W Class A from a 12V rail

Richard Burfoot's 32 watt Class-A push-pull FET power amplifier is inductance loaded, achieving a power efficiency of around 45%.It is fully symmetrical and delivers full power when running from a car battery and a small NiCd pack for biasing.

Class-A amplifiers are generally regarded as inefficient but what, precisely, does this mean? One definition of efficiency is the ratio of maximum output power to input power, expressed as a percentage. It can range widely from, say, 2% for a loudspeaker to close to 100% for a heating element.

Efficiency can usually be determined with good accuracy, but inefficiency is a subjective matter. Take for instance the ubiquitous *RC* coupled Class-A common-emitter amplifier of **Fig. 1a**). What would its efficiency be; 25%? 12.5%?

In fact it cannot better 8.33%, and that is assuming idealised components. This amplifier is very inefficient yet it is probably the most widely used amplifier in electronics.

Another measure of efficiency is voltage efficiency. This is the ratio of maximum peak-peak output voltage – the output compliance – to supply voltage. Again this is expressed as a percentage. Voltage efficiency for the common-emitter amplifier is 67%. This is a lot better than its power efficiency but, as I will show, there is still room for considerable improvement.

A simple modification of Fig. 1a) is to replace the resistor with an inductor, as in Fig. **1b**). What then is the power efficiency of this amplifier, 12.5%?, 25%? No, it is in fact 50%. And it isn't only power efficiency that improves. Voltage efficiency also goes up, from 67% to a full 200% – not quite as spectacular but still a very considerable improvement and even more surprising perhaps.

There is, of course, nothing new about inductive loading. It is very common in RF design, but it is rarely used for audio.

Inductance loading in the input stage

There is a price to pay to this performance increase. In **Fig. 2**, Tr_2 and Tr_6 form part of the input differential amplifier, biased of course for Class-A operation. Conventionally, these transistors would be loaded by a pair of resistors at a cost of 5 pence or so, but in this case they are loaded by L_1, which retails at £10.55.

As specified, L_1 is actually a transformer, which accounts, to some extent for its high cost. A purpose-designed – or even off-the-shelf – inductor should cost less, but even so it would take a very long time indeed for any efficiency improvement to offset this order of cost difference.

It would seem then that inductance loading for the input stage cannot be justified on efficiency grounds alone. In truth, this part of the design started out as just as a bit of fun. Nonetheless I retained it because it confers yet another desirable attribute, and that is an inherently low output offset voltage.

The weak point of differential amplifiers – output offset voltage – is a secondary effect of unbalanced collector current. It appears as a difference in the voltage across the two resistors. Replacing these resistors with a low resistance inductor removes the voltage, and with it the difference and its troublesome drift. It does so without the need for additional control.

As a bonus, inductance loading permits considerable freedom in setting and controlling the differential-amplifier output voltage. In this case R_6 and R_8, in conjunction with current source $Tr_{3\text{-}4}$, set the output voltage, which in turn is stabilised by $Tr_{8\text{-}9}$ via Tr_5.

Emitter followers Tr_1 and Tr_7 then drive the output pair $Tr_{10\text{-}11}$. Note that, because the differential amplifier is a push-pull stage, voltage efficiency is further doubled to a theoretical 400%. The actual figure is academic but it does provide the potential for Tr_2 and Tr_6 collectors to swing far below the –12V rail. This in itself is not a requirement of the design but it does mean that the stage operates well inside its linear range.

Inductance loaded output stage

Class-A output stages having efficiencies approaching the theoretical maximum of 50% have been published before, but many only achieve 25% with some as low as 8%. Such poorly specified amplifiers doubtless have their uses but their high heat flux must compromise their application in the domestic situation.

The output stage of Fig. 2 provides its power with an efficiency of about 45%; idealised components would provide for 50%. It does this with a voltage efficiency of some 377% and this in turn has implications for the choice of power supply, as discussed later.

Drive voltage from followers Tr_1 and Tr_7 is set such that Tr_{10} and Tr_{11} draw 3A each, giving a quiescent current of 6A. These currents are supplied via L_2. Ignoring for now the small voltage drop across its windings, all three terminals of L_2, along with the loudspeaker terminals, are at 0V, i.e. at the positive level of battery B_2.

How it works

Suppose that a signal drives Tr_{10}'s gate positive. Its drain current will increase above 3A but this additional current cannot be sourced from L_2 because it is an inductor and will not permit such a change at the frequencies of interest. Instead the additional current must come from the loudspeaker, the resulting voltage drop then pulling Tr_{10}'s drain towards –12V.

Simultaneously Tr_{11}'s gate is driven negatively. Its drain current will fall below 3A, but again L_2 will not permit any change to the 3A quiescent current. Instead this current is diverted into the loudspeaker, the resulting voltage generated now pulling Tr_{11}'s drain up towards +12V.

At the limit, Tr_{10} will saturate at 6A drain current, with 3A coming from L_2 and 3A from the loudspeaker. Transistor Tr_{11} will cut-off with the 3A from its side of L_2 going to the loudspeaker. At this point, one side of the loudspeaker will be at –12V with the other at +12V for a total peak voltage of 24V. Note that at no point does signal current flow in L_2, which continues to supply its quiescent 6A as should be expected.

On the following half cycle, the general situation reverses, for a peak loudspeaker voltage of –24V, giving a peak-to-peak output of 48V. From a 12V supply this represents a voltage efficiency of 400%, though as stated earlier only about 377% is achievable in practice. The general performance can now be checked.

Input power, P_{in}	12V×6A=	72W
Output power, P_{out}	$V_{pp}{}^2/8R_L$=$(48V)^2/64$=	36W

Power efficiency is $P_{out}/P_{in}\times100\%$, i.e. 50%. Again these are idealised figures. In practice, achievable power is some 32W for an efficiency of 45%.

Inductors and other considerations

At the lower –3dB point, of, say, 40Hz, L_2 needs to have a reactance equal to the load, LS. So,

$$L_2 = 2\pi \times \frac{40}{8\Omega} = 32\text{mH}$$

and of course the winding resistance should be as low as possible.

Knowing the value of the inductor is of limited help though because such components are neither readily available nor easily made. Instead I used a spare 15V-0-15 V 50VA mains transformer secondary, leaving the primary unconnected. This gives a surprisingly good result and rewinding the transformer as a choke – i.e. no mains primary – with a heavier gauge wire did not appear to improve the performance. Nevertheless this would be the first item to optimise given the necessary resources.

Prospective builders should be advised that a mains transformer, used in this way, will develop very high voltages at

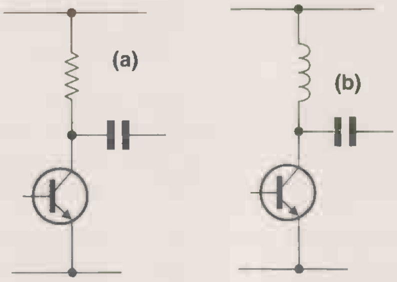

Fig. 1. Fitting a Class-A amplifier with an inductive load instead of a resistance gives a dramatic increase in power efficiency.

the unconnected primary. This in itself is hazardous and might also result in failure of the winding insulation, though I have not experienced problems with any of the transformers that I have used.

Similarly with L_1, because of the difficulties in specifying procuring and testing inductors, I chose it from Farnell's audio transformer selection, part number 149-840.

A feedback loop is used to establish the 6A bias current. The small voltage drop across L_2 referred to above is proportional to the bias current. This voltage is picked off by R_{18-19} and is compared, by differential amplifier Tr_{8-9}, with a reference voltage set by R_{13-14}.

Output from Tr_9 passes to Tr_5, as previously indicated, to control the input differential amplifier and hence the FET drive, completing the loop. Overall feedback, taken from Tr_{10}, is applied to the input differential amplifier via R_{10} and R_{12}.

Input bootstrapping is not used because I did not want the complication of the switch-on delay relay needed to mask the thump. Instead, I prefer a low but acceptable input impedance of about 15kΩ at a sensitivity of 750mV.

Output protection should not be necessary because the output topology featured poses no threat to the loudspeakers.

Powering the design

During initial experiments, a 9V NiCd battery supplied the differential amplifier bias, now shown as B_1. I used a battery because I wanted to be quite sure that hum was not coming in from the front end.

Still dissatisfied with the hum level, and having spent more than enough time and effort on the AC mains power supply, I also replaced this supply with a battery. Shown as B_2, this was an old lead-acid car battery.

The original aim was simply to allow the basic design to move forward but I became seduced by the total absence of all supply noise. There now seems to be no good reason to go back to AC mains power. The prospect of introducing hum along with transformer buzz and even more cooling problems is not at all attractive.

The original car batteries have since been replaced by 75Ah lead-acid Leisure batteries as used in caravans. There's one per channel. These are more tolerant of the occasional deep-discharge and have the capacity to provide for some 5 to 6 hours use between charges.

In its turn, battery power has implications for packaging. Free from a central power source, the attractions of organising the amplifiers as monoblocs arises. This permits mounting each amplifier to the back of its loudspeaker. This eases the cooling problem since the heat sinks can be mounted vertically in free cooling air.

It also eliminates the speaker cables and because the only electrical correction is 0V, cross-talk is reduced to zero.

Living with the amplifier

This amplifier has been designed entirely for my own domestic use. Decisions such as the choice of battery power and monobloc construction undoubtedly make it commercially unattractive, but they also undoubtedly make it sound better, which is what I consider important.

The prototype has been in use for two years now. It is uncannily quiet and smooth playing a Mahler Adagio – as I write this. But it can also deliver the power and dynamics required for thumping Afro-Celt.

Distortion when it comes is a bit of a surprise. I had expected the soft clipping and gentle low frequency roll-off associated with valve amplifiers. After all they both feature large output inductors, and MOSFETs are reputed to sound like valves.

Certainly the clipping is soft from the mid-bass upwards, but it does not roll-off as expected. If a test signal of, say 400Hz, driving the amplifier close to clipping, is slowly reduced in frequency, full output is maintained until a point is reached, at around 80Hz, when L_2 can no longer supply the required current for the required time. The current first runs out at the end of each half cycle, the voltage falling abruptly to 0V and looking a bit like crossover distortion. This distortion increases as the frequency is reduced unless the drive is also reduced.

I describe this behaviour in detail not because it is a problem, but because it is unusual. In practice, none of my CD collection will excite it before normal clipping intrudes. Besides, using the amplifier in this way misses the point.

The inconvenience of battery power may be a cause for concern but it really should not be. The results are well worth the little effort required to connect and switch on the chargers. The batteries can even be left charging while the amplifier is in use. I find that the inevitable bit of hum is acceptable when you know you can stop it at will.

Construction and performance

This is not a constructional article, but prospective builders may like to know that components shown within the dotted lines can be assembled on a piece of strip-board measuring 97 by 67mm.

The wiring shown in bold should be generously rated and kept short. My prototype heat sinks were recovered from a redundant power supply. I guess they would be rated at about 0.6°C/W to 0.7°C/W per channel.

Quiescent current should be set to 6A by select-on-test adjustment of R_{13}, the actual value depending greatly on the inductor chosen for L_2. Selection of R_{13} should start with low values. A couple of limiting resistors of, say, 8Ω by 25W, in the drains of Tr_{10-11} would be a sensible precaution when making initial adjustments.

Specific distortion figures are not provided because I do not have the equipment necessary to obtain them. However, the amplifier auditions well and is not at all difficult or expensive to build. Why not try it?

Finally I must credit Mullard Ltd, which published details around the early 1960s of a 5W single-ended inductance loaded amplifier using an *AD149* p-n-p germanium transistor. I have tried for some time to replace my lost copy but without success. This was a truly delightful little gem and was the inspiration for the amplifier presented here. ■

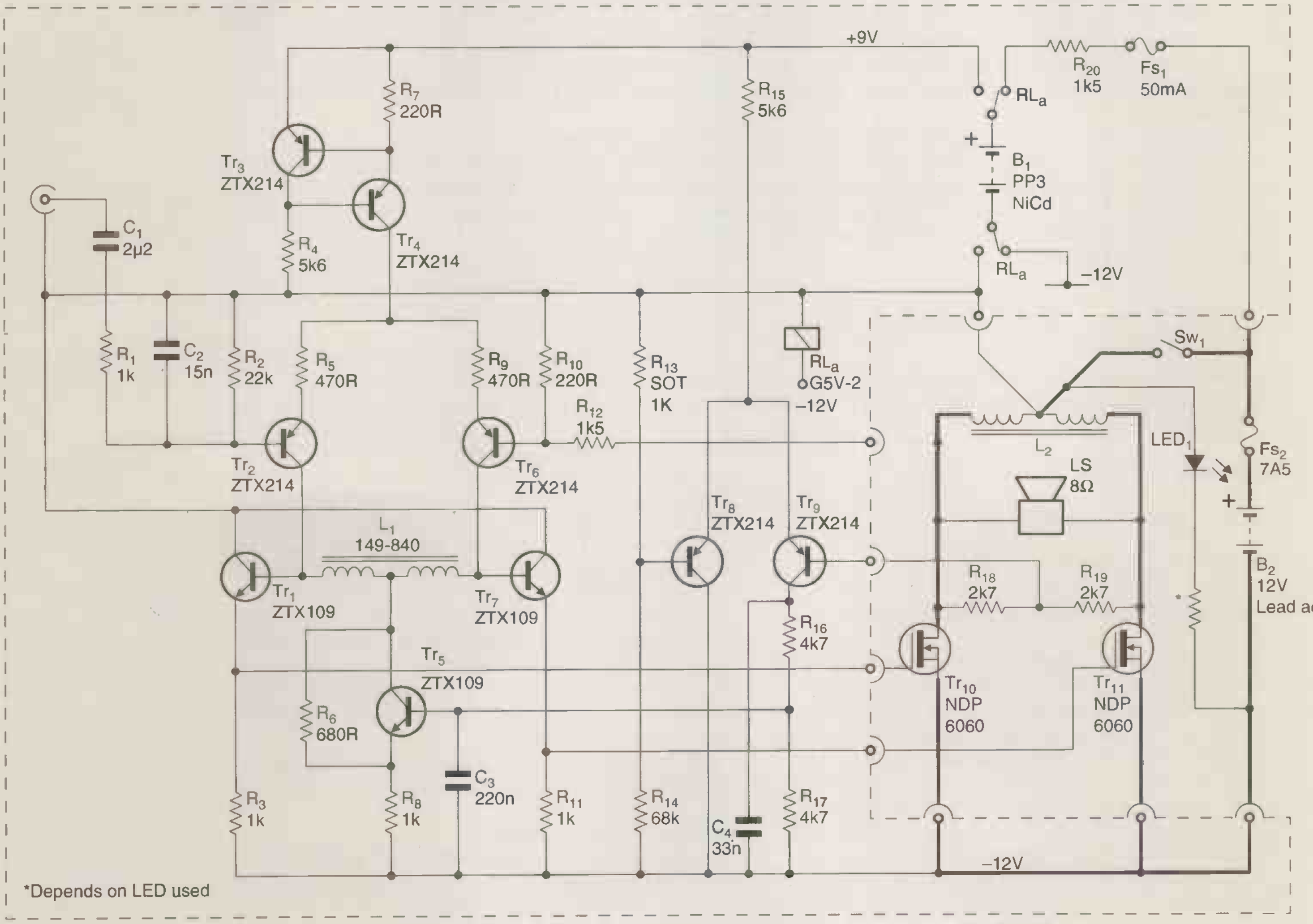

Fig. 2. Complete Class-A power amplifier with inductive loading. Both inductors were transformers since they are much easier to obtain than large inductors. If you use transformers, make sure you insulate the unused mains side of L_2. It can give a nasty shock. Turning S_1 on activates the relay, switching the NiCd battery from 'charge' to 'on'

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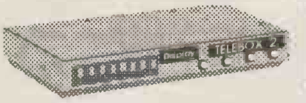

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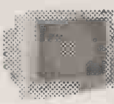

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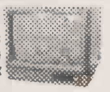

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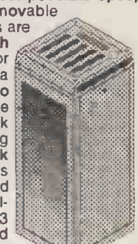

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The dirt on switching

Switches can do some really odd things. Most engineers learn this soon after connecting a switch or relay to a digital system. Switches don't make and break cleanly on the time scales of digital systems. Instead, a typical switch makes multiple transitions during the tens of milliseconds required to open or close it. Commonly called switch bounce, this behaviour is an inescapable fact of life, as John Wettroth explains.

After connecting a standard switch to a digital counting circuit, you can observe several counts on opening and several counts on closing, **Figs 1-2**. This erratic action can wreak havoc on data, because the exact number of counts does not necessarily repeat in the long term.

Switch bounce is not consistent from unit to unit, lot to lot, or even over the life of an individual switch. Membrane switches and some other types appear not to bounce when new, but all mechanical switches bounce sometimes. Nothing can ensure that another switch of the same type will act the same way, or that a particular switch will remain bounce-free as it ages.

In addition to bounce, switches and digital systems have other annoying habits. Strange things happen, for example, when you run switch wiring in a noisy industrial environment. An open switch has high impedance by definition, so interfering signals have an easy load to work against. Any noise impulse that is capacitively or inductively coupled to the switch wiring can cause phantom switch closures.

Imagine a PLC, i.e. a programmable-logic controller, switching a motor through a hefty relay. A limit switch placed near the motor provides position feedback to a digital input on the PLC.

When the PLC tells the motor to start, a surge of current flowing to the relay and motor can couple to other conductors in the long wiring runs, causing ground bounce or a capacitively coupled spike in the digital input. If not properly designed, the PLC may interpret this spike as a premature switch closure and shut down the operation.

Similar things can happen when the PLC turns the load off, due to the effect of wiring capacitance, wiring inductance, and the inductive kick of the relay and motor. If the PLC and its digital inputs are not properly designed these spikes and transients can cause erroneous readings on the digital inputs.

The digital and analogue inputs on equipment used in the home, office, and industry are subject to the effects of overvoltage, voltage transients and ESD strikes. Improper wiring, miscellaneous fault conditions and power-supply sequencing – in which one box with power off is connected to another with power on, even temporarily – cause overvoltage.

Voltage transients are often associated with capacitively or inductively coupled spikes, as discussed above. ESD can strike a connector, an operator console, or a terminal strip during installation. Any of these transients can cause destruction if the system latches up. If not destructive, they can cause CPU resets, watchdog overflows, and other erratic operation.

System designers should be aware of these problems and the methods used to combat them. One solution for such interface problems is a new series of ICs. Available in low-cost, easy-to-use configurations, these devices offer foolproof, software-free debouncing along with protection against overvoltage and ESD.

This article highlights the application of IC switch debouncers while describing the classic methods for thwarting overvoltage, voltage/current spikes, switch bounce, and ESD.

Switch bounce

If asked, most engineers would say that switches are debounced in software, and that debouncing is no problem. Both assumptions are true if you pay proper attention to the details. Software debouncing takes care of the bounce, but does not address the problems of overvoltage, ESD, or other transients.

Debouncing with resistors and capacitors is also possible. In general, you need a pull-up resistor, a resistor and capacitor in series, a resistor to the input of a Schmitt-trigger buffer, and often a diode to ensure that the capacitor charge doesn't force lots of current through the buffer's input-protection network during power-down.

The resulting parts count can be unwieldy for multiple-input systems, **Fig. 3**, so this approach will not be covered in any detail.

Debouncing via software

Debouncing via software is the primary method in use today. A good debouncing routine is actually real-time software that acts like a simple low-pass digital filter. Non-switch digital inputs are often routed through debounce filters as well. That technique can eliminate short transients at the input by ensuring a stable state before reporting the input open or closed.

The pseudo code in **List 1** illustrates a software-debounce routine for one input. It accommodates multiple inputs if you generalise the routine and use pointer-based variables, etc. Though a mediocre approach at best, this type of routine is often used in spite of the problems and flaws outlined below.

The routine debounces switch closures, but it will accept 'open' as a legitimate state even when the switch is bouncing. Though unintentional, this asymmetrical operation might be acceptable in keypads and other systems that take action on closures but not on opens. For general-purpose inputs, you should debounce both edges.

Another shortcoming is that this routine assumes the switch is open if not closed, thereby ignoring a third state in which the switch is unstable – i.e. still bouncing. A better routine should therefore report the last non-bouncing state until the switch reaches a new debounced state. This action can also cause problems, however. In such cases, the software should recognise a third state of 'changing.'

Sampling wastes

Many debounce routines sample the input repeatedly, waiting for it to remain in the same state for a prearranged number of samples. If the switch changes state during that interval, the routine tests the new state for stability in the same way. This action can cause large delays that eat up a lot of CPU time.

As an extreme case, a programmable logic controller with high frequency applied to one of its general-purpose input ports – whether inadvertent, on purpose, or due to failure – would completely hang the processor. A watchdog timer might bring back the processor, but the problem would recur indefinitely; not a robust design.

Further, you need a lot of memory and code to debounce a large industrial system with lots of inputs, such as a PLC or general-purpose input board.

Each input requires a closed counter, an open counter, and two bits to define its state.

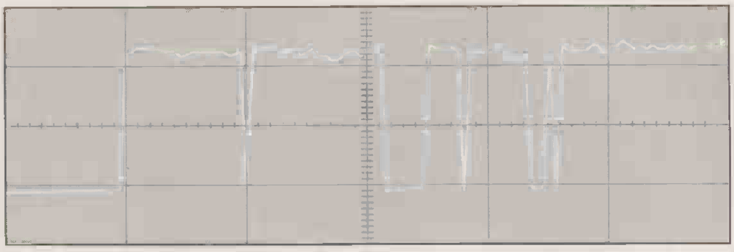


Fig. 1. This rising-edge switch bounce for a small push-button switch shows an approximate 5ms bounce interval that includes ten transitions. Like a bouncing ball, the switch-action frequency increases towards the right.

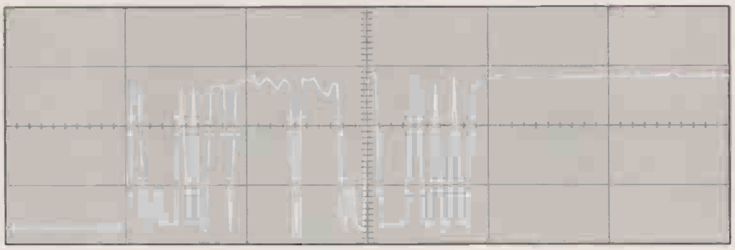


Fig. 2. Another rising-edge switch bounce for a 5A contact relay shows an approximate 5.5ms bounce interval that includes 20 full-amplitude transitions and a few smaller ones.

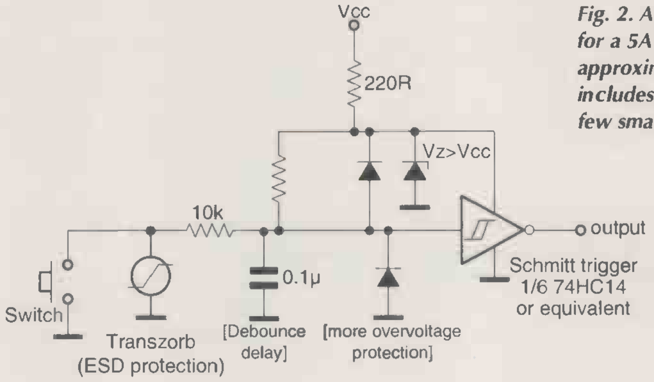


Fig. 3. Discrete components can provide debouncing with protection against ESD and overvoltage, but a properly designed discrete interface that accounts for all likely faults is unwieldy for more than a few inputs.

Transient and ESD suppression

The standard prevention for ESD is a transient suppressor or MOV device at each external input.

Quad and octal Tranzorbs, for example, are straightforward and relatively inexpensive devices that can reduce clutter and real estate requirements, but care must be taken to avoid cross coupling of fault currents. This approach is common in industrial and automotive systems, where engineers understand the peril of omitting such protection.

A good practice is to connect a 220Ω resistor in series with the V_{CC} line for port input devices. A common CMOS input device like the octal 74HC244 or 74HC573, for example, draws very little current. Should it latch up, the 220Ω resistor limits the current and power dissipation to a safe level.

Power cycling may still be necessary, though. In general, you should not connect the port pins of a microcontroller to the external inputs directly. Latchup is a problem, but radiated EMI is likely to be even worse.

Because a part cannot latch up unless sufficient current is injected into one of its pins, some designers believe that resistors in series with CMOS digital inputs prevent these problems. Indeed, the threshold for SCR latchup in modern CMOS ICs can exceed 50mA.

This high current threshold, covered in the next section, actually protects against overvoltage to some extent, but is not necessarily effective for ESD. A 15kV ESD strike can force significant currents through parasitic paths and around resistors, and it can force a large current even through 100kΩ.

Overvoltage protection

Overvoltage protection enables a system to withstand continuous and longer-term-transient inputs that extend beyond the rails.

As an example, an IC with no V_{cc} applied has 24V from an external source applied to the inputs. Such applied voltage often 'backdrives' the protection networks, forcing voltage onto the power rail inside a system.

One effective countermeasure is a

List 1. Action sequence for debouncing using software.

Action	Comments
1. Input timer: expired?	A timer bit is polled in the main routine
2. Return if no timer	Go do something more useful
3. Get input bit	The "bouncy" input
4. Count++ if high; clear else	Increment a counter if input is high
5. If count >4 state=1, else 0	Check counter and clamp it at 4
6. Return input state	State is debounced

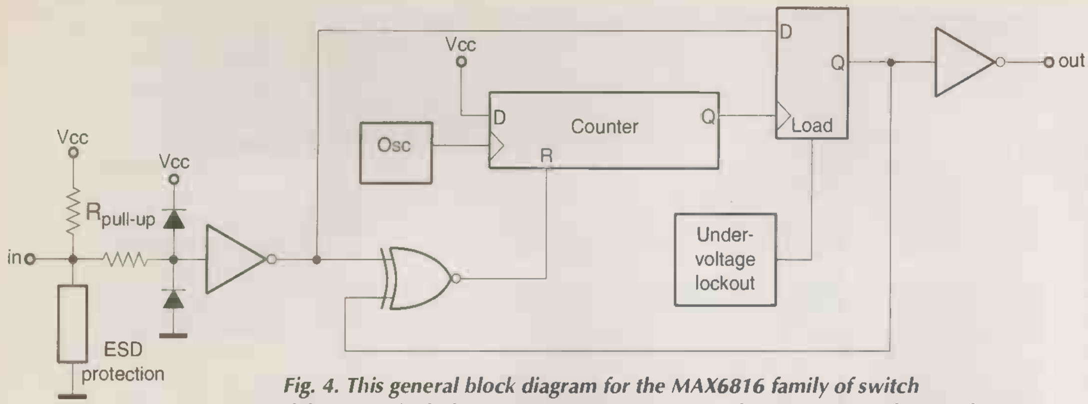


Fig. 4. This general block diagram for the MAX6816 family of switch debouncers includes an input structure protected against ESD and overvoltage, followed by a digital filter that debounces the input and applies undervoltage lockout.

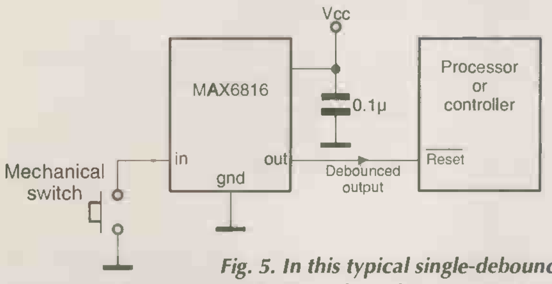


Fig. 5. In this typical single-debouncer application, the only components are a small bypass capacitor and the 4-pin SOT-23 package.

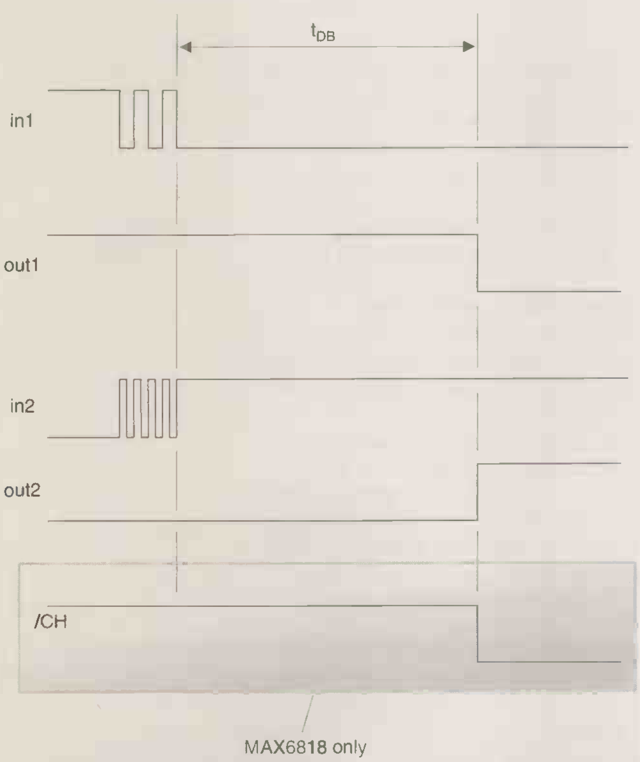


Fig. 6. Timing diagram for the MAX6816 switch-debouncer family shows that the outputs change state about 40ms after the inputs become stable. An additional MAX6818 output that indicates a change of state for any of the inputs. The $\overline{CH}$ output reduces polling overhead, especially in multiple-input systems.

resistor in series with the input, acting against protection diodes tied to the rails. A zener diode across the V_{cc} rails of the input port will also help.

To ensure that the protection circuits won't fail under worst-case conditions, you should calculate the maximum power dissipation of this zener and the series input resistors.

Integral debouncing ESD protection

Several years ago, Maxim engineers saw the need for a simple interface device capable of debouncing a switch while protecting it against ESD and overvoltage.

Some customers were using the manual-reset input of microcontroller-supervisory ICs like the *MAX811* just to obtain the single-channel debouncer function in a SOT-23 package. Others were using ESD-protected RS232 transceivers as general-purpose digital-input devices.

Customers were attracted to the RS232 ICs because they could handle low-voltage transitions while withstanding high voltage and ESD. Putting these facts together, Maxim produced a line of switch debouncers that incorporated ESD protection and robust input features, **Figs 4-5**.

The *MAX6816* is a single-switch debouncer in a four-pin SOT-23 package, while the *MAX6817* is a dual-switch debouncer in a six-pin SOT-23 package. They provide debounce logic and a digital filter, input overvoltage protection to ±25V, and ESD protection to ±15kV for harsh industrial environments.

Operating on single supply voltages in the range 2.7V to 5.5V, they draw typical supply currents of only 6μA. They also provide undervoltage-lockout circuitry that ensures correct output states on power-up.

Because the proprietary ESD-protection structure at each input includes an overvoltage clamping diode and 63kΩ pull-up resistor, these ICs provide a direct interface to the switch without external components. Their nominal debounce delay of 40ms ±20ms masks the bounce produced by even the ugliest of switches, **Fig. 6**.

An octal debouncer

The *MAX6818* octal-switch debouncer is designed for data-bus interfacing, **Fig. 7**. It monitors eight switches, providing a change-of-state output, $\overline{CH}$, and three-stated data-bus output in addition to the debounce and input-protection features of the single and dual parts. In particular, its $\overline{CH}$ output greatly simplifies the polling and interrupting of microprocessors.

Each time the system reads the data outputs by driving the enable line low, the IC resets high. Output $\overline{CH}$ then goes low when any input changes state. The *MAX6818* is pin-compatible with the 74HC573 and other standard, 20-pin octal logic devices. It handles multiple inputs with ease.

These new switch debouncing ICs solve multiple problems associated with connecting digital systems to noisy, transient-prone, 'bouncing' inputs. They make systems more robust and reliable by simplifying design, reducing CPU time and overhead, and replacing multiple passive components. ■

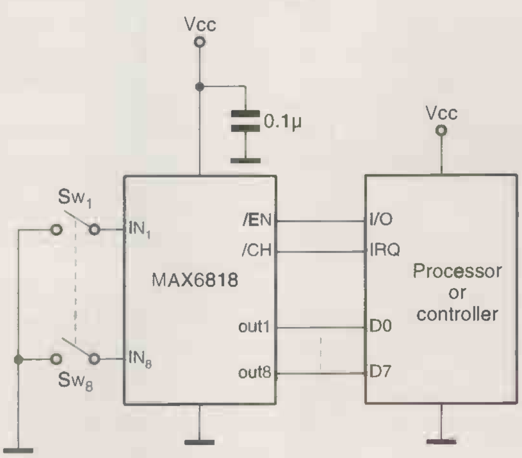


Fig. 7. In a typical application, the MAX6818 data outputs remain three-stated until $\overline{EN}$ is pulled low. The change output, $\overline{CH}$, is reset high following each read, and set low following a change of state at any input. It can either be polled by the system or tied to an interrupt as shown.

Measuring RF power

Joe Carr explains presents a backgrounder to RF power and describes a number of circuits for measuring it.

Radio-frequency power measurements are made for a variety of purposes. In this pair of articles, several different topics will be discussed: the nature of the power being measured, methods of measuring power, error sources in RF power measurement, and typical commercial instruments used for RF power measurements.

The assumption is that the RF power is being measured to determine the output level produced by a radio transmitter, or some associated circuit or device.

What is Power?

Electrical power is defined as energy flow per unit of time. The internationally accepted standard unit of power is the *watt*, abbreviated to W of course, which is defined as an energy flow of one joule per second.

Other electrical units are defined in terms of the watt. One volt for example is one watt per ampere of current flow. The watt is the product of the electrical potential and the current flowing,

$$P = V \times I$$

Other expressions of power include,

$$P = I^2 \times R$$

and,

$$P = \frac{V^2}{R}$$

Where P is power in watts, V is electrical potential in volts, I is current in amperes and R is resistance in ohms.

Decibel notation of power units

It is common practice to express power relationships in terms of decibel notation, which allows gains and losses to be added and subtracted, rather than multiplied and divided, somewhat simplifying the arithmetic.

For relative power levels,

$$dB = 10\log\left[\frac{P1}{P2}\right] \qquad (4)$$

And for absolute power levels 50Ω load,

$$dBm = 10\log\left[\frac{P_w}{0.001}\right] \qquad (5)$$

or,

$$dBm = 10\log P_{mw} \qquad (6)$$

where dBm is power level relative to one-milliwatt in a 50Ω load, $P1$ and $P2$ are two power levels (same units), P_W is power in watts and P_{mW} is power in milliwatts.

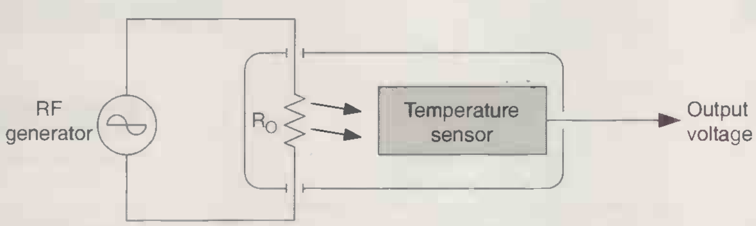


Fig. 1. Simple bolometry/calorimetry approach to measuring RF power.

Table 1. Power relationships between various modulation waveforms. System impedance is 50 Ω. Waveforms are compared with 100W RMS unkeyed CW power.

Waveform Description	PEV	PEP (PEV^2/Z_o)	Heating power
Continuous wave (CW)	70.7	100W	100W
Amplitude modulation (100%)	141.4	400W	150W
Amplitude modulation (75%)	122.3	300W	127W
Single-sideband (one-tone)	70.7	100W	100W
Single-sideband (two-tone)	70.7	100W	50W
Single-sideband (voice)	70.7	100W	(Note 1)
TV black level	70.7	100W	60.1W
Pulse (10% duty cycle)	70.7	100W	10W
Multiple carriers (Note 2)	282.8	1600W	400W

NOTES:
Note 1: Depends on voice modulation characteristics
Note 2: Four 100W RMS CW carriers

Types of RF power measurement

Measuring RF power is essentially the same as measuring low frequency AC power, but certain additional problems present themselves.

For a continuous wave, or CW, signal, the issue is relatively straight forward because the signal is a series of equal amplitude sine waves. For on-off telegraphy, the problem gets somewhat more difficult because the waves are not constant amplitude. The RF power depends on the ratio of on time to off time.

In the case of the sine wave, a peak reading instrument, such as a diode detector, can be calibrated for root-

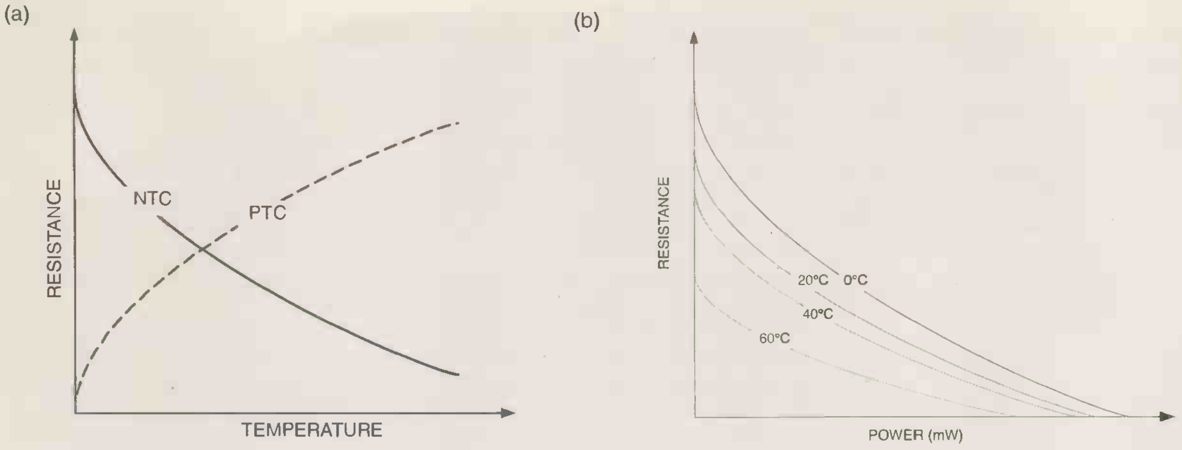
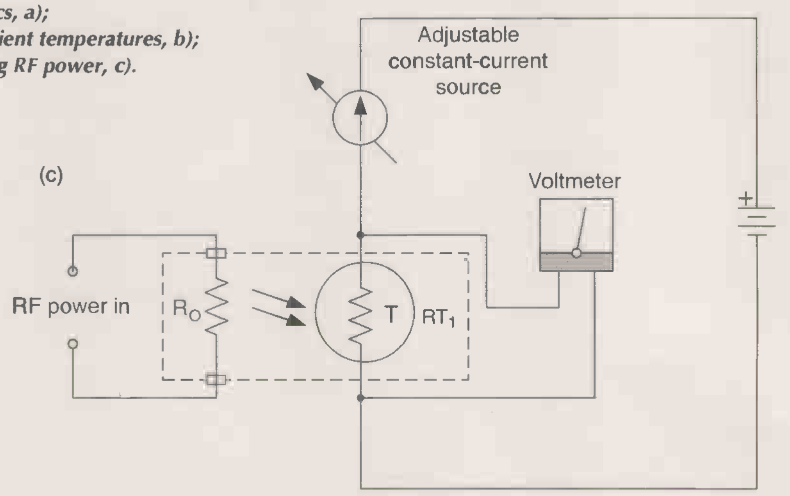


Fig. 2. Thermistor R-versus-T characteristics, a); R-versus-P characteristics at different ambient temperatures, b); and thermistor sensor mount for measuring RF power, c).



mean-square (RMS) power by the simple expedient of dividing the indication by the square root of two, which is 1.414.

If the meter is inherently RMS reading – it is, for example, a thermally based instrument – then the power measurement of the complex waveform is inherently RMS.

Table 1 shows the power relationships for assorted modulation waveforms. The figures are arbitrarily based on the peak envelope voltage (PEV) in each waveform, a 50Ω system impedance, and are compared with 100W RMS unkeyed CW power.

Methods for measuring RF power

RF power meters use a number of different approaches to making the measurement. Some instruments measure the current or the voltage at a resistive load, and depend on the equations I^2R or E^2/R.

Other methods are based on the fact that power dissipated in a resistive load is converted to heat, so the temperature change before and after the RF power is applied can be used as the indicator of RF power. This approach has the advantage of finding a DC equivalent RMS power.

Figure 1 shows the basic scheme. A load resistor, R_o, with a resistance equal to the system impedance, is enclosed in an isolated environment with some sort of temperature sensor.

Theoretically, you could place a dummy load resistor in a workshop room, and then use a glass mercury thermometer and stop watch to measure the rise in temperature and elapsed time to find the power. That's hardly practical though.

The basic idea is to find a sensor, such as a thermistor or thermocouple, that will convert the heat generated in the load resistor to a DC, or low-frequency AC, signal that is easily measured with ordinary electronic instruments.

In the case shown in Fig. 1, the temperature sensor produces a voltage output that is proportional to the applied RF power level.

Thermistor RF power meters

A thermistor is a resistor that changes its electrical resistance with changes in temperature. Although all conductors exhibit some 'thermistor behaviour' actual thermistors are usually made of a metallic oxide compound.

Figure 2a) shows the resistance versus temperature curve for a typical thermistor device. A negative temperature coefficient, or NTC, device will decrease resistance with increases in temperature. A positive temperature coefficient device, or PTC, device is the opposite: resistance rises with increases in temperature.

Bolometers

Figure 2b) shows a family of resistance versus self-heating power curves for a single thermistor operated at different temperatures.

The resistance is not only nonlinear, which makes measurements difficult enough in its own right, but also the shape and placement of the curve varies with temperature. As a result, straight thermistor instruments can be misleading.

Bolometry is a method that takes advantage of this problem to create a more accurate RF power measurement system.

Self-heating power is caused by a DC bias flowing in the thermistor. Figure **2c)** shows how self-heating can be used in bolometry. Thermistor RT_1 is adjusted to a specified self-heating point when no RF power is applied to the dummy load R_O. The resistance of thermistor RT_1 can be read from the voltmeter because the current from the constant-current source remains the same once it is adjusted to a set point.

When RF power is applied to the dummy load, heat radiated from the load causes the resistance of RT_1 to decrease. The bolometer current source is then adjusted to decrease the bias until the resistance rises back to the value it had before power was applied. This point is indicated by returning the meter reading to the same point as before.

The change of bias power required to restore the thermistor to the same resistance is therefore equal to the power dissipated in the dummy load.

Self-balancing bridge instruments

The Wheatstone bridge circuit, **Fig. 3**, is used in a number of instrumentation circuits. In the null condition, when V_O is zero, the ratios of the resistors are equal: $R_1/RT_1=R_2/R_3$.

It is not strictly necessary that $R_1=RT_1=R_2=R_3$, only that $R_1/RT_1=R_2/R_3$. If one of the resistors is a thermistor, then the temperature can be measured by the unbalance of the bridge. Similarly, if the thermistor resistance is such that the equality $R_1/RT_1=R_2/R_3$ is satisfied, then you can infer the resistance of the thermistor by the null condition.

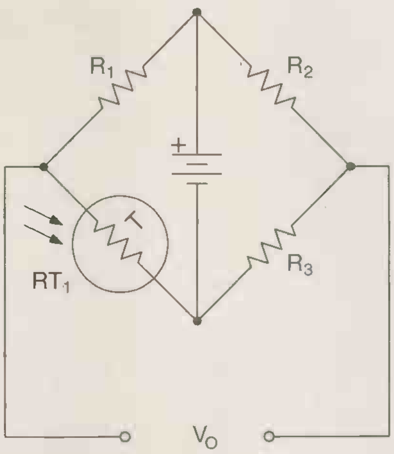


Fig. 3. Wheatstone bridge circuit with a thermistor in one arm.

The self-balancing – also known as autobalancing or autonull – bridge shown in **Fig. 4** uses a Wheatstone bridge thermistor to perform bolometry measurement of RF power. The thermistor mount sensor assembly contains a dummy load and a thermistor, R_T. The null condition is created when $R_1/R_T=R_2/R_3$.

The self-balancing bridge uses a differential amplifier A_1 to perform the balancing. A differential amplifier produces an output voltage proportional to the difference in two input voltages.

When the Wheatstone bridge is in balance, then the output of the differential amplifier is zero. The bias for the Wheatstone bridge, hence the thermistor in the bolometer sensor, is derived from the output of the amplifier.

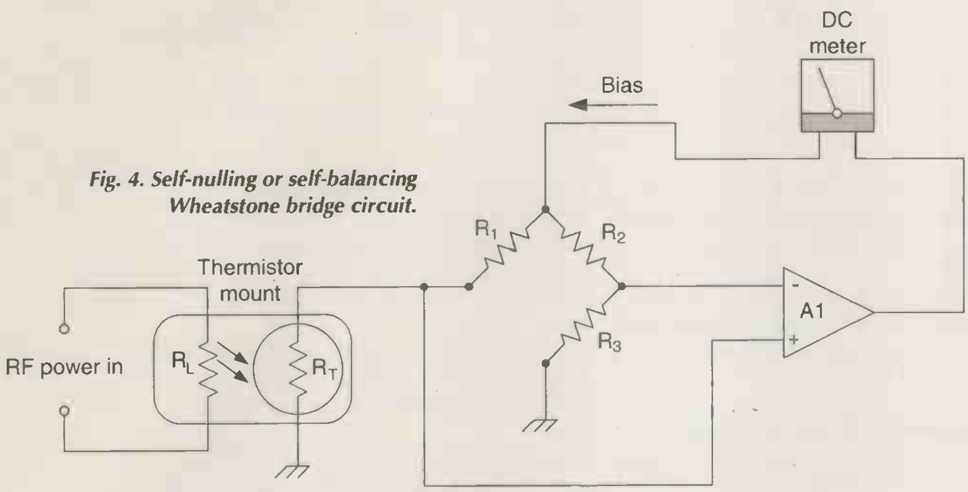


Fig. 4. Self-nulling or self-balancing Wheatstone bridge circuit.

A change in the resistance of the thermistor unbalances the bridge, and this moves the amplifier's differential input voltage away from zero. The amplifier output voltage goes up, thereby changing the bias current in an amount and direction necessary to restore balance. Thus, by reading the bias current, the power level that changed the thermistor resistance can be inferred.

Because the thermistor will have a different characteristic curve at different ambient temperatures, it is necessary to either control the ambient temperature, or correct for it. It is very difficult to control the ambient temperature. Although it is done, it is also not terribly practical in most cases. As a result, it is common to find RF power meters using two thermistors in the measurement process, **Fig. 5**.

One thermistor is mounted in the thermistor sensor mount used to measure RF power, while the other is used to measure the ambient temperature. The readings of the ambient thermistor are used to correct the readings of the sensor thermistor.

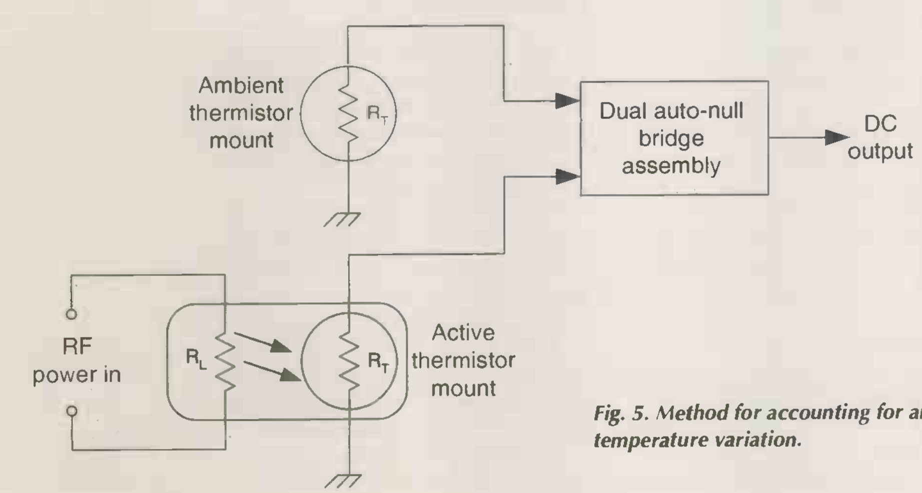


Fig. 5. Method for accounting for ambient temperature variation.

Thermocouple RF power meters

The thermocouple is one of the oldest forms of temperature sensor. When two dissimilar metals are connected together to form a junction, and the junction is heated, then the potential across the free ends, V_T, is proportional to the temperature of the hot junction. This phenomenon is called the Seebeck effect.

A thermocouple RF ammeter is constructed using thermocouples and a small value resistance heating element, **Fig. 6**. The meter will have a small wire resistance element in close proximity to a thermocouple element. The thermocouple is, in turn, connected to a DC meter.

When current flows through the resistance heating element, the potential across the ends of the thermocouple changes proportional to the RMS value of the current. Thus, the RF ammeter measures the RMS value of the RF current.

If the RF ammeter is used to measure the current flowing from an RF source to a resistive load, then the product I^2R indicates the true RF power. These meters can measure RF current up to 50 or 60MHz, depending on the instrument.

Thermocouples and thermistors share the ability to measure true RF power. Although thermocouple RF ammeters have been used since the 1930s, or earlier, the use of thermocouples in higher frequency and microwave power meters started in

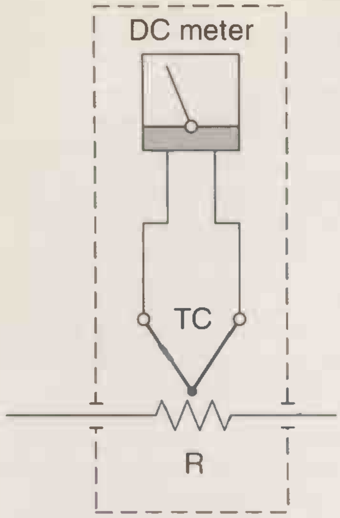


Fig. 6. The thermocouple RF ammeter is useful for measuring frequencies to about 60MHz.

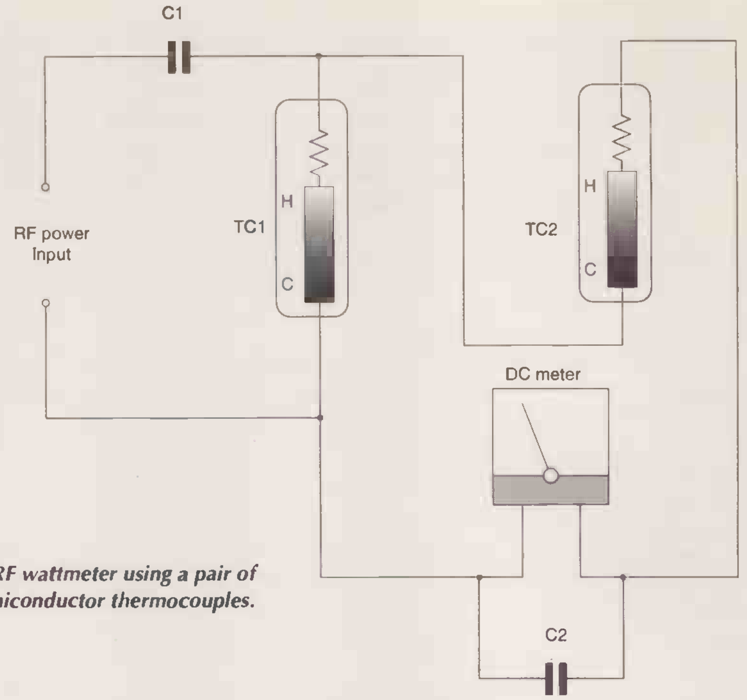
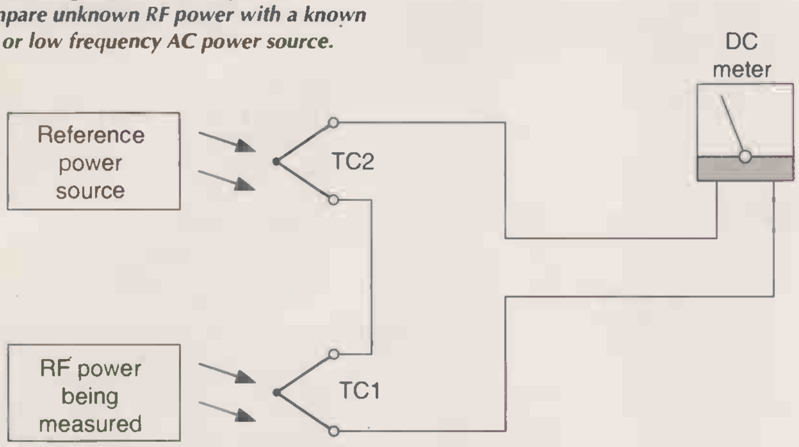


Fig. 7. RF wattmeter using a pair of semiconductor thermocouples.

Fig. 8. Using two thermocouples to compare unknown RF power with a known DC or low frequency AC power source.



the 1970s. Thermocouples are more sensitive than thermistor sensors, and are inherently square-law devices.

Figure 7 shows a solid-state thermocouple sensor that can be used well into the microwave region. Two semiconductor thermocouples are connected such that they are in series for DC, and parallel for RF frequencies. Thus, their combined output voltages are read on the DC voltmeter. Because of the capacitors, however, they are in parallel for RF frequencies, and if designed correctly will make a 50Ω termination for a transmission line.

Thermocouples suffer the same reliance on knowing the ambient temperature as thermistors. **Figure 8** shows a method for overcoming this problem. A pair of thermistors is used. One is used either in a bolometry circuit or as a terminating sensor to measure the unknown RF power. The other sensor is used to measure a highly controlled reference power source.

Depending on the implementation, the reference power might be DC, low frequency AC or another RF oscillator with a highly controlled, accurately calibrated output power level.

Diode detector RF power meters

Rectifying diodes convert bi-directional alternating current

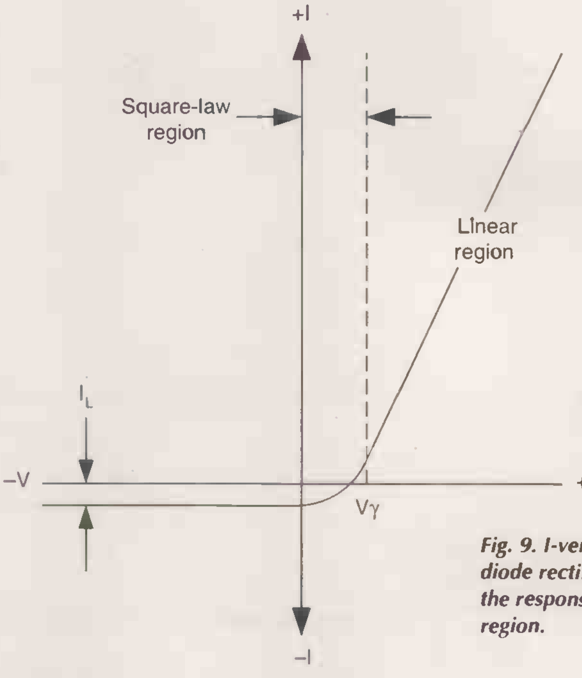


Fig. 9. I-versus-V curve of diode rectifier. Note that at Vγ, the response enters a linear region.

to unidirectional pulsating DC. When filtered, the output side of a diode is a DC level that is proportional to the amplitude of the applied AC signal.

Figure 9 shows the unidirectional action in the form of the *I-versus-V* curve. When the applied bias is positive – i.e. forward bias – the current begins to flow, but not proportionally. At some point between 200 and 300mV in germanium diodes or 600 and 700mV in silicon diodes, marked V_γ in Fig. **9**, the response enters a linear region. This response is termed ohmic because it follows Ohm's law.

When the applied voltage reverse biases the diode, the current flow ceases, except for a very small leakage current, I_L. One indicator of the quality of diodes is that I_L is minimised on higher quality units.

The nonlinear region of the *I-versus-V* curve is called the square-law region. In this region the rectified output voltage from the diode is proportional to the input power **Fig. 10**. This behaviour is seen from power levels of –70 to –20dBm.

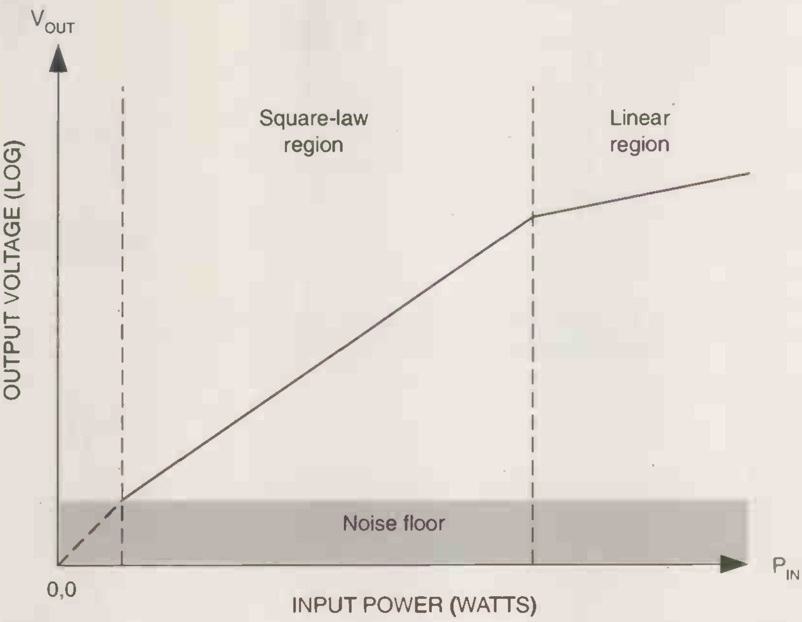


Fig. 10. Output voltage versus RF input power curve. In the square-law region, rectified output from the diode is proportional to input power.

In low-cost RF power measuring instruments, silicon and even germanium diodes are often used, but these are not highly regarded for professional measurements. Low-barrier Schottky diodes are widely used up to well into the microwave region.

For higher frequencies in the microwave region, planar doped barrier (PDB) diodes are preferred. They work up to 18GHz or better, and power levels of –70dBm. It is claimed that PDB diodes are more than 3000 times more efficient than thermocouple detectors.

Circuits

Figures 11a) and **11b)** show two similar circuits using a diode detector. Resistor R_L in Fig. 11a) is a dummy load that has a resistance value equal to the characteristic impedance of the transmission line connecting the system – 50Ω for example.

Diode D_1 is the rectifier diode, while capacitor C_1 is used to filter the pulsations at the rectifier output into pure DC. A problem with that circuit is that it is limited to power levels consistent with the native characteristics of the diode.

Figure 11b) shows the same circuit with a resistor voltage divider, R_1/R_2, to reduce the voltage associated with higher power levels to the characteristic of the diode. The actual voltage applied to the diode will be $V_{RL}\times[R_2/(R_1+R_2)]$. This circuit is similar to the metering circuit built into a number of low-cost amateur radio dummy loads in the past.

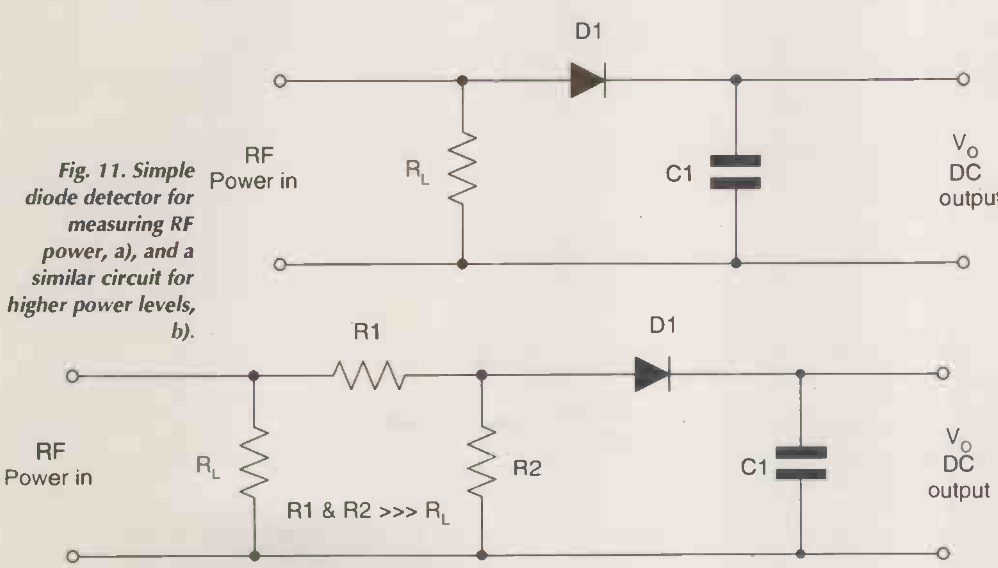


Fig. 11. Simple diode detector for measuring RF power, a), and a similar circuit for higher power levels, b).

Practical in-line bridge circuits

Low-cost in-line RF power meters, **Fig. 12**, are available using a number of different forms of bridge circuit. These are superior to the classical Wheatstone bridge because they can

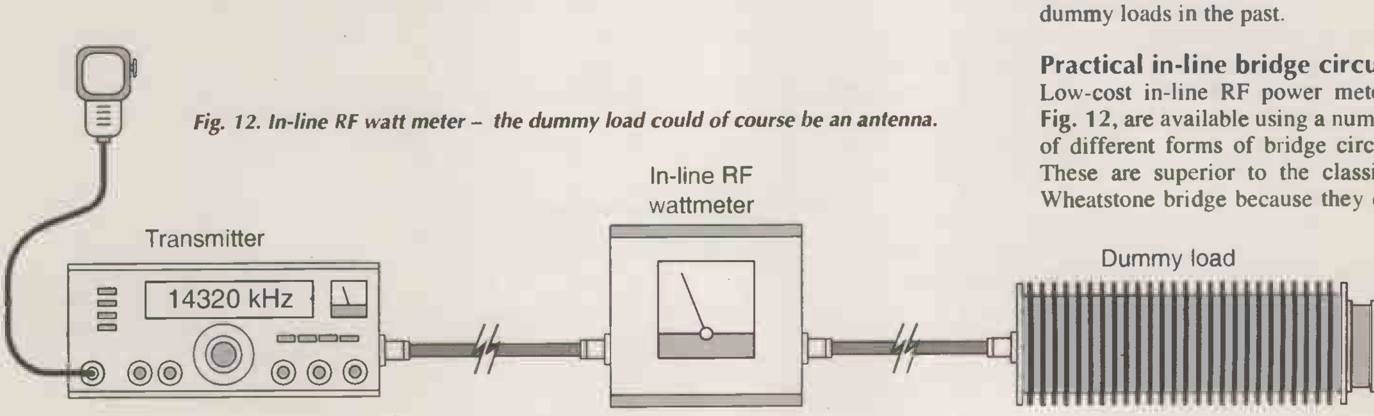


Fig. 12. In-line RF watt meter – the dummy load could of course be an antenna.

be left in-line while transmitting. Although the illustration in Fig. 12 shows a dummy load, it could just as easily use a radiating antenna.

Micromatch. One form of in-line RF power meter is the micromatch circuit of Fig. 13. This device is similar to a Wheatstone bridge in which the antenna impedance represents one arm, and a pair of capacitive reactances X_{C1} and X_{C2} represent two other arms.

Output voltage of the bridge is rectified by D_1, and filtered by R_2/C_3, before being applied to a microammeter. Note that this may be any meter from 100μA to 1mA full-scale.

The bridge consists of X_{C1}, X_{C2}, R_1 and R_L – the antenna or load resistance. The null condition exists when $X_{C1}/X_{C2}=R_1/R_L$. For 50Ω antenna systems the ratio R_1/R_L is 1/50, so a value of C_2 around 15pF is needed to produce the correct C_1/C_2 ratio.

For a 75Ω system, about 10pF is

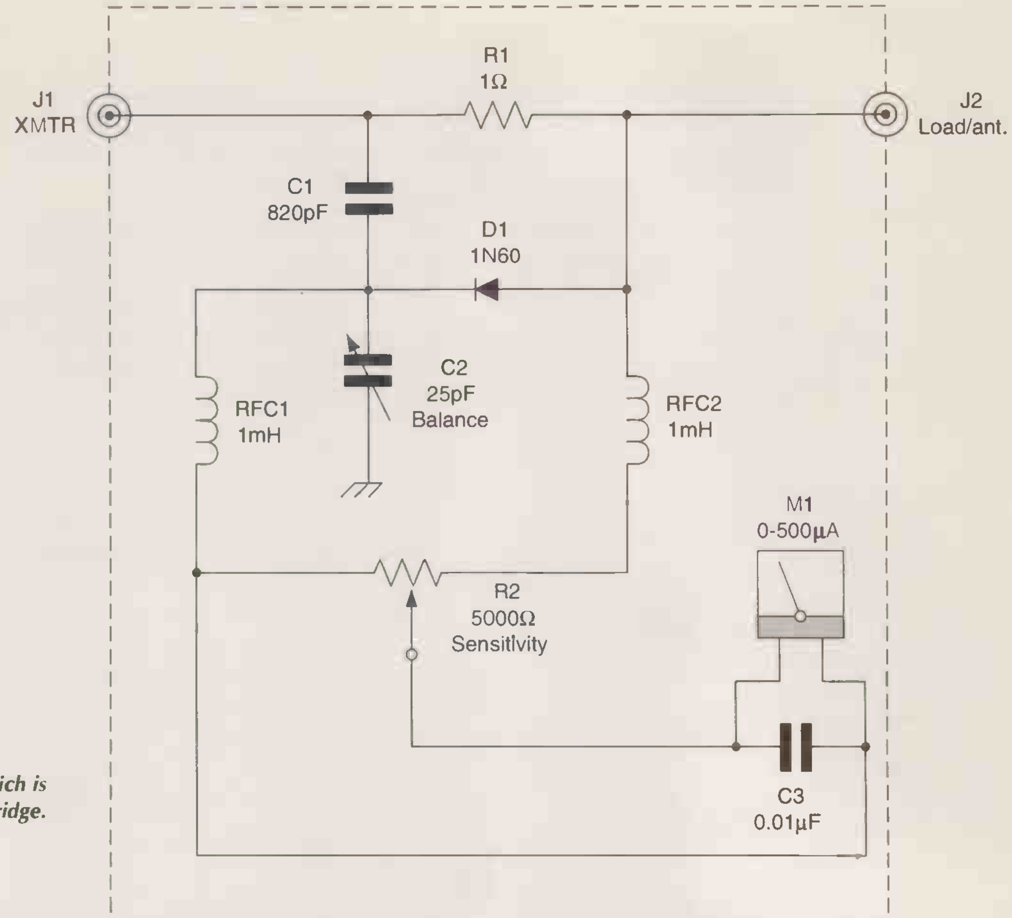


Fig. 13. Micromatch RF watt meter, which is similar to a Wheatstone bridge.

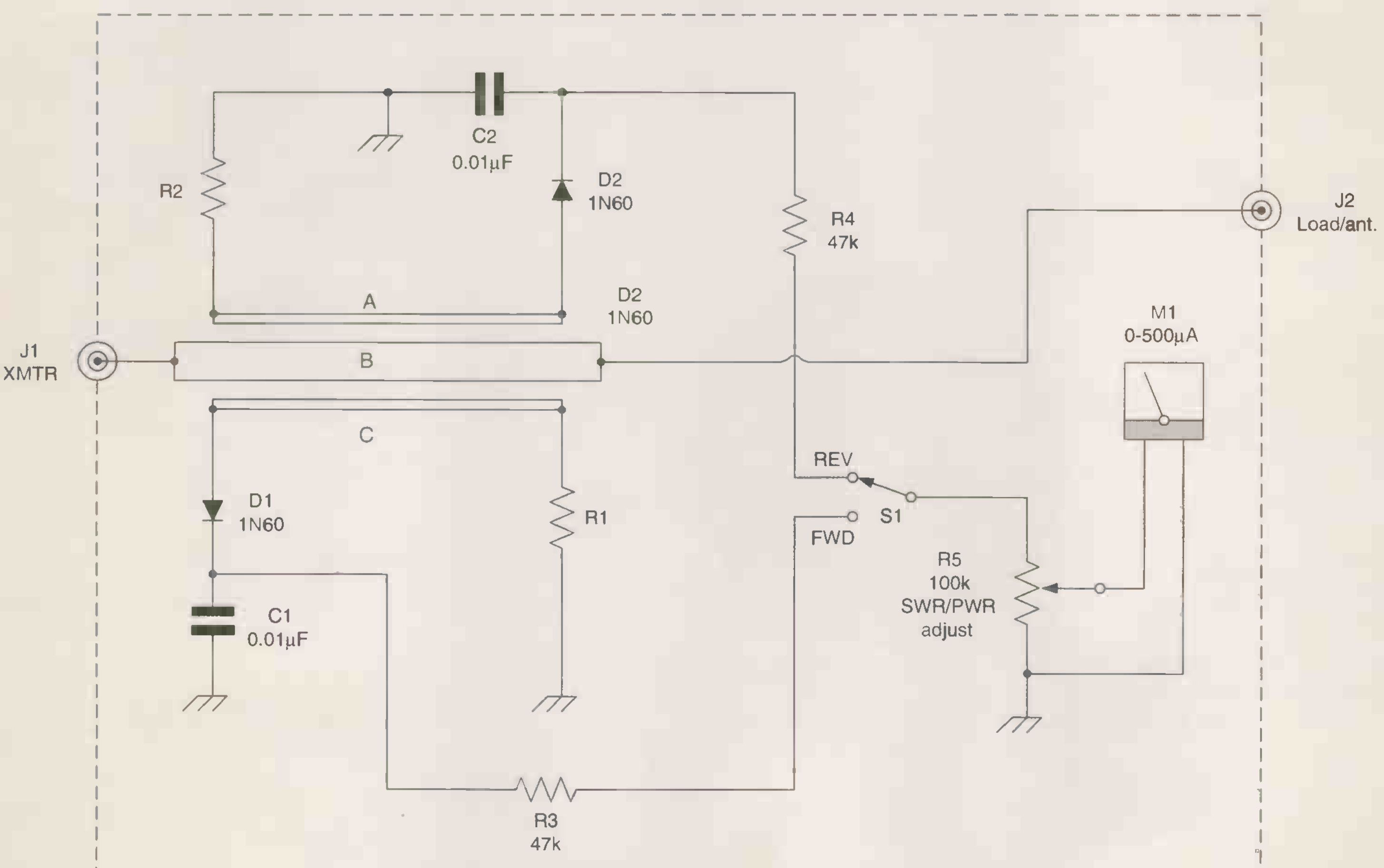


Fig. 14. Printed circuit mono-match RF wattmeter useful from HF to VHF.

needed. A number of people prefer to make a compromise by assuming a 68Ω load, so the capacitance needed in C_2 for a 1/68 ratio is about 12pF.

Series resistor R_1 is a one-ohm unit. In commercial micromatch RF wattmeters, this resistor is made using ten 2W, 10Ω resistors connected in parallel.

The RF power level is calibrated by adjusting the sensitivity control, R_2. In at least one commercial micromatch, there are actually three switch selectable sensitivity controls. These are calibrated for 10W, 100W and 1000W ranges.

Monomatch. The classical transmission line monomatch RF wattmeter is shown in **Fig. 14**. It can be used in the HF through VHF ranges. It consists of three printed transmission line segments – A, B and C – connected as a directional coupler.

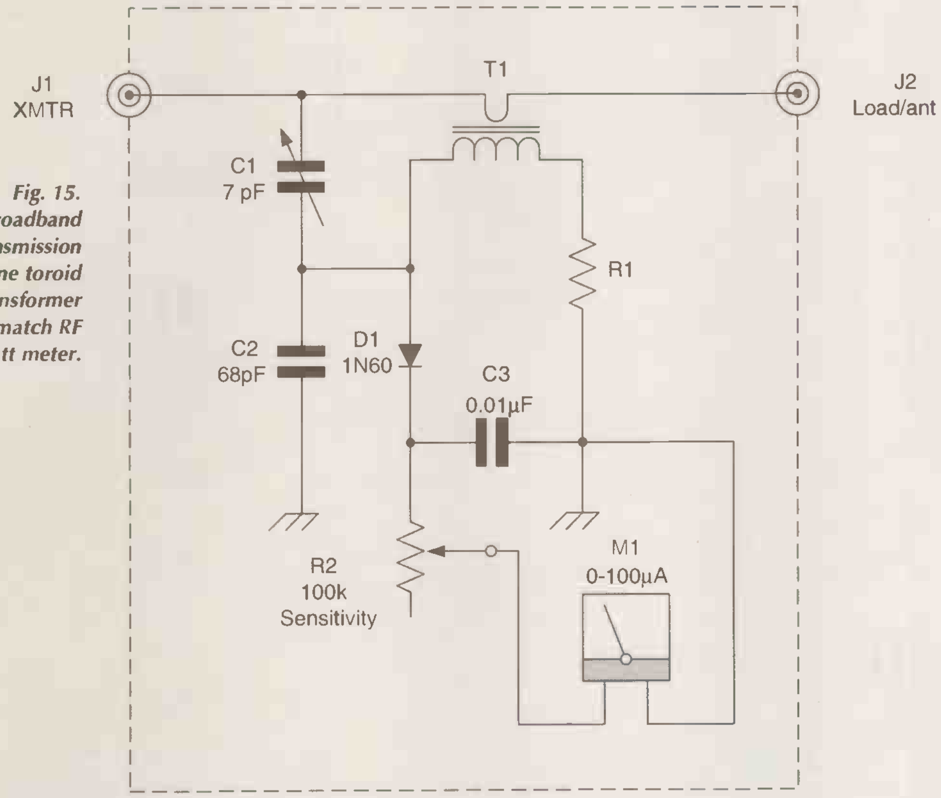


Fig. 15. Broadband transmission line toroid transformer mono-match RF watt meter.

In older instruments, the transmission line directional coupler was made using a length of RG-8/U coaxial cable with a pair of thin enamel insulated wires slipped between the shield and inner insulator. In more recent instruments the three transmission line segments are etched on a printed circuit board.

Sampling lines A and C are terminated in either 50Ω or 75Ω noninductive resistors such as carbon composition or metal film types. Again, a compromise value of 68Ω is often seen, so that either 50Ω or 75Ω antennas can be measured with only a small error.

Figure 15 shows an alternative monomatch system that uses a broadband transmission line transformer, T_1, made using a ferrite or powdered iron toroidal core. This circuit is usable throughout the HF range.

Detail for implementing the transformer is shown in **Fig. 16**. A 12 to 40mm toroid core is wound with 10 to 30 turns of #22 through #30 enamelled wire, leaving a gap of at least 30° between wire ends. A rubber grommet is inserted in the hole to receive the through transmission line.

Small diameter copper or brass tubing can be used, provided that it is a snug fit to the grommet.

In my second article on RF power measurement, I will discuss a commercial in-line RF wattmeter, and a calorimetry method used for high power RF measurements. I also intend to cover the problems of low-power measurement, and several error and uncertainty sources found in RF power measurements. ■

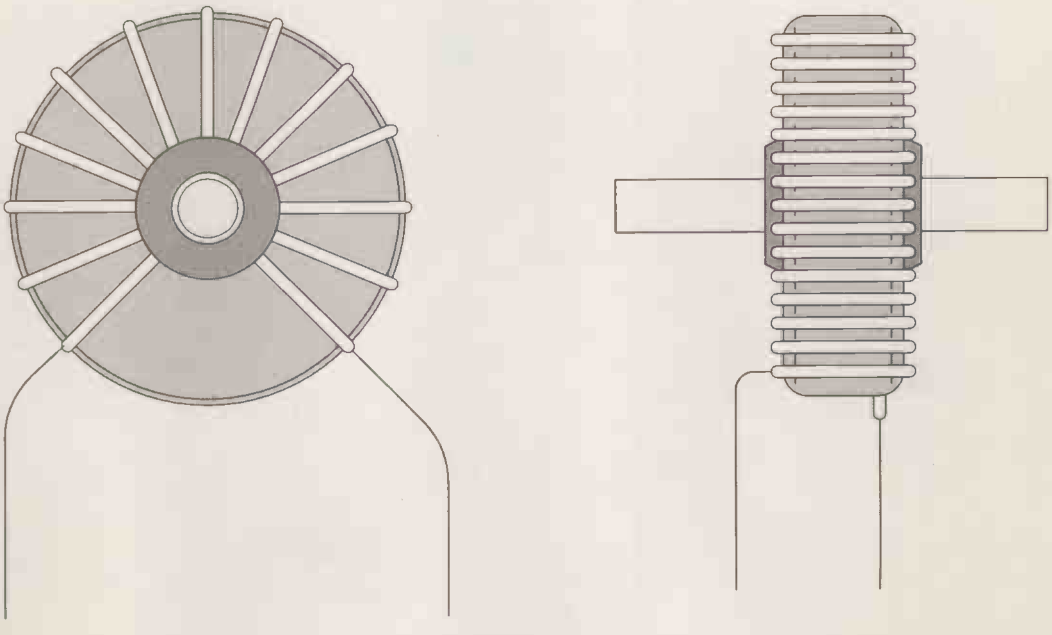


Fig. 16. Detail of transformer assembly for T_1 of Fig. 15.

CMOS output stages capable of swinging from rail to rail – even when powered by a single 1.5V cell – are discussed by leading researchers in the field of low-voltage, low-power analogue ICs. This third article includes an outline of why traditional current mirrors are unsuitable for low-voltage operation – and presents an entirely new solution to the problem.

Low voltage design III

When designing analogue building blocks that can handle rail-to-rail signal swings, classical output stages have to be abandoned. Take **Fig. 1**, representing a common-drain output stage. Clearly, the output voltage can only swing to within one gate-to-source voltage of the supply rail. If the circuit is operating from a 1.5V cell, the loss of dynamic range is intolerable.

In a low-voltage operational amplifier, the simplest output stage that can be used is the common-source stage. For this stage is, the minimum required supply voltage is,

$$V_{sup(min)}=V_{gso}+V_{d(sat)} \qquad (1)$$

where V_{gso} is the gate-source voltage of the output transistor M_{n1}, and V_{dsat} is the voltage across the current source I_{b2}.

This means that in a low-voltage environment, the common drain output stage has to be replaced by a common source output stage, as in **Fig. 2**. This has the advantage of giving an additive voltage gain, but it increases the output impedance.

Output voltage swing of this circuit is nearly rail-to-rail.

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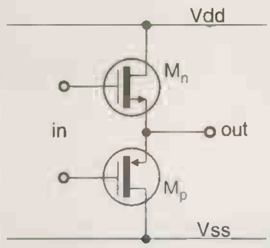


Fig. 1. Traditional common-drain output – not particularly useful for low-voltage op-amps since the nearest the output can swing to either rail is one gate-to-source voltage.

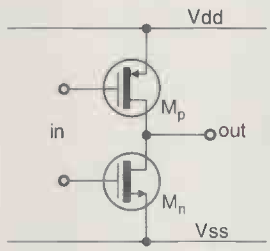


Fig. 2. In low-voltage designs, the common-source output stage allows rail-to-rail output swings and voltage gain at the expense of output impedance.

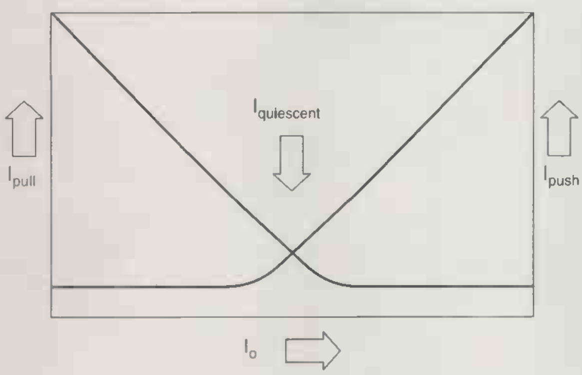


Fig. 3. Output transistor current versus current flowing into the load in a Class AB amplifier stage.

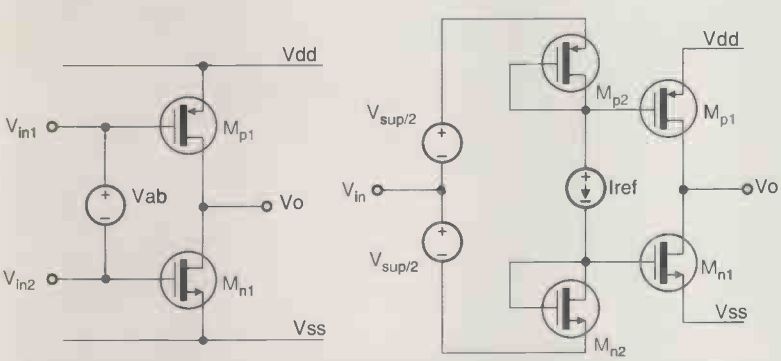


Fig. 4. Basic Class AB , left, in which generator V_{AB} provides a constant voltage difference between the gates to make sure that both devices remain turned on at all input voltages, and a method for implementing it, right.

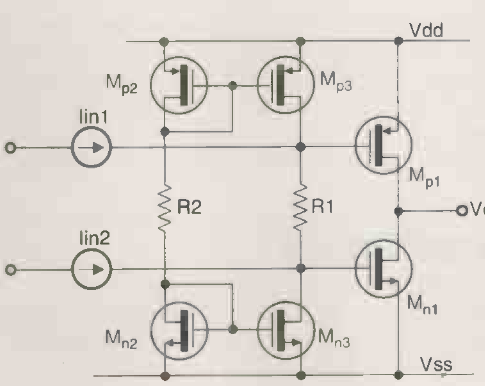


Fig. 5. Rail-to-rail output stage with resistive feed-forward Class-AB control. This configurations is limited in that minimum supply voltage has to be equal to two stacked gate-source voltages.

Class-AB output stages

To efficiently use the supply power, an output stage should combine a high maximum output current with a low quiescent current. To fulfil this requirement, a class-B output stage can be used, because its quiescent current is nearly zero.

Using class-B however leads to a large cross-over distortion. To overcome this drawback, class-A output stages could be used, but the maximum output current of a class-A biased output stage is equal to its quiescent current. This leads to low power efficiency – only 25% for a rail to rail output sine wave.

For these reasons, and in order to achieve a good compromise between distortion and quiescent dissipation, the output stage has to be biased between class-A and class-B. This solution is not surprisingly called a class-AB output stage. Its behaviour is characterised by **Fig. 3**.

The class-AB transfer function can be realised by keeping the voltage between the gates of the output transistors constant. In order to make this relationship independent of both the supply voltage and process variations, the voltage source has to track these parameters. To achieve this, the circuit technique of **Fig. 4** is used[1].

It is easily proved that,

$$\sqrt{I_{push}} + \sqrt{I_{pull}} = 2\sqrt{I_q} \qquad (12)$$

where the quiescent current I_q , given by

$$I_q = \frac{\left(\frac{W}{L}\right)_{M_{n1}}}{\left(\frac{W}{L}\right)_{M_{n2}}} \times I_{ref} \qquad (13)$$

is consequently insensitive to process and supply variations.

Feed-forward. A straightforward implementation of a class-AB biasing is shown in **Fig. 5**.

The diode-connected transistors, together with resistor R_2, build up a reference chain that generates a bias current I_{ref}. This current, copied by current mirrors, is fed into R_1. This resistor, in turn, sets the voltage between the gates of the output transistors.

Feedback[1]. The feed-forward class-AB control suffers from the limitation that the minimum supply voltage has to be equal to two stacked gate-source voltages and one saturation voltage. This prevents this circuit to operate under extremely low-voltage conditions.

The aforementioned limitation of feed-forward class-AB control can be overcome by using feedback class-AB control. In contrast to feed-forward control, this type of biasing does not directly control the current of the output stage, but the push and pull currents are first measured and then regulated in a class-AB manner. This allows the output stage to run on extremely low supply voltages. **Figure 6** shows a straightforward implementation of a feedback class-AB controlled output stage.

In order to obtain a quiescent current that is insensitive to process and temperature variations, resistor R_3 has to match R_2 and R_1, the current source I_{b1} has to be half the value of I_{b2}, and the *W/L* of M_{pb} has to be half the $M_{p7\text{-}8}$ aspect ratio. Using these values, it can be calculated that the quiescent current is given by,

$$I_q = \frac{\left(\frac{W}{L}\right)_{M_{p1}}}{\left(\frac{W}{L}\right)_{M_{pb}}} \times \frac{R_3}{R_2} I_{b1} \qquad (14)$$

Frequency compensation

An operational amplifier for use in mixed-mode systems has to be able to operate in a wide range of conditions. Among these conditions are unpredictable awkward loads and temperature fluctuations. Process variations also play a role.

Frequency compensation is a fundamental topic in op-amp design. It can be performed in different ways. Important considerations are Miller compensation and realising the pole splitting.

In low-voltage operational amplifiers, two different issues have to be considered. One is realising a constant-g_m input stage that allows simple compensation in all the operating regions of the input pairs.

The second is that, in general, a larger number of stages is needed relative to traditional topologies. This implies an increased accuracy in the frequency compensation, and, in particular, the need for a large number of capacitors. Nested Miller compensation[2] has been developed in order to solve this problem.

Next, the main aspects of Miller compensation and pole splitting are discussed.

Current generators designed in CMOS technology are covered in the panel entitled 'Biasing CMOS in low-voltage designs'. In particular, two low-voltage current mirror structures are shown.

In a subsequent article, we present a low-voltage rail-to-rail constant-g_m operational transconductance amplifier. We will also show examples of novel circuits working at low supply voltages. Among these are the adaptive biasing technique and switched op-amps.

There will also be an outline on transistors models used in commercial Spice in low-voltage applications.

Commercial products operating at low supply voltages.

	Name	Type	Supply voltage
Motorola			
	Power PC 750	Microprocessor	1.9V
	MC 12147	VCO	2.7 - 5.5V
National Semiconductor			
	LMC 6582/84	Op-amp	1.8V (lowest)
	LM 6142/44	Op-amp	1.8V (lowest)
	LMC 6442	Op-amp	2.2V
	LMC 6085/86	Op-amp	2.7V (lowest)
	LMC 272	Op-amp	2.7 - 15V
	LMC 7101	Op-amp	2.7V (lowest)
	LMC 6482/84	Op-amp	3 - 5V
	LMC 6462/64	Op-amp	3 - 5V
Analog Devices			
	OP 191/291/491	Op-amp	2.7 - 12V
	AD 627 +2.2	Instr. Amp.	2.2 - ±18V
Philips			
	NE/SA 5230	Op-amp	1.8 - 15V
Gennum			
	LC/LD505	Class A amp	1.3V
	LC/LV506	Pre-amp	1.1 - 1.55V
	LC/LD/LV549	Power-amp	1.0 - 1.6V

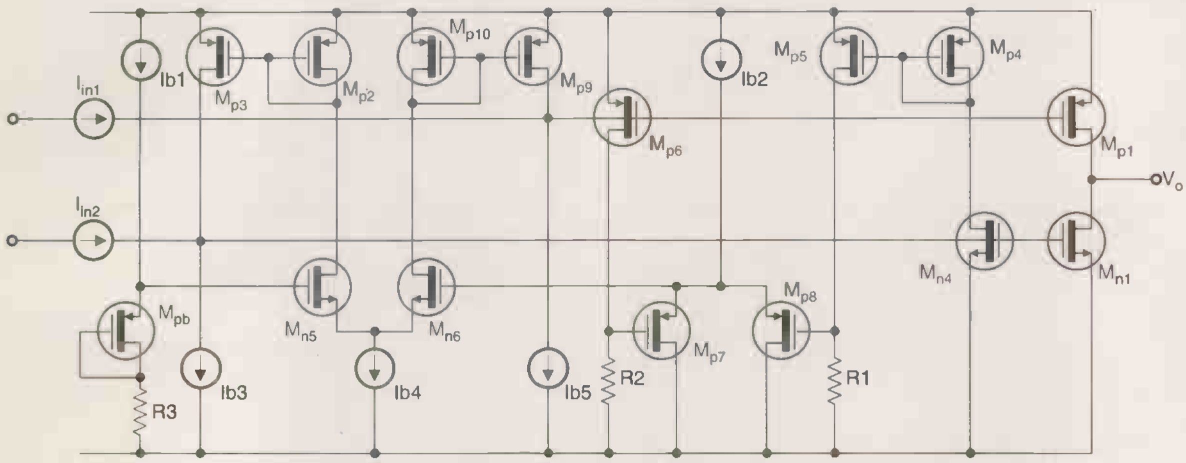


Fig. 6. Feedback-biased Class-AB rail-to-rail output stage. Currents through the output transistors are measured by resistors.

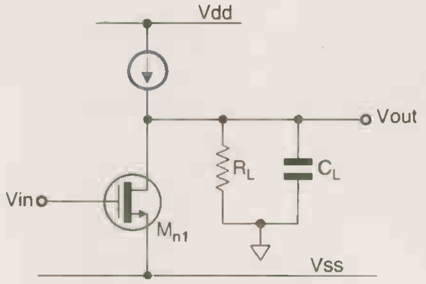


Fig. 7. Simplified single-stage amplifier – stable, and popular in many analogue building blocks. Sadly though, not the ideal solution for low-voltage designs.

Splitting poles

Pole splitting is a useful way to compensate the operational amplifier and make it stable in frequency operation. Here we look at the main pole-splitting compensation mechanisms, related to different numbers of stages for the op-amp.

This analysis involves CMOS circuits, but it applies equally to bipolar circuits.

Single-stage amplifiers are popular in VLSI circuits because of their excellent high-frequency behaviour. This makes them highly suitable for use in high-performance switched-capacitor circuits and analogue-to-digital converters. Because of the single-stage topology, there is no need for frequency compensation and the amplifiers are inherently compact.

A simplified schematic of a single-stage amplifier is shown in **Fig. 7**. The amplifier has only one dominant pole and is, therefore, always stable.

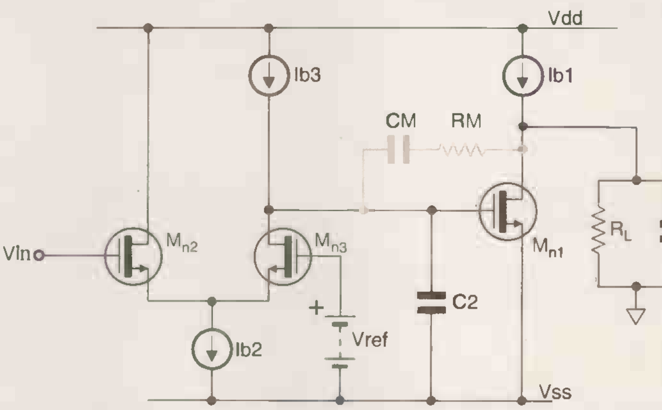


Fig. 8. Two-stage amplifier – more gain, but the additional pole complicates

Biasing CMOS in low-voltage designs

In CMOS integrated circuits, a stable current generator capable of realising current references of any value is needed. Consider the topology universally used to generate DC currents, which can be a multiple of the reference current value. The circuit, – the current mirror – is shown in **Fig. A1**.

Here, the mirror comprises two transistors, matched with respect to their threshold voltages, but having different *W/L* ratios. Transistor M_{n1} is biased by reference current I_{ref}. Output current is taken via the drain of M_{n2}, which has to be in saturation.

Neglecting the effect of channel length modulation, for transistor M_{n1},

$$I_{ref}=\frac{KP}{2}\left(\frac{W}{L}\right)_1\left(V_{gs}-V_T\right)^2$$

where V_{GS} is the gate-source voltage corresponding to a drain current equal to I_{ref}.

Since M_{n2} in connected in parallel with M_{n1}, it has the same V_{GS}. So, neglecting the effect of channel-length modulation,

$$I_{out}=\frac{KP}{2}\left(\frac{W}{L}\right)_2\left(V_{gs}-V_T\right)^2$$

From the two equations above,

$$I_{out}=I_{ref}\frac{\left(\frac{W}{L}\right)_2}{\left(\frac{W}{L}\right)_1}$$

Ideally, I_{OUT} is a multiple of I_{ref} and its value is only determined by the geometry of the device.

In practice, this value of I_{OUT} can be precisely obtained if the drain-voltages of M_{n1} and M_{n2} are equal. However, the output impedance that MOS devices exhibit has not been taken into account; remember that a drain-voltage variation produces an I_{OUT} variation. The shortening of *L* is taken into account in the effect of channel-length modulation as,

$$I_{DS(sat)}=\frac{KP}{2}\times\frac{W}{L}\left(V_{gs}-V_T\right)^2\times\left(1+\lambda V_{ds}\right)$$

where $1/\lambda$ directly fixes the slope of the output characteristic of the transistor. Note that λ is inversely proportional to *L*

Transistor output impedance can be calculated as the inverse ratio of the output current with respect to output voltage, expressed by,

$$r_o=\frac{1}{\lambda I_{DS(sat)}}$$

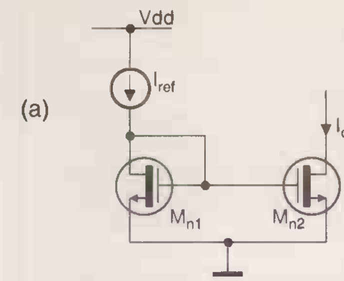
This shows that r_o increases proportionally to *L*. So an optimal current mirror can be realised if two transistors have equal drain voltages and if the drain-voltage change of the output transistor is made as small as possible.

An easy way to increase the output impedance is use a cascode current mirror, **Fig. A2**. It is easily shown that,

$$r_o=\frac{V_x}{I_x}=r_{o4}+r_{o2}+g_{m4}r_{o4}r_{o2}\approx\left(g_{m4}r_{o4}\right)r_{o2}$$

The cascode connection between M_{n2} and M_{n4} increases the output resistance from r_{o2} to $(g_{m4}r_{o4})r_{o2}$, with a gain factor equal to $g_{m4}r_{o4}$.

Similar results can be obtained with the Wilson circuit, shown in **Fig. A3a)**. This topology needs fewer MOS devices, but suffers



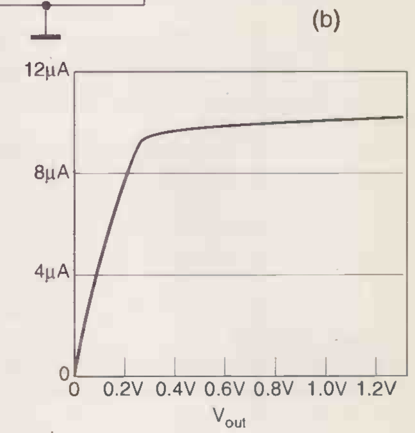
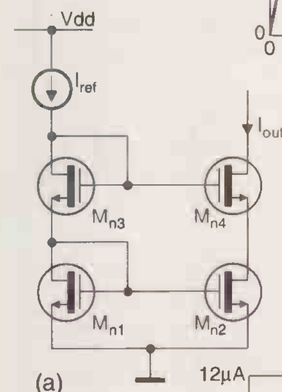


Fig. A1. Basic scheme of current mirror. Current I_{ref} is set to 10μA.



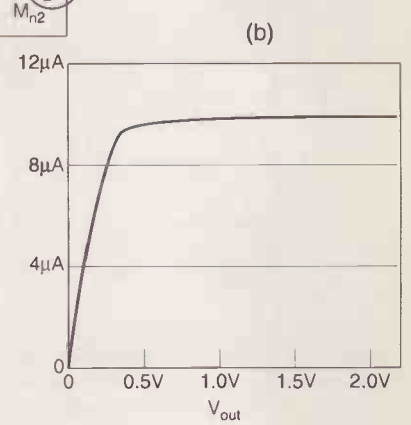


Fig. A2. Cascode arrangement improves the performance of the current mirror. This is the circuit and its characteristic.

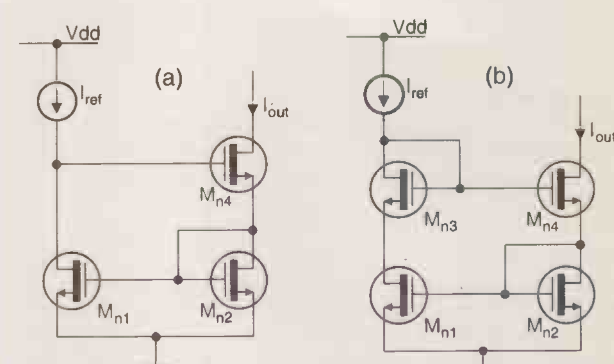


Fig. A3. Similar in performance to the cascode arrangement but with fewer components is the Wilson current mirror, a). An improved Wilson current mirror is shown in b).

from the asymmetry of $M_{n1,2}$ biasing voltages; this causes the output current to be different from the reference current.

It is possible to overcome the asymmetry drawback using the structure shown in **Fig. A3b)**, where the diode-connected transistor M_{n3} ensures a better symmetry of the circuit.

Though the cascode current mirror improves the output resistance, it has a smaller output dynamic range. As the graph of Fig. A2**b)** shows, the minimum voltage drop on the real current generator is quite large. This voltage is $V_T+2V_{ds(sat)}$.

Even if $V_{ds(sat)}$ can be made very small by using both large width values and biasing transistors with low current, the threshold-voltage term represents an unacceptable loss of dynamic range – especially in low-voltage circuits.

A biasing scheme that improves the output range involves transistor M_{n2} of Fig. A2**a)** being biased near to its saturation region limit. This can be achieved by connecting a voltage translator in series with the gate of M_{n4}. The voltage translator is implemented with a source follower, M_{n5} in **Fig. A4a)**.

The lowest voltage that the circuit works at is $2V_{ds(sat)}$. However, the circuit suffers from the fact that M_{n2} has a drain voltage entirely different from that of M_{n1}. This introduces an error in the output current, Fig. **A4b)**.

To solve this problem, the circuit of **Fig. A5a)** is proposed. Here, the gate voltage of M_{n1} and M_{n2} is connected to the drain of M_{n3}. By a suitable choice of V_{bias}, the output voltage range can be reduced to about $2V_{ds(sat)}$. In particular, it has to be,

$$V_{bias} - V_{gs}(M_{n4}) \approx V_{ds(sat)}(M_{n2})$$

If $V_{bias}=V_d(M_{n3})$, the current mirror of **Fig. A6a)**, is obtained. This configuration works better in the subthreshold region, as shown in plot Fig. **A6b)**.

As with all current mirrors, the drain voltage of M_{n2} needs to be kept above a certain value. In the subthreshold region, letting $V_{ds}(M_{n2})$ be larger than a few times U_T is sufficient. In this region the MOS current equation is given by,

$$I_d = \frac{W}{L} I_{d0} \exp\left(\frac{V_{gs}}{nU_T}\right)$$

where I_{d0} is reverse saturation current, n is the subthreshold slope factor, and $U_T=kT/q$ is the thermal voltage.

This opportune voltage can be obtained with a suitable sizing of transistors, since the following relationships are valid,

$$V_{ds(M_{n2})} = V_{gs(M_{n2})} - V_{gs(M_{n4})} = V_{gs(M_{n2})} - V_{gs(M_{n3})}$$

$$= n\frac{kT}{q} \ln\left[\frac{\left(\frac{W}{L}\right)_{M_{n3}}}{\left(\frac{W}{L}\right)_{M_{n2}}}\right]$$

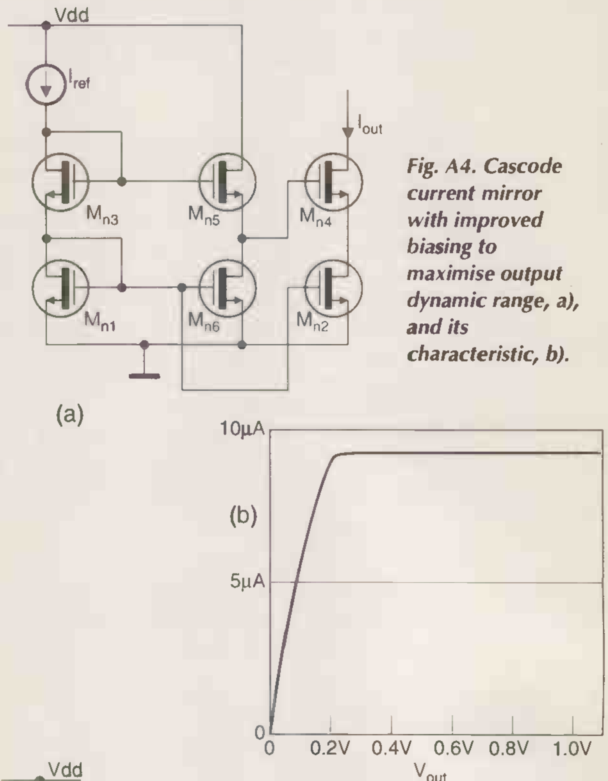


Fig. A4. Cascode current mirror with improved biasing to maximise output dynamic range, a), and its characteristic, b).

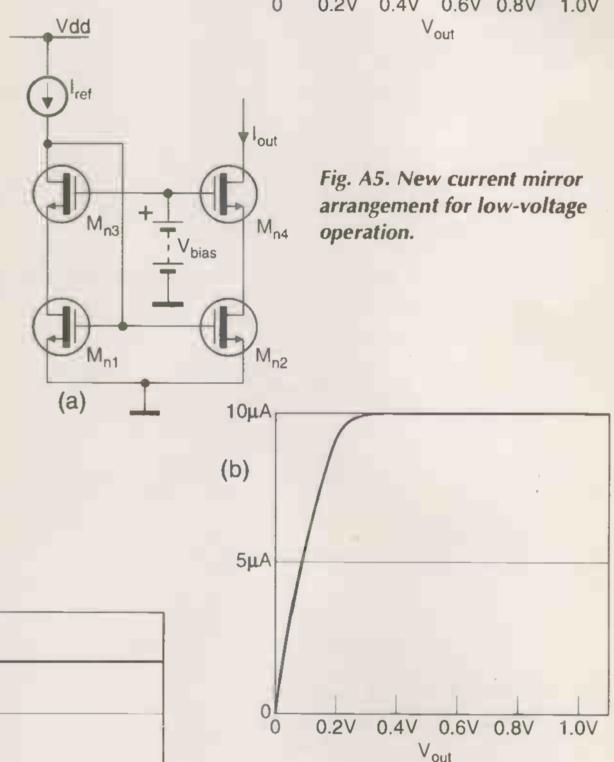


Fig. A5. New current mirror arrangement for low-voltage operation.

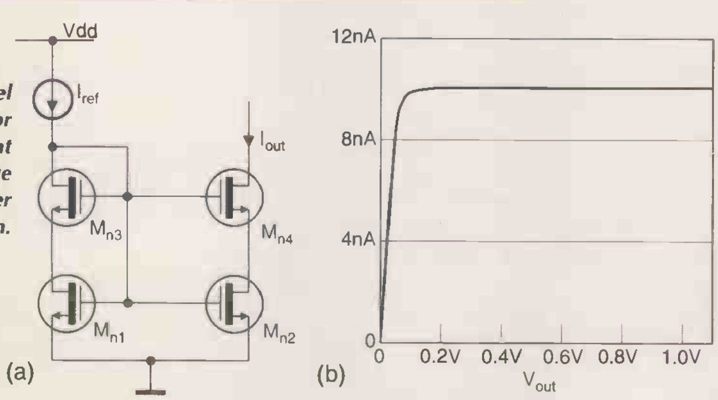


Fig. A6. Novel current mirror arrangement for low-voltage and low-power operation.

The unity-gain frequency of this amplifier, GBW, is given by,

$$GBW = \frac{g_m}{2\pi C_L} \tag{15}$$

The dc voltage gain of the amplifier is determined by the transconductance g_m and the load resistance R_L, which also incorporates the output impedance of the transistor.

$$A_0 = g_m R_L \tag{16}$$

The gain of this amplifier is usually about 40dB. It is possible to add gain by using a transistor cascode or, for even more gain, the gain-boosting technique.

In many applications, the gain of a single stage is too low, especially when the output is heavily loaded. In addition, the existing cascoded stages – folded cascode for example – are not suitable for very low voltage electronics.

In those cases, using two gain stages increases the overall gain of an amplifier, **Fig. 8**. Each of these gain stages introduces a dominant pole at its output. Consequently, the amplifier behaves like a two-pole system.

The bandwidth of the two-stage amplifier is equal to the geometric mean of the two stages. For CMOS stages operating in strong inversion, the bandwidth is given by,

$$GBW = \frac{\beta\left(V_{gs} - V_T\right)}{\pi\sqrt{2C_2C_L}} \tag{17}$$

The poles of the uncompensated amplifier are situated at the output of each stage. The first pole, at the output of the amplifier, is located at,

$$p_1 = \frac{1}{R_L C_L} \tag{18}$$

The output resistance of the input stage, r_{o2}, and the gate-source capacitor, cgs1, of the output transistor, determine the second pole, located at the output of the input stage. It is given by,

$$p_2 = \frac{1}{r_{o2}c_{gs1}} \tag{19}$$

In order to ensure stability in feedback configurations, the amplifier has to act like a one-pole system up to its unity-gain frequency. It is possible either to insert a Miller $R_M C_M$ network or to apply a parallel $R_P C_P$ network.

The drawback of the parallel alternative is that the compensation method relies on matching with the load impedance, which is often not defined. On the other hand, the Miller technique is robust against parameter variations. This makes it the best compensation technique for two-stage amplifiers.

In order to appreciate Miller compensation, consider the effect of inserting only capacitor C_M. If the loop is closed, poles p1 and p2 are split apart, as in **Fig. 9**. This clearly follows from the root locus of the two-stage amplifier for a varying Miller capacitor.

The root locus starts from the poles of the uncompensated amplifier, when C_M equals zero, and ends at the zeros, when C_M is infinite, or very large.

In practice, such large capacitors cannot be realised. Consequently, the non-dominant pole of the compensated amplifier ends up at a lower frequency. Simple calculations show that this non-dominant pole (see **Fig. 10**) ends up at a frequency of,

$$P_1' = -\frac{g_{m1}}{C_L\left(1 + \frac{c_{gs1}}{C_M}\right) + c_{gs1}} \tag{20}$$

The Miller capacitor also gives a direct feed-forward path to the output at high frequencies. This results in a zero situated in the right plane, therefore an additional phase-shift is introduced. This zero is positioned at,

$$z = \frac{g_{m1}}{C_M} \tag{21}$$

However, this zero can be exactly cancelled if resistor R_M, inserted in series with the capacitor C_M, is dimensioned such that,

$$R_M = \frac{1}{g_{m1}} \tag{22}$$

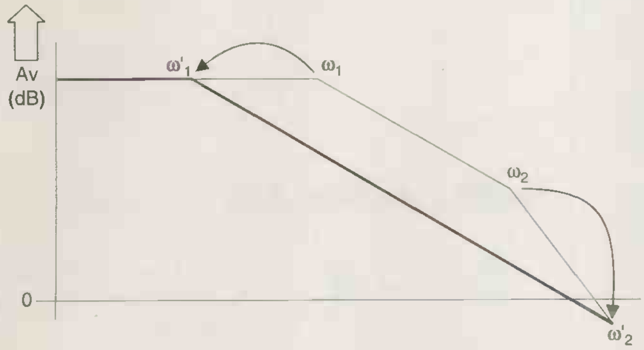


Fig. 9. Bode plot of the two-stage amplifier with Miller compensation.

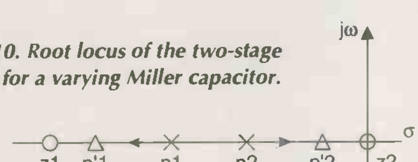


Fig. 10. Root locus of the two-stage amplifier for a varying Miller capacitor.

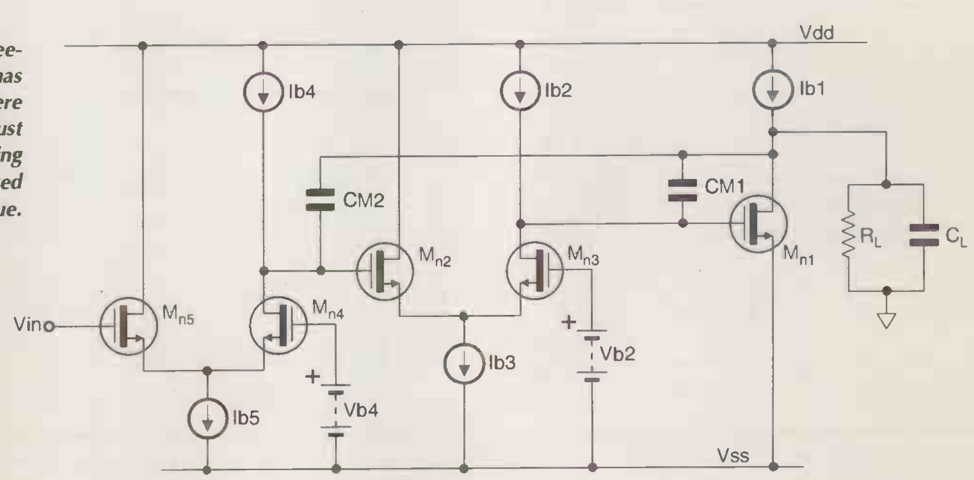


Fig. 11. Basic three-stage amplifier has three poles, but there is a simple and robust way of compensating using the nested Miller technique.

Three-stage amplifiers

Placing an additional gain stage between the input and output stages can further increase the gain of a two-stage amplifier.

The basic topology of a three-stage operational amplifier is shown in **Fig. 11**. This amplifier contains an input stage, M_{n4-5}, an intermediate stage, M_{n2-3}, and an output stage, M_{n1}. Each stage introduces a dominant pole at its output.

A simple and robust method to compensate this amplifier is the nested Miller compensation technique.[2]

Figure 12 explains the principle of this compensation technique. The output and intermediate stages can be conceived as a two-stage amplifier with two dominant poles, f_1 and f_2.

Capacitor C_{M1} closes the first Miller loop, which split f_l and f_2 to f'_1 and f'_2 respectively. The pole f'_1 is 3dB below the unity-gain frequency. Therefore, the intermediate and output stage can now be treated as one stage with one dominant pole f'_2.

The Miller splitting can be repeated by inserting C_{M2}. This capacitor splits the pole f_3 and f'_2, resulting in one dominant pole at f''_2. ■

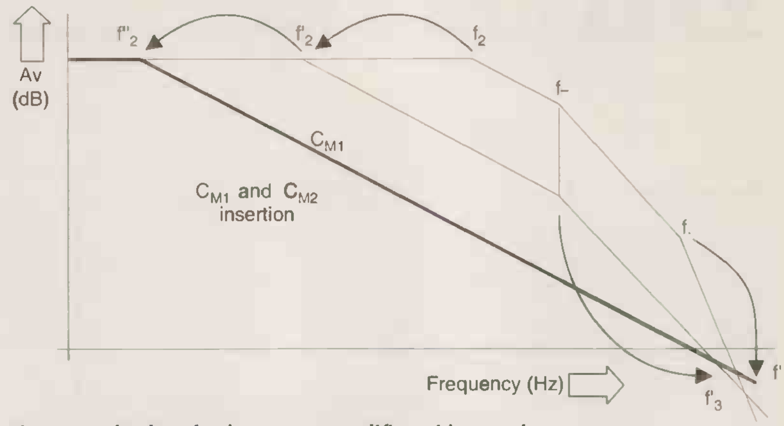


Fig. 12. Bode plot of a three-stage amplifier with nested Miller compensation.

Our next article on this topic, planned for the December issue, explains why switched-capacitor techniques are not suitable for use at low voltages – together with an alternative that is. A new OTA for very low voltages is presented, together with a brief discussion of Spice in relation to low-voltage design.

References

1. R. Hogervorst, J.H. Huijsing, 'Design of low-voltage low-power operational amplifier cells', Kluwer Academic, 1996
2. G. Ferri, W. Sansen, V. Peluso, 'A low voltage fully differential constant-G_m CMOS operational amplifier', *Analogue Integrated Circuits and Signal Processing* Vol. 16 No 1 April 1998, pp. 5-15.

A comprehensive list of references was presented with the first of these articles, in the September issue.

LETTERS

Letters to "Electronics World" Quadrant House,
The Quadrant, Sutton, Surrey, SM2 5AS
email jackie.lowe@rbi.co.uk

Line-powered fault

The circuit idea for a line powered telephone monitor submitted by a reader from Holland on page 370 of the May 1999 issue will probably cause problems if connected to a System X exchange of the UK telephone network.

Such exchanges carry out automatic line testing, and one of the tests appears to be a leakage test – a momentary line polarity reversal at a lower voltage than 48V.

The inclusion of the diode bridge in the circuit will ensure that current will be drawn during this test. This is detected as unwanted leakage and the test is repeated, apparently, *ad infinitum*. This is manifested as a slow flashing of the LED indicator.

The cure is to omit the diode bridge, and ensure that the line polarity is correct.

A very similar circuit – without bridge – can be found in *Telephone Installation Handbook*, (Roberts 1997) ISBN 07506 34278, page 147.

Those proposing to use the circuit should also be aware that the connection of unapproved devices to the Public Switched Telephone Network is not permitted.

Steve Roberts
Bude
Cornwall

12 or 24V supplies accepted

A small comment on the '12V or 24V supplies accepted' circuit idea published in August 1999 *EW*. This circuit regulates a 24V input or passes a 12V input to give a 12V output, regardless. There is a small flaw that when 24V is asserted on the input, the circuit delivers 24V on the output for the few milliseconds it takes the relay to open. This transient may be enough to destroy the following circuit.

Tim Herklots
Via e-mail

Pied pipings

The flute on the front cover of August 1999 issue appears to be a mirror image. There is no such thing as left handed flute.

Is this a subtle visual pun to illustrate the effect of phase inversion on stereo image perception or did someone simply get the photograph the wrong way round?

Allan Winsor
Via e-mail

Please would you publish details of where to obtain the left handed flute illustrated on the front cover of the August 1999 issue of *Electronics World*. Please also pass my sympathies to the grossly deformed player whose left hand little finger is clearly twice as long as a normal finger and excessively arched. Just so you are in no doubt at my amusement, the illustration is laterally transposed, and whoever assembled the flute for the illustration set the foot joint about 90° away from the normal position. It would certainly be a challenge to play as illustrated.

You're not the first journal to have musical egg on their faces. I can never understand why normally technically competent people get musical instruments so wrong.

John Holt
Via e-mail

I mentioned it to the photographer John, and he said, "Has he nothing better to do?" Needless to say, I reprimanded him severely. Ed.

Dropper memories

Regarding series valve heater chains fed by capacitors, I plead guilty to Bob Pearson's charge of naivety in failing to consider switch-on transients.

However, in my humble job I was in no position to recommend anything – let alone 2μF capacitors.

In fact I was naive enough to assume that the technique must be faulty because it was not commercially used at the time, in spite of the fact that none of the several portable radios I built ever suffered from blown heaters. I remember only being uneasy about what peak voltage was applied across the capacitor terminals and the lack of a specific terminal to case rating.

I was supremely unqualified for this work, having passed out as a conscript RAF Air Radar Mechanic only about a year before and having no experience of 'civvy' radio or TV (such as it was) other than wiring up a crystal set. I was equipped with little more than a clumsy soldering iron, a copy of the 1938 Admiralty Handbook of Wireless Telegraphy Vol.1, and permission to stay in the Radar Section after working hours doing PJs (private jobs).

From page 500 of your sister magazine 'Television' this month, I learned for the first time that Thorn fed the valves in their 960 series portable TVs via 'Wattless' capacitor droppers. This belated use would have soon been rendered obsolete by solid-state devices, but I remain proud to share my naivety with such a renowned manufacturer.

John Norman MSERT
Via e-mail

Further to the correspondence on capacitor droppers for heater supplies, the last Rediffusion monochrome TVs to use any valves (Marks 12/13 for those who remember) had series heaters for the tube, line output pentode, and efficiency diode.

If my memory serves, these were 300mA @ 63V in total. This was provided by an auto transformer which was of course quite bulky and expensive. We looked briefly at a capacitor dropper but the problem we hit was to keep the current within the limits – especially for the tube. The capacitor's tolerance was if I remember rightly 4%, which coupled with 250V AC working would have probably cost more than the auto-transformer. It would not have been much smaller either.

Bryan Simmons
Surbiton
Surrey

With reference to the heater debate. I have been a service engineer for over thirty five years and Ferguson has used both a capacitor and auto transformer for 300mA heater chains in tvs.

Ferguson 900 series used an auto transformer. Ferguson 980, a portable mono set, used a series capacitor. The same set even had a delay in which one or two valve heaters were in the negative leg of the set's ht voltage to produce a warm up delay.

The capacitor or transformer did not have any unduly high failure rates.

P. Edenbrow
Sleaford
Lincolnshire

Where did the weekend go? Take 2

31 December 1999 will be a Friday.
1 January 2000 will be a Saturday.
1 January 1900 was a Monday.
Will computers that have not been modified revert to the 1900 calendar and hence jump from Friday to Monday losing two days?

R N Soar
Doncaster

Sorry Mr Soar. Fingers crossed I got it right this time. Ed.

Engineering women

With reference to your response to John Phillips' letter 'Engineering women' in the August issue, I would accuse you of being more naive than patronising. Your metaphor either implies that women have just recently been allowed to become engineers, or even worse forget everything that they learn after a few days!

The women engineers that I have met are generally more competent than men. Unfortunately there are too many companies that share your view encouraging a predominantly male engineering workforce.

Jon Fuge
Via e-mail

Eh? Ed.

Aether way...

Recently, a book entitled 'Handbook of Wireless and Telelgraphy – 1938 Volume 1', by the Admiralty came into my possession. Judging by its difficulty level, I would say it is equivalent to today's BTEC level 4 or first year HND.

The book kicks off by explaining waves, etc., and finally refers to what should be electromagnetic waves as 'Aether' waves. The book devotes two long paragraphs to the subject.

There are two very interesting sentences which I quote from the book, *"All experience goes to show that light and electromagnetic energy generally are transmitted through space as a wave motion, and we are led to the supposition that all space is occupied by a medium which conveys the energy, and that this medium has properties different from those possessed by ordinary matter. We call this medium the 'Aether'. The medium called the 'Aether' must necessary be universally diffused and must inter-penetrate all matter. It cannot be exhausted or removed from any place, because no material is impervious to it."*

It is pointless for me to describe the countless experiments that have been done that agree accurately with both the Special and General theories of Relativity. As an example, electronic ballistics – as in oscilloscopes – are aptly described with use of Special Relativity theory.

What I find strange is that Michelson and Morley proved that no Aether exists as far back as 1887. The Special Theory of Relativity was first published in 1905, solving many problems in physics that Newtonian theory simply could not solve. It seems that the theory of Relativity had not yet filtered through to Admiralty – even after 33 years. Or had it? Was the Admiralty defiant of the new theory for various reasons?

Can anyone shed some light on the subject?

Darren Heywood
Buckley
Flintshire

Where did Otto go?

What happened to Dr Otto Schmitt? This is a mystery that I have never managed to solve. His eponymous circuit remains in use while other names have faded into history; Miller, Puckle, etc.

I have a copy of his original article, *A Thermionic Trigger*, published in 1938. The circuit was simply and accurately described and I remember studying it in my valve days. I even adapted it to use a transistor and phototransistor to act as a light operated switch in 1956.

The phototransistor was on a probe and was used for timing the operation of photographic printers.

Electronics World's series of articles on great inventors in the field of electronics and radio (wireless) did not include Schmitt. Although he was working in a British university when the article was published, he may have been of German nationality and returned home at the outbreak of war.

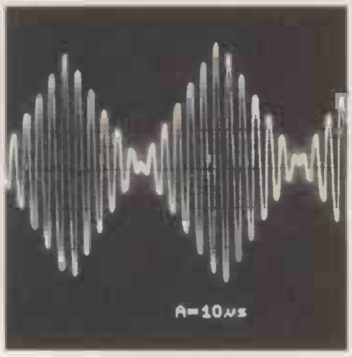
I am sure other readers would be interested in hearing what happened to him as many use the name without realising the origin.

Guy Selby-Lowndes
Billingshurst
Kent

How did I do it?

Sorting through my many papers, I came across the attached. It does not relate to anything that I have done for at least five years and I cannot remember how I did it! Can anyone suggest how they would set about creating such a snappy waveform simply?

Nick Wheeler
Sutton
Surrey



Magnet memories

I am an ancient electronics person. There is much to be said for the letter from Adrian Jansen, Letters August 1999, regarding speaker coils. I remember speakers with magnets from tungsten, cobalt alloy and Alnico. I think these materials provided not only smaller magnets, but stronger fields.

Could it be that the better quality of Neodimium magnet assemblies is mainly due to extra damping of unwanted resonance?

Copper plating is sometimes used on soft iron or steel to prevent rusting which could result in particles touching the moving coil. Surely the coating is too thin for any other purpose?

A J Quinton
Cranbourne
Australia

What's all the fuss about?

The EU is about to remove one of the last Member State trading barriers; the requirement for radio and telecommunications apparatus to be independently tested and certified.

So where is the problem in this? After all, this is only the same as we currently have for EMC, LVD etc. and these work fine.

The pundits claim that the market will be flooded with substandard goods. Yes, they might, but what about the following?

Have you ever tried to sell a substandard product to a customer? Save yourself the effort. Consumers are not idiots, and they will see through your sales spiel instantly. From personal experience, you will be left with shelves full of a product, that fails to move, until you drop the price to some ridiculous level.

Even when you are lucky (unlucky) enough to make a sale, the resulting service returns will wipe out any profits you make on that, and the rest of your products.

Reputable manufacturers and distributors will still test their products to the relevant ETSI regulations before making their declaration of conformity. Whether they have the products independently tested, or continue to send them to a test lab, the cost is approximately the same, so there are no great savings there.

Ok, so a few substandard products may get onto the market. As with the LVD, EMC directives, the policing procedure will be complaint driven, and it really does work.

Do not waste your time making unsubstantiated complaints to your local TSO; they are unlikely to take any action. If you believe a product is non-compliant, obtain a sample, carry out a complete test, and furnish Trading Standards with irrefutable proof of non-compliance. The speed of the results will astound you.

Kevin Aston
Via e-mail

Crossover error

The circuit on page 572 of the July issue is very interesting.

The author has, however, failed to verify the calculations. To achieve 0dB gain and the frequency responses shown, conductances are required in the calculations, so that the sum of these becomes 10μmho.

Thus for the average filter (a) instead of five 20kΩ resistors being required five 2μmho conductances are required – i.e., five 500kΩ resistors.

Thus, for the optimised average filter (c), the resistance values become – R_1 (and R_5) 1.37MΩ (0.732μmho), $R_{2,4}$ 401.6kΩ (2.49μmho), R_3 280kΩ (5.37μmho) – with similar recalculations being required for the other cases. (Response has been verified for the optimised average filter with those values shown using the Tina simulator).

It is also not strictly true to call

this a finite-impulse-response filter as the individual delay sections actually have an infinite impulse response. They are also only approximately phase-linear – and then only within a narrow range of frequencies.

It is, however, one of the more interesting circuits that you have published recently and adequate for the purpose intended.

When audio amplification becomes digital, then true FIRs will be possible, approximating ideal loudspeaker crossover networks!

Trevor Blogg
Via e-mail

Summing up crossovers

Ian Taylor wrote in August to claim that loudspeaker crossover networks have been done wrong all along, and that they should cross over with a gain of –3dB, or $1/\sqrt{2}$ in amplitude, because we are summing the power from the two loudspeakers which is proportional to the square of the drive voltage, hence the sum is 2 times $(1/\sqrt{2})^2=1$.

Life would indeed be strange if we perceived sound levels from loudspeakers as the square of the drive voltages. But perhaps this is what interests people in psychology departments?

Imagine what would happen if two sine waves were presented to the loudspeaker: call them $\sin(\omega_1 t)$ and $\sin(\omega_2 t)$. For simplicity, they are both of unity amplitude. If the ear responded to the square of these signals, we would hear

$$[\sin(\omega_1 t)+\sin(\omega_2 t)]^2$$

which is,

$$\sin^2(\omega_1 t)+\sin^2(\omega_2 t)+2\sin(\omega_1 t)\sin(\omega_2 t).$$

The $2\sin(\omega_1 t)\sin(\omega_2 t)$ term is interesting because my maths textbook says that this can be rewritten as

$$\cos(\omega_1 t-\omega_2 t)-\cos(\omega_1 t+\omega_2 t).$$

In other words, sum and difference frequencies. These appear, along with highly distorted $\sin^2$ signals of similar amplitude.

The fact that in practice we hear mostly $\sin(\omega_1 t)$ and $\sin(\omega_2 t)$ – with a small amount of distortion caused by non-linear effects – shows that we definitely hear a sound proportional to the signal levels going in from the loudspeakers.

This being the case, then when two speakers are sharing the signals at the crossover point, the drives should each be a half voltage, or –6dB. Or should they? In practice, it depends on the phase relationship between the two signals.

Quite commonly, the two signals – one high-pass filtered, and one low-pass filtered – are not in phase. If they are 90° out of phase at the crossover frequency – which happens, for example, with third-order Butterworth networks – then to get unity amplitude as a sum, the two filtered signals must each be $1/\sqrt{2}$, or –3dB. They are orthogonal, and so sum as the square root of the sum of the squares.

Inductance of a single wire: Incidentally, I have what claims to be the full formula, derived from theory, for the inductance of a single piece of wire – i.e. where the return path is at a great distance. It comes from Inductance Calculations, Frederick Grover, Van Nostrand 1946. For circular cross-section wires, it is,

$$L(\text{nH})=0.2l[\log_e(2l/r)-1+\mu/4]$$

where r is the radius in mm, l is the length in mm, and the wire has a relative permeability of μ. This does give a value close to 1μH per metre or 1nH per millimetre, for wire sizes of around 1mm.

Brian Pollard
Cranbrook
Kent

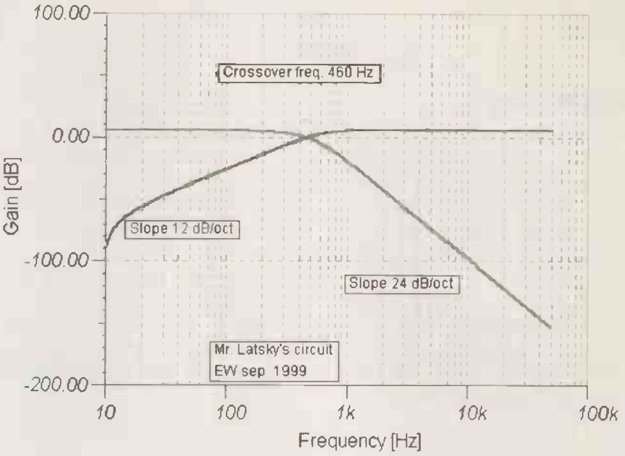
Phase-linear crossover

Mr Latsky's circuit on page 779 of the September issue suffers from a serious drawback: the slope of the low-pass output is 24 dB/octave, as should be the case for a fourth-order network, but the high-pass output rolls off at only 12dB/octave. This is shown in the graph above, calculated by the circuit simulator Tina Plus, using ideal op-amps.

This behaviour is common to all filter circuits where the second output is formed by subtracting the first from the input and it can be mathematically explained by developing the formula for the transfer function for the high-pass part from that for the low-pass:

$$F(s)=1-\frac{1}{s^4+a.s^3+b.s^2+c.s+1}$$

$$F(s)=\frac{s^4+a.s^3+b.s^2+c.s}{s^4+a.s^3+b.s^2+c.s+1}$$



Using the appropriate coefficients for the filter type shows zeros (in the numerator) nearly coinciding with poles (in the denominator) and therefore very nearly cancelling each other. This reduces the response from fourth to second order.

The consequence is that in practice the high-frequency speaker must be able to handle frequencies down to at least two octaves below the crossover frequency, because bass signals can contain large transients, especially when bass boost is applied.

The same goes for Mr Van Dormael's third-order circuit in the August issue, where his graph shows that the high-pass output, after a peak of 4dB (and not 2!) at crossover, rolls off at 6dB/octave, as a first-order circuit.

Exactly the same scheme has, indeed, been described before under the name 'asymmetrical crossover', in National Semiconductor's 'Audio Handbook', first published in 1976.

It has never become popular because of the abovementioned drawback.

S. de Boer
Veghel
Netherlands

Receiver radiation

In his worthwhile article *RF Mixers* in the February 1999 issue Joe Carr asks anyone with first hand knowledge of detecting receiver radiation to contact him.

If he has not yet read *Spy Catcher* by Peter Wright, he may find the reference to RAFTER interesting.

I found out about receiver antenna leakage when trying to get 20 UHF receivers fed from one antenna to work properly. I hope Mr Carr can offer help on this topic in a later article.

P Robinson
Tadley ■

Corrections

0-10 led display for digital input

On page 648 of the August issue, pin 7 of the leftmost 555 should have been connected to the junction of R_2 and the 1kΩ resistor, just to the left of pin 7.

Josef Holecek
Prague

Phase-linear crossover

In the September issue, page 779, resistor R_{45} should be 464kΩ and not 46k4 as marked. The incorrect resistor would give a stage gain of 1.84 which is not a fourth-order Bessel filter. If for some reason someone does not want to use such a high value for R_{45} they can leave it at 46.4kΩ, but then they have to scale R_{46} down to 3.9kΩ. The chart's second 1kHz should be 10kHz of course.

Peter Latsky

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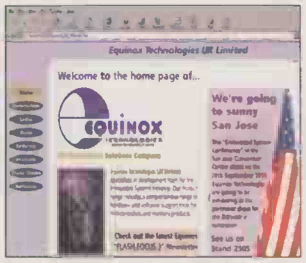

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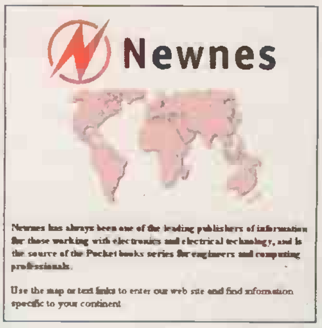

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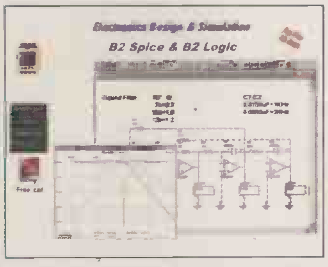

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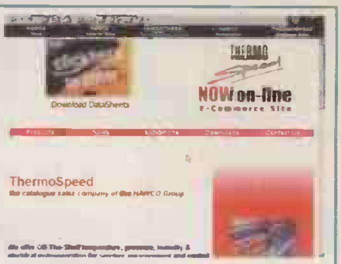

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Working voltage	600V DC or pk-pk AC
Switch position 'Ref'	
Probe tip grounded via 9MΩ, scope i/p grounded	

Richard Brice's de-jittering circuit for digital audio can be inserted between a CD player and external d-to-a converter for improved fidelity, particularly at low frequencies. It also removes copy code and can be modified to convert between formats.

Improve your digital audio

Jitter on a digital audio signal is known to cause appreciable signal degradation. All the more irksome then, that its elimination is extremely difficult by means of classical PLL-style digital audio interface receivers. This is especially so when the modulation is at a relatively low frequency, such as that caused by power-supply induced coupling.

This article describes a practical circuit for a digital interface unit that may be used to remove low-frequency jitter from a digital audio signal. Its use between a CD output and an external converter is described.

The unit has a number of useful ancillary provisions that allow it to be modified to transcode between the SPDIF consumer interface and the various AES/EBU interfaces. It also strips copy-code, allowing direct digital copies to be made.

Table 1. Aperture effect – even when the sampling pulse width equals the sampling period, loss at the pass-band edge is only –3.9dB.

T_s/T_o	Attenuation at pass-band edge
1	3.9dB
0.5	0.9dB
0.25	0.2dB
0.2	0.1dB
0.1	0.04dB

Background

The quality of digital audio is mathematically definable in terms of; the sampling frequency employed, the bit 'depth', the sampling-pulse aperture and time uncertainty. Expressions for the first two are well known. The effect of the latter two parameters is less well appreciated.

Sampling pulse width (as a proportion of sampling period) simply has an effect on frequency response as defined in the expression,

$$20\log sinc\left(\frac{\pi}{2}\times\frac{f}{f_n}\times\frac{T_s}{T_o}\right)\text{dB}$$

where, T_s is the duration of the sampling pulse (aperture) and f_n is the Nyquist frequency limit. Note that *sinc* is shorthand for sin*x*/*x*. This is termed 'aperture effect' and is actually relatively benign.

As **Table 1** indicates, even when the sampling pulse width is equal to the sampling period, the loss, at the band edge, is only –3.9dB. Provided T_s<$0.2T_o$, the effect is pretty negligible.

In any case, frequency response 'droop' can always be made up in the design of the reconstruction filter following the d-to-a converter – where it is often referred to as sin*x/x* correction.

Why is jitter a problem?

The effect of sampling-pulse time uncertainty or 'jitter' is much more destructive. Because all signals change their amplitude with respect to time, the effect of a slightly misplaced sampling point has the effect of superimposing a distortion on the original waveform, effectively reducing available dynamic range.

This next equation defines the limit of sampling uncertainty, d*T*, for a digital system of *n* bits,

$$\frac{dT}{T_o} = \frac{1}{(\pi \times 2^{n-1})}$$

Working through an example, a sixteen-bit audio system with 48kbit/s sampling must have a jitter performance of less than 200ps in order to preserve the theoretical dynamic range available from the 16-bit system. In other words the jitter must be just 0.001% of the sampling period!

Even if this requirement has been met in the recording stage, for absolute fidelity to be preserved, this value must be 'recreated' in any subsequent conversion to analogue for playback.

Phase-locked loop receivers

Most digital-audio converters rely on a phase-locked loop front-end to extract clock from the self-clocking AES/EBU or SPDIF digital-audio interface and to use this in the reconstruction of the analogue signal.

Several very good chips exist for this purpose, one of the most famous being the *CS8412* from Crystal Semiconductor. Should there be any high-frequency jitter on the interface, the PLL type receiver does a very good job in rejecting it. But, at low frequencies, it has no effect whatsoever, as **Fig. 1** shows.

This is unfortunate for the audiophile because jitter very often exists at much lower frequencies, usually due to the interaction of other analogue or digital signals or to power-supply induced effects.

Experiments have shown that the effect of substantially monotonic jitter indicates that the limits defined in the second equation still apply – even on modern over-sampling a-to-d and d-to-a converters.

Asynchronous sample-rate conversion

The construction of high-frequency phase-locked loops with low-frequency rejection is no mean task. Effectively the circuit must behave like a resonant circuit with a Q of thousands; a design constraint that usually compromises lock-time and centre frequency variability without recourse to complicated multi-stage designs.

Fortunately there exists an alternative, in the form of a family of chips from Analog Devices. These are based on asynchronous sample-rate conversion, or ASRC, technology.

There is more than one way to describe the nature of asynchronous sample rate conversion. The easiest to understand is the interpolation-decimation model in which the input signal is over-sampled to a much higher rate, digitally low-pass filtered and re-sampled at the output sample frequency.

Unfortunately, while easy to understand, the interpolation-decimation model is not a suitable basis for a practical system. This is because the output of such a system is only the nearest appropriate value in a temporal sense.

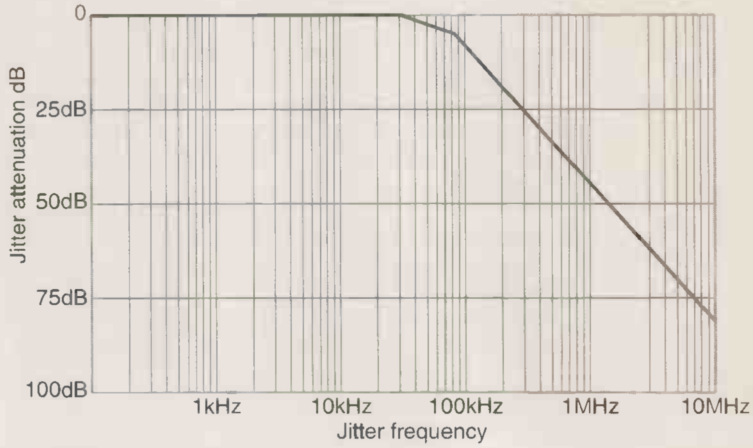


Fig. 1. Phase-locked loop receivers like the CS8412 do a very good job of removing jitter at high frequencies, but they do nothing for low frequency jitter.

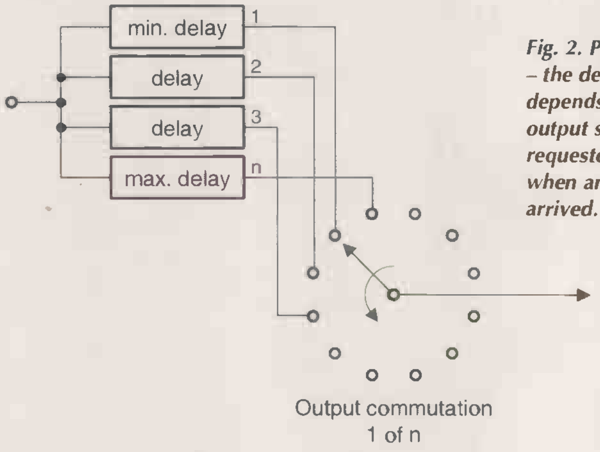


Fig. 2. Polyphase filtering – the delay chosen depends on when an output sample is requested in relation to when an input sample arrived.

There is no theoretical reason why the interpolation shouldn't be carried out at a fast enough rate to make this viable. But there exist some very good practical reasons why it is not.

For instance, in order to achieve a reasonable performance – and this means, to achieve 16-bit levels of THD+N across the 0 to 20kHz audio band – the interpolation up-sample frequency would need to be over 3GHz! Clearly, this is an impracticable rate for a low-power IC, so the Analog Devices chips use a less commonly known method of sample-rate conversion called polyphase filtering.

Polyphase filtering

In the polyphase filter ASRC, the digital audio sample sequence is over-sampled – but at a manageable rate of megahertz. It is then applied to a digital FIR low-pass filter in which the required impulse response – 20kHz cut-off – is

Thinking of prototyping it?

For those interested in building the circuit, PCBs, are available. For more details contact Richard Brice via e-mail richard@perfect-pitch.demon.co.uk.

itself highly over-sampled.

The filter is 'over-sampled' in the sense that it comprises many times the required number of coefficient sample taps to satisfy the Nyquist criterion. This means that, at any given moment, only a sparsely sampled subset of coefficients of this filter need be chosen to process the input samples.

These subsets of coefficients, create a kind of 'sub-filter', each possessing an identical 0 to 20kHz magnitude response but with a fractionally different group delay – hence the term 'polyphase'.

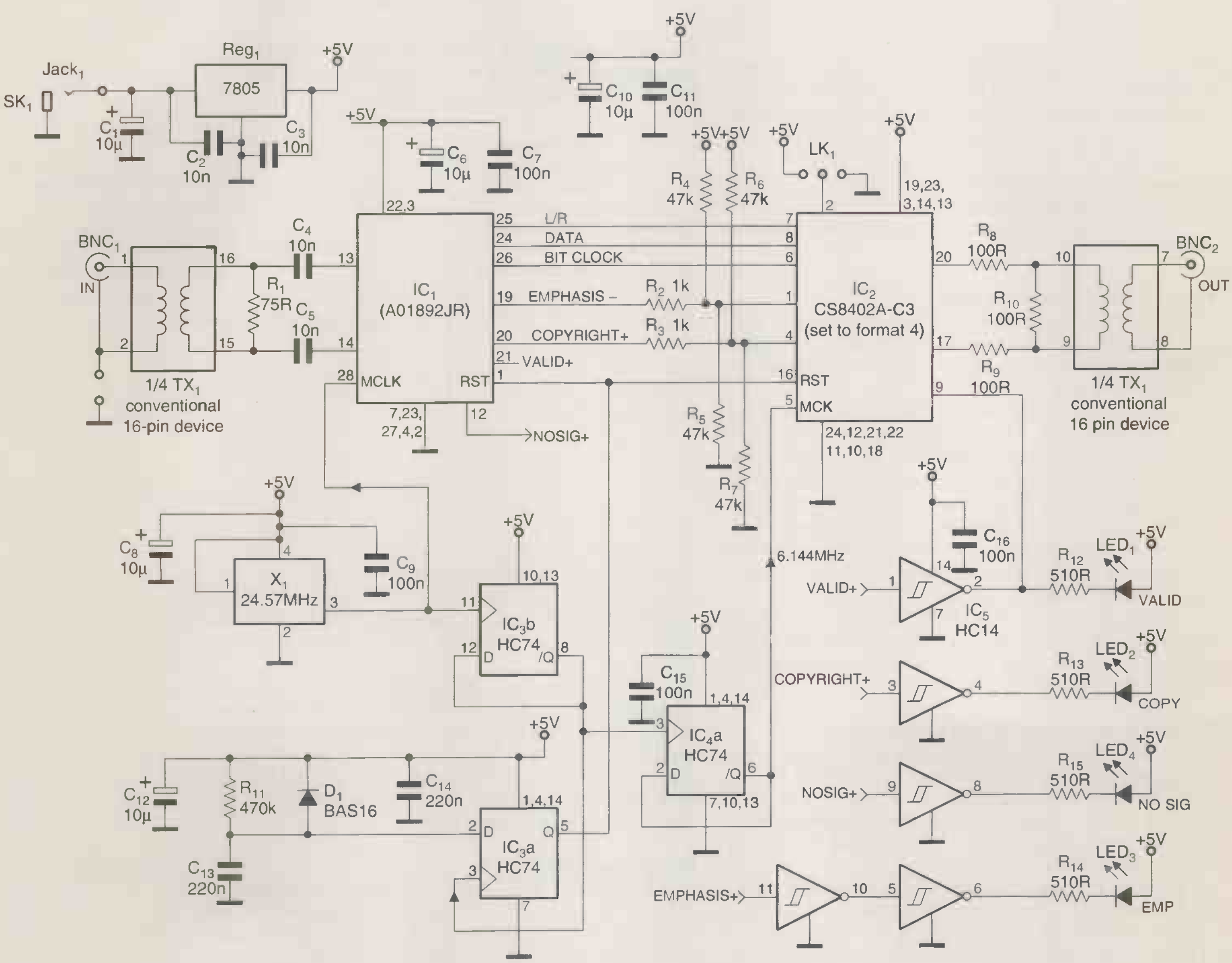
It is as if the input signal was being applied to a very great number – i.e. thousands – of digital delay-lines, each with a slightly differing delay. This is shown greatly simplified in **Fig. 2**.

The sample-rate conversion process works like this; if a request for an output sample occurs immediately after an input sample has arrived, a polyphase filter is chosen that imposes a short group delay. If a request for an output sample occurs late in the input-sample period, a polyphase filter that imposes a long group delay is chosen. In this fashion, the amplitude of the output sample is precisely computed at the desired output sample frequency.

Looking at the output commutator in Fig. 2, it's possible to imagine that, provided the relationship between the input and output frequencies is not completely random, there will be a pattern to the switch selection when looked at over a certain period of time.

Indeed, provided the input and output frequency are relatively stable, you can imagine the commutator revolving at the computed difference frequency between the input and output sample frequency.

This process is controlled, within the Analog Devices parts,



Note:
1 in PRO mode, EMP not indicated, flag **ALWAYS** set
2 in PRO mode output **ALWAYS** flagged stereophonic
3 consumer flags - 48k; COPY and EMP are transparent, so is validity

Fig. 3. Complete jitter remover for digital audio signals. It can be modified to transcode between SPDIF and various AES/EBU interfaces and it strips copy-code – allowing direct digital copies to be made.

by an on-chip, digital servo-control system that bases its commutation decisions not an instantaneous measurement, but rather on a digitally filtered ratio.

It is the effect of this powerful, low-pass filtering mechanism that greatly reduces any jitter that may be present on the sample clocks – even when the jitter frequency is just a few tens of hertz.

Implementing the design

Figure 3 is a practical implementation of the *AD1892* used as a jitter rejection device for use between the output of a CD player and the input of an outboard d-to-a converter. The *AD1892* is not just an ASRC. It is also an AES/SPDIF interface receiver too, so the circuit implementation is very simple.

The *1892* has some limitations, the most severe of which is that it only retains its performance over a limited range of upward sample-rate conversion and a very limited range of downward rate conversion.

For this design, I decided to use an up-conversion from 44.1kHz to 48kHz. The part works well at these two rates and the master oscillator – which must be 512 times output sample rate; 24.576MHz – is relatively easy to source at this frequency.

The SPDIF signal arrives at TX_1 – one part of a 16-pin, four transformer data-bus isolator – and is terminated, on the far side of the transformer by R_1. The signal is applied directly to the *1892* via coupling capacitors, C_4 and C_5.

The master output clock is derived from a small 24MHz crystal oscillator. Having been broken down into separate clocks and data by the Analog Devices part, the composite AES/SPDIF signal is put back together again by the Crystal Semiconductor *CS8402* transmitter chip. This too requires a master clock, but at one quarter of the frequency of the *AD1892*, hence the inclusion of the divide-by-two bistables IC_3 and IC_4.

SPDIF output is via transformer TX_1, which is another part of the same data-bus isolator used for the input. Note resistors $R_{8,9,10}$: these produce an output impedance of 75Ω at a level of about 2V. This is above that specified for SPDIF, which is 1V and is therefore a bit non-standard.

I made the choice for two reasons. Firstly, I have found that outboard d-to-a converters like to have a bit more level. Secondly, by changing the position of LK_1, the circuit may be used to encode a digital signal to the unbalanced form of the professional AES/EBU digital interface. This requires the higher output level.

Such provisions make the circuit useful if you need to interface a non-professional CD player in a digital studio. The output is also quite suitable for driving symmetrical a 110Ω AES-style interface, *mutatis mutandis*.

User indications

The circuit includes several user LEDs to indicate; validity, copy-code, pre-emphasis and signal loss. These are derived and decoded by the *AD1892*. The LEDs are driven by an HC14 and are primarily there for amusement since no user intervention is required.

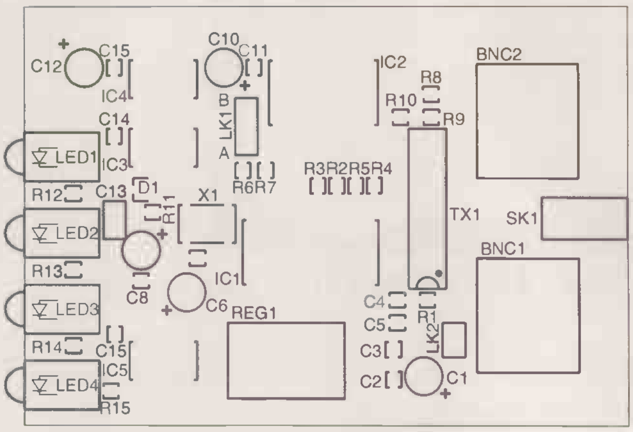
Emphasis state and Copyright prohibit are decoded and re-coded by the *CS8402*. Pull-up, pull-down resistor positions are provided here to allow for various options. The most useful of these is the removal of R_3 and R_6, which strips copy-code and allows direct digital copies to be made.

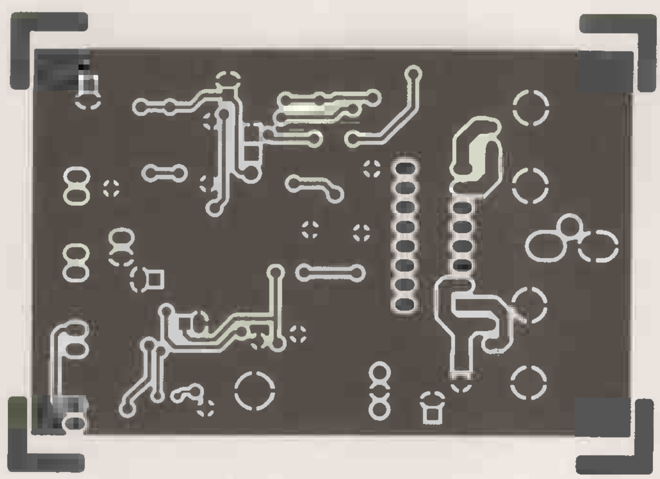
The layout of my prototype is shown in **Fig. 4**. Note the extensive use of ground-plane. Note that the signal inputs and outputs are on BNC as I prefer this connector enormously to the RCA phono alternative.

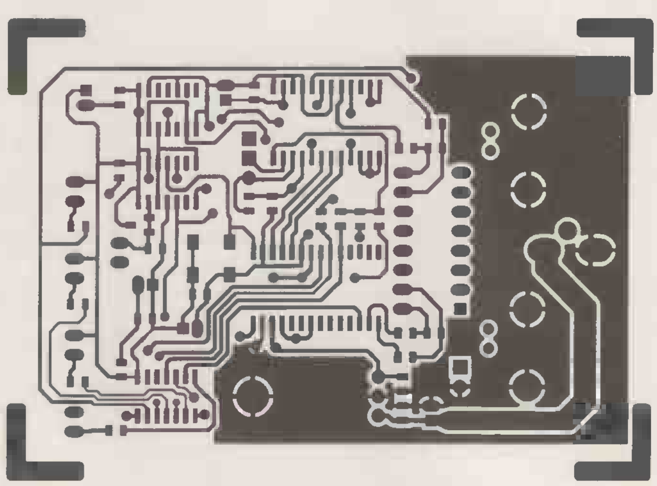
The power supply input is squeezed between the input and output and the whole circuit is enclosed in a little anodised, aluminium extrusion box, no bigger than a household box of matches. It is ideally suited for sitting on top of a CD player or d-to-a converter.

Although it's unwise to be adamant in this area, everyone who has listened to the circuit has been amazed by the improvement in quality that it yields; especially in definition at the bass-end of the spectrum. ■

Fig. 4. Layout of the author's prototype PCB.







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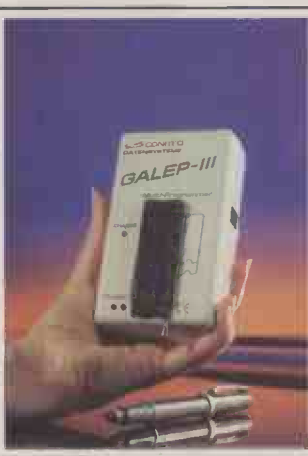

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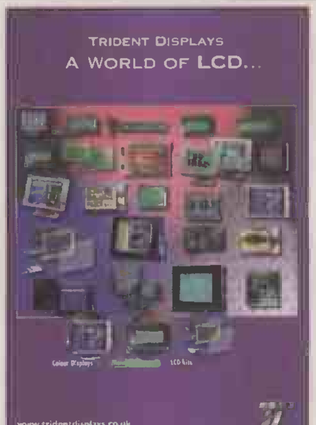

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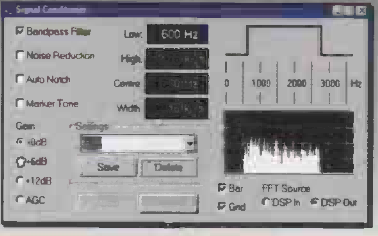
The DSP applet provided with the WR3100i spectrum monitor ISA card (£995+VAT) allows continuous control of audio bandwidth and other signal conditioning functions.

ONLY £81.07 inc vat
(requires SoundBlaster 16 compatible sound card)



WiNADiO™ PC RECEIVERS

Available as either an internal ISA card that slips inside your PC, or as an external (portable) unit. WiNRADiO combines the power of your PC with the very latest, and greatest, synthesised receivers.

YOU CAN USE WiNRADiO™ SCANNiNG PC COMMUNiCATION RECEiVERS FOR:
Broadcast, media monitoring, professional & amateur radio communications, scanning, spot frequency, whole spectrum monitoring, instrumentation surveillance and recording.

If you're after the ultimate receiver-in-a-PC with full DSP then smile and say, "Hello" to the new **WR31000i-DSP** with its hardware for real-time recording, signal conditioning and decoding applications. It's all you need.

NEW EXTERNAL MODELS

EXTERNAL WiNRADiO™
We are now able to offer you a complete range of stand-alone WiNRADiO comms systems:

- **WR1000e - £359 INC VAT**
- **WR1500e - £429 INC VAT**
- **WR3100e - £1169 INC VAT**

Each stand-alone unit connects to your PC through either the basic RS232, or through an optional PCMCIA adapter (for high speed control).
The units are powered through either your existing 12v supply, or through an (entirely optional) NiMH rechargeable 12v battery pack.

"It's software is excellent.. more versatile and less idiosyncratic than the Icom IC-PCR1000"
WRTH 1999 Review

"Five stars for its mechanical design"
WRTH 1999 Review

"Most Innovative Receiver"
WRTH 1998 Awards

Model Name/Number	WR-1000	WR-1500	WR-3100
Construction of internals	WR-1000i/WR-1500i-3100iDSP- Internal full length ISA cards		
Construction of externals	WR-1000e/WR-1500e - 3100e - external RS232/PCMCIA (optional)		
Frequency range	0.5-1300 MHz	0.15-1500 MHz	0.15-1500 MHz
Modes	AM,SSB/CW,FM-N,FM-W	AM,LSB,USB,CW,FM-N,FM-W	AM,LSB,USB,CW,FM-N,FM-W
Tuning step size	100 Hz (5 Hz BFO)	100 Hz (1 Hz for SSB and CW)	100 Hz (1 Hz for SSB and CW)
IF bandwidths	6 kHz (AM/SSB), 17 kHz (FM-N), 230 kHz (W)	2.5 kHz(SSB/CW), 9 kHz (AM) 17 kHz (FM-N), 230 kHz (W)	2.5 kHz(SSB/CW), 9 kHz (AM) 17 kHz (FM-N), 230 kHz (W)
Receiver type	PLL-based triple-conv. superhet		
Scanning speed	10 ch/sec (AM), 50 ch/sec (FM)		
Audio output on card	200mW	200mW	200mW
Max on one motherboard	8 cards	8 cards	3-8 cards (pse ask)
Dynamic range	65 dB	65 dB	85dB
IF shift (passband tuning)	no	±2 kHz	±2 kHz
DSP in hardware	no - use optional DS software		YES (ISA card ONLY)
IRQ required	no	no	yes (for ISA card)
Spectrum Scope	yes	yes	yes
Visitune	yes	yes	yes
Published software API	yes	yes	yes (also DSP)
Internal ISA cards	£299 inc vat	£369 inc vat	£1169.13 inc
External units	£359 inc vat	£429 inc vat	£1169.13 inc (hardware DSP only internal)

PCMCIA Adapter (external): £69.00 inc when bought with 'e' series unit (otherwise: £99 inc)
PPS NiMH 12v Battery Pack and Charger: £99 inc when purchased with 'e' series unit (otherwise: £139 inc)
The WiNRADiO Digital Suite: £74.99 inc when purchased with a WiNRADiO receiver (otherwise: £81.05 inc)

To receive your completely free (no obligation) info pack and WINRADIO software emulation demo disk all you have to do is get on the internet and go to our website at http://www.broadercasting.com. If you don't yet have easy access to the internet then by all means feel free to telephone us or send a fax.

Please send all your enquiries to: info@broadercasting.com or Telephone: 0800 0746 263 or +44 (0)1245 348000 - Fax: +44 (0)1245 287057
Enformatica Limited, Unit B, Chelford Court, Robjohns Road, Chelmsford, Essex, CM1 3AG, United Kingdom

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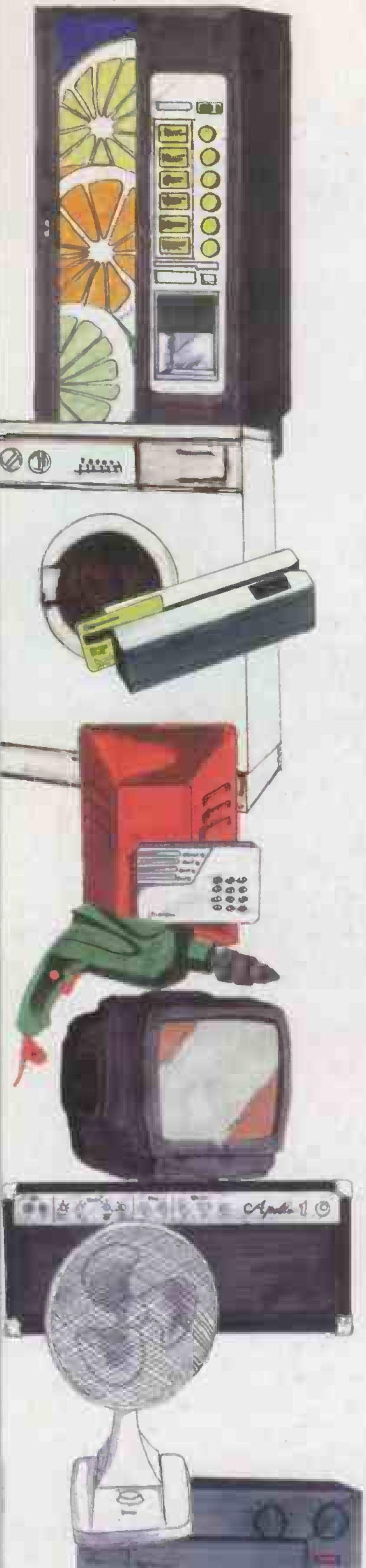
CIRCLE NO. 102 ON REPLY CARD



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