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Joules Watt on wave motion

Two approaches to risc processor design

Sequency-division multiplexing

Pseudo-science in sound

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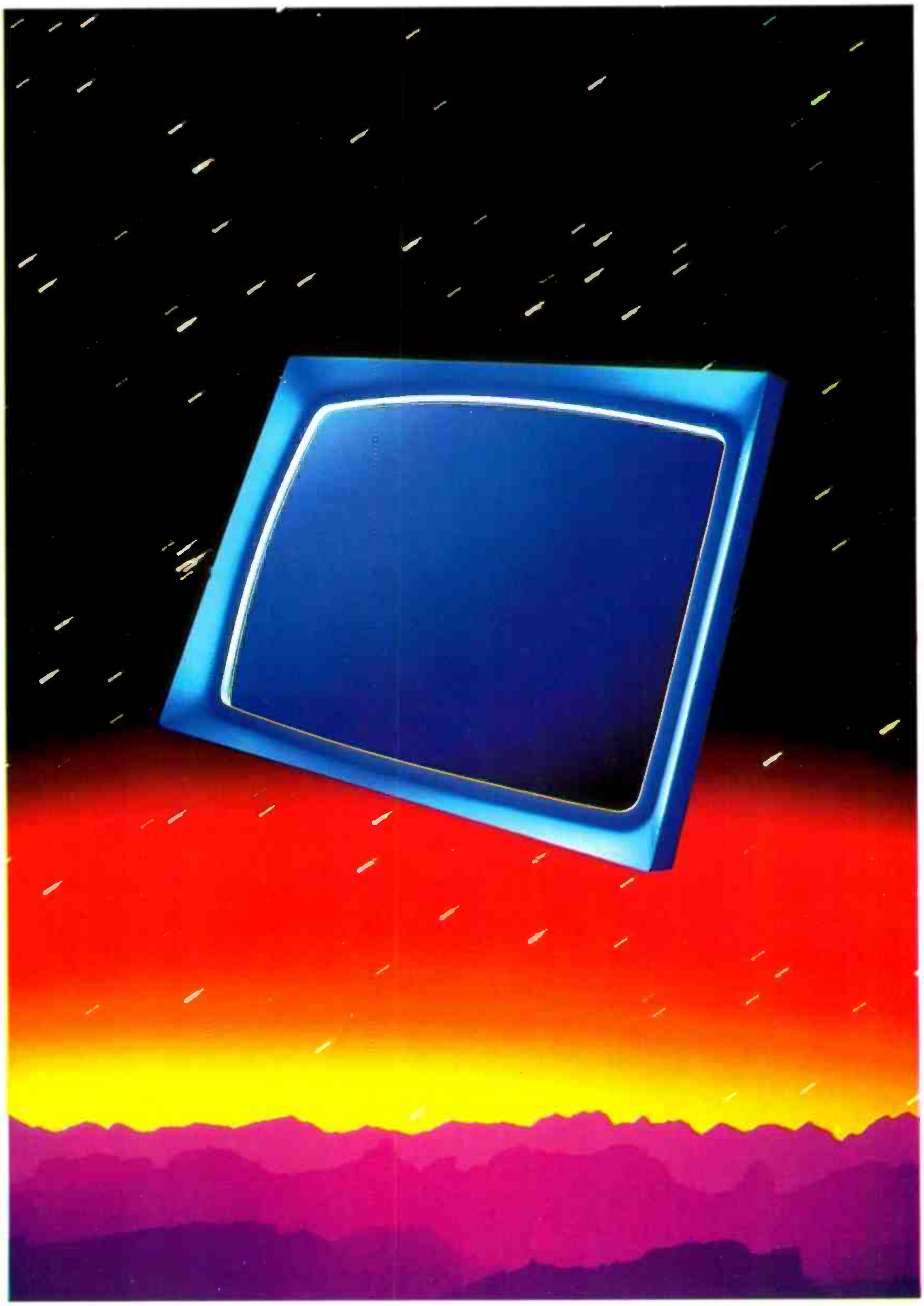

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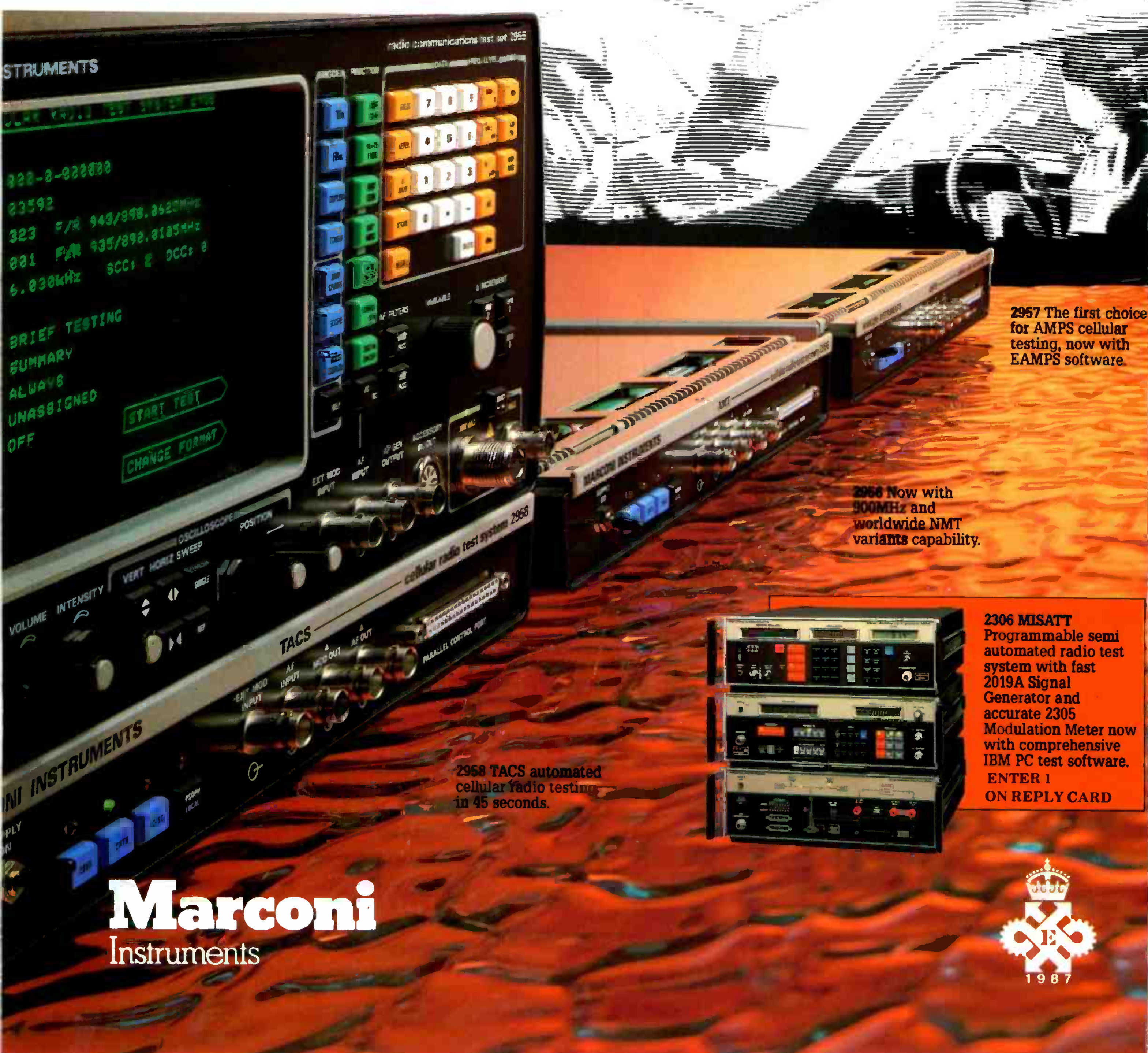

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A broader view

In a report published by the Engineering Council, "A comparison of the statistics of engineering education: Japan and the United Kingdom", some remarkable figures are presented to account for the success of the Japanese in the years since the war.

For example, 150 000 Japanese students are studying technology until they are eighteen and over, while only around 35 000 do so in the UK, including those who leave school at 16+ to take National Certificates. Ninety per cent of the Japanese population take mathematics to 18+, compared with 18% here. In fairness, it must be pointed out that the report was commissioned to challenge earlier Government figures published in the December *Employment Gazette*.

While all the statistics are indeed thought-provoking – and a little depressing – it is the very last sentence in the whole report that seems to carry at least as much significance as any of them. "...their system ensures that... the important decisions on both technical and managerial matters at all levels up to President of the major industrial undertakings are generally made by engineers".

There is a clear difference here between the common practice in the UK, where engineers advise managers who then make the decisions, and the state of affairs in Japan, where the decisions are evidently made directly by engineers.

The report also makes mention of the kind of education provided in university engineering schools, which "includes languages and humanities/social sciences, mathematics, physics and chemistry as well as engineering".

In an editorial on this topic in the May issue, the phrase "mid-career diversion to financial and business affairs" was used. It is quite clear from this EC report that the Japanese view the education of engineers in a much more liberal light, so that the managers in an engineering company do not need to change direction: they have been educated in the relevant disciplines from the start, while still having their engineering background and therefore being able to draw from a more complete knowledge base when making the decisions. It means, too, that engineers lower down the scale should possess a more enlightened, broader view of their own efforts and the consequences thereof.

British industrial history over the last forty years at least is littered with abandoned projects, loss-making developments and a failure to capitalize on ideas. Not all of these disasters can be laid at the door of "captains of industry", but one is left with the feeling that, if their decisions on whether to go ahead with certain developments had been informed by a better education in all the subjects which possess relevance to an engineering question, we might not be looking to the Far East for inspiration on how to compete in an increasingly technological world.

EDITOR
Philip Darrington

EDITOR – INDUSTRY INSIGHT
Geoffrey Shorter, B.Sc.
01-661 8639

DEPUTY EDITOR
Martin Eccles
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COMMUNICATIONS EDITOR
Richard Lambley
01-661 3039

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CLASSIFIED SALES EXECUTIVE
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01-661 3033

ADVERTISING PRODUCTION
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01-661 8648

Clare Hampton
01-661 8649

MARKETING EXECUTIVE
Rob Ferguson
01-661 8679

PUBLISHER
Shobhan Gajjar
01-661 8452

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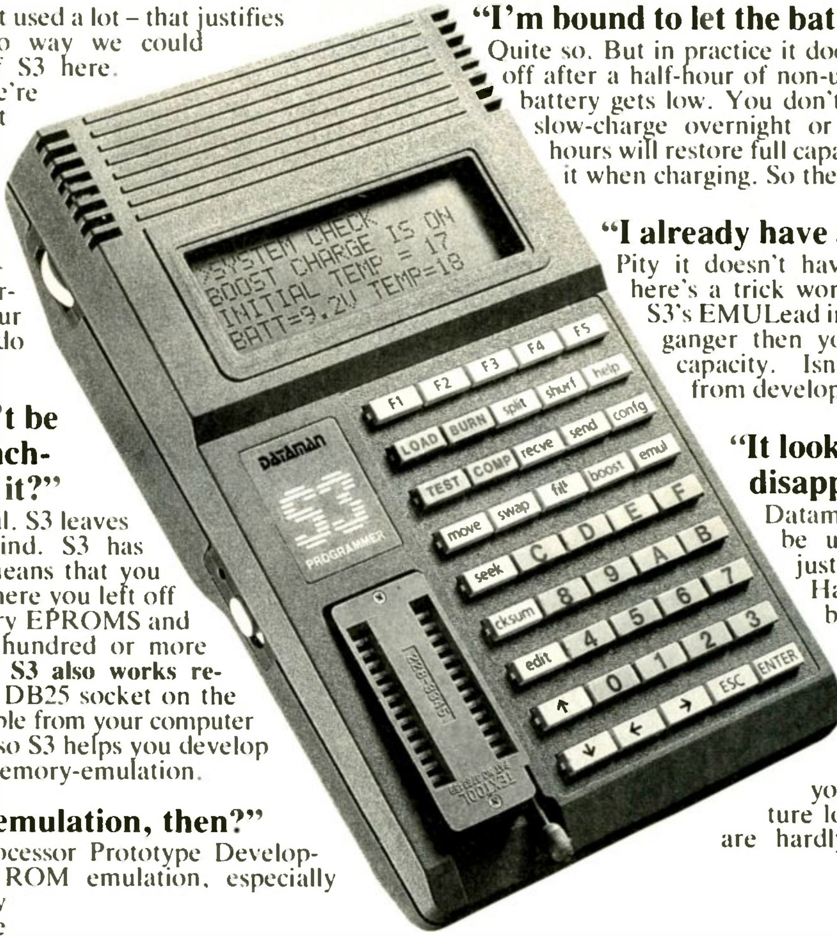
"It's a pretty small battery-powered PROM programmer – so what?"

Tools which are convenient get used a lot – that justifies their existence. There is no way we could explain all the usefulness of S3 here. Instead, if you're interested we're going to let you see it, use it and evaluate it in your own workshop. We went to a lot of trouble to design S3 just the way it is – no other PROMMER is all CMOS and all SMT. So we must be convinced that S3 would be a formidable addition to your armoury. Now all we have to do is to convince you.



"Such a little thing can't be powerful, like a big bench-programmer – er – can it?"

Yes, it can. It is more powerful. S3 leaves other prommers streets behind. S3 has continuous memory, which means that you can pick it up and carry-on where you left off last week. S3 has a huge library EPROMS and EEPROMS. S3 can blow a hundred or more PROMS without recharging. **S3 also works remotely, via RS232.** There's a DB25 socket on the back. All commands are available from your computer (through a modem, even). Also S3 helps you develop and debug microsystems by memory-emulation.

"What's this memory-emulation, then?"

It's a technique for Microprocessor Prototype Development, more powerful than ROM emulation, especially useful for single-chip "piggy back" micros. You plug the lead with the 24/28 pin header in place of the ROM/RAM. You clip the Flying-Write-Lead to the microprocessor and you're in business. The code is entered using either the keyboard or the serial interface. Computer-assembled files are downloaded in standard format – ASCII, BINARY, INTELHEX, MOTOROLA, TEKHEX.

Your microprocessor can WRITE to S3 as well as READ. You can edit your variables and stack as well as your program, if you keep them all in S3.

S3 can look like any PROM up to 64K bytes, 25 or 27 series. Access is 100ns – that's really fast. Memory-emulation is cheap, it's universal and the prototype works "like the real thing".

"I'm bound to let the battery go flat."

Quite so. But in practice it doesn't matter. S3 switches off after a half-hour of non-use anyway, or when the battery gets low. You don't lose your data. Then a slow-charge overnight or boost-charge for three hours will restore full capacity. You can keep using it when charging. So there really is no problem.

"I already have a programmer."

Pity it doesn't have S3 features, eh? But here's a trick worth knowing. If you plug S3's EMULead into the master socket of a ganger then you get an S3 with gang capacity. Isn't production separate from development anyway?

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S3 loads its working programs out of a PROM in its socket, like a computer loads from disk. Software expansion is unlimited. Upgrades will come in a PROM. Programs can be exchanged between users. How's that for upgradability?

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THE MOTOROLA APPROACH

MC88100 risc

Design goals for the 88100 were simple – a microprocessor capable of 17Mips that could be manufactured within the limits of existing design and wafer-fabrication technology.

DAVID JONES

Two approaches to risc

Two of the most recent reduced-instruction-set computers are described in this article and the one on page 689 – the MC88100 from Motorola and the MA29000 from AMD. Both can run at 17Mips.

Over the past three years, Motorola has been researching advanced micro-processor architectures, in particular those used for reduced-instruction-set computers. The first product resulting from that research, the MC88100, was announced on April 18. It has a fully 32bit internal architecture, a 17Mips execution speed and 51 instructions, a group of which covers IEEE specified floating-point operations.

Although the 88100 is dubbed a reduced-instruction-set computer, it contains many architectural extensions for, say, multiprocessor, fault-tolerant and graphics applications. Perhaps the most important extension is the use of multiple execution units, each with its own pipeline.

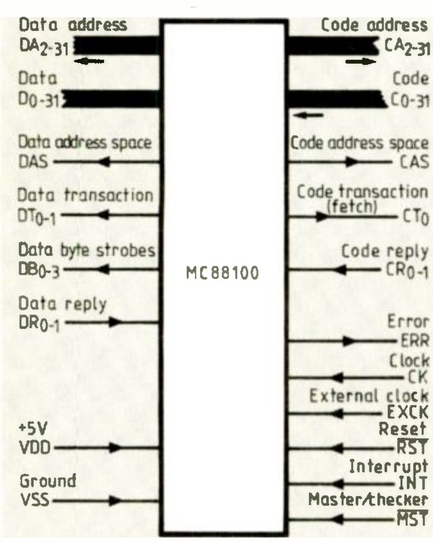
The idea and use of pipelining is not new; microprocessors with between three and eight levels of pipelining are already available. Unlike the 88100 though, existing processors generally have only one pipeline for instruction-address calculation, fetching and decoding. The penalties for change of flow instruction are obvious. Some form of invalidation or clearing of the pipeline needs to be performed, causing a processor delay. Rather than having just one pipeline, the 88100 has four, which eases the problem of change of flow instructions and increases overall parallelism.

Having the functions of the processor split between a number of execution units allows a high degree of parallelism. The 88100 is capable of executing up to five floating-point instructions and six floating-point or integer-multiply instructions at the same time as up to three data memory accesses and three instruction fetches.

Pipeline problems are reduced by allowing the processor to perform delayed branching

One set of buses for code and one for data allows very fast simultaneous accesses of instructions and data, below left.

RISC OUTLINE



To fully explain the philosophy behind the term reduced-instruction-set computer would take at least a long article. This section simply outlines risc philosophy, and how it is implemented in the 88100.

Rather than having instructions with complex addressing modes designed into on-chip microcode, a simpler approach is adopted. This involves defining a small number of optimum instructions and designing them directly into the logic on the chip. When complex instructions and addressing modes are needed, they are simply built up from the simpler instructions.

Instructions of a risc processor, besides executing faster than those of a complex-instruction-set computer because of the removal of the decay through several layers of microcode decoding logic, take just one clock cycle to execute.

Perhaps the third most important aspect of the risc philosophy is that the processor's architecture should be designed for high-level languages. In other words, a large number of general-purpose registers should be available to aid compiler design and optimization.

If the 88100 is to execute code in only one clock cycle then considerable attention needs to be given to the process of fetching data and instructions. In fact this consideration gave rise to the first architecture extension implemented during the 88100's design.

Bus and control signals of the 88100 are shown in the diagram. You can see that the Harvard-style architecture which started to appear internally in 32-bit processors like the 68030 and NS32532 now extends outside the chip. Instructions and data can be fetched simultaneously in one clock cycle, and instructions can be fetched while data is being written.

Adopting Harvard-style rather than Von Neumann architecture increases throughput on the external buses, but it creates a problem for the system designer. Accessing data and instructions in only one clock cycle is fine, but it makes external decoding logic and memory difficult to design and expensive. For a processor running at 20MHz, only 50ns is available for decoding and accessing memory. In the future, the 88100's clock frequency will increase, leaving even less time. To make sure that memory technology does not impede the processor's progress, a cache memory-management unit has been designed.

so that the instruction fetched after the branch instruction can be executed before the branch is taken. In this manner the instruction fetched after the branch instruction is executed while the branch target instruction is fetched from memory by another section of the pipeline. This eliminates a "bubble", or delay in the pipeline.

Separate execution units permit concurrency; in other words each of the execution units can operate independently allowing instruction decoding and source-operand fetching to be performed at the same time by the same execution unit, and branch-instruction decoding or branch-target address calculations to be performed in parallel with sequential address generation.

Obviously with such extensive internal bus traffic some form of internal routing is required. The processor contains three internal 32bit buses, allowing the 88100 to perform *simultaneous* rather than sequential op-code fetching.

For multiprocessor applications a memory exchange instruction, XMEM, has been added. When the instruction is executed the contents of two memory locations are exchanged in a locked bus cycle so that semaphore register bits can be updated without the risk of interruption from another processor.

Bit-field instructions are included in the instruction set for applications requiring manipulation of data that does not correspond to a particular standard format like, for example, frame buffers in graphic applications.

The method used by the 88100 in fault-tolerant systems, for example, p.a.b.x. military and medical applications, is shown in Fig.1. Two 88000 processing nodes are wire-Ored together. By programming the processor, one node is assigned to be a master while the other acts as a checker. The checker floats all its outputs except for the error pin, ERR. Its outputs are then monitored as inputs. The checking processor executes the same instructions as the master, comparing its internally generated output signals with information on the floated pins (outputs actually produced by the master). If at any time there is a mismatch then the checker asserts the ERR pin to signal to the external logic a fault on the node.

In addition the master checks that the internal value of the signals which it is driving are the same as those on the output of the device. In other words the inputs and outputs of the processor's drive buffers are compared. If an error occurs it is likely that some form of external short circuit has occurred. Again the ERR pin is asserted, this time by the master. The external controller can then act on the output of the error signal to report or replace the faulted processing node.

INTERNAL ARCHITECTURE

A block diagram of the internal architecture of the 88100 is shown in Fig.2. Comparing this to a typical complex-instruction-set computer highlights how clean the architecture of a reduced-instruction-set computer is. Rather than having complex areas of control logic, a number of separate execution units are implemented, each containing their own complex but *independent* logic.

A simple load-and-store architecture is supported, allowing most of the instructions to be executed within the program registers. Triadic addressing, where two source registers are acted upon to yield a result for a separate destination register, is used extensively. This form of architecture supports non-destructive data manipulation of the source registers. Accessing three registers simultaneously in only one clock cycle is achieved via three parallel buses, two for the sources and one for the destination.

The three buses are responsible for routing data to the execution units and the registers, all of which are 32 bits wide and completely general purpose. Bus activity is monitored and controlled by the sequencer unit.

Dedicated instruction-execution units have been implemented, namely the integer unit and the floating-point unit. Pipelining these units provides simultaneous instruction execution with data and instruction fetching. For speed, integer arithmetic is performed on both the integer unit and the floating-point unit, whose instructions conform to the IEEE P754 specification.

Independent execution units control the operations on the code and data buses of the 88100. Data is fetched to or from memory or the c.m.m.u. immediately into a pipeline in the data unit. This allows prefetching of data into the device so that it is fetched on the external bus before the processor requires it. Effectively, this hides the time required to fetch or write data outside the processor. The instruction unit has a similar pipeline for instruction fetching; in this case some predecoding can be done on the instruction before it is required by the processor.

The remaining portion of the architecture is assigned to future execution units or special-function units. When a new family of microprocessors is developed it is very difficult to foresee the requirements that will be placed upon it in the future. Therefore it is common to include in the original design some elements which can be designated as "for future expansion".

This approach has been adopted for the 88100, which has "room" for expansion in the form of special-function units. The architecture allows up to eight special-function units to be designed into the processor; at present only one is used, namely the floating-point unit. Accesses to a new special-function unit will be the same as an access to the floating-point unit, the only difference being the bit-pattern change in the upper 32 bits of the opcode word that addresses a different unit.

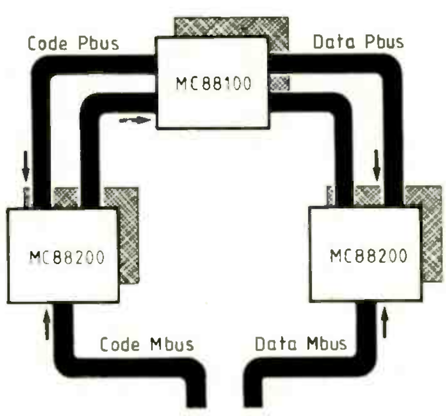


Fig.1. In fault-detection mode, one 88000 node is a master and the other a checker. The checker monitors the activities of the master and asserts the ERR pin should a fault be detected.

Floating-point unit. As its name implies, the floating-point unit is responsible for all internal floating-point operations within the microprocessor. Indeed, this unit not only performs all the floating-point arithmetic instructions, but also integer/floating-point conversions and integer multiply and divide instructions.

When an integer unit exists within the 88100, why does the floating-point unit need to perform integer multiplication and division? The reason is that it allows the integer unit to perform many instructions in only one clock cycle, leaving the multiple-cycle instructions to execute in the more complex a.l.u.

Obviously, the inclusion of floating-point instruction-execution on chip implies that there are a number of instructions that the 88100 can execute which do not complete in only one clock cycle. Single-precision add, subtract and convert instructions require five cycles while the multiply requires six. Divide instructions execute by looping through one stage of the pipeline for each bit of required accuracy. To compensate for these delays the floating-point unit contains two pipelines, one of five stages and the other of six.

By proper scheduling of floating-point and integer add and divide instructions, the processor can be made to execute a number of floating-point instructions on each consecutive clock cycle.

Integer unit. Remaining integer arithmetic, bit field, logical instructions, etc, execute under control of the integer unit, which is also responsible for all the operations required for memory addressing calculations and program sequencing. During exception processing sections of the integer-unit control registers are used for storage of program status information.

The integer unit actually contains three separate functional units to increase throughput. All bit-field instructions are executed in a bit-field unit, alleviating the main a.l.u. from this task. Integer arithmetic instructions execute in a 32-bit a.l.u. and allow a wide range of logical and arithmetic operations. A separate branch unit is used to calculate the branch offset, permitting the implementation of delayed branching for all branch instructions.

This method of separating tasks in the integer unit allows a considerable amount of parallelism to be achieved in that three of the above types of operations can be executing simultaneously in different registers. A compiler designed to monitor register usage very carefully can take considerable advantage of this type of architecture.

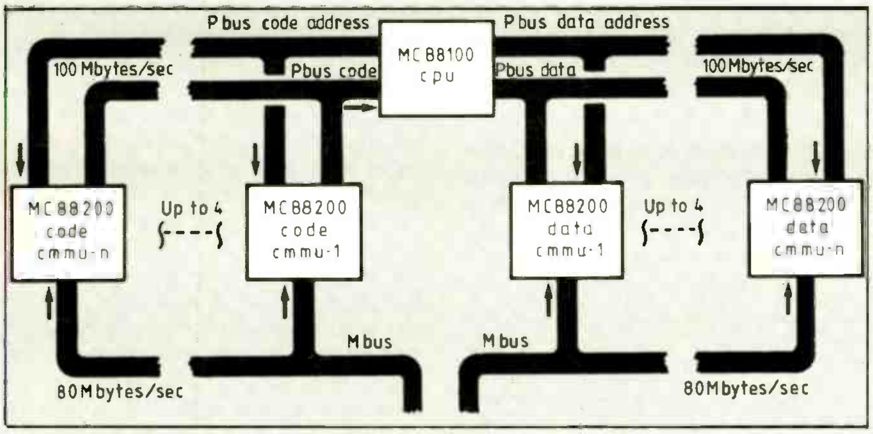
Register file. The programmer's access to all the internal execution units is via programs

CACHE MEMORY MANAGEMENT REDUCES THE PROBLEM OF MEMORY SPEED

When 88100s are produced that run faster than 20MHz, it will be difficult for memory technology to keep pace. For this reason, a cache memory-management unit has been designed for use with the processor. It contains two sections: a four-way set-associative cache of 16Kbyte operating without waiting states, and a full demand-page virtual memory-management unit.

The c.m.m.u. is designed to operate on both code and data sides of the 88100 buses; indeed a typical 88000 node would consist of a minimum of one c.m.m.u. on each of the 88100 buses. A maximum of four c.m.m.us can be configured on either the code or data buses, allowing the sizes of the caches to be increased and the memory-management unit's address range to be increased.

Simulation results for the c.m.m.u. are showing greater than 95% hit rates in the caches when running typical Unix type code. With hit rates of this order, there is little need for two external buses linking to external memory so the memory ports on the c.m.m.us are multiplexed together to simplify external memory design.

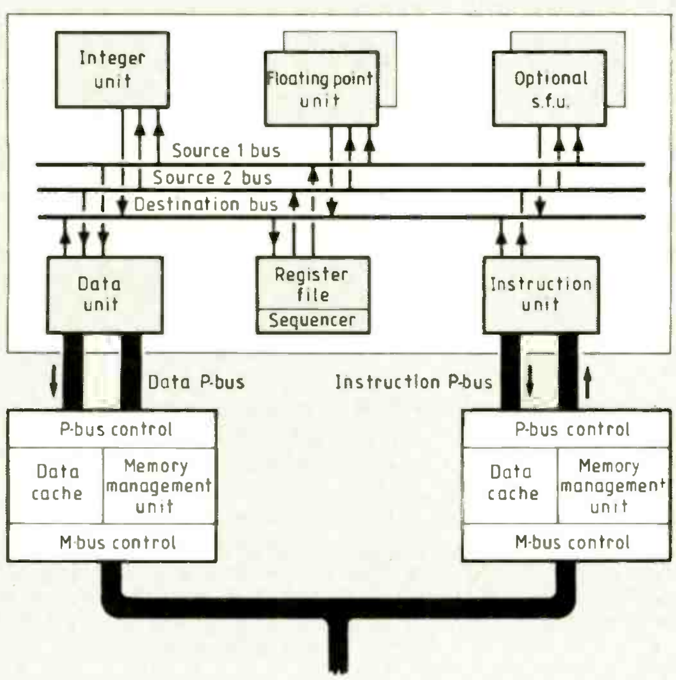


For zero wait-state access to memory, the 88100 interfaces to up to four cache memory-management units on each bus. These contain a 16Kbyte four-way set-associative cache and a demand-page virtual m.m.u.

executing in the register file containing 32 general-purpose user registers, r_{0-31}. The only restriction to the use of these registers is in the use of r_0 and r_1. Register r_0 always returns zero when read. This allows very quick clearing of register contents for initialization code since r_0 can be moved to any of the other registers in the register file. The return address of a branch to a subroutine is held in r_1 by the processor.

To allow simultaneous access to the register file by the three internal buses they have three ports: two take input from the two source buses and one outputs to the destination bus. For multiple register usage a method of feeding foward information has been implemented for the internal buses. This places the destination result from the current operation required for execution by the next instruction on one of the source buses to ensure that the unit waiting for data receives it at the earliest possible time. The same applies if both source registers are in use: both source operands are obtained by forward feeding from the destination bus.

Fig.2. Internal architecture of the 88100 consists of a number of execution units that operate concurrently.



A method of internal arbitration is performed by the register file for controlling internal control and storage register write protocols. The scheme is implemented to prioritize register writes from the execution units, the highest priority being the integer unit followed by the floating-point then the data unit.

Sequencer. Rather than a section of microcode to control all the function of the 88100 a hardware sequencer is used. It controls register access, arbitration of the internal buses, exception processing and the generation of control signals for the integer and special-function units.

The main problem the sequencer has is knowing what register is actually being used at any one time. Due to the nature of this parallel concurrent processor several execution units have independent access to the register file. For this reason each of the general-purpose registers in the register file has an associated scoreboard bit to indicate usage. When a particular register is being used by one of the execution units a bit corresponding to that register is set in the scoreboard register. This ensures that parallel instruction execution does not use any source operand registers that have not yet been computed from a previous instruction. In other words the processor is not restricted to executing each individual instruction to completion before starting a new one.

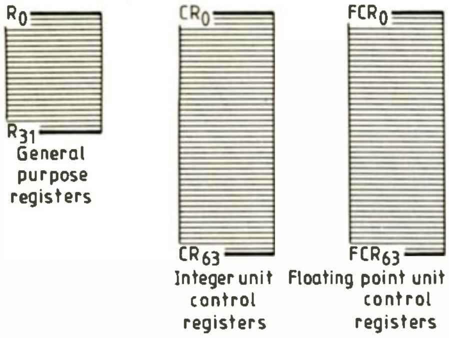
To understand the role that the sequencer plays in handling exceptions the exact definition of an exception on the 88100 is required. There are two types of exceptions, precise and imprecise. A precise exception is one in which the exact processor context and exact causes are know. These include interrupts and bus errors. These types of exception contain all the information required by the processor for a complete recovery. Imprecise exceptions are those caused spuriously when executing instructions, for example, when the floating-point unit overflows. In this case the whole context is not required for recovery if the operation in progress was known.

The sequencer's responsibility for handling exceptions can affect performance: to minimize the effects, the sequencer has the ability to act on all pending precise exceptions before handling imprecise exceptions. Imprecise exceptions require investigation work by the exception handler.

Instruction unit. Operations required of the instruction unit are to fetch instructions form the instruction data bus and to perform the first stages of the instruction decoding before sending them to the integer unit. This operation is performed in only one clock cycle where possible.

The instruction unit has an integrated three-stage pipeline containing the value of three instruction pointers, the first of which is the execution instruction pointer (x.i.p). This pointer is the address of the instruction currently being executed in the instruction or floating-point unit. The second stage contains the next-instruction pointer (nip) which points to the instruction currently being received from memory or the

Fig.3. The programming model consists of 32 general-purpose 32bit registers and 128 control registers, also 32bit wide, in the integer and floating-point units. Note that there are unused control registers for future expansion.



c.m.m.u. Due to the external bus operation, data is received on the clock after the address, therefore during this phase the fetch-instruction pointer (fip) contains the memory address location of the next instruction to be accessed. The three-stage pipeline caters for the prefetching of instructions before the processor is ready to use them and for recovery from an erroneous bus access.

Data unit. Within this execution unit, for controlling aligned 32bit data accesses, is a 30bit arithmetic unit for address calculations. A three-stage pipeline improves data fetching operations.

In the first stage of the pipeline the required fetching address is calculated, or for a store operation, data is fetched from the internal register file. During stage two the external address bus is driven to fetch the data; in the store operation the data bus is also driven. In the final stage of the pipeline the response of the external memory is monitored. For a store operation the data bus is read and for a write operation a register in the register file is loaded.

PROGRAMMING MODEL

Figure 3 shows the program and control register of the 88100. After the reset sequence the processor operates in supervisor mode. In this mode all the registers are available for programming and the hardware of the processor can be configured for the particular application area. In user mode, control registers are protected from being accessed to prevent system corruption, and the general purpose registers are used to program the processor.

Supervisor mode is normally restricted to kernel type software so only a small amount of very fast exception handling code, etc, resides in this mode. While in supervisor mode the 88100 can be programmed to perform a number of functions. For fault-tolerant applications the processor can be programmed to be a master or a checker. Big Endian or little Endian* modes can be enabled to fetch data, which is useful for interfacing the 88100 to a number of different vendors' bus systems.

Serialized mode can be enabled for debugging code. In this mode no instruction can start execution until all the bits of the scoreboard register have been cleared. This ensures no overlap with the pipelines for concurrent instruction execution. To facilitate arithmetic operations the carry flag can be enabled or disabled for add and subtract instructions, and the floating-point unit can be enabled or disabled. The processor can be configured to generate an exception on a misaligned data access. Interrupts can be enabled or disabled as can the shadow registers.

The user has access to all the general purpose registers although r_0 and r_1 have special functions as explained earlier. Access to the floating-point unit by the user program is through the floating-point control and status registers. These are the only control registers accessible to the user.

No restriction is placed on the remaining general-purpose registers. The registers are not allocated special tasks, unlike for example, the 68000 family's data and address registers. There is no stack pointer or indeed condition-code register. If there is no stack in either the supervisor or user modes then obviously there must be some other method for handling interrupts etc. With no condition code register there must also be another method of handling branches and compares.

During the evaluation of this project these items were looked at with respect to performance. In many ways it was the old style architectures which contained both stacks and condition-code registers that hindered the performance of these processors.

The 88100 solves the problem of having to spend processor time staking registers during an exception in the following manner. Integrated into the interger unit are a number of "shadow" and "exception time" registers. The idea behind shadowing is to record the contents of the internal pipeline registers during instruction execution. In other words a copy of these registers is taken on each clock cycle. Information held in these registers is the three instruction pointers and the scoreboard register.

At the time of an exception the contents of the shadow registers are frozen, thus leaving a copy of instructions that the processor was about to act upon at the time of the exception. The exception-time registers are used to save other internal information at the time of the exception. With the contents of the shadow and exception-time registers the complete internal context of the processor is know at the time the exception was taken.

In addition, the user can create a stack for other register information and enable nested exceptions if required, utilizing the general-purpose registers or external memory. This method of exception handling in hardware reduces the interrupt latency time considerably compared to that of stack-oriented processors.

When the features of parallel architecture in a microprocessor are analysed the inclusion of a fixed condition-code register is detrimental to the processor's performance. If the condition-code register were fixed, evaluation of, for example, the result of a compare instruction would imply that one instructon takes priority over another for access to the results of the condition-code bits. If this were the case then the second instruction would be required to halt execution until the first had operated on the bits in the condition-code register.

Fig.5. While the clock is high, the processor drives the addresses and the bus type indicating a read or write operation. Data is read on the next rising edge and the response is read on the next falling edge.

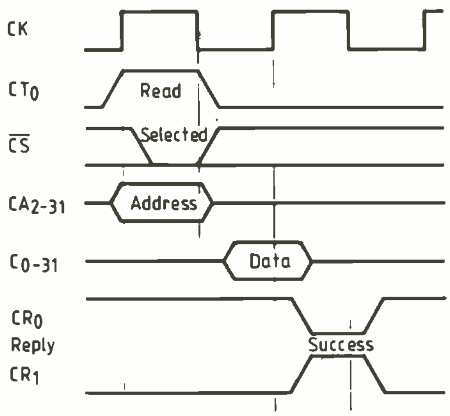


Fig.6. Write cycles are similar to read cycles, but byte strobes are asserted at the same time as the addresses and the data is driven immediately after the address.

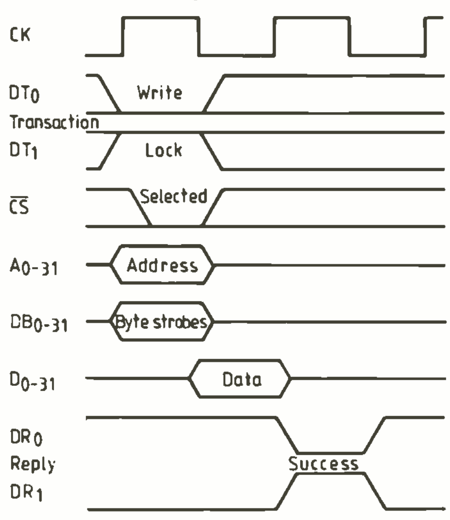
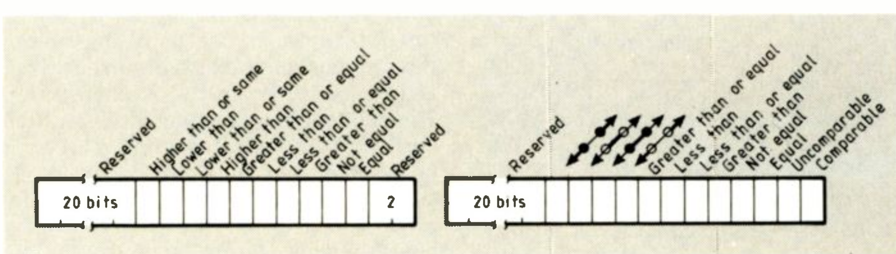


Fig.4. Rather than use a condition-code register, conditions, or predicates, are evaluated in a general-purpose register. All the predicates are shown here.



* Most-significant data first or last.

In the 88100 this type of operation is impractical as it would have serious consequences on the performance of the entire machine. So rather than a fixed condition-code register the 88100 uses a string of predicates which can be evaluated in any of the general-purpose registers. As long as the destination register that the predicates are being evaluated in is not used by the instructions that follow then a string of compared or branch-like instructions can execute without delay. The full range of predicate values are shown in Fig.4.

It is not uncommon to see more than 128 general-purpose registers in a risc processor architecture, as opposed to just 32 as with the 88100. The number of registers used within the microprocessor will obviously effect the performance of the final device. During the design phase of this project, multiple register usage was simulated. It was found that 32 registers proved to be an optimum number for high-performance applications.

There were a number of findings that helped Motorola arrive at this conclusion, one of then being an analysis of high-level language programs running on register-based microprocessors. It was found that an overflow situation occured which affected the performance of the compliers when controlling data allocation on such a large number of registers. It was physically impossible to keep track of such a large number of registers – with above about 25, complier performance actually starts to decrease as the number of registers increases.

Physical limitations of the device also come into play; drive capability, or fan out of the decoder is affected by capacitive loading as the number of registers increases. This restricts register access times and hence slows down the processor clock frequency.

Another consideration is that the number of bits in the opcode word required to decode the register becomes greater, hence placing a restriction on the size of immediate data values.

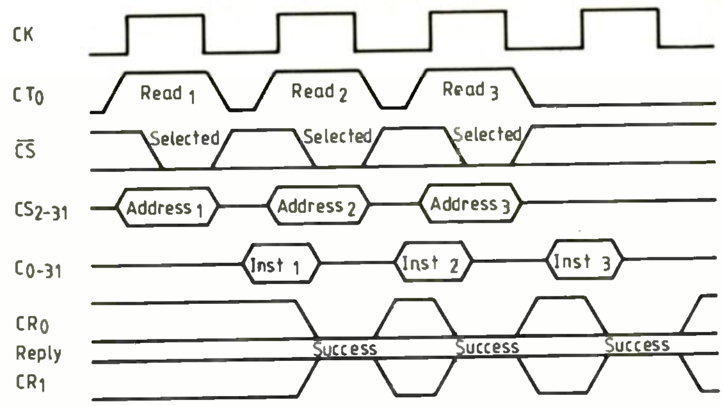


Fig.7. In pipeline mode, the 88100 can read and write data on consecutive clock cycles.

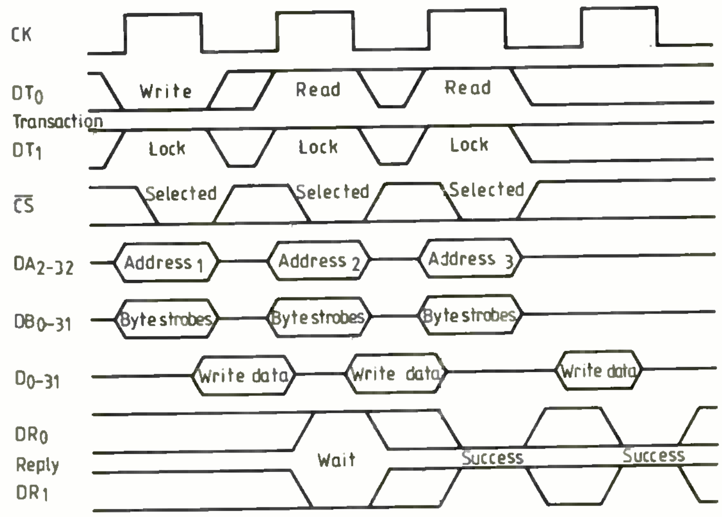


Fig.8. When a fault or wait-reply condition is read by the processor, the bus cycle is either repeated or ignored. When the reply is read, the second bus cycle has already started so the processor does not wait for slow responses from memory.

INSTRUCTION SET AND ADDRESSING MODES

Bit field instructions allow the 88100 to operate on individual bits of information such as individual pels in a frame buffer of information. This is just one data size that the processor can work on. Single and double precision floating-point numbers conforming to the IEEE specification are operated on by the floating-point unit. The operating size of a word for the processor is 32 bits, but it can also act directly on 16bit quantities, known as half words, double words of 64 bits, and bytes of information.

Although not as complex as those of a complex instruction-set computer a number of addressing modes are present in the 88000 architecture for accessing instructions, data and the internal registers.

Instructions are fetched using three types of addressing mode. The first of these is called 'register with a 9bit vector table index'. It is used to place a value in the fetch-instruction pointer corresponding to the exception-vector address during exception processing. The addressing mode concatenates the 9bit vector offset for the exception with the vector base registers to produce the final fetch-instruction pointer. At reset, the vector base register is programmed to an address corresponding to where the exception vectors are placed in memory. For a trap-on-condition exception, five bits in the opcode word are tested and only if they are true is a new value loaded into the fetch-instruction pointer.

The second instruction-addressing mode is 'register with 16bit field'. This mode is used by branch and trap instructions for target-address and test-condition generation. In this mode source register S_1 is tested. If the branch or trap condition is true then the 16bit offset is added to the execution instruction pointer and placed in the fetch-instruction pointer.

Branch instructions use the third instruction-addressing mode, '26bit branch offset', to specify the branch target address. The 26bit offset in the opcode word is added to the execution-instruction pointer to produce the fetch-instruction pointer.

Of the data-memory addressing modes register-direct with unsigned offset is the first. In this data fetching mode, content of source register S_1 is added to a 16bit literal field in the opcode word to produce an external memory address for data access.

The second data-addressing mode, called 'register direct with index' is similar to the first. In this case a second source register replaces the literal value to index the first to produce the memory address where data is to be fetched or written to.

In the third data-addressing mode, 'register indirect with scaled index', contents of the second source are scaled by a factor of zero, one, two or three. Note that in this case, 'one' is the same as the second data-addressing mode.

Of the internal register-addressing modes, the first is triadic register. In this mode source registers S_1 and S_2 are acted upon by the integer or floating-point units to produce a result for a separate destination register. Note that no register contents are destroyed.

For bit-field instructions, the second internal register-addressing mode, 'register with 10bit field', is used. The ten-bit literal field in the bottom bits of the opcode word contains the width and offset of the bit field to be acted upon in source register S_1.

In the final addressing mode, 'register with 16bit field', source register S_1 contents are added to a 16bit literal value of the opcode word to produce the destination result. This form of addressing is used by arithmetic and logical instructions requiring an immediate value.

THE P-BUS

Access to both code the data memory is via two processor buses called p-buses. These are very fast synchronous buses operating in one processor clock cycle. Signals controlling access over these buses are shown in the first panel. The top of this diagram shows signals associated with the data and code p-buses.

Data accesses are made via a 30bit data-address bus, $DA_{2\text{-}31}$, and 32bit data bus, $D_{0\text{-}31}$. However since the 88100 only accesses data as aligned words this corresponds to an addressing range of 4Gbyte.

Controlling accesses to data memory are the following control signals; DAS, $DT_{0,1}$, $DB_{0\text{-}3}$ and $DR_{0,1}$. Signal DAS is used to separate user from supervisor accesses to data memory. This signal could be used in the external control logic to double the size of directly accessible memory, that is to allocate 4Gbyte for user data and 4Gbyte for supervisor data. In addition supervisor data can be protected from corruption by user programs. When this pin is low, user data space is selected.

Signals $DT_{0,1}$ indicate the type of data access being performed. Signal DT_0 is equivalent to the read/write pin on other processors, except that it is asserted at the same time as the address bus is driven. Signal DT_1 is asserted during a locked bus cycle. The XMEM instruction causes DT_1 to be asserted for the duration of the memory-exchange operation.

Response to a data access is reported by signals $DR_{0,1}$. Depending on the decoding of these signals they will indicate either a successful memory transfer, a memory fault, or a memory wait cycle. Signals DB_{0-3} select individual byte locations within the fetched word of data.

Code accesses on the code p-bus are similar to those on the data bus. The differences are: there are no byte strobes since all opcodes are 32 bits wide and no extension words are supported; the code data bus, C_{0-31}, is unidirectional; the CT_0 is used to indicate that an instruction fetch is in progress.

To support accesses in one clock cycle the buses used for both code and data need to operate in only one clock, therefore a new type of fast bus was designed. Most microprocessors operate by first driving the address bus, qualifying the addresses with a strobe signal, reading from or writing to memory and then finally waiting for an acknowledgement from the memory device before terminating the cycle. Obviously with this type of bus operation the processor needs to wait for a period greater than one clock cycle between accesses.

The p-bus can assert addresses on each clock and drive or read the data bus on each clock. Latching of the acknowledgement occurs after the cycle so that the processor is not slowed down. Only when a fault exists does a delay occur. All p-bus activity is synchronized to the 88100 processor clock, therefore the requirement to supply strobing qualifying signals is eliminated, simplifying the external control logic.

Figure 5 shows how the p-bus performs a read cycle over the code bus. At the same time as the address bus is driven, just before the rising edge of the clock, CT_0 is asserted to indicate an instruction fetch. Next, the address is decoded by the external memory/cache/c.m.m.u. and if accessed places the opcode word on $C_{0\text{-}31}$ before the next rising edge of the clock. Response to the processor is read on the falling edge of the clock. For correct bus operation the only requirement is that the signals meet the set-up and hold times with respect to the processor's clock. The only difference between this cycle and a data-read cycle is that the byte strobes are also driven at the same time as the addresses.

During a write operation the 88100 places data on the data bus just before the falling edge of the first clock in the cycle. Timing of a data write operation is shown in Fig.6. Instructions are read-only so there is no instruction write cycle.

Pipelining the accesses increases the throughput of the 88100, making one-cycle instruction execution possible. The pipelined bus protocol for an instruction fetch is shown in timing diagram Fig.7. This method allows two instruction reads to be outstanding on the bus at any one time. Therefore the processor can initiate each transfer before the proceeding transaction has completed. In effect this allows the processor to fetch instructions and/or data on every rising edge of the clock. Pipelining is also supported for data-bus writes.

In many cases it may not be possible to access code or data in only one clock cycle. In this case the external devices on the p-bus must send back a wait signal on reply pins, $DR_{0,1}$, as shown in Fig.8. When the processor reads this response it repeats the second address on the next rising edge of the clock. The first address that caused the fault should be latched externally. If a wait signal is encountered, then this latched address can be used to access the external devices. The p-bus operation allows enough time to latch the first address externally.

On an erroneous bus cycle the processor ignores the second bus cycle and continues execution on the third cycle. Enough time is given to buffer the erroneous access. This allows a memory-management unit to load a page containing the faulted address and begin execution by retrying the buffered address. Virtual memory systems, can thus be designed on both code and data buses.

CONCLUSION

Hardware and software features of the 88100 have been described in this article and indications of how high performance can be achieved from the architecture have been given. The reduced and simple instruction set makes designing code at assembly level very difficult, as with most risc microprocessors. For speed of software development there will therefore be a shift toward high-level languages.

The architecture of the 88100 is designed for very high throughput of code and data. Therefore the overhead of writing code in C for example is minimal. However the design of highly optimized compilers for the processor is important if all the advantages of the architecture are to be used.

To demonstrate how compilers can be used to develop highly optimized code an example is shown in List 1. This is a C source

Continued on page 691

```
FFT(xr, xi, n, m, wr, wi)
single xr[ ], xi[ ], wr[ ], wi [ ];
unsigned n, m;
{
    unsigned n1, n2, ia, ie, i, j, k, l;
    single c, s, xrt, xit;

    n2 = n;
    ie = 1;
    for (k = 0; k< m; k++)
    {
        n1 = n2;
        n2 = n2 >> 1;
        ia = 0;
        for ( j = 0; j < n2; j++ )
        {
            c = wr[ia];
            s = wi[ia];
            ia = ia + ie;
            for ( i = 0; i <n; i+=n1)
            {
                l = i + n2;
                xrt = xr[i] - xr[l];
                xr[i] = xr[i] + xr[l];
                xit = xi[i] - xi[l];
                xi = xi[l] + xi[l];
                xr[l] = c*xrt - s* xit;
                xi[l] = c*xit + s*xrt;
            }
        }
        ie = ie << 1;
    }
}
```

List 1. Fast Fourier transform in C.

```
L3:
    add         r16,r15,r8      * l = i + n2
    ld          r19,r1,r15      * xr[i]
    ld          r20,r1,r16      * xr[l]
    fsub.sss    r17,r19,r20     * xrt[i] - xr[l]
    fadd.sss    r19,r19,r20     * xr[i] = xr[i] + xr[l]
    st          r19,r1,r15      * xr[i] = xr[i] + xr [l]
    ld          r21,r2,r15      * xi[i]
    ld          r22,r2,r16      * xi[l]
    fsub.sss    r18,r21,r22     * xit = xi[i] - xi[l]
    fadd.sss    r21,r21,r22     * xi[i] = xi[i] + xi[l]
    st          r21,r2,r15      * xi[i] = xi[i] + xi[l]
    fmul.sss    r23,r11,r17     * c*xrt
    fmul.sss    r26,r12,r17     * s*xrt
    fsub.sss    r20,r23,r24     * c*xrt - s*xit
    st          r20,r21,r16     * xr[l] = c*xrt - s*xit
    fmul.sss    r24,r12,r18     * s*xit
    fmul.sss    r25,r11,r18     * c*xit
    fadd.sss    r22,r25,r26     * s*xit + c*xrt
    st          r22,r2,r16      * xi[l] = s*xit + c*xrt
    add         r15,r15,r7      * i = i + n1
    cmp         r27,r15,r3      * i < n
    bb1         lt,r27,L3
```

List 2. Assembler output of the inner loop of the FFT from a C compiler with no optimization.

```
    add         r16,r15,r8      * l = i + n2
    ld          r19,r1,r15      * xr[i]
    ld          r20,r1,r16      * xr[l]
    ld          r21,r2,r15      * xi[i]
    ld          r22,r2,r16      * xi[l]
    fsub.sss    r17,r19,r20     * xrt[i] - xr[l]
    fadd.sss    r19,r19,r20     * xr[i] = xr[i] + xr[l]
    fsub.sss    r18,r21,r22     * xit = xi[i] - xi[l]
    fadd.sss    r21,r21,r22     * xi[i] = xi[i] + xi[l]
    fmul.sss    r23,r11,r17     * c*xrt
    fmul.sss    r26,r12,r17     * s*xrt
    fmul.sss    r24,r12,r18     * s*xit
    fmul.sss    r25,r11,r18     * c*xit
    fsub.sss    r20,r23,r24     * c*xrt - s*xit
    fadd.sss    r22,r25,r26     * s*xit + c*xrt
    add         r15,r15,r7      * i = i + n1
    cmp         r27,r15,r3      * i < n
    st          r19,r1,r15      * xr[i] = xr[i] + xr [l]
    st          r21,r2,r15      * xi[i] = xi[i] + xi[l]
    st          r20,r21,r16     * xr[l] = c*xrt - s*xit
    bb1.n       lt,r27,L3
    st          r22,r2,r16      * xi[l] = s*xit + c*xrt
```

List 3. If the load-and-store instructions are grouped together at the start and end of the program and all floating-point operations are also grouped, a better result emerges. The code produces the same result, but it is arranged to take advantage of the processor architecture.

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H.f. developments

Notes from the Fourth International Conference on H.f. Radio and Techniques, held at the IEE in April.

GEORGE SHORT

In these days of overcrowding and deliberate jamming in the short-wave broadcast bands, the domestic listener needs a little help from the engineer. The directional null property of a loop aerial is attractive as a potential weapon against co-channel interference. To be of use to the non-technical listener with minimal constructional skills any design must be simple to make and use and should preferably be small for use indoors.

Two ingenious new designs which both meet these requirements and break new ground were described. One of them is a notable advance in that it works against interference by sky-waves, where the nulling power of a traditional loop is seriously impaired by multi-path propagation. The other, which is effective against ground-wave interference from local transmitters, is particularly easy to make and use. Both derive from the ideas of Professor O.G. Villard of SRI International, California.

HORIZONTAL LOOP ANTENNA (HLA)

The HLA[1], developed by G. Hagn and C.A. Hagn (Fig.1), is a gapped rectangle of metal plates or foil, laid on a table top or any similar rigid nonconducting horizontal plane. A variable capacitor (medium-wave type) in the gap tunes to the wanted signal.

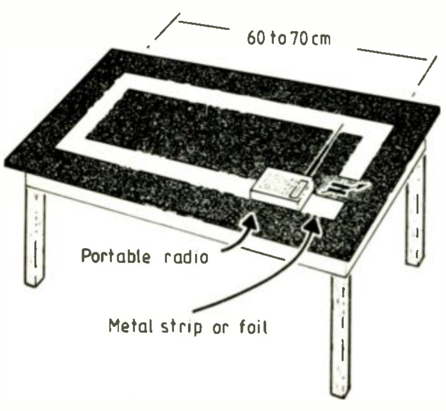


Fig.1 Horizontal Loop Antenna. This responds to the vertical magnetic field of a horizontally polarized wave. The loop is blind to vertically polarized groundwave interference because this has no vertical magnetic field. The loop inductance is low (1μH) and the resulting low impedance eliminates hand effects. Ways of improvising the tuning capacitance from domestic materials are described in the IEE paper.

Being horizontal, the HLA has minimal response to vertically polarized signals, which makes it blind to any vertically polarized local transmitter, irrespective of its direction. Sky waves from distant transmitters can be received, because even if the original transmission is vertically polarized, horizontal components are generated in the ionosphere. The HLA therefore discriminates against local jamming and interference while accepting distant transmissions. It is virtually omnidirectional and works well with signal-arrival angles 3-60 degrees above horizontal. Attenuation of unwanted ground waves is about 30dB.

A portable receiver with a non-conducting cabinet can be capacitively coupled to the HLA simply by laying it on one side of the gap and extending its built-in whip aerial to the opposite side. Actual electrical contact is not necessary.

A tuning capacitor can be improvised by overlapping the plates and inserting a sheet of dielectric between them; waxed paper and plastic film are suitable. A compression tuning effect is obtainable by slightly bowing the upper plate and moving a heavy object (book, etc.) along the plate over the dielectric. This suggests the possibility of hiding the HLA beneath a tablecloth and tuning by positioning a tray, teapot, samovar, etc. In areas of very strong groundwave field this may be impracticable, because for optimum rejection the table must be tilted to compensate for the tilting of the groundwave front by the earth. In such cases the HLA is best mounted on a separate, tiltable support plane.

Signals may also be extracted from the loop by a one-turn coupling loop of coaxial cable. They can then be taken to a fixed receiver. The Hagns report that, even when there is no interference, the HLA still often gives better reception than the receiver's pull-out whip. This is partly attributable to the Q of the loop (typically 30-40) which locally magnifies the signal field and also gives a useful amount of preselection. Loops like this, made from wide metal strip, have low inductance (1μH) and low impedance, making them free from hand effects.

COPLANAR TWIN LOOP (CTL)

To those familiar with the figure-of-eight bidirectional response of a simple loop aerial it comes as something of a shock to learn that loops can be arranged to give a unidirectional cardioid response pattern with a single broad acceptance lobe at the front and a sharp null at the back. This feat of technical legerdemain is performed by Professor Villard's Coplanar Twin Loop (CTL)[2].

The CTL exploits a little-known property of high-impedance loops (Fig.2), where L is large and C small (10pF). Such a loop shows a significant amount of response to the electric (E) field as well as to the magnetic (H) field.

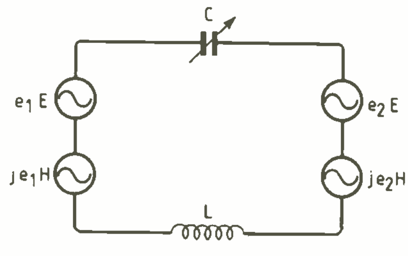


Fig.2 High-impedance loop. A one-turn wire loop is inductance-loaded (L) so that the required tuning capacitance C is small (10pF). Significant amounts of signal voltage are then produced by the electric (E) component of an incident wave as well as by the magnetic (H) component. Left-to-right waves produce different voltages from right-to-left waves. This directional effect is exploited in Villard's Coplanar Twin Loop to give a unidirectional response.

Assuming vertical polarization and a wave travelling left to right in the plane of the loop, e.m.fs as shown are induced in the vertical sides. Since, in this orientation, the 'electric' e.m.fs (e1E, e2E) are unequal, a net e.m.f. is available to drive current round the loop. The same goes for the quadrature 'magnetic' e.m.fs je1H, je2H. If the direction of the wave is now reversed, the instantaneous sign of the electric e.m.fs is unchanged, but that of the magnetic e.m.fs is reversed. Hence the loop's net response is changed when the wave direction is changed.

This effect is not noticeable in a low-impedance loop made from a broad band of metal and tuned with a relatively large capacitance (e.g. by using a medium-wave tuning capacitor). In this case the loop impedance is so low that the electric e.m.fs are virtually shorted and only the magnetic response remains.

In the CTL, (Fig.3) a small, low-impedance magnetic loop is positioned inside a larger high-impedance loop. The two are kept in the same place by attachment to a rigid support such as a sheet of plywood. E.m.fs are induced in the inner loop in two ways: directly by the incident wave and indirectly by the magnetic field set up by the current circulating in the outer loop. As we have seen, part of this current is due to the electric field of the wave. The resulting e.m.fs in the inner loop oppose one another. If the coupling between the loops is adjusted

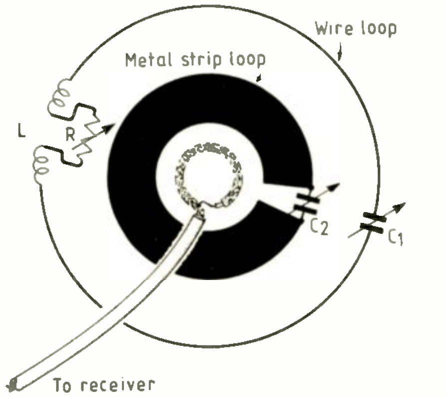


Fig.3 Coplanar Twin Loop aerial. A low-impedance metal strip loop is positioned inside a high-impedance loop. Both are tuned to the same signal. When coupling is adjusted by the damping resistance R a unidirectional null is obtained.

a setting can be found where they cancel.

However, this nulling effect holds good only for one loop orientation. If the coupling is set to produce a deep null for a wave approaching from the left then there is no null for one coming from the right. It turns out that when the loop is oriented to null a signal from the left then the response to signals from the right is maximum.

SKY WAVE NULLING

Since the two loops are close together, changes in the wave field caused by ionospheric variations affect both equally and the null holds good. This is just what is needed for combatting skywave interference. Provided there is enough angular separation between a wanted and an unwanted signal, the unwanted one can be nulled out, or at any rate reduced by more than 20dB. The sort of directional patterns seen in field tests of a CTL are indicated in Fig.4, with the response of a simple loop superimposed for comparison. The degradation of the nulls of the simple loop by multipath effects is typical and shows why such a loop is useless for nulling skywave interference.

The CTL is not too difficult to set up. The inner loop is tuned to mid-band then left alone. Next, the outer loop is tuned. It needs

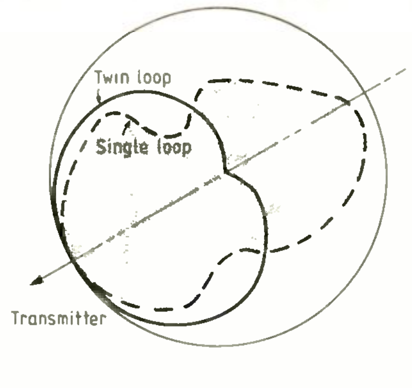


Fig.4 Directional response. The cardioid response of a CTL shows a broad acceptance lobe one way and a 20dB null the other way, when receiving even multi-hop skywave signals. The poor skywave performance of a single loop is shown for comparison.

readjustment if the frequency is changed by more than a few kilohertz. Coupling is adjusted by means of a damping resistance in the outer loop; one coupling adjustment per band is usually enough.

Diversity reception. The CTL lends itself to a form of diversity reception. Under normal multipath conditions it is possible to null out the direct paths to one receiver while a second receiver still responds to them. Thus the receivers respond to two directions of signal. Differences in the path lengths can be large enough to create a pseudo-stereo effect.

Clearly, the CTL can also be set to null groundwave interference, but its main attraction is for skywaves, where its relative insensitivity to arrival angle can be very useful.

H.F. RADAR

Historically, h.f. radar is quite venerable. The earliest practical system of 'radiolocation' (the CH system) used transmissions at 20-30MHz. In those days, however, it was logical to use horizontal polarization, to maximize reflections from aircraft wings. Today, one important type of h.f. radar, ground wave (surface wave) radar (Fig.5) invariably uses vertical polarization. This enables transmitters on the sea coast to take advantage of the refraction produced when vertically-polarized waves are launched over sea water. Such waves propagate far beyond the microwave horizon and can therefore extend the useful range of ground-based radars.

Suitable transmitter sites may be hard to find. A paper by Marconi researchers[3] spells out the requirements. The aerial must be at or very close to sea level, which rules out cliff-top sites. It must be on level ground soaked by salt water; fresh water has insufficient conductivity to refract decametric waves adequately. Since returning echoes are weak the receiver should be remote from man-made noise.

The logical answer is to site the system on a lonely salt marsh — a bleak prospect which might be slightly mitigated by engineering the ground[4]. The required vertical monopoles can be erected over a wire ground screen. Research at Birmingham University has demonstrated that in one special case a further technique can be used. If the transmitter site consists of a stratum of dielectric material over an underlying high-conductivity layer (e.g., dry sand over brine-soaked sand) the dielectric layer can with advantage be inductively loaded by driving into it vertical metal rods. The researchers used over 600 aluminium rods each two metres long and obtained an improvement in field strength at ground level of around 3dB.

The fixed transmitting aerial of an h.f. radar is used to illuminate a wide sector (e.g. 90°) of ocean. The main directivity of the radar is obtained from the receiving antenna and its associated electronics. Since physical aerial movement is impracticable for large arrays at decametric wavelengths, this

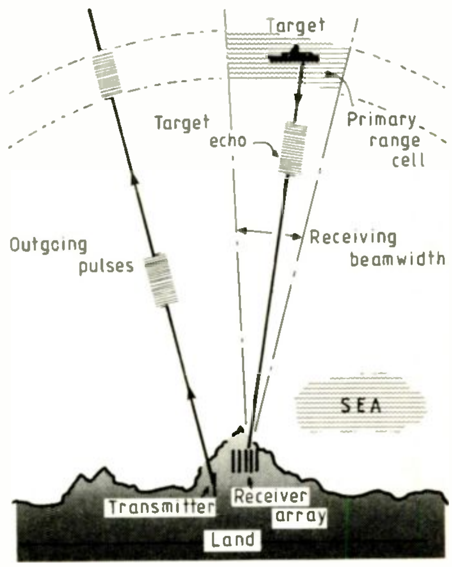
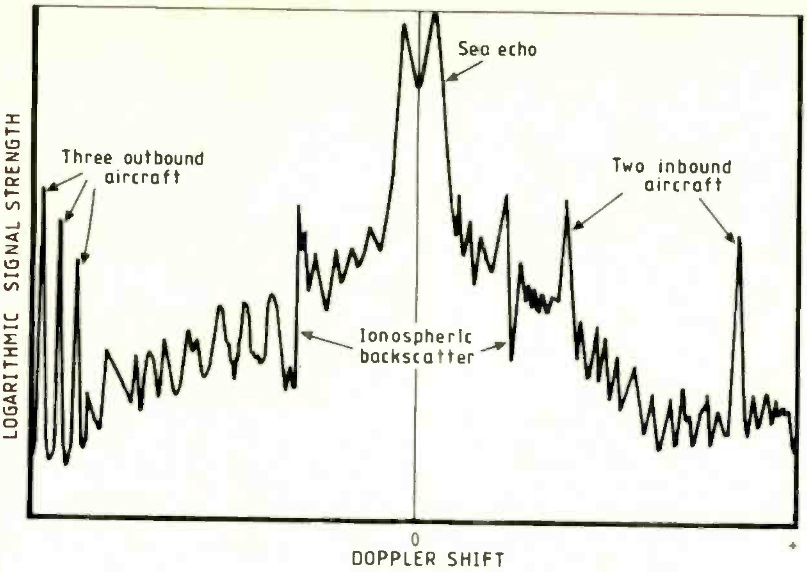


Fig.5 Essentials of a groundwave h.f. radar. The radar site is on the sea coast, so that decametric waves are refracted over the surface. Directivity is provided by a phased array capable of synthesising many narrow beams. Angular resolution can be improved by overlapping beam-widths by 50%. Adapted from ref. 4. Not to scale.

means using a phased array with many elements. Processing gives this the response equivalent to a number of beams each a few degrees wide. The receiving array is separate from the transmitting array (over 200dB of isolation between the two is needed[3]), but both may be on the same site.

What can this heroic engineering achieve? The attraction of ground-wave radar to the military mind lies in its promise of detecting ships and low-flying aircraft at ranges which are not usually indicated by researchers but which must, to be useful, run at least to hundreds of kilometres. Civil applications include remote sensing of rough sea conditions, long-distance monitoring of ship and aircraft traffic and perhaps the early detection of tidal waves.

The problem of detecting small targets such as aircraft and ships amounts to picking out special target-echo features from an intense background of sea clutter and noise.



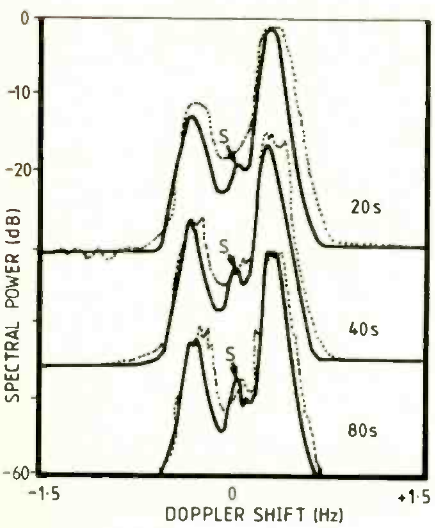


Fig.6 Doppler spectrum of signal returns from groundwave radar. The main sea response is central. Aircraft echoes show large Doppler shifts whose sense indicates the motion of the target towards or away from the radar. Adapted from ref.3.

Fig.7 Skywave radar Doppler spectra. Improved target detection after allowing for frequency modulation by changes in the F2 layer. Solid lines, corrected signals; dotted, uncorrected, for three integration times. The two large peaks at about ± 0.5Hz are the Bragg lines. The small peak (S) is the return from a ship. Adapted from ref.3.

Reflections from ocean waves have been much investigated. They show a small but characteristic Doppler shift, which takes the form of two unequally sized 'Bragg lines' surrounded by a continuum of lower-level Doppler noise. The Bragg lines may be less than 1Hz apart. Returns from aircraft show much greater Doppler shifts (Fig.6) but ships may produce very small shifts which lie between the Bragg lines in the thick of the noise.

In a pulse Doppler system, the need to limit the effective bandwidth (e.g.to 20kHz) calls for longish pulses (50µsec) and lowish repetition rates (a few kHz). To average out noise, integration times from a few seconds for aircraft up to a few hundred for ships are needed. Resolution is relatively poor; the pulse length limits the range resolution and the beamwidth the angular resolution. Clever processing may reduce the size of the primary range cell fixed by these quantities, but with aerials of moderate size it is good going to fix a target within a 10km square.

SKYWAVE RADAR

If no convenient ocean is available, or target areas are too distant, h.f. radar must rely for its range on reflections by the ionosphere. In practice, this may mean using one hop via the F2 layer. A paper from the Chinese People's Republic gives details of a practical system[6] capable of detecting aircraft at 700-1600km.

The transmitted pulses are frequency-modulated (linear sweep up to 30kHz), up to 3.5ms wide and with a repetition frequency down to 60Hz. Pulse compression based on Fast Fourier transforms is used at the receiver along with much other signal processing.

These measures ensure the *detection* of targets. But the great problem with skywave radar is to say just where your targets are. The ionosphere is notoriously variable. Under civil conditions fixed radar landmarks or transponders may be available to calibrate range but this may not be so in the military arena.

The Chinese system attempts to allow for the state of the ionosphere. A swept-frequency backscatter sounder (using the radar transmitting array on a time-shared basis) generates ionograms from which corrected range figures are calculated. Test runs illustrated in the paper show range errors of less than 4% (target located within about 30km). Old ionosphere hands are likely to have their doubts about calculated corrections, but the paper shows that the system can work, at any rate on a good day in China.

A paper from France[7] shows how both ionospheric and sea-surface-induced errors can be allowed for. The basic idea is that variations in the go and return paths produce a frequency modulation of the carrier. This appears as f.m. noise on the Doppler signal. By estimating the amount of f.m. and applying a correction, improved target detectability can be achieved. Target-induced shifts of a small fraction of 1Hz are detectable at long range (Fig.7).

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(Page numbers refer to IEE Conference Publication No 284: Fourth International Conference on HF Radio Systems and Techniques.)

1. G.A. Hagn: The Wide-Strip Horizontal Loop Antenna: an Effective Solution for Ground-Wave Interference to Shortwave Reception. (p. 145)
2. Villard, O.G., Jr: Portable Unidirectional HF Receiving Aerial for Reducing Co-Channel Multi-hop Sky-Wave Interference. (p. 141)
3. J. Bodonyi and T.W. Pegram: HF Groundwave Radar Theory and Practice. (p 241)
4. R.G.L. Evans, D.J. Bagwell and T.S. Maclean: Launching of HF Surface Waves. (p250)
5. A.M. Ponsford and D.J. Bagwell: Receiver Design for HF Groundwave Radar. (p254)
6. Jiao Peinan and Du Junhu: The Localization Technique for the Experimental HF Skywave Radar. (p259)
7. J. Parent and A. Bourdillon: Enhancement of Remote Sensing Capability of HF Skywave Radar by Correcting the Signal for Frequency Modulation. (p268)

Summer school in broadcasting

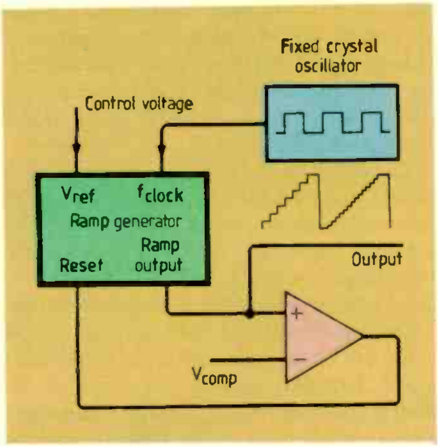
If you move fast, you might still be in time to take advantage of a week's crash course in new broadcasting technologies, organized by the IEE. Listed in the provisional programme are sessions on high-definition television, component coding, digital audio and video, MAC, v.t.r. standards, satellite and terrestrial broadcasting, RDS (Radio Data System), measurement techniques, the EBU/SMPTE control system, and stereo sound for tv. Among the speakers are well-known names from manufacturing industry and from the broadcasters' engineering departments.

The vacation school, to be held at the University of Southampton, runs from midday on 18 July to 2 p.m. on 22 July. Residential accommodation is available at the University. Cost is £355, including meals, accommodation, an evening social function and v.a.t; non-residential participants pay £250 (£50 less for IEE and IERE members in each case). Two bursaries are offered to IEE members who work for small companies or in the academic field and cannot obtain alternative finance.

Further details and a booking form are available from Lorna Richardson at the IEE, Savoy Place, London WC2R 0BL (tel. 01-240 1871 ext. 330).

CIRCUIT IDEAS

Precision digital ramp generator using switches



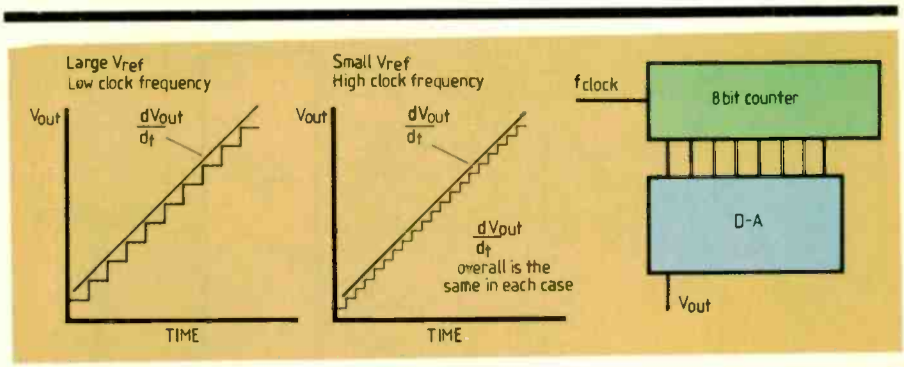
Unlike ramp generators made from digital-to-analogue converters, this circuit allows accuracy of the ramp to be varied. With a small reference voltage and high clock frequency, a staircase with very small steps results. Increasing reference voltage and decreasing clock frequency cause the steps to become larger. Digital-to-analogue converters, connected as shown right of the graphs, produce a fixed number of steps.

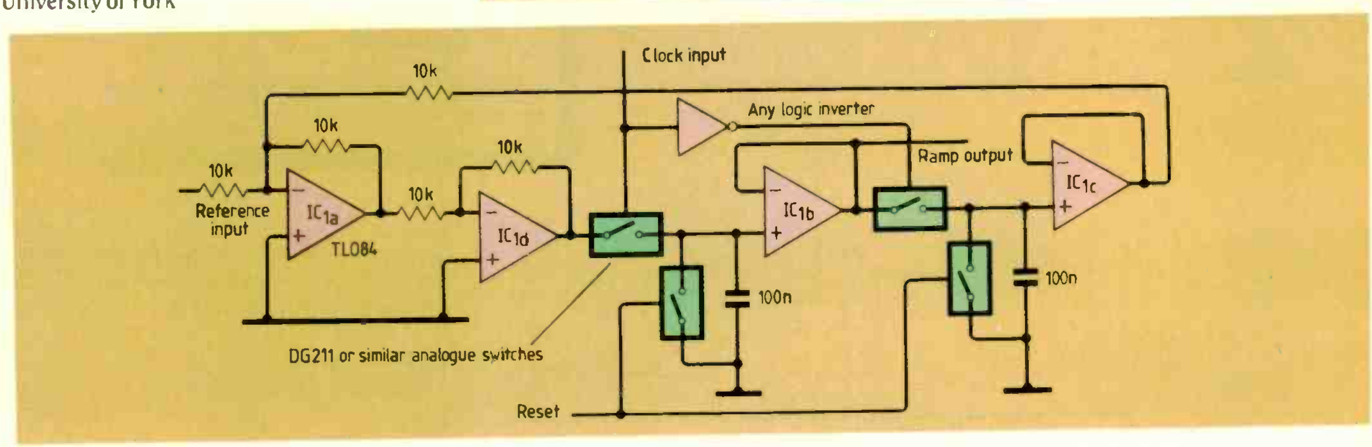
On each clock pulse, output increases by V_{REF} so voltage at time t is thus $V_{REF} \times t \times f_{CK}$. Op-amp IC_{1a} forms an adder and $IC_{1b,c}$ form two sample-and-hold circuits. Output of IC_{1b} is the running total and thus final output.

Originally, the circuit was used as a sweep oscillator running under microcomputer control. Providing a crystal-based clock oscillator makes the circuit suitable for the basis of a high-stability v.c.o. in which a comparator would provide a reset pulse when the ramp reaches a specific level. Output frequency is,

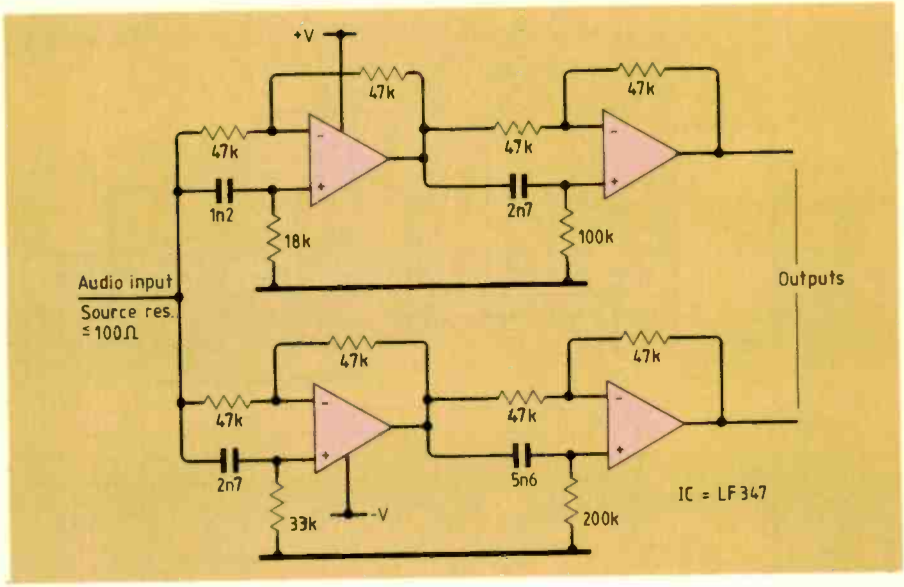
$$\frac{V_{REF} \times f_{CK}}{V_{COMP}}$$

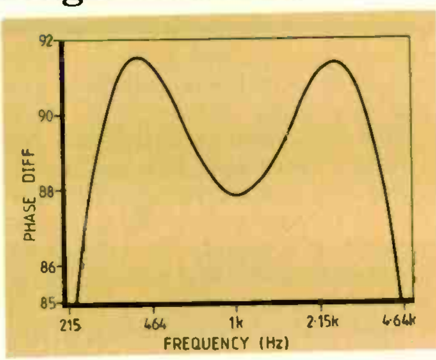
Peregrine Andrews
University of York





Phase shifter for single sideband



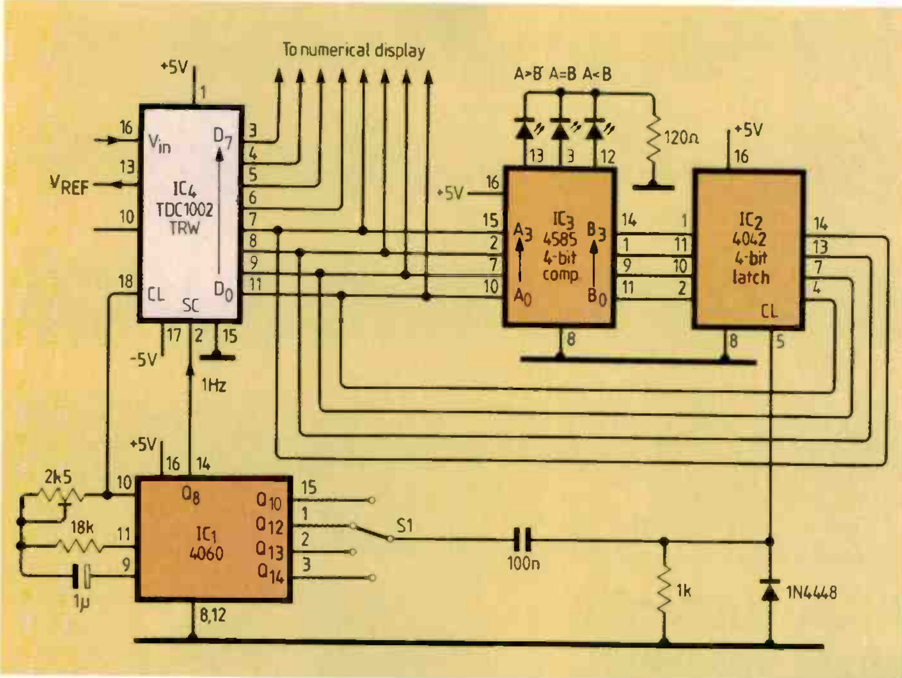


Using the second, or 'out-phasing', method of s.s.b generation eliminates the need for steep filters for reducing the unwanted sideband but requires an audio signal with a constant phase shift of 90° across the band. The two ouputs of this circuit are in quadrature within 3° between 270Hz and 4kHz, which could theoretically produce up to 25dB rejection of the unwanted sideband.

Capacitor values are critical and I recommend Mullard 1% polystyrene 425 types; these should not be subjected to defluxing agents. The resistor values should also be within 1%.

J.R. Charlesworth
Kirkbymoorside
North Yorkshire

CIRCUIT IDEAS



Digital tendency indicator

Three leds and three c-mos i.cs connected to the output of an analogue-to-digital converter provide an indication of the direction of change in an electrical signal representing, say, temperature. Eight-bit binary data from the converter is also available for decoding and display driving.

Converter output bits D_{0-3} feed one set of inputs to a four-bit comparator. The other four comparator inputs are fed with the previous converter reading held in a latch. Comparator outputs A=B, A>B and A<B drive leds to indicate the direction of change.

Display updating occurs once a second since the converter is driven from a 256Hz clock. Oscillator-divider IC_1 provides the 256Hz converter clock, the start-conversion signal and the latch pulse, which can be set to latch data for 4, 16, 32 or 64 seconds. Clock output and the latch pulse can be altered to suit the application.

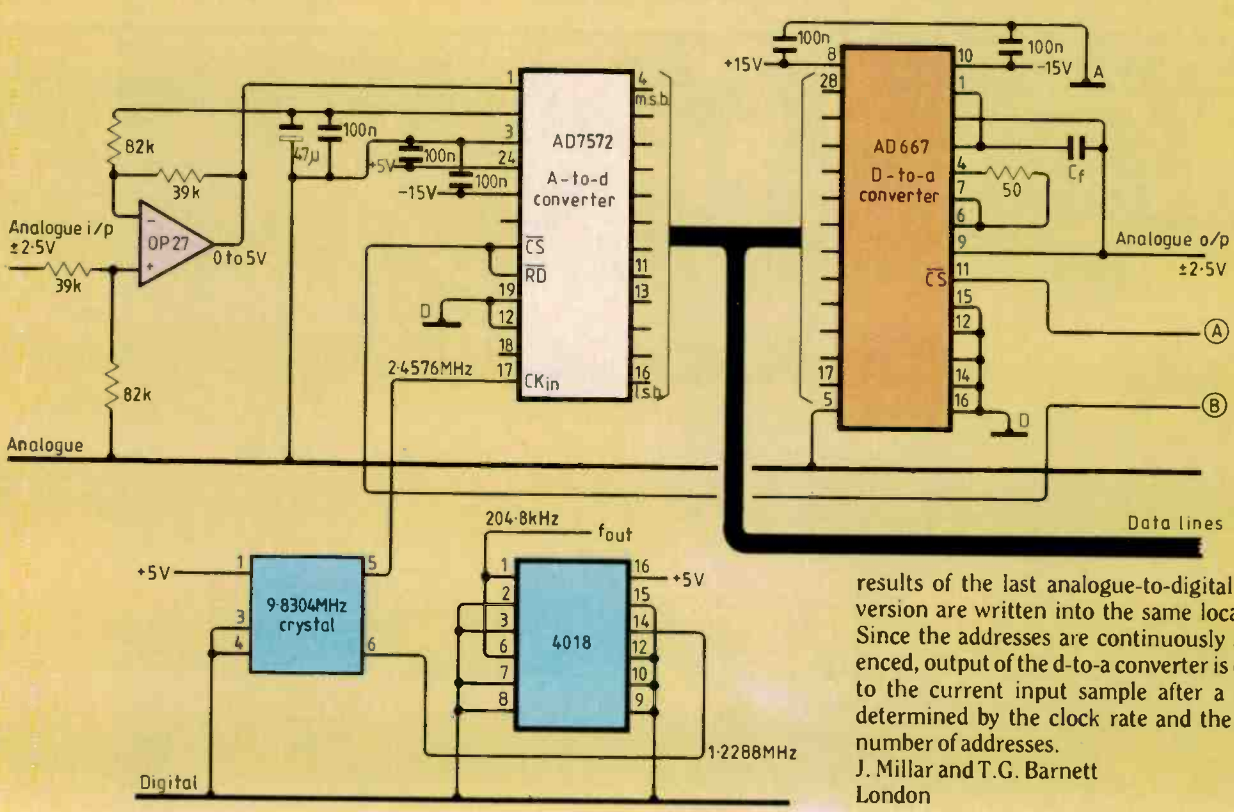
Analogue delay using a 12bit a-to-d converter

Delay lines in which an analogue signal is digitized and stored for later retrieval and conversion back to analogue form are widely used. Most published circuits though are either slow or use low-resolution a-to-d and d-to-a converters.

With a 5µs conversion time and 12bit resolution, the AD7572 analogue-to-digital converter is both fast and accurate. Shown here is a circuit that can delay analogue signals from 20ms to 40.96s in binary steps at bandwidths of 10kHz and 5Hz respectively.

Input is taken through an op-amp to the a-to-d converter's analogue input. A crystal oscillator provides the 2.4576MHz required to clock the converter and, after division through counters, the memory address lines. Monostable i.cs, synchronized by pulses from the oscillator circuit, produce chip-select, read and output-enable signals for the a-do-d/d-to-a converters and ram.

The rams are accessed sequentially. At each address, the contents are read and the



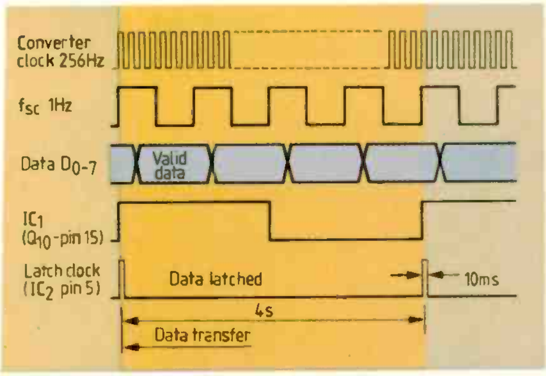
results of the last analogue-to-digital conversion are written into the same location. Since the addresses are continuously sequenced, output of the d-to-a converter is equal to the current input sample after a delay determined by the clock rate and the total number of addresses.

J. Millar and T.G. Barnett
London

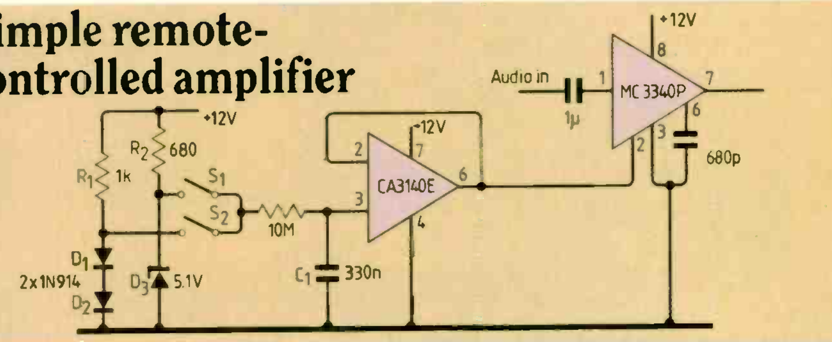
CIRCUIT IDEAS

To prevent undesirable effects from noise and small voltage fluctuations, the comparator inputs are simply moved up one or two bits, i.e. connected to D_{1-4} or D_{2-5}. Replacing the comparator with an a.l.u. for subtraction, and connecting its output to an appropriate numerical driver will give a temperature gradient reading.

Petrovic Tomislav
Nis, Yugoslavia



Simple remote-controlled amplifier

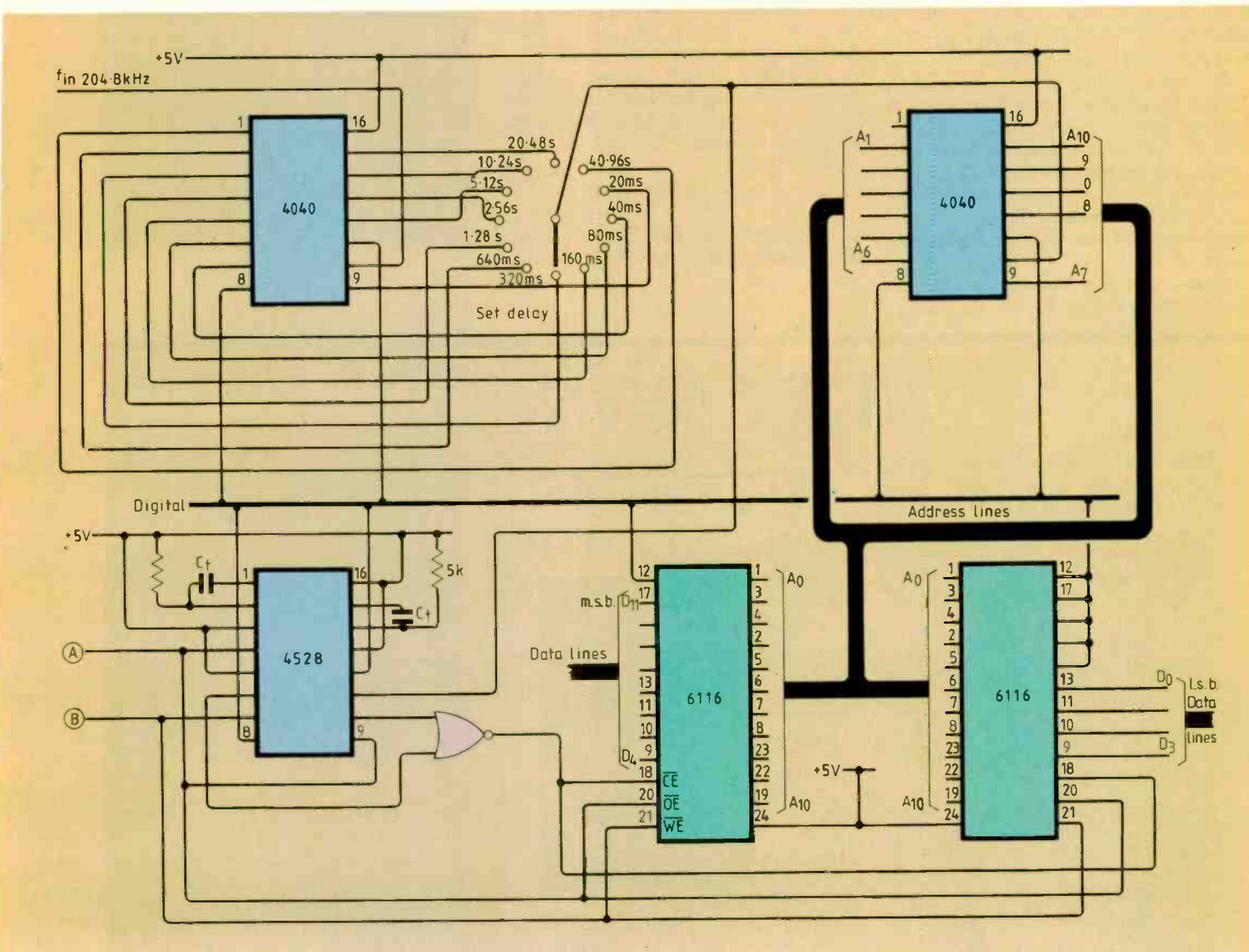


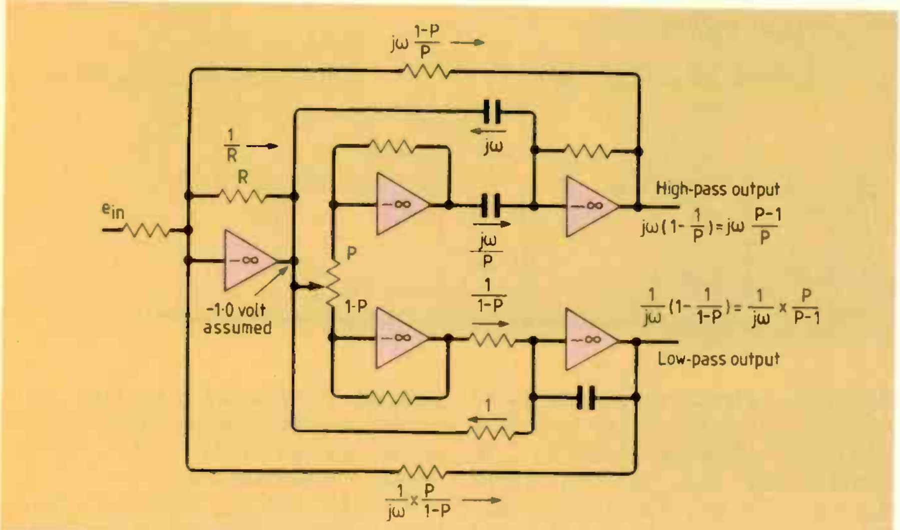
This is a cheap alternative to the usual digital circuitry required for remote control. It relies on the voltage across capacitor C_1 to control the gain of an MC3340 electronic attenuator. The capacitor, a low leakage type, must be buffered by a mosfet-input op-amp such as the CA3140 to ensure a very long discharge time.

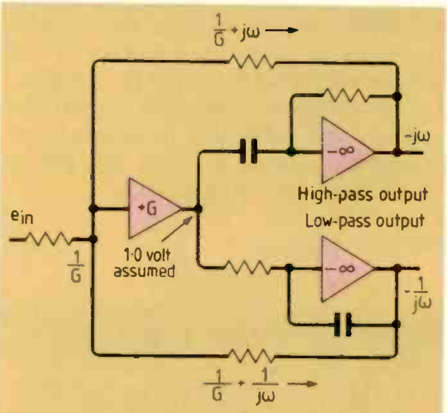
Closing S_1 causes C_1 to charge, resulting in a decrease in volume, while closing S_2 increases volume. Both switches should preferably be miniature relays since semiconductor switches like the 4016 allow small currents to flow in the off state, thereby altering the capacitor voltage.

Diodes D_{1-3} with $R_{1,2}$ set upper and lower limits of the capacitor voltage range for a 12V supply. With metallized polyester-film capacitors we observed a 90mV fall over a 13 hour period; over an hour, less than 2mV.

C. Anyanwu and J. Agada
University of Nigeria







Crossover filter

These two circuits are combined high and low-pass filters useful as low-level crossover networks. In the first circuit, right, all values except gain G are normalized to unity. Output voltage of this amplifier is taken to be 1V and transmissions T_{LP} T_{HP} are the ratios of output to input.

As shown, with the proper choice of G, T_{LP} and T_{HP} are Butterworth functions with common cut-off frequency $\omega=1.0$. This is the usual choice for crossover networks. Crossover frequency can be scaled by simultaneously varhing either both capacitors or both of their associated resistors so the network can be tuned with a duel potentiometer.

Tuning can also be accomplished by changing the gains in the two channels in such a way that their product is a constant of, say 1.0. How a single potentiometer can be connected so that the resistance from wiper to one end controls one gain and resistance from wiper to the other end the other gain is shown in the second diagram. Overall resistance of this potentiometer is 1.0, with P being the setting and resistor R playing much the same role as gain G.

As is evident, (P-1)/1 is simply a multiplier on ω showing that the potentiometer controls the crossover frequency in the proper way. Contant,

$$e_{in}=\frac{1}{R}+j\left(\omega-\frac{1}{\omega}\right)$$

so following the analysis in the first diagram, outputs are maximally flat when,

$$R=\frac{1}{\sqrt{2}}$$

NEXT MONTH

Digital signal processing. Alan Sewards raises and answers some of the questions that commonly occur to engineers facing d.s.p. for the first time. The aspect covered in this case is the sampling of an analogue signal, converting it by a-to-d and performing a fast Fourier transform.

Smith's chart. Philip Smith wrote of his Transmission-line Calculator in *Electronics* in 1939. It rapidly became known as simply the Smith Chart and has been in widespread use by r.f. engineers ever since. Joules Watt describes the background and application.

MosMarx. P.E.K. Donaldson, of the Medical Research Council, describes an alternative to the Cockroft-Walton voltage multiplier – the Marx generator. This modern variant incorporates mosfets and Schottky diodes in place of spark gaps and resistors and is able to provide a d.c.-to-d.c. conversion efficiency of nearly 90%. As an illustration, a voltage quintupler to give 45V at 250mA is shown.

Measuring by ultrasound. It isn't very often that worthwhile development comes about as a result of a "blinding flash", it is more often the outcome of a long process of development work. R.J. Redding describes the development of a technique for ultrasonic flow measurement.

Pioneers – Faraday. Michael Faraday, a formidable and totally practical investigator, has been called the "plain man's scientist". Tony Atherton continues his series on the pioneers of communications with a look at one of the men who started it all.

Filter design using a microcomputer

Steps required in the design of active and passive Butterworth and Chebyshev filters

S. NIEWIADOMSKI

The modern approach to filter design, developed in the 1950s, greatly simplified the process of producing practical filters, making it possible for non-experts to design complex filter circuits. Briefly, the process depends on the use of pre-calculated catalogues of normalized component values for low-pass filters, which are then scaled to the desired frequency and impedance level. Filter responses other than low-pass can be implemented by applying a transformation to the low-pass prototype and then performing the scaling process. Over the years many books (see the bibliography) and articles have been published on this approach to filter design, and the method has been very successful. There are, however, three drawbacks to the catalogue method, namely:

- catalogues can only document a finite number of combinations of filter parameters, such as order and passband ripple, and so inevitably some compromise is required to match the filter specification with the designs available in the catalogue;
- catalogues tend to be published in expensive books which are not easily accessible;
- manual searching through catalogues for a suitable design, the extraction of data and scaling to the final frequency and impedance level are not entirely satisfactory processes today.

The underlying design equations which enabled the catalogues to be produced required formidable computing power thirty years ago, but can easily be tackled by even the cheapest personal computer today. It seems logical then, that these equations are now made available to enable programming into personal computers, liberating us from the restrictions of the catalogue method.

This article describes the steps required to design passive and active Butterworth and Chebyshev filters and also gives listings of programs which perform the necessary calculations. It would be possible to write a single program to handle the whole of Butterworth and Chebyshev filter design: the program could be menu-driven, taking the user through all the choices available at each step, and turning a specification into a final practical filter design. The drawback with this approach is that not all users will want such a comprehensive program and may require only a small part. The onus is therefore put on those who want a comprehensive package to merge together the individual programs presented here.

The programs are written in Basic for the Sinclair Spectrum computer, although the author recognises that many other types are in use. To aid the conversion of programs to other computers or languages, statements are avoided which might cause problems. Similarly, the graphical display of results, which again can make conversion difficult, is avoided. Details of the specific dialect of Basic implemented on a particular machine will be found in the language manual accompanying it. Some guidance on converting Basic programs to run on different machines can be found in Reference 1.

FILTER ORDER ESTIMATION

The first step when designing a filter to meet a given specification, whether a passive or active implementation is to be used, is to determine the order of filter required. Traditionally, nomographs (Reference 2) have been used which effectively transform the algebraic process of solving an equation into the geometric process of drawing lines on a graph. Clearly, the algebraic process is the more suitable for implementation on a computer. The nomograph method can also be inaccurate and in some instances results in the use of a more complex filter than strictly necessary.

With a Butterworth filter, only the stopband filtering requirement needs to be specified to enable the order to be calculated, as all Butterworth filters have an identical passband specification. Equation 1 calculates this order.

$$n=\frac{\log_{10}\sqrt{10^{A/10}-1}}{2\log_{10}\left(\frac{\omega}{\omega_c}\right)} \qquad (1)$$

where n is the order of the Butterworth filter,
A is the stopband attenuation in dB,
ω_c is the cut-off frequency,
ω is the frequency at which the attenuation is specified.

Estimating the order of a Chebyshev filter is slightly more complex because a new variable, the passband ripple, has been introduced. Equation (2) shows how the order can be calculated from the passband ripple and the stopband performance.

$$n=\frac{\cosh^{-1}\sqrt{\frac{10^{A/10}-1}{\epsilon^2}}}{\cosh^{-1}\left(\frac{\omega}{\omega_c}\right)} \qquad (2)$$

where n is the order of the Chebyshev filter,
A is the stopband attenuation in dB,
ω_c is the cut-off frequency,
ω is the frequency at which the attenuation is specified,
ε is the ripple factor.

Since inverse hyperbolic functions (such as $\cosh^{-1}x$) are not usually available on home computers, a useful relationship is:

$$\cosh^{-1}x=\ln(x+\sqrt{x^2-1}) \qquad (3)$$

In equation (2), the passband ripple is expressed in terms of the ripple factor, whereas a more common format is in terms of the maximum value of the ripple in dB. Equation (4) shows how the ripple factor can be calculated from the ripple in dB for use in Equation 2.

$$\epsilon=\sqrt{10^{Ap/10}-1} \qquad (4)$$

where Ap is the passband ripple in dB.

Program 1 (ORDER) implements these equations. If the answer "b" is given after the

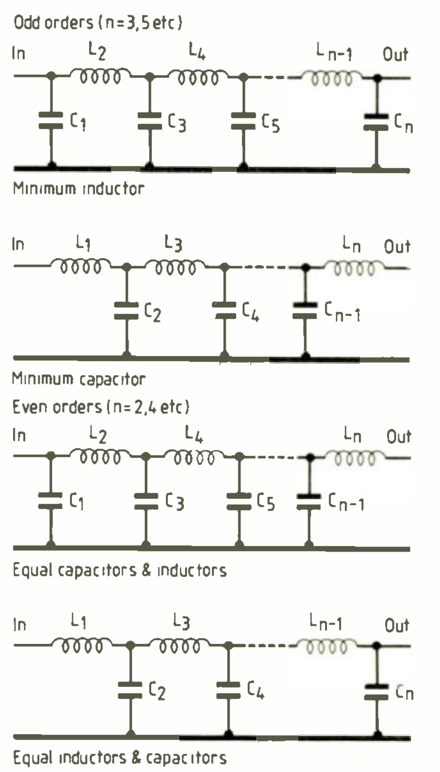


Fig.1. Passive Butterworth low-pass filter implementations

prompt for the filter type, the program works out the order of a Butterworth filter after further prompting for the appropriate values. The cut-off frequency and the frequency at which the stop-band attenuation is specified can be in rad/s or Hz, as long as they are both in the same units. The Butterworth section of the program takes advantage of the fact that the value of ratio of the logarithms of two numbers is independent of the base of the logarithms. Therefore, although the logarithms are taken to base 10 in equation (1), the Spectrum program uses logarithms to base e, which are all that are available in Spectrum Basic.

If the answer "c" is given after the prompt for the filter type, the program works out the order of a Chebyshev filter from equation (2). One additional value for the filter specification, the passband ripple, is required in the Chebyshev case. As for the Butterworth filter, the cut-off frequency and the frequency at which the stopband attenuation is specified can be in rad/s or Hz.

As well as printing the exact value of filter order needed to satisfy a specification, the program also rounds the order upwards and outputs this value. The orders estimated are equally valid for passive and active implementations of the filters.

PASSIVE LOW-PASS BUTTERWORTH FILTER DESIGN

Figure 1 shows the possible configurations for passive Butterworth low-pass filters. For odd orders, there is a minimum-inductor and a minimum-capacitor implementation which are duals of each other: that is, the responses of the duals are identical in all respects. When a low-pass filter is required, the minimum-inductor version is usually chosen, but if the low-pass filter is merely a prototype for transformation to a filter with a different response, the minimum-capacitor version can lead to a better transformed filter. Even-order filters have an equal number of capacitors and inductors, and therefore there is often little to choose between the duals.

Calculation of component values for passive Butterworth filters with equal source and termination impedances is fairly simple and equation (5) is presented here to enable this to be performed if required (reference 3).

$$COMP_k = 2\sin\frac{(2k-1)\pi}{2n} \text{ for } k=1,2\ldots n \quad (5)$$

where $COMP_k$ is the Kth component value, being either a shunt capacitor (in Farads) or a series inductor (in Henries) of the circuits shown in Fig. 1, and n is the order of the filter.

If filters with unequal source and termination impedances are required, the calculations are not so simple: this section describes the design of such filters and gives a program to perform the calculations. In fact, equation (5) can be ignored as the more general method can be used with equally terminated filters by entering 1 ohm as the terminating impedance.

Figure 2 shows two implementations of passive Butterworth low-pass filters: note

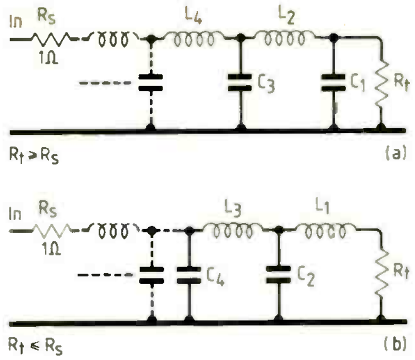


Fig.2 Passive low-pass Butterworth filter implementations showing the component designations produced by the general design method.

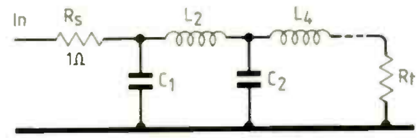


Fig.3 Passive low-pass Chebyshev filter circuit diagram showing component designations produced by the design method.

that in these circuits the filter components are not numbered in the conventional way, that is starting at the input and working towards the output. This is a peculiarity of the method of design described in this section. Circuit (a) has a shunt capacitor, C_1, closest to the output, whereas circuit (b) has a series inductor, L_1, closest to the output. Both circuits have source impedances of 1 ohm. With filters having equal source and termination impedances, the circuits are duals of each other and either implementation can be used. When a filter with unequal source and termination impedances is required, only one implementation is suitable: circuit (a) can be used where the termination impedance is greater than or equal to the source impedance, and circuit (b) can be used where the termination impedance is less than or equal to the source impedance.

The first step in the calculation of the component values (reference 4) is to determine λ

$$\lambda=\left(\frac{R_t-1}{R_t+1}\right)^{1/n} \text{ for } R_t\geq R_s \quad (6)$$

or

$$\lambda=\left(\frac{1-R_t}{1+R_t}\right)^{1/n} \text{ for } R_t\leq R_s \quad (7)$$

where R_t is the termination impedance, R_s is the source impedance (1 ohm), and n is the order of the filter.

The value of the first component can now be calculated from the expressions

$$C_1=\frac{\alpha_i}{R_t(1-\lambda)} \text{ for } R_t\geq R_s \quad (8)$$

$$L_1=\frac{\alpha_i R_t}{(1-\lambda)} \text{ for } R_t\leq R_s \quad (9)$$

where α_i is obtained by substituting i=1 into the expression:

$$\alpha_i=2\sin\frac{\pi i}{2n} \quad (10)$$

The remaining component values are now calculated by stepping m from 1 to (n-1)/2 when n is odd, or to n/2 when n is even, in the following equations.

$$C_{2m-1}L_{2m}=\frac{\alpha_{4m-3}\alpha_{4m-1}}{1-\lambda\beta_{4m-2}+\lambda^2} \quad (11)$$

and

$$C_{2m+1}L_{2m}=\frac{\alpha_{4m-1}\alpha_{4m+1}}{1-\lambda\beta_{4m}+\lambda^2} \quad (12)$$

for $R_t\geq R_s$, or

$$L_{2m-1}C_{2m}=\frac{\alpha_{4m-3}\alpha_{4m-1}}{1-\lambda\beta_{4m-2}+\lambda^2} \quad (13)$$

and

$$L_{2m+1}C_{2m}=\frac{\alpha_{4m-1}\alpha_{4m+1}}{1-\lambda\beta_{4m}+\lambda^2} \quad (14)$$

for $R_t\leq R_s$, where α_i is given by equation (10) and β_i is given by

$$\beta_i=2\cos\frac{\pi i}{2n} \quad (15)$$

Program 2 (BUTPASS) implements these equations. A major portion of the program is concerned with determining whether a particular component is a capacitor or an inductor and printing out the normalized component value preceded by a component designation. If the component designations are not required, the program can be simplified. The designations produced by BUTPASS correspond to Fig.2: that is, the value of the component closest to the output is generated first and is either shunt capacitor C_1 or series inductor L_1. If an equal source and termination impedance filter is specified, the print out indicates that each component value refers to either a shunt capacitor or a series inductor.

PASSIVE LOW-PASS CHEBYSHEV FILTER DESIGN

The design of passive Chebyshev filters (reference 5) is similar to the Butterworth case, but is slightly more complex because of the incorporation of an extra variable, the passband ripple.

The first step is to calculate the quantity a, from the termination impedance and ripple factor, using the expression:

$$a=\frac{4R_t}{(R_t+1)^2} \text{ when n is odd} \quad (16)$$

or:

$$a=\frac{4R_t}{(R_t+1)^2}[1+\epsilon^2] \text{ when n is even} \quad (17)$$

where R_t is the termination impedance, ε is the ripple factor, and n is the order of the filter.

In the case where the filter order is even, a test must be carried out on the value of a, and the specification can only be implemented by this method if a is less than or equal to 1.

The by letting

$$\alpha_i=2\sin\frac{\pi i}{2n} \quad (18)$$

$$\beta_i = 2\cos\frac{\pi i}{2n} \qquad (19)$$

$$\gamma = \left[\frac{1}{\epsilon} + \sqrt{\frac{1}{\epsilon_2} + 1}\right]^{1/n} \qquad (20)$$

$$\delta = \left[\sqrt{\frac{1-a}{\epsilon^2}} + \sqrt{\frac{1-a}{\epsilon^2} + 1}\right]^{1/n} \qquad (21)$$

$$x = \gamma - \frac{1}{\gamma} \qquad (22)$$

and

$$y = \delta - \frac{1}{\delta} \qquad (23)$$

The value of the first capacitor (the component closest to the input of the filter in this case) can be calculated from the expression:

$$C_1 = \frac{2\alpha_1}{x-y} \qquad (24)$$

The remaining component values are now calculated by stepping m from 1 to (n-1)/2 for n odd, or to n/2 for n even, in the following equations. Figure 3 shows the component designations to which the results apply.

$$C_{2m-1}L_{2m} = \frac{4\alpha_{4m-3}\alpha_{4m-1}}{b_{2m}(x,y)} \qquad (25)$$

and

$$C_{2m+1}L_{2m} = \frac{4\alpha_{4m-1}\alpha_{4m+1}}{b_{2m}(x,y)} \qquad (26)$$

where the function $b_i(x,y)$ is defined by

$$b_i(x,y) = x^2 - \beta_{2i}xy + y^2 + \alpha^2_{2i} \qquad (27)$$

Program 3 (CHEBPASS) performs these calculations, after prompting for the appropriate values. The program gives component designations as in Fig. 3.

ACTIVE LOW-PASS FILTER DESIGN

Nowadays, active filters, using op-amps as the active elements, tend to be more popular than passive filters, especially for audio frequency applications. Whether a Butterworth or a Chebyshev (or Bessel, for that matter) response is required, a similar approach to the design of active filters can be adopted. The method relies on the fact that active sections with first and second responses can be cascaded to form filters of any complexity. Because of the inherent isolation of output from input of an op-amp, an active section retains its individual response even when cascaded with other sections. Figure 4 shows how active filters of orders from 1 to 4 can be built from first and second-order sections. Higher-order filters are implemented by simply adding further sections. The required overall response for the filter is obtained by selecting the correct design parameters for each of the sections.

The simplest implementation of a first-order section is an RC network. By adding an op-amp to the output of the network, a section which can be cascaded without loading effects is obtained. This type of first-order section is ideal for use in multi-section active filters. Figure 5 shows two possible implementations of first order active sections. Figure 5(a) is a unity gain (at d.c.) section whereas the circuit in Fig. 5(b) can have its d.c. gain set to any reasonable value by adjusting the ratio of R_2 to R_3. The component values shown are for a cut-off frequency normalized to 1 rad/s. All that is required to change this cut-off frequency is to divide the capacitor value by a factor ωφ, so that for values of ωφ greater than 1, the value of C_1 is decreased and so the cut-off frequency of the section is raised, and conversely when ωφ is less than 1 the cut-off frequency is lowered.

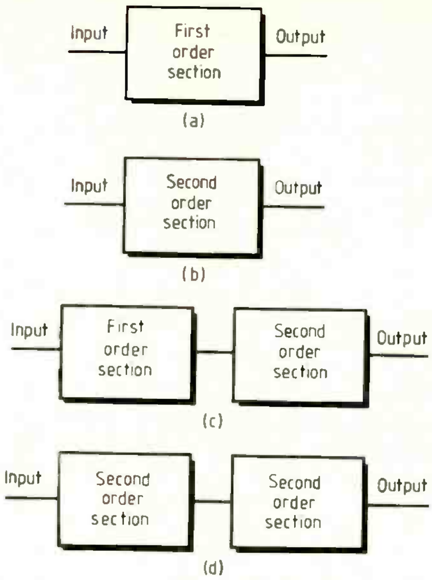


Fig.4 Active low-pass filters are built up by cascading first and second order sections. Adjustment of the parameters of the individual sections gives the desired overall response. At (a) is a first-order filter, (b) is a second-order filter, (c) is a third-order filter and (d) is a fourth order filter

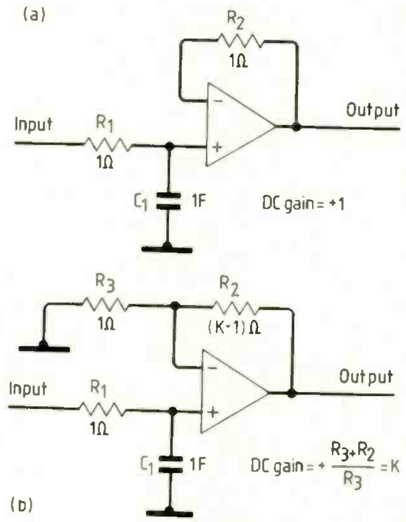


Fig.5 Simple normalized, first-order, low-pass active filter sections, where (a) has a d.c. gain value of 1, and (b) has a d.c. gain of $\frac{R_3+R_2}{R_3}$

Many different implementations for second-order sections have been developed over the years, each optimized for a particular aspect of performance. Typical parameters which can influence which type of section is used include the gain (if any) required from the section, the performance demanded for the op-amp for a given section specification, the ease of tuning the section to an exact cut-off frequency if required and the sensitivity of the section to small deviations of the component values which are inevitable when a section is built using real components. Figure 6(a) shows a normalized unity-gain Sallen and Key second-order section. The expressions for the values of C_1 and C_2 can be seen to contain the variable d, which is the reciprocal of the section Q. Figure 6(b) shows a Sallen and Key second-order section with equal capacitor values. With this circuit, the section gain cannot be set independently, but depends on the value of d chosen.

Figure 7 shows how the value of d affects the amplitude response close to cut-off of a unity gain section which has been scaled to 1 kHz. Lower values of d, and hence higher section Q, result in a greater peak before the response starts to tail off and tend towards infinite attenuation. With d=1.414, the attenuation at 1 kHz is 3.01dB, giving a Butterworth response: when d=1.045, 0.886 and 0.766, Chebyshev responses with peak gains of 1, 2 and 3dB respectively are obtained.

The amplitude response of a normalized second-order section can be moved up and down the frequency axis by dividing the capacitor values by the quantity ωφ, in the same way that the cut-off frequency of a first-order section can be varied. This allows the 3.01dB frequency (for Butterworth sections) or the peak gain frequency (for Chebyshev sections) to be set to suit the overall requirement of any filter into which they fit. To summarize, by choosing the appropriate values for d and ωφ, the cut-off frequency of a second-order section can be altered, and Butterworth or Chebyshev type responses can be obtained.

Figure 8 shows a different implementation of a second-order section. This circuit is a normalized, unity-gain, state-variable section (reference 6) which has some advantages over the rather simpler Sallen and Key circuits. One interesting feature is that it has high-pass and band-pass outputs as well as its low-pass output, which could be useful in some applications. Figure 9 shows a normalized state-variable section with gain. This circuit allows the section gain to be set independently of the value of d, which was not the case with the Sallen and Key circuits. These three and four op-amp sections do not necessarily have to cost a great deal more or occupy much more board area than the single op-amp circuits. Quad op-amp packages are cheaply available and allow easy implementation of these circuits.

Design of an active filter therefore consists of selecting a value for ωφ for the first-order section (if the overall filter order is odd), and d and ωφ for each of the second-order sections. Gain can be obtained

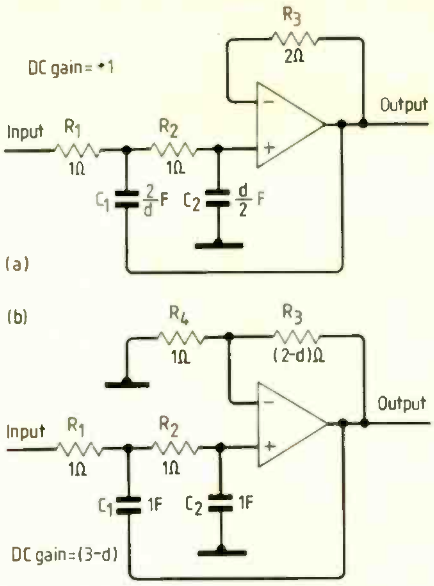


Fig.6 Second-order Sallen and Key filter sections. At (a) is a unity-gain section and at (b) an equal-component-value section.

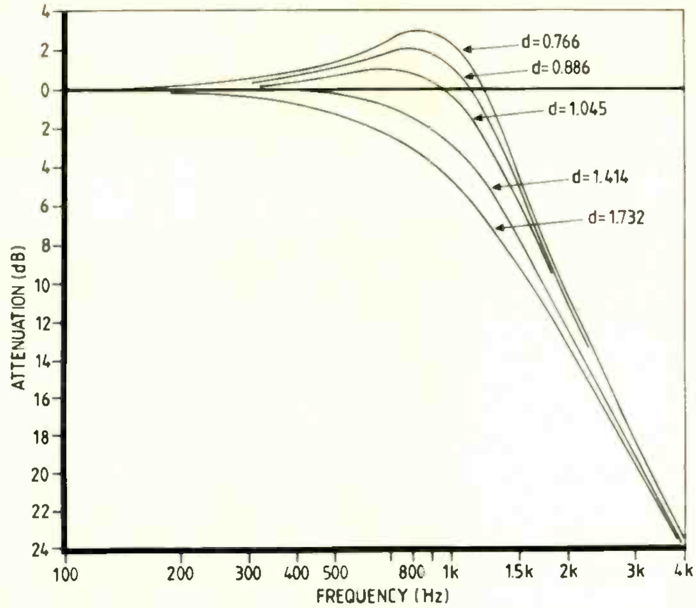


Fig.7 Amplitude response of a unity-gain, second-order, low-pass active section for various values of d. The response has been scaled to 1kHz.

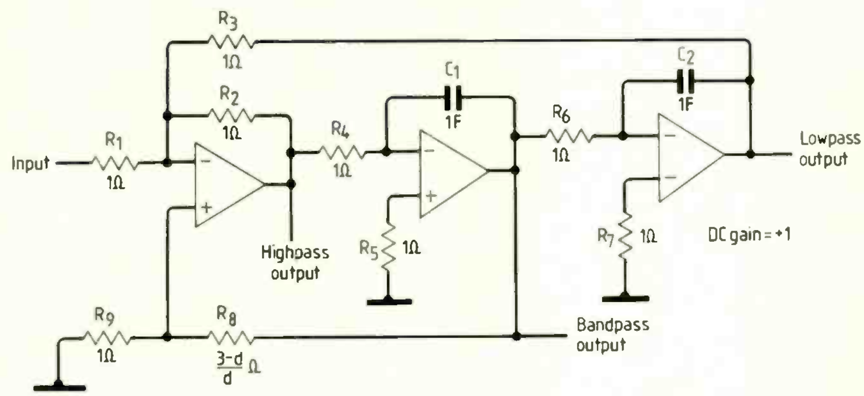


Fig.8 The normalized, unity-gain, second-order, state-variable active section. Note that this section has high-pass and band-pass outputs, as well as a low-pass output.

if required by selecting the appropriate implementations of the sections. In the same way that catalogues of normalized component values for passive filters are available, there are also catalogues of d and $\omega\phi$ for active filters (reference 7). Some catalogues (reference 8) express this information in terms of the roots of the equations describing the responses in the complex plane, which can easily be converted into the d, $\omega\phi$ format. The methods of determining values of d and $\omega\phi$ for Butterworth and Chebyshev active filters will now be described.

ACTIVE LOW-PASS BUTTERWORTH FILTER DESIGN

The first step in obtaining the values of d and $\omega\phi$ for each of the sections in a Butterworth active filter is to determine the locations of the poles in the complex plane for the required response. Only the order of the filter needs to be specified to calculate these pole positions, using equations (27), (28) and (29) (reference 9).

$$\alpha_k = -\sin\theta_k \tag{28}$$

and

$$\beta_k = \cos\theta_k \tag{29}$$

where

$$\theta_k = \frac{(2k-1)\pi}{2n} \text{ for } k=1 \text{ to } (n+1)/2 \tag{30}$$

For even values of n, there will be n/2 pairs of poles which are complex conjugates of each other. For odd values of n, there will be one purely real pole (with d=2) and (n-1)/2 pairs of complex conjugates poles.

The complex pole format can now be converted into the d, $\omega\phi$ format by using equations (31) and (32).

$$\omega 0 = \sqrt{\alpha_k^{\,2} + \beta_k^{\,2}} \tag{31}$$

$$d = \frac{-2\alpha_k}{\omega 0} \tag{32}$$

Program 4 (BUTACT) implements these equations, prompting for the filter order, and printing the values of d and $\omega\phi$ for each first and second-order section in the filter. In general, it is best to place the first-order section (if the filter order is odd) closest to the input and the second-order sections in order of descending d. This ensures that the greater peak gains obtained from higher-Q sections are less likely to result in overloading of subsequent sections.

ACTIVE LOW-PASS CHEBYSHEV FILTER DESIGN

The first stage in calculating d and $\omega\phi$ values for Chebyshev filters (reference 10) is to determine the quantity a using the formula:

$$a = \frac{1}{n}\sinh\frac{1}{\epsilon} \tag{33}$$

where n is the filter order, and ϵ is the ripple factor.

A useful formula for determining the inverse hyperbolic sine function on a personal computer is:

$$\sinh^{-1}x = \ln(x + \sqrt{x^2+1}) \tag{34}$$

The locations of the poles in the complex plane for the response are given by

$$\alpha_k = -\sin\theta_k \sinh a \tag{35}$$

and

$$\beta_k = \cos\theta_k \cosh a \tag{36}$$

where

$$\theta_k = \frac{(2k-1)\pi}{2n} \tag{37}$$

for k=1 to n

For even values of n, there will be n/2 pairs of poles which are complex conjugates of each other. For odd values of n, there will be one purely real pole (with d=2) and (n-1)/2 pairs of complex conjugates poles.

sinh a and cosh a are calculated from the relationship

$$\sinh a = \frac{e^a - e^{-a}}{2} \tag{38}$$

and

$$\cosh a = \frac{e^a + e^{-a}}{2} \tag{39}$$

The complex pole format can again be converted into the d, $\omega\phi$ format by using equations (31) and (32).

Program 5 (CHEBACT) implements these

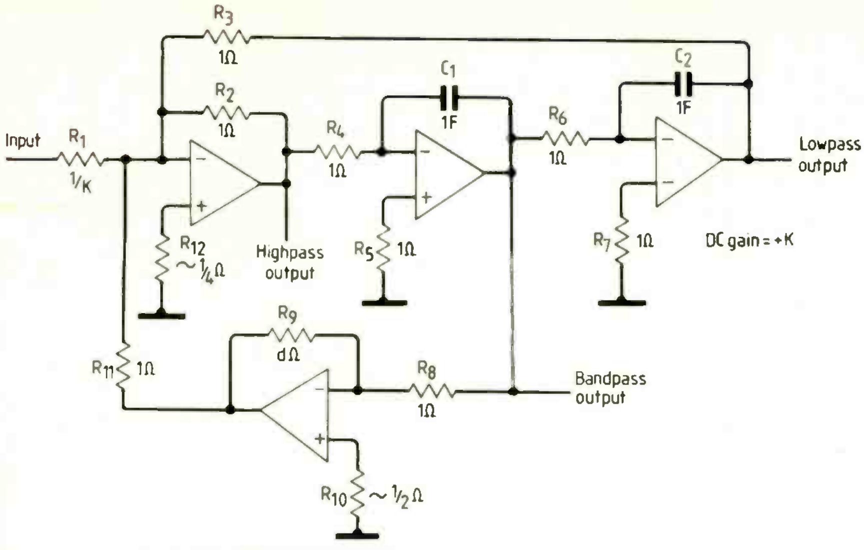


Fig.9 The normalized, variable gain, second-order, state-variable active section.

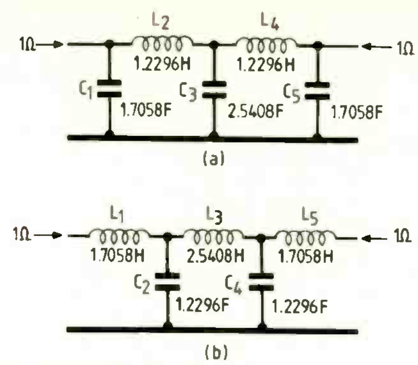


Fig.10 Normalized component values produced by CHEBPASS for the design example.

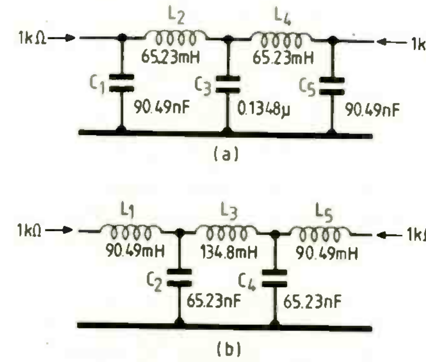


Fig.11 Scaled (3kHz, 1KΩ) values produced by SCALE for the design example.

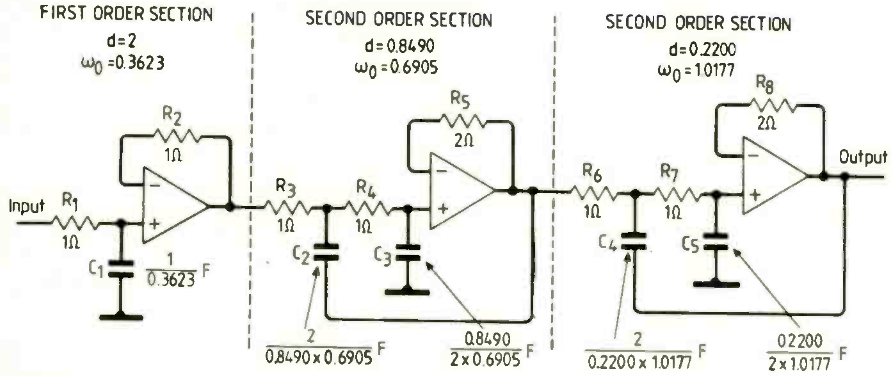


Fig.12 The normalized Sallen and Key implementation of the active-filter design example.

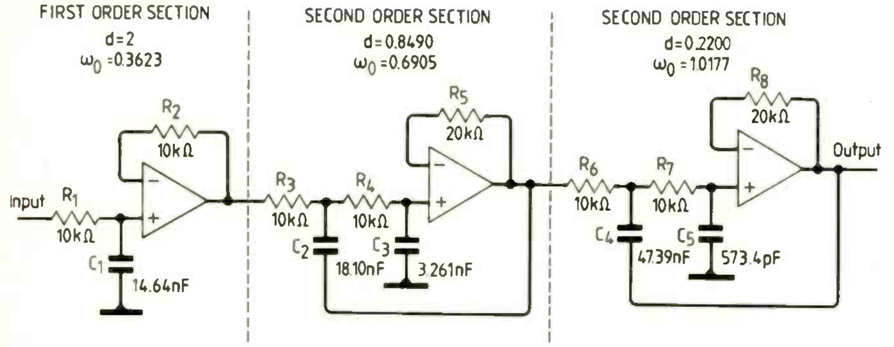


Fig.13 The active-filter design example with component values scaled to 3kHz, 10kΩ.

equations, prompting for the passband ripple and the filter order, and outputting the values of d and ωϕ for each and second-order section in the filter.

COMPONENT SCALING

The scaling of filter components from their 1 rad/s, 1 ohm values to their final frequency and impedance is a vital part of the modern filter design process. Equations (40) and (41) show how capacitors and inductors are scaled.

$$C = C_n \frac{1}{2\pi f Z} \quad (40)$$

$$L = L_n \frac{Z}{2\pi f} \quad (41)$$

where C is the final capacitor value, L is the final inductor value, C_n is the 1 ohm, 1 rad/s capacitor value, L_n is the 1 ohm, 1 rad/s inductor value, Z is the final impedance, and f is the final frequency.

In active filters, resistors are scaled using the simple formula:

$$R = R_n \times Z \quad (42)$$

where R is the final resistor value, R_n is the 1 ohm, 1 rad/s resistor value, and Z is the final impedance.

Scaling capacitors and inductors can be a tedious process, especially if many components have to be scaled and several different final impedance and frequency combinations tried. SCALE (Program 6) performs this scaling process. As well as scaling the components, the program adjusts the scaled value and prints it pF, nF or μF for capacitors, and μH, mH or H for inductors. Note that the parts of the program which adjust and print the values are constructed as subroutines, starting at line 1000 (for capacitors) and line 2000 (for inductors). The statements following the subroutine returns cause the prompt for another normalized component value to be printed. If the final frequency and impedance need to be modified each time the scaling process is carried out, lines 260 and 360 can be changed to GO TO 100.

EXAMPLES

To illustrate the use of the techniques described in this article, two worked examples are now presented:

Obtain passive and active implementations of Chebyshev low-pass filters with the following specifications:

	Passive	Active
passband ripple	0.5dB	0.5dB
cut-off frequency	3kHz	3 kHz
stopband attenuation	40dB	40dB
at a frequency of	6 kHz	6 kHz
system impedance	1 kohm	10 kohm

Note that the filter specifications are identical except for the system impedances, and therefore the same order will be required for both filters. Running FILTORD in the "c" option gives exact and rounded orders of 4.82176 and 5 respectively.

Tackling the passive implementation first: the system impedance for the filter is specified as 1 kΩ, implying equal source and terminating impedances. When CHEBPASS

Program 1 (FILTORD)

```
5       REM This program is called FILTORD (Spectrum version)
100     INPUT "Butterworth or Chebyshev (b/c):";f$
110     IF f$ = "b" THEN GO TO 200
120     IF f$ = "c" THEN GO TO 300
130     REM invalid input
140     PRINT "invalid filter type, try again"
150     GO TO 100
200     REM Butterworth filter selected
210     INPUT "cut-off frequency:";fcut
220     INPUT "required attenuation (dB):";atten
230     INPUT "at a frequency of:";fatten
240     LET n = LN(10↑(atten/10)-1)/(2*LN(fatten/fcut))
250     GO TO 1000
300     REM Chebyshev filter selected
310     INPUT "ripple cut-off frequency:";fcut
320     INPUT "required attenuation (dB):";atten
330     INPUT "at a frequency of:";fatten
340     INPUT "passband ripple (dB):";ripple
350     LET etasqu = 10↑(ripple/10)-1
360     LET x = SQR((10↑(atten/10)-1)/etasqu)
370     LET y = fatten/fcut
380     LET n = LN(x+SQR((x*x)-1))/LN(y+SQR((y*y)-1))
390     GO TO 1000
1000    REM order rounding and print routi
1010    LET nround = INT(n)-1
1020    PRINT "exact order =
1030    PRINT "rounded order
1040    STOP
```

Program 2 (BUTPASS)

```
5       REM This program is called BUTPASS (Spectrum version)
100     INPUT "filter order:";n
110     INPUT "termination impedance (ohm):";rl
200     IF rl>1 THEN LET lambda = ((rl-1)/(rl+1))↑(1/n)
210     IF rl<1 THEN LET lambda = ((1-rl)/(rl+1))↑(1/n)
220     IF rl=1 THEN LET lambda = 0
300     PRINT "Component designations are from output to input"
310     REM first component value
320     LET a = 2*SIN(PI/(2*n))
330     IF rl>1 THEN GO TO 400
340     GO TO 500
400     LET comp = a/(rl*(1-lambda))
410     PRINT "C1=";comp;" F"
420     GO TO 600
500     LET comp = (a*rl)/(1-lambda)
510     IF rl<1 THEN PRINT "L1=";comp;" H"
520     IF rl=1 THEN PRINT "C1/L1=";comp;" F/H"
600     REM calculating other components
610     FOR m=1 TO n/2
620     LET comp = (4*SIN((PI*(4*m-3))/(2*n))*SIN((PI*(4*m-1))/(2*n)))/
        ((1-lambda*2*COS(PI*(4*m-2)/(2*n))+lambda↑2)*comp)
630     IF rl>1 THEN PRINT "L";2*m;"=";comp;" H"
640     IF rl=1 THEN PRINT "L";2*m;"/C";2*m;"=";comp;" H/F"
650     IF rl<1 THEN PRINT "C";2*m;"=";comp;" F"
700     IF n<=2*m THEN STOP
710     LET comp = (4*SIN((PI*(4*m-1))/(2*n))*SIN((PI*(4*m+1))/(2*n)))/
        ((1-lambda*2*COS(PI*4*m/(2*n))-lambda↑2)*comp)
720     IF rl>1 THEN PRINT "C";2*m+1;"=";comp;" F"
730     IF rl=1 THEN PRINT "C";2*m+1;"/L";2*m+1;"=";comp;" F/H"
740     IF rl<1 THEN PRINT "L";2*m+1;"=";comp;" H"
800     NEXT m
```

Program 3 (CHEBPASS)

```
5       REM This program is called CHEBPASS (Spectrum version)
100     INPUT "filter order:";n
110     INPUT "passband ripple (dB):";ap
120     INPUT "termination impedance (ohm):";rl
200     LET refl = SQR(10↑(ap/10)-1)
210     REM test for n odd
220     IF n = 2*(INT(n/2)+0.5) THEN GO TO 400
300     REM n must be even
310     LET o$ = "even"
320     LET a = (1+refl↑2)*4*rl/((rl-1)↑2)
330     IF a<=1 THEN GO TO 500
340     REM a> 1, therefore can't implement this specification
350     PRINT "Can't implement this specification"
360     STOP
400     REM n is odd
410     LET o$ = "odd"
420     LET a = 4*rl/((rl+1)↑2)
500     LET x = (1/refl+SQR((1/refl↑2)+1))↑(1/n)
510     LET y = (SQR((1-a)/refl↑2)+SQR(((1-a)/refl↑2)+1))↑(1/n)
520     LET x1 = x-1/x
530     LET y1 = y-1/y
540     LET num1 = PI/(2*n)
600     REM first component value
610     LET comp = (4*SIN(num1))/(x1-y1)
700     IF o$="odd" AND rl>1 THEN PRINT "L1 = ";comp;"H"
710     IF o$="odd" AND rl=1 THEN PRINT "C1/L1 = ";comp;"F/H"
720     IF o$="odd" AND rl<1 THEN PRINT "C1 = ";comp;"F"
730     IF o$="even" AND rl<1 THEN PRINT "C1 = ";comp;"F"
740     IF o$="even" AND rl>1 THEN PRINT "L1 = ";comp;"H"
800     FOR m=1 TO n/2
810     LET b = x1↑2-2*x1*y1*COS((4*m-2)*num1)+y1↑2 +(2*SIN((4*m-2)*num1))↑2
820     LET comp = (16*SIN((4*m-3)*num1)*SIN((4*m-1)*num1))/(b*comp)
830     IF o$="odd" AND rl>1 THEN PRINT "C";2*m;"= ";comp;"F"
840     IF o$="odd" AND rl=1 THEN PRINT "L";2*m;"/C";2*m;"= ";comp;"H/F"
850     IF o$="odd" AND rl<1 THEN PRINT "L";2*m;"= ";comp;"H"
860     IF o$="even" AND rl<1 THEN PRINT "L";2*m;"= ";comp;"H"
870     IF o$="even" AND rl>1 THEN PRINT "C";2*m;"= ";comp;"F"
900     IF n<=2*m THEN STOP
910     LET b = x1↑2 - 2*x1*y1*COS(4*m*num1) + y1↑2 + (2*SIN(4*m*num1))↑2
920     LET comp = (16*SIN((4*m-1)*num1)*SIN((4*m+1)*num1))/(b*comp)
930     IF o$="odd" AND rl>1 THEN PRINT "L";2*m+1;"= ";comp;"H"
940     IF o$="odd" AND rl=1 THEN
        PRINT "C";2*m+1;"/L";2*m+1;"= ";comp;"F/H"
950     IF o$="odd" AND rl<1 THEN PRINT "C";2*m+1;"= ";comp;"F"
960     IF o$="even" AND rl<1 THEN PRINT "C";2*m+1;"= ";comp;"F"
970     IF o$="even" AND rl>1 THEN PRINT "L";2*m+1;"= ";comp;"H"
1000    NEXT m
```

Program 4 (BUTACT)

```
5       REM This program is called BUTACT (Spectrum version)
100     INPUT "required order:";order
200     PRINT "Values of d and w0 are:"
210     PRINT
300     FOR k = 1 TO (order-1)/2
310     LET thetak = ((2*k-1)*PI)/(2*order)
320     LET real = -SIN(thetak)
330     LET imag = COS(thetak)
340     LET w0 = SQR(real*real + imag*imag)
350     LET d = -(2*real)/w0
360     PRINT "  d = ";d
370     PRINT " w0 = ";w0
380     PRINT
390     NEXT k
400     STOP
```

Program 5 (CHEBACT)

```
5       REM This program is called CHEBACT (Spectrum version)
100     INPUT "passband ripple (dB):";ap
110     INPUT "required order:";order
200     LET eta = SQR(10↑(ap/10)-1)
210     LET inveta = 1/eta
220     LET a = (LN((inveta)+SQR((inveta↑2)+1)))/order
230     LET sinha = (EXP(a)-EXP(-a))/2
240     LET cosha = (EXP(a)+EXP(-a))/2
300     FOR k=1 TO (order+1)/2
310     LET thetak = ((2*k-1)*PI)/(2*order)
320     LET real = -SIN(thetak)*sinha
330     LET imag = COS(thetak)*cosha
340     LET w0 = SQR(real*real+imag*imag)
350     LET d = -2*real/w0
400     PRINT "   d = ";d
410     PRINT "  w0 = ";w0
420     PRINT
430     NEXT k
440     STOP
```

Program 6 (SCALE)

```
5       REM This program is called SCALE (Spectrum version)
100     INPUT "final frequency (Hz) :";freq
110     INPUT "final impedance (ohm):";imped
120     INPUT "capacitor or inductor (c/l):";c$
130     IF c$ = "c" THEN GO TO 200
140     IF c$ = "l" THEN GO TO 300
150     REM invalid component type
160     PRINT "invalid component, try again"
170     GO TO 120
200     REM capacitor scaling routine
210     INPUT "normalised capacitor value (F):";ncap
220     LET scap = ncap/(2*PI*freq*imped)
230     REM now get it printed
240     LET prcap = scap
250     GO SUB 1000
260     GO TO 120
300     REM inductor scaling routine
310     INPUT "normalised inductor value (H):";nind
320     LET sind = nind*imped/(2*PI*freq)
330     REM now get it printed
340     LET prind = sind
350     GO SUB 2000
360     GO TO 120
1000    REM capacitor adjusting and print subroutine
1010    IF prcap<1e-9 THEN GO TO 1100
1020    IF prcap<1e-7 THEN GO TO 1200
1030    GO TO 1300
1100    REM value will be printed in pF
1110    LET prcap = prcap*1e12
1120    PRINT "scaled capacitor value = ";prcap;"pF"
1130    RETURN
1200    REM value will be printed in nF
1210    LET prcap = prcap*1e9
1220    PRINT "scaled capacitor value = ";prcap;"nF"
1230    RETURN
1300    REM value will be printed in uF
1310    LET prcap = prcap*1e6
1320    PRINT "scaled capacitor value = ";prcap;"uF"
1330    RETURN
2000    REM inductor adjusting and print subroutine
2010    IF prind<1e-3 THEN GO TO 2100
2020    IF prind<1 THEN GO TO 2200
2030    GO TO 2300
2100    REM value will be printed in uH
2110    LET prind = prind*1e6
2120    PRINT "scaled inductor value = ";prind;"uH"
2130    RETURN
2200    REM value will be printed in mH
2210    LET prind = prind*1e3
2220    PRINT "scaled inductor value = ";prind;"mH"
2230    RETURN
2300    REM value will be printed in H
2310    PRINT "scaled inductor value = ";prind;"H"
2320    RETURN
```

prompts for the terminating impedance, 1 is therefore entered. CHEBPASS produces the following results:

$C_1/L_1 = 1.70577$ F/H
$L_2/C_2 = 1.22963$ H/F
$C_3/L_3 = 2.54083$ F/H
$L_4/C_4 = 1.22963$ H/F
$C_5/L_5 = 1.70577$ F/H

These values correspond to the dual circuits shown in Fig. 10. SCALE can be run with a final frequency of 3000 Hz and a final impedance of 1000 ohm, giving the scaled values shown in Fig. 11.

The active implementation will now be designed. Running CHEBACT gives the following values (rounded to four decimal places here) for the three sections required (one first and two second order sections)

$$d = 0.2200$$
$$\omega\phi = 1.0177$$

$$d = 0.8490$$
$$\omega\phi = 0.6905$$

$$d = 2$$
$$\omega\phi = 0.3623$$

If unity-gain sections are used to built the filter, the circuit of Fig. 12 is the Sallen and Key implementation. Note that the capacitor values contain the d and $\omega\phi$ results, and the first-order section (with d=2) is placed nearest the input, then the d = 0.8490 and finally the d = 0.2200 section.

The circuit is scaled to 3kHz, 10kΩ by multiplying the resistors by 10000 and using SCALE on the capacitor values. Figure 13 shows the final component values.

Computer simulation of the circuits in Figs 5, 11 and 13 give identical results, as expected. The 0.5dB passband ripple and 3 kHz cut-off frequency specifications are met exactly, and the attenuation at 6 kHz is 42dB, better than specified because of the rounding upwards of the filter order.

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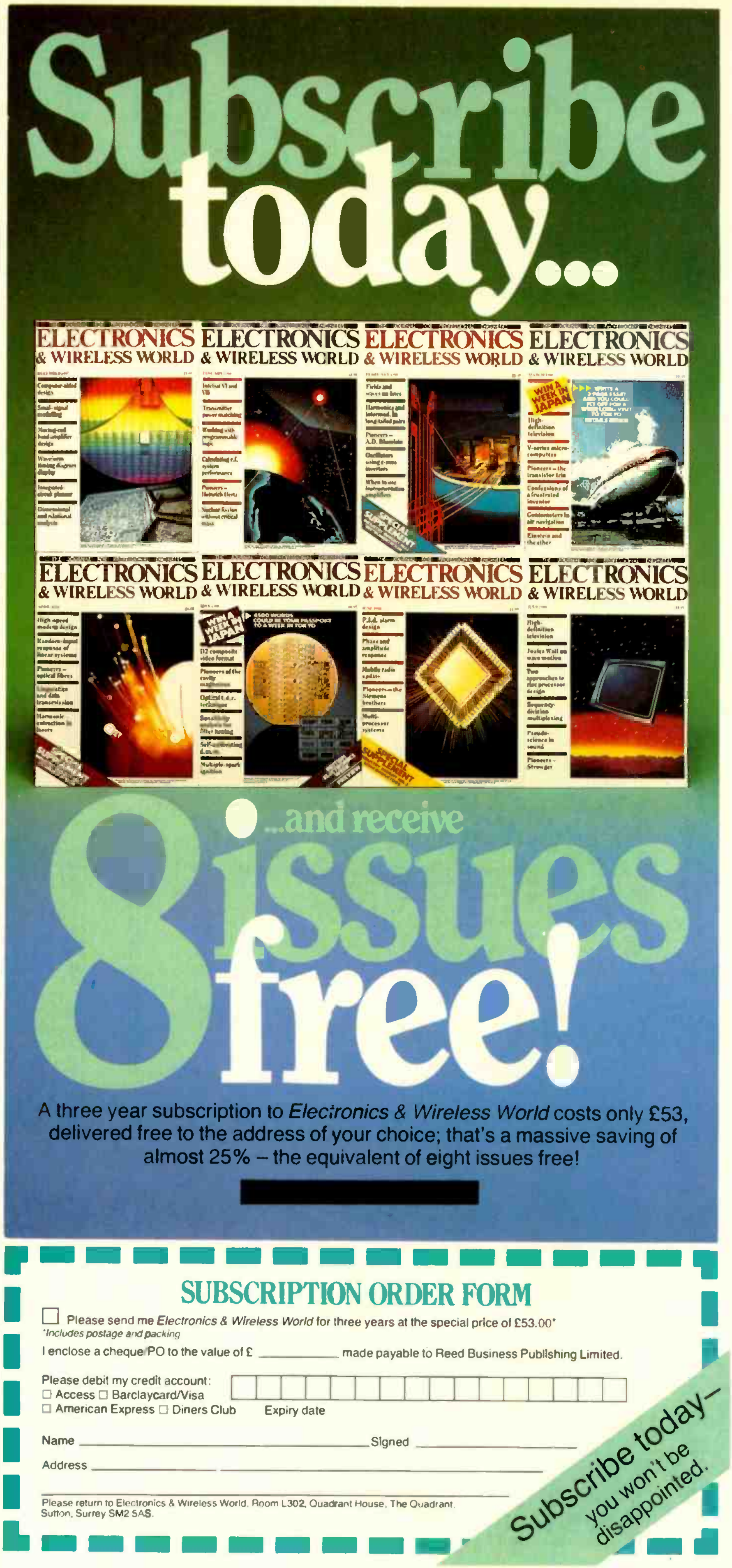
Filter Design and Evaluation, G E Hansell, published by Van Nostrand Reinhold.

References

1. Amateur Radio Software, J. Morris, published by RSGB, 1985. Appendices: BASIC Program Adaptions, pages 319 to 324.
2. Handbook of Filter Synthesis, A.I. Zverev, published by Wiley, 1967. Pages 140, 141.
3. Analog and Digital Filters: Design and Realisation, H. Y-F. Lam, published by Prentice-Hall. Page 238..
4. As reference 3, pages 236 to 238.
5. As reference 3, pages 261 to 262.
6. Active Filter Cookbook, D. Lancaster, published by Sams, 1980. Page 156.
7. As reference 6, pages 73 to 76, 79.
8. Electronic Filter Design Handbook, A.B. Williams, published by McGraw-Hill, 1981. Pages 12-1, 12-19 to 12-23.
9. As reference 3, page 235.
10. Analog Filter Design, M.E. Van Valkenburg, published by Holt, Rinehart and Winston. Page 235.



Sequency-division multiplex

The use of Walsh functions offers a successful alternative to frequency-division multiplex or time-division multiplex.

CHARLES LANGTON

The problem is as old as the concept of electrical communication itself. How can several (perhaps many) stations at each end of a link communicate with each other when there is only one channel connecting them?

The solution, of course, is to multiplex, or interleave all the transmitting stations in some way, and send this composite signal down the line. At the receiving end the signals have to be disentangled or de-multiplexed, so that ultimately the separated signals are each presented to their appropriate reproducer. Figure 1 shows the general idea of multiplexing.

In a correctly adjusted system, each receiver will be able to listen to its appropriate transmitter and be oblivious to the fact that other messages are sharing the same link.

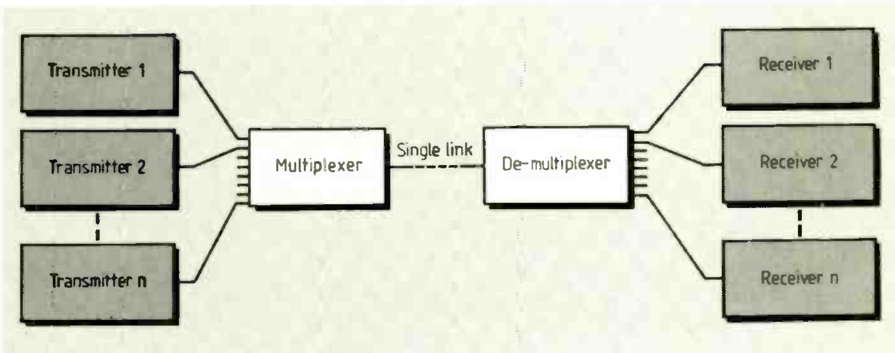


Fig. 1. General system of multiplexing.

TRADITIONAL METHODS OF MULTIPLEXING

The two most well known methods of multiplexing are frequency-division multiplex (f.d.m.) and time-division multiplex (t.d.m.).

Frequency-division multiplex was certainly in force in the early days of wireless communication. It was necessary for each transmitter to impress its modulation upon a carrier frequency unique to that station. All such carriers were propagated through the same link (free space, alias aether). Upon reaching a receiving aerial, tuned circuits or filters selected the desired modulated carrier from the multitude of unwanted signals, thereby resulting in intelligible communication via one common medium.

This principle has been most elegantly refined to the extent that single-line links or radio beams are in normal use, each carrying very many channels of information simultaneously.

Time-division multiplex is an alternative system which allows many channels of information to be sent along one single link. The development of t.d.m. was accelerated by feasibility studies into pulse-code modulation (p.c.m.) by the international telecommunication authorities. P.c.m. was declared good but expensive, and could only be introduced if costs could be reduced by multiplexing many channels along each link. T.d.m., being a pulsed system, fitted in with p.c.m. ideally.

In practice, a t.d.m. system operates by sampling the audio signal from each channel in turn at a rapid rate; the samples are sent down the link consecutively. This is done under clock control, so that the same clock may synchronize the re-assembly of the samples at a receiver. So long as the sampling rate is at least twice as great as the highest harmonic frequency to be transmitted, no information will be lost by this process.

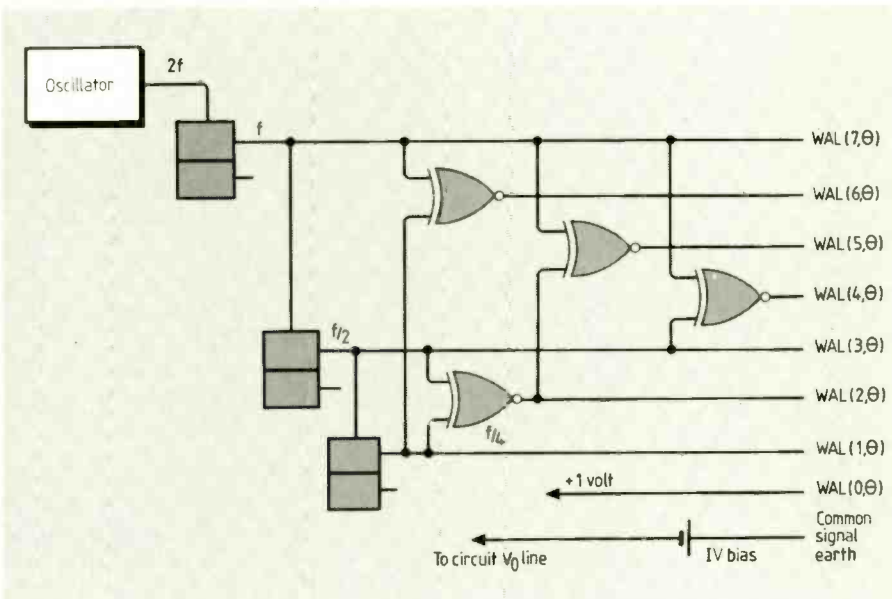


Fig. 2. Circuit to generate one set of eight Walsh functions

SEQUENCY-DIVISION MULTIPLEX

A third method, sequency-division multiplex (s.d.m.), makes use of Walsh functions, and this is the subject of this article. In some ways this system is similar to f.d.m., in that the audio (or data) signal modulates a carrier.

In other ways, as Walsh functions are pulses, the s.d.m. system resembles a t.d.m. system. The input signals may be analogue, such as audio waves, or they may first of all be sampled and the sampled pulses multiplexed. Alternatively the input may be data pulses in pure binary or other code.

Each signal modulates one particular Walsh function from the set of Walsh functions. The complete set of modulated Walsh functions is now added together, and the resultant composite waveform is transmitted along a single link to the receiver. At the receiver, de-multiplexing is accomplished simply, because Walsh functions are orthogonal. This means that any desired signal may be recovered from the composite by multiplying this waveform by the Walsh function used as the carrier of the original signal. We will now look in a little more detail at the s.d.m. method.

Walsh functions. Walsh waves may be generated fairly simply in a suitable waveform generator. They must be generated in sets, each set containing N functions, N being a binary number. The order of functions in a set is designated by n, which may be any integer from $n=0$ to $n=(N-1)$. The number of time-slots per function is also equal to N. Such time-slots are designated θ, which again may be any integer from $\theta=0$ to $\theta=(N-1)$. In general, the instantaneous value of any Walsh function n during any time-slot θ is

$$WAL(n,\theta)$$

This can only have a value of $+1$ or -1, as Walsh waves are rectangular waves having an amplitude of ±1. Figure 2 shows the circuit diagram of a Walsh function generator where N is equal to 8. There are thus eight output wires, the functions being

$$WAL(0,\theta), WAL(1,\theta).........WAL(7,\theta)$$

If the t.t.l. is employed in the above circuit, the output pulses would normally have levels of about zero volts and two volts. The addition of a one-volt bias in series with the common lead results in output levels of $+1$ and -1 volts. Figure 3 shows the set of waveforms produced. $WAL(0,\theta)$ is a constant having a value of $+1$ volt.

MULTIPLEX TRANSMITTER

Each of the eight waves in the set has a different sequency, where sequency is defined as half the average number of zero crossings per second. As an example, if T seconds is the total time duration of 8 time-slots, we can see that the sequency of $WAL(5,\theta)$ is

$$0.5(6/T) = 3/T \text{ zero-crossings per second}$$

This is exactly analogous to the concept of frequency in more conventional waveforms, in which the period of the wave is constant cycle by cycle.

Walsh functions generated in this manner can now be used as carrier waves. However, to simplify our description, consider a system which uses only four Walsh carrier waveforms, namely $WAL(0,\theta)$, $WAL(1,\theta)$, $WAL(2,\theta)$, and $WAL(3,\theta)$. Each of these carriers may be modulated by one input signal. Hence the total of four input signals will be designated S_0, S_1, S_2 and S_3. The basic schematic of this system is shown in Fig. 4 (transmitter) and Fig. 5 (receiver). The main transmitter waveforms are shown in Fig. 6.

Sequency-division multiplex transmitter. With reference to Fig. 6, the input signals S_0-S_3 in this diagram are three-bit binary numbers. In each case, the first (least significant) digit is transmitted during time period T_1. The middle digit occurs during T_2, and the most significant digit during T_3. Note that, as we read these time intervals from left to right, the digit positions appear in the reverse order to that normally accustomed. For example, the numerical value of S_1 is 3, although at first glance at the diagram it looks like 6.

The signals from each channel pass into a modulator, and modulate a corresponding Walsh carrier. This modulator circuit is a multiplier, and therefore gives an output which is the product of the input signal S_n and the Walsh function $WAL(n,\theta)$

$$\text{Modulator output} = S_n \times WAL(n,\theta)$$

However, S_n is a binary digit having a value of 1 or 0, so the multiplication process is reduced to its simplest level.

During T_1, then, there will be four activities in progress and the four modulated output carriers will be as shown under the T_1 column on the right hand side of Fig. 6. Furthermore, during each of the time-slots within T_1 the four modulated carriers are

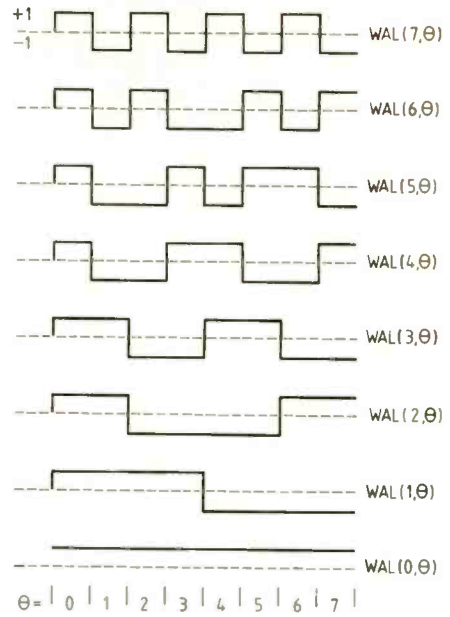


Fig. 3. Repetitive cycle of Walsh waveforms produced by the circuit shown in Fig. 2.

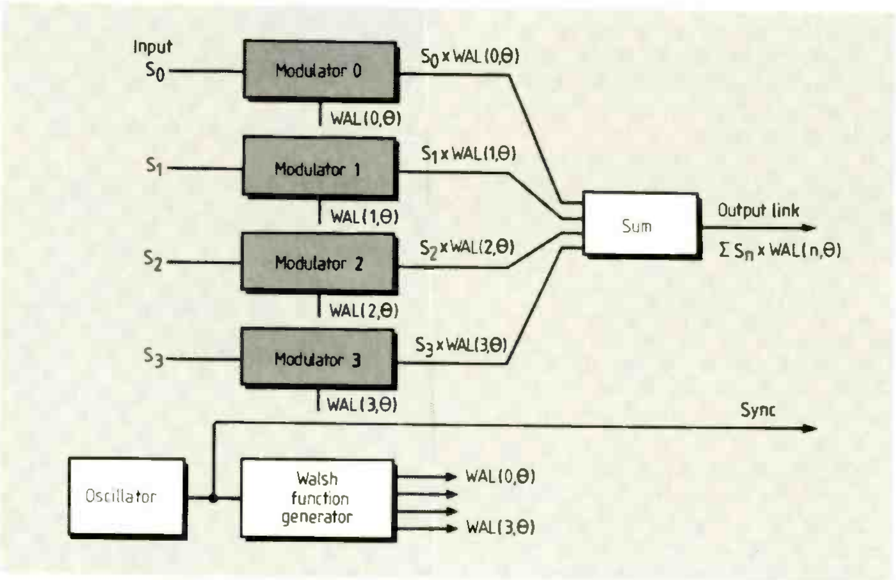


Fig. 4. Four-channel Walsh multiplexer.

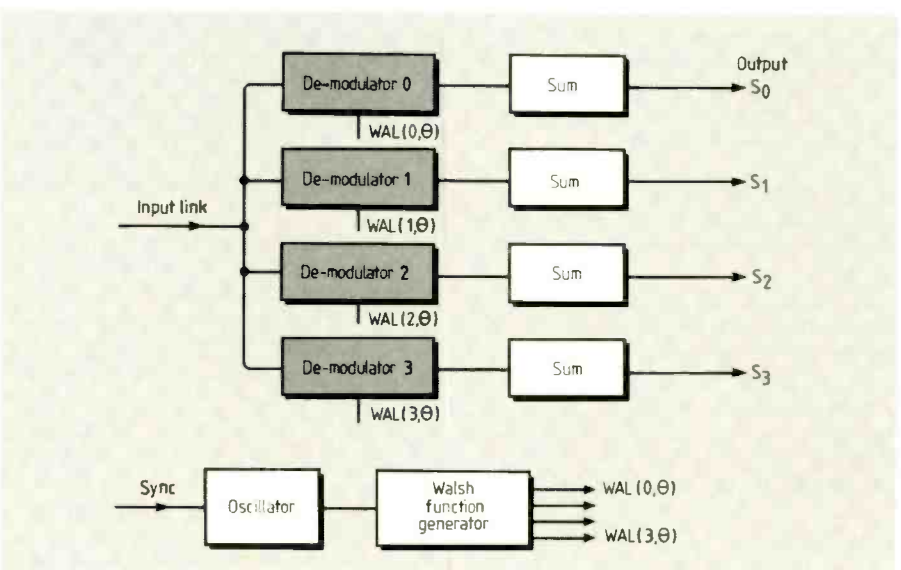


Fig. 5. De-multiplexer, or receiver, for four channels.

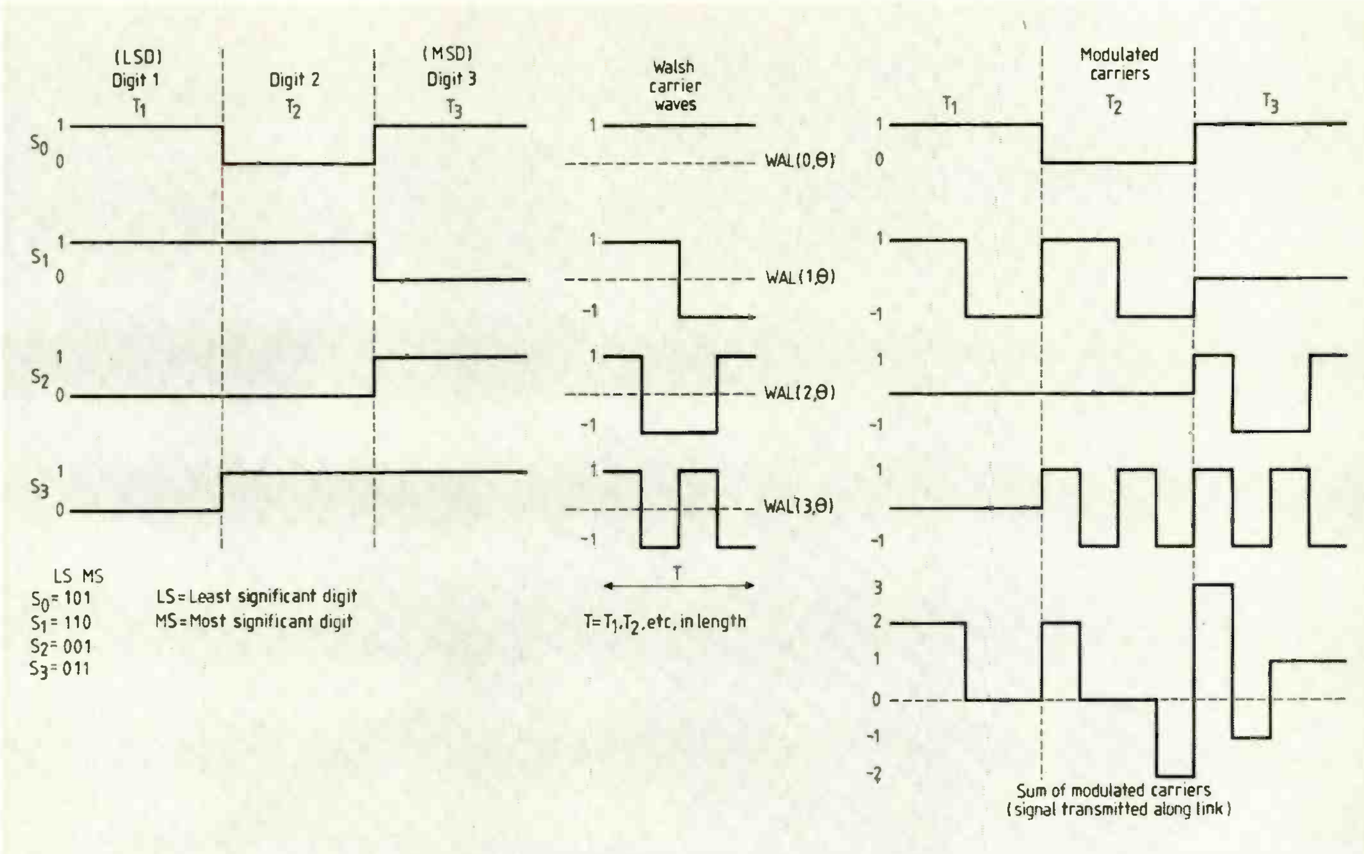


Fig. 6. Waveforms in the multiplexing process.

being instantaneously added, the resultant summed waveform being shown at the lower right of Fig. 6. This is the transmitted multiplexed signal produced during period T.

The total action is repeated for each subsequent digit during periods T and T. By the end of T, each of the four channels will have completed the transmission of one 3-bit word.

Synchronizing. In addition to transmitting the multiplexed signal, it is necessary to transmit a clock pulse to synchronize an identical Walsh function generator at the receiver. This is because at the receiver, identical Walsh functions are required to de-multiplex the signal.

RECEIVER

The incoming multiplexed wave is fed to input 1 of each of four demodulators, each of which is a multiplier. Input 2 of each is driven from the appropriate Walsh function, the Walsh function being generated at the receiver and synchronized by the incoming clock pulse.

Consider channel zero. During each digit period (T, T, and T) there are four time slots, θ=0, 1, 2, and 3. The amplitudes of pulses emerging from the demodulator are added together, giving a total value for each digit period. If this total value is finite, the value of the digit is 1. If zero, the digit is 0. In practice, the total value will be 4 or zero. Had we used eight channels (and eight Walsh functions) the total value would have been 8, and so on.

Similar remarks apply to channels 1, 2, and 3. Needless to say, if the system is working correctly, the output digits per channel will be the same as those at the input.

The system described above accommodates only four channels and three-digit words. In practice, many channels of any reasonable word length can be multiplexed in this way, subject only to the usual factors which limit high-speed working such as bandwidth, device switching time, and signal to noise ratio. The number of channels must be matched by the number of Walsh functions in the set, and this must be a binary number.

All Walsh waves consist of rectangular pulses having values of +1 or −1, so that processing (such as multiplication) is reduced to very simple operations.

An important advantage of Walsh multiplexing compared with frequency-division multiplex is that modulating a Walsh carrier only produces a single-sequency sideband, whereas modulating a sinusoidal carrier produces two, the upper and the lower sidebands.

Furthermore, a Walsh system is less sensitive to noise than a time-division multiplex system. This is because any noise affecting the Walsh carrier is distributed evenly among all the multiplexed signals. The effective noise intensity per digit is therefore diminished.

Further reading

Gill, S.M., "The use of Walsh functions in data multiplexing", University of York, 1979.

Langton, C.H., "Orthogonality and Walsh functions", *Electronics & Wireless World*, March, 1987.

Leetronex '88

The North of England's electronics show moves to new premises this year: for the first time, it is to be staged in the newly-refurbished University of Leeds Exhibition Centre. All stands in the centre's two halls have been taken by exhibitors, and the show promises to be one of the best yet.

Leetronex, organized by the electrical and electronic engineering department of the University of Leeds, is now in its 25th year. Subject areas covered by the show include semiconductor products, hybrid circuits, instrumentation, data acquisition and control systems, signal conditioning, fibre optics, computer peripherals, p.c.b. artwork and manufacture, power supplies, technical training aids and recruitment. Many product launches have been promised.

To accompany the exhibition is a programme of seminars. This year's topics are

- Surface-mount technology, presented by Peter Grundy of Siemens Ltd, who will introduce some of the latest developments in circuit production.
- The Transputer and occam: Dr John Kerridge of the National Transputer Support Centre offers a practical guide to the Transputer and the Government's strategy for promoting its use.
- Mobile and cellular radio: Professor Peter Matthews of Leeds University examines the limitations on these systems imposed by Nature and considers the ways in which they affect system design and service to the user.

Dates for Leetronex are 5-7 July, 1988. Further information from the organizer's office on 0532-420339.

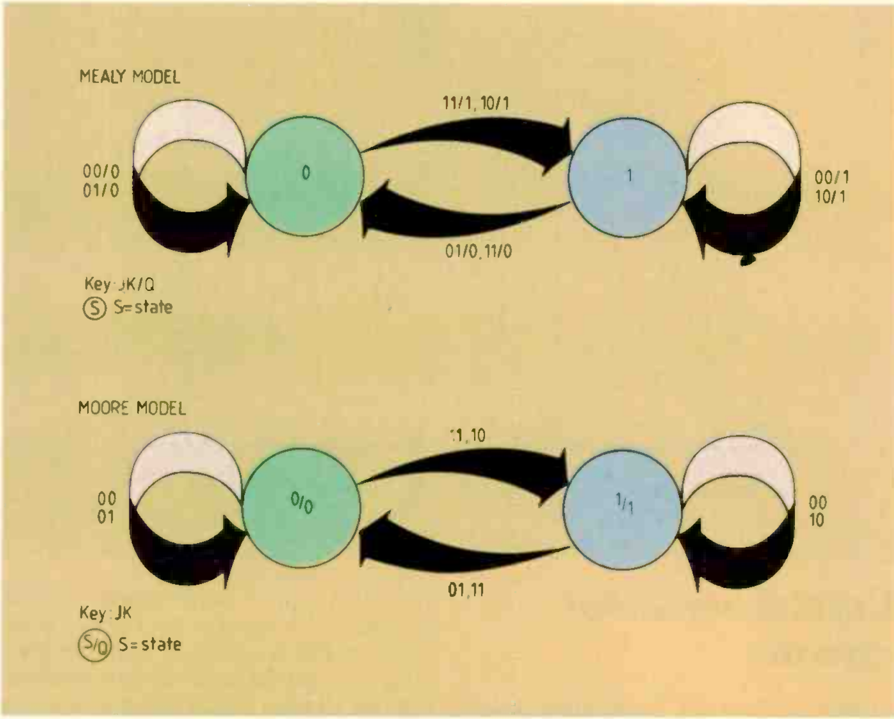
APPLICATIONS SUMMARY

State machines explained

Mealy and Moore models are commonly used to simplify state-machine design. According to a note called 'State machines explained' from AMD, which of the two models is used depends to some extent on the designer's preference but there are also benefits of one method over the other that vary according to the application.

In a Mealy model, state-machine output is dependent on the inputs and the current state whereas in the Moore model, the outputs only depend on the current state. Being more general, the Mealy model tends to be more popular.

Although the seven-page note gives an example of each model, it leaves the reader to decide what the advantages and disadvantages of the two are. It briskly steps from simple state diagrams to Mealy and Moore models using the JK register as an example. Finally, it presents high-level language control software for a drinks machine using sequential synchronous logic.



Telephone and terminal i.cs

Brief details of how to link together various integrated-services digital-network i.cs from Siemens are given in 'ICs for ISDN telephone sets and terminals'.

Divided into two sections, the note first outlines an ISDN 'feature telephone' and explains how this apparatus's X interface as specified by the Deutsche Bundespost* can be in two forms. Simple X interfaces allow only an auxiliary speaker on recording equipment to be connected to the port but the more complex form can take various equipment such as automatic diallers, alarm systems and measurement/control units.

Voice and data terminals are covered in the note's second section. Such terminals, which include network terminators, repeaters, multiplexers and exchange systems, can incorporate many features including packet-switched data links and slower data links for Teleservices.

In order to fully understand the seven-page note, you will need a prior knowledge of ISDN abbreviations.

*There is no similar UK specification for the X interface.

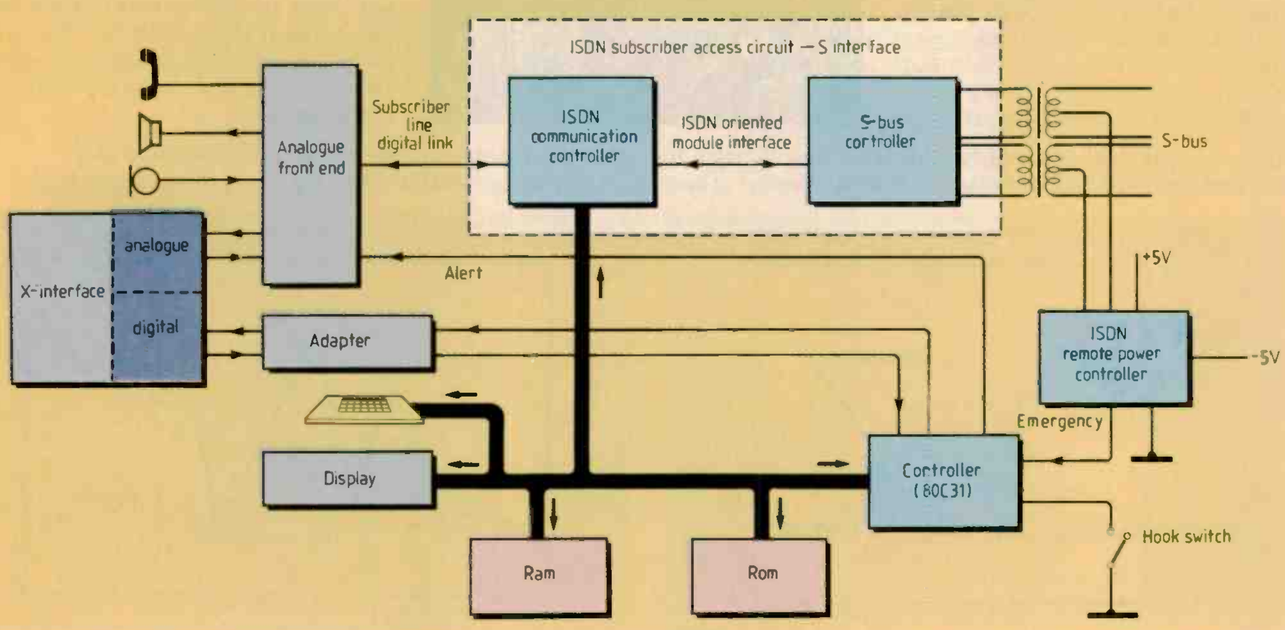
ADDRESSES

AMD
Applications Department
AMD House
Goldsworth Road
Woking
Surrey GU21 1JT
04862 22121

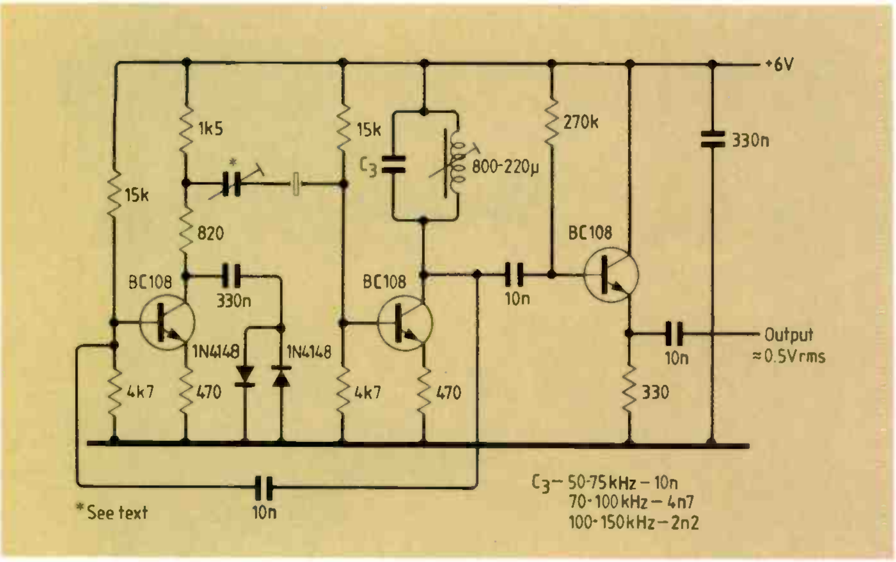
Siemens
Siemens House
Windmill Road
Sunbury-on-Thames
Middlesex TW16 7HS
0932 752626

IQD
Hitek
Ditton Walk
Cambridge CB5 8QD
0460 74433

Electrolube
Blakes Road
Wargrave
Berkshire RG10 8AW
073 522 4031



APPLICATIONS SUMMARY



Crystal oscillator circuits

Quartz crystals for frequencies below 150kHz, having a relatively high equivalent series resistance, need a high-gain oscillator circuit. Cascading two common-emitter stages as shown in IQD's Crystal Catalogue provides sufficient gain for crystal frequencies down to 50kHz.

Diodes in the first transistor's collector circuit limit the crystal drive level to avoid damage and the tuned circuit provides selectivity. Parallel resonance is set by the trimming capacitor; for series-resonance calibration, the trimmer should be replaced with a 1nF fixed capacitor and the inductor adjusted instead.

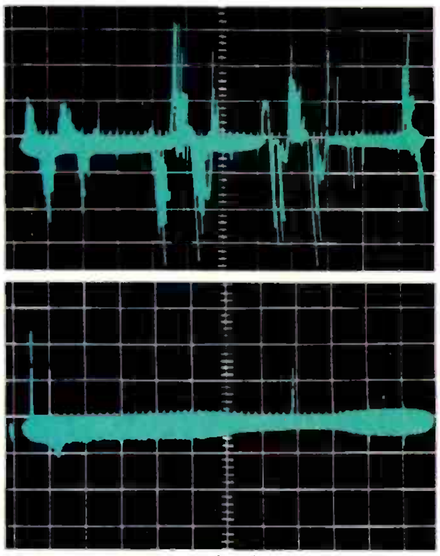
In the applications section of the catalogue is a number of other discrete and integrated-circuit oscillators, including a discrete-component circuit for up to 200MHz.

Contact lubrication

In a discussion of the merits of contact lubrication, Electrolube says that keeping contact surfaces clean and dry does not necessarily remove the problems of noise caused by high contact resistance.

Four main factors influence contact resistance — contact surface condition, contamination, frettage and contact bounce. Contact lubricant, says Electrolube, evens out the peaks and troughs of the contact surface, thus increasing the contact-surface area. This in turn reduces contact resistance, minimizes the formation of hot spots and reduces surface deformation due to friction wear.

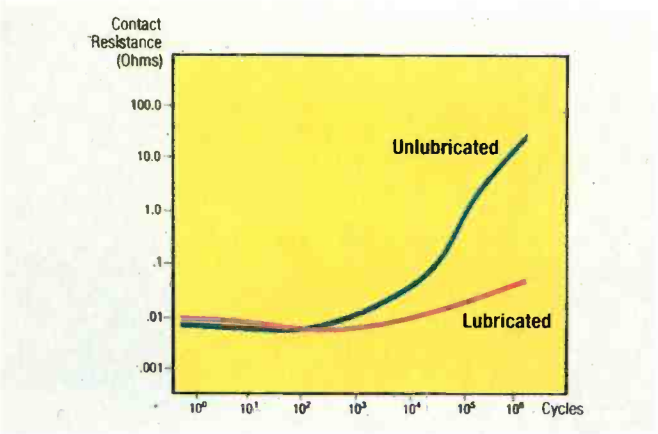
Arcing, occurring when the contact opens, causes airborne contaminants to react with the contact metal and form a surface film. It also causes electrochemical reactions in the air between the contacts which result in the formation of nitric acid, and it causes metal transfer between the contacts due to ionization, which is a particular problem when switching direct current (the pip-and-crater effect). Lubrication reduces the effects of arcing since the lubricant effectively forms a bridge between the partially-open contacts to prevent current concentration at the contact-surface peaks and reduce current flow more gradually.

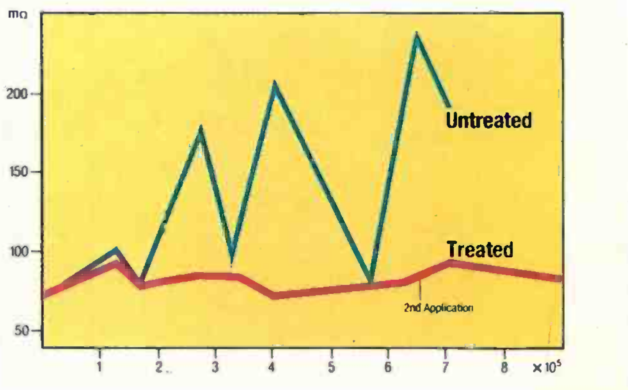


Noise on untreated (top) and treated contacts; Y is 0.15V/div., sweep is 20ms/div.

More precise information on how contact resistance is increased and prevented is included in the Contact lubricant brochure, and charts are used to illustrate the difference between the company's various lubricants. Two other new brochures cover conformal coatings and cleaning products in a similar way.

Peaks and troughs on a copper surface at x1000 magnification, top, gold-plated contacts under fretting conditions (left) with contact pressure of 50-60g, and resistance versus number of operations for sliding contacts of a wire-wound potentiometer (right).





Upgrading from 68000 to 68020/68881

Bob Coates has designed a Kaycomp-compatible platform board holding a 68020 processor and 68881 maths co-processor; it should plug into any single-master 68000 system that can keep up with the 68020's shorter bus cycle.

BOB COATES

My design objectives were to produce a platform board for the 68020 microprocessor and 68881 maths co-processor, together with interface logic, which could plug into the 68000 processor socket on the Kaycomp board, Fig.1. Interface logic is necessary because of differences between the 68000 and 68020, like the number of data-bus bits and the control-bus signals.

The result is a six-layer board measuring 80 by 100mm holding the two processors, two 20-pin Pals and one t.t.l. device – ten components in all. It plugs directly into the Kaycomp but it should also be suitable for other 68000 systems provided that there is no bus master other than the main processor and that the memory and peripherals can cope with the 68020's shorter bus cycle.

In order to significantly reduce component count and hence board size, I have used programmable-array logic for the interfacing. A problem for small organizations wishing to use such devices can be the cost of specialized equipment required for development and programming. In this article, I will show that programmable logic devices can be designed without any capital outlay.

Both of the processors have 32bit data paths and communication between the two is over the full 32bit bus. Communications with the rest of the Kaycomp board is over a 16bit bus, making use of the 68020's ability to dynamically adjust its bus size. For readers with small pockets, I have made the 68881 maths co-processor optional. Later upgrading is possible by simply plugging in the i.c. and breaking a link.

Kaycomp's firmware has been extended to cope with the new processors. The monitor has been improved to accommodate some of the new facilities on the 68020 and also to display all the extra registers. I have removed the 40-column option since there are too many registers for a 40-column display.

Both the assembler and disassembler have been extended to cope with all the new addressing modes and instructions, including all the maths instructions for the 68881.

Installation of the upgrade consists merely of altering a link on the Kaycomp board to configure the system for 27256 eproms, swapping the eproms, removing the 68000 and plugging in the platform board. Having only ten components, the board is easily and quickly assembled. Details of how to obtain the board, etc., will be given later.

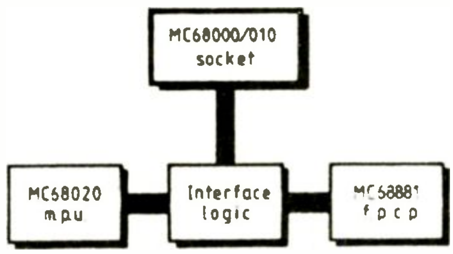


Fig.1. Platform board outline. Interface logic is needed because of differences between the 68000 and 68020.

An application note published by Motorola describes a platform board that appears to the host board to be identical to a 68000 as far as functionality and bus timing is concerned, which guarantees that the platform board will work with any host board designed for the 68000. However, such a board would be excessive for a simple board such as the Kaycomp and there are certain features which can be simplified or left out providing that the design of the target board is known.

There are six basic areas that must be dealt with by the interface logic, these are:

- Dynamic bus-sizing considerations
- Bus-cycle timing modification
- M6800-family device communication
- Interrupt processing
- Bus arbitration
- Co-processor selection

Dynamic bus-sizing considerations. A new feature available on the 68020 is the ability to transfer byte, word and long word operands using its dynamic bus-sizing capabilities, allowing mixing of different-width memory and peripherals within one system. To allow this, the data-transfer acknowledge (DTACK) signal found in the 68000 is replaced by two data-transfer and size-acknowledge ($DSACK_{0,1}$) inputs. The two data strobe outputs (UDS, LDS) are replaced by a single data strobe (DS) and two size outputs ($SIZ_{0,1}$) to indicate the type of transfer.

Since any 68000 host board always uses 16bit data transfers, the inputs are configured to always initiate 16bit transfers, except for accesses to the 68881 coprocessor when full 32bit transfers can be used.

Bus-cycle timing modification. There are some minor differences between the 68000 bus-cycle timing and that of the 68020. Assuming no wait states, a normal 68000 bus cycle takes four clock periods whereas a normal 68020 bus cycle takes three. To correctly emulate the timing of a 68000 would therefore require a modification to the bus timing. However, if it is known that the host board is able to cope with the shorter bus cycle then this removes the need for some of the interface logic, simplifying the design, and speeding up the system.

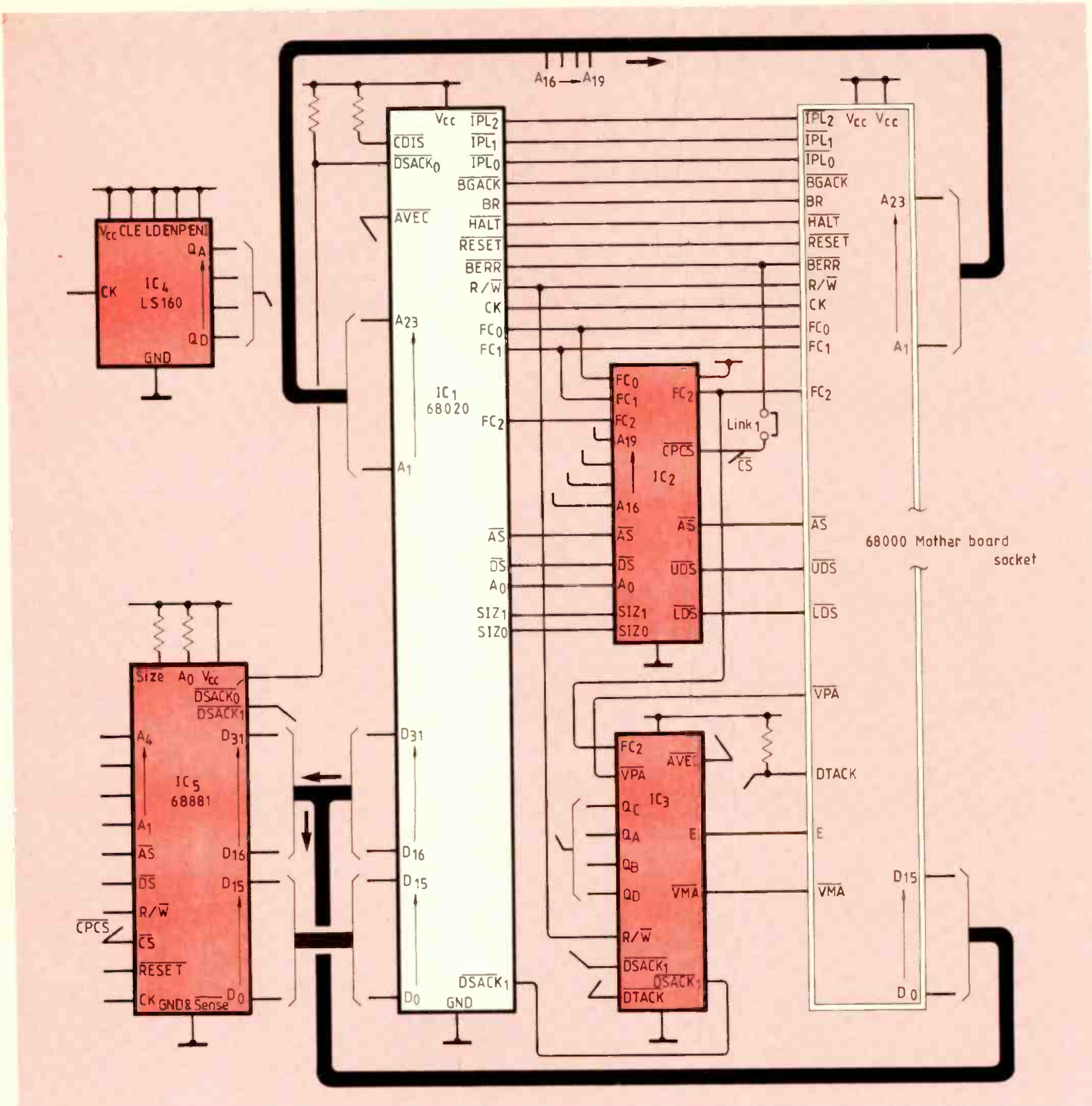
Kaycomp can cope with the shorter bus cycle provided that memories with an access time of 200ns or less (8MHz clock) are used, so I decided not to modify the 68020 bus cycle timing.

M6800 family device communication. On the 68000 the facility was available to interface to 6800-family peripheral devices, which require a synchronous bus transfer. This was initiated by the address decoding asserting the VPA input when accessing a 6800 device area. On the Kaycomp this type of transfer is used for the G64 bus interface.

This synchronous bus interface is no longer available with the 68020 and so an equivalent function must be implemented in hard logic in order to allow the G64 bus interface to continue to be used. This function accounts for most of the interface logic on the platform board.

Interrupt processing. On the 68000 an interrupt-acknowledge cycle is indicated by the three function-code outputs, FC_{0-2}, being all logical one and this is used to produce the IACK signal on the Kaycomp. With the 68020 however, all ones on the function-code lines indicates a c.p.u.-space cycle, which can indicate a breakpoint acknowledge, access-level control, co-processor communication or interrupt acknowledge cycles. Which it is is encoded in address lines A_{16} to A_{19} which have a different binary code allocated for each c.p.u.-space cycle type. The code for an interrupt-acknowledge cycle is all ones on the four address lines.

As the Kaycomp decodes only the function-code lines and not the address lines, FC_2 from the 68020 is gated with the four address lines before going to the Kaycomp and so this modified FC_2 signal will



only go high if all of FC_2 and A_{16-19} go high.

The only other c.p.u.-space cycle used in this design is the co-processor communication cycle.

On the 68000, the valid-peripheral address input, VPA, as well as being used for the M6800 communication mentioned previously is also used to indicate that an autovectored interrupt response is required. This pin is no longer used on the 68020 since synchronous transfers are not supported. It is replaced by AVEC which serves the same purpose as VPA for initiating autovectored interrupts.

Bus arbitration. On the Kaycomp board there is only one possible bus master, the 68000 processor itself. Consequently the platform board design was simplified by leaving out the interface-logic functions required to allow multiple bus masters. The only bus arbitration implemented is that

Fig.2. Two programmable logic i.cs greatly reduce the number of decoding components needed. Both the 68020 and 68881 are pin-grid arrays; their pin references have been omitted for clarity.

required to support co-processor communication. Address strobe signal AS must not be asserted on the host board while co-processor communication cycles are taking place, and so the AS output from the 68020 is gated with the co-processor chip-select signal CPCS produced by the interface logic before being sent to the host board.

Co-processor selection. A co-processor communication cycle is classed as a c.p.u.-space cycle and indicated by all ones on the three function-code lines, as mentioned previously. Whereas the code on address lines A_{19-16} is all ones for an interrupt-acknowledge cycle it is '0010' for a co-processor communication cycle. A co-processor chip-select signal (CPCS) is produced by the interface logic when the code appears.

In conclusion, although designed for the Kaycomp, this platform board should work with any 68000 board provided there can be no bus master on the host other than the processor and that the memory and peripherals can cope with the shorter bus cycle.

CIRCUIT DESCRIPTION

Figure 2 shows the full circuit diagram of the platform board. You can see that the complete circuit consists of the two processors IC_1 and IC_5, two programmable-array logic (Pal) devices IC_2 and IC_3, a t.t.l. decade counter IC_4, a few pull-up resistors, and the adaptor, SKT_1, which connects the board to the host's 64-pin dil processor socket.

The object of this board is that the interconnector, SKT_1, appears to the host board as if it were a 68000, even though a 68020 processor is actually connected. In the case of the majority of the processor pins this means simply connecting a pin on the 68020 to its equivalent pin number for the 68000 on SKT_1.

Address pins A_{1-23} connect straight through, but note that A_0 cannot be routed through as this pin does not exist on a 68000. It is used though in generating the data strobe pulses. The upper address pins of the 68020, A_{24-31}, are ignored as these cannot be used in a 68000 system.

Sixteen of the thirty-two lines of the data bus are also connected straight through, but note that the most significant bit on the 68020, bit 31, should be connected to the most significant bit on the 68000, bit 15, to provide the simplest access mechanism.

Some of the control lines are also connected straight through, namely the interrupt lines, CLK, FC_0, FC_1, BGACK, BR, HALT, RESET, BERR and R/W. An extra connection may also be made to BERR from the co-processor chip-select via a link. This link is inserted on the Kaycomp to allow the firmware to detect the presence or otherwise of the co-processor, which is optional. Remaining processor outputs, FC_2, AS, UDS, LDS, E and VMA are all produced by the interface logic as there is no direct equivalent available from the 68020, and likewise for the two inputs DTACK and VPA.

The interface logic consists of IC_{2-4} and provides two basic functions, control-line modification and M6800 synchronous bus transfers.

Control line modification. A 16L8 PAL device, IC_2, handles the c.p.u.-space bus-cycle decoding. As discussed previously, on the 68000 when all function-code outputs (FC_{0-2}) are at logical one it indicates an interrupt-acknowledge cycle, but on the 68020 it can also indicate a co-processor communication cycle (and other conditions not used in this design). Kaycomp however assumes that if all three lines are one it is an interrupt acknowledge cycle and so it is presented with a modified FC_2 signal which only goes to one if the other conditions are correct for this particular cycle.

These other conditions are that A_{19-16} should also be at logical one and so these address lines, along with the three function-code lines, are all Anded together so that the modified FC_2 signal at IC_2 pin 12 only goes to logical one if all these seven lines are one.

These same address and function-code lines are also used to indicate a co-processor communication cycle which requires the generation of a co-processor chip-select signal (CPCS) to enable the 68881 maths co-processor. Code on address line A_{19-16} in this case though is '0010' and the logic generates a low at IC_2 pin 16 when this condition exists and the function code lines are logical one. Address lines A_{15-13} also indicate which co-processor is being accessed, but as there is only one in this design these are ignored.

When the main 68020 processor is accessing the 68881 co-processor then valid addresses appear on the address bus and the AS line is asserted. It must be ensured though

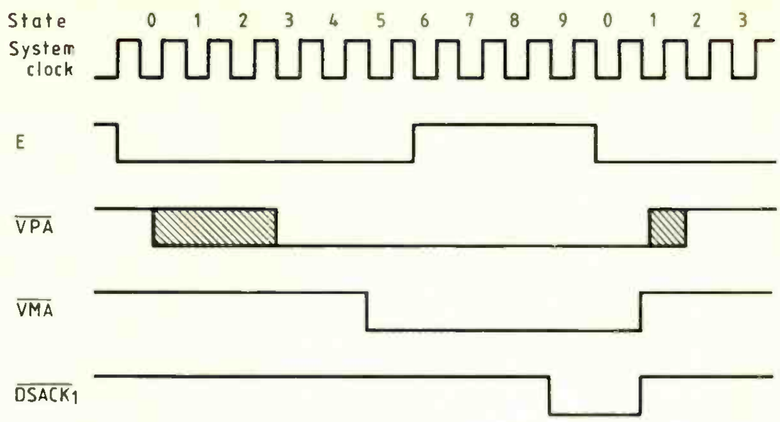


Fig.3. Timing for E-clock, valid memory address, valid peripheral address and data-transfer/size-acknowledge signals.

that the host board does not attempt to place data on the data bus at this time so the AS signal is gated with CPCS so that AS is not asserted if CPCS is.

The remainder of IC_2 is concerned with generating the two data-strobe signals, UDS and LDS, with the 68000 family, unlike some other 16bit processors, each address in the memory map signifies one 8bit byte and not a 16bit word. Consequently, as the 68000 can only access quantities of 16 bits, there is no need for an A_0 signal, the even address byte being considered to be the upper eight bits of the data bus and an odd address byte the lower eight bits. These quantities are strobed in and out of the processor by the upper and lower data-strobe lines (UDS, LDS) respectively.

However, the 68020, with its dynamic bus-sizing capabilities, can access bytes, words, three bytes, or 32bit long words at any base address, even or odd and so the A_0 signal is required and the actual framing of the external data bus connections indicated by the two size signals, SIZ_0 and SIZ_1, and finally strobed by a single data-strobe line DS.

This mechanism is fully explained in the 68020 User's Manual and so I shall just conclude that the equations for generating the UDS and LDS signals which are implemented in part of IC_2 are;

$$\overline{UDS}=DS+A_0$$

$$\overline{LDS}=\overline{DS}+A_0\cdot\overline{SIZ_1}\cdot SIZ_0$$

Remaining modifications of control lines are carried out in IC_3, also a 16L8 Pal device. Production of the E and VMA signals for M6800 synchronous bus cycles will be considered later. Signal AVEC, which is an input to the 68020 indicating that an autovectored interrupt service is required, is taken from the VPA signal from the host board, to which it is equivalent, but also gated with the modified FC_2 signal. This ensures that AVEC is only asserted during an interrupt acknowledge cycle and not during a M6800 cycle which is also indicated by asserting VPA in a 68000 system, this pin serving a dual purpose.

The data-transfer acknowledge signal from the host board, DTACK indicates the termination of a normal bus transfer. There are two equivalent pins on the 68020, $DSACK_0$ and $DSACK_1$, the combinations of codes on the two pins indicating an 8, 16 or 32bit port size. All transfers to and from a 68000 board must be 16bit and this is indicated by $DSACK_0=1$ and $DSACK_1=0$. Consequently $DSACK_0$ is pulled high by a resistor and $DSACK_1$ is driven by DTACK, but Anded with the two other possible sources of $DSACK_1$, co-processor IC_5 and the MC6800 transfer-generation logic also in IC_3.

This leaves us with the production of E and VMA to implement M6800 synchronous bus transfers. Consider how this is done on the 68000. When this type of transfer is required, to access a synchronous peripheral device say, the address decoding asserts low the 68000 VPA input. After recognition of VPA, the 68000 assures that E, which is produced by dividing the system clock by ten with a 60:40 low-to-high ratio, is low by waiting if necessary and subsequently asserts VMA until after E has gone through the next high-to-low transition.

To emulate this, a system clock divide-by-ten circuit is required which is provided by IC_4, and other logic contained in IC_3. The state machine produces the two 68000 signals E and VMA and the $DSACK_1$ signal to the 68020 to terminate the transfer, according to the state table, Table 1. The E clock is produced quite simply from three of the divider outputs, Q_b, Q_c and Q_d.

Table 1. Decoding state table for E, VMA and $DSACK_1$.

Output (Q) d	c	b	a	E	Action
0	0	0	0	0	
0	0	0	1	0	Negate VMA and $DSACK_1$
0	0	1	0	0	
0	0	1	1	0	
0	1	0	0	0	
0	1	0	1	0	Assert VMA if VPA asserted
0	1	1	0	1	
0	1	1	1	1	
1	0	0	0	1	
1	0	0	1	1	Assert $DSACK_1$ if VMA asserted

Two RS type bistable devices are used to produce the two outputs, VMA which outputs on IC_3 pin 17, and $DSACK_1$ which is Anded with other sources of $DSACK_1$ before appearing as an output at pin 18. Both of these bistable devices are reset at the first state, that is when $Q_a=0$ and $Q_c/Q_d=1$, negating (logic 1) the two outputs. At the fifth state, if the VPA input is asserted (logic 0) then the VMA bistable device is toggled, asserting (logic 0) VMA. At the ninth state, if VMA is being asserted, then the $DSACK_1$ bistable device is also toggled, asserting $DSACK_1$. One clock cycle later, after the counter has reset to state zero, both signals are once again reset. Resulting waveform timings are clarified in Fig.3.

CO-PROCESSOR INTERFACE

The 68881 is flexible in how it can be interfaced to the main processor, allowing 8, 16 or 32bit connections. Although an 8 or 16bit bus between the two would have reduced the number of tracks required, clearly maximum processing speed will result if all 32 data bits are used. Thus all 32 data-bus pins on the 68881 (IC_5) are connected to their equivalents on the 68020 (IC_1). The SIZE and A_0 pins of IC_5 are tied to a logical one (V_{cc}) which configures the device in 32bit transfer mode.

Since the 68020 has a built in co-processor interface, other connections are relatively simple; just a few address lines to select the various registers, control lines and a chip-select signal (CPCS) – in fact very similar to interfacing a normal peripheral chip to a processor. Both DSACK pins on the 68020 are driven by the 68881, a low on both indicating to the 68020 a 32bit bus transfer.

The beauty of this system is that all this is totally transparent to the software designer. Attaching the maths co-processor means to the programmer that a new set of instructions and registers have been added to the 68020's. The fact that they are contained in a separate chip is not apparent to the programmer, who just uses the additional instructions as if they were all part of the main processor.

PAL IMPLEMENTATION

When the prototype for this design was first built, it was done using standard t.t.l. for the interface logic, using similar basic circuits to those shown in the blocks for the Pals, Fig.7. Unfortunately it took about 10 t.t.l. devices to implement the interface logic, which made the board rather large and complicated. Programmable-array logic (Pal) devices were therefore used for the final design which reduced the chip count to three and considerably reduced the complexity of the board.

The principle problem with using programmable logic devices can be the equipment required, which could perhaps be difficult to justify for the small development laboratory or educational establishment. I found that with a little time and effort and a Pal distributor who could program the devices from paper it was quite possible to produce the designs with no specialized equipment at all.

PAL CONCEPT

The Pals used on the 68020 upgrade board are one of the simpler types, the 16L8, whose designation indicates logic low or inverted outputs (L) and 16 inputs and 8 outputs. For a 20-pin device, this number of input and outputs means that some pins are dual purpose, programmable as inputs or outputs, as can be seen from the block diagram of the device, Fig.4.

A full explanation of the workings of Pals can be obtained from the various manufacturers' data, but I will briefly explain the basics. A Pal implements the familiar sum of products logic by using a programmable And array whose output terms feed a fixed Or array. Since the sum of products form can express any Boolean transfer function, the Pal's uses are only limited by the number of terms available in the And-Or arrays.

Basic structure of a cell within the array is shown in Fig.5. Working backwards, this shows the Or gate which drives the output, being driven by two And gates, each of which has four inputs connected through programmable fuses to the true and inverted states of two inputs. In an unprogrammed Pal these fuses are intact, but by programming may be selectively blown, opening the circuit to that particular And input, which then assumes a permanent one state.

In practice, the arrays are much larger than this. With the 16L8 device, each And gate has 32 inputs, the true and inverted version of each of the device input pins. Driving each of the eight Or gates there are seven individual And gates. An eighth And gate enables an output inverter-buffer which drives the output pin from the output of the Or gate. A quick calculation reveals a total of 2048 individual fuses. This would be difficult to show by way of a circuit diagram and so a diagrammatic representation is used as is shown in Fig.6.

In this representation, at the point where the horizontal line to the And input (product term) crosses a vertical input line there is a fuse. We need to mark on this diagram crosses where a fuse is to be left intact and no mark where a fuse is to be blown. This 'fuse-plot' can then be handed to the supplier/programmer of the device to pro-

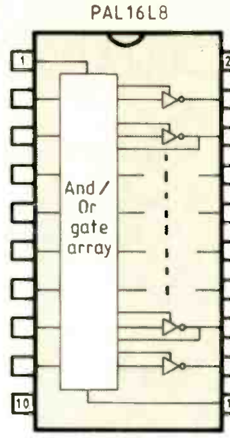


Fig.4. Up to 16 inputs and 8 ouputs are available on the 16L8 programmable-array logic i.c.

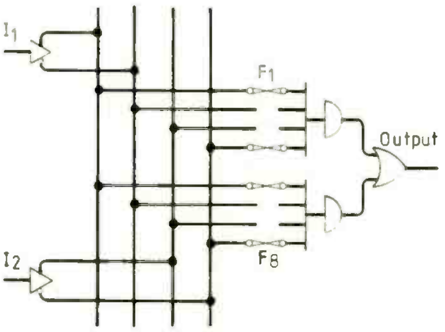
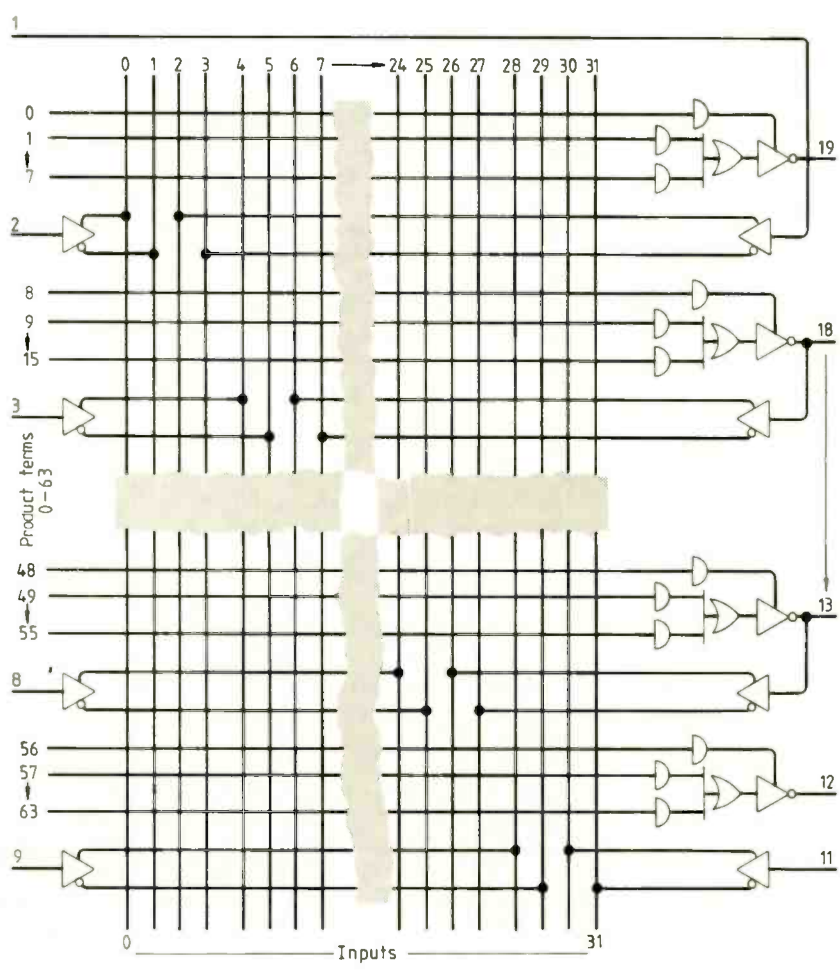


Fig.5. Basic Pal structure for a two-input, one output logic segment.

Fig.6. Logic diagram for the PAL16L8. Eight of the lines on the right can be programmed either as inputs or outputs.



duce a Pal to your custom design. Clearly this will need to be entered into the device programmer manually, which takes time, and hence money. For each of the devices used for this project a charge of around £24 was made. This is a non-recurring cost though as when further devices are required, the first device can be used as a master, saving this charge.

DERIVING THE FUSE-PLOT

We now have a circuit diagram and this has to be translated to a fuse-plot. There are 'schematic capture systems' available to run on various computer systems which can take a circuit diagram and produce the necessary output to the device programmer from which the Pal can be programmed – clearly the best solution, but expensive. A simpler, more readily available software package is a Pal assembler (Palasm) which runs on IBM PC compatibles. This takes the Boolean expressions which define each output and produces the information for programming the device. Even these days though, not everyone has an IBM PC, let alone PALASM, so for this project a manual system had to be devised.

The first question to be answered is can the interface logic be contained in one Pal or if not how many? The main restriction will be the number of inputs and outputs required. I soon saw that more than one would be needed, and after some juggling around the logic was fitted into the 16L8 Pals and one t.t.l. decade counter. The decade counter too could have been implemented in Pal, using a registered output device, such as the

MONITOR ENHANCEMENTS

In order to accommodate the 68020's enhancements and the maths co-processor, the original 68000 monitor with line-by-line assembler has been supplemented with the following commands and modifications.

AS line-by-line assembler. Modifications – 'size' attribute for branch instructions is now 'B', 'W', 'L' for 8, 16 and 32bit offsets respectively, instead of 'S' and 'L'. Default start address is now 400600_{16}. All ram below this address is used by firmware.

CA examine/alter cache address register. Same as An command but for cache register address.

CC examine/alter cache control register. Same as An command but for cache control register. Note that only bits 0-3 are used.

DF examine/alter destination function-code register. Same as An but for destination f.c. register. Only bits 0-2 are used.

Fn examine/alter floating-point data register. Same as An command but for the eight f.p. data registers. Note that this is a 96bit value (24 hexadecimal characters).

FA examine/alter floating-point instruction address register. Same as An command but for the f.p. instruction-address register.

FC examine/alter floating-point control register. Same as An but for the f.p. control register. Only bits 0-15 are used.

FD disable floating-point register printing.

FE enable floating-point register printing.

FS examine/alter floating-point status register. Same as An command but for f.p. status register.

RD register display. Modifications – displays extra registers available on the 68020 and the 68881 registers provided that it is enabled by the FE command.

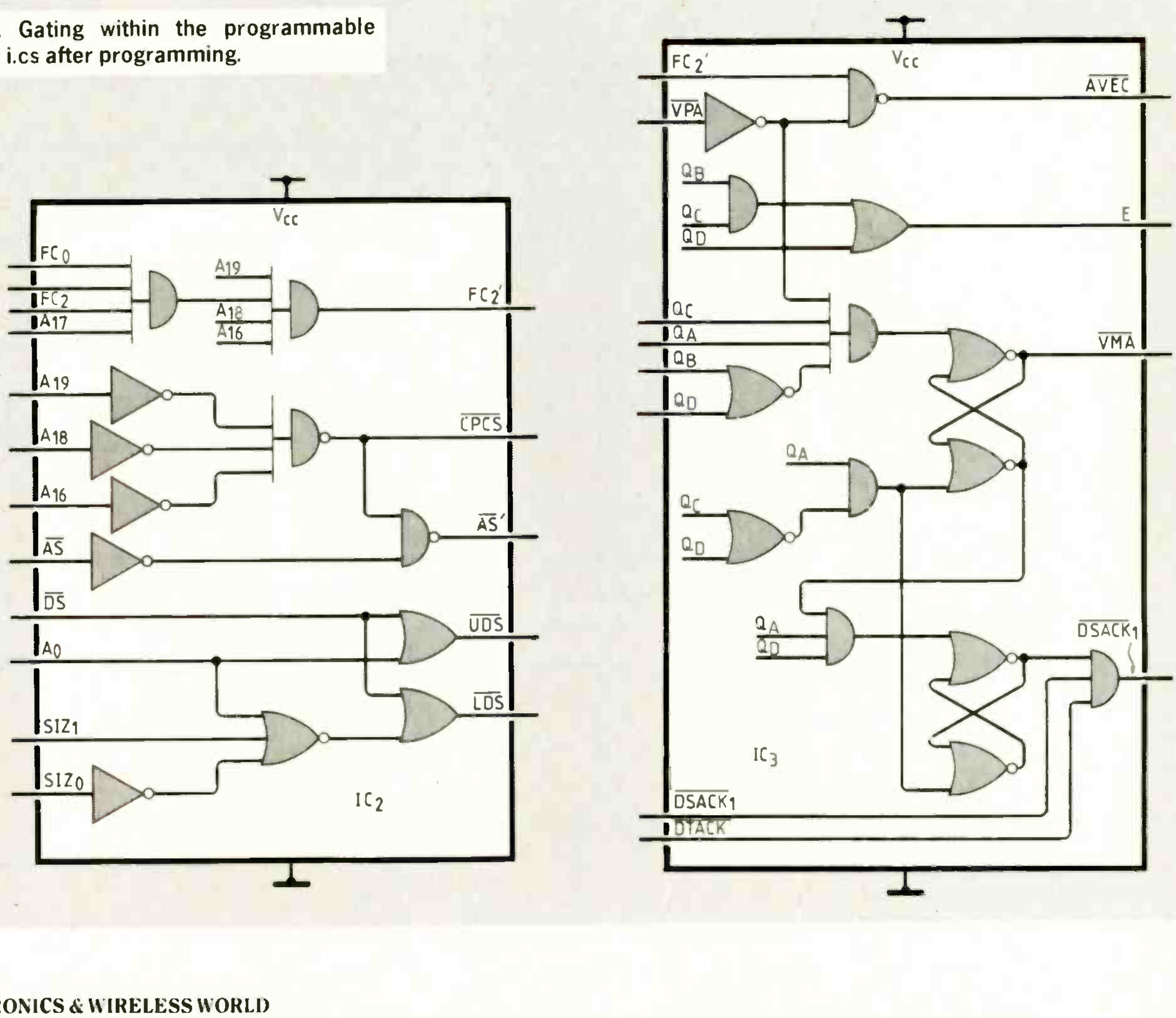
SC single-step on flow change. Similar to SS command but only stops and displays registers on a program flow change, i.e. when a jump or branch is executed.

SF examine/alter source function-code register. Same as An command but for source function-code register. Note that only bits 0-2 are used.

TC trace on flow change. Similar to TR but only displays registers on program flow change.

VB examine/alter vector base register. Same as An command but for the vector base register.

Fig.7. Gating within the programmable logic i.cs after programming.



16R6, but the configuration used proved to be just as simple and more cost-effective in this instance.

The circuit required to produce one of these Pals, IC_2, is shown in Fig.7(a). You can see that 12 inputs and 5 outputs are needed, which is within the capabilities of the device. Next I had to decide which pins were to be allocated to each function. One of the great things about programmable logic is that you can assign the pins to suit your p.c.b. layout, making this easier and more flexible. The allocation chosen is shown in Table 2.

The circuit diagrams we have for each output of the device though is not in a form

Table 2. Allocation of pin functions for Pal IC_2.

I/O	Function	Pin	Pin	Function	I/O
I	FC_0	1	20	V_{cc}	–
I	FC_1	2	19	$\overline{AS}'$	0
I	FC_2	3	18	$\overline{UDS}$	0
I	A_{16}	4	17	$\overline{LDS}$	0
I	A_{17}	5	16	CPCS	0
I	A_{18}	6	15	LDS	I/O
I	A_{19}	7	14	SIZ_0	I
I	$\overline{AS}$	8	13	SIZ_1	I
I	$\overline{DS}$	9	12	FC_2'	0
–	GND	10	11	A_0	I

which reflects the internal structure of the Pal and so the next process is to translate the circuit diagram for each output of Fig.7(a).

Internally, this Pal has four gates in series, the appropriate inputs of each one of which must pass through to produce the output. These are, first either an inverting or non-inverting gate, an And gate, an Or gate, and finally an inverting buffer which may be tri-stated if desired. In this design all used outputs are permanently enabled.

Taking the simplest block first, the UDS output, from Fig.7(a) you can see that this output must go low if both DS and A_0 are low, and high at all other times, i.e. an Or gate. How this is implemented in the Pal is shown in Fig.8. Two of the inputs to the And gate are used and driven from DS and A_0 after each is inverted. Unused And inputs must all be tied permanently high for the gate to function and this is done by blowing all the remaining fuses in that product term.

Thus the And output will be high only if both DS and A_0 are low. To get the required output from this merely requires inversion, the Or gate not being required. If the output from the And function is applied to one input of the Or gate and all the other inputs are low, then the Or gate will act as a non-inverting buffer and have no effect. The output inverter is permanently enabled, producing the required output.

With the circuit diagram in this form, we have the logic function defined in the same manner as the internal structure of the Pal and the fuse plot diagram can now be marked up. Figure 9 shows the section of the fuse plot for the UDS output. The top And gate drives the output buffer which is to be permanently enabled. This is done by blowing all fuses on its input, denoted by no x's on that product term line.

For the next And gate shown, the one that is used, x's on inverted input lines from DS and A_0 indicate these fuses are intact, while all other fuses are blown. The remaining six And gates are not used and must output a logic low to the Or gate. If the fuses for both

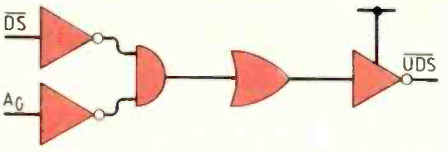


Fig.8. Pal equivalent circuit for the upper data-strobe.

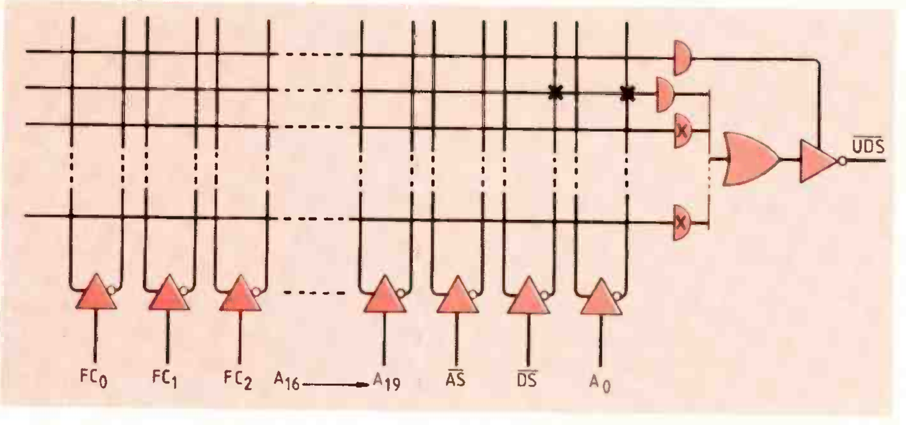


Fig.9. Fuse plot for upper data-strobe signal decoding.

the true and inverse of an input are left intact to an And gate then the output must always be false (low) whatever the state of the input. So all fuses to the unused And gates are left intact and this is indicated by placing an x inside the And-gate symbol.

This output was a fairly simple one, but the LDS output is a little more complicated, Fig.10. You can see that the first output gives the inverse of what is required and so this output, which is also an input, is fed back in and inverted to finally give the required function. The first output pin is left unconnected externally. Figure 11 shows the fuse-plot equivalent of this.

FIRMWARE MODIFICATIONS

In order to accommodate the new processors that can now run on the Kaycomp board, the firmware has been extended to cover the extra features. The assembler and dis-assembler have obviously had to be extended to cover the extra instructions and addressing modes. And the monitor has had to be altered considerably to enable the extra registers to be accessed individually by various extra commands and also displayed by the 'RD' command.

Since the 68881 maths co-processor is optional, the monitor also has to determine

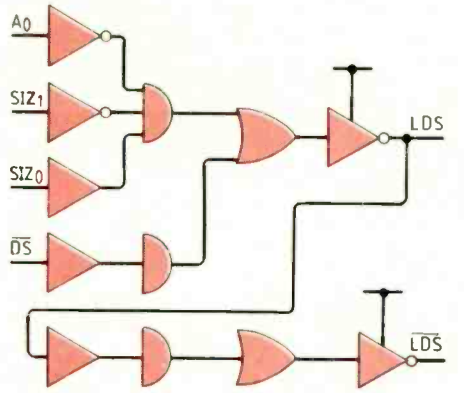
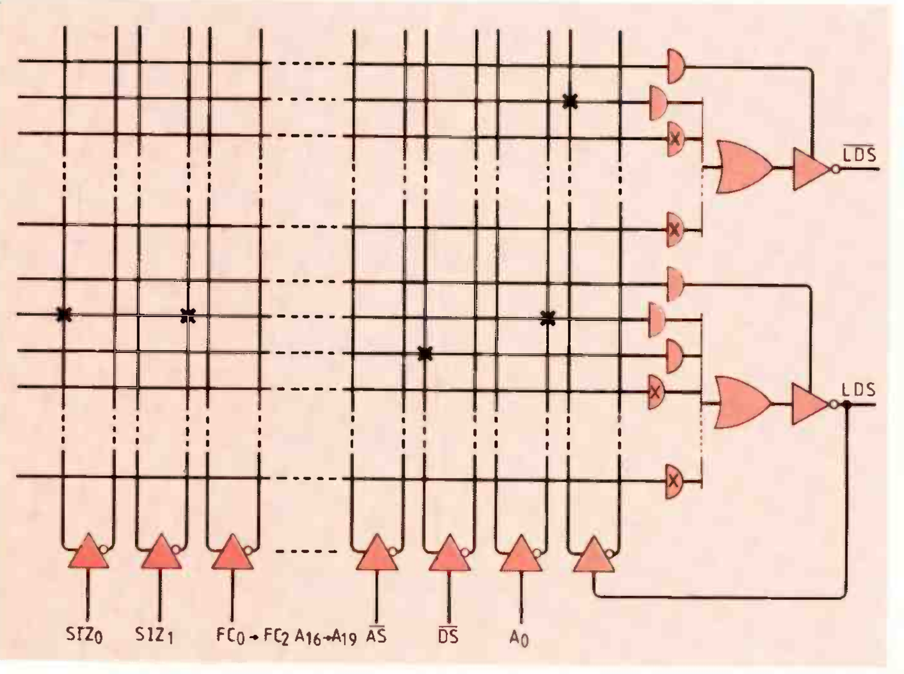


Fig. 10 Pal equivalent circuit for the lower data strobe.

Fig.11. Lower data-strobe output fuse plot.



at reset whether or not it is fitted. If it is not fitted, a link from the co-processor chip-select line to the 68020 bus-error input, BERR, must be fitted (LK_1) which causes a bus error if a co-processor access is attempted. At reset this is intercepted and used to prevent further access by the monitor of the co-processor, instead of causing normal bus-error exception processing.

Otherwise the monitor has been kept the same wherever possible. One alteration is that the 40-column display mode has been removed, the reason being the extra number and size of the registers that have to be displayed make it really unworkable with anything less than an 80-column display. Also consider the maximum length of instruction that now has to be handled by the assembler and disassembler, for instance,

```
00400612 33DO19370012345600056789213300ABCDEF00FEDCBA\
MOVE ([$123456,A0],D1.L,$56789),([$ABCDEF,A1,D2.W]$FEDCBA)
```

which is difficult enough to make clear with an 80-column display!

All these extensions have virtually doubled the size of the Kaycomp firmware package which is now about 60Kbyte. Eproms therefore must be at least 32kbyte each, so 27256 devices must be used.

ASSEMBLY AND INSTALLATION

Because of the use of Pals the board itself is quite simple and requires very little assembly. Parts for this design are available from Magenta Electronics Ltd whose address is given later. These include a high-quality six-layer plated-through-hole p.c.b. which makes the design very compact and reliable. There are only ten components in all to be mounted on the board and their positions are shown by a silk-screened legend. Remember though that the 64-pin board-to-board adapter is fitted on the underside of the board and soldered on the top (component) side. This adapter has pins which are a larger diameter at one end than the other. The larger diameter ends must be soldered into the new board. Also, if the 68881 is not being fitted at this stage, wire together the two holes of LK_1.

To fit the kit, the new 27256 Eproms should be fitted first after altering the links on LK_1 on the Kaycomp to suit 27256 eproms (see November 1985 issue, page 53). Next remove the 68000 i.c. from the Kaycomp and in its place plug in the new board, ensuring it is the correct way round. The boards are now ready for powering up.

Getting the new upgraded Kaycomp working is probably the easiest part of upgrading. The harder part is upgrading the user to be able to program the two new processors, particularly the maths co-processor which requires new ways of working for the programmer but has great potential for real-time mathematical applications[1].

Two books therefore will be essential to the new user. These are the manufacturer's user manuals for the 68020[2] and the 68881[3] processors.

Magenta Electronics is at 135 Hunter Street, Burton-on-Trent, Staffordshire DE15 0AB, tel: 0283 65435.

References

1. Using 68020 co-processors, Burns and Jones, *E&WW* May 1987, pp.535-537.
2. Motorola MC68020 32-Bit microprocessor user's manual, part number MC68020UM/AD.
3. Motorola MC68881 floating-point coprocessor user's manual, part number MC68881UM/AD.

Bob is now Development Manager for computer systems and consultancy company, Barron McCann.

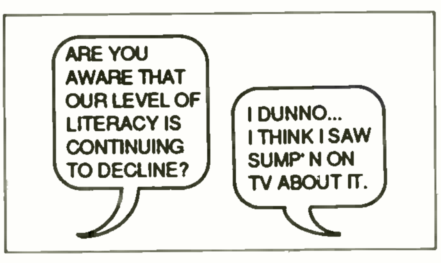
BOOKS

Numerical recipes in C: the art of scientific computing, by William H. Press, Brian P. Flannery, Saul A. Tukolsky and William T. Vetterling. Cambridge University Press, £27.50. Comprehensive handbook of scientific computing in the C language, the follow-up to an earlier work devoted to computing in Fortran and Pascal. Its 17 chapters and 200-odd program examples cover such areas as linear equations, differential equations, interpolation and extrapolation, random numbers, sorting, finding roots, eigensystems, FFTs, statistical methods and data modelling. Program examples are commented and fully explained in the accompanying text. In keeping with the informality of their title, the authors present their material in a sleeves-rolled-up, practical, and very readable style, and are not above tossing in the occasional touch of humour (illustrating one of their sections on C control structures is a program called 'badluk', which calculates occurrences of Friday the 13th when the Moon is full). Useful reading lists follow each section. C programmers will undoubtedly find this book invaluable; and in view of the sheer bulk of information it contains, they will probably think it very reasonably priced too. Hard covers, 755 pages. Recipes and examples are available from the publishers on disc (IBM PC format).

Electronics and computer acronyms (revised second edition) by Phil Brown. Butterworths, £24. Glossary of electronic technology from A for atto (10^{-18}) to Zr for zirconium: 2500 entries covering every branch of the art, including the commonplace, the esoteric and a few surprises. For example, none of us in the office knew until now that BNC (as in connector) stood for Bayonet Nut Coupling. This sent us scurrying to look up SMA – but alas, it wasn't included. One error we spotted is that CTS, in serial communications links, stands for clear-to-send, not clear-to-stand. Useful, if expensive. Hard covers, 272 pages.

Computer Lib, revised edition, by Ted Nelson. Tempus (Microsoft Press, Penguin Books), £14.95. "This book is a multi-user, high-resolution, demand-paged, hardware-software, read-mostly memory, retrieval and display system with real-time interaction, tactile interface, audio and video feedback." Get the idea? Its cheerfully anarchic typography occupies territory somewhere between early *Private Eye* and *The Face*, and the scrapbook-style pages are a diverting mixture of computer facts, fantasy, ideas, opinion, jokes, cartoons, eccentricity, wisdom, whimsy and aphorisms (sample: "Wordstar is the second hardest video game yet made").

THE DECLINE OF MIND



In fact, it's *two* scrapbooks – the second half, entitled "Dream Machines", is printed upside down, working from the back cover inwards, which means that two readers can enjoy the book at once (this must be the multi-user angle). Besides the humour, there's plenty of serious analysis of the computer world: a favourite target is the industry's *éminence bleue*, IBM, which comes in for a good deal of criticism over its business policies. "A computer cult classic returns", says the blurb-writer, and who could dispute that? Enormous fun to dip in to. Large format paperback, 235×247mm, 305 pages.

• Following the appearance of our Research Notes item, "Building with atoms" (May 1988, page 452), a reader draws our attention to a book by an MIT visiting scholar, Eric K. Drexler, called "Engines of Creation".

This, he writes, discusses the full implications and implementation of this technology, in all fields, from mechanical engineering through computing and telecommunication to medicine.

Nanotechnology, the building of devices using individual atoms and molecules as working parts, is likely to make a major and profound change to humanity. Drexler discusses its use for good or ill. A real age of plenty could result. The medical implications are profound, leading to cures of cellular diseases such as cancer and ultimately to the immortality of the individual and possible revival to youthful good health of those being frozen upon death today.

"Engines of Creation" also discusses possible abuses of the technology, and the implications of its effects on present problems of society. The book is published by Doubleday at just under $18. If any reader has difficulty in getting a copy through the usual channels, a limited number of copies are available at £11.20 each, post free, from John de Rivaz at West Towan House, Porthtowan, Truro, Cornwall TR4 8AX. (Despatch by first class post within 28 hours; your money back if sold out.)

Opto first from Ericsson

Ericsson's researchers in Stockholm have developed a new integrated optical switching device that can be interfaced directly to standard single-mode optical fibres. This polarization independence is claimed by Ericsson to be a world first and to represent a significant advance on earlier switching matrices.

A further interesting feature of the new switch is that it can be used as a non-blocking point-to-point switch; but the switch architecture means that it can also be used as a broadcast switch, with one input channel able to broadcast to all outputs.

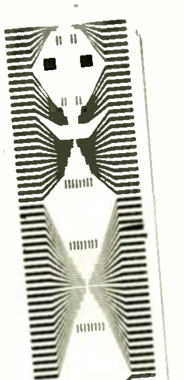
The device, demonstrated as a laboratory version, is a 4×4 channel switch matrix, implemented in lithium niobate and operating at the telecom industry standard wavelength of 1300nm. Switching is executed entirely in the optical domain, under the control of electronic signals.

Ericsson's new device uses a completely different switching mechanism from the first 8×8 channel integrated optical switch matrix unveiled by the company two years ago, although that too was implemented in lithium niobate.

The new device uses an optical switch technique that can be electronically tuned so that high performance is achieved with lower fabrication tolerances. This, coupled with an extremely low crosstalk between channels of less than −35dB (optical), augurs well for the development of more complex photonic switching networks.

At present the data stream on an optical fibre transmission route must first be converted to an electronic signal before it can be switched to another route. This imposes a severe restriction on the bandwidth and coding format of the information stream to be switched.

The attraction of development work into optical switching is the long-term promise of communication networks where transmission and switching takes place entirely in the optical domain. Optical space switching promises virtually unlimited bandwidth, and transparency to coding format. The ultimate aim is what are called "coherent optical networks".



System X enhancement

Under the System Enhancement Programme 2 (SEP2), the present installed base of System X telephone exchanges is being enhanced by the manufacturer GEC Plessey Telecommunications (GPT) to allow the new integrated services digital private branch exchanges to interoperate with the digitized public network, via a signalling system known as Digital Access Signalling System 2 (DASS 2). Other benefits of the upgrade include X.25 direct computer-to-computer interfacing links for more efficient network management and quicker reaction to operating statistics.

SEP2 is possibly one of the most complex improvement programmes undertaken in telecommunications. It is believed to be the world's largest progressive modification ever attempted to a real-time on-line computer-based telephone network.

Hull telephone department (now known as Kingston Communications p.l.c.) is the largest System X user outside British Telecom, and was in the forefront of this enhancement programme. It achieved a complete and smooth transition by the end of December 1987.

In total around 50 processor exchanges have been modified so far and this will rise to 230 by the end of 1988. SEP2 forms part of the progressive series of enhancements which are an inherent feature of System X's design evolution strategy known as 'future proofing'. This allows the incorporating of new facilities and function as they become technically achievable or are demanded by the customer.

One of the many private steam railway companies in the UK is on the Isle of Purbeck in Dorset, where the Swanage Railway Company is busy restoring the old line from Swanage to Corfe Castle and Wareham. One of its main areas of work lies in the establishment of modern signalling and communications systems, not only for normal operation of the trains, but to satisfy the Department of Transport's safety requirements. These include, for example, the installation of railway company telephones on both sides of every level crossing and farm crossing, as well as train-activated platform bells and all conventional signalling needs.

The system being installed along the trackside uses Duratube & Wire's Dropwire No.10 – a twin-pair telephone cable fitted with steel strength members for overhead distribution capability and sheathed with solid polyethylene to BS6234.

Telecom Pocketbook

Philips Business Systems has published 'The Philips Pocketbook of Telecommunications'. It is intended to be a serious reference book covering virtually every aspect of the UK industry and technology, and is published at a time when many changes are happening in telecommunications.

Topics covered range from telephones and p.a.b.xs to signalling and i.s.d.n. Both cellular and data communications are given prominence, while information on installation and maintenance will be of practical help to users. The various public networks and operators are described together with regulatory authorities and the regulations they supervise. Of particular significance and usefulness is the section on approvals.

Copies of the booklet are available free of charge from Philips Business Systems, Elektra House, Bergholt Road, Colchester, Essex, CO4 5BE; telephone 0206-575115.

Fibre link record

A record-breaking £6 million optical fibre cable between Britain and the Isle of Man has been inaugurated by a video call between London and Douglas. The cable spans the entire 90km (56 mile) route without intermediate boosting, making it the

TELECOMMS TOPICS

longest 140Mbit/s unregenerated system in Europe – and probably the world – in commercial service; though not for long (see "UK-Ireland opto link", below).

The new cable, supplied by STC Submarine Systems, is the latest step in the programme to convert the Isle of Man network to digital operation by 1990. Manx Telecom Ltd, the British Telecom subsidiary responsible for running the Isle of Man's telecommunications services, has already installed 126km of optical fibre in links between its exchanges on the island.

Within the cable are six fibre pairs, five of which are being brought into immediate use at 140Mbit/s – equivalent to nearly 2000 simultaneous phone calls per pair. The cable augments analogue and digital links with the Isle of Man.

It is the first application by STC of 1535nm distributed feedback single-line lasers in submarine system terminals. These lasers have a very narrow spectral width, and may be used at this longer operating wavelength where the fibre attenuation is much less. As a result, systems well over 100km in length may be constructed without the need for submerged regenerators.

To minimize any risk of damage to the link by activities such as trawling, the cable was buried during laying by using British Telecom's special purpose seabed plough.

UK-Ireland opto link

A submarine fibre cable has been successfully laid by British Telecom between the UK and Ireland, at a cost of £7.2M.

When it comes into service this summer, it will become the longest optical unrepeatered fibre cable in use. It runs 126km (80 miles) between Porth Dafarch, near Holyhead, Anglesey, and Portmarnock in the Irish Republic.

Jointly owned by British Telecom and Telecom Eireann, the cable has been supplied by French company Submarcom. It is being installed under modernization programmes which are bringing to customers the benefits of digital services and will carry speech, data, text, graphics and facsimile.

The cable has 12 fibres of which six will be put into use immediately the cable comes into service. Each pair of fibres will support a system operating initially at 140Mbit/s. It will use the latest single-mode technology with high performance 1500nm lasers.

BT to improve packet service

British Telecom is to enhance its public packet data services, Packet SwitchStream (PSS), by introducing a range of new customer facilities during the year. Further benefits will come from improvements to be made to the International Packet Switching Service (IPSS).

Enhancements to PSS will follow BT's implementation of the 1984 recommendations for expanding the CCITT X.25 protocol. The principal enhancements initially include: increasing the permitted length of the facility field in call set-up and call clear packets for data terminating equipment (DTE) from 63 to 109 bytes, providing capacity for additional facilities; enhancement for fast select calls so that user data can be carried in the call request packet, even when the call has entered the data transfer phase; and notification of call redirection, by which a called DTE is informed that it is chosen as an alternative destination for a call.

Other enhancements include DTE-originated cause codes, expanded interrupt user data field, hunt group address replacement, called line address replacement.

Payphone monopoly abolished

Oftel has published its technical requirements for the approval of payphones. Up to now, payphones installed on private premises such as shops, restaurants, garages and public houses could only be obtained from BT. These new requirements effectively abolish BT's monopoly and will enable payphones to be evaluated for type-approval and – if approved – offered directly to users by suppliers other than BT.

Southwestern Bell Telecom UK, a subsidiary of the regional Bell operating company in the USA, is aiming to be the first company to take advantage of this change.

Its PP1000 desktop programmable payphone (pictured below) will, once it has been approved, be on sale in high street shops at £215+v.a.t. It does not require any special "payphone" tones from the exchange and so can be plugged into any BT telephone socket. It is supplied pre-programmed, but the user can readily alter the charges. For example, with a payphone installed by a company for staff to make private calls, or in a hospital for the convenience of patients, the charges would probably be set much lower than when the unit is sited in a pub or wine bar.

Currently some 300 000 private payphones are rented from BT, but according to SBT-UK's managing director, Stephen Carter, the total potential market could be as large as 800 000. This estimate is based on an analysis of the total number of shops, garages etc. of a suitable size to employ such a device. However, he expects sales to reach about 50 000 units over the next year, with his company supplying over half.

V.42 protocol finalized

After more than three years of extensive development and negotiations between interested parties, CCITT V.42, error-correcting procedures for modems using asynchronous to synchronous conversion, has been finalized. It will go before the full CCITT for formal adoption next November in Australia, following agreements recently reached by the Plenary Modem Study Group XVII in Geneva.

Known as LAPM (link access procedure-modem), the primary protocol in the V.42 standard is based on the link access procedure-D channel (LAPD), used on the i.s.d.n. 'D' signalling channel) international standard. Both LAPB (used for error control in present Hayes V-series modems) and LAPD are based on the high-level data link control (HDLC) procedures.

Basing the primary protocol for modem error-control on the LAPD protocol adopted by ISO and the CCITT supports the position Hayes and many others favoured during the development of the V.42 standard. By developing LAPM, the CCITT has provided a firm foundation for continuing work on advanced modem features such as data compression, remote configuration and use of multiple virtual circuits. The new V.42 and the existing X.25 packet standard provide direction for the continuing evolution of dial-up data communications.

At the same time V.42 includes support for the well-established Microcom MNP protocol. This ensures that existing users of this *de facto* standard are not left out in the cold even though the progression is via the HDLC route.

Telecomms Topics is compiled by Adrian Morant.

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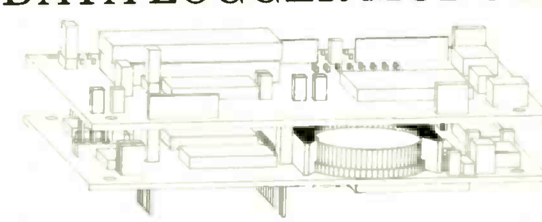

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74H20N	NSC	.19
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74S08PC	FSC	.08
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74S240N	SIG	.11
SN74S241N	TI	.15
UA7812UC	FSC	.09
UA7912UC	FSC	.09
N8T245N	SIG	.11
N8T26AF	SIG	.11
MC8T26AP	MOT	.11
N8T97N	NSC	.11
N9401N	SIG	.81
96LS02DC	FSC	.30
MICROPROCESSORS		
WD1015PL-00-02	WD	6.75
FDC1797	SMC	1.62
WD1943M00	WD	1.35
D2147D2	NEC	.46
COM2601	SMC	.54
AI2625-5	BURR BROWN	.38
SCB2673BC5N40	SIG	.81
D2758	INT	.27
TMS2764-25JL	TI/SGS	1.54
TMS27C256-25JL	TI	2.16
TMS2732A30JL	TI	1.72
AM27S181DC	AMD	.65
TBP28L22N	TI	.97
AI4437	AMICON	1.62
MM5290N4	NSC	.15
SSI580CP	SSI	2.16
R6500-IEAB3	ROCK	14.04
R6503AP	ROCK	1.30
MK68000P6A	MOSTEK	2.70
S6802P	AMI	1.35
HD6845P	HIT	.97
AD7503JN	AD	1.08
D8048C	NEC	.46
M110B1	SGS	2.16
Z80HCPU	SGS	.97
P8224	INT	.65
D8259AC	NEC	.54
P8287	AMD	.70
P8748H	INT	2.97
N8X305N	SIG	4.32
TRANSISTORS & SCR's		
TIC206D	TI	.15
TIC216D	TI	.15
TIP2955	TI	.23
GI752	GI	.11
DIODES		
1N4001	ITT	.01
1N4005	ITT	.01
1N4006	ITT	.01
1N5225BRL	MOT	.05
1N5818	MOT	.06
1N6263	ITT	.04
1N753A		.02
OPTOS		
H21A2	GE	.41
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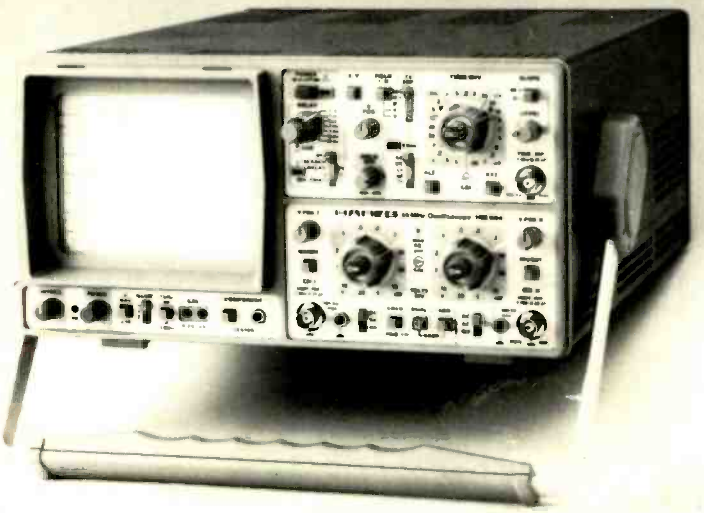

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Pioneers

19. Almon Brown Strowger (1839-1902): inventor of the automatic telephone exchange.

W.A. ATHERTON

The "girl-less, cuss-less, out-of-order-less, wait-less telephone", was how Strowger's invention of the automatic telephone exchange was described. "I am often told that the telephone girls will be angry with me for robbing them of their occupation", Strowger said at a public demonstration of the first such exchange. "In reply I would say that all things will adjust themselves to the new order . . . The telephone replaced the messenger boy as this machine now displaces the telephone girl".

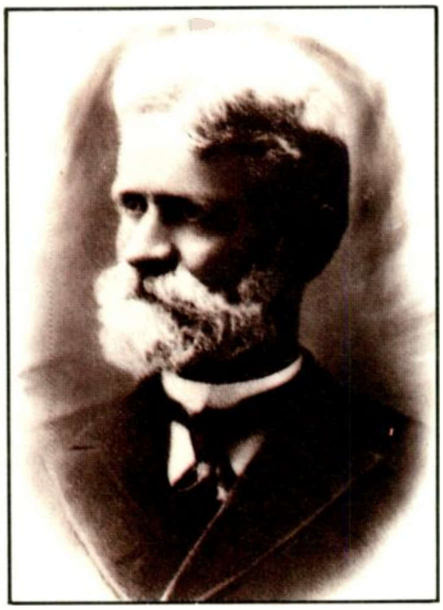
But not every telephone girl gave up without a fight, as one salesman discovered in North Carolina – where a town attorney feared for his girl-friend, the local telephone operator. Strowger's man lost the order because, the salesman was told, the automatic telephone "had no sex appeal".

The popular story behind the invention of the automatic telephone exchange has been told often. Strowger, at that time an undertaker in Kansas City, read in the local newspaper of the death of a friend and learned with surprise that the funeral arrangements were to be carried out by a rival. Always a man of quick temper he jumped to the conclusion that the lady telephone operator had connected the grieving relatives to his rival when they had really wanted to be connected to himself. This he judged to be either because his telephone was engaged at the time or for some more mischievous reason. His complaints brought denials and no satisfaction; and he turned to wondering how telephone callers might automatically connect themselves to their desired number without needing an unreliable operator.

Perhaps having seen a manual exchange, he soon conceived, in 1887, the basic idea of what he wanted and made a mock up by sticking pins into a circular cardboard collar box. (If you do not know what a collar box was, ask your grandfather!) Soon he had ten rows and ten columns of pins, representing contacts to 100 telephones. He envisaged a pawl and ratchet mechanism driven by pulsed electromagnets which would lift a wiper (which was connected to the caller's telephone) to the required row and then move it round to the required column. There the wiper would make contact with the distant telephone. A US patent for a 1000-line "automatic telephone exchange" was applied for on 12 March, 1889 and granted almost exactly two years later, number 447,918.

Strowger had little if any knowledge of electricity or mechanics and so his nephew, Walter, was given the task of making a working model. Even with the help of a competent jeweller, a 1000-line switch proved beyond him. Instead, the first Strowger automatic switch consisted of just 100 contacts arranged in a single row inside a cylinder. It was exhibited at the Bell telephone exchange in Kansas City.

Fortune now smiled on Almon B. Strowger. Having failed to interest financial backers, Walter Strowger was approached by an enterprising salesman called Joseph Harris, on the look out for ideas which he could promote at the forthcoming World's Columbian Exposition. This was to be held in Chicago in 1892.



Almon B. Strowger (GTE Communications Systems).

Two years before the Exposition, Harris persuaded Strowger and his nephew to go to Chicago. There the Strowgers, Harris and a friend, M.A. Meyer, agreed to develop the system. Further financial backing was found, and on 30 October, 1891, the Strowger Automatic Telephone Exchange was incorporated in the state of Illinois. With a salary of $100 a month, Strowger's days as an undertaker were over.

The automatic switch was exhibited at the Chicago offices of a railway company where Meyer worked. A firm of model makers was asked to make samples and work began on further development of the switch. Perhaps more importantly a telegram was sent to Baltimore to one Alexander Ellsworth Keith, electrician, asking him to come to Chicago at their expense. Keith came, saw, understood, and accepted an appointment as chief engineer. He proved to be a good choice.

NO NEED FOR GIRLS

The world's first automatic telephone service was opened in the autumn of 1892 at LaPorte in Indiana.

In 1890 the Bell Telephone Company had won a suit against the telephone system in LaPorte for infringement of the Bell patents. The judge ordered that all the telephones be burned! Presumably for the next two years LaPorte had no telephone service, enabling Strowger to obtain a franchise there in 1892.

For $25 a month the Strowger Automatic Telephone Exchange rented premises above a shop, sub-letting spare space for $15 a month. It was a reasonable start to business. By September wiring work had begun, and in mid-October the first telephones and automatic switches were installed. The service was up and running, serving a mere 52 subscribers.

On 3 November, 1892, was held a grand public demonstration. Formal invitations were sent out and a special train was laid on to bring guests from Chicago. About 50 or 60 were met by the mayor and the city band. At the telephone exchange they saw shelves carrying 80 of the new Strowger switches together with batteries. There were no "hello girls" (operators). The reporter from the Chicago *Daily Inter Ocean* could not get over the fact that the solitary employee present had "nothing, absolutely nothing, to do".

Even if *he* had nothing to do, the poor subscribers certainly had. From 12.30 to 4p.m. they were busy answering their telephones "very promptly" as the invited guests tried out the system. For Strowger and his companions the day was hugely successful and the Chicago newspapers duly reported the invention. "No need for girls" was a typical headline.

The dial telephone was an invention made by Strowger's engineers and it stood the test of time, until eased aside recently by push-button telephones. It is an historical curiosity that Strowger's original telephones also had push buttons. With less than a hundred subscribers the Strowger team simply provided two buttons. To call, say, number 37 the caller pressed the 'tens' button three times and then the 'units' button seven times. Although it worked well it was prone to errors as callers lost count!

Even as the LaPorte system was being publicly demonstrated, a second installation was under construction at Fort Sheridan in Illinois. By the end of 1896 Strowger exchanges had been installed in other towns in Indiana, and in Iowa, Minnesota, New Mexico, Colorado, New York State and Wisconsin.

In those early years almost every installation was a field trial, but by 1895 Strowger engineers (for Strowger himself played relatively little further part) had settled on the basic step-by-step 'up and round' motion. This came to characterize the Strowger switch which, of course, underwent many improvements. At LaPorte five wires were used: three for signalling, one for releasing and one for the conversation.

The dial was introduced in 1896 to reduce the number of mistakes made by callers, and the first dials used flanges instead of the later finger holes. Even dials gave problems. German settlers in Wisconsin pronounced '21' for example, as "one and twenty". So they dialled 1, then 2 and then possibly zero. An education programme on how to count in English solved the problem.

TRUNK DIALLING

In 1897 the concept of 'transfer trunking' was introduced whereby calls were directed through a series of switches. This meant that the size of the exchange was no longer limited to the size of the individual switch, which previously had provided contact to every telephone on the exchange.

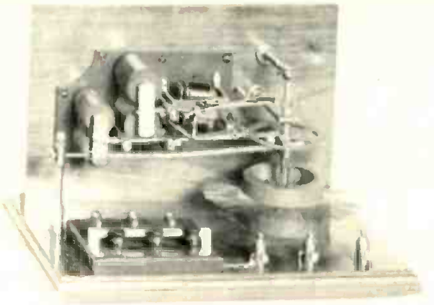
Strowger's love affair with automatic telephony was short but very sweet. This major invention had been made by a man in his 48th year. (Who says you are past it at 40?) By the late 1890s his health was beginning to fail and he retired from the company which had been founded on his invention. The company, however, went from strength to strength and by the turn of the century it had opened operations abroad. In 1901, in effect, it became a new company: Automatic Electric. In 1955 Automatic Electric became part of General Telephones and Electronics and it is now part of GTE Communication Systems.

There was still life in Strowger himself however: for on his retirement he married his nurse, Susan Bellanger, and they moved to the healthier climate of St Petersburg in Florida. No-one now seems quite sure whether Susan was his fourth or fifth wife. It was in St Petersburg that he died on 26 May, 1902, aged 62. He was buried beneath a tombstone which read simply, "Lieut. A.B. Strowger, Co. A, 8 N.Y. Cav.". Forty-seven years later, 110 years after his birth, representatives of the telephone industry placed a bronze plaque on his grave commemorating his pioneering work on automatic telephony.

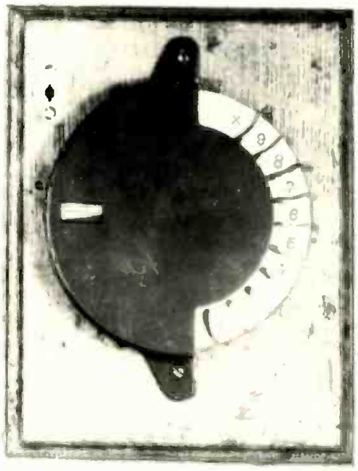
AN ARDENT BOOSTER

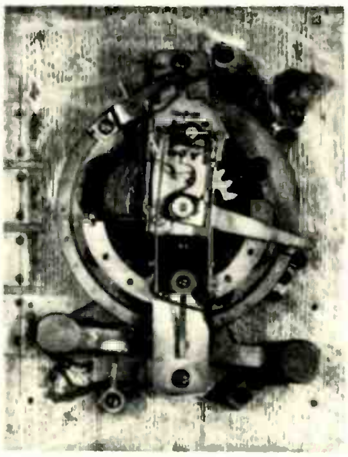
Life began for Almon Brown Strowger at Penfield in the state of New York on 19 October, 1839, one of a family of six boys and six girls. On his 22nd birthday, at the start of the American Civil War, he enlisted in the 8th New York Cavalry and served until the end of 1864. By this time he had been commissioned as second lieutenant. Shortly after, he married his first wife, who bore him two girls.

Strowger's career then moved from Civil War soldier to peace-time schoolteacher. He became principal of the school he had attended as a boy in Penfield. He taught in other schools in Illinois, Michigan and Kansas before buying an undertaking business in North Topeka, some 60 miles west of Kansas City. There he lived at the shop with his second wife, his first wife having died.



Original pre-patent Strowger switch, 1889 (GTE Communications Systems).





"No sex appeal": an early Strowger dial, 1898 (Museum of Independent Telephony).

"All things will adjust themselves to the new order": the first Strowger switch installation at LaPorte, Indiana, 1892 (GTE Communications Systems).

What happened to this second wife is not certain. Whether widowed again or divorced, Strowger married again in 1886, this time to Alice Marie Hill who bore him his only son. They moved house and business to Kansas City where he continued his work as an undertaker. It was there, as we have seen, that he conceived the idea of an automatic telephone. He was not actually the first to have the idea, but he *was* the first to produce a working model which could be developed into a commercially successful product.

Strowger's retirement in St Petersburg, Florida, lasted four or five years to his death.

When his eyesight failed, a young man, R.H. Sumner, used to read to him. Sumner, who died in 1949, left this description of A.B. Strowger in his final years. "He was an ardent booster for the city and was very active in all political and civic affairs. He was not known locally as an inventor but rather as an investor. He gave generously of his time and money when the occasion demanded. Strowger became a familiar figure on the unpaved streets of St Petersburg as he drove a pure white horse and a shiny black buggy, and he was always on the move. He was a very outspoken, high tempered man and if he didn't like the way the city was doing certain things he did not mind telling somebody how it should be done and oftimes would take over and do it himself".

The author would like to thank GTE Communication Systems Corporation of Northlake, Illinois, and the Museum of Independent Telephony, Abilene, Kansas, USA, for their help in the research for this article.

Next in this series of pioneers of electrical communication will be Michael Faraday, the 'patron saint' of electrical engineers.

Further reading

1. J. Harwell Jones, *Telephony*, 15 October 1949, p11-13,34.
2. "Big day in '92'" GTE Communications Systems Corporation, Northlake, Illinois.

FEEDBACK

Wagner, Debussy and electro-magnetism

As is well known, Richard Wagner conceived the notion that it would have a wonderful effect if, everytime a character in one of his operas appeared on stage, the orchestra played that character's signature tune. The resulting mélange led Claude Debussy to remark that Wagner's music reminded him of the inhabitant of a lunatic asylum who, every time he met you, presented you with his visiting card!

Much the same thing prevails in modern books on electromagnetism, which start by pointing out that the magnitude, F, of the force between two charges separated by a distance, r, is given by

$$F = k_e \frac{q_1 q_2}{r^2} \qquad (1)$$

where q_1 and q_2 are the moduli of the charges and k_e is a constant (which we may call the Coulomb constant). The practice in modern exposition is to remark, immediately after Coulomb's Law has been stated thus, that if another constant — the permittivity, ϵ_0, of free space is defined by the equation

$$\epsilon_0 = \frac{1}{4\pi k_e}$$

$$k_e = \frac{1}{4\pi\epsilon_0}$$

$$F = \frac{1}{4\pi\epsilon_0} \cdot \frac{q_1 q_2}{r^2} \qquad (2)$$

From then on all the analysis is done using $1/4\pi\epsilon_0$. Flick through most modern texts and you will see $1/4\pi\epsilon_0$ on practically every page – sometimes many times on a page. I have seen one page in which $1/4\pi\epsilon_0$ appears line by line thirteen times.

The situation immediately conjures up the vision of a madman wandering anachronistically through the Royal Society and engaging everyone he meets with the remark "Ah! Good morning. You know, that fellow may be the brother of The Earl of Cork but I — I am the Coulomb Constant and I am related to The Permittivity of Free Space.

The reasons given for the above procedure are varied and some bizarre in the extreme. I was told when I was a student that it made the expression of all spherically symmetrical problems less complex — with (1) and (2) staring one in the face! A well-known and very popular book actually says

"...it (k_e) is usually written in a more complex way as $1/4\pi\epsilon_0$"! Often it is stated that ϵ_0 is "a more important" or "a more fundamental constant" than k_e — whatever these expressions mean. Even if they had a meaning and they were true that would hardly justify the procedure adopted!

Many authors state the value for ϵ_0 — viz.

$$\epsilon_0 = 8.854\,\mathrm{p\,N^{-1}\,m^{-2}\,C^{2}}$$

and it is not clear to me whether some expect you painstakingly to go through the calculation $1 \div (4 \times \pi \times \epsilon_0)$ every time the Coulomb constant is required. But many people (not only students) actually do this! (I have seen a student's tutorial notebook on one page of which this calculation was carried out six different times in all its excruciating and agonizing detail, (who can blame her innocent little heart!). In recent years some authors have gone half-way by always using (2) but calculating $1 \div (4 \mathrm{x} \pi \mathrm{x} \epsilon_0)$ once for all at the beginning. Thus

$$\frac{1}{4\pi\epsilon_0} = 8.988\,\mathrm{G\,N\,m^2\,C^{-2}}.$$

In any calculation, this number is substituted immediately.

One would like to notice in passing that while (1) is superior to (2) for the mathematical reasons given above it has other glorious properties. Thus, putting numbers in (1)

$$F = \frac{\frac{q_1}{C}\,\frac{q_2}{C}}{\left[\frac{r}{m}\right]^2} \qquad (8.988\,\mathrm{G\,N})$$

Again the equation for the potential, V, at a point distant r from a charge, q, is given by

$$V = k_e \frac{q}{r}$$

$$= \frac{q/C}{r/m} \quad \left(8.988\,\mathrm{G}\,\frac{\mathrm{Nm}}{\mathrm{C}}\right)$$

$$= \frac{q/C}{r/m} \quad (8.988\,\mathrm{G\,V})$$

i.e. V = pure number x unit of potential.

Thomas R. Boag
Department of Physics
University of Strathclyde
Glasgow

Atomic fission

I was surprised to find from H. Aspden's April letter that he doesn't know what the Mössbauer effect is. It refers to the phenomenon that in solids resonant absorption of nuclear gamma radiation can sometimes occur with *no recoil shift*. In Mössbauer measurements Doppler tuning is used to compensate, not for energy losses arising from nuclear recoil, but for minute changes in the energy of the absorbed or emitted radiation arising from the different environments of the emitting and absorbing atoms within different solids. For this purpose the velocities available from loudspeaker cones are adequate. Resonance measurements have been made on a few of the many nuclei which show no significant Mössbauer effect, but these measurements required the use of a something like a high speed rotor to compensate for the nuclear recoil. Unfortunately the Doppler techniques are effective only over very small solid angles.

He goes on to talk about '... two species of *atom* that are driven into instability...'. However, nuclei which are isotopes of the same element are just as varied in their binding energies and energy level structures as are nuclei containing the same number of neutrons, but different numbers of protons, and have as strong a bias in favour of having even numbers of protons and neutrons. Anyone familiar with nuclear structure will be aware that the protons and neutrons in nuclei show independent shell closure phenomena analogous to those shown by atomic electrons. However the numbers associated with the closure of 'shells' of protons and of neutrons bear no simple relationship to those associated with the closure of the electron shells of atoms, and it is the latter numbers which determine the chemical classification of the elements embodied in Mendeléeff's Table. A particular nucleus is likely to be stable if it is more tightly bound than the nuclei which are related to it by exchanging one proton for a neutron, or the converse. Even if the ether oscillations Dr Aspden postulates were shown to exist, to affect proton binding, and not to affect neutron binding, I doubt whether they would produce more than a tiny fraction of the effects associated with the nuclear shell structure

C.F. Coleman
Grove
Oxfordshire.

Crossover

McKenny W. Egerton's three-way crossover network ("Circuit ideas" p.63 *EWW* January 1988) is very pleasing in its symmetry and simplicity. I believe however that the design's starting point as stated by Mr Egerton is flawed by a misconception, and yet despite this the filter design may be adequate even if not quite as perfectly adequate as Mr Egerton appears to believe.

The misconception is clearly stated by Mr Egerton in his preamble: "...this means that ignoring phase angle, the total power delivered to the two channels is constant". The job of a loudspeaker crossover network is not to keep the total power constant*, but to keep the vector sum of the drive voltages constant. (Just imagine, if two speakers both produce the same power, but with opposite polarity drive signals, the result will be quite different from that obtained if the drive signals have the same polarity). Hence the

FEEDBACK

phase angle cannot be ignored. It is voltages that add, in effect, rather than power levels, since the conversion from drive voltage to sound pressure is essentially a linear process.

Now what does all this mean as regards Mr Egerton's design? He begins by showing, correctly, that the combined effect of maximally flat ($Q=1/\sqrt{2}$) two-pole, low-pass and high-pass filters with the same resonant frequences is to keep the sum of the power levels in the two channels constant. Looking instead at the vector sum of the two channel voltages, the result is quite different, giving $1-\omega^2/(1-\omega^2+j\sqrt{2}\,\omega)$ (ω here is normalized radian frequency). Notice there is a complete loss of output, or notch, at $\omega=1$ (the crossover frequency). This can be rectified by reversing the polarity of one channel, to give a vector sum of $1+\omega^2/(1-\omega^2+j\sqrt{2}\,\omega)$. This response has a 3dB 'hump' at the cross-over frequency, and a rapid 180 degree phase shift around the crossover frequency. Between the Scylla and Charybdis of the notch, and the rapid phase shift-plus-hump, there is no respite, and both have been demonstrated to be audible.[1]

From this unpromising beginning, Mr Egerton's three-way circuit actually leads to a considerable improvement. Taking first his wrong approach of summing the squares of the amplitudes (proportional to power) of each channel, the result from his circuit for A=4 is

$$\left[1-\frac{\omega^2(1+\omega^4)}{2.5(1+16\omega^4+\omega^8)}\right]^{-1}$$

which is +0.1±0.1dB for all frequencies.

Taking the correct vectorial addition of voltages gives

$$\frac{4-\left(\omega^2+\frac{1}{\omega^2}\right)}{6.24-\left(\omega^2+\frac{1}{\omega^2}\right)-j3.52\left(\omega-\frac{1}{\omega}\right)}$$

if the mid-range speakers connections are reversed with respect to the other two channels. This response shows a rise of 2.5dB at 0.1 and 10 times the centre frequency, reaching 3dB beyond that range.** For a centre frequency of 1kHz, for instance, this results in a relative boost of 2.9dB at 50Hz and 20kHz. This will be audible with comparative listening tests, though for at least some listeners, this may be felt to be an improvement rather than otherwise. The phase shift is quite modest (30 degrees maximum) and is unlikely to be audible.

In summary, an elegant circuit, though requiring further minor corrections for best results, and potentially causing a 50% drop in speaker efficiency (second footnote); a fair result considering the error on which the design appears to be based!

Brian J Pollard
Watford
Herts.

Reference

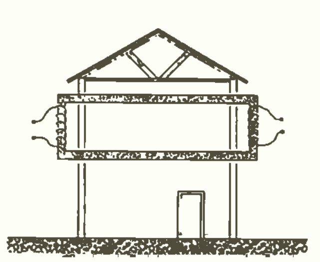
1. Ashley J.R. "On the transient response of ideal crossover networks" *J.A.E.S.* vol.10 No 3 July, 1962 pp.241-244.

* Constant power is usually a crossover and network design objective, but in this context it means making the power levels of each speaker the same at their crossover frequency.

** More accurately, this response results in partial cancellation at mid-band frequencies, causing a 3dB reduction in net output at mid-band. These effects will in practice be mitigated by any tendency of the low and high frequency speaker responses to naturally fall off at mid-band frequencies.

Poynting the way

Joules Watt's article "Poynting the way" in the February issue of *EWW* reminds me of an exercise given to us (students at the Technological University of Delft) in 1963 by Professor J.L. Bordewijk. He told us that this exercise appeared in the technical literature in his student days which must have been during or shortly after World War II.



Imagine a laboratory as shown in the diagram. By means of two iron bars, part of a huge transformer, energy is transported through the walls of this laboratory. The laboratory assistant is asked to find out whether the iron bars convey energy and if so in what direction. He has available all the equipment that can be found in a well-equipped physics laboratory.

This puzzle is easy to solve after reading Joules Watt's article.

Ir. P. van der Wurf
Geldorp
The Netherlands

Relativity

Thank you for the provocative article published in the February edition of your publication where the author expressed his rejection of Relativity. As students of physics we were smugly expecting a flood of angry reply letters but instead, March's contributions to 'Feedback' contained two replies: a letter assuring everyone that the formulae actually work, and a gentleman who constructed a new paradox in argument against Special Relativity.

First a response to February's article. We do not wish to involve ourselves with the so-called logical contradictions mentioned, as we believe it is subject to personal opinions and would distract from the main point we wish to make: this being that many physicists are confident that Relativity — especially Special Relativity — is a fact of life and, contrary to the concluding remark, is *not* a theory with no genuine experimental support. Some of the phenomena in support of Special Relativity which were not mentioned in the article include muon decay lifetimes, the concept of particles 'spin'[1], mass-energy equivalence, Boucherer's 1909 experiment on the velocity dependency of an electron's mass[3], and transverse Doppler broadening. More can be mentioned, but anybody can find an undergraduate text on Special Relativity for more examples. Of the above list, stress must be placed on Dirac's work on the idea of 'spin' which came from a Special Relativistic treatment of Quantum mechanics, and is a central feature of Quantum theory. What all this means of course, is that any attempt to disprove Special Relativity will require an awful lot of work and thought in order to explain numerous experimental results. Compared to these, rewriting Quantum mechanics should be relatively straightforward.

We have deliberately separated Special Relativity from General Relativity because the latter requires a mathematical grasp beyond our present competence. However, evidence does exist in support of this theory apart from the famous Eddington experiment quoted. For example, a variation on the Eddington experiment has been done using radio waves from quasars to an accuracy sufficient to discount rival theories. Then there is the well known confirmation of Mercury's orbital precession (a procedure later repeated for Venus) which is considered a major triumph of General Relativity, yet to our surprise it is absent from Dr Essen's article.

As an aside we wish to comment on Dr Essen's interesting choice of Bertrand Russell to support his assertions, since he is acknowledged to have written an excellent introductory book on Special Relativity. Furthermore, it is puzzling what Dr Essen reads into the remark that Special Relativity is all contained in the Lorentz-Fitzgerald transformations. In reply, so what? One may equally say that Newtonian mechanics is all contained in F=ma, and yet this is hardly the height of contention.

In reply to Mr Hobden's paradox presented in March's Feedback, we shall use R.P. Feynman's definition of a paradox as a scenario where the physics has been misunderstood. If the inquisitor is sufficiently clever, then the misunderstanding may lie in the theory and awards and adulation will shortly follow. Unfortunately for Mr Hobden, the

FEEDBACK

misunderstanding lay elsewhere.

Before we substantiate that last statement, we wish to emphasise one of the two postulates of Special Relativity: physics is the same in all inertial frames and no one is preferred (the other being that the speed of light in vacuo is the same for all observers), where an inertial frame is a non-accelerating frame of reference — i.e. either stationary or moving at a constant velocity. The consequence of this is that the Galilean idea of an absolute reference frame simply does not exist.

With this in mind, one can see that the horologist "having read 'Relativity'", is entirely incorrect in thinking that as "the clock velocity increases, the mass increases". Relative to a stationary observer, the mass of the clock (and the entire vehicle) has increased; but in the inertial frame of the clock itself the mass of the clock is invariant. Hence the ingenious horologist's clock will have no increase in mass to compensate, for if there is a measurable increase then it would be possible to tell if one is stationary or merely moving at a constant velocity — i.e. the existence of an absolute reference frame and violating the above postulate. We are sorry to say this, but the physics in this compensating clock 'paradox' is badly flawed and appears to display a gross misunderstanding of reference frames — a construct that is prevalent in Special Relativity.

As for the accusation levelled against Einstein that he had little understanding of clocks (!), it ought to be realised that Special Relativity does not specify what the clocks are. They could be based on atomic transitions, spring-wound torsional oscillations, piezo-electric effect, or even massless photons. It does not matter! They are all physical systems and as such are covered by physical theories such as General and Special Relativity, Quantum mechanics, thermodynamics, and (sometimes) Newtonian mechanics. Compared to these deliberately general theories, worries about the non-linearities in the workings of clocks are extremely misleading and irrelevant.

To conclude, it is conceivable that Dr Essen and Mr Hobden might have made errors in their understanding of Special Relativity, but Dr Essen's dogmatic statement of Relativity as a theory that ". . . lacks any genuine experimental support" is frankly incredible. Such a sweeping statement requires careful consideration and cannot be concluded after quoting three experiments and the cavalier dismissal of two of them. A man of Dr Essen's credentials will be aware of the everyday use of Special Relativity by a lot of physicists and yet he has neither quoted an experiment which blatantly violates Special Relativity, or given a reason why everybody else is wrong. Anybody can stand on a soapbox and declare the world is flat, but a scientist is expected to prove this, and show why all the prior data in fact supports one's claim in preference to the previous model. This is part of the scientific method, and is still the best objective way to describe the universe Mankind has discovered.

M.R. Taylor and H.W. Yau
Whalley Range
Manchester
Lancashire

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1. P.A.M. Dirac, 'The Principles of Quantum Mechanics' Fourth edition, 1958. Oxford University Press.
2. D. Kleppner and P.J. Kolenkow. 'An Introduction to Mechanics' 1978. McGraw-Hill, inc.

L. Essen is quite mistaken when he writes that Bertrand Russell criticized the theory of relativity (*EWW* February 1988). Russell only pointed out the rather obvious fact that it was all contained in the Lorentz transformations to which I also draw attention following one of the articles of Essen on relativity (*WW* October 1978 and April 1979).

Russell wrote an excellent book on relativity (The ABC of relativity) which I can strongly recommend to anyone who wants to understand the theory.

Also it seems that Essen is oblivous to the decision in 1983, of the General Committee on Weights and Measures, that the speed of light is now one of the basic constants to define the unit of length, the metre, and to blame Einstein for having "anticipated" this development (Feedback December 1986).

J J Bleeker
Geneva
Switzerland

Sinusoidal oscillators

Reading Damljanović's article (February 1988) describing a sinewave oscillator, reminds me of a plausible alternative if stability is the criterion at sub-audio frequencies: a time-sampled model of a dual integrator.

Visualize a 16 bit microchip driving a digital-to-analogue converter with the following program (rendered in generic Basic for simplicity):-

```
1 : REM INITIAL CONDITIONS
2 : ESIN = 0 : ECOS = 1 : AMP
    = 2 ↑ 15 – 1
3 : DELTATIME = 0.001
4 : REM SIN/COS INTEGRATOR
    LOOP
5 : ESIN = ESIN + ECOS *
    DELTATIME
6 : ECOS = ECOS – ESIN *
    DELTATIME
7 : OUTPUT = INT (ESIN*AMP
    + 0.5)
8 : GOTO 5
```

The loop consisting of statements 5 to 8 computes sin and cos to better than 1 part in a million, and estimates the value of pi through the following relation:-

Number of loop iterations per sinewave = 2π/DELTATIME. The 'OUTPUT' value is scaled to a 16 bit integer whose most significant bit is the sign.

The output waveform error is ± 1 l.s.b. of the converter. Its stability is controlled by the crystal clock driving the micro.

Accuracy falls with increasing DELTATIME; at the limit of DELTATIME = 1, a cycle of 6 iterations produces a trapezoidal wave. For an execution speed of 1 Basic statement per millisecond, the upper frequency limit is then 167Hz.

Very low frequencies are produced, preferably, by inserting a delay loop within the main loop, rather than arbitrarily reducing the value of DELTATIME. (This would be defeated by the finite resolution of the arithmetic in lines 5 and 6) In a 'bare' micro using assembler, no sine table is needed, but double precision arithmetic is desirable.

Brian Whatcott
Garland
Texas
USA

Seven per cent

Ivor Catt, in his article on data compression (*EWW* April), is right; there is no need to use a 15 bit code for compressing English text. Neither is there any need to invent a new 7 bit code, since a 6 bit code with extensive text compression is already in wide use.

It is called Braille.

J.R. Heath
Southampton
Hampshire

A rule which leads to Ivor Catt's "7% Rule" (March) was put forward by Zipf[1] in the late forties, and its application to information storage and retrieval was considered by Schuegraf and Heaps[2] some fifteen years ago. The rule states that if the words in a particular text are arranged in order of decreasing frequency then the relative frequency of the Nth word is approximately Z/N, where Z is a constant near 0.07. For dictionaries of more than a couple of thousand words it would give a total relative word frequency of more than one, so that it must break down somewhere. Ten years ago I had occasion to determine the word frequencies in five legal texts, the largest of about half a million words, and found by plotting the Zipf 'constant' for the Nth word against logN that in each case it began to decrease almost linearly with logN when N exceeded about a thousand. As a result the Catt 7% rule fails when the fraction of all text words accounted for rises above 90%.

For the rule to be useful in

FEEDBACK

data compression it must be shown that for different documents the same words tend to occur in about the same places in the frequency ordered lists. Analyses such as those applied to the documents contained in the Lob Corpus show that this is broadly true for the everyday words. However the frequency sequence Ivor Catt quotes *(the, of, and, to, a, in, that, is)* already departs from the corresponding sequence in the legal texts *(the, of, to, or, in, and, be, any)*. The ratios of the maximum and minimum frequencies of the forty commonest everyday words over the five legal documents ranged from 1.1 for *not* to values near three for *any, such, made, so,* and *all*.

Certainly the frequencies of the nouns and other words which convey the particular meaning of a text vary wildly from one passage to another, and I have seen a specific word come as high as fourth in the frequency ordering. For one of the legal texts I estimated that there was better than a fifty percent chance that any passage of three hundred words would contain one word whose relative frequency taken over the whole document was less than one in ten thousand! Words which occurred once only accounted for between 20 and 30% of the dictionary entries for the legal documents (the Zipf rule gives 50%).

Ivor Catt's information on dictionary costs is already outdated. The word processing package PROTEXT marketed for the Amstrad PCW computers includes spell-checking facilities and a dictionary of 30,000 words and costs £60. The designers recognise that many users will want either to extend their dictionaries, or to generate supplementary dictionaries, and provide facilities for doing so.

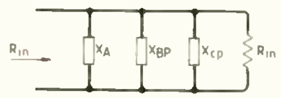
1. G.K. Zipf, 'Human Behaviour and the Principle of Least Effort', Addison Wesley, Cambridge, Massachusetts (1949)
2. E.J. Schuegraf and H.S. Heaps, Information Storage and Retrieval *9*(1973)697

C.F. Coleman
Grove
Oxfordshire

"Lossy Ells for Pie Tea"

The loaded Q of a low-pass pi matching network is not given by R_{in}/X_A (Joules Watt, January 1988 issue), but by $(R_{in}/X_A + R_L/X_C)$ (refs. 1 and 2). This may be derived easily using the equivalent shunt circuit shown in Fig.1.

The equations for X_A and X_C in Fig.1, are from Everitt (ref.3) and show that for the radical $\sqrt{R_{in}R_L - X_B^2}$ to be real, X_B^2 should not exceed the product $R_{in}R_L$. The maximum value of loaded Q (as in any tuned circuit) is limited in practice, by its effect on the efficiency of the network which is equal to $P_o/P_{in} = (1-(Q_{LOADED}/Q_o))$ where Q_0 is the unloaded Q of the network with R_L and R_g disconnected. The higher the loaded Q, the greater will be the power loss through the network.

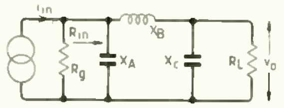


$$X_{BP} = \frac{R_{in}}{X_B}\left(\frac{X_C^2 R_L}{X_C^2 + R_L^2}\right)$$

$$X_{CP} = X_C \frac{R_{in}}{R_L}$$

$$R_{in} = \frac{X_B^2}{R_L} + R_L\left(\frac{X_B}{X_C} + 1\right)^2$$

$$\frac{1}{X_A} + \frac{1}{X_{BP}} + \frac{1}{X_{CP}} = 0$$



$$X_B = \frac{R_{in} + R_L + 2\sqrt{R_{in}R_L - [1/Q(R_{in} - R_L)]^2}}{Q + 4/Q}$$

$$X_A = \frac{-R_{in}X_B}{R_{in} + \sqrt{R_{in}R_L - X_B^2}}$$

$$X_C = \frac{-R_L X_B}{R_L + \sqrt{R_{in}R_L - X_B^2}}$$

If the generator feeding the network has internal resistance R_g (which may or may not equal R_{in}), then the presence of this resistance will reduce the loaded Q of the network.

The frequency response of the networks is a function of the ratios of R_{in}/R_L, R_g/R_{in} and loaded Q; the variation of Z_{in} with frequency also is a function of the same parameters.

Finally, back to Joules Watt's pi network: it works perfectly in transforming the load impedance to 75 ohms, but, and assuming that the generator resistance greatly exceeds 75 ohms, the loaded Q is 17.3 and not 15.

J.E. Diggins
South Ascot
Berkshire

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1. J.E. Diggins, "Tuned Radio Frequency Impedance Transformers", Racal Electronics Plc, International Paper, June 1982.
2. Elmer A. Wingfield, "New and Improved Formulas for the Design of Pi and Pi-L Networks", *QST*, August 1983 p23.
3. L. Everitt, "Communication Engineering", 2nd edition (McGraw-Hill Book Company Inc., 1937).

Invention

Heinz Lipschutz (*EWW* March, 1988) raises a number of interesting questions in his discussions of inventors as an endangered species. He is correct in his belief that neither governments nor industry appreciates the need to succour inventive genius wherever it may be found. He does, however, confuse invention and development. He implies, for instance, that his 1939 system for inertial navigation was similar to that used today. The possibility of an i.n.s. was considered well before 1939 and for long-range use was critically dependent on the work of Schuler who, in 1923, described a gyroscopic method of establishing a true vertical. But to produce gyroscopes and computers with anything like the precision needed to navigate an aircraft to a town in Germany required the development of manufacturing and data handling techniques simply not available in the 1940s. Most engineers can look back over their careers and see a collection of inventions that might have been. For example, about 30 years ago I developed a very clever device that responded to the proximity of either ferrous or non-ferrous metals. It was cheap, reliable and required no adjustments. When it had served its purpose — in a biomedical application — it was junked. I did not publish, protect or market it as I did not foresee a world where every traveller must be regarded as a possible terrorist.

In electronic and computer engineering the first need is for a rational patent system. At present the paper work is burdensome, much time is consumed and in the last analysis your patent grants you only the right to defend the invention in the courts. For a true invention (safety pins, zippers and 'cat's eyes' are examples from the recent past) the present tests of patentability are probably adequate but when vast numbers of concepts 'known to practitioners of the art' are combined (e.g. colour tv, digital recorders, inertial navigation systems) to make a 'new' product the process is one of development, not invention, to which patent law is scarcely relevant.

As a final thought, a visitor from outer space looking over the entire spectrum of human activities might well find "bashing a ball into a hole" one of mankind's least harmful competitive endeavours.

Harold W. Shipton
St Louis
Montana
USA

A question of waves

It is known that longitudinal waves all obey the same rules in an elastic medium. Might I ask the readers of *EWW* whether transverse waves in an elastic medium also all obey the same rules?

Alex Jones
Swanage
Dorset

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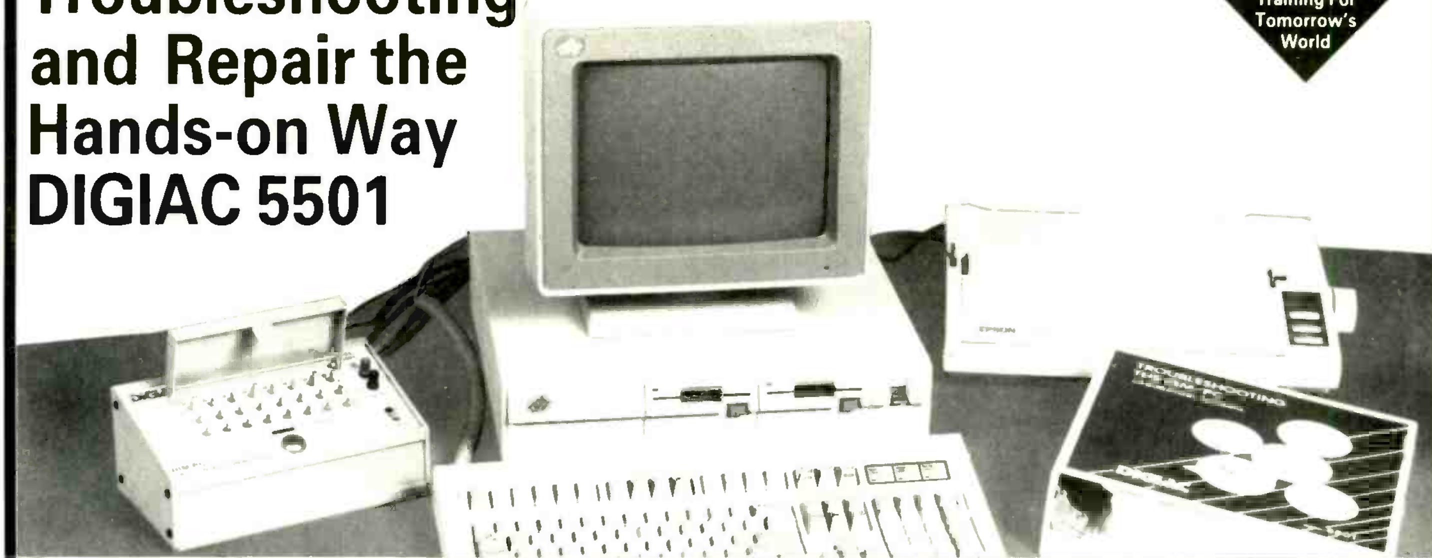

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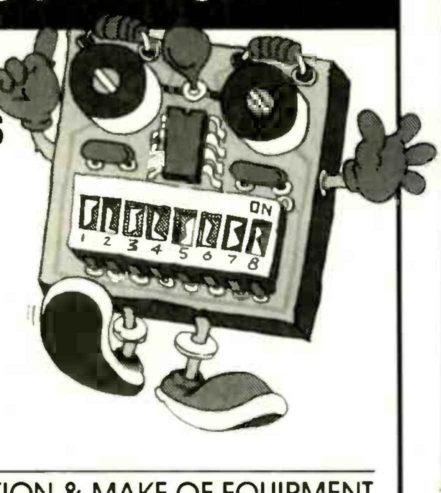

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Marconi TF144/H 10Hz 72MHz	£75
Marconi TF1064B/5 68-470MHz	£100
Philips PM5234 100KHz-110MHz	£375
Marconi TF2333 Transmission Test Set	£350
Feedback FG600 Function Generator	£80
Lyons PG73N Bi Polar Generator	£85
HP202C LF Oscillator	£95
HP 8620A Sweep Generator 1.2 to 12GHz	£3,000
HP3311A Function Generator	£175
HP333A Distortion Analyser	£350
Marconi TF2330A Wave Analyser	£300
Levell TG200B RC Oscillator	£40
HP654A Test Oscillator	£495
Fluke 6160A Freq Synthesizer	£275
Texscan WB11 500MHz Sweeper	£475
Radford LD02 DMS2 The Pair	£235
Marconi TF2012 400 to 520MHz	£175
Tektronix 184 Time Mark Generator	£175

GENERAL TEST EQUIPMENT

HP 1600A 1607A Logic Analyser	£550
HP 1600A Logic Analyser	£350
Gould LA500 50MHz Logic Analyser	£750
Gould G200 Biomation Logic Analyser	£200
HP 1601A Logic Analyser Plug in	£375
HP 8170A Logic Pattern Generator	£1,250
Data Lab DLGOS Transient Recorder	£150
Marconi TF2300 Mod Meter	£200
Marconi TF2300B Mod Meter	£400
Marconi TF2700 LCR Bridge	£150
Radiometer MM2 LCR Bridge	£175
Marconi TF1313A ·1% LCR Bridge	£200
Marconi TF1245/1246 Q Meter	£195
Bird Termaline 6254 100MW 30/500MHz	£75
Marconi TF2501 RF Power Meter	£135
Marconi TF6460-6420 Head	£450
HP141A Storage Main Frames Good Tubes	£120
Ferrograph RTS2 Audio Test Set	£250
Kerco Bipolar PSU −20V-220V+ 20A-20+	£200
HP3400A RMS Voltmeter	£325
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Racal Dana 9915M	£200
Racal Dana 9917A	£235
Racal Dana 9916	£225
HP 5381A 80MHz Counter	£175
HP 5300A 5304 Counter	£195
Schlumberger 7201 Locators + 2 Pods	New
Schlumberger 4980 ATT	£250
Brandenburg Alpha 30Kv PSU	£150
HP8052 Sound Level Meter	£100
Marconi TF1026 Freq Meter 500/1000MHz	£15
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Thermo-electric temperature controller

Using a Peltier module to cool or heat components or laboratory specimens

PETER TURNER

The range of applications for this thermo-electric module controller is limited only by the imagination. Uses which I have come across are as diverse as cooling a charge-coupled device (c.c.d.) in vacuum to maintain constant operating conditions while the c.c.d. is under test to cooling a milk sample in a test laboratory while quality control tests are carried out. With the recent advances in superconductor technology bringing operating temperatures ever closer to ambient, researchers could soon be using thermo-electric modules to cool their samples rather than liquid helium or nitrogen.

PELTIER EFFECT

The Peltier effect is the phenomenon discovered by Jean Peltier in 1834 whereby the passage of an electric current through two conductors of dissimilar material causes heat to flow from one side of the junction to the other, depending on the direction of current flow. The rate of heat flow is proportional to the amount of current flowing. Of course the Peltier effect will also work in reverse, i.e. if a temperature difference is maintained across the junction a potential difference will be created proportional to the temperature difference (the thermocouple effect). The Peltier effect can be used to heat or cool an object depending on the direction of current flow.

THERMO-ELECTRIC MODULES

Peltier heat pumps are made up of several p and n-type semiconductor couples connected in series electrically and in parallel thermally.

Three sizes of heat pump are obtainable from RS Components, having 4, 31 and 127 couples with a heat pumping capacity of 0.29W, 19W and 75W respectively. The controller was designed to operate with the

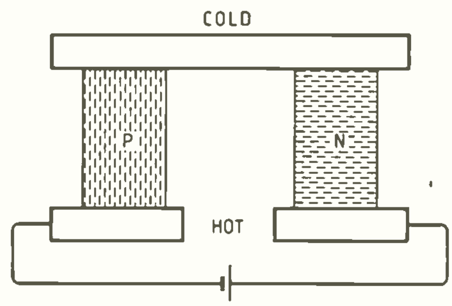
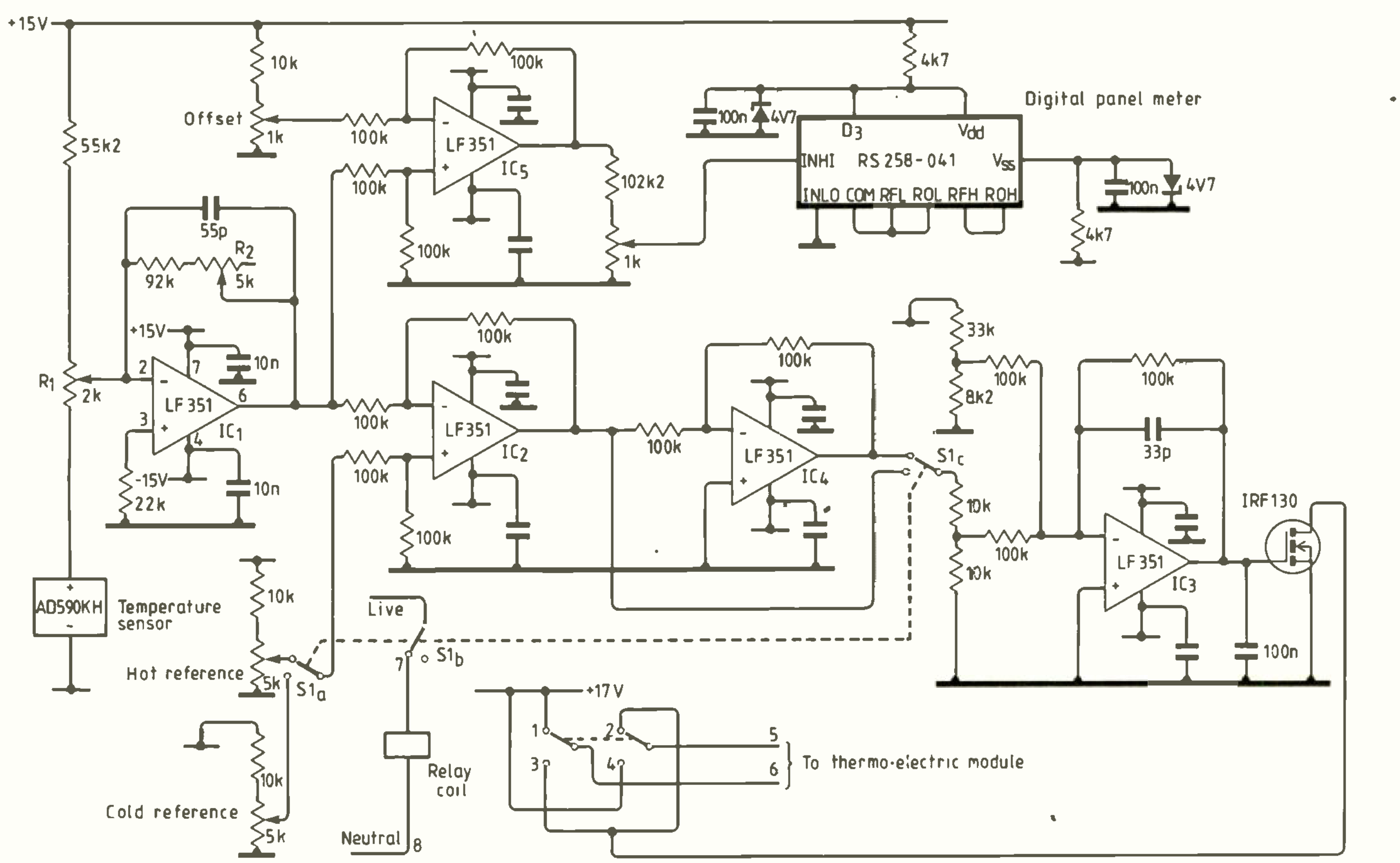


Fig.1. A single semiconductor couple connected as a heat pump.

Fig.2. Controller circuit diagram. The instrument was constructed on pin-board.



75W module although it could easily be modified for use with either of the other two.

CONTROLLER

The controller provides a current proportional to the difference between the set point and the actual temperature of the item being cooled up to a maximum of 9A. The temperature of the item is sensed by an Analogue Devices AD590KH temperature-sensor i.c. and the current regulated by an IRF130 power mosfet. An l.c.d. panel meter module provides a temperature indication. The control circuitry is shown in Fig.2. The AD590KH temperature sensor, acts as a high-impedance current source passing 1μA per degree Kelvin. The op-amp IC_1 effectively converts this current into a voltage and provides a two-point temperature trim, R_1 setting the offset and R_2 the gain. Thus, at the output of IC_1 we have a voltage which varies with temperature at the rate of 100mV per °C. The capacitor between pins 2 and 6 on IC_1 prevent the circuit bursting into oscillation, as it was initially prone to do!

The op-amp IC_2 is configured as a differential amplifier and compares the reference voltage with the voltage from IC_1, which is proportional to the actual temperature, and produces an error voltage. When cooling, this error voltage is taken directly to IC_3, which is a summing amplifier. It adds an offset voltage to the error voltage and produces a voltage at the gate of the power mosfet somewhere in the conduction range of the transistor (see Fig.3).

As an example, consider when the actual temperature is 30°C and the cold set points is −40°C. There will be a voltage of +3V at the output of IC_1 and −4V at pin 3 of IC_2, producing an error voltage of −7V at the output of IC_2. One input to the summing amplifier is, therefore, 3.5V, the other being the output of another potential divider at a voltage of approximately −3V. The mosfet gate voltage is about 6.5V, giving a drain current of 9A. As the temperature falls the error voltage reduces and the voltage at the gate of the mosfet is +3V, which is the cut-off voltage of the device.

Again, the 33pF capacitor is to prevent oscillations, as is the 100nF capacitor to ground at the output of IC_3, which is particularly necessary to prevent oscillation when switching from cool to heat.

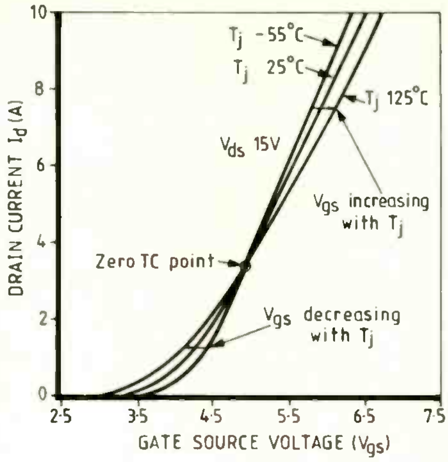


Fig.3. Gate voltage-drain current curves for the IFR130 power mosfet, showing cut-off at around 3V.

To heat the sample, a 3-pole switch switches from cold reference to the hot reference; energizes a relay, which reverses the current flow through the thermoelectric module; and switches in a unity-gain inverting amplifier IC_4, since the output from IC_2 is now in the reverse sense.

The remaining circuitry, IC_5 and the digital panel meter provide a constant readout of the temperature of the sensor. IC_5 is arranged again as a differential amplifier to provide some offset adjustment.

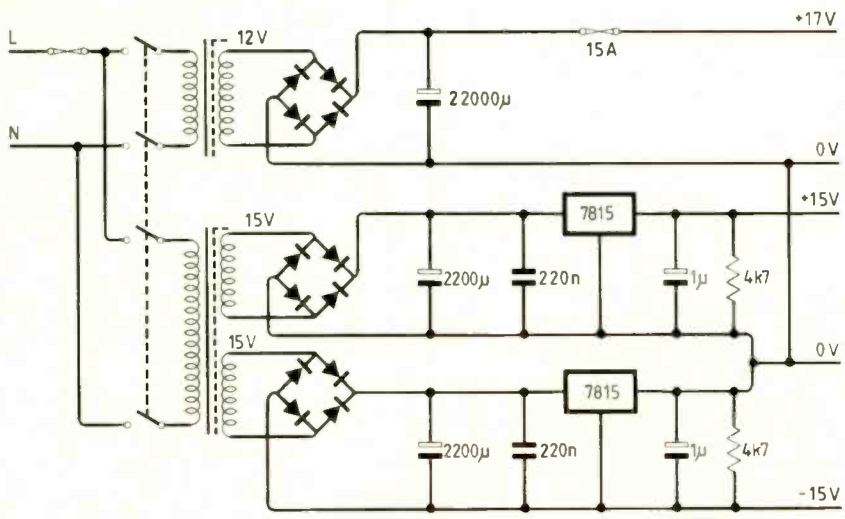
The power mosfet was mounted on a 1.1°C/W heatsink along with the diode bridge for the high current supply.

To achieve temperatures mentioned in the text it would be necessary to enclose the module and item to be cooled in a vacuum. This cryostat enclosure could be maintained at, say, 0.5 of atmospheric pressure using a vacuum pump to prevent condensation of water vapour and heat gain by conduction and convection by air.

Of course, such an elaborate set up would not be necessary if using the mini-module to cool an i.c. or the larger ones for a lower temperature reduction.

Mr Turner is an electronics research technician at the University of York.

Fig.4. Simple suggested power supply for the controller.



Music database on CD-rom

The world's first integrated audio database on CD-rom, carrying both recorded music and computer data, has been launched by the British company Nimbus Records. CD-roms can store some 600M-byte of data, though in this case Nimbus has used only the first minute or two of the disc for data: the remainder consists of dozens of excerpts from the company's music recordings.

This electronic catalogue can be read on any personal computer with a CD-rom drive plus the appropriate operating system extensions. With the help of its user-friendly retrieval software, record buyers can search the database for the music of their choice; and, having come across an interesting-looking item, they can listen to a sample of it in full CD quality. Besides listing the contents of each record, the database carries supplementary information, such as reviews by music critics, and a digitized image of the cover design.

Besides producing its own-label recordings, Nimbus manufactures discs in both large and small numbers for outside customers. The CD mastering equipment, designed by the company's own staff, last year won it a Queen's Award for Technological Achievement. Nimbus is now the UK's largest manufacturer of Compact Discs, with its factories at Wyastone Leys, Monmouth, and Cwmbran producing some 20 million discs per year. A further plant, at Charlottesville, Virginia, in the USA, was completed last September. A newly-established division now specializes in the exploitation of CD-rom, from data compilation to the production of finished discs.

Further information from Nimbus Records on 0600-890682.

Second-generation risc processor

A new design from AMD overcomes some of the throughput limitations of earlier risc devices.

M. A. LEHRER

With current semiconductor technology, it is possible to make chips which can process at incredible speeds. Unfortunately, much of this potential is never realised in a practical system because most of c.p.u's latent power is dissipated in decoding instructions, in communicating with the outside world via too few pins, or in housekeeping operations.

This article describes one approach to solving this problem. The Am29000 is the first microprocessor AMD has designed; and an overriding objective during the project was to ensure that as much as possible of the potential speed of the chip should be available to the system in which it is installed.

First of all, a decision was made to employ the reduced instruction set concept (risc) which enables instructions to be decoded and executed in the minimum possible time. By employing only the most commonly-used instructions, risc removes the need for the microcoded logic required for the large instruction sets of conventional microprocessors and allows most instructions to be executed in a single cycle.

The first generation of risc processors failed to exploit the technique fully because a high instruction-execution rate cannot be sustained without fast access to program data, and proper pipeline design. Lacking these, a processor can achieve only a fraction of its potential speed, because it has to wait for external data accesses or to refill the pipeline after a branch.

In addition, the external interface must be designed to achieve the shortest possible cycle time. To achieve high system throughput, the Am29000 has been designed for maximum performance by optimizing the product of instructions per task, cycles per instruction and time per cycle – not by minimizing one factor at the expense of others.

Two approaches to risc

But a microprocessor cannot achieve this alone: compiler, operating system, application environment, arithmetic accelerator, system development and debug methodology all affect the speed of the device. Resources of the Am29000 include a 192-word register file to eliminate data access delays, on-chip branch target cache, 40ns cycle time, on-chip memory management unit and a high-performance channel interface.

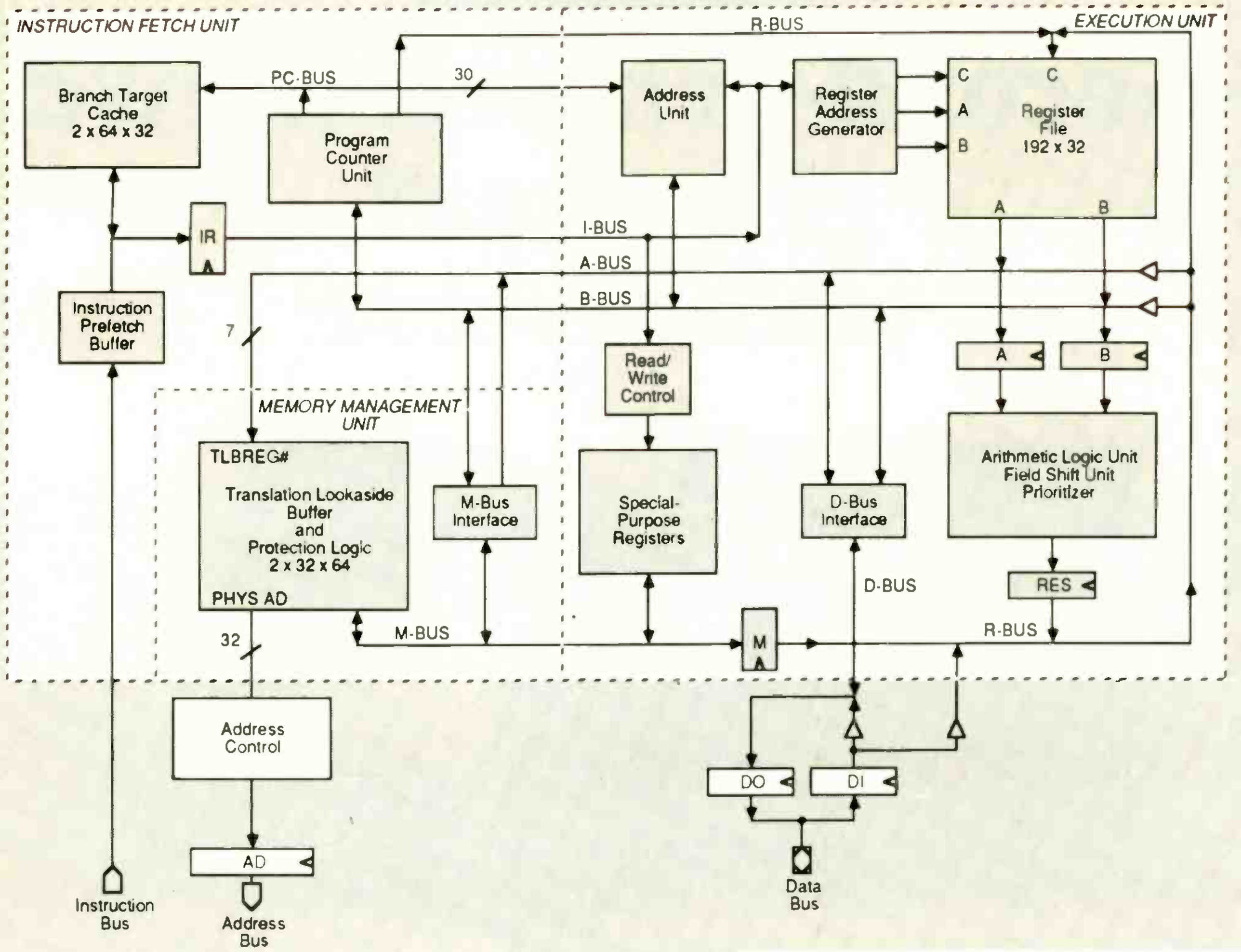
Because efficient software is necessary if a system is to give its best, support is included for procedural languages, real-time operating systems, optimizing compilers, multitasking operating systems and demand-paged virtual memory.

MAXIMIZING THROUGHPUT

Within the microprocessor, which operates at 25MHz, is a four-stage pipeline which can execute an instruction on every 40ns cycle and can sustain a speed of 17Mips continuously. To realise this throughput, a new instruction must be delivered to the pipeline on each cycle.

When on-chip resources cannot cope with simultaneous instruction and data requests, the bus must have a 200Mbyte/s capacity. This is provided by the chip's external interface, which consists of three 32-bit buses and associated control signals: input-only instruction bus, bidirectional data bus and output-only address bus. This permits simultaneous instruction and data access even though the address bus is shared, because the address is pipelined. This arrangement offers the benefits of a full Harvard architecture (which is familiar to bit-slice designers), at a greatly reduced pin count and without the cost of four full 32-bit buses.

All 192 32-bit general purpose registers in the Am29000 are capable of storing variables, addresses and operating system values. In multi-tasking applications they can be used to hold processor status and variables for as many as eight different tasks. As a result, most instructions can be fetched without the delay of an external access. Compare this with microprocessors which



Data flow within the Am29000 risc microprocessor.

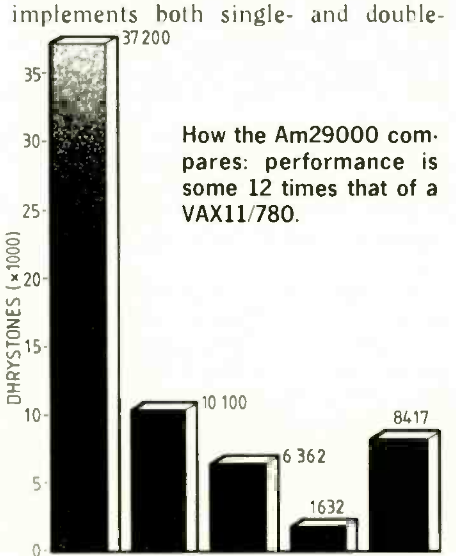
have only 16 general-purpose registers: these necessitate more frequent external memory accesses, making the instruction cycle time longer.

In multi-tasking applications when real-time response is most important, the register file can be banked into 12 blocks of 16 registers, each block dedicated to a separate process. A task switch can then occur in as few as 17 cycles, or 680ns.

Fastest transfer of instruction and data to sequential addresses is achieved through 'burst mode' accesses. Only the first address is sent: subsequent requests for information can occur at the rate of one access per cycle, without requiring additional address transfers. Using burst mode accesses, the Am29000 can load or store data at 100Mbyte/s, which is considerably faster than both FDDI and Ethernet. This ability makes it an ideal processor for high performance communications and local area networking.

Key features of the integral memory management unit (m.m.u.) include a 64-entry, two-way set associative translation lookaside buffer (t.l.b.) which performs the virtual-to-physical address translation. To make this more efficient, operation of the t.l.b. is pipelined to run parallel with other processor operations. Other features of the m.m.u. include 4Gbyte of virtual memory per process for up to 256 processes; pipelined address translation to reduce latency; software t.l.b. reload allowing user-defined memory-management architecture for maximum flexibility; least-recently-used hardware to assist reload and memory protection. Software t.l.b. reload allows system designers to choose a memory management scheme best matched to the particular environment: for example, page selection of 1K, 2K, 4K or 8Kbytes. T.l.b. reload performance is enhanced by least-recently-used hardware as well as a low trap overhead.

A floating-point coprocessor, Am29027, interfaces directly to the microprocessor. It implements both single- and double-precision floating point operations, offering also a complete range of integer and conversion operations. When one operand comes from the register file and the other from the input register, the Am29027 reduces operations by performing computations on one single-precision and one double-precision operand.



How the Am29000 compares: performance is some 12 times that of a VAX11/780.

The accelerator can be used in pipelined or non-pipelined (flowthrough) mode, depending on system requirements. Pipelined mode maximizes the rate at which the accelerator produces new results; flowthrough mode minimizes the overall executive time for scalar operations.

Benchmarks can be run with various memory configurations using the Am29000 architectural simulator, to allow comparisons to be made with other processors. The diagram (left) shows that the Am29000 offers approximately 12 times the performance of the VAX11/780, and three to five times that of the 68020.

APPLICATIONS

With its high performance, the Am29000 is suitable for many applications where real-time or super fast data processing is required – such as in embedded controllers, engineering and scientific workstations, desktop publishing, super minis, industrial robots, artificial intelligence and speech recognition. In ISDN networks, the Am29000 provides the high-speed switching control node.

Controllers based on this will give workstation networks access to large databases ten times faster than that of existing Ethernets. A multi-font laser printer built around an Am29000 is four times faster than printers based on a 68030.

Applications likely to benefit especially from the accelerator include graphics, robotics, simulation and array processing.

UNIX MACHINES

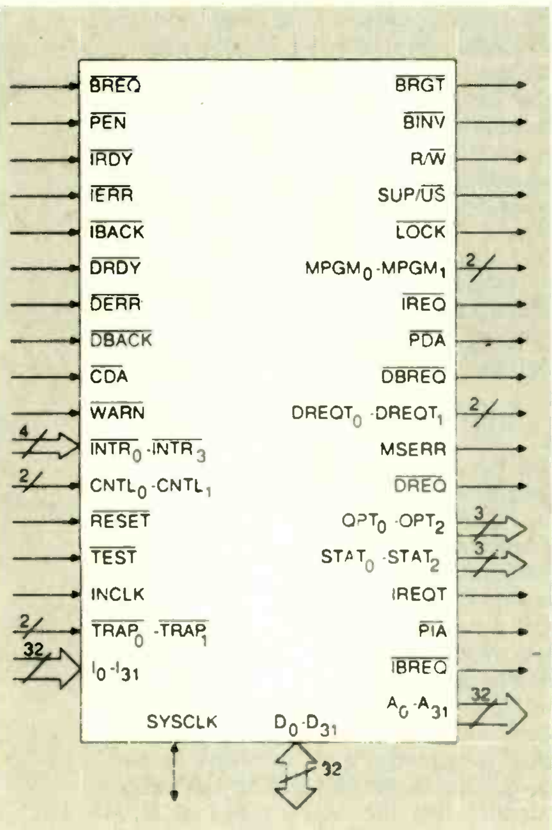
The Am29000 addresses Unix by virtue of an architecture that supports the C programming language. This is done by using part of the register files to increase the speed of C function calls by five to ten times. Optimizing compilers are supported by a three-address load store architecture. Branch conditions are based on Boolean data stored in general-purpose registers instead of using condition codes. Within the C compiler are a number of optimization levels to suit the individual application. For example, when debugging software written in C, minimum optimization will be required because readability of the code will be of prime importance.

DEVELOPMENT TOOLS

Included in the Am29000 development package are hardware design tools such as an interactive hardware simulator of Am29000-based systems. Software development tools currently include an optimizing compiler for C, high performance maths libraries and an assembler-linker. Fortran, Pascal, and Ada will be available later. Debugging tools include the XRAY29K, a source-level symbolic debugger for C; a target-resident monitor; and ADAPT29K, the Advanced Development and Prototyping Tool.

To help the designer obtain the highest performance/cost ratio from his proposed 29000 system, an architectural simulator is provided. It gives the designer a detailed, clock-for-clock simulation of the behaviour of the entire system, exclusive of input/output. It allows the user to specify the size, speed and organisation of both cache and main memory. Results from the simulation will tell the designer what the program 'mix' is – for example, what the hit ratio of the translational look-aside buffer and the branch target caches are, and the overall system speed in Mips.

More advanced technology will allow even higher levels of integration and faster processing speeds. One option will be to integrate the arithmetic code-processor with the c.p.u. while maintaining software compatibility. In the next generation of products will be a 30MHz device, and development will eventually lead to a device capable of sustaining 37Mips.



Above: control signals and buses of the Am29000 processor.

Table 1. Typical simulation for a C program.

```
Time in cycles is 4246
Simulation complete -- failure termination
Statistics of "long" simulation:

User Mode:           8901 cycles (0.00035604 seconds)
Supervisor Mode:     2797 cycles (0.00011188 seconds)
Total:              11698 cycles (0.00046792 seconds)

Instructions executed:     8733

Simulation speed: 18.66Mips (1.34 cycles per
instruction)

-----------Configuration---------------------------
Inst memory decode time = 0
Inst memory access time = 1
Inst memory burst  time = 1
Inst memory pipelined   = TRUE
Inst memory burstable   = TRUE

Data memory decode time = 0
Data memory access time = 1
Data memory burst  time = 1
Data memory pipelined   = TRUE
Data memory burstable   = TRUE

Inst rom    decode time = 0
Inst rom    access time = 2
Inst rom    burst  time = 2
Inst rom    pipelined   = FALSE
Inst rom    burstable   = FALSE

------------Pipeline-------------------------------
25.35% idle pipeline
20.61% instruction fetch wait
  2.24% data transaction wait
  0.51% page boundary crossing fetch wait
  0.02% unfilled cache fetch wait
  0.23% load/store multiple executing
  1.07% load/load transaction wait
  0.67% pipeline latency

-----------Branch target cache---------------------
Branch cache access:           3750
Branch cache hits:             2549
Branch cache hit ratio:      67.97%

----Translation look-aside buffer------------------
T.l.b. access:                    0
T.l.b. hits:                      0
T.l.b. hit ratio:             0.00%

----------Bus utilization--------------------------
Inst bus utilization:      67.38%
                       7882 instruction fetches

Data bus utilization:      12.00%
                       724 loads
                       690 stores

---------Instruction mix---------------------------
  1.66% Calls
15.94% Jumps
  8.29% Loads
  7.79% Stores
  7.82% No-ops

---Register file spilling/filling------------------
  0 spills
  0 fills
```

Mark Lehrer is a field applications engineer with Rapid Silicon.

MC88100 risc

continued from page 642

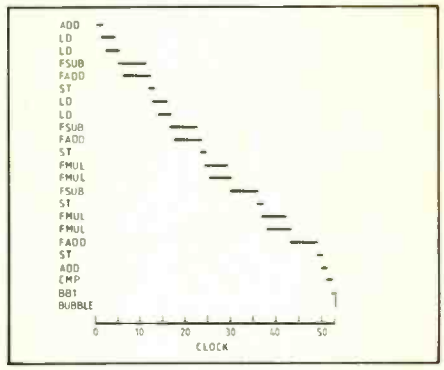


Fig.9. Cycle-by-cycle analysis of the unoptimized assembler-code sequence. Execution of 22 instructions takes 54 clock cycles. For a 20MHz processor, this corresponds to 3.7Mflops.

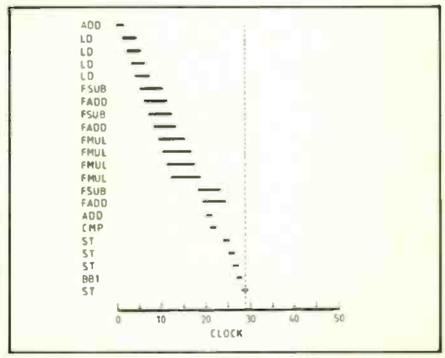


Fig.10. Optimized FFT code reveals better performance. There is more concurrent instruction execution. For the same 20MHz processor, the rate is now 6.9Mflops.

program for a fast Fourier transform. If this code is compiled using a simple C compiler containing no optimizing techniques, then straight-line code would result. List 2 shows a portion of this code for the inner loop of the program. A clock-by-clock analysis of this is shown in Fig.9, yielding a performance of 3.7 million floating point operations per second (Mflops) from a 20MHz clock.

While 3.7Mflops is a respectable performance from any microprocessor, if the timings are analysed carefully there are a number of areas in which improvements can be made. If the load and store instructions are grouped together then the use of the pipelines in the data and instruction units can hide the extra clock cycles required for data accesses. In addition, grouping the floating-point instructions together takes advantage of two pipelines, allowing a considerable amount of concurrency to take place.

Assembly code in List 3 demonstrates the output from a compiler that has an optimizer to produce carefully ordered code that takes advantage of the 88100 architecture. Executing this code on the processor reveals an almost doubling in performance as shown in Fig.10. Therefore highly optimized compilers can give performance figures that are very close to those of assembly language programs, yet enable software development to proceed much faster and easier.

David Jones is a microprocessor applications engineer at Motorola's East Kilbride plant.

Science v. subjectivism in audio engineering

Douglas Self fights his way through the undergrowth of pseudoscience and magic that surrounds the subject of audio engineering.

D.R.G. SELF

Audio engineering is a subject in a singular position. It must be the only branch of engineering science today in which a significant minority of its practitioners, and a very vocal majority of those commenting on it, profess to have no idea how to set about it. Mercifully very few of my comments will apply to the professional audio field, where an intimate acquaintance with the original sound, and the need to earn a living with reliable and economical equipment, have provided an effective barrier against most of the irrational influences.

Most fields of technology have precise measures of excellence: car makers compete to improve miles per hour and miles per gallon; computer manufacturers quote millions of instructions per second and cost per kilobyte, and so on. Any improvement in these objective quantities is regarded as unequivocally a step forward. In the field of audio design, many people seem to have difficulty in deciding which direction forward is.

Working as a professional audio designer, I often encounter opinions, some rational, some not, which while they may be part of the mainstream of hi fi are treated with ridicule by practitioners of other branches of electrical engineering. In the course of publishing some preamplifier designs I have often triggered debates, usually concentrating on subjectivist claims, and sadly ignoring the technical merits or otherwise of each design.

I have often been told that audio is not far removed from witchcraft, that no-one truly knows what they are doing, that I am certainly no exception.

I have even been told at exhibitions that the operation of the human ear is so complex that its interaction with measurable parameters lies forever beyond human comprehension, though this is perhaps an extreme position.

This article deals mainly with the controversies surrounding amplifier design, as here the objective side of the argument is most closely defined; transducers such as cartridges and speakers are inherently subject to greater uncertainties due to the large number of variables involved in electro/mechanical/acoustical conversion.

Those pursuing the correspondence on 'capacitor sound' in the letters section of this journal and others will know that, having studied the subject from the viewpoints of electronic design, psychoacoustics, and my own humble musical efforts, my own view is one of scepticism towards most of the subjectivist hypotheses. (Throughout this article I shall use the word 'hypothesis' in its proper sense as an idea or supposition which has no necessary connection with the truth; in contrast a 'theory' is a logical and self-consistent explanation of real experimental results.) Although often invited to jettison the knowledge gained from these fields, I am not likely to do so purely because some people passionately assert that logic-defying acoustic differences are real for them.

At this point I would like to make it quite clear that I have no doubt that most of the esoteric opinions are held in complete sincerity. Also, if hitherto unsuspected dimensions of audio quality are ever shown to exist, then I plan to be one of the first to make use of them.

It is not my purpose simply to propagate a currently unfashionable point of view; I am attempting a dispassionate analysis of a situation that is perhaps unique: an engineering discipline that seems to have wilfully lost itself in confusion.

THE SUBJECTIVIST MANIFESTO

A concise definition of the subjectivist position on amplifier design might perhaps read as follows.

- Objective measurements of an amplifier's performance are unimportant compared with the subjective impression gained by listening tests. Where the two contradict the objective results may be dismissed.
- It therefore follows that degradation effects exist in amplifiers that are unknown to orthodox engineering science, and that are not revealed by the usual objective tests.
- Considerable latitude may therefore be employed in suggesting hypothetical mechanisms of audio impairment, such as mysterious capacitor shortcomings and subtle cable defects, without reference to the plausibility of the concept, or the gathering of objective evidence of any kind.

I hope that this is considered a reasonable statement of the situation; in passing one might mention that the great majority of the paying public go on buying 'midi' rack systems, and ignore the whole esoteric high-end sector where the debate is fiercest.

In contemplating the state of hi-fi today, I looked for parallels in other fields. The destruction of the study of genetics under Lysenko in the USSR[1], or perhaps the study of parapsychology seemed to be the only candidates. Lysenkoism is arguably irrelevant because external pressures (the need for the Soviet government to increase agricultural production as fast as possible) made it possible for Lysenko, a scientifically illiterate charlatan, to hi-jack the debate about his claims simply by accusing his opponents of obstruction. Opposition was not only futile but also dangerous; however I am assuming I will survive the publication of this article. Parapsychology as a quasi-serious subject is in deep trouble because after some 100 years of controlled investigation it has not uncovered the ghost (sorry) of a repeatable phenomenon[2]. This has a familiar feel to it. It could, however, be argued that parapsychology is also a poor analogy because most people would accept that there was nothing there to study in the first place, whereas nobody would assert that objective measurements and subjective sound quality have no correlation at all; one need only pick up the telephone to remind oneself what a 4kHz bandwidth and 10% or so t.h.d. sounds like. Parapsychology also labours under the crippling disability of being riddled with outright fraud.

SHORT HISTORY OF SUBJECTIVISM

The early history of sound reproduction is interesting, if only for the number of times that observers report that a given acoustic gramophone gave results indistinguishable from reality. This would certainly not be the case today, but the mere existence of such statements throws an interesting light on how mind-set affects subjective impressions. As interest in sound reproduction intensified in the post-war period and technical standards such as DIN 45-500 were set, though they soon came to be criticized as too permissive. By the late 1960s it was widely accepted that the requirements for hi-fi would be satisfied by something along the lines of "t.h.d. less than 0.1%, with no crossover distortion worth mentioning, frequency response 20-20kHz, and as little noise as possible, please". The early 1970s saw this expanded to include specification of slew-rates, properly behaved current-limiters and the like, but the approach was always scientific and it was quite usual to read amplifier reviews in which lab. test results were dissected but no mention at all made of listening tests.

Following the growth of subjectivism through the pages of one of the leading subjectivist magazines (*HiFi News*), the first intimation of what was to come was the commencement of Paul Messenger's column "Subjective Sounds" in September 1976, in which he said "The assessment will be (almost) purely subjective, which has both strengths and weaknesses, as the inclusion of laboratory data would involve too much time and space, and although the ear may be the most fallible, it is also the most sensitive evaluation instrument". Subjectivism as expedient rather than policy. Significantly, none of the early instalments contained references to amplifier sound. In March 1977, an article by Jean Hiraga was published vilifying high levels of negative feedback and praising the sound of an amplifier with 2% t.h.d. In the same issue, Paul Messenger stated that a Radford valve amplifier sounded better than a transistor one, and by the end of the year the amplifier-sound bandwagon was rolling. Hiraga was back in August 1977 with a highly contentious set of claims about audible speaker cables, and after that no hypothesis was too unlikely to receive attention.

THE LIMITS OF HEARING

Before evaluating the subjectivist position, it is relevant to consider the known abilities of the human ear. Contrary to the impression given by some commentators (who call for more psychoacoustical research at the drop of a tone-arm) a vast amount of hard scientific information already exists on ths subject, and some of it may be briefly summarized as follows.

● The smallest step-change in amplitude that can be detected is about 0.3dB for a pure tone. In more realistic situations it is 0.5 to 1.0dB[3]. This is about a 10% change.

● The smallest detectable change in frequency of a tone is about 0.2% in the band 500Hz-2kHz. In percentage terms, this is the parameter for which the ear is at its most sensitive[4].

● The least detectable amount of harmonic distortion is not an easy figure to set, as there is a multitude of variables involved, and in particular the continuously varying level of programme means that the level of t.h.d. introduced is also dynamically changing. However, a lot of careful work has been done on the subject and in normal circumstances (i.e., mostly low-order harmonics) the just-detectable amount is about 1%, though crossover effects can be picked up at 0.3%. There is certainly no evidence that an amplifier producing, say 0.001% sounds any cleaner than one producing .01%[5].

I would not dispute that t.h.d. measurements, done with the usual notch-type analyser, are of limited use in predicting the subjective impairment a given audio path will create. When music is in question, intermodulation effects are demonstrably more important. However, t.h.d. tests have the unique advantage that visual inspection of the distortion residual can give an experienced observer a great deal of information about the cause of the non-linearity.

Other distortion tests exist which, while yielding less information to the designer, exercise the whole audio bandwidth at once and correlate well with properly-conducted tests for subjective impairment by distortion. The Belcher intermodulation test (the principle is shown in Fig.1) deserves more attention than it has received, and may become more popular now that digital signal-processing i.cs are readily available.

One of the objections often made to t.h.d. tests is that their resolution does not allow verification that no non-linearities exist at very low level — presumably a sort of micro-crossover distortion. Hawksford, for example, has stated "Low-level threshold phenomena . . . set bounds upon the ultimate transparency of an audio system"[6] and Duncan[7] has put on record his belief that some metallic contacts consist of a net of 'micro-diodes'.

In fact, this kind of hypothesis can be disposed of using t.h.d. techniques. Having evolved a method of measuring t.h.d. down to 0.01% at 200 microvolts r.m.s. I applied it to large electrolytics, connectors of varying provenance, and lengths of copper cable with and without allegedly magic properties. The method required the design of an ultra-low noise (EIN= −150 dBμ with a 1Ω source resistance — is this a record?) and very low t.h.d.*. The basic method is shown in Fig.2; using an attenuator with a very low value of resistance keeps the Johnson noise to a minimum. In no case was any distortion at all detected, and it would be nice to think that this red herring at least has been laid to rest.

● Interchannel crosstalk can obviously degrade stereo separation, but the effect is not detectable until it is worse than 20dB, which would be a very bad amplifier indeed[9].

● Phase and group delay have been an area of dispute for a long time, and it would take a big book to give all the contending arguments a fair crack of the woofer. As Stanley Lipshitz *et al*[10] have pointed out, these effects are obviously perceptible if they become gross enough; if an amplifier was so heroically misconceived as to produce the top half of the audio spectrum three hours after the bottom, there would be little room for argument. In more practical terms, and leaving aside specialized laboratory test-signals, concern about phase problems has centred on loudspeakers and their cross-overs, as this would seem to be the only place where a phase-shift might exist without an accompanying frequency-response change to make it obvious.

Lipshitz[10] appears to have demonstrated that a second-order all-pass filter (one that gives phase-shift without frequency-response deviation) is audible, whereas BBC findings, as reported by Harwood[11] indicate the opposite, and the truth of the matter is still far from clear.

Fortunately, this controversy is of little importance to amplifier designers, as it would take spectacular incompetence to produce a circuit that included an unsuspected all-pass filter. Without such filters, the phase response of an amplifier is completely defined by its frequency response, and vice-versa; in control theory, this is called Bode's Second Law[12], and it deserves to be much more widely known in the hi-fi world than it is. A properly designed amplifier has its response roll-off points not far outside the audio band, and these will have accompanying phase-shifts; there is however no evidence that these are in any way perceptible[7].

The picture of the ear that emerges is not that of a precision instrument. Its ultimate sensitivity, directional capabilities and dynamic range are far more impressive than its ability to measure level changes or detect

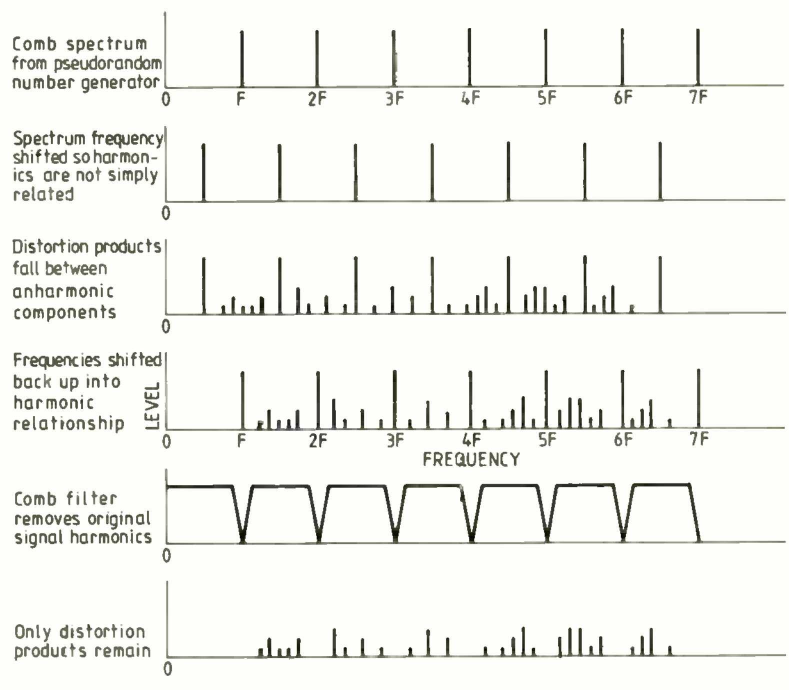


Fig.1. Basic principle of Belcher intermodulation test.

correlated low-level signals such as harmonic distortion products. This is unsurprising; from an evolutionary viewpoint the functions of the ear are to warn of approaching danger (sensitivity and direction being paramount) and for speech communication. For the latter, the perception of formants, the resonance-emphasized bands from the vocal-chord pulse excitation, and vowel/consonant discriminations, are far more important than the usual hi-fi parameters.

Presumably the whole existence of music as a source of pleasure is an accidental artefact of our remarkable powers of speech perception: why it should be able to act as a direct route to the emotions, however, remains profoundly mysterious.

All of the alleged effects listed below have received considerable affirmation in the audio press, to the point where some are treated as facts, though the reality is that none of them has in the last ten years proved susceptible to objective confirmation; a sad record that is perhaps equalled only by students of parapsychology. I hope that the brief statements below are considered fair by their proponents. If not I have no doubt I shall soon hear about it.

- Capacitors affect the signal passing through them in a way invisible to distortion measurements.
- Passing an audio signal through cables, p.c.b. tracks, or switch contacts causes a cumulative deterioration. Precious metal contact surfaces alleviate but do not eliminate the problem. This too is undetectable by tests for non-linearity.
- The sound of valves is inherently superior to that of any semiconductor, despite intractable problems of linearity, reliablity and the need for intimidatingly expensive iron-cored transformers.
- Negative feedback is inherently a bad thing; the less it is used, the better the amplifier sounds, without qualification.
- Tone-controls cause an audible deterioration even when set to the flat position. This is usually blamed on "phase-shift".
- The design of the power supply has subtle effects on the sound, quite apart from ordinary dangers like ripple injection.
- Monobloc construction (i.e. two separate power amplifier boxes) is always audibly superior, due to the reduction in crosstalk.

There is not space to comment properly on all of these viewpoints in this article; each deserves an article to itself, and in any case the list is far from exhaustive. However I will just make a few points. Several people have praised the technique of subtracting pulse signals passed through two different sorts of capacitor, claiming that the non-zero residue proves that capacitors can introduce audible errors. My own view is that these tests expose only well-known capacitor shortcomings such as dielectric absorption and series resistance, plus perhaps the vulnerability of the dielectric film in electrolytics to reverse-biasing. No-one has yet shown how these relate to capacitor audibility in a properly designed amplifier.

A touching lack of faith in cables is widespread, but it can be said with confidence that there is as yet not a shred of evidence to support it. Any piece of wire passes a sinewave with unmeasurable distortion, and so simple notions of inter-crystal rectification can be discounted, quite apart from the fact that such behaviour is absolutely ruled out by materials science. No plausible means of detecting, let alone measuring, cable degradation has ever been proposed.

The valve sound is one phenomenon that seems to have a real existence; it has after all been known for a long time that listeners often prefer to have a small amount of second-harmonic distortion mixed in[13], and all too many valve amplifiers do just that, due to grave difficulties in providing good linearity with modest amounts of feedback. While this may well sound nice, hi-fi is supposedly about accuracy, and if the sound is to be thus modified it should be controllable from the front panel by a 'niceness' knob.

Negative feedback is not inherently a bad thing; it is an absolutely indispensable principle of electronic design, and if used properly has the remarkable ability to make just about every parameter better. However, like all powerful techniques, it can be misused, and too high a feedback factor can require heavy dominant-pole compensation, restricting maximum slew rates. Normally this is not too hard to avoid. In fact slew rates do not need to be very high for disc and f.m. programme sources; Baxandall[14] argues convincingly that a clean full output at 2.2kHz is sufficient, though most people would prefer a good safety-margin over this. The arrival of digital audio has also rendered this figure rather dubious, and there seems no reason why all audio equipment should not be designed to reproduce an undistorted 20kHz sinewave at maximum amplitude.

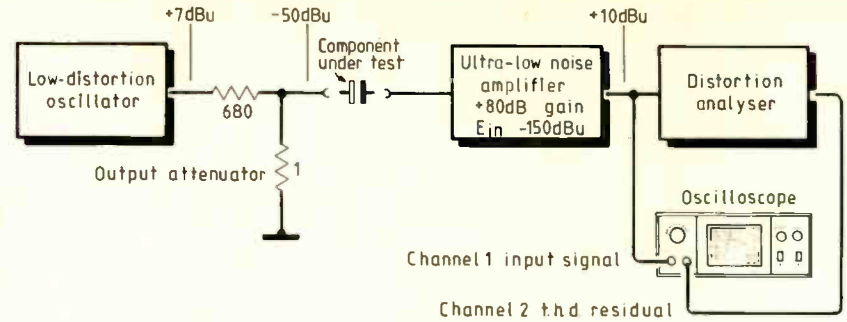


Fig.2. T.h.d. measurement at very low levels.

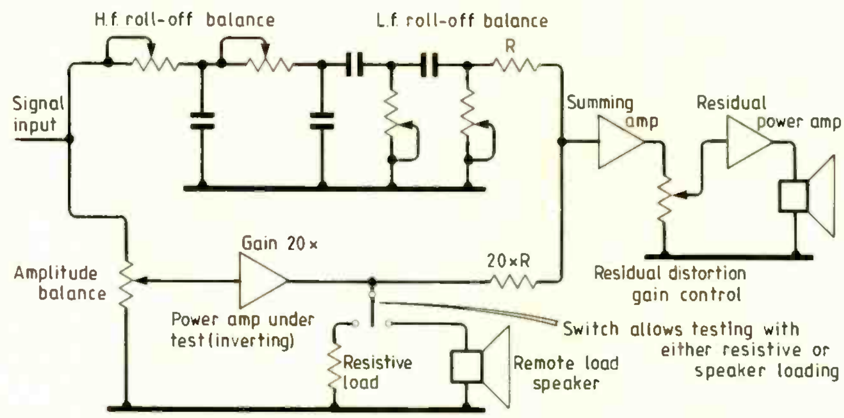


Fig.3. Baxandall cancellation technique.

At the time of writing, tone controls on a preamp badly damage its chances of street (or rather sitting-room) credibility, for no good reason. As pointed out above, well-designed tone-controls set to 'flat' cannot possibly contribute any extra phase-shift and must be inaudible. My own view is that they are indispensable for correcting room acoustics or tonal balance of the material, and that a lot of people are suffering for their devotion to fashion. It might be interesting to try and correlate minimalism in music with minimalism in the equipment designed to reproduce it.

All good amplifier stages ignore imperfections in their power supplies, op-amps in particular excelling at 'power-supply rejection-ratio'. More nonsense has been written on the subject of subtle p.s.u. failings than on most audio topics; recommendations of hard-wiring the mains or using gold-plated 13A plugs would seem to hold no shred of rationality, in view of the usual processes of rectification and regulation that the raw a.c. undergoes.

There is no need to go to the expense of monobloc power amplifiers in order to keep crosstalk under control, even when making it substantially better than the −20dB that is actually necessary. The techniques are conventional; the last preamplifier I designd managed an easy −90dB at 10kHz without anything other than the usual precautions. In this area dedicated followers of fashion pay dearly for the privilege, buying chassis, mains transformer, etc. twice over.

If the above hypotheses are taken as true, and yet their effects are not detectable by conventional measurement, then there are certain implications which need further exploration. Firstly, it can presumably be taken as axiomatic that for each hypothetical defect *some* supposed change occurs in the pattern of pressure fluctuations applied to the ears, and that therefore a corresponding modification has occurred to the electrical signal passing through the amplifier. Any other starting point supposes we are dealing with magic or forces-as-yet-unknown-to-science, and mercifully no commentator has (so far) suggested this. Hence there must *be* defects in the signals, but these are not revealed by the usual tests.

There seem two possible explanations for this failure of detection: the standard measurements are relevant, but of insufficient resolution, although there is no evidence whatsoever that such micro-deviations are audible under any circumstances; or the standard measurements miss the point by, for example, failing in sinewave t.h.d. tests to excite subtle distortion mechanisms that react only to music. This assumes that the music-only distortions are also left undisturbed by multi-tone intermodulation tests, and even the complex pseudo-random signals used in the Belcher distortion test[15]. The Belcher method effectively tests the audio path at all frequencies at once, and it is hard to conceive of a real defect that could escape it. The general principle was shown in Fig 1.

Most damagingly, these alleged music-only mechanisms are not even revealed by music; the devastatingly simple technique of subtracting before-and-after amplifier signals and showing that nothing audibly detectable remains would, it might be thought, have finally shown as non-existent these elusive degradation mechanisms. The technique was proposed in a reasonably sophisticated form by Baxandall in 1977[16]. The principle is shown in Fig.3; note that careful adjustment of the rolloff-balance network prevents minor variations in bandwidth from swamping the true distortion residual. It is fair to say that the subjectivist camp made no effective reply.

A simplified version has been proposed recently by Hafler[17] which, while less sensitive, has the advantage that there is less electronics in the signal path to argue about. See Fig.4. A prominent subjectivist reviewer, on trying this test, was reduced to claiming that the passive switchbox used to implement the Hafler test was causing so much sonic degradation that all amplifier performance was swamped[18]. I do not feel that this is a tenable position. So far all experiments such as these have been brushed aside by the subjectivist camp; no attempt has been made to answer the extremely serious objections that this sort of demonstration raises. Similarly, in the ten years that have elapsed since the emergence of the Subjectivist Tendency, no hitherto unsuspected parameter of audio quality have emerged.

THE LENGTH OF THE CHAIN

One apparently insurmountable objection to obsessive concern with non-measurable amplifier quirks is that recorded sound of almost any pedigree has passed through a complex mixing console at least once; prominent parts like vocals or lead guitar will almost certainly have passed through at least

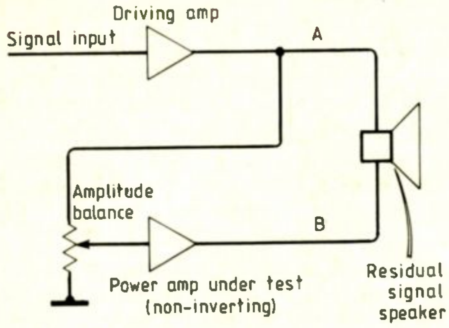


Fig.4. Hafler 'straight-wire' differential test.

twice, once for recording and once at mixdown. More significantly, it must have passed through the potential quality-bottleneck of an analogue tape machine or the ninth-order elliptical filters of digital storage. In its long path from here to ear the audio will have passed through at least a hundred op-amps, dozens of connectors and several hundred metres of ordinary screened cable. If mystical degradation can occur, it defies reason to insist that those introduced by the last 1% of the path are the critical ones.

THE IMPLICATIONS

The confused state of the art of amplifier design has many consequences, few of them good.

Firstly, if equipment is reviewed with results that appear arbitrary, and which are in particular incapable of replication or confirmation, this can be grossly unfair to manufacturers who lose out in the lottery. Since subjective assessments cannot be replicated, the commercial success of a given make can depend entirely on the vagaries of fashion. While this is fine in the realm of clothing or soft furnishings, the hi-fi business is still claiming accuracy of reproduction as its raison d'être, and therefore you would expect the technical element to be dominant. This unfair discrimination between manufacturers is complicated by a discernable xenophobia that seems to give British products an inside track.

It is reasonable to advocate the purchase of home products, if you make it clear that that is the reason why they get a better press. Most people active in the audio field remember vividly the Japanese destruction of the original British hi-fi industry, (I will not mention motorcycles) and it is perhaps not surprising that in times of recession there is a subconscious tendency to strike back. This may be more difficult if you make objective tests your criterion.

Another consequence of the dominance of subjectivity over measurements is that it places designers in a most unenviable position.

No degree of ingenuity or attention to technical details can ensure a good review, and the pressure to adopt fashionable and expensive expedients (such as linear-crystal internal wiring) is great, even if the designer is certain that they have no audible effect for good or evil. Most designers would seem to be faced with a choice between swallowing the subjectivist credo whole or keeping very quiet and leaving the talking to the marketing department.

While the double-think here seems trivial compared with, say, the activities of an uninhibited used-car salesman, it should be remembered that engineering aspires to be a profession, with the implication that certain moral as well as technical standards are observed. I for one would feel uncomfortable about foisting expensive and non-functional parts on an unsuspecting customer. It is hard to avoid the conclusion that the man who pays thousands of pounds for a preamp can only be described as a Hifi Victim.

Since objective measurements are given little weight, it would appear inevitable that therefore poor amplifiers will be produced, some so bad that their defects are certainly audible. Reading recent reviews[19], it was easy to find a £795 preamplifier (Counterpoint SA7) that boasted a feeble 12dB disc overload margin, (another preamp costing £2040 struggled up to 15dB — Burmester 838/846) and another, costing £1550 that could only manage a 1kHz distortion performance of 1%; a lack of linearity that would have caused consternation ten years ago (Quicksilver). However, by paying £5700 one could get this down to 0.3% (Audio Research M100-2 monoblocs).

This does not of course mean that it is impossible to buy an 'audiophile' amplifier that does measure well — another example would be the preamplifier/power amplifier combination that provides a respectable disc overload margin of 31dB and 1kHz rated-power distortion below 0.003%; the total cost being £725 (Audiolab 8000C/8000P). I believe this to be a representative sample, and we appear to be in the paradoxical situation that the most expensive equipment provides the worst objective performance. Whatever the rights and wrongs of subjective assessment, I think that most people would agree that this is a strange state of affairs.

Finally, it is surely a morally ambiguous position to persuade non-technical people that to get a really decent sound they have to buy £2000 preamps and so on, when all technical orthodoxy indicates that this is quite unnecessary.

THE REASONS WHY

It is important to try to understand why hi fi has reached the pass that it has; some tentative conclusions are possible. I believe one basic reason is the difficulty of defining the quality of an audio experience; you can't draw a diagram to communicate what something sounded like. In the same way, acoustical memory is more evanescent than visual memory.

It is far easier to visualize what a London bus looks like than to recall the details of a musical performance. Similarily, it is difficult to 'look more closely' — turning up the volume is more like turning up the brightness of a tv picture; once an optimal level is reached, any further increase becomes annoying, then painful.

In the study of experimental psychology, particularly in experiments about perception, it has been universally recognised for many years that people tend to perceive what they want to perceive. This is often called the 'experimenter expectancy' effect; the history of science is littered with the wrecked careers of those who failed to guard against it. It has most often occurred in fields such as biology, where although the raw data obtained may be uncompromisingly numerical, there is no mathematical theory to allow checking. Where the only 'results' are subjective impressions, the danger is clearly much greater, no matter how absolute the integrity of the experimenter. Thus great care is necessary in the use of impartial observers, double-blind techniques, and rigorous statistical tests for significance. The majority of subjectivist writings ignore these precautions, with predictable results. Only in relatively few cases have properly controlled tests been done, and at the time of writing those that have been resulted in different amplifiers appearing to be indistinguishable. I believe the conclusion is inescapable that experimenter expectancy has played a large part in the growth of subjectivism.

It is also interesting that in subjectivist audio the 'correct' answer is always the more expensive or inconvenient one. Electronics is rarely as simple as that.

A major improvement is more likely to be linked with a new circuit topology or new type of semiconductor, than with mindlessly specifying more expensive components of the same type; cars do not go faster with platinum pistons. It would be difficult to produce a rigorous statistical analysis, but it is my strong impression that the reported subjective quality of a piece of equipment correlates far more with the price than with anything else. There is perhaps here a resonance of the Protestant Work Ethic — you must suffer now to enjoy yourself later.

Another reason for the relatively effortless rise of subjectivism is the 'me-too' effect; many people are reluctant to admit that they cannot detect acoustic subtleties — nobody wants to be labelled as insensitive, outmoded, or just plain deaf. It is also virtually impossible absolutely to disprove any of the claims made, as the claimant can always retreat a fraction and say that there was something special about the combination of hardware in use during the disputed tests, or that the phenomena are too delicate for brutal logic to be used on them. In any case, most competent engineers with a taste for objectivity probably have better things to do than dispute every tendentious report. Under these conditions, vague claims tend, by a kind of intellectual inflation, to become regarded as facts. Manufacturers have some incentive to support the subjectivist camp as they can claim that only they understand a particular non-measurable effect; they may of course get their come-uppance if the dice fall badly in a subjective review.

CONCLUSION

It seems unlikely that subjectivism will disappear for some time, given the momentum that it has gained, the entrenched positions that some people have taken up, and the sadly uncritical way in which people accept an unsupported assertion as the truth simply because it is asserted with frequency and conviction. In an ideal world every such statement would be greeted by loud demands for evidence. However, the recent history of the real world, and of British politics in particular, sometimes leads one to suppose pessimistically that people will believe anything. By analogy, one might suppost that subjectivism would persist for the same reason that parapsychology has; there will always be people who will believe what they want to believe rather than what the hard facts indicate.

I hope that people will be moved to comment on the foregoing article. However, it is perhaps worth pointing out that anecdotes of the "I tied a knot in my speaker cables and it sounded much better" school do not constitute a refutation.

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Writing for EWW

Notes for potential authors

Fairly often, people write in or ring the editorial office to suggest a possible article. When we have discussed the project and made any suggestions that come to mind, the potential author asks whether there are any requirements about presentation he should bear in mind.

Well, there are a few, but they are not rigorous and articles are not rejected because the typing is not double-spaced or because it is on green paper.

First of all, the style. Different people write in different ways and there can be no changing that; varying styles have their attractions and there is no "correct" one. We prefer the simple kind, in the active voice; there is no need to use stilted company-report English and the kind of writing that is designed to exhibit your own knowledge rather than to explain the subject.

If you are not used to writing technical articles, there is still little need to worry, because the editorial staff is ever-present to knock things into shape, if required, although we do try to preserve the original as far as is possible. The essentials are that the piece should contain *all* the information, including illustrations and diagrams and, if the article refers to some obscure component, what it is and where it comes from.

Diagrams must be clear enough for our draughtsmen to interpret, but need not be masterpieces, since they will be re-drawn in our own style. It is helpful to have drawings and photographs kept separate from the rest of the article and to have captions on a separate sheet. Photographs can be negatives, glossy prints or transparencies. Typing should, ideally, be double-spaced with wide margins, so that we can scribble notes to the typesetters.

If you have relevant qualifications, please tell us what they are and also let us have a telephone number for daytime use – since we sometimes need to talk to authors in a hurry, particularly on press day!

SATELLITE SYSTEMS

Integrating space and ground navaids

From the variety of radio navigational aids now in existence, we could well end up with a preferred system utilizing a partnership of Earth- and space-borne transmitters. A recent conference at the City University in London revealed that both of the world super-powers are considering the integration of satellite and terrestrial navigation systems. The idea is for each technique to back up and compensate for the limitations of the other, thus providing a combination that meets as many operational requirements as possible.

The great advantage of satnav systems, of course, is that they give global coverage, whereas terrestrial radio navaids only serve particular areas. But the terrestrial systems provide better repeatability, and possible accuracy, for certain purposes (e.g. on fishing vessels). Furthermore the latest satnav schemes (GPS/Navstar and Glonass – see April 1987 issue, pp. 377-378) are still in a pre-operational phase and not yet complete anyway. Another possibility now being explored is the integration of satellite and inertial navigation systems.

At the Royal Institute of Navigation's 1988 international conference, W.L. Polhemus said that a plan for integrating the terrestrial hyperbolic system Loran-C and the satellite Global Positioning System (GPS) had been prepared for use throughout the USA. This plan was the work of members of Congress and specialists from the Department of Transportation. It would integrate the two systems in a "non-dependent but synergistic way" to ensure reliable, redundant signals for "virtually 100% of the year".

Although GPS is expected to have a signal availability of 99.8 to 99.9% when complete and similarly Loran-C provides 99.7 to 99.9%, these figures are several orders of magnitude removed from the desired level of availability. They represent a possible signal absence rate of as much as 1 500 minutes per year. For aeronautical use in the USA a radio navaid must provide, among other things, a continuous signal availability of 99.998% before it can be certified as a 'sole means' of position determination. According to Polhemus, a partnership of Loran-C and GPS would in fact increase their combined availability to this level.

In addition, he said, such an integrated system would provide a means for independent integrity checking and for redundant operation in the event of transmitter failures. It should also eliminate diurnal gaps in coverage and give protection against various forms of interference.

Loran-C, now installed in 11 countries, is the latest and most widely used version of the Loran pulsed hyperbolic navaid. The original system was developed in the USA, like Decca in the UK, for military purposes during the 1939-45 world war. Loran-A, a more advanced version, followed after the war, while a Loran-B was developed but never implemented.

Two speakers from maritime organizations in the Soviet Union, G.I. Moskvin and V. Sorochinsky, outlined a different terrestrial-celestial combination being tried in that part of the world, but first expressed doubts on the advisability of combining Loran-C with a satnav system. Their objection seemed to be one of operational incompatibility: "pulse-phase systems are known to produce more accurate position data within their coverage areas" than that obtainable with satnav systems in those areas. But "other combinations may turn out to be more promising".

This small GPS satnav receiver, the MX4400 from Magnavox Systems Ltd, will determine position with two visible GPS satellites when the user's altitude is known and time is fed in from an external atomic frequency standard. It is a two-channel receiver (see main text) and works from the so-called L_1 signal transmitted from the satellites at 1575.42MHz. Data ports on the unit also allow Loran-C terrestrial and Transit satellite position data to be fed in. Position fixing accuracy is claimed to be better than 15m when the receiver is moving or 5.1m when it is stationary and half hour averaging is used. Land, sea and aircraft L-band antennas are available. Power consumption is 20W and weight is 7.2kg.

Moskvin and Sorochinsky said that the Soviet integrated system, called Biryuza, combined the use of the Russian Cicada and American Transit satellites (April 1987, p.377) with a v.l.f. terrestrial navaid. Quasi-differential techniques were used to calculate corrections to navigational measurements produced by the v.l.f. system. The correction values were derived from a simple algorithm and then refined as soon as any of the Cicada or Transit satellites became available.

By comparison of the actual and measured values, a more accurate forecast of corrections was achieved in the intervals between successive usable satellite transits. This resulted in some improvement in position finding accuracy over that obtainable by other methods. But what they felt to be more important was that "a mutual check-out of position data" became possible because of the availability of two series of position fixes produced by data from two independent systems.

The Russian contributors added that equipment had been developed in the USSR to allow position fixing from just the two satnav systems, Cicada and Transit. An obvious advantage was some 1.8-fold reduction of the intervals between satellite observations. A four-channel shipborne receiver had been designed to operate with both systems – two channels for Cicada and two for Transit. It allowed selection of signals in both frequency and time, so that choices could be made of the most suitable satellites for obtaining position fixes if several were in view simultaneously.

Multiple channels in satnav receivers improve operational speed and flexibility. A single-channel receiver (e.g. as shown in the April 1987 issue, p. 378) will only work sequentially from satellite to satellite, whereas a multi-channel set will operate with a number of satellites simultaneously. The GPS receiver in the photograph above (on show at the conference) has two channels. One of these performs

Fig. 1. Comparison of satellite and terrestrial navaids.

	Terrestrial systems			Satellite systems			
	Decca	Chaika	Loran-C	Cicada	Transit	Glonass	GPS
Coverage (miles)	250–300	1000	1200	global	Up to latitude of ±88°	global	
Positioning accuracy (miles)	0.2–0.5	0.05–0.01	0.05–0.01	0.05–0.3		0.05	
Availability of position fixes	continuous	continuous	continuous	every 40	–110 min	continuous	
Duration of communications	3 fixes/min	10–20 fixes/min	10–20 fixes/min	6–16 min	8–18 min	1 s	

Source: Moskvin and Sorochinsky, RIN 1988 International Conference, London.

SATELLITE SYSTEMS

a general 'housekeeping' function, searching for whatever satellites are available, while the other channel handles the actual position finding process.

J.H. Beattie of Racal believed that there would be an increasing use of hybrid navaid receivers in the future. He mentioned, for example, the combinations GPS/Decca, GPS/Loran-C, GPS/Transit and GPS/Glonass. Regarding the last-mentioned, the two Russian speakers felt that serious consideration should be given to using both of these new satnav systems in conjunction. They said that there will be areas of oceans where the stated accuracy of GPS could not be maintained for some 40 minutes twice a day. Moreover, in the event of failure of one or more satellites, gaps could occur in providing positioning needs throughout overall areas.

A combined use of GPS and Glonass would have the advantage of preventing the appearance of areas of reduced position-finding accuracy. It would also allow a 1.4-times better accuracy of determination of position fixes and velocity vectors. There would be no need for either co-ordination of the two systems or unification of the transmitted signal parameters. It would be quite enough, they felt, to design a shipborne receiver capable of handling signals from both systems.

The accompanying table is the two Russians' comparative assessment of the basic characteristics and performance of current satellite and terrestrial navaids. Note that Transit and Cicada, with their long intervals between position fixes and dependence of accuracy on user velocity, are due to be phased out in about 1996, leaving the field clear for GPS and Glonass. The Chaika (Seagull) system is a Russian terrestrial hyperbolic navaid similar to Loran-C.

• The Institute of Navigation in the USA is organizing a conference this year on "GPS around the world" at the Broadmoore and Antlers Hotels, Colorado Springs, Colorado, USA, September 19-23 1988. It will include a session on integration of GPS with other navaids, as well as sessions on technical developments in equipment and applications.

In brief

Following the failure of the German direction broadcasting satellite TV-Sat 1 last year, the French authorities have asked Arianespace to postpone the launch of their sister d.b. satellite TDF-1 (also built by Eurosatellite) from the originally booked date this summer until the autumn. Now, Ariane flight V24, planned for July 1988, will carry instead of TDF-1 the European ECS-5 comsat and the Indian multi-purpose Insat-1C (January 1988 issue, p. 57). The French d.b.s. craft is due to be launched now on flight V26 in October. Meanwhile a replacement German satellite, TV-Sat 2, is being built for delivery in the spring of 1989.

• First flight of the new, more powerful European rocket Ariane-4 is likely to be No. V22 this summer. Designed to carry heavier payloads, its first task will be to launch three satellites together – Meteosat P2, Pan American Satellite 1 and Amsat IIIC. The Ariane-4 design encompasses six versions with various combinations of engines and boosters for different tasks. These allow payloads of 1900kg to 4200kg for putting satellites into geostationary transfer orbits and up to 7000kg for injection into low circular orbits.

• University of Surrey is building a third UoSAT-OSCAR spacecraft to carry engineering, science and communications experiments arranged collaboratively between professional engineering and amateur radio communities. Organizations likely to take part in this UoSAT-C mission include Amsat-UK, Amsat-NA (USA and Canada), Vita, Quadron, NASA, BNSC and ESA. The new spacecraft is expected to be launched in late 1988, by NASA using a Delta rocket, into a circular orbit of 500km altitude and 43° inclination.

• A career civil servant, Arthur Pryor, is the new director-general of the British National Space Centre. He succeeds Jack Leeming, who took over for a while from Roy Gibson and then retired in February. Mr Pryor joined the old Board of Trade in 1966 and held a variety of positions in the DTI. His last job was as regional director of the West Midlands DTI.

Radio engineering terms in satellite links

Carrier-to-noise ratio

An important performance figure in all communication channels is signal-to-noise ratio (s:n). But in many radiocommunication systems s:n depends on the characteristics of the signal itself, and engineers find it more appropriate, in working out the gain and loss performance of the r.f. parts of links, to use *carrier-to-noise ratio* (c:n). This is particularly relevant with the f.m. widely used in satellite links, as the carrier is transmitted at a constant amplitude which is not affected by the signal waveform. In the receiver's limiter the received carrier frequency is held at a constant amplitude and the noise power goes up and down according to how c:n is affected throughout the link. Of course, the s:n achieved overall by the link can be measured after f.m. demodulation in the receiver.

So the c:n at the receiving end of the complete satellite downlink consists basically of the equation for carrier r.f. power given earlier (May, page 489, under 'power budget') divided by the noise power as discussed. In the logarithmic form of dB, gains are added while loss and noise terms are subtracted to arrive at the carrier-to-noise ratio. To start with, we have the following relationship in decibels:

$$C/N_0 = \text{e.i.r.p.} - L - A + G - 10\log_{10}kT \text{ dBHz}$$

Here, N_0 is the *noise spectral density*. It is not noise power because the right hand side of the equation does not yet contain any term representing the noise bandwidth. Thus C/N_0 means carrier over noise power per Hz, written dBHz. The term $10\log_{10}kT$ (where k is Boltzmann's constant, see below) represents the *noise power density* in the receiving system. The equation can be rearranged to give

$$C/N_0 = \text{e.i.r.p.} - L - A + G_R/T - 10\log_{10}k.$$

G/T in general is a ratio of receiving antenna gain to total noise temperature from all sources, thermal and otherwise, collected by the receiver antenna system (e.g. from sky, earth, low-noise amplifier, antenna and waveguide structure). It is a *figure of merit* for the antenna performance, analogous to signal/noise ratio, and has become a standard performance figure for receiving stations in satcom engineering. Its value is expressed in dB/K (even though there is no natural relationship between a gain and a temperature). Thus, for a receiver antenna with a gain of 40dBi and a system noise temperature of 350K, the logarithmic form required for a value in dB is:

$$G_R/T = 10\log_{10}(10^4/350) = 14.54\text{dB/K}$$

The symbol k on the right hand side of the C/N_0 equation is *Boltzmann's constant* written in dB form. As a constant, its purpose is to convert the noise temperature T in the equation into noise power density. Being a very small number in the denominator of the carrier-power/noise-power-density expression, it therefore makes a very large contribution to the size of the overall c:n. In this C/N_0 equation, the term $10\log_{10}k$ represents −228.6dB.

As an example, using values mentioned above, a downlink with an e.i.r.p. of 60dBW, a free space loss (L) of 205.1dB, an atmospheric attenuation (A) of 5dB, and a receiver antenna G/T of 14.5dB/K would have a carrier to noise ratio given by

$$C/N_0 = 60 - 205.1 - 5 + 14.5 - (-228.6) \text{ dBHz}$$
$$= 93.0\text{dBHz}$$

But, as mentioned above, this figure does not take account of the noise bandwidth, and a term representing this factor in dB must be subtracted. Thus, if the noise bandwidth is 27MHz, or 74.3dBHz, the carrier-power/noise-power ratio C/N is: 93.0dB − 74.3dB = 18.7dB.

In practice, C/N ratios in a range of about 7 to 14dB are found to be adequate in downlinks, depending on the modulation method and the application.

I would like to thank IBA Engineering Information for very helpful discussions during the preparation of this material.

Satellite Systems is compiled by Tom Ivall.

Making waves

J W propagates some views on wave motion and doesn't even try to avoid algebra.

JOULES WATT

After struggling for a number of weeks trying to get a group of people to see at least roughly how radio waves propagate, I gradually realised that the very idea of waves was so imperfectly understood that most of the young trainees in the group would never see how the ionosphere altered the velocity of propagation. At best this was unlikely without the elementary knowledge I was assuming.

"Aha", I though, "more evidence of our neglect of basics in education, in favour perhaps of some fashionable 'black box' or another". As expected, my enquiries elicited that hardly anyone had studied waves, I mean *real* waves, although one young man said he had heard that everything was made of waves. He had been told that a chap called Schroedinger, or someone, had 'proved' it.

Once again an old radical like me found myself in a reactionary position claiming that you can't study radio, propagation, transmission lines — or understand reflection 'glitches' in computer interconnections for that matter — without a good grasp of classical wave theory. And when you consider that half of all physical science, with nearly all communications engineering theory turns out to be rooted firmly in wave motion, I thought my assertion hard to refute, in spite of being out of fashion.

MATHEMATICS AGAIN

Judging by further comments, wave phenomena suffer from a reputation of being a 'mathematical' subject. Once again, in the increasingly innumerate climate as people continue to give up teaching maths to any depth in schools, such a reputation is a damning indictment. How often we see small books make a glowing claim that the author has taken a 'non-mathematical approach', or by some vast verbosity or another 'has reduced mathematics to a minimum'.

I would make two points here. The first is that studying waves is not necessarily 'more mathematical' than, say, mastering circuit design. The second point is that simple mathematical formulation is so economical in explanatory power and induces such precision of thought, that mastering it confers a considerable advantage. Besides, mathematical modelling that works well enables you to design systems and predict performance. Finally, whatever has happened to the fun and pleasure of learning and solving numerical problems. "Education" turns out a miserable lot these days.

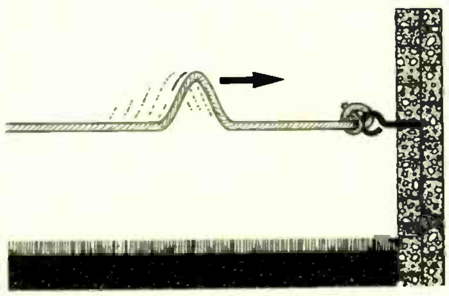


Fig.1. A pulse travelling along, say, a rope shows how little the rope moves, but how fast the energy travels by. A typical wave situation.

However, I would be the last to hint that maths is somehow mystical in its power, that it should be doled out in vast amounts or otherwise employed to give a false air of learnedness. Taking that line moves into the world of the "maths pushers", criticized by other authors. No, the maths should be an applicable tool for the engineer — clear, pragmatic and doing a job of work.

FUNDAMENTALS

Further thought showed how far the basic fundamentals have become neglected. No one in my group had studied co-ordinate geometry, for example. This was confirmed later by a teacher who answered my question with, "That's correct, co-ordinate geometry has been reduced greatly. We cover Boolean algebra and set theory now. But there is considerable work on relations and the more restricted set of these — the functions".

I struggled for a long time to try to explain a little about waves with Boolean algebra, relations and functions but no, readers will have to put up with a quite ordinary approach, I'm afraid.

You should have noticed wave motion all around. The hoary old classic of water ripples which always used to turn up as an analogy when introducing radio transmission[1] is still studied by all the children I know (apparatus: stones and a pond). There are many other wave phenomena which do not go away. The universal international standards of length remain based upon the wave number of a certain Krypton line.

Many a discussion of waves starts by a consideration of 'vibrations'. The oscillator, and theory of simple harmonic motion (s.h.m.), receive a detailed development, which you should continue to find useful[2].

Then waves are introduced via the idea that these vibrations might travel away (through some 'medium'). The oscillations propagate with a certain velocity — and so on. This is fine and anyone who has studied it all most probably would not be reading this.

Yet I have always wondered about such an approach, when many of the most interesting waves are not vibration at all, but are *pulses*, such as the kink travelling along a stretched rope in Fig.1 (again, children still regularly study this phenomenon), or perhaps as digital pulses going along a data bus.

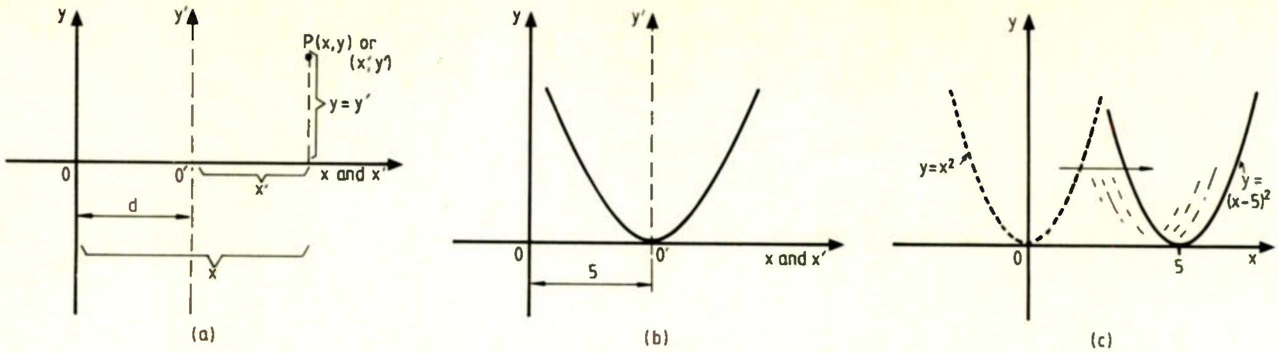
One of the earliest mathematical difficulties many beginners seem to experience when they start to look at waves centres on the fact that the wave definitely *travels*, so that energy propagates from place to place, but the medium itself moves very little. Then later, to compound the problem, you may find that some waves 'stand' or remain stationary after all. The condition in which the wave moves means that the position of the wave's outline or *profile* varies as a function of time as well as being a function of distance.

Because of this, the wave function has two independent variables, a fact that immediately induces many people to assume a great increase in complexity. But, in my previous perambulations[3], we often met functions of several variables. Agreed, any differentiation of functions of several variables means *partial* differentials, yet most people feeding back views concerning the earlier discussions said that these were 'pretty straightforward'. The thing to do every time, is to keep all the other variables constant while varying the one you are interested in. Then to get the total result, add all the partial bits thus obtained; at least, it is as simple as that in *linear* systems.

There are two ways of looking at a profile of a plotted function which varies in its position on the axes: either you have a moving function (going "along the axis"); or you move the position of the origin instead. No doubt it is best to imagine the wave as a moving function plot on fixed axes, whereas in co-ordinate geometry we often keep the function fixed and move the origin.

Now physically, all motion is relative, so in effect either of the above scenarios says the same thing; either you sit on the origin and see the wave go by, or you can sit on the wave and travel with it and watch the origin recede.

Take a look at Fig.2(a). This is just a quick revision of the coordinate origin shift (along



one axis for simplicity). If you place point P at (x,y) relative to origin 0 and therefore at (x',y') relative to origin 0', then with 0' at distance d from 0, the coordinates relate according to $x = x' + d$. The y coordinate remains unchanged for this one-dimensional shift.

For example, if you plot the quadratic function $y = x^2 - 10x + 25$ on the original axes and then shift the origin to 0' where $d = 5$, then the new form this function takes relative to 0' is,

$$y' = (x' + 5)^2 - 10(x' + 5) + 25$$

by putting in the $x' + 5$ in place of x. Tidying up gives the new form in terms of x', namely, $y' = x'^2$. You can see that 0' sits right on the apex of the parabola, as Fig.2(b) shows.

The other way of looking at this problem means that you can consider that the apex of the parabola $y = x^2$ has travelled away from 0, out to $x = 5$ as in Fig.2(c). For y to equal 0 in this second position, x has to equal 5 and placing $x = 5$ into $y = (x - 5)^2$ duly makes it so. In a small book "Waves"[4], C.A. Coulson gives alternative reasoning as follows.

"Imagine taking a snapshot of the wave profile at t = 0. You observe y = f(x). Another snapshot taken t seconds later shows an identical picture, but with the profile moved ct metres to the right. If we took a new origin at the point ct and let distances measured from this be called X, then x = X + ct. Now y = f(X) and looks exactly the same (with this origin) as the first snapshot. But referred to the original fixed origin y = f(x – ct)".

In general, *subtracting* the shifting quantity τ, say, from the independent variable yields how far your function has moved to the *right*. In symbols this means $f(x - \tau)$ is function x moved a distance τ to the right.

Now consider any function. If we plot it we get a *profile*. From an engineering point of view, most functions of interest produce profiles which drop off towards zero in both directions — we do not have much to do with those that rise to infinity. This kind of shape that starts off small, rises up to a peak and then drops off again forms a typical pulse. If you imagine a function f(x) of this form and think of it moving towards the right with a velocity c, then after a time t it will be at a position ct along the axis, according to the above reasoning. In other words, the function will vary with x and t according to $f(x - ct)$. If the profile moves to the left, then the description becomes $f(x + ct)$, see Fig.3.

Now whatever the profile, and for our purposes they are reasonable ones like a Gaussian shape, or perhaps a rectangular

Fig.2. In (a) you see the standard way axes enables reference to a point from various origins and the relation between co-ordinates. In (b), the origin has been shifted to the point (5,0) relative to the original. The situation in (c) says the same thing, but now we consider the origin fixed and the plotted function moved by the same amount.

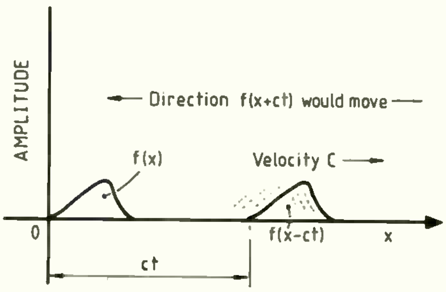


Fig.3. If the shifting term depends on a velocity, then as time ticks away the function travels off to the right at this velocity. If you change the sign of the shifting term, the function travels away in the opposite direction.

pulse, we often assume at this point in the discussion that the shape does not change as it travels along x. This rather dubious assumption (in general) means the *form* of f(x) stays the same. In addition we assume no dissipation or damping occurs. These simplifications yield functions we can differentiate both with respect to time and with respect to distance. Take $y = f(x - ct)$, and differentiate with respect to x,

$$\frac{\partial f}{\partial x} = f'(x-ct) \text{ and again } \frac{\partial^2 f}{\partial x^2} = f''(x-ct)$$

(keeping t fixed of course).

Now keep x constant and differentiate with respect to t,

$$\frac{\partial f}{\partial t} = \frac{df}{d(x-ct)} \cdot \frac{\partial(x-ct)}{\partial t} = -cf'(x-ct)$$

and again,

$$\frac{\partial^2 f}{\partial t^2} = c^2 f''(x-ct)$$

So whatever f, if it is behaving all right under differentiation, it fully satisfies the partial differential equation.

$$\frac{\partial^2 f}{\partial^2 x} = \frac{1}{c^2} \frac{\partial^2 f}{\partial^2 t}$$

as you can show by eliminating $f''(x - ct)$, putting the c^2 on the "t" side and dividing through.

We have here the standard equation that all non-dispersive wavemotion obeys, whatever is doing the waving and in whatever way it is doing it. We made no assumption about f except that it was well behaved from an engineering point of view and that it did not lose its shape, which is the meaning of non-dispersive. The wave velocity c is always obtained straight away, so that often you will not even need to solve the differential equation to get useful information from it.

MAXWELL VINDICATED

In my discussion of James Clerk Maxwell's remarkable work[5], his derivation of just such an equation, showing that **E** and **H** fields moved with a wave motion whose velocity c turned out to be the same as that for light, was the work that revolutionized electromagnetic theory. The equation,

$$\frac{\partial^2 \mathbf{E}}{\partial x^2} = \frac{1}{\mu_0 \epsilon_0} \frac{\partial^2 \mathbf{E}}{\partial t^2}$$

shows that electric waves travel in a vacuum at a velocity

$$c = \frac{1}{\sqrt{\mu_0 \epsilon_0}},$$

where $\mu_0 = 4\pi \times 10^{-7}$ henries per metre is the permeability and $\epsilon_0 = 10^{-9}/36\pi$ farads per metre is the permittivity

$$\therefore \quad c = 3 \times 10^8 \text{ ms}^{-1} \text{ nearly.}$$

From the relations between **E** and **H**, an accompanying magnetic wave always appears, so that the wave always has an electromagnetic nature with the **E** and **H** vectors remaining at right angles with each other and with the direction of propagation. EM propagation forms a *transverse* wavemotion. In a similar way, while discussing transmission lines a wave equation arose there[6]. Again the wave velocity appeared naturally in the form of the wave equation.

ANOTHER LOOK AT THE EQUATION

The wave propagation equation often receives the title of the mathematician who solved it in a particular way. This was d'Alembert. He obtained the two functions $f(x - ct)$ and $f(x + ct)$ from it, and therefore showed that waves travelling in both directions form solutions to it. I have already mentioned the interesting fact that vibration or periodicity has not necessarily appeared. Of course, periodic waves such as sinusoidal variations interest us greatly, but they are only special or simple cases of what is a very general result.

Yet harmonic waves, because of their

simplicity, turn out to be solutions to many more complicated wave type equations than d'Alembert's. If we stick to harmonic waves for a moment and write,

$$y = A\sin\beta(x - ct)$$

then the wave has a sine form (profile) travelling to the right without loss, as in Fig.4. Its peak value (crest) is A units above the mean level and its shape reproduces itself every 2 radians, thinking in terms of angle. Because of this periodicity, the phase, which is the angle $(x-ct)$, repeats every now and then if you look along x at a fixed time (i.e. take a snapshot of the wave) or it goes through repeated cycles at the same point (fixed x) over a period of time as you watch. Thus you find periodic variations in space or wavelength associated with the motion, as well as repeatability in time — in other words frequency — as you watch.

PERIODIC WAVES

In the case of $y = A\sin\beta(x-ct)$, you can find a number of parameters that describe the motion. They arise from the 2π periodicity.

For example, at a certain instant, the wave profile repeats at a distance λ metres further on. This defines the *wavelength*. In other words, $y = A\sin[\beta(x-ct)+2\pi]$ assumes the same value as the original y.

$$\therefore \quad y=A\sin\beta\left(x+\frac{2\pi}{\beta}-ct\right)$$

which shows that adding $2\pi/\beta$ onto x reproduces the wave profile.

We put $2\pi/\beta=\lambda$ so that $\beta(=2\pi/\lambda)$ is the number of waves in a distance of 2πmetres. Some people call β the *wave number*, others define wave numbers as $k=1/\lambda$, ie. the number of waves in a metre.

Electronic engineers usually call β radians m^{-1}, the *phase constant*, because it gives you the number of radians shift in travelling a distance d along the wave.

$$\beta=\frac{2\pi}{\lambda}d \text{ radians.}$$

Suppose now you observe over a period of time at a fixed point x, then,

$$y=A\sin\beta\left(x-c\left[t-\frac{2\pi}{c\beta}\right]\right)$$

by inserting the 2π into the time term and re-arranging in standard form. The wave therefore repeats after a time $T=2\pi/c\beta$. We identify this time T as the *period*, so that you have immediately from its reciprocal the *frequency* f Hz of the oscillating wavemotion. This means,

$$f=\frac{c\beta}{2\pi} \text{ or } \frac{c}{\lambda}$$

By simple transposing, the other way of writing this turns out to be important in later work, namely,

$$c=\frac{\omega}{\beta},$$

where ω is the familiar $2\pi f$. This shows if you find β, and know ω, then dividing them gives the wave, or *phase* velocity, c.

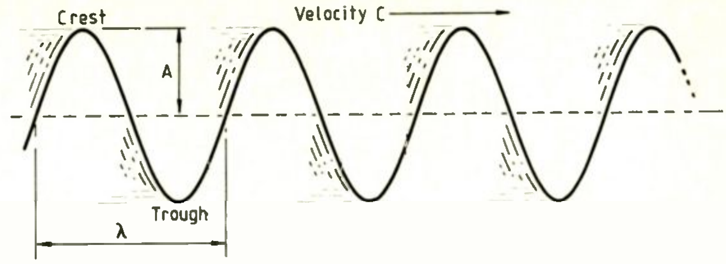


Fig.4. A periodic wave, in this example a harmonic function, shows other properties we associate with waves. The "crest" is obvious, the wavelength λ is well known and so is the meaning of the amplitude A.

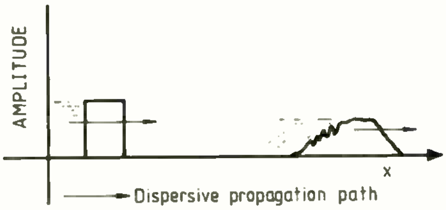


Fig.5. One of the obvious and troublesome effects of dispersion is that the harmonic components of a complex wave travel at different velocities and arrive at different times.

EULER AGAIN

As soon as we deal with harmonic waves, Euler's identity soon comes into the picture again because of the ease with which complex quantities describe harmonic variations.

$$e^{j\theta} = \cos\theta + j\sin\theta$$

Therefore, wave equations have a complex exponential solution of the form,

$$yAe^{j(\beta x-\omega t)}$$

You could show that this certainly is a solution of d'Alembert's non-dispersive wave equation by differentiating it with respect to t and then with respect to x a couple of times and inserting into the differential equation. In general, what all this says is that such a solution exists for any single-frequency disturbance starting in a medium. Once started, a harmonic disturbance continues along as such, although it might fade away eventually, i.e. damping might be present.

It looks as though y is a solution of any equation where the differential operators $\partial/\partial t$ and $\partial/\partial x$ appear only as 'powers' in a polynomial function g operating to the right on y. In other words,

$$g\left(\frac{\partial}{\partial t},\frac{\partial}{\partial x}\right)y=0$$

forms a whole family of wave equations. For example, if

$$g=\left(\frac{\partial}{\partial x}\right)^2-\frac{1}{c^2}\left(\frac{\partial}{\partial t}\right)^2, \text{ then } \frac{\partial^2 y}{\partial x^2}-\frac{1}{c^2}\frac{\partial^2 y}{\partial t^2}=0$$

which is our familiar d'Alembert equation.

DISPERSAL TIME

Now we can handle dispersion, which simply means that the velocity of the wave varies with its frequency (or alternatively you could say, with its phase constant). You would quite rightly see this as a possibly awkward problem. As all communications signals cover a band of frequencies, those at the low end will turn up at a different time compared to those at the high end. For example, the low-frequency components of someone's telephone conversation might turn up at a different time compared to the high-frequency parts. Data pulses might start off erect and squared off but could arrive looking a very sorry state, having been smeared out by the dispersive effects, as in Fig.5.

Oliver Heaviside investigated dispersion on lossy telegraph cables and suggested ways of equalizing the delays of all frequency components, thus overcoming the distortion problem. As usual, no one took much notice of him at the time, but shortly afterwards he was vindicated by actual experiments that Professor Pupin carried out in the USA, showing that 'loading coils' would indeed reduce voice distortion on long telephone lines.

For example, if you had this rather fearful looking wave equation turning up in some work,

$$\frac{\partial^3 y}{\partial x^3}+3\frac{\partial^3 y}{\partial x^2\partial t}=2\frac{\partial y}{\partial t}$$

then at least the harmonic solution would work. This is because of the particularly simple way exponentials differentiate,

$$y=Ae^{j(\beta x-\omega t)}$$

$$\frac{\partial y}{\partial t}=-j\omega y,\ \frac{\partial^2 y}{\partial t^2}=\omega^2 y, \text{ etc. up to } \frac{\partial^n y}{\partial t^n}=(-j\omega)^n y$$

and

$$\frac{\partial y}{\partial x}=j\beta y,\ \frac{\partial^2 y}{\partial x^2}=-\beta^2 y, \text{ etc. up to } \frac{\partial^n y}{\partial x^n}=(j\beta)^n y$$

This enables us to rewrite the above equation,

$$-j\beta^3 y+3j\beta^2\omega y=-2j\omega y$$

Cancelling through the y and the j results in,

$$\frac{\omega}{\beta}=\frac{\beta^2}{(2+3\beta^2)}=\text{the wave velocity } v.$$

This shows large dispersion, with the velocity being very small for small phase constants or what amounts to saying the same thing, for long wavelengths. The velocity becomes constant as β grows larger as Fig.6 shows.

PHASE AND GROUP VELOCITY

The problem with dispersive media, in that the various components of a complex signal waveform turn up at different times, means that a point of constant phase in the wave moves faster than expected, while the energy flow becomes slower. The *phase velocity* agrees with what I have been calling the wave velocity, which we will now call v_p, and the rate of energy flow gives rise to another velocity, v_g, called the *group velocity*.

You can see from $y = A\sin(\beta x - \omega t)$ that $\beta x - \omega t$ (the phase) must remain constant on a phase front. Differentiate this with respect to time,

$$\beta\frac{dx}{dt} - \omega = 0,$$

$$\therefore \quad \text{phase velocity } v_\phi = \frac{\omega}{\beta}\,\text{ms}^{-1}$$

as before. If v_ϕ is a function of frequency, as it will be in a dispersive medium, then if you consider all the fairly close frequencies in a complex wave train, and further, that the wave group 'peaks' at the point where all the components come into phase, we have at this point $(\beta x - \omega t) = \Phi$.

The rate of change of Φ with *frequency* (not time, as before) is zero — we are 'at the peak', then,

$$\frac{d\Phi}{d\omega} = t - x\frac{d\beta}{d\omega} = 0$$

Therefore the velocity of the peak of the group as it moves forward is,

$$v_g = \frac{x}{t} = \frac{d\omega}{d\beta}\,\text{ms}^{-1}$$

You will find a good example exists in our subject illustrating these effects. Electromagnetic waves in a waveguide travel in a dispersive system. Well known results in waveguide theory show that various modes of propagation exist in both transverse electric (TE) and transverse magnetic (TM) wave types[7]. Each propagation mode 'cuts off' at certain critical frequencies, ω_c, at the low end of the band and energy ceases to flow along the pipe. The phase velocity turns out to vary strongly with frequency according to,

$$v_\phi = \frac{c}{\sqrt{1 - \frac{\omega_c^2}{\omega^2}}}$$

for a particular mode in an unloaded (air filled) guide.

This shows, at first sight, the peculiar result that v_ϕ appears to be much greater than the speed of light c, when ω_c is near the cut off frequency ω_c. Indeed the phase travels down the guide at a tremendous rate near cut off. The guide wavelength becomes correspondingly very long. Yet hardly any energy travels down the pipe. We see this by looking at the group velocity.

$$v_\phi = \frac{c}{\sqrt{1 - \frac{\omega_c^2}{\omega^2}}}$$

By squaring this and re-arranging, we get,

$$\beta^2 = \frac{\omega^2}{c^2} - \frac{\omega_c^2}{c^2}$$

which you can now differentiate with respect to ω, thus obtaining,

$$\frac{d\beta}{d\omega} = \frac{1}{v_g} = \frac{v_\phi}{c^2}, \text{ or } v_g v_\phi = c^2$$

This shows that as v_ϕ gets greater and greater as cut-off approaches, v_g becomes smaller and smaller, as shown in Fig.7. At cut-off, the propagation of energy (i.e. the signal) ceases because the wave just bounces up and

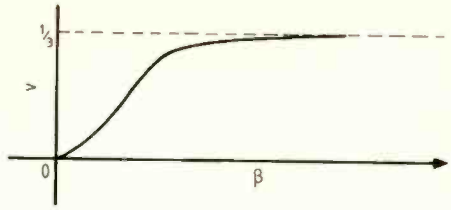


Fig.6. The example here shows a rapid change of velocity as a function of phase constant. Signals travelling in this medium would arrive very distorted.

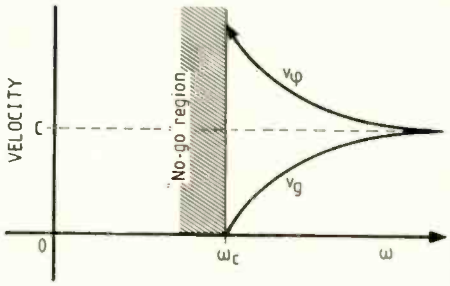


Fig.7. The rapid rise in the phase velocity of EM waves travelling in a waveguide as the frequency falls, is matched by a large decrease in the group velocity. Beyond cut-off no energy propagates at all.

down across the guide, with no propagation along it.

Of course, all the problems of signal dispersion turn up in optical-fibre design and equalizers need careful design to maintain the integrity of the data streams. In earlier line transmission systems, the parameters varied with the weather, and line dispersion changed all the time. Telecoms engineers had to expend much design ingenuity in developing *adaptive equalizers* to compensate this headache.

Thus, wavemotion turns out to be no trivial matter in understanding communications systems. Particular applications change all the time, but the basics hardly change at all — and these I have discussed here. Yet how interesting it all turns out to be again. There is no doubt about it, the fundamental natural phenomena behind the engineering become most worthwhile to study.

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Multiprocessor systems

Multiple-instruction/multiple-data-stream architectures are discussed in this second article.

ALAN CLEMENTS

As discussed last month, multiple-instruction/multiple-data-stream architecture, m.i.m.d., is the most general purpose of the four main multiprocessor-system structures.

While the array or pipeline processor is likely to be constructed from very special units, the more general m.i.m.d. architecture is much more likely to be built from widely available devices like the 68000. Therefore, the major design consideration in the production of such a multiprocessor concerns the topology of the system, which is a measure of the way in which the communications paths between the individual processors are arranged.

Figures 3 to 7 depict the five classic m.i.m.d. topologies, which are (apart from the hypercube) incidentally, the same as those available to the designer of local area networks. Multiprocessor structures are described both by their topology and by their interconnection level. The level of interconnection is a measure of the number of switching units through which a message must pass when going from processor X to processor Y. The four basic topologies are the unconstrained topology, the bus, the ring, and the star, although, of course, there are many variants of each of these pure topologies. The principal features of each of these arrangements are defined as follows.

The unconstrained topology is so called because it is an ad-hoc arrangement in which a processor is linked directly to each other processor with which it wishes to communicate, Fig.3. It takes very little thought to appreciate that the unconstrained topology is not really practicable for any but the simplest of systems. Clearly, as the number of processors grows, the number of interconnections, i.e. buses, becomes prohibitive. The advantage of such a system is the very high degree of coupling that can be achieved. As the buses are dedicated to communications between only two processors, there is no conflict between processors waiting to access the same bus.

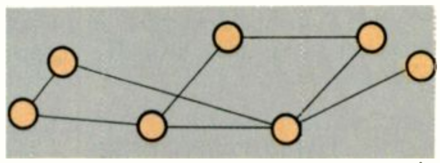


Fig.3. Unconstrained multiprocessor topology is not really practicable for any but the simplest of systems. Clearly, as the number of processors grows, the number of interconnections – buses – becomes prohibitive.

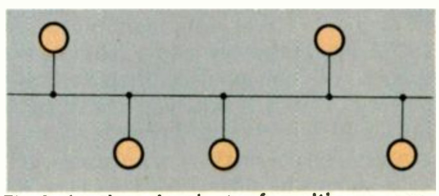
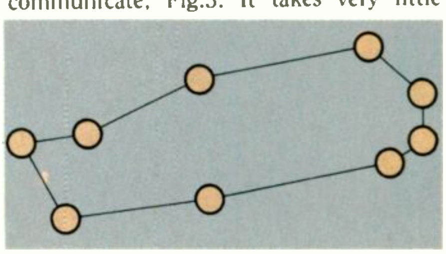


Fig.4. In the simplest of multiprocessor topologies, each processor is connected to a single common bus. Bus systems are simple in that they avoid the problem of how to route a message from processor X to processor Y.



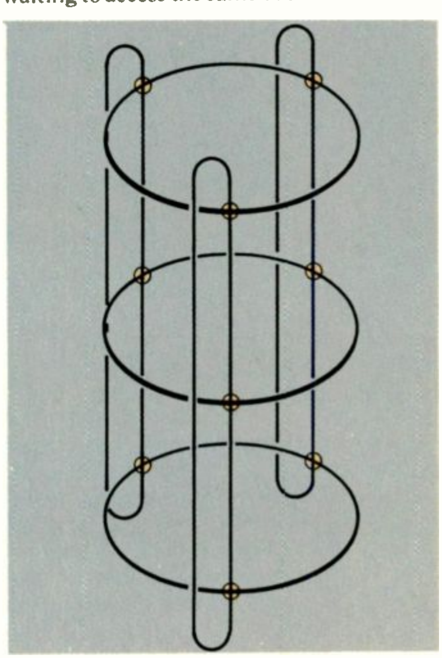


Fig.5. In ring topology, each processor is connected only to its two nearest neighbours. Information flows round the ring in one direction only, and a packet of information passes through each of the processors in the ring. Ring topology has advantages in some classes of loosely-coupled multiprocessor network and represents one of the most popular forms of local area network. Fig.5(b) shows a two-dimensional ring with nine processing elements and three horizontal plus three vertical rings.

THE BUS

The bus is the simplest of topologies because each processor is connected to a single common bus, Fig.4. The bus is a simple topology not least because it avoids the problem of how to route a message from processor X to processor Y. In fact, the interprocessor bus is little more than an extension of the bus (back-plane) found in any conventional computer. Indeed, it is entirely possible to use the existing bus of many microcomputers to create a multiprocessor system. All that is required is a mechanism enabling control to be passed from one microprocessor to another.

The disadvantage of the bus as a method of implementing a multiprocessor system lies in the problem of controlling access to the bus. As only one processor at a time can use the bus, it is necessary to design an arbiter to determine which processor may access the bus at any time. Arbitration between two or more contending processors slows down the system and leads to bottlenecks. A bus offers a relatively high degree of coupling but is more suitable for schemes in which the quantity of data exchanged between processor is small. Because of the economic advantages of the bus, we concentrate on the MIMD bus architecture. In my final article I will consider the design of a suitable m.i.m.d. architecture for the 68000 microprocessor.

THE RING

In a ring, each processor connects only to its two nearest neighbours, Fig.5. One neighbour is called the upstream neighbour and the other the downstream neighbour. A node receives information from its downstream neighbour and passes it on to its upstream neighbour. In this way, information flows round the ring in one direction only, and a packet of information passes through each of the processors in the ring. Information passed to a node contains a destination address. When a node receives a packet, it checks the address and, if the packet address corresponds to the node's own address, the node reads the packet. Similarly, a node is able to add packets of its own to the stream of information flowing round the ring.

Ring topology offers certain advantages for some classes of loosely-coupled multiprocessor network and represents one of the most popular forms of local area network. It is less widely used as a method of interconnecting processors in a tightly-coupled m.i.m.d. architecture. It is, in fact, possible to construct two-dimensional ring networks. Figure 5(b) demonstrates a two-dimensional ring with nine processing elements and three horizontal plus three vertical rings. The two-dimensional ring offers multiple paths between nodes and is not catastrophically affected by the failure of a single node.

STAR

Star topology, Fig.6, employs a central processor as a switching network, rather like a telephone exchange, between the other processors which are arranged logically (if not physically) around the central node. The advantage of the star is that it reduces bus contention, as there are no shared communication paths, and it does not require the large number of buses needed by unconstrained topologies.

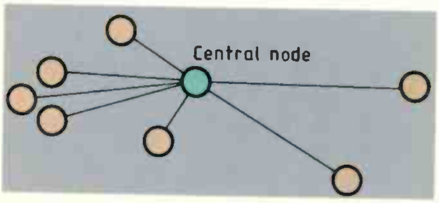


Fig.6. Star multiprocessor topology reduces bus contention, eliminates shared communication paths and needs only a small number of buses – but it is only as good as its central node. Its central processor acts as a switching network, switching between the other processors arranged logically around the central node.

On the other hand, the star network is only as good as its central node. If this node fails, the entire system fails. Consequently, the star topology does not display any form of graceful degradation. Moreover, the central network must be faster than the nodes using its switching facilities, if the system is to be efficient. In many ways, the star topology is a configuration better suited to local area networks, where the individual signal paths are implemented by serial data channels, rather than by the parallel buses of the tightly-coupled multiprocessor.

HYPERCUBE

An n-dimensional hypercube multiprocess connects together $N=2^n$ processors in the form of an n-dimensional binary cube. Each corner (vertex or node) of the hypercube consists of a suitable processing element and its associated memory. Because of the topology of a hypercube, each node is directly connected to exactly n other neighbours. Figure 7 illustrates the hypercube topology for n=1, 2, 3 and 4.

Each processor in a hypercube has an n-bit address in the range $0...00_2$ to $1...11_2$ (i.e. 0 to 2^{n-1}) and each of the n nearest neighbours of a particular node has an address that differs from the node's address by only one bit. For example, if n=4 and a node has binary address (0100), its four nearest neighbours have addresses (1100), (0000), (0110) and (0101).

A hypercube of dimension n is constructed recursively by taking a hypercube of dimension n-1 and prefixing all its node addresses by 0 and adding to this another hypercube of dimension n-1 whose node addresses are all prefixed by 1. In other words, a hypercube of dimension n can be subdivided into two hypercubes of dimension n-1, and these two subcubes can, in turn, be divided into four subcubes of dimension n-2 and so on.

The hypercube is of interest because it has a topology that is particularly suited to certain groups of algorithm. In particular, it is well-suited to problems involving the evaluation of fast Fourier transforms. The first practical hypercube multiprocessor was built at Caltech in 1983. This was called the Cosmic Cube and was based on 64 8086 microprocessors plus 8087 floating-point coprocessors.

HYBRID TOPOLOGIES

In addition to the above pure network topologies, there are very many hybrid topologies, some of which are described in Figs 8 to 11.

Figures 8(a,b) both illustrate the dual-bus multiprocessor, although this topology may be extended to include any number of buses. In Fig. 8(a) the processors are split into two groups, with one group connected to bus A and one to bus B. A switching unit connects bus A to bus B and therefore allows a processor on one bus to communicate with a processor on the other. The advantage of the dual-bus topology is that the probability of bus contention is reduced, because both buses can be operated in parallel (i.e. simultaneously). Only when a processor connected to one bus needs to transfer data to a processor on the other does the topology become equal to a single-bus topology.

The arrangement of Fig. 8(b) also employs two buses, but here each processor is connected directly to both buses via suitable

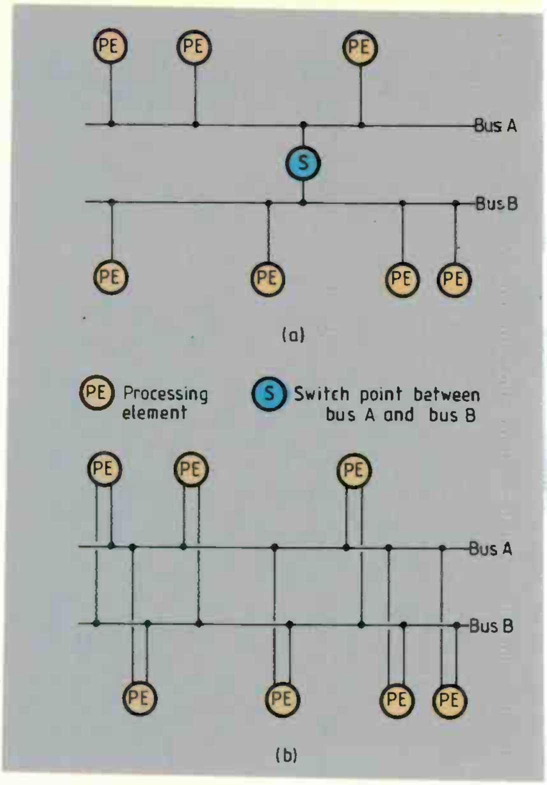


Fig.8. These two examples of multi-bus multiprocessor structures have two buses, but the method can be extended to include any number of buses. Because both buses can be operated in parallel, the probability of bus contention is reduced. In (a) the processors are split into two groups. A switching unit connects bus A to bus B and therefore allows a processor on one bus to communicate with a processor on the other. In (b), each processor connects directly to both buses via suitable switches and two communication paths always exist between any pair of processors.

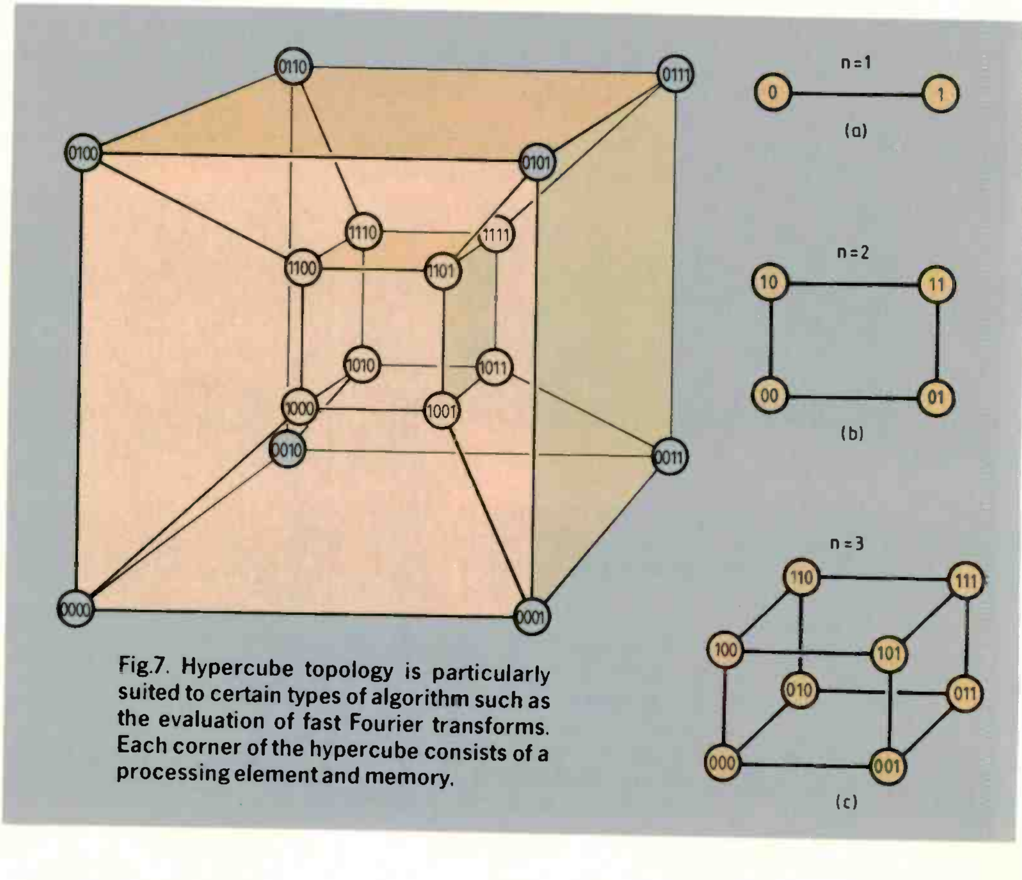


Fig.7. Hypercube topology is particularly suited to certain types of algorithm such as the evaluation of fast Fourier transforms. Each corner of the hypercube consists of a processing element and memory.

switches. Two communication paths always exist between any pair of processors; one using Bus A and one using Bus B. Although the provision of two buses reduces the bottleneck associated with a single bus, it requires more connections between the processors and the two buses, and more complex hardware is needed to determine which bus a processor is to use at any time.

Another possible topology described in Fig.9 is the so-called crossbar switching architecture, which has its origin in the telephone exchange where it is employed to link subscribers to each other.

The processors are arranged as a single column (processors P_{c1} to P_{cm}) and a single row (processors P_{r1} to P_{rn}). That is, there are a total of m + n processors. Each processor in a column is connected to a horizontal bus and each processor in a row is connected to a vertical bus. A switching network, $S_{r,c}$, connects the processor on row r to the processor on column c. Note that there are m×n switching networks for the m+n processors.

The advantage of the crossbar matrix is the speed at which the interconnection between two processors can be set up. Furthermore, it can be made highly reliable by providing alternative connections between nodes, should one of the switch points fail. Reliability is guaranteed only if the switches are fail safe and always fail in the off or no-connection position.

If the switches at the crosspoints are made multi-way (vertical-to-vertical, horizontal-to-horizontal or horizontal-to-vertical), you can construct a number of simultaneous pathways through the matrix. The provision of multiple pathways considerably increases the bandwidth of the system.

In practice, the crossbar matrix is not widely found in general-purpose systems, because of its high complexity. Another penalty associated with this arrangement is its limited expandability. If you wish to increase the power of the system by adding an extra processor, you must also add another bus, together with its associated switching units.

An interesting form of multiprocessor topology is illustrated in Fig.10. For obvious reasons this structure is called a binary tree.

Any two processors (nodes) in the tree communicate with each other by traversing the tree right-to-left until a processor common to both nodes is found, and then traversing the tree left-to-right. For example, Fig.10 shows how processor P_{0110} communicates with processor P_{0100}, by establishing backward links from P_{0110} to P_{01} and then forward links from P_{01} to P_{010} to P_{0100}.

The topology of the binary tree has the facility to set up multiple simultaneous links (depending on the nature of each of the links), as the whole tree is never needed to link any two points. In practice, a real system would implement additional pathways to relieve potential bottlenecks and to guard against the effects of failure at certain switching points. Note that the failure of a switch in a righthand column causes the loss of a single processor, while the failure of a link at the lefthand side immediately removes half the available processors from the system.

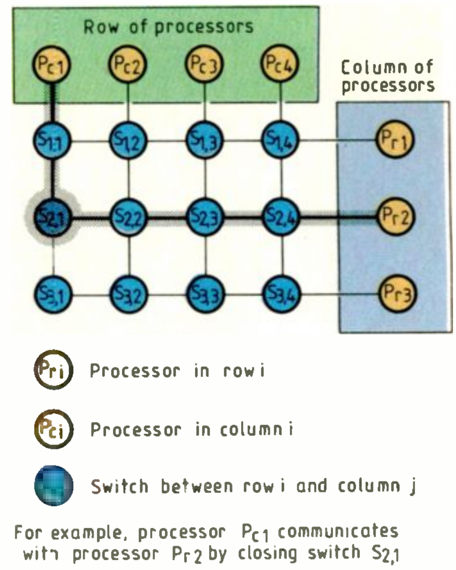


Fig.9. Crossbar switching is a multi-bus multiprocessor architecture similar to the structure used in telephone exchanges. Processor interconnections can be set up very quickly and the structure can be very reliable but in practice expansion is difficult and the structure is complex.

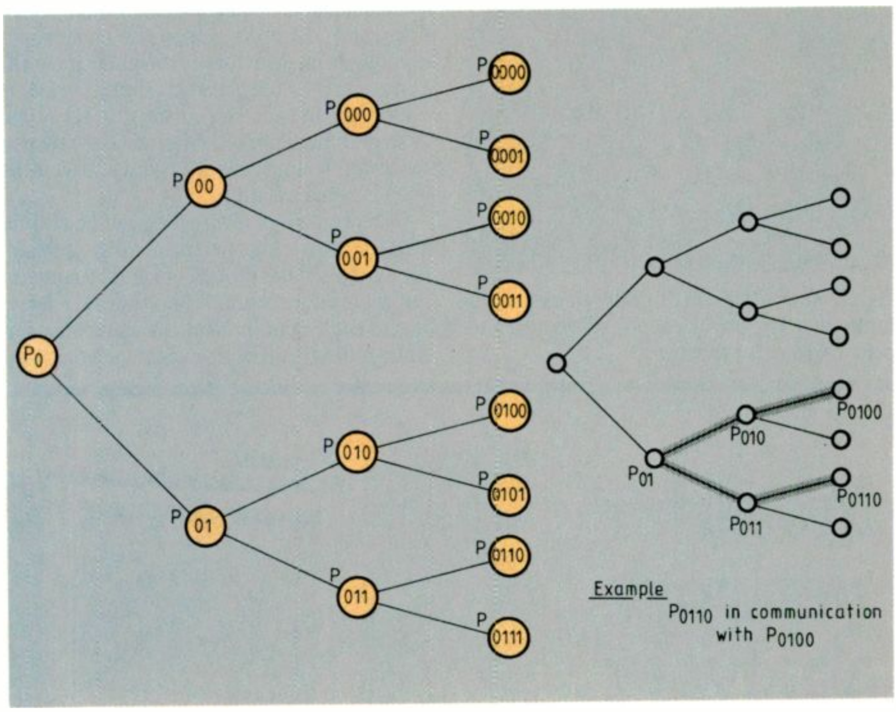


Fig.10. In binary-tree structured multiprocessor system, any two processors communicate with each other by traversing the tree right-to-left until a processor common to both nodes is found, and then traversing the tree left to right. Potentially, multiple simultaneous links can be set up since the whole tree is never needed to link any two points. A disadvantage is that failure of a switch in a righthand column causes the loss of a single processor, while the failure of a link at the lefthand side immediately removes half the available processors from the system.

In some ways, the structure of Fig.10 can be found to exhibit interesting properties. However, due to its complexities, it is likely to remain an intriguing topology and almost never to raise its head above the pages of somebody's PhD thesis.

Finally, and more down-to-earth, Fig.11 illustrates the cluster topology which is a hybrid star-bus structure. The importance of this structure lies in its application in highly-reliable systems. Groups of processors and their local memory modules are arranged in the form of a cluster. Figure 11 shows three processors per cluster in an arrangement called triple modular redundancy. The output of each of the three processors is compared with the output of the other two processors in a voting network. The output of the voting circuit (or majority logic circuit) is taken as two-out-of-three of its inputs, on the basis that the failure of a single module is more likely than the simultaneous failure of two modules.

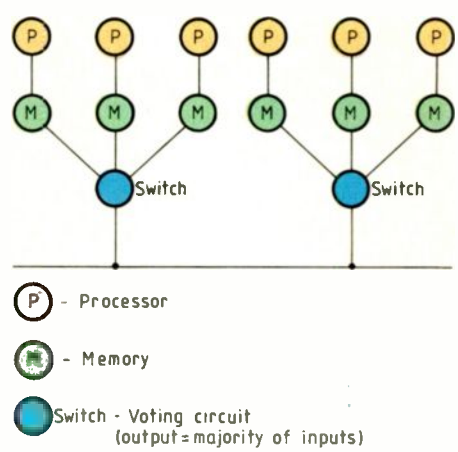


Fig.11. Cluster topology is a hybrid star-bus structure for use in highly-reliable systems. In this three-processor example, each output is compared with the output of the other two processors in a voting network.

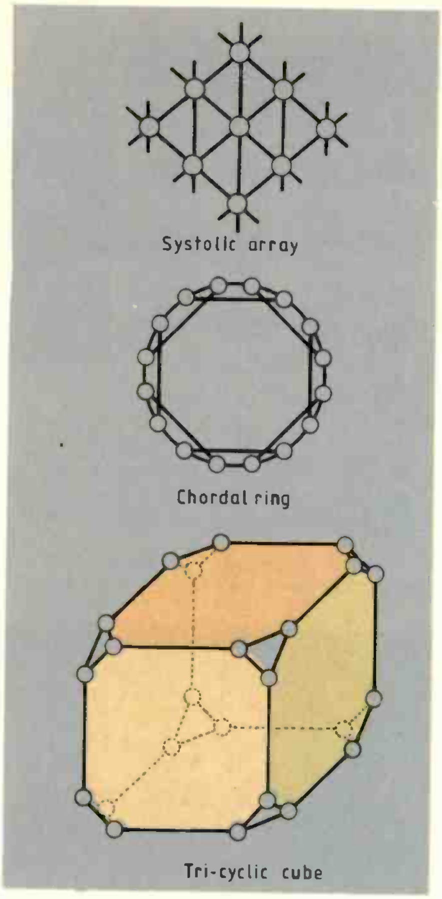


Fig.12. Topologies for multiprocessor systems are legion; these three topologies provide food for thought.

Although the clusters in Fig.11 communicate with each other via a bus, it is possible to use any other suitable mechanism to link them.

The design of a clustered triple modular redundancy system is not as easy as might be first thought. One of the major problems associated with modular redundancy arises from a phenomenon called divergence. Suppose that three identical processors have identical hardware and software and that they receive identical inputs and start with the same initial conditions at the same time. Therefore, unless one processor fails, their outputs are identical, as all elements of the system are identical.

In actual fact, the above statement is not true! In order to create truly redundant systems, each of the three processors in a cluster must have its own independent clock and i/o channels. Therefore, events taking place externally will not be 'seen' by each processor at exactly the same time. If these events lead to conditional branches, the operation of a processor in the cluster may diverge from that of its neighbours quite considerably after even a short period of operation. In such circumstances, it becomes very difficult to tell whether the processors are suffering from divergence or whether one of them has failed.

The problem of divergence can be eliminated by providing synchronizing mechanisms between the processors and by comparing their outputs only when they all wish to access the system bus for the same purpose. Once more it can be seen that, although the principles behind the design of multiprocessor systems are relatively straightforward, their detailed practical design is very complex due to a considerable degree of interaction between hardware and software. As I have already pointed out, topologies for multiprocessor systems are legion. Figure 12 provides examples of three further topologies – just to provide food for thought.

My next article looks at how processors communicate with each other and describes some of the techniques that can be used to link together two or more 68000s to create a multiprocessor. In particular, I will look at the way in which 68000s are able to communicate with each other via shared memory. In the final article I introduce the VMEbus and discuss how it is used to implement practical 68000-based multiprocessor systems.

Alan Clements, BSc., PhD., is a reader in the School of Information Engineering at Teesside Polytechnic.

Modular network analyser

Tektronix has launched a modular PC-based telecommunications network analyser. Designated the TC-2000, its functions can include protocol analysis, PCM/BERT (pulse code modulation/bit error rate testing), lan testing, trunk testing, remote communications, signalling and TIMS (transmission impairment measurement system) with voice band spectrum analysis. By using an 8088 for housekeeping and keyboard driving in conjunction with a series of individual tester modules using more powerful multiple processors, the unit is able to meet a wide range of requirements.

Its modularity allows a customer to purchase modules as and when needed and thus permits better control of costs and equipment usage. In addition, it allows the same basic test equipment to be used all over the world, specific modules being available to meet the testing requirements of different markets.

For example, the technique for the measurement of the audio frequency response differs between the USA and Europe. In the USA, the Bell standard test requires the swept frequency test signal to be sent down the pair under test at the same time as being compared with a reference signal being transmitted via another pair. On the other hand, a technique developed by Wandel & Goltermann has been adopted in Europe. Here, using just the one pair, the signal on the line is switched alternately between test and reference frequencies. This obviates the need to take a second pair out of service.

Options available include X.25, p.c.m. (both Bell and CEPT versions) and well as a number of local area network protocols including Ethernet 1.0 and 2.0 as well as IEEE803.2. The unit has seven slots; and since most analysers occupy only one slot, a single instrument can generally accommodate all analysers required.

A.J.M.

NEW 8051 DEVELOPMENT CARD

The new Cavendish Automation development card carries a full symbolic Assembler and text editor as well as the MCS-BASIC 52 package. It will allow the user to write applications programmes in either BASIC or Assembler.

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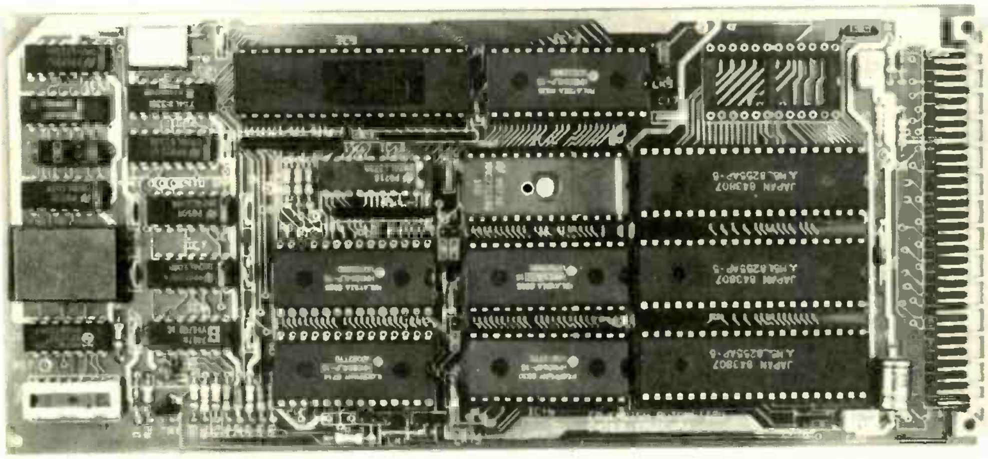

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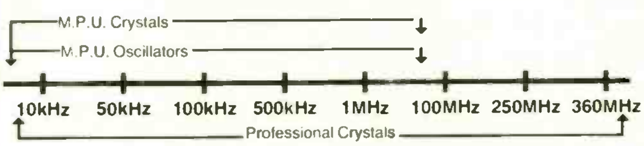

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Multi-standard h.d.tv camera

This BTS design, suitable for studio or outside broadcasts, makes extensive use of automation.

WOLFRAM KLEMMER*

In developing this camera, the aim was to secure the maximum image quality possible today. For this reason, the camera includes mixed-field high resolution camera tubes, a dynamic knee processor, and a digital two-dimensional aperture and contour correction unit. All critical analogue signal processors are stabilized by reference pulses. Registration and shading alignment is done fully automatically with the help of a test projector in the zoom lens, and lateral chromatic errors are compensated according to the lens setting.

Since the high-definition television production standard is still undefined, this development is based on a flexible concept allowing adaptation to various scanning formats. Some examples of possible scanning modes are 1050/59.94/2:1, 1125/60/2:1 and 1250/50/2:1.

The KCH 1000 is equipped with all the usual studio audio and intercom features. For its computer-aided operation it uses a serial data bus, which can be extended for the operation of several cameras from one master control panel.

DESIGN BACKGROUND

Between 1982 and 1984, an experimental h.d.tv camera[1] was developed in the advanced development department of BTS (previously the Television Systems Division of Robert Bosch GmbH). Prototypes of this camera were supplied to various German institutes and used there as high-resolution picture sources.

With this earlier project too, there was no fixed scanning mode: the cameras were delivered with various line and field frequencies, and aspect ratios, or equipped with conversion sets to enable changing of the scanning format.

This first camera was developed for operation in a laboratory setting, and can thus be considered as a concept study. But the new BTS h.d.tv camera is intended for the future h.d.tv studio – which will nevertheless at first have an experimental character – and for h.d.tv o.b. vans. Important additions and even basic modifications were therefore needed to meet operational requirements.

At the same time, picture quality has been further improved by use of highest-quality camera tubes and digital signal processing.

The flexible scanning mode has been retained, since such a concept continues to be absolutely essential. Hence, the camera can be adapted to different scanning standards within the physical limits set by signal-to-noise ratio, flyback time, and the resolution of the camera tubes.

*BTS Broadcast Television Systems GmbH, Darmstadt, Federal Republic of Germany.

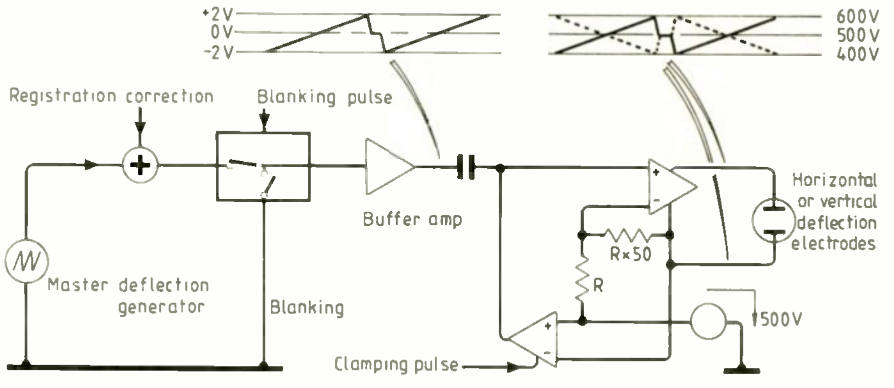


Fig.1. Deflection circuit. To prevent random modulation of line spacing, a signal-to-noise ratio of 100dB must be achieved.

CAMERA TUBES

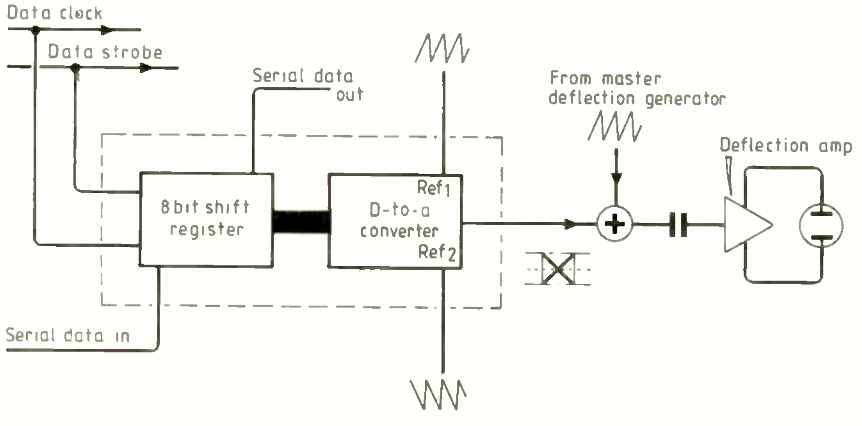
The choice of camera tubes determines the picture quality of a television camera, especially a camera for h.d.tv use. In fact, all aspects of performance are limited by the quality of the pick-up sensor. Two-dimensional resolution, dynamic resolution and lag, signal-to-noise ratio, sensitivity, stability of registration and highlight characteristics are some of these parameters; moreover, they interact and thus cannot be considered separately.

Signal-to-noise ratio is particularly important in this regard, since it is directly dependent on the layer sensitivity at a given lighting level, and is essentially determined by the target capacitance. But any attempt to compensate for insufficient static or dynamic resolution would worsen the signal-to-noise ratio.

Compensation of dynamic resolution, in particular for low-light lag effect through temporal high-pass filtering, is not practical because of its high technical complexity. The lag characteristics of tubes must therefore be given particular consideration, especially as the discrepancy between static and dynamic resolution in previous h.d.tv systems has been greater than in normal television.

BTS has been examining several tubes from European and Japanese manufacturers for possible use in the new h.d.tv camera. These tubes have various types of photoconductive layer (Saticon II, Saticon III, Plumbicon) and different electron guns (triode system, diode-mode system), but all have

Fig.2. Principle of coarse registration adjustment and serial distribution of correction data.



electrostatic deflection with magnetic focusing.

This type of beam deflection offers a number of advantages over the conventional method (magnetic/magnetic):

- uniform resolution
- low geometric distortion, and thereby
- low shading
- shorter flyback, since the stored field energy is lower than with magnetic deflection
- no eddy-current losses, which in case of magnetic deflection have a negative effect on flyback and line-start linearity
- high stability of registration, since there is no thermal drift of the deflection electrodes.

Generating the necessary deflection voltages (200—300Vpk-pk), however, is far from simple. The horizontal and vertical deflection amplifiers must likewise have a bandwidth of over 500kHz to be capable of transmitting the digitally-generated fine-correction signal.

At the same time – at the bandwidth specified – a signal-to-noise ratio of 100dB must be achieved in order to prevent random modulation of the line spacing. This applies in particular to the vertical amplifier.

Deflection voltages must continue to rise to the deflection bias voltage (grid 3) of about 400 — 600V. Naturally, it is necessary to superimpose fixed voltage components on this high voltage level for position shifting. In the BTS h.d.tv camera this problem is solved through capacitively coupling the high-voltage to the low-voltage level, with subsequent clamping in the high-voltage range. This requires a reference potential to be keyed into the deflection signal on the low-voltage level beforehand (Fig.1).

AUTOMATION

Automatic functions in the new camera can be broken down into three groups: basic adjustment, pre-operating adjustment and continuously automated functions.

- Automatic basic adjustment:
  - geometry, fine
  - image registration, fine
  - white shading
  - black shading

- Pre-operating adjustment:
  - white balance
  - automatic filter-wheel selection
  - cable-length equalization

- Continuous automatic functions:
  - automatic iris
  - centring of red and blue channels
  - dynamic lens-error correction

With the exception of cable-length equalization, all automatic functions can be called up from the master control panel, either individually or in groups by menu control.

Automatic basic adjustment, i.e. geometry, registration, and shading adjustment, is done with the help of a special test slide which can be inserted into the optical path in the zoom lens. When the camera is first

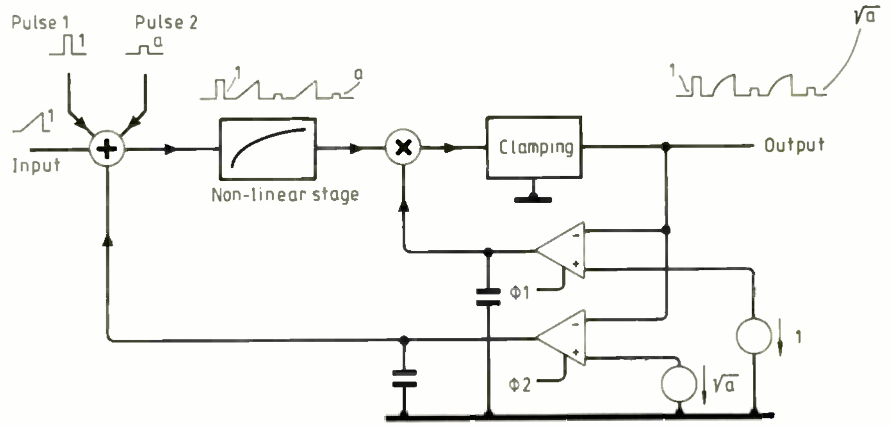


Fig.3. Video stages are stabilized and controlled throughout by the use of reference pulses.

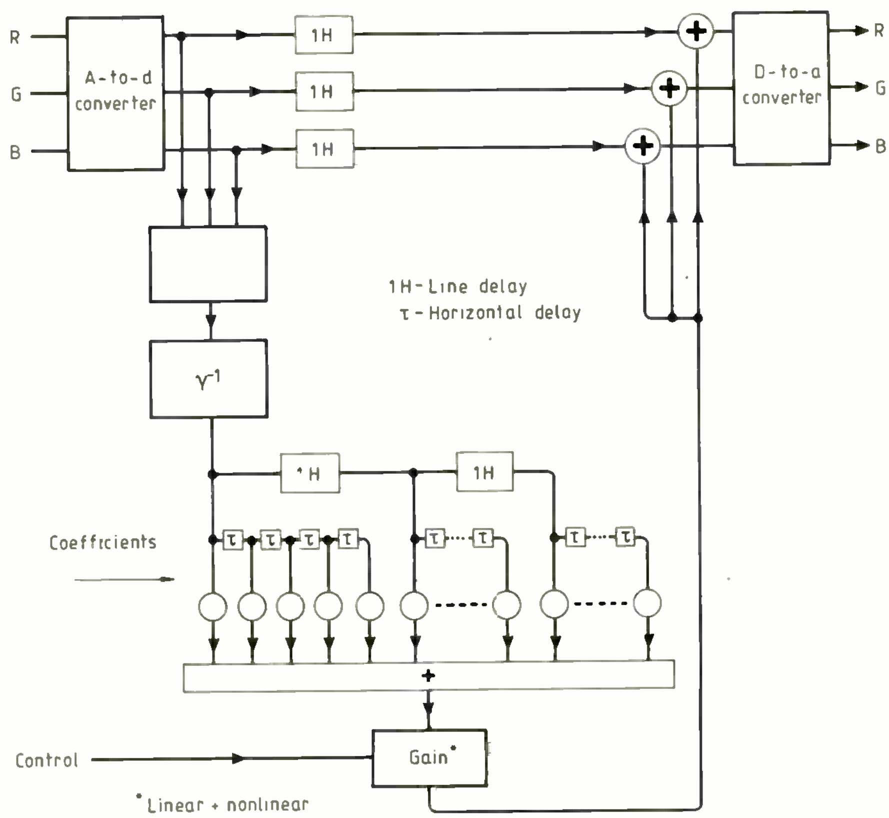


Fig.4. Block diagram of the two-dimensional processor for aperture and contour correction.

used, or after a tube change, this fine adjustment should be preceded by a coarse adjustment of registration. Coarse adjustment is manual, but menu-controlled from the master control panel with digital potentiometers. Here the highest stability is necessary, and so the signal is transmitted digitally to the analogue operating point.

This principle is illustrated in Fig.2, using manual adjustment of size for horizontal and vertical deflection as an example. The hybrid d-to-a converter is used here as a digitally-controlled potentiometer. At the same time, the example gives an idea of the general data transmission concept of the camera: data transmission through the camera cable as well as data distribution in the camera head and camera control unit (c.c.u.) take place serially in the vertical blanking interval.

The test pattern projector in the lens simulates errors of the main lens for a fixed focal length and focusing distance.

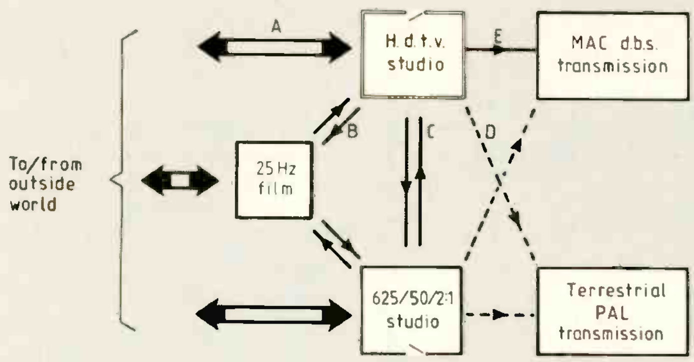
The remaining chromatic errors and geometric errors of the lens are eliminated to a great extent by the dynamic lens error correction. Depending on the set values for focal length and distance, the following variables are affected:

- h, v position and size in red channel
- h, v position and size in blue channel
- pincushion correction for geometry

Correction data for different lens types can be stored.

Each time the camera is switched on the cable-length compensation system checks the cable length using a burst signal generated in the camera head. The resolution of measurement and compensation is 10m for

Continued on page 710.



The h.d.tv studio

M.D. WINDRAM and G.J.TONGE*

In the context of UK broadcasting any high-definition studio or source must feed existing and new networks. A d.b.s. studio complex of the future may be configured as illustrated above in simple diagrammatic form. Five classes of interface with the h.d.tv studio standard are identified.

A. Interface with a (possibly different) h.d.tv standard used in other parts of the world.

B. Interface with motion picture film with a frame rate of 24Hz (or 25Hz as used in conjunction with current UK television). As a source, film is likely to form a significant proportion of h.d.tv broadcasts, especially in the early stages of an h.d.tv service. In addition electronic h.d.tv production is likely to be used increasingly in the making of feature films.

C. Interface with the 625/50/2:1 studio. As an input to the h.d.tv studio, archive material stored on the 625 line standard will be required from time to time to be included in h.d.tv productions. In the other direction, many h.d.tv productions will also be down-converted to a 625 line studio format, to be included in 625 line productions.

D. There may be a need (e.g. during prestige sporting events) to broadcast the same programme both in h.d.tv and over the terrestrial PAL network.

E. It is seen as essential that h.d.tv broadcasts are compatible with the 625/50/2:1 MAC transmission system (Eureka project EU95).

Some comments on the benefits and disadvantages connected with these interfaces are collected together in the Table (right). In considering these comments it soon becomes clear that there is only one interface (A) where the UK broadcaster may gain some advantage by adopting a 60Hz h.d.tv system, assuming that other parts of the world adopt 60Hz. For all the other interfaces a 50Hz h.d.tv studio standard provides both a more economic and a substantially higher-quality solution. In particular, a standard of 1250/50/1:1 (with initial equipment operating on 1250/50/2:1) is the ideal for studios related to UK broadcasting.

An obvious question arising from this argument is "What happens with the single world-wide h.d.tv studio standard?" There are two possible answers to this:

i) Adopt a 50Hz standard worldwide. This would create problems of conversion in the 60Hz countries.

ii) Adopt h.d.tv studio standards as higher members of the extensible family of digital standards of CCIR Recommendation 601 (i.e. 1250/50 and 1050/59.94).

If we assume that option (i) will not be followed, then the advantages offered by option (ii) are:

- Easier commonality of equipment between the two h.d.tv standards.
- Simpler compatible broadcasting options for USA and Japan etc.
- Further establishment of the already agreed Recommendation 601 standard.

*Independent Broadcasting Authority

Summary of benefits and deficits of 50 and 60Hz h.d.tv for the UK studio shown in the diagram.

	50Hz h.d.tv	60Hz h.d.tv studio
A	H.d.tv field-rate conversion may be necessary for exchanged material.	H.d.tv field-rate conversion may not be necessary for exchanged material.
B	Simple, high-quality conversion to and from 25Hz film.	More complex lower quality conversion to and from 24Hz film.
C,D	Simple conversion with high quality.	Complex conversion with motion artifacts.
E	Compatible MAC transmission format possible.	H.d.tv field-rate conversion need for all programme material prior to compatible MAC transmission.

Further reading

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Continued from page 709.

a maximum length of 300m multi-core cable.

VIDEO PROCESSING

Video signal processing was redesigned with with the emphasis on extending signal dynamics. A module for automatic beam-current control for diode-gun beam generators is a prerequisite for this. The camera head furthermore incorporates a stage for selecting static gain and a dynamic knee processor.

This knee processor enables compression of signal levels up to 400% to the normal video level range of 100%, with the knee-point (controlled by the picture content) lying between 80% and 100%. To ensure the transmission of fine image structures even in the compressed signal range (low differential slope), a special knee detail signal is mixed in.

Critical stages in the analogue video signal processing are stabilized with the help of controlled or fixed reference pulses in the blanking intervals. This guarantees complete independence from all types of drift and from component tolerances. Furthermore, the characteristic parameters of these stages can be defined through external control voltages.

This concept is illustrated is Fig.3, using the gamma processor as an example. The use of both additive and multiplicative control ensures the equality of the reference pulse for grey and white with the set values, thus generating the desired characteristic curve.

A unit for two-dimensional aperture and contour correction is an important feature of the main camera amplifier. The unit forms an independent processor with RGB interface as well as synchronization and control interface. However, this processor is fully integrated into the system, so that the user is not aware of it except in terms of great improvement in picture quality.

Figure 4 shows the equivalent total structure of the internal fully-digital equipment. The central feature is a 5×3 coefficient matrix for the generation of the two-dimensional detail signal. This matrix has proved extraordinarily flexible, and allows:

- smoothing (or, if desired, peaking) of the total frequency response
- suppression of diagonal frequency ranges for reasons of signal-to-noise ratio
- symmetrization of unequal diagonal resolution of h.d.tv camera tubes [2,3,4].

SYSTEM OUTLINE

The mechanics of the camera-head housing are broadly taken from the BTS studio camera KCM 125[5]. The interface for the zoom lens is absolutely identical, and the camera likewise includes a double filter wheel. Other components and assemblies taken from the KCM 125 are the power supply and the sound and intercom system.

Two high-quality microphone channels and three intercom communication paths are provided: producer intercom, engineering intercom, and programme sound. Headsets for the cameraman, reporter and dolly driver can be connected to the camera head.

A fault-diagnosis system allows remote monitoring of the video signals (test operation), the deflection sawtooth waveforms, the operating voltages and pulses, and the temperature in the prism mounting.

The control interface of the c.c.u. is designed so that multi-camera operation is possible. Data linkage between all participants is through a bidirectional bus with a transmission rate of 10Mbit/s. The main cable also includes the bus intercom and the power supply for the master control panel. Cross-connections provide intercom and power supply services to the remote control panels. Each system participant includes an 80186-based control computer with special data bus interface.

Correction data is stored in the camera head, whereas operation-related data is stored in the control computer of the c.c.u. This means that the same picture is maintained even after the camera heads have been exchanged and the system turned back on. The master operating unit ensures access to all operating and adjustment functions of all cameras, with menu-controlled and software-protected adjustment.

Interfaces for the control oscilloscope and control monitor (monochrome) are likewise bus-capable. The output signals of all cameras can thus be monitored from the master operating unit.

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HIGH-DEFINITION TELEVISION

Prospects and politics

Where is the search for a standard leading? Peter Wilson of Sony Broadcast, the 1125/60 system's leading proponent, presents his view.

Sony's second-generation h.d.tv camera, type HDC 300, seen in film-style mode.

Some time has now passed since the slogan "evolution or revolution" was bandied about. This has naturally died a death as technology seems to need a revolution to achieve evolution. The latest slogan in h.d.tv is "compatibility or convertibility".

Players in this game fall mainly in two camps: those who make a living from hardware, predominantly domestic (and are interested in compatibility); and those who make a living from software/programme production, and are concerned with convertibility – predominantly the broadcasters and production companies.

High-definition television is a system which gives the viewers a much more powerful impression of being there. This is achieved by increasing the size of the picture and giving approximately five times the information seen on conventional tv.

Add to this a change in the aspect ratio from 4:3 to 16:9, an increased contrast ratio, more faithful colorimetry and an absence of coding artefacts, and the sum total results in simply amazing pictures.

Sony feels that this major advance in visual perception should not be subverted. Sensing that programme producers were becoming worried about the future implications of h.d.tv on their business, the company has adopted a "seeing is believing" policy for programme production, especially over the very heated issue of 50Hz versus 60Hz.

The people we have been working with are European broadcasters. We have made productions, both long and short, with companies in many countries.

Of these broadcasters, Radiotelevisione Italiana is the most committed. It has made two test productions and one commercial production, *Julia and Julia* (Italian name, *Linea de Confine*). *Julia and Julia* featured the actors Kathleen Turner and Sting and

EUROPEAN H.D.TV PRODUCTIONS MADE WITH SONY EQUIPMENT	
SFP, France	Test production
RAI, Italy	Two test productions, one commercial feature film
Channel 5, Italy	Test (musical)
SRG, Switzerland	Test (general)
ORF, Austria	Test (opera, *The Magic Flute*)
ZDF/ARD, West Germany	Test (*Wetten das*)
NDR, West Germany	Test (documentary on Hamburg)
BR, West Germany	Test (drama, *Rendezvous*)
BBC, UK	Training production (wildlife); test production
HTV, UK	Technical test
TVE, Spain	Test (documentary on the architect Gaudi)

was technically acclaimed in its 35mm film version.

All these broadcast companies are enthused with h.d.tv and found the technical problems much slighter than they anticipated. An opinion appears to be forming that although broadcasters are apprehensive about down-conversion from 1125/60Hz h.d.tv to PAL for Europe, this is merely a technical problem to be solved with a black box. Far more frightening to them is the loss of revenue caused by a multiplicity of originating formats – that is, by the existence of more than one h.d.tv production standard. Broadcasters in the 1990s will only survive by programme exchange deals and co-productions.

BUSINESS PROSPECTS

Sony has already sold 100 analogue high-definition video tape recorders. These are used in the main as studio mastering and post-production machines in the 1125/60 format. Sony has also sold around 40 high-definition cameras.

The European market area has two commercially operating high-definition studios, in Paris and in West Berlin. Today, 14 Sony recorders and five cameras are in operation in Europe. BTS Stystems in Germany has several orders for its new multi-standard h.d.tv studio camera and is working on v.t.r. and telecine. Quantel in the UK has shipped around 30 high-definition Paintboxes to the graphics industry. These units output standard 1125/60 h.d.tv pictures or magnetic computer tape which is turned into printing plates, giving a very fast turn-around time for high resolution colour graphics. Rank Cintel is developing high-definition telecine machines and Avitel (UK) has delivered many distribution amplifiers.

FUTURE TECHNOLOGY

Both Sony and BTS Systems exhibited at the NAB show in America real second-generation high-definition television cameras. The Sony camera is multi-role (film style or television studio) and the BTS camera is studio style. Sony demonstrated again its experimental digital high-definition v.t.r. with a video bit-rate recorded on tape of 1.188 gigabits per second and with eight digital audio channels. This machine uses metal tape in a reel-to-reel format.

Already in most European countries (certainly in the UK) there exists enough digital telephone network capacity to take wide-band high-definition signals for conferencing or closed circuit links. Both UK national carriers are taking a very healthy interest in h.d.tv, as is the Bundespost in West Germany. The concept of seeing a prize-fight in the local cinema on a big screen is very attractive to many people. It could become a reality very soon.

Most Japanese manufacturers are now working on a half-inch wideband cassette v.t.r. for the industrial market. Sony has introduced a wideband video disc player for museums, exhibitions and general display purposes.

THE POLITICS

Returning to my starting point, what about the question of compatibility or convertibility? There are now three players in the h.d.tv production standard stakes.

1. 1125/60: 1125 lines, 60 fields – known alternatively as Hi-Vision or HDVS (High Definition Video System, a Sony trade name).
2. 1250/50: 1250 lines, 50 fields – this is one form of the Eureka proposal.
3. 1050/59.94: this was recently suggested in USA by North American Philips (MAC 60) and the David Sarnoff Research Centre — AC TV (Advanced Compatible Television).

We can regard system 1 as convertible and systems 2 and 3 as compatible although not with each other. So let us examine these systems more closely.

SYSTEM 1: 1125/60

Parameters of this system were laid down after exhaustive tests by NHK's laboratories in Japan, where it was found that 1125 lines were the best compromise between resolution and perception of resolution. The field rate of 60Hz was found very suitable for removing large-area flicker and the aspect ratio of 5:3 (modified to 16:9 by ATSC and SMPTE) gave a very high level of scene involvement at a viewing distance of 3 to 3.5 times the screen height. This standard is subject to a recommendation by SMPTE and ATSC in the USA.

Advantages: system was laid out after exhaustive evaluations, has shown spectacular performance and has proved itself capable of highest quality production, on a par with 35mm film. System is convertible to all current broadcast and film interchange formats.

Disadvantage: system is not PAL-compatible without complicated down-conversion (NTSC is easy).

SYSTEM 2: 1250/50

This is the Eureka proposal. The system currently under discussion is a revamp of the original HD-MAC proposals. The original aim was to upgrade studio systems gradually to an h.d.tv level. Upgrading was clearly unrealistic, and the plan has been replaced with an h.d.tv proposal down-convertible to PAL/SECAM or MAC-components. Several down-conversion phases are planned, starting with regular PAL and MAC and progressing to EMAC.

Advantages: system is easily down-converted to PAL or MAC, but not to NTSC or MAC 60.

Disadvantages: virtually impossible to convert to 1125/60 or 1050/59.94; suffers from large-area flicker. Broadcasters may dislike the spectre of conversion from 1125/60 to PAL/SECAM/625 MAC. But they will not accept a succession of interim conversions, from PAL to component origination to MAC to 1250/50 origination to EMAC. Nor will the consumer enjoy buying a new decoder box every three years and then a wide-screen display device.

SYSTEM 3: 1050/60

The North American proposal. This system is designed to lessen the worries of the large American broadcasters.

Advantage: converts easily to NTSC.

Disadvantage: does not convert easily to 1125/60, 1250/50 or 625/50. Two primary players in 1050/59.94, NBC and CBS, have recently commissioned 1125/60 commercial productions with ZEBIC Productions and 1125 Productions, respectively, both of New York.

THE FUTURE

High-definition television is accelerating at a frightening pace. It cannot be stopped now. The two evolving scenarios are:

1. A 1250/50 h.d.tv system where programme producers get left in the cold. This will mean increased production and transcoding costs, since 1125/60 already exists and will continue to do so. Beneficiaries will be European and foreign-owned tv set manufacturers based in Europe, and one or two local studio equipment manufacturers.
2. The 1125/60 system. Programme producers get what they always wanted, a universal tape interchange format. European industry has to sharpen up its act to compete in a new worldwide market.

Footnote: just as Europe imagines the Japanese threat from consumer manufacturers, so America worries about the European threat from within. So watch this space.

Glossary

ATSC: Advanced Television Systems Committee (USA).

NAB: National Association of Broadcasters (USA).

SMPTE: Society of Motion Picture and Television Engineers (USA).

Eureka: European Technical Development Projects (Project no95).

NEW PRODUCTS

Frequency analysis workstation

Several instruments are combined to make an integrated analysis tool with uses in a variety of applications. The foundation of the Schlumberger 1220 is a two-channel f.f.t. analyser. Built-in software caters for four-channel sampling and the extra channels are added when the optional expansion card is plugged in. Any two of the four channels can be measured interactively through the 4 by 4 averaging matrix which, it is claimed, gives the equivalent of six two-channel analysers. The instrument operates over the frequency range 0 to 50kHz with up to 1000 lines frequency resolution and the ability to zoom in on any detail in 'real time'. The standard memory holds 256k samples. Optionally, this goes up to 1Msample. Other options include floppy-disc drive(s) and a built-in printer.

Incorporated within the 1220 are the facilities of a digital-storage oscilloscope, a transient recorder, a signal generator and synthesizer and several other functions. The instrument works from 0 to 50kHz and, because the signal generator is used to excite the system under test, can often be used by itself where formerly a cluster of instruments would be required.

Vibration analysis, acoustics and noise, loop response of control systems, speech and music analysis are among the areas of use for the instrument. With the 1Msample memory, it can record up to two minutes of speech and offer several ways to analyse it, including a "waterfall" which offers time, frequency and spectrum plots combined into one display.

The green display (there is also an RGB output to a colour monitor) can be divided into two or four, or used to display a single channel. Channels may be superimposed. All measuring parameters are clearly labelled.

Similar care has been put into the design of the keyboard with all keys labelled with understandable words, so that the instrument is easy to use. Built-in amplifiers allow most transducers to be connected directly without further signal-conditioning equipment. Schlumberger Instruments, Victoria Road, Farnborough, Hants GU14 7PW. Tel: 0252 544433.

Processor board tester

Simple debugging of complex boards can be carried out by the Polar B3T tester which is claimed to be easy to operate. No programming effort by the operator is required as the instrument runs pre-set test sequences. An interface pod plugs into the processor socket on the board and the test sequences are run through this.

Additional user ports enable the instrument to extend the test to include peripheral devices, edge connectors and external circuitry. Results are displayed on the instrument's l.c. display or output on its printer.

All buses and memory devices are exercised and tested. Rom contents can be disassembled using processor mnemonics. A logic/frequency probe is provided along with a power supply for the board under test. B3T can be controlled by an external computer through an RS232 port. Test sequences are stored internally on non-volatile memory or can be downloaded through a rom socket.

The instrument is said to be particularly useful in field service engineering, low-volume production testing, and prototype debugging. It can also be used to complement other test systems. Antron Electronics Ltd, Hamilton House, 38 Kings Road, Haslemere, Surrey GU27 2QA.
Tel: 0428 54541.

Kit for prototyping surface mounts

'All that you could ever need' is the claim by Electrolube for its kit of chemicals and tools for working on surface-mounted components. Major components are solder paste, adhesive, syringes for applying both, multi-purpose cleaner, spray-on conformal coating, tweezers and even a little pot to store components. For reworking there is a spray flux, desolder braid and a cleaning brush.

The solder paste is a mixture of low-oxide solder powder in a flux binder. Small particle size permits use for screen printing and automatic dispensing. Heat cures the adhesive but it is sticky enough to hold the components before curing, and is flexible enough after curing not to stress the components. Conformal coating offers protection against moisture, solvents and lubricants. It can be soldered through for reworking. Before de-soldering, a quick squirt of spray flux makes it easier and helps the replacement part to adhere. The solvent cleaner provided is non-flammable and the aerosol spray is sufficiently powerful to penetrate under components and blow away any debris.

All parts can be re-ordered individually. Electrolube Ltd, Blakes Road, Wargrave, Berks RG10 8AW. Tel: 073 522 3104.

Circuit breakers for p.c.b. protection

Ranges of circuit breakers have been developed which combine circuit protection with very small size. This makes them easy to design into complex p.c.b. layouts. Two types are available: the 808 series has 26 current ratings from 0.006A to 3.25A; and the 104 series with 24

ratings from 0.05A to 8A. The 108 Series is designed for vertical or horizontal mounting and can save as much as 70% board space compared with other breakers. ETA Circuit Breakers Ltd, Broadfields, Bicester Road, Aylesbury, Bucks. Tel: 0296 20336.

NEW PRODUCTS

Synthetic flux for wave soldering

Traditionally, p.c.b. manufacturers have maintained a loyalty to rosin-based fluxes, which although they produce little corrosion, are not self-cleaning. Many synthetic fluxes have given contamination problems.

Main features of a new synthetic flux from Fry's are its minimal residue and its ability to encapsulate fumes created during wave soldering and render them non-corrosive.

According to Fry's, these factors make this SM88/2X flux ideal for high-volume wave soldering, and especially for those who automatically test boards in their production process. Fry's Metals Ltd, Tandem Works, Christchurch Road, London W19 2PD. Tel: 01-468 7020.

Microprocessor-board tester

By 1990, 80% of all digital p.c.bs will include a processor. So says Fluke, who have produced the Series 90 tester for use with such boards. This is fitted with a clip that plugs over the resident processor on the board under test. Switch-on initiates a number of tests including the system clock, the operation of memory access protocol, wait and reset states. Other tests, of memory and bus functions and of the input/output facilities are carried out from commands entered through the instrument's keyboard. Further tests use the tester's synchronized logic probe and a QuickTrace facility can identify and display the location of nodes to which the probe is connected.

Connectors are fitted for an external power supply (normally, the instrument is powered from the supply of the board under test), and for an RS232C interface. This adds further fault-finding power by allowing extra facilities, such as setting processor breakpoints, down and uploading the contents of memory. Even more tests can be automated through the RS232C link and a personal computer with a suite of software called QuickTools.

Each Series 90 tester is dedicated to a specific processor. Three versions are currently available for Z80, 6809 and 8085 processors. They cover a high proportion of all such processor cards especially as all the second-source and variations of the processors are included, so long as the pin-out remains the same.

An alliance between Fluke and Philips means that Philips are now the European agents for Fluke and details of all Fluke products can be obtained from Philips Test and Measurement, Colonial Way, Watford, Herts WD2 4TT. Tel: 0923 240511.

Not just an i.c. tester

An addition to the ABI range of i.c. testers is the DIT-24XP which is used for in-circuit testing. It not only tests the functions of the chip under test but also identifies open circuit pins caused by broken tracks, poor soldering or plated-through holes that aren't. Short-circuits to the ground or supply pins and links between any pins are also detected. All faults are noted on the outline of the device displayed on the built-in c.r.t. — a feature which ABI claims unique.

Power to the circuit under test is provided by the instrument. The i.c. type number is entered on the keyboard and a test clip is hooked onto the device. If a device is unidentified the tester has a search routine to check its functions against its internal library and identify the i.c. It makes no difference how the device is configured in a circuit: automatic compensation is included in the tester to cope and still requires a single test pattern for each device in the library. Loop testing modes are used to identify intermittent faults.

An additional facility is the ability to save the test results from a known 'good' board and then compare other boards. Differences lead to rapid diagnosis of any faults. Sockets on the front of the instrument check components before insertion.

Software supplied with the instrument for a variety of devices including t.t.l, c-mos, memory and interface chips. An update service adds new devices. ABI Electronics Ltd, Mason Way, Platts Common Industrial Park, Barnsley, South Yorks S74 9TG. Tel: 0226 350145.

New liquid crystal displays

Epson are confident that their new neutralized twisted nematic (NTN) liquid crystal displays will replace c.r.ts for computers. Super-twisted nematics (STN) always had the disadvantage of being, usually, yellow and blue; of not having a good contrast and a relatively narrow viewing angle. NTN displays are black and white, offer twice the contrast of STNs and a wider viewing angle. Other advantages are a low power consumption, a compact (25mm thick) size and no flicker or electromagnetic radiation.

Electronics for driving the display are mounted on a flexible film p.c.b. leaving room behind the display to install a cold-cathode fluorescent backlight. Three displays are due for release in Autumn: 640 by 200 dots, 640 by 400 and 640 by 480. These correspond with the PC CGA, EGA and VGA display standards respectively. Prices have not been announced, but are expected to be less than one-and-a-half times those of STN equivalents. Epson UK Ltd, 388 High Road, Wembley, Middlesex HA9 6UH. Tel: 01-902 8892.

Optical communications light sources

Laser and led plug-in modules are accommodated in an Ando Electric 'mainframe' light source controller and modulator. AQ-4137 can house one laser source or two leds while the AQ-4141 accommodates up to three laser or six leds or a combination of these. Both models can attenuate the output by up to 6dB and they drive the sources in carrier-wave or chopped mode. Laser sources can be modulated by an external signal between 0.3kHz and 100kHz.

Wavelengths of the led modules are 660nm and 850nm unfiltered, or 1300nm and 1550nm with a choice of filtered output. One unit is switchable between 1300nm and 1550nm. The longer wavelengths are used for single or multi-mode operation.

Laser units have 850nm, 1300nm and 1550nm wavelengths and another can also switch between the two longer wavelengths.

All sources are temperature compensated to maintain stability. GPIB interface is provided on both models. Soon to be released is an additional range of high-output, high-stability led modules for use with the controllers. Available through Aspen Electronics Ltd, 1 Kildare Close, Eastcote, Ruislip, Middlesex HA4 9UR. Tel: 01-868 1311.

Hybrid power circuits

Different circuit-building methods are combined by General Hybrid, who is able to partition a circuit and apply the specific method to a particular part of a circuit. This leads to improved thermal management and reduces the circuit size considerably. One design can combine direct heavy-wire bonding of chips onto a copper substrate, thick-film resistors, surface-mounted and conventional p.c.bs. Such circuits can carry up to 300A, and, depending on the provision of heatsinks, dissipate over 100W/in^2. Similar circuits have been used for motor-drive controllers, robotics, d.c.-to-d.c. converters, uninterruptible power supplies, programmable voltage regulators, telecommunications systems and process control.

The same techniques are used to combine electronic control circuitry within power circuits to provide what Hybrid calls Smart Power. General Hybrid, Lawson Hunt Industrial Park, Broadbridge Heath, Horsham, West Sussex RH12 3JR. Tel: 0403 40400.

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Let us send you data on the product featured above and update you on the extensive Kenwood instrument range.

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PE17 4BG. **Tel: (0480) 63570**

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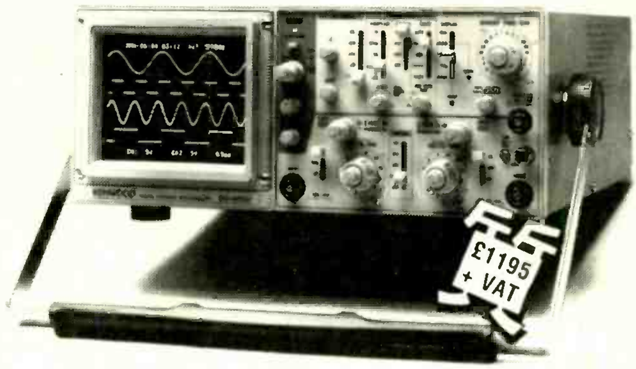

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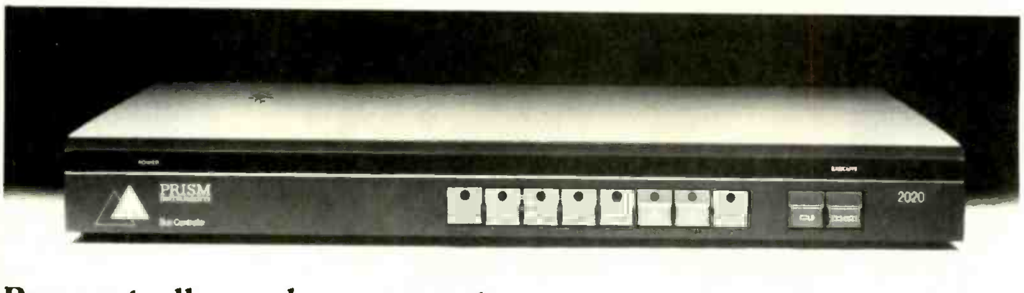
NEW PRODUCTS

Microwave amplifier

Gallium arsenide transistors are used in the MMA 4500 series of linear amplifiers for use in the 14GHz band. Gain and output power are up to 65dB and 3W respectively. Modular microstrip circuitry conforms to the relevant British standard and incorporates isolators to ensure stability and v.s.w.r. under all operating conditions.

The modular construction also allows a high degree of flexibility in providing a number of options, which include: power monitoring, automatic level control, gain control, built-in test equipment, and higher output powers. Units can be built to meet the requirements of fixed or mobile ground-based transmitters for satellite communications. Densitron Microwave Ltd, 112 South Street, Braintree, Essex.
Tel: 0376 551717.



Bus controller replaces computer

Dedicated control systems can take advantage of the facilities offered by the 2020 GPIB bus controller. This has the ability to store up to eight independent control programs. Each program is held in a 16Kbyte eprom or in 8Kbyte battery-backed ram. Two independent IEEE-488 buses may be controlled for up to 28 linked devices.

Simple command language instructions allow easy control of devices on the bus including sending and receiving data, simple arithmetical operations and message handling, conditional testing and the logging of data to the RS232 output. Further commands adjust and read the internal facilities of the 2020, i.e. real-time clock, audible alarm, and status indicator lamps. Control programs can be prepared on a standard microcomputer and downloaded to the 2020 through the RS232 interface. Initially they are held in ram but when fully operational and debugged, it is recommended that they be made more permanent in eprom.

At a price of £595, the 2020 can replace a more expensive computer normally used as an IEEE-488 bus controller. Prism Electronics Ltd, Burrel Road, Industrial Estate, St. Ives, Huntingdon, Cambs PE17 4NF. Tel: 0480 62225.

Take a look inside

It is very annoying to have to dismantle equipment to find a simple fault that is not visible otherwise. That problem can be solved by an endoscope and two are made by Scholly in Germany. These are designed for industrial use but the principle is the same as those used by surgeons to look inside people without taking them apart: light is conducted through optical fibres and an image is reflected through a bundle of thousands of fibres.

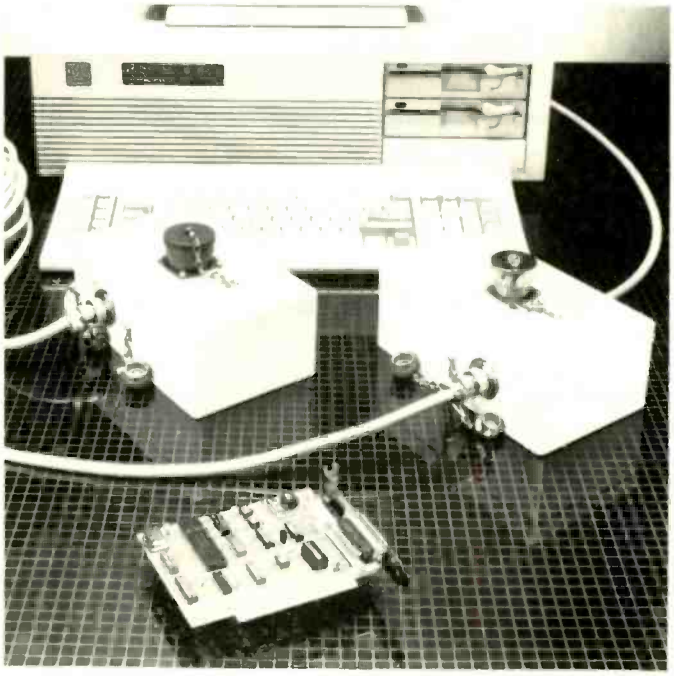
Miniflex uses rigid light guides and mirrors can be attached to the end to provide different viewing angles. The probes have diameters from 0.96mm, claimed to be the world's smallest, up to 2.7mm.

Fibrescope has a flexible light guide, interchangeable tips and mirrors are also available. Sizes start from 2.2mm up to 10.8mm.

Both instruments can be linked to a tv camera. Finlay Microvision Co. Ltd, Unit 6, Southfield Road, Kineton Road Industrial Estate, Southam, Warwicks CV33 0JH.
Tel: 092 681 3043.

Complex continuity testing by computer

Some cable harnesses can have many branches and be very difficult to test for continuity. Software has been developed for the IBM PC which can program a universal continuity tester and make the task much easier. The system can 'learn' a wire list which is created and edited on the computer. Test sequences run on the tester are controlled from the computer and offer an automatic documentation for quality-assurance testing. Short or open circuits, intermittent faults and incorrect wiring for up to 28,000 points can be detected at a rate of 1250 connections in a second. Expansion boards can be easily added to the UCT-1000 tester to enable the testing of even more points if needed. The software is called PC/UCT. Omnitest Ltd, Highcliffe House, 411 Lymington Road, Highcliffe, Christchurch, Dorset BH23 5EN. Tel: 04252 77731.



Communications multiplexer

Synchronous multiplexing of four different data streams is possible with the Feshon Datamizer II. Each of the four may be half or full duplex at a different data rate and all are transmitted through a single high-speed data link.

Compression of data to a quarter of its normal length eliminates the need for additional leased lines and can reduce line charges by up to 75%. Feshon Systems Ltd, Resicon House, London Road, Sevenoaks, Kent TN13 2DN. tel: 0732 560088.

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MOTOROLA M1000-100 5" CRT black & white compact chassis monitor measuring only cm 11.6h, 12w, 22d, ideal for CCTV or computer applications. Accepts standard Composite video or individual H & V syncs. Operates from 12v DC at appr x 0.8a. Some units may have minor screen marks, but still in very usable condition. Fully tested with 30 day guarantee & full data **Only £29.00 (C)**
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SAE for full list of other S100 boards and accessories.

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VISA

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NEW PRODUCTS

Three from HP

Low-cost digitizing oscilloscope
Low-price, general-purpose, benchtop oscilloscopes account for about half of the total oscilloscope market, says Hewlett Packard, who has decided to provide an oscilloscope at £2499 for that half. Several integrated circuits have been developed specially for the HP 54501A four-channel, 100MHz digital oscilloscope, to produce all the controlling circuitry on one p.c.b. It offers a sampling rate of 10Msamples/s. Other key features are its ease of use, unattended measurement, its triggering facilities and the HPIB link. An 'autoscale' button will automatically lock onto a signal and adjust parameters to give a clear trace on the screen. Specifications include 5mV sensitivity and 8-bit vertical resolution.

New features on the instrument include a measurement limit test which allows pass/fail testing between preset parameters; statistical analysis of measurements; and dual timebase windowing, which gives facilities similar to the delayed timebase on an analogue instrument for close up viewing of part of a trace.

The triggering facilities are particularly noteworthy. They include tv signal and inter-channel triggering. The ability to set a time window on a trigger means that the instrument can be set to trigger only on signals that are unexpected so that any glitches are automatically recorded.

The front panel has been designed to be easy to use with most parameters altered by a single keystroke or a knob.

Several accessories are offered as optional extras and include a tv/video sync. module with a clamped video output; and a number of probes, probe sockets, clips and adaptors.

Signal analyser with built-in Basic
A fast Fourier transform (FFT)-based analyser is claimed to provide easier test solutions, faster, and at lower cost. HP 35660A two-channel dynamic signal analyser includes many test and automation features that traditionally have required the use of an external computer.

The HP 35660A provides spectrum analysis from 0 to 102.4kHz and network analysis from 0 to 51.2kHz. The FFT provides 400 lines of resolution in both one and two-channel modes. The analyser has two input channels with 70dB dynamic range and a signal source that provides signals for stimulus-response testing. It measures linear spectrum, power spectrum, frequency response, gain/phase, group delay, time history and power spectral density.

Test engineers can use the analyser's new test-automation language, HP Instrument Basic. A built-in, 3.5-in disc drive stores traces, tables and Basic programs. Users can create algorithms and format results to provide answers not directly available from the standard analyser.

Tables of limits can be built into the memory for repetitive testing. Formerly measurements had to be transferred to a computer for comparison with specifications.

Using Instrument Basic, the HP 35660A can serve as a controller for HPIB (IEEE-488) test systems. The systems might include peripherals such as hard discs, printers and plotters, and instruments such as switch matrices, voltmeters and signal generators. This makes it easy to automate smaller systems without the need for an external computer. Instrument Basic is a subset of HP Basic. It adds decision-making, branching and i/o facilities, including the control of other instruments. One very useful facility is keystroke recording. This automatically creates a Basic program as the user makes measurements from the front panel. A test sequence can be recorded and saved with no 'manual' programming required. Basic programs also can be developed on HP workstations and then transferred to the analyser on discs. This Basic is also useful for normal computer-aided tests with an external computer. Keystroke-recorded routines can be merged with the main program being written on the computer.

In addition to user-developed software there is also commercially available software. HP offers the 35681A analysis pack, which provides enhanced features for spectrum and network analysis in electronics. Third-party software suppliers also are providing application solutions: Entek Scientific has converted its Motor Monitor program to the instrument and SMS is writing several packages specifically for the HP analyser, the first of which is the ME Toolkit.

Trigger to speed-up oscilloscopes
Many oscilloscopes offer the facility of external triggering where the internal trigger timebase might not be fast enough. HP has produced the 54118A trigger which runs from 500MHz to 18GHz. Moreover it actually runs at this speed and is based on a high-speed flip-flop linked to a tunnel diode without using count-down techniques. It can trigger on a leading or trailing edge. Its intended use is as an add-on for the HP 120 series of oscilloscopes in very high speed applications.

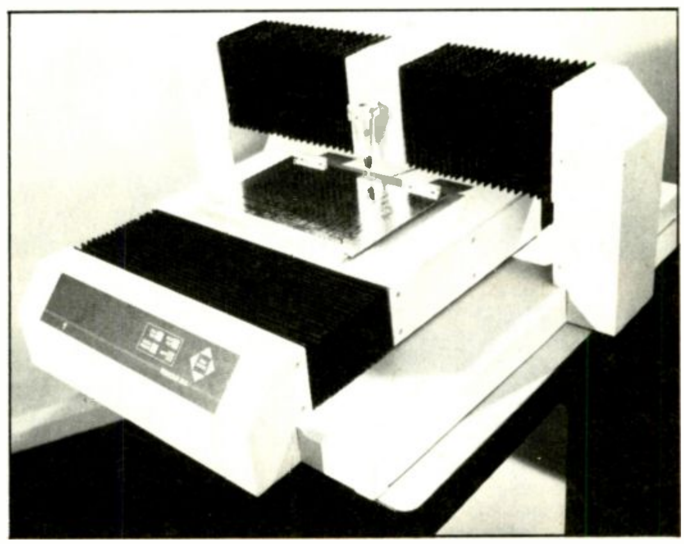
Details of all these HP products come from Hewlett-Packard Ltd, Winnersh Triangle, Wokingham, Berks RG11 5DZ. Tel: 0734 696622.

Speedometer chip

The mechanical odometer or mileometer may soon become relegated to the motor museum if Siemens' non-volatile, non-resettable, counter that records 'mileage' is successful. Called the SLE4501, the eeprom can store more than four million events, the equivalent to some 250,000 miles (400,000km). Conventional eeproms are usually limited to much less but an additional chip, the SLE 4502 c-mos pre-scaler, permits the electronic odometer to be adapted to any type of car.

The SLE 4501 incorporates a shift register, a 22-bit binary counter, two non-volatile 22-bit registers, a programming power source, a sequence controller, and for the command code, a two-bit shift register, and the 64byte eeprom, programmed by the car manufacturer according to wheel and tyre sizes and gearbox and differential gear ratios.

Counting is performed entirely on-chip and is immune to outside interference. The signals produced by the speedometer pulse generator are fed to the prescaler, the divider ratio of which is programmable from 1 to 65,000. The chip also contains a data protection facility that prevents tampering. What is more, prospective buyers of a used car can be more certain that the mileage displayed is correct. Siemens Ltd, Windmill Road, Sunbury-on-Thames, Middlesex TW16 7HS. Tel: 0932 75 2323.



Design computer for p.c.bs

By the time you read this, you should be able to get your hands on a Versatron 2000 p.c.b. development system that was on pre-launch demonstration at British Electronics Week. There are two main components; a computer dedicated to the task, and a plotter/drilling machine that produces prototype boards under the direct control of the computer. Combined, they are intended to ease the transition between hand drafting and cad by being easy to use and low in cost — 'about the price of a desk-top'.

Central to the system is the purpose-designed 16-bit, 10MHz computer with 1.2Mbyte of ram, colour monitor and a disc drive. Graphics have been developed to give smooth scrolling of images, and rapid zooming and panning. As a whole design is retained in memory these operations are very fast and do not suffer from the more usual redrawing delays experienced in such systems. Similarly, any erasures produce instant redrawing of the affected part of the design. The disc is only used to store component libraries and completed layouts.

The software is specifically designed to be easy to use with pop-up menus and highlight selection of options, all driven by a mouse. Component pad patterns can be designed and added to the library provided. Once components have been positioned, a multi-pass autorouting program draws the tracks. In many cases it will complete the whole process with no supervision and is claimed to be very fast.

Making a board is carried out by the prototyping machine which plots the pattern directly onto the bare copper board in etch-resistant ink. It also drills the holes automatically. The board is then etched and optionally plated ready for component insertion.

Refinements to be added during this year are circuit design software and programs to produce net lists and check rules. Versatronics Ltd, Mardy Road, Cardiff CF3 8EQ. Tel: 0222 770488.

Amplitude from phase

Calculation of a network's amplitude response when its phase response is known.

D.V. MERCY

In the last issue, I described a numerical method by which the phase response of a network can be calculated when the amplitude response is known. This article describes the reverse procedure, for obtaining the amplitude response. Some of the equations and figures referred to were shown in the first article.

AMPLITUDE RESPONSE FROM PHASE RESPONSE

In this section the equations to be considered are equns. (9) and (10). It can be seen that the answer is either referred to the amplitude value at zero frequency (A_0), or at infinite frequency (A_∞). These formulas assume, therefore, that either $\ln A_0$ or $\ln A_\infty$ will be finite. This is not always true, but measures can be taken to deal with this problem when it arises and these are discussed later.

An additional point to consider when the insertion loss (transmission) characteristic of a network is being derived is that it is not possible uniquely to define the amplitude response from the phase data, since any number of amplitude characteristics, differing only by a fixed number of dB, have the identical phase response. This means that the value of A_0, A_∞, or the gain at another frequency can be chosen with complete freedom in this case.

It is helpful in the first instance to plot the phase data against frequency, with frequency on a log-scale. Enough data is needed so that the detailed behaviour of the characteristic can be seen as well as its asymptotic behaviour at low and high frequencies.

With a network defined by equn.(1) and Fig.3, the general behaviour of the phase response follows a recognisable pattern. First of all, the phase of the function will be asymptotic to the value $\{-(P-Z).\frac{\pi}{2}\}$ as ω tends to infinity. Secondly, at very low frequencies, the phase will be asymptotic to the value $L.\pi/2$. Thus, inspection of the phase curve will indicate the value of (P-Z), the excess of poles over zeros, and L, the number of zeros at the origin.

As in the previous section the integration is divided into three regions, with analytic expressions derived for the high and low-frequency regions and with numeric integration for the mid-frequency region.

Amplitude response, Referred to zero frequency. Equation (9) is applicable in this case. However, before proceeding with the evaluation of the integral, it is necessary to remove any zeros present at the origin. Of course, this will not be necessary if the phase curve is that of a low-pass filter, but is certainly necessary if the network is a high-pass filter and probably necessary in the case of a band-pass filter. (Note that in the case of a high-pass filter, the use of equation (10) may be more appropriate and this is dealt with later).

To eliminate the zeros at the origin, the zero-frequency asymptote of the phase curve ($=L.\pi/2$) is first determined and then this value is subtracted from all the phase data that is to be used. As a result of this procedure, the high-frequency asymptote of the "new" phase curve will have the value $\{-(P-M).\frac{\pi}{2}\}$, where M is the number of zeros of the function not at the origin.

In others words, it is the function g(s), whose phase is now to be considered, where

$$g(s)=K.\frac{(s+a_1)(s+a_2)\ldots\ldots\ldots\ldots(s+9m)}{(s+b_i)\ldots\ldots\ldots\ldots\ldots\ldots\ldots(s+b_p)} \quad (33)$$

Once the amplitude curve has been obtained by the procedure outlined below, the effect of the zeros at the origin, if any were present, can be reintroduced by adding an amplitude characteristic with a slope of 20.L dB per decade, or 6.L dB per octave, to the results.

High-frequency contribution. If a frequency ω_h is selected in the high-frequency asymptotic region of the response, then the integral I_h can be expressed,

$$I_h=\frac{-2\omega_x^2}{\pi}\int_{\omega_h}^{\infty}\frac{\psi/\omega-\psi_x/\omega_x}{\omega^2-\omega_x^2}.d\omega \quad (34)$$

Now as $\omega\longrightarrow\infty$, then $\psi\longrightarrow-(P-M).\ \pi/2$ and equation (34) becomes

$$I_h\approx\frac{-2\omega_x^2}{\pi}\int_{\omega_h}^{\infty}\frac{-(P-M)\pi/2\omega-\psi_x/\omega_x}{\omega_2}.d\omega \quad (35)$$

$$I_h=(P-M).\omega_x^2\int_{\omega_h}^{\infty}\frac{d\omega}{\omega^3}+\frac{2\omega_x\psi_x}{\pi}\int_{\omega_h}^{\infty}\frac{d\omega}{\omega^2}$$

$$\frac{9_x/\omega_x}{\omega^2-\omega_x^2}.d\omega$$

$$I_h=\frac{\omega_x^2}{\omega_h^2}\left\{(P-M)+\frac{2\omega_h}{\pi}.\frac{\psi_x}{\omega_x}\right\}. \quad (36)$$

To determine (P-M), estimate the value of the phase as $\omega\longrightarrow\infty$

$$(P-M)=-\psi_{hf}.\frac{2}{\pi} \quad (37)$$

As (P-M) should be an integer value, it is permissible to round it up or down as necessary.

Low-frequency contribution. It is first necessary to consider the behaviour of the phase curve at low frequencies. It is shown in the Appendix that, for functions of the type given by equations (1) or (33), the phase curve always has a constant slope at low frequencies, which is given by

$$\left.\frac{d\psi}{d\omega}\right|_{\omega\rightarrow 0}=k \quad (38)$$

The frequency range over which this condition holds true depends on the actual pole-zero pattern of the function. To determine the extent of the linear phase region the phase curve should be plotted, using a linear frequency scale this time, for low freqencies.

The integral for the low-frequency region is given by

$$I_l=\frac{-2\omega_x^2}{\pi}\int_{0}^{\omega_l}\frac{\psi/\omega-\psi_x/\omega_x}{\omega^2-\omega_x^2}.d\omega \quad (39)$$

$$I_l\approx\frac{-2\omega_x^2}{\pi}\int_{0}^{\omega_l}\frac{k-\psi_x/\omega_x}{-\omega_x^2}.d\omega$$

The phase angle is zero at $\omega=0$ and is linear for a range $0\leqslant\omega\leqslant\omega_l$ i.e. $\psi=k.w.$ for small ω.

Therefore,

maths

and finally

$$I_l=\frac{2.\omega_l}{\pi}\left\{k-\frac{\psi_x}{\omega_x}\right\} \quad (40)$$

As discussed above, k is the slope of the phase curve near to $\omega=0$.

Mid-frequency contribution. The expression to be evaluated is

$$I_m=\frac{-2\omega_x^2}{\pi}\int_{\omega_l}^{\omega_h}\frac{\psi/\omega-\psi_x/\omega_x}{\omega^2-\omega_x^2}.d\omega \quad (41)$$

and, as before, it is necessary to use numeric integration to evaluate this expression.

Since only a limited number of values of phase are likely to be available to plot the phase curve, interpolation will be a necessary part of the procedure. However, a method similar to the one used earlier gives disappointing results, i.e. where points on a plot of phase against log(ω) are connected by a series of straight lines and where logarithmic interpolation is used to obtain phase values at intermediate frequencies.

A better procedure is to use a frequency scale normalized with respect to a "pivot" frequency ω_0, in the same way as was shown in Fig.1(b). This pivot or normalizing frequency is best chosen from inspection of the initial plot of phase against $\log(\omega)$. A good choice, for example, is a frequency about which the response is approximately skew symmetrical (if this is possible).

An incidental advantage of the new plot, because it has a linear frequency scale up to ω_0, is that it can be used to determine the slope of the phase curve at low frequencies, as required in for the low-frequency contribution.

Interpolation between points is carried out in a linear manner on the new scale. Thus, if the initial frequency values are ω_1 to ω_n, then the normalized frequency values u_1 to u_n are given by

$$u_m=\omega_m/\omega_0 \text{ if } \omega_m\leq\omega_0 \tag{42}$$

$$\text{and } u_m=\omega_0/\omega_m \text{ if } \omega_m>\omega_0 \tag{43}$$

so if two adjacent points on the phase curve have phase values ψ_a and ψ_b at frequencies ω_a and ω_b respectively, and the interval is subdivided into (n-1) equal intervals, then the intermediate values are given by

$$\psi_m=\psi_a+(\psi_b-\psi_a)\frac{m}{n} \tag{44}$$

for $o\leq m\leq n$

$$u_m=u_a+(u_b-u_a).\frac{m}{n} \tag{45}$$

for $o\leq m\leq n$

From (45), together with (42) and (43) the corresponding values of ω can be found, as required.

If the amplitude value is currently being calculated at frequency ω_x (or u_x) and the phase value there is ψ_x, then the value of the integrand function is

$$BD_m=\frac{\psi_m/\omega_m-\psi_x/\omega_x}{\omega_m^{\ 2}-\omega_x^{\ 2}} \tag{46}$$

The area of each sub-interval, obtained by the trapezium rule, is

$$\text{Area}\,\Big|_m^{m+1}=\tfrac{1}{2}.(BD_{m+1}+BD_m)(\omega_{m+1}-\omega_m) \tag{47}$$

and the total area between these two adjacent nodes is the sum of the sub-areas as m is stepped from one to n. This process is repeated for all the segments in the mid-frequency range.

As before, however, there is a problem when $\omega_m=\omega_x$. Equation (47) cannot be used directly at this frequency, and it is necessary to find the limiting value of the expression as $\omega\rightarrow\omega_x$

$$\text{i.e. } BD_x=\lim_{\omega\rightarrow\omega_x}\frac{\psi/\omega-\psi_x/\omega_x}{\omega^2-\omega_x^{\ 2}}\approx\frac{1}{2\omega_x}.\frac{d}{d\omega}\left(\frac{\psi}{\omega}\right) \tag{48}$$

Two values of the limiting expression have to be found, one as ω approaches ω_x from the low-frequency side and one as ω approaches ω_x from the high-frequency side.

With these additional values, all the required values of BD can be calculated, so all areas for the mid-frequency region can be calculated. Then

$$I_m=-\frac{2.\omega_x^{\ 2}}{\pi}\sum_{\omega_{11}}^{\omega_{h2}}\text{areas}$$

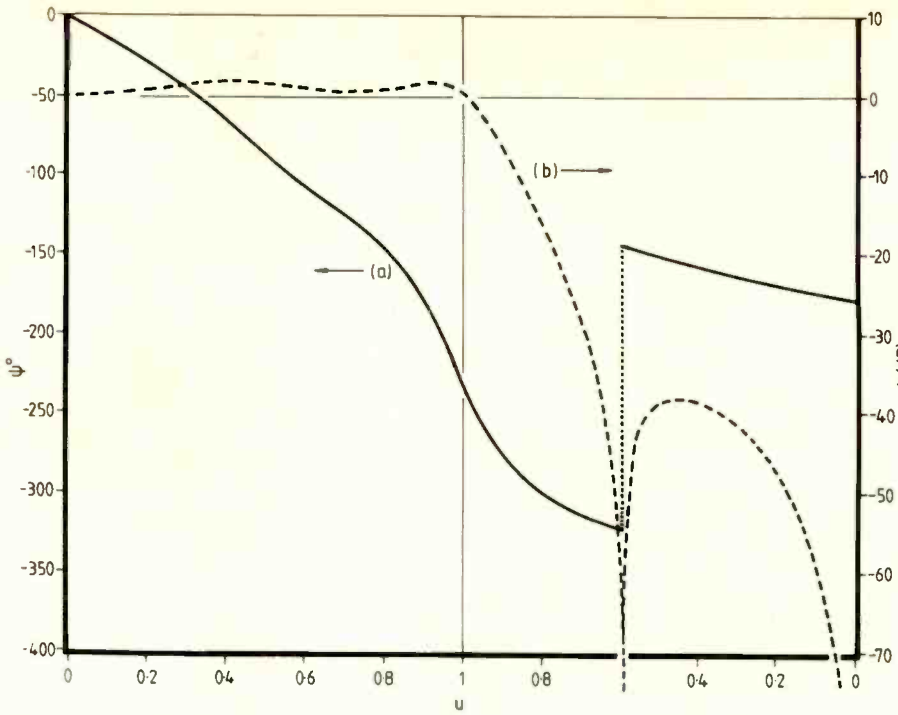


Fig.8. Amplitude (b) derived from phase (a).

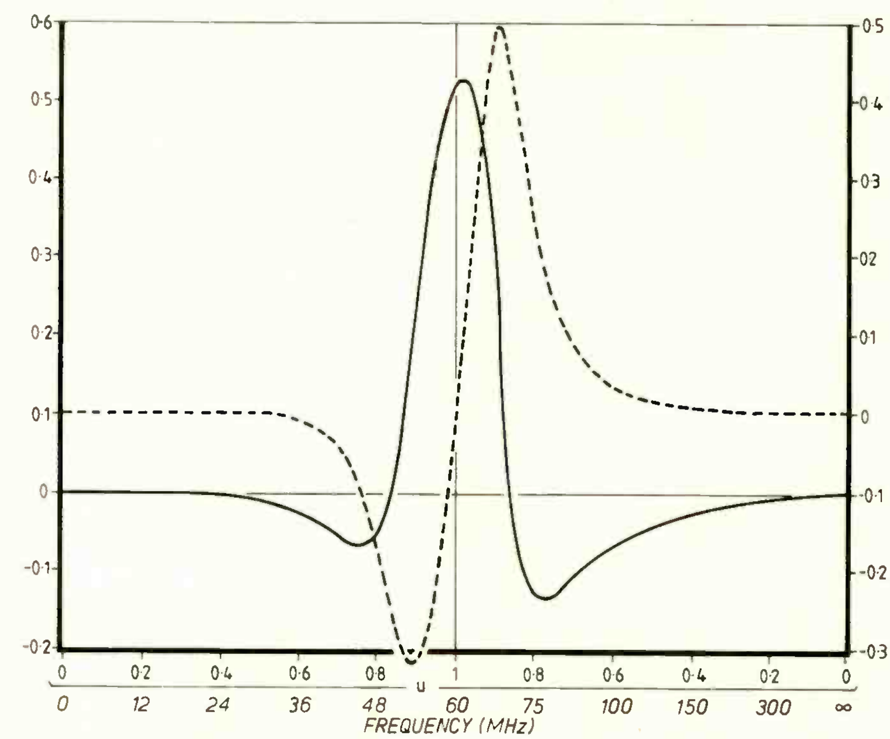


Fig.9. Real part (b) of transfer characteristic of band-pass filter, with its imaginary part at (a).

Sum of all contributions. At a given ω_x, the sum of the three contributions gives the required answer, in nepers.

$$\text{i.e. } \ln A_x-\ln A_0=I_l+I_m+I_h \tag{49}$$

The results can be easily converted to dB, as mentioned earlier. Results at other values of ω_x can be carried out as required.

Examples. Two examples are given in this section to demonstrate the procedure. As before the interval between adjacent nodes has been subdivided into 10 parts when carrying out the numerical integration.

Example 4. The phase plot for this example is given at (a) in Fig.8 and the analytic solution for the amplitude is the full line (b). The plots are those of a fourth-order, elliptic,, low pass-filter with a pass-band ripple of 1.25dB and a stop-band attenuation of about 38dB.

Triangles on the first curve (phase) are the input data. Circles on the second curve (amplitude) show the initial output data, with additional interpolated results shown by squares.

Example 5. In this example the real part of a

Fig.10. Filter used in example of Fig.9.

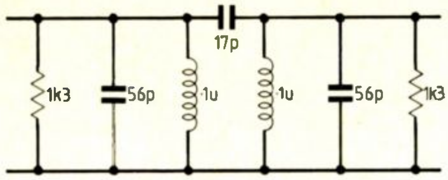
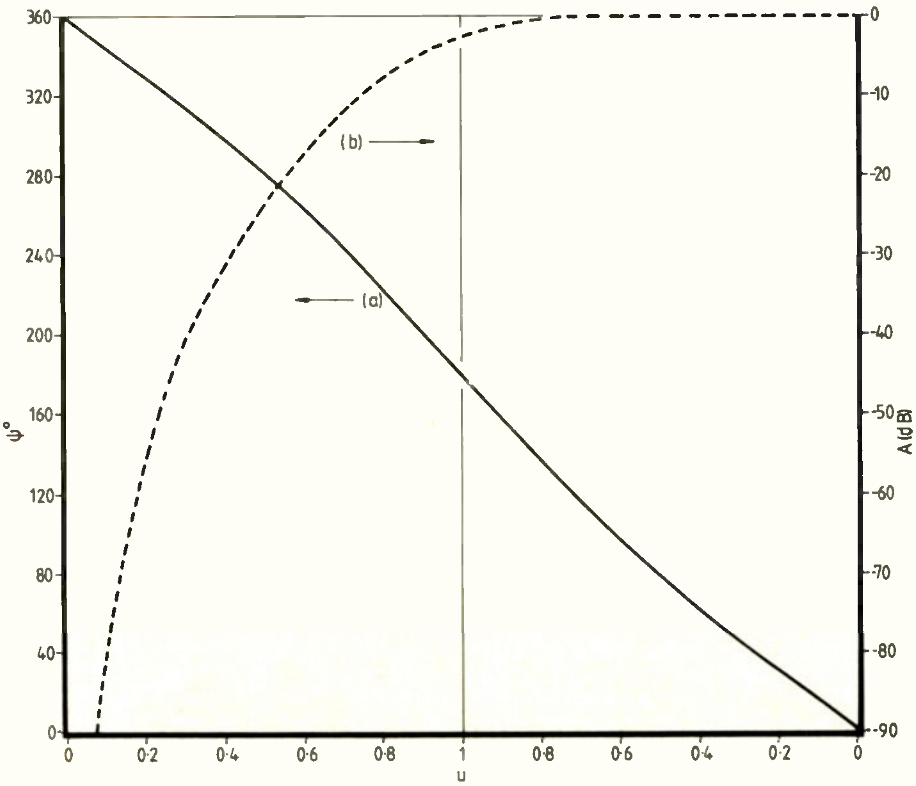


Fig.11. Phase response (a) of a 4th-order high-pass filter and its calculated amplitude response (b).



network function is derived from data on the imaginary part. Figure 9 gives the two analytically derived responses, which are the characteristics of the top-capacitively-coupled band-pass filter shown in Fig.10. The results are accurate to 0.02 worst case, with most of them well within this margin.

Amplitude response, referenced to infinite frequency. Equation (1) is the one to be considered. It should only be used in cases where the phase response is asymptotic to zero at high frequencies, which is the case, for example, when the network is a high-pass or band-stop filter. Although the procedure could perhaps be made more general, there is little point in doing so since all other situations can be dealt with by the methods described earlier.

As in the preceding sections, the integral is divided into three frequency bands. The procedures are very similar, so the results are given this time with little discussion.

At a given ω_x, the amplitude A_x is given by

$$\ln A_x - \ln A_\infty = I_l + I_m + I_h \qquad (50)$$

where

$$I_h = \frac{2}{\pi\omega_h}.\psi_x.\omega_x \text{ provided that } \omega_h \gg \omega_x \qquad (51)$$

$$I_l = \frac{2\omega_l}{\pi\omega_x^2}\left(\frac{k.\omega_l^3}{3} + \frac{\psi_l.\omega_l}{2} - \omega_x\psi_x\right) \qquad (52)$$

provided that $\omega_l \ll \omega_x$ (52)

and where I_m is obtained by numerical integration methods.

In determining I_m, values of the integrand have the form

$$BD_m = \frac{\psi_m.\omega_m - \psi_x.\omega_x}{\omega_m^2 - \omega_x^2} \qquad (53)$$

except where $\omega_m = \omega_x$, when

$$BD_x = \frac{1}{2\omega_x}.\frac{d}{d\omega}(\psi.\omega) \qquad (54)$$

and where, as before, two values at each ω_x are required. The summation of all the areas in the interval ω_l to ω_h calculated in a similar manner to those of the earlier sections, allows the value of I_m to be determined, since.

$$I_m = -\frac{2}{\pi}\sum_{\omega_l}^{\omega_h} \text{areas} \qquad (55)$$

Example 6. Figure 11 is the phase response (a) of a Butterworth fourth-order high-pass filter and the circles and squares on (b) show the calculated amplitude response points.

Example 7. The imaginary part of the network function shown in Fig.9 is asymptotic to zero at high as well as low frequencies, so it can be investigated by the methods of this section, using the same input data as was used in example 5 (i.e. the triangles in Fig.9).

APPENDIX 1 Phase response of minimum phase shift networks at low frequencies.

The function being considered has L zeros at the origin, with M zeros and P poles in the left half plane. If the left-half plane zeros have positions given by

$Z_i = \sigma_i + j\omega_i$ for $1 \leq i \leq M$

and the poles by

$P_i = R_i + j\omega_i$ for $1 \leq i \leq P$

then the phase value at a given frequency is given by

$$\psi(\omega) = L.\frac{\pi}{2} + \sum_{i=1}^{M} \tan^{-1}\frac{\omega+\omega_i}{\sigma_i} - \sum_{i=1}^{P} \tan^{-1}\frac{\omega+W_i}{R_i}$$

The slope of the phase curve is

$$\frac{d\psi}{d\omega} = \sum_{1}^{M} \frac{1}{1+\left(\frac{\omega+\omega_i}{\sigma_i}\right)^2}\frac{1}{\sigma_i} - \sum_{1}^{P} \frac{1}{1+\left(\frac{\omega+W_i}{R_i}\right)^2}\frac{1}{R_i}$$

which at low frequencies approximates to

$$\left.\frac{d\psi}{d\omega}\right|_{\omega\to 0} = \sum_{i=1}^{M} \frac{\sigma_i}{\sigma_i^2+\omega_i^2} - \sum_{i=1}^{P} \frac{R_i}{R_i^2+W_i^2}$$

= a constant for a given pole-zero pattern
= k

The frequency range over which the slope remains constant is determined by the location of the poles and zeros of the function. In some cases the linear region extends over a considerable portion of the passband.

Program listings for the procedure described in the two articles can be supplied from this office. Please write in and enclose an A4 stamped and addressed envelope. Mark your communication AMPHASE.

David Mercy is with Thorn EMI Electronics Ltd, Hayes, Middlesex.

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TELEVISION BROADCAST

Counting the costs

At a time when broadcasters everywhere are seeking to economize and are placing greater emphasis on the total costs of ownership of their engineering facilities, including staff costs, it may seem strange that there is so much talk of system changes. The days have gone when engineers alone largely determined, primarily on the grounds of technical quality, equipment procurement policies. The slimming-down of engineering costs and staff costs is now a priority objective of many companies.

The impact on the broadcast equipment industry is considerable. Major firms in Europe, North America and even Japan are finding the going tough, with the marked retreat from systems (turnkey) contracts. It is not surprising that engineers as well as accountants are increasingly concerned with the amount of money needed to continue the development of digital-component, h.d.tv and d.b.s. equipment which may, or may not, translate into major sales in this century. It is recognized that it is vitally important to invest in r&d but equally vital to pick potential winners.

Despite, or perhaps because of, such doubts, large industry exhibitions continue to expand and multiply. The Brighton IBC88 (September 23-27) will see more than 200 exhibitors filling a record amount of space – including a temporary marquee 'pavilion' near the West Pier. There will be a special Eureka (EU95) h.d.tv presentation, plus about 100 solidly engineering papers (of some 200 offered) stemming from ten countries.

But there has apparently been a retreat from the presentation of at least one keynote paper devoted to broadcast economics: one recalls Michael Checkland's prophetic "The economics of television production" in 1984; David Reay's similarly titled address in 1986; and R.V. Arnaboldi's thought-provoking "Consumer electronics – the next five years" in 1982, in which he stressed that the improvements in systems must be clearly demontrated in the High Street. All the five keynote addresses this year appear, from the provisional programme, to concentrate on technology rather than economics. Perhaps few people are prepared to contemplate too realistically the short-and long-term commercial prospects for h.d.tv or d.b.s. Yet rather more than 30 of the papers bear upon the technology of h.d.tv and e.d.tv.

The IBA has recently demonstrated to the press a closed-circuit 625-line, wide-screen e.d.tv system using an eight-foot wide Sony projection display (priced at £50 000, not exactly geared to domestic use). For such demonstrations the IBA has made an attractive 20-minute video, "Images from Winchester" (director, Peter Sykes). This proved (at least to me) that 625/50 "component" video, when displayed as 100Hz progressive scan by means of frame-store memories, is capable of providing sufficient resolution, flicker reduction and absence of cross-colour effects for all presently-conceivable large domestic receiver displays likely to be available at consumer prices this century. It also showed that such material can be originated satisfactorily by modifying standard studio equipment, including cameras and v.t.r. machines.

There could, admittedly, be a role for 1125 or 1250-line systems for electronic cinematography, video-to-film transfers and for some master tapes where the picture quality does not have to be degraded by compression of the 50MHz or so baseband for transmission. Yet the degree to which r&d work is being concentrated on systems to provide improved pictures in the home is underlined in the February issue of the *IEEE Transactions on Consumer Electronics*. This 278-page issue is devoted entirely to "advanced television systems", including high definition (over 1000 lines) systems, "enhanced" (also called "extended quality") television and "improved composite television" (sometimes called "improved NTSC") together with questions relating to their digital transmission over the future broadband integrated services digital network (B-ISDN).

A recent listing of currently proposed advanced systems relating to 525-line, 60-field areas shows that many of the proposals rely on the use of frame-store memories in the receivers. Among the proposals are: the NHK wide-screen studio system (1125/30/60); seven "single-channel" systems capable of being transmitted in the FCC 6MHz-wide broadcast channels: NBC/Sarnoff 5:3 "ACTV"; MIT 4:3 bandwidth-efficient system (1200/60/60); Fukinaki 4:3 system (1050/60/60); 5:3 HD-NTSC (Iredale); MIT's 16:9 receiver-compatible system; 4:3 Super-NTSC (Faroudja); and a 4:3 Yasumoto system. Additionally there are three proposals for compatible systems that would require two or one-and-a-half terrestrial channels: Bell Laboratories (Rzeszewski) 5:3, 6+6MHz, 1050/30/60 system; William Glenn's 16:9, 6+3MHz, 1125/30/60 system; and North American Philips 16:9, HD-NTSC, 1050/60/60. Three wideband (satellite channels) proposals are NHK's Hi-Vision with MUSE (10MHz baseband) 16:9, 1125/30/60; Scientific Atlanta's HDB-MAC (10.7MHz) 4:3, 1050/30/60; and HD-MAC-60 by North American Philips, 9.5MHz, 16:9, 1050/60/60 with a four-field sequence.

Although a majority of the 45 member-organisations forming the Advanced Television Systems Committee (USA) have voted for the adoption of the NHK 1125/60 system for the proposed h.d.tv *production* standard, both the influential National Association of Broadcasters (NAB) and the Association of Maximum Service Telecasters (MST) were among the 11 organizations opposed (26 voted for, eight abstained). NAB and MST believe the decision is too restricted and pressed for delaying a decision "in view of new information about serious technical issues that have not been evaluated to our satisfaction."

O.b. vehicle trends

The first transportable production centres, mobile control rooms, scanners, call them what you will, for outside broadcasts ("remotes") date back to the 'thirties. Over the years there has been a wide variety of approaches, from the mammoth 'Type 5' BBC pantechnicons to battery-operated single-camera general-purpose units. The Royal Television Society's London Centre recently devoted a discussion meeting to attempting to estimate trends likely to affect the next generation of o.b. vehicles now being planned. Chaired by John Jarvie (BBC), panellists included Arthur Duff (Thames), Bob Warren (Thames) and Roger Jephcott (BBC) and an audience including many o.b. practitioners and representatives of specialist manufacturers including Ampex, Brabury and Sony.

The need to respond to programme requirements, involving more camera channels, more use of mobile recorders, character generators, digital video effects, stereo sound and longer programmes demands more and more on-board facilities. But sports o.b. programmes, for example, remain cost-effective, with an average cost (BBC) of about £24 000 per hour, less than half the average of all programmes. Bob Warren believes that fleets will evolve towards flexible super-large units with up to ten or a dozen camera channels, supported by more small units of one or two cameras, but with less use of "middle ground" units with three, four or five cameras.

A current problem is in deciding the extent to which o.b. operations will tend to component rather than composite video. Thames will soon have a small unit with a c.c.d. camera and M.II v.c.r. which will, in effect, be a component unit; but for large scanners an all-component approach would involve considerable extra cost, extra weight and extra equipment volume, and it seems unlikely that all-component working will come before h.d.tv is required.

There seems little prospect of large organizations buying off-the-shelf vehicles from a standard catalogue specification; most organizations cherish their own way of doing things and demand custom-built vehicles. Production staff as well as engineers have specific requirements and "in no way would use a standard vehicle".

Television Broadcast is written by Pat Hawker.

RADIO COMMUNICATIONS

Role changes for h.f.

The selective nature of the resurgence of interest in the h.f. spectrum was evident at the IEE's well-attended (over 330 delegates from 25 countries) h.f. radio systems and techniques conference. A breakdown of the 161 UK delegates shows that 82 were from industry (almost all concerned with defence); 28 from universities and polytechnics; 22 from Government and Service research establishments; 11 from GCHQ, HMGCC and MoD; 9 from BBC External Services; 7 miscellaneous, including technical press; and just two from British Telecom International, with none from Cable & Wireless. My mind went back 25 years to the pre-satellite communications era to a March 1963 IEE "Convention on H.F. Communication" which enjoyed major participation by the Post Office and Cable & Wireless, with the industry still regarding civil communications as major markets.

One consequence of the change in role is that there is no longer the same drive towards universal standards and systems; instead organizations are much more interested in doing their own thing and keeping their traffic, modes and frequencies to themselves. It was perhaps unfortunate that a French slide unwittingly revealed that they operate their experimental Valensole skywave radar on 14.147MHz, right in the middle of one of the 'exclusive' amateur bands. Dr G.F. Earl assured the conference that the Australian 100kW Jindalee facility has a frequency management protocol that prohibits any incursion into exclusive amateur allocations.

The increasing use of h.f. skywave communications for medium distance (one hop) links has brought about more interest in antennas having a vertical radiation pattern which maximises near-vertical incidence. The need for frequency agility has directed attention towards horizontally-polarized transportable broadband antennas, rather than the customary vertical whip antennas, with their upward-pointing nulls. Dr B.A. Austin and Andre Fourie described their resistive-loaded broadband wire antenna (see this column, September 1987). C&S Antennas Ltd has also developed "A wideband transportable antenna for n.v.i.s. links" based on a three-wire fan dipole with resistive loading near the ends of the dipole arms. The complete system, including a 12m, 8kg carbon-fibre telescopic mast, weighs only 16kg and has been erected single-handed in 15 minutes.

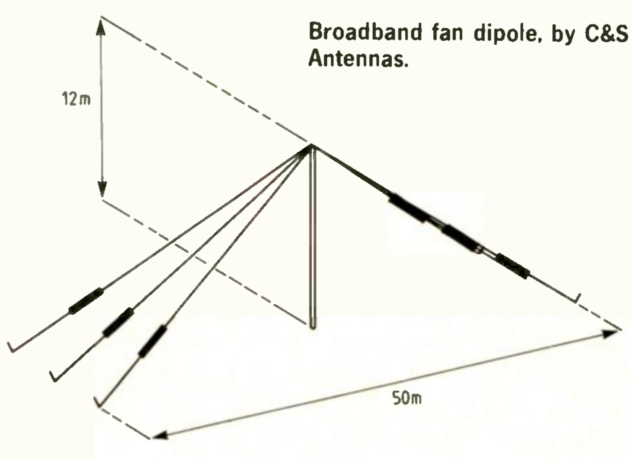
Efforts by a team at RAE to decide whether space diversity reception does or does not significantly improve the throughput of a meteor burst communications system had so far proved indecisive and "further study is necessary before a definitive statement can be made". Their basic "Blossom-A" system with 2400 baud bursts at 36MHz (other trial frequencies were 46, 70 and 71MHz) over the 813km North/South path from Wick in northern Scotland to the RAE's radio station at Corbett Hill, near Farnborough in Hampshire, has provided equivalent data rates of from about 5 to 45 baud averaged over ten-minute periods and from about 6 to 20 baud averaged over 60 minute periods. Maximum throughput is at about 10 hours GMT with an afternoon minimum at about 1700.

Featured on the first day were a number of ingenious adaptive and real-time channel evaluation systems, including the Plessey system developed to provide MoD with an "unattended" network of 100 terminals at up to 80 sites (see Radio Communications, October 1987). The prime objective of these systems is to reduce the need to train morse and technical operators, though one wonders whether keyboard systems will ever instill into users the same sense of personal involvement in getting messages though under adverse conditions. On the other hand there is no doubt that r.t.c.e. techniques involving constant sounding of several frequencies lead to more effective use of the h.f. spectrum, including the use of much higher frequencies for medium distance links during sporadic E propagation conditions. There appears to be a growing disillusion about the use of monthly average predictions made many months in advance, even when supported by d.i.y. computer systems such as Minimuf. The behaviour of the new sunspot cycle 22 has been virtually unpredictable in advance, moving the critical frequencies up and down erratically.

Keith Thrower (Racal-Chubb) in a keynote address pinpointed the current interest in adaptive systems and digital signal processing; the h.f. problems imposed by fading, multipath cochannel and the variable propagation conditions; the military desire to avoid detection and to defeat jamming; the constraints imposed on transportable systems in terms of power, antennas, size and weight; the desire of users to minimize data errors and improve voice quality. He noted that adaptive signal processing permits microprocessor control of the bandwidth of filters, type of demodulator, data rate and choice of frequency. He emphasized that adaptive and related techniques have become more viable with improved digital signal processing chips.

Keith Thrower noted the use of direct-conversion (zero i.f.) receivers which at first sight appear very attractive for the implementation of d.s.p., but he felt that this does not negate the need for a superhet-type first mixer. The technique greatly reduces the problems of designing frequency synthesizers having to cover the five octaves from 1 to 30MHz with the high local oscillator output needed to increase dynamic range, but it also results in the leakage of the oscillator signal to the antenna. A super-d.c. 1-30MHz digital receiver could result in a synthesizer covering 42.4 to 71.4MHz (less than an octave) with only one high-quality mixer stage followed immediately by a-to-d conversion.

• *See also the article 'H.f. developments', page 644.*



Broadband fan dipole, by C&S Antennas.

Mobiles win v.h.f. argument

The DTI has officially confirmed that the UK will continue with its policy of using v.h.f. Bands I and III for mobile radio communications rather than re-introduce v.h.f. television broadcasting. This would be possible under the international Radio Regulations although the DTI believes it could be "very difficult and costly to negotiate with neighbouring administrations who would regard a policy reversal by the UK as a serious disruption of the international understandings on which their domestic planning over the past few years has been based".

The DTI has also reported, following the initial meetings of the Council of Ministers' working group on the EMC Directive, that "progress is likely to be slow, in view of the detailed difficulties we and other Member States have raised, and there is now no prospect of Council final agreement by June 30". Initially the DTI found no support for the UK view that the Directive was unworkable where no national standards existed, but later the Italians expressed strong support. The Greeks have proposed that the Directive should not enter into force before the end of 1992, but this seems unlikely to achieve majority support.

Radio Communications is written by Pat Hawker.

Beefing up h.f.

Under its External Services improved audibility project, the BBC is installing two more high-power transmitters at its Ascension Island relay base. Originally fed by an s.s.b. link, this relay has for several years had a high-quality satellite feed. It serves primarily West Africa and South America. These recent pictures by BBC engineer Gerry Mills show the receiving dish for programme feeds (note the almost vertical elevation, due to the site's equatorial latitude) and, below, the antenna feed lines leaving the switching matrix.

The BBC has now virtually completed the modernization of one of its main UK h.f. transmitter sites at Rampisham, Dorset. This site was originally opened in 1941 during the second world war with four 100kW Marconi SWB18 senders, each of which needed to be looked after by some eight technical operators per shift. Since 1981, about £30 million has been spent in installing eight fully-automatic 500kW transmitters and the complex computer-controlled multi-processor programme-unit, frequency and antenna changing system implemented last year at the two-transmitter site at Hong Kong (see Radio Broadcast, January 1988). With the Rampisham complex operating 24 hour schedules, the major running expense must now be electric power consumption. This costs an astonishing £6 per minute, or some £8500 per day.

Satellite frequencies sought

The European Broadcasting Union has announced that it is still seeking, as a priority objective, the allocation of an exclusive frequency band, from 50 to 100MHz wide in the range from 500MHz to 2GHz, for satellite sound broadcasting. The Administrative Council has recommended that "EBU members should use every means at their disposal to convince their Administrations of the paramount importance of the allocation of sufficient suitable frequencies for satellite sound broadcasting to ensure the unimpeded development of the sound broadcasting services of the national broadcasting organizations to meet developing and future consumer requirements".

An EBU planning exercise has suggested that a band between 50 and 100MHz wide is needed on a world-wide basis to provide with frequency modulation from five to ten programmes per country or, with an advanced digital system, up to 16 stereo programmes per country. The EBU does not consider that s.h.f. bands, such as 12GHz or higher, would be suitable for sound broadcasting reception on portable or car-radio (mobile) receivers but only where listeners used relatively expensive fixed installations. Services in the region of 1GHz could overcome such limitations.

Whether large numbers of listeners could be persuaded to use direct satellite radio is a moot point in view of the many years during which listeners have failed to respond to the "listen on f.m." exhortation. The problem facing the BBC in its current campaign to persuade more f.m. listening, in advance of the proposed phasing out of simulcasting on both a.m. and f.m. on BBC network radio, is underlined by recent IBA audience research which shows that while f.m. reception is now available in some 86% of UK homes, about 50% of listeners claim always to use a.m., 25% f.m., 15% both, with more than 10% indicating that they had no idea which waveband they use.

A.m. stereo in doldrums

Although some 550 American medium-wave stations now transmit a.m. stereo – about 450 using the Motorola C-QUAM system and about 100 using the Kahn-Hazeltine independent sideband system – most listeners remain virtually unaware of, or not interested in, the service. With more a.m. stations turning to talk and news rather than music formats, customer demand seems likely to remain minimal although Motorola claims to have sold about 12 million C-QUAM-only decoder chips.

Leonard Kahn claims that Motorola has blocked the marketing of sets incorporating multi-standard stereo decoders by alleging infringement of a patent that he claims was "illegally obtained". Kahn's allegations are being investigated by the US Patent Office but future prospects for his system seem bleak. At present only Kenwood Electronics is marketing a "multi-standard" receiver range; other firms, including Sony, have now pulled out of the multi-standard market. One of the pioneer users of the Kahn-Hazeltine i.s.b. system, WTIC-AM, Hartford has, rather reluctantly, switched to C-QUAM although its director of engineering insists that the i.s.b. system performs much better under night-time sky-wave conditions than C-QUAM.

In brief

A biography is soon to be published in the United States ("The Genius at Riverhead", by Alberta Wallen) of one of the surviving pioneers of long-distance radio communication: Harold H. Beverage, now 94 years old. He joined General Electric in 1915 on graduating from the University of Maine after deciding to take a $560-per-year engineering job with GE rather than play trombone in a theatre orchestra at night to permit him to continue his post-graduate studies. He feared he would fall asleep during the day-time. With RCA, which split away from GE in 1922, Beverage pioneered the long, low directional wave antenna that still bears his name. In 1927, with Harold Peterson, he found that diversity reception, with spaced antennas, could greatly reduce the effects of fading. He came to London just before D-Day, 6 June, 1944 to direct air support communications and hurriedly persuaded the British that horizontal polarization would be more effective than vertical for communicating from the UK with aircraft over Normandy.

Radio Broadcast is written by Pat Hawker.

RESEARCH NOTES

Silent keys and cancer

Studies attempting to link human diseases with exposure to non-ionizing electromagnetic radiation have been as inconclusive as they have been numerous. For this reason safety limits are set rather arbitrarily in different countries. The problem is quite simply that of extrapolating from experimental results on the very edge of statistical significance. What does it prove for example if three cases of an extremely rare disease all occur in the same street?

To try and harden up the evidence for a possible link between exposure to radio energy and cancer, Samuel Milham Jr of the Washington State Department of Social and Health Services conducted a massive survey of nearly 68 000 US radio amateurs (*American Journal of Epidemiology* vol.127 no 1). The assumption was that such people probably suffer a much greater degree of exposure to high electromagnetic fields than the general public.

Milham investigated a computerized search of public records in the states of Washington and California to discover how many of these amateurs had died between 1971 and 1983 and what diseases they'd succumbed to. Given a reasonably large total number of deaths (2485), he then analysed the causes of death and compared the incidence of each disease with that for the general population in the same two states.

Interestingly enough, radio hams turned out to be a fairly healthy bunch. They had a lower than average incidence of respiratory infection – probably because they tend to smoke less. Cancer rates as a whole were surprisingly similar to those of the general population except in the case of relatively rare cancers of the blood and lymph systems. For leukaemia, for example, the actual number of deaths among the hams was 24 – almost twice the expected figure.

Obviously there's no cause for alarm: 12 extra cases among 68 000 people over a period of 12 years, even if electromagnetic radiation *is* to blame, doesn't make amateur radio a risky hobby. You're much more likely to be killed by flying golf balls.

But even if the latest figures are more statistically significant than hitherto, they still don't necessarily implicate radiation. Radio hams are exposed to several other potentially carcinogenic agents such as soldering fumes, polychlorinated biphenyls etc. So, interesting though Milham's latest study is, there still remains a lot of investigative work to be done before amateur licences carry a government health warning.

. . . and more biological effects of radiation

Evidence such as the above comes rather late in the day if you're trying to discover what, if any, are the precise effects of non-ionizing radiation on biological cells. If such radiation does indeed cause cancer, there's still no clue as to what happens on the cellular level, maybe 20 years before the disease becomes apparent.

Surprisingly the same criticism applies in lesser degree to most of today's bench-top experiments on cell cultures. In nearly all the published investigations, biological samples are studied *after* exposure to the radiation. As a consequence such experiments are only able to detect permanent damage and not any transient, but possibly still significant changes.

A recent project (*Electronics Letters* vol.24 no 7) by Massa and Scaglione at the universities of Naples and Salerno respectively, set out to overcome this deficiency by studying the optical effects on enzyme systems contained entirely within a waveguide. The set-up consisted of a 8.2 – 12.4GHz signal generator, boosted by a t.w.t. and fed into a specially-made section of waveguide designed to hold a 4ml sample. The waveguide was then terminated with a reflectometer.

Two small holes were drilled into the waveguide to permit a beam of monochromatic light to pass through the sample and be measured using a photo-diode. Other parameters such as temperature were also measured and controlled externally.

Initial results using power densities of around 50 – 200 mW/ml showed changes in optical denity corresponding to a lower enzymatic activity, i.e. a lessening of biochemical activity. The results also prove the capability of this type of experimental set-up to distinguish temporary from permanent biological effects of radiation.

The enzymes used in this particular experiment were not intended to reveal any specific temporary effects; only to validate the methodology. Nevertheless it does seem that the Italians have developed a tool that could go a long way to answering many of the mysteries that still surround the biomedical effects of non-ionizing radiation.

113 GHz transistor

A transistor capable of more than 10^{11} switching cycles per second, the fastest ever achieved, has been developed by a team of engineers from Cornell University and the Siemens Research and Technology Laboratories in Princeton, New Jersey.

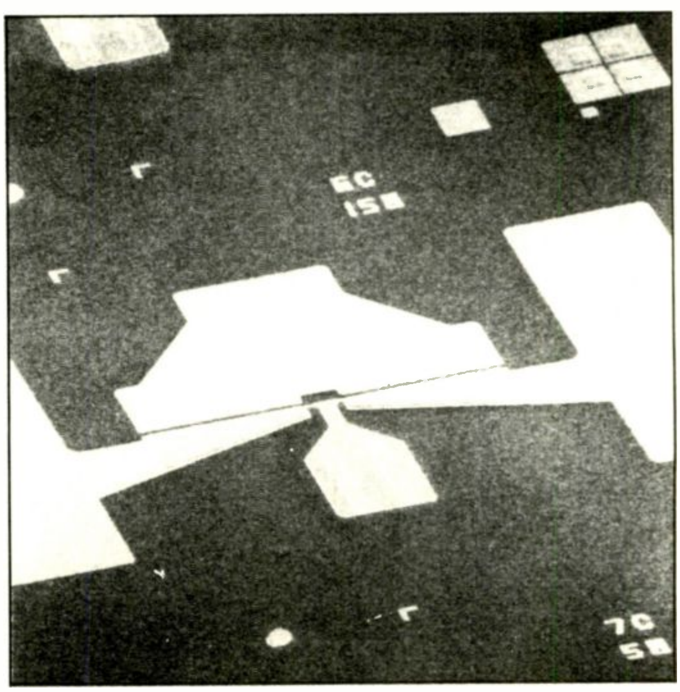
The new transistor, described in *Electronics Letters* (vol.24 no 6), is an improved modulation-doped field-effect transistor (modfet) which consists basically of a multiple-layer sandwich of silicon-doped gallium arsenide and aluminium gallium arsenide. Each layer is a few millionths of a centimetre thick, and the sandwich was created layer by layer using molecular beam epitaxy.

One key to this development is the control of the silicon doping in the layers to achieve the optimum electronic characteristics of the material. The ability to control the silicon doping means that critical regions of the transistor can be made with very few defects or scattering centres, which might otherwise reduce the speed of the electron flow.

Another aspect is the construction of an extremely small gate. This is fabricated using electron beam lithography and is 10^{-7}m wide, about the width of 300 gold atoms.

The Cornell-Siemens transistor represents the first demonstration that a smaller gate size can mean an inherently faster device. Previously, the highest reported speed was 80GHz by engineers at Hewlett-Packard. Computer modelling of the new transistor indicates a potential unity gain cut-off frequency as high as 160GHz.

The next step for the Cornell-Siemens group will be to use



RESEARCH NOTES

their improved processing to construct a modfet that will have the higher power-gain characteristics required of commercial operating devices. Such a further engineered device is likely to include layers of indium gallium arsenide in the sandwich.

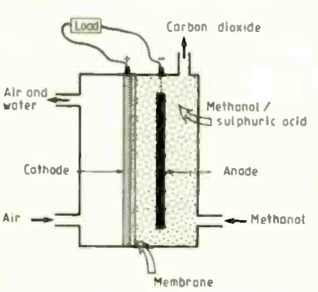
Methanol fuel cells

While many of us would dearly love something that would convert electricity into alcohol, industry is more interested in an efficient device for doing the reverse. Developments reported recently (*Platinum Metals Review* vol. 31 no.4) indicate that such a device may not be far away. The alcohol in this case is methyl alcohol (methanol) and the device a fuel cell.

Fuel cells convert chemical energy into electricity without any intermediate steps and have found many applications where a compact high energy power source is needed, for example on space vehicles. For such purposes the chemical energy is stored as oxygen and hydrogen and the waste product is water which can be used for drinking.

For more down-to-earth used compressed hydrogen is inconvenient, even if the oxygen can be obtained freely from the air. Fuels like methanol which are liquid at ordinary temperatures would be much more convenient if they could be used instead of hydrogen. Fuel cells based on methanol would have potential applications in portable high-current devices and for vehicle propulsion.

The main problem in the 25 years during which they have been researched is the catalyst that forms the anode and on which the methanol is oxidized. Several systems have been developed by organizations such as Shell, Esso, Hitachi and by the US Army and different catalysts have shown varying levels of efficiency. A platinum/ruthenium one has proved among the most active.



Devising suitable catalysts has proved something of a black art because it is by no means clear how they work. Several different theories now exist on the precise details of the electro-oxidation mechanism.

This and other difficulties such as the practical construction of the electrodes are now the subject of a collaborative research project involving organizations in France, Germany, Ireland and the UK. Initiated by the Commission of the European Communities, the programme is due for completion in late 1989.

Measuring 100 million degrees

A system for computing real-time temperature profiles of the plasma within a nuclear fusion reactor has been developed by a group at ERA Technology Ltd in Leatherhead. It's intended for use at the JET (Joint European Torus) facility at Abingdon, where temperatures approaching 10^8 degrees Celsius are being created as part of the effort towards the long-term goal of generating energy by the fusion of hydrogen atoms.

In the JET experiments it is vital to have accurate information on the electron temperatures across the plasma in which the fusion reactions take place. This is important because plasma instabilities can occur on the scale of hundredths of a second and can, in certain situations, damage the plant. The object of the research at ERA has therefore been to compute complete temperature profiles in 10^{-2}s. These can then be used either to generate shut-down signals or to produce graphic displays for later analysis.

Surprising as it may seem, the easiest bit of the whole operation is the sensing element. Temperature information is derived from millimetric electromagnetic emissions from the plasma. The algorithms required to transform this raw data into temperature profiles are, however, complex and demand considerable computation.

In order to make best use of the computing hardware, ERA has developed algorithms optimized for high-speed parallel operation, including a technique for calculating a 4096-point discrete Fourier transform in less than 3ms. Computer-based simulation has shown that the system operates effectively, particularly when presented with noisy or defective data.

ERA Technology Ltd believes that the benefits of high-speed real-time analysis are applicable not only to advanced projects like JET, but to a whole variety of scientific and industrial operations where complex changes take place rapidly.

Meanwhile. . .

Casual readers of this column may have got the impression (*mea culpa*) that all research is a highly motivated and thoroughly down-to-earth search for exploitable knowledge. Indeed a quick scan of the national dailies may even dupe the unwary into believing that all worthwhile work is directed to solid commercial results at the earliest opportunity. At the risk of accelerating the brain drain, I've collected a few examples of research to show that adventurousness, original thinking and sheer uncommercial curiosity are still alive, albeit overseas.

First in this category, as an example of brilliant lateral thinking, is a scheme devised by the US military. They're going to build battle tanks out of uranium. . .

This clever idea (an improvement on shooting your own feet) is the result of years of behind-the-scenes research by the Pentagon into new materials that will survive the heat of the battle. True it *is* ^{238}U, it *is* the densest material available and it *is* pretty good at absorbing the odd stray neutron. But what I wonder would happen if a neutron bomb went off nearby? (Take another look at the theory of fission-fusion-fission weapons).

If that idea might just appeal to the British government's notion of worthwhile pursuits, what about this next achievement from the Houston Area Research Centre in Woodlands, Texas? Walter N. Colquitt (*Science News* vol.133 no 6) has just discovered the third largest Mersenne Prime Number. This consolation prize-winning contribution to mankind's thirst for knowledge is the result of 11 minutes of number-crunching by a NEC SX-2 supercomputer.

Mersenne primes are expressed in the form 2^p-1 where p is itself a prime number, though of course not all Mersenne numbers are prime. They do however offer something of a short cut in the search for truly huge primes. Up till now searches have covered exhaustively every potential Mersenne prime up to $2^{103\,000}-1$ and have found 30 of them.

Colquitt and his colleagues wrote a Fortran programme in the hope of finding a record-breaking prime: bigger, that is, than $2^{216\,091}-1$ (all 65 050 digits!). What they ended up doing was finding one that other workers had overlooked . . . its exponent a mere 110 503. But please don't ask me to print it — the Editor might conceivably have better use for three whole pages of the journal.

Finally some evidence that *perestroika* has done nothing to halt the progress of good fundamental research in the USSR. You only have to scan these columns to discover gems like electronic bird-scarers, the passionately serious search for extra-terrestrial intelligence or attempts to create element 111. But were I the Soviet equivalent of Senator William Proxmire, my 'Golden Fleece-ski' award would go without question to the recently-formed Yeti Research Group led by Dr Pavel Belenitsky. Sponsored by the Soviet Ministry of Culture and inspired by a recent sighting of the Abominable Snowman, the indomitable Dr Belenitsky is about to set off on an expedition to discover more about the creature's natural habits and psychology . . . Shy and retiring, at a guess.

Research Notes is compiled by John Wilson of the BBC External Services science unit at Bush House, London.

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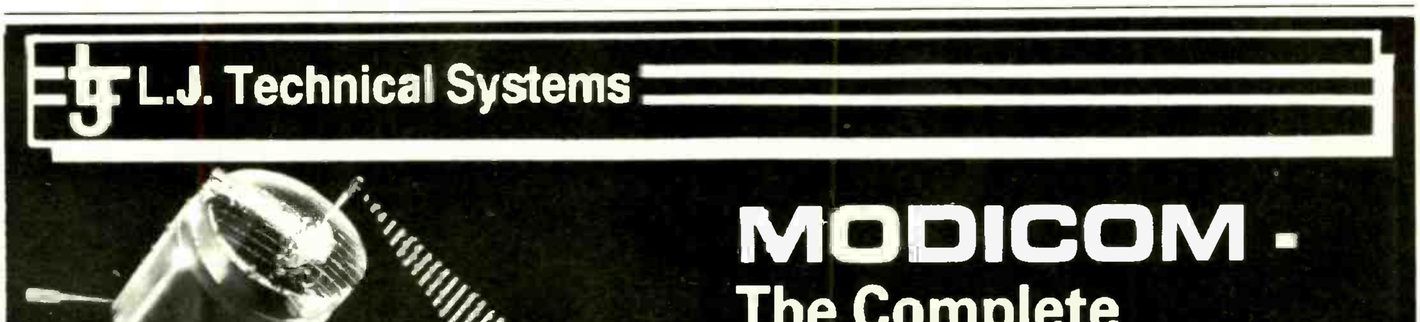

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