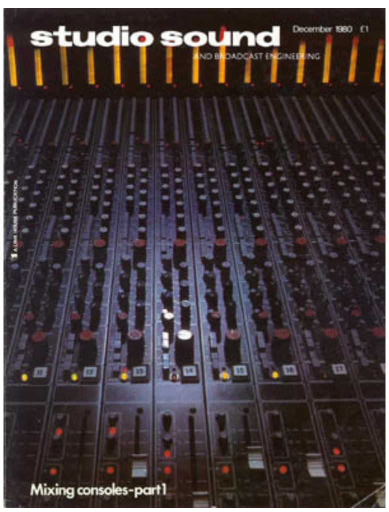
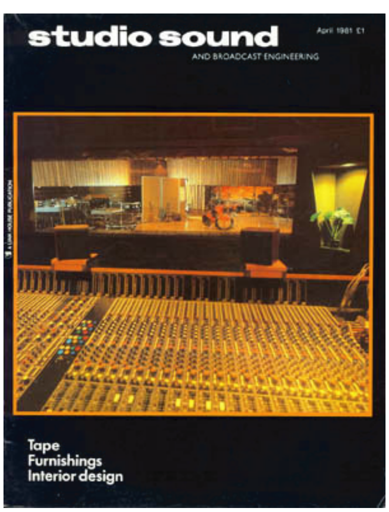
# Designing a Professional Mixing Console

By Steve Dove

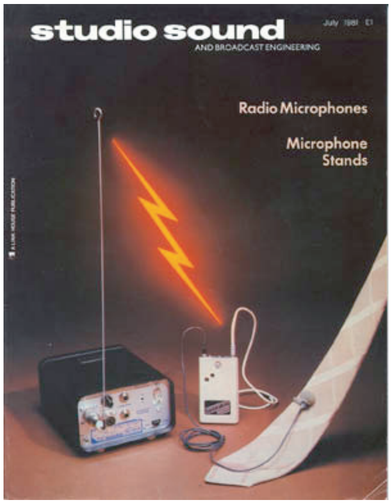




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# Designing a professional mixing console

Steve Dove

## Part One ~ Introduction and Recording/PA Console Description

**This series of articles was written in the hope of explaining in relatively straightforward and non-mathematical terms, the processes involved in the conception and design (both systems and electronic) of cost-effective consoles to today's upper-bracket commercial standard. Along the way a lot of ill-founded mystique about what goes on under the knobs will be attacked mercilessly, and a few hypotheses as to the future direction of audio control thinking will be mooted. Please see the editorial on Page 3 for further comments on the series.**

AS PART of this project, a considerable amount of electronic design culminating in the construction of current professional standard modules was undertaken. Circuitry of these modules and much else is to be published in subsequent issues. In their most advanced form they exemplify the kind of circuitry found in consoles that are evolving toward complete programmability of all functions and control statii used in the mixdown process. This is primarily to give an insight into how the latest brand of 'magic' can be implemented. A discussion of how a console based around these modules can be used in conjunction with a proprietary automation system—in this case the Valley People's *Fadex* system—is dealt with in depth and that particular issue will probably be of use and interest to current owners of a *Fadex* or similar system based upon the Allison Research *65K Programmer*.

A simplified but no less well performing module based on a good, solid conventional 'buttons and knobs' format is described and this particular design (perhaps more than the 'all-singing' one) may be of interest to constructors, being that it was originally designed for a commercial mixer and hence is very cost-effective and easily producable. The card layout and terminations are such that they may be single front panel mounted *or* channel mounted, dependent on your willingness to play at metalworking!

Two distinct considerations interplay in determining the ability of a console to fulfil a given application. These two, the system and the electronics, have entirely differing parameters which need to be refined, but are nevertheless completely indivisible. The design approach taken in this instance was to devise an operationally workable (many aren't) multitrack recording and monitoring channel based upon the experiences of studios. Sufficient access is made to all the individual elements within the channel to enable them to be reconfigured to the extent of allowing establishment of almost any of today's conventional recording console formats.

The electronics, as much as being designed to perform the required functions, have been very carefully designed *not* to be a major influence on the 'sound' of the console—most causes of sonic disturbance can be attributed or predicted and these are mitigated in design, with *still* dubious circuit configurations avoided altogether. To the shock of some purists, commonly available integrated circuit operational amplifiers are used throughout, the reasons why (other than the obvious convenience) together with the reasons they acquired a bad reputation are treated in depth later in this series.

Operational amplifiers, known to their friends as op-amps, have in recent years revolutionised the concepts and systems capability of full performance audio consoles. Allowing system elements to be thought of, designed and implemented as building bricks, simplifies matters considerably but also entertains the valid criticism that console design can be relegated to 'do it by numbers' status. Fortunately, device idiosyncracies, subtleties and the entirely separate science of getting heaps of individual system elements to behave successfully as a total console, prevent design falling into the precincts of bureaucrat, marketing persons and other genres noted for inappropriate motives, insensitivity and general idiocy. This leaves it largely, at least for now, still in the hands of the people who know and care; intelligent user engineers and the tiny number of sympathetic electronic engineers and manufacturers.

### Manufacturers

Fortunately for the industry, a very large proportion of current console manufacturers started off in life as small bunches of studio engineers furtively constructing a mixer for their own ends in a garden shed or the managing director's loft—grass roots system design owing everything to immediate operational needs. Continuing in this vein in production, listening to and, most importantly, relating to customer requirements because they've played this game for themselves. Take now the few notable cases, no names mentioned, on both sides of the Atlantic of large prestigious manufacturers in which the system design people are career-jockeys and the electronic engineers probably haven't even set foot in a recording studio. A caricature, maybe, but not a million miles from the truth. The product, as beautifully made as it may be, probably had the maintenance people at the studios who took delivery of the first couple sweating nights to iron out the system gaffes and inadequacies.

The worst cases, though, are the 'rubber-stamp' console manufacturers who break out into a shifty-eyed sweat at the mention of anything other than a 'standard console'. Perfectly reasonable if you're selling 8 into 2s, but 40 by 32s? They, unfortunately are also the people who by the large quantity of product they place through carefully manipulating sales techniques, create a customer expectation climate facilitating ever increasing technological, hence monetary, inflation. It is a business. It has increasingly little to do with recording music.

Retaliation is partially what these articles are about—if as a result of reading them you feel more knowledgeable and better equipped to understand more fully what mixers are really about and not have to rely on sales guff, pretty pictures and hearsay as much, a worthwhile strike has been made. Even if you don't reach the extreme of constructing your own console, at least you will be more in tune with manufacturers' thinking and stand a better chance of finding one who cares as much about mixers as profiting from them.

### History

Once upon a not so distant time, systems didn't exist. Mixers as such didn't exist. All the bits of electronics used in the control room sat there with all their inputs and outputs accessible, by way of a jackfield if you were prosperous, or by small screwdriver and sore knees if you weren't.

Mixing sources was accomplished by directly paralleling amplifier outputs (possible because all the old valved gear had a finite and predictable output impedance usually arranged to be a conventional balanced 600Ω) and either hoping or arranging that the destination had enough gain in hand to make up the accrued loss. Crude as that may seem today, from an engineering viewpoint it has a sheen of pure elegance. An amplifier was just that—a box that had balanced 600Ω source and termination impedances, maybe an alternative 'bridging' (>10kΩ) input term, a selectable amount of gain, and, of universal application from mic-amp through mix-amp to headphone amp. If you wanted to do more things, you got more boxes. Equalisers and limiters, a treasured few if there were any, were similarly universally applicable. Variable level control was again by true balanced 600Ω source and termination via studded rotary attenuators. The utter beauty of the systemless studio was that anything could go to anywhere via anything else and be mixed or distributed at any point on the way.

Soon enough, amplifiers were hard-wired to attenuators and designated specifically 'microphone amplifier' or whatever—a system had been created. Some of these together with a mixing gain make-up amplifier were thrown in a box. The mixer was born.

Its been downhill all the way since, with ever-increasing numbers of system elements being tied together in increasingly circumloquacious manners in order to maintain some kind of flexibility—a system can be defined as a means of reducing the ultimate versatility of its constituent parts.

Once a 'mixer' was accepted as a system element itself, the rot set in further. There was no need to provide for connection of its internal interconnections to the outside world so (conveniently) the balancing transformers disappeared, and more economic alternatives to the stud attenuators operating at more convenient internal impedances evolved. By a more positive token, the electronics were becoming gradually optimised for specific functions to which they were designated, mic-amp, mix-amp or whatever. (The question nags whether a universal amplifier, by now all but obsolete, could be optimised for all the varying requirements.) Still, at least all the inputs and outputs of the mixer were still conventional. This held true until the slow demise of valves in professional audio.

Transistors were justifiably unpopular for a long time because of the numerous limitations they placed upon design. Headroom was severely limited because of the low rail voltages that could be applied to the early devices, they were noisy, the lower operating impedances and differing modes to valves took some getting used to and when they clipped, they actually clipped, rather than the graceful 'bending' people had known, and frequently taken advantage of, characteristic to valves. In order to realise a reasonably low stage distortion, many transistors in compound configurations using heavy amounts of negative feedback were used—a far cry from a single valve stage operating virtually open with little feedback. This gave rise to a peculiar phenomenon that sounded as if it hailed from science fiction—zero impedance.

It was possible by virtue of the mechanism of the heavy negative voltage feedback employed to render an amplifier's output insensitive (in terms of output voltage) to varying load impedances—obviously within the current handling capabilities of the output structure. Goodbye termination problems with the worry of compensating in level for differing load hook-ups. With the exception of long-line feeds, 600Ω terms were dead. High level balanced inputs were now almost exclusively 'bridging'. For better or worse, it has become the conventional studio interconnection technology.

It has taken until fairly recently for an accepted distinction and separate level specifications for the two technologies to be accepted.

The original transmission line level specification referred to a power level of 1mW at whatever the impedance was. It was a universal specification applicable to any signal of any frequency being transmitted along any bit of wire for any purpose at any rated impedance and is used extensively in radio-frequency work and other things entirely unrelated to audio—the dBm definition is sacred and can't be changed just because it doesn't suit us anymore. Zero dBm in a 600Ω load works out at about 0.775V rms this having also been adopted as the reference for use in general audio work. With zero impedance technology, although the working voltage is specified, the impedance varies so the power varies; 0.775V rms across say a 100Ω load works out at +7.78dBm, whilst across 10kΩ it would be −12.22dBm. Confusing to the point of insanity.

The reference level for zero impedance thinking is a voltage, and the one chosen is that familiar 0.775V rms that everyone was used to dealing with. That voltage is distinguished as 0dBu. Some lunatics have tried to impose a new universal reference for audio, based around a voltage level of 1V, called the dBV, which made some sums nice and easy, looked neat and proved confusing to anyone brought up on the dBm. Adding 2.2dB to everything was a dreadful bore. Or was it subtracting?

## Monitoring

With the exception of disc mastering suites, most professional audio work ends up on magnetic tape, the replay of which is required often simultaneous to its recording. Source/return or A/B monitoring is as crucial a subsystem as the live recording chain itself. Until the advent of multitrack, monitoring was a fairly straightforward business consisting of, in essence, a switch that fed the monitoring chain from the desk output, the outputs of however many machines you had, the Light Programme or whatever. It was totally passive as far as the recording chain was concerned since any overdubs took place whilst the appropriate machine was being replayed through the recording chain and being mixed with the additional source(s). With little variation, this technique was used extensively in 2-track also, the final master representing only the first generation of the last overlay. (In retrospect, that is an advantage over contemporary multitracking where the master is at best the second generation of everything.)

Initially, as the number of tracks per machine increased so did the number of mixer groups correspondingly. Each group had its own A/B switch relating to that track output and the associated machine return, with a level and pan control feeding an altogether separate stereo mixing buss from the recording chain. This independent stereo mix appeared as another source on the main monitor selector. This, alas, was insufficient. Foldback pre-fade mix feeds became no longer a luxury but a necessity, since the desk stereo output or a derivation thereof could no longer be relied upon to be roughly what the artiste needed to hear—there was no proper desk stereo output at any time other than mixdown. In order not to clog up the input channels appropriate to the multitrack returns just so that pre-recorded tracks could be made accessible to the foldback busses, foldback feeds were added to the monitor system on each group. Effect sends also, just to let the monitoring sound pretty.

The monster had split itself amoeba-like into two entirely separate signal processing systems, with the curious situation that the mix used for monitoring during the original multitrack recording had to be transferred over to another system entirely some time for mixdown.

Perhaps the first major rationalisation (which occurred long after many 'X' input, 24 group, 24 monitoring consoles had been made) was a result of the realisation that you don't actually need 24 group faders sitting there full up, collecting coke and fag-ash. This instantly avoided a normally unwanted gain variable stage in the signal path which, if maladjusted, could upset noise or headroom performance.

A much smaller number of stereo mixing subgroups which could again be routed to any of the multitracks together with the individual channel outputs proved easily as flexible. But still there was duplication of monitor busses and main stereo mixing busses both with their attendant effects and foldback feeds—rarely being used simultaneously. At last the dawning of realisation that the pair, monitoring and stereo mastering busses, could be one and the same thing.

In-line monitoring recording systems had come to fitful fruition.

A potted 'action replay' of console evolution is an impossibility—well, maybe not an impossibility but it would make an excellent basis for a comedy series. We all have to be thankful for the cranks and visionaries along the way (often the same) who have manipulated or shocked the industry into grudgingly lurching back into step with technology's capability. These milestones represent significant plateaux of thinking that point the way to today's console concepts.

The designs published in this series were evolved around a full function in-line monitoring recording system, together with all its attendant frantic system juggling to make it operationally feasible. Each, or any, of the system 'modes' almost certainly is directly appropriate to other mixer formats and conventions, so little flexibility is compromised as a result of this approach.

## Applications

Four distinct system requirements will be considered in order to demonstrate how the basic elements provided can be rearranged to suit the specific differing needs:—

**(a)** a full scale 32-track capable multitrack console;
**(b)** a (nominally) 12 input, 8 group system optimised for smaller scale multitracking;
**(c)** a (nominally) 32 input, 8 subgroup, stereo output mixer intended for large-scale PA;
**(d)** an on-air broadcast control console.

Broadcast technology traditionally owes little or nothing to recording, consequentially it will be regarded separately, although crossbreeding (both ways) has come up with interesting approaches to some critical applications, notably in monitoring.

The full, completely stacked, bells, whistles and foghorns (all transients faithfully reproduced) in-line monitoring and main signal path modular system is used in entirety on the 32-track capable system. As the complete module contains all the system elements that would be used in other module functions (such as stand alone groups, effect returns, or subgroups with subsidiary mixing and routing) all modules in the recording class utilise the same basic module designs with unrequired bits and controls left off or relabelled as needed. This 'bits left off' thinking as well as being superbly efficient for the scheduling and manufacture of a console, additionally is ergonomically delightful in operation—similarly acting functions on differing module types are found in just the same physical location, reducing 'knob-grovel' (searching for control functions) greatly.

The necessary trade-off between control density (ie cramming as many knobs into as small a space as possible) and ease of operation is simply resolved, if the designer has ever had to use one of his own creations!

Exceptions to this 'identical module/variable format' concept are obviously monitoring control for studio and control room, studio communications (talkback) and master foldback/effects send controls. Multi-destination routing, whether it be of control of the 32-track electronic switching matrix or of local conventional 8-group switching obviously differ and are (usually, but not necessarily always) mutually redundant, but both are allowed for within the concept of the universal modules.

Absentees from convention in-

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# Mixing console

clude a dedicated multi-input effects return module, since in nearly every studio with which the author has had dealings, they are disused in favour of additional full-function input channels. In fact, this is a most often quoted rationale for having an excess of channels over the size of the multitrack recorder in use. Admittedly, this absence could be a hang-up in a mobile recording situation where inevitably all the input channels get used up simultaneously on 'real' sources, the effect returns (unless it is a 'hot' to disc, 2-track or broadcast) almost always just end up as a guide in the monitoring. A way around this is to bring the effect returns back up into the 'machine return' inputs on channels not appropriate to the multitrack—those channels will in order to be recorded have to be routed elsewhere anyway, leaving their 'B' monitor chains free. Access in the monitoring module is made to the stereo buss for external extension of the monitoring capability for this, or any purpose in any case.

## The channel (fig 1)

Three source input stages exist on the module intended as front ends for the main signal path and multitrack machine send and return monitor paths. The microphone amplifier which may be gain reduced and padded to act as a line-input amplifier if need arises, heads the main signal path, fixed-level electronically balanced differential inputs acting as the machine send and return (or 'A' and 'B') inputs to the monitoring chain. No level adjustment is available to these stages as monitoring returns since the levels returning from the multitrack are (!) conventional and consistent. The microphone amplifier is gain adjustable sufficiently to enable most common microphone types to be used for most given circumstances.

Around the second stage of this amplifier is a gain-reduction element for limiting coupled from a peak-detector side chain. The detector is selectable to sense either pre-equaliser (ie post mic-amp and highpass filter) or after the post-equaliser insert point (in order to catch any extra level incurred during eq or inserts). The detector level is switchable between 'clipping' (2dB before supply rails) or an operational level, nominally +8dBu but tweekable up or down. Whether or not reaching the selected level activates the limiter is a switchable choice, but an indicating LED shows regardless — so a choice exists whether to use the limiter as protection, an operational effect or not at all, but still keeps a useful indication of channel level status or impending clipping.

The switch shown on the main module block diagram following the machine return input amplifier and mic preamp is part of the module status switchery (which will be described fully in due course). This particular switch disables the microphone preamp and selects the machine return into the main signal path in the 'mixdown' mode—incidentally this switch and all the others concerned with status or routing may be electronic or mechanical, dependent on which design is utilised, but for the purpose of the block diagram conventional switch symbols are shown.

Following the second stage of the main path is a variable highpass filter of second-order response with an ultimate 12dB/octave roll off with the turnover frequency adjustable between 20 and 250Hz or bypassable by means of an end-stop switch on the control. The output of this is a line-amp capable of feeding any normal studio-type load in the case that the pre-equaliser insert point is selected and used, whilst the input to the equaliser section is a ground-free electronic differential input to simplify potential ground-path problems at this insert point. Similarly the return from the post-equaliser break point encounters a differential input.

Equalisation should at this point just be regarded as a 'black box' since again its specifics vary with the design used, being discussed in a subsequent article in this series.

The two basic variants of the equaliser both contain high and low frequency shelving of selectable turnover frequencies, with the lf shelf curve selectable to a 'bell' shape—otherwise meaning that the response falls back to unity below the turnover frequency selected. In addition either one or three sections of 'parametric type' curve generation are present, each with variable centre frequency, resonance sharpness (Q) and differential level with respect to unity gain. The output of the equaliser is again line-drive capable for the purpose of the post-equaliser insert point.

In the module's basic form, no provision is made for transformer balancing the break point outputs. The assumption is made that it is unlikely an insert point is likely to be required to feed anything beyond the confines of the control room and little untoward can happen to a high level, low impedance (if unbalanced) signal given that constraint.

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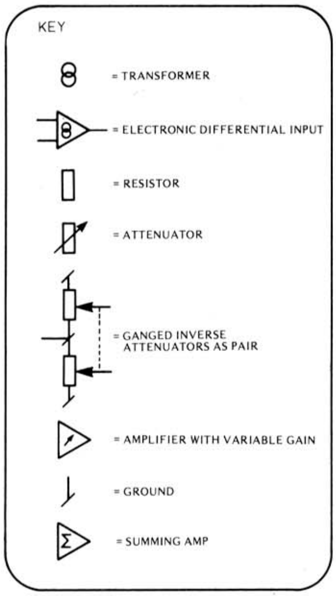
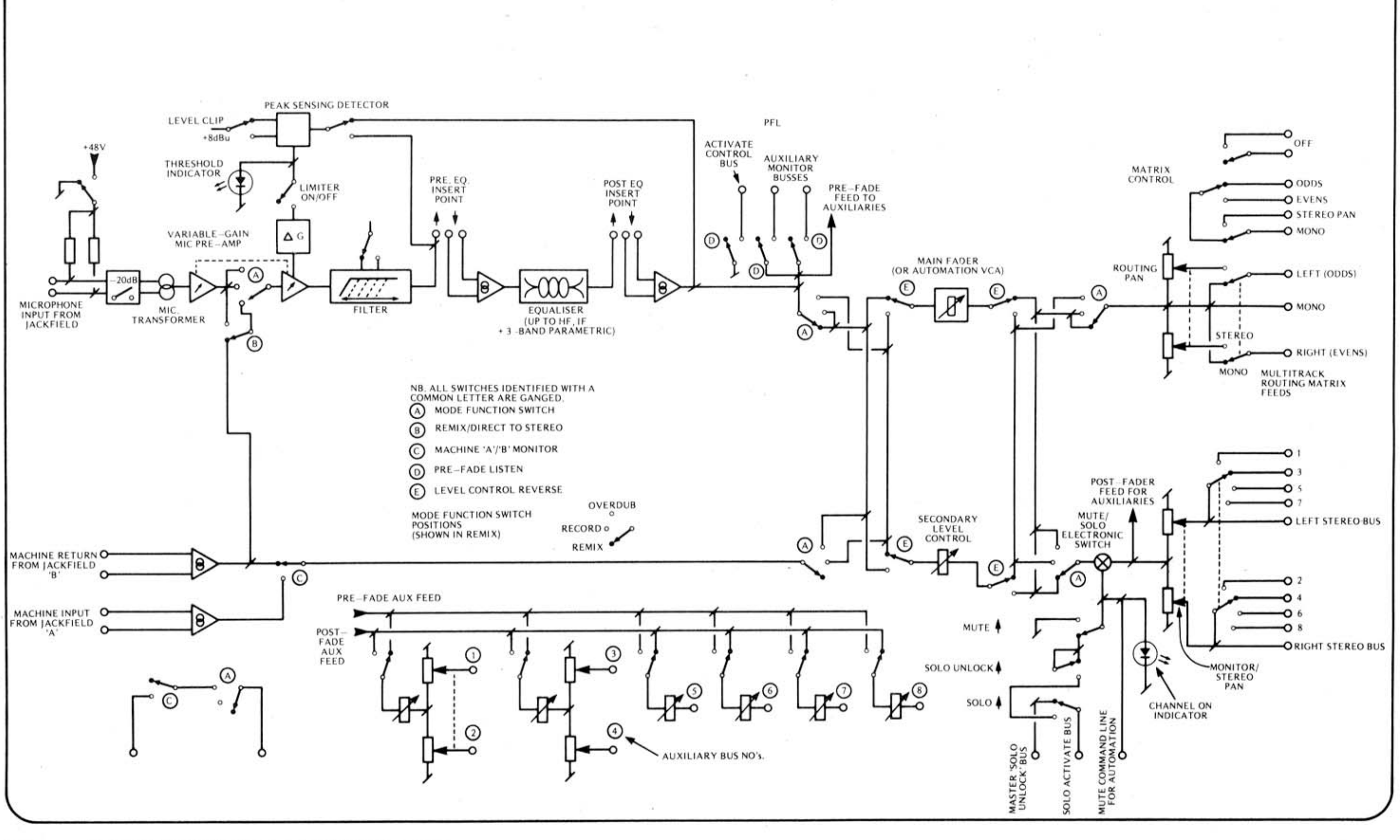



FIG.1 BASIC SCHEMATIC, IN-LINE MULTITRACK RECORDING AND MONITORING CHANNEL



# Mixing console

Whatever is being fed has its electronic ground referred to the mixer whilst the return, whether balanced or unbalanced, faces a ground free differential input eliminating grounding problems at least within that particular loop. There is easily sufficient common mode voltage swing capability in the differential input to cope with anything that may occur under normal operational circumstances.

Both the input and output of the equaliser section are available on the module card connector allowing, if required, full jackfield pre- and post-eq insert points. Under normal conditions only the post-eq position is necessarily available for a variety of reasons.

The greatest application for a pre-eq break point is to insert a limiter on 'untidy' sources or those with a high peak to mean level difference. There is a limiter available that can be wrapped around the mic-amp built into the channel. Should any access be required pre-eq during a tape replay mode, it already exists in the form of 'tape machine return/channel input' jackfield normalled/broken access and insert points.

The facility is present not only to fulfil this possible requirement, but to contribute to the systems' versatility as a whole.

FIG2a-c SKETCH SCHEMATIC OF THE DIFFERING MODES OF IN LINE CHANNEL

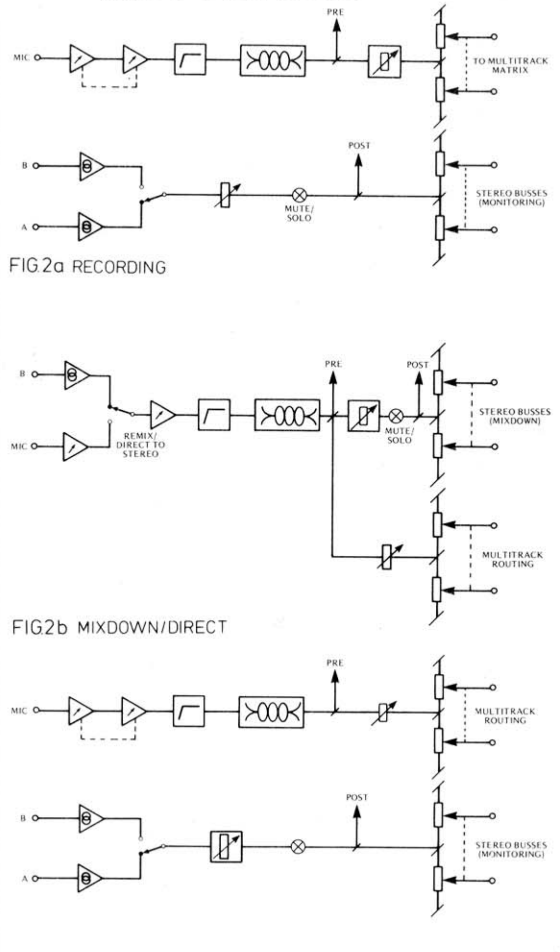



FIG2a RECORDING

FIG2b MIXDOWN/DIRECT

FIG 2c OVERDUB

## Various modes

Unless you possess a mind warped in similar fashion to a railway enthusiast's, the status switchery following the post-equaliser break point will seem totally unfollowable. Logic diagrams are like that. To make life simpler, further sketch system drawings of the blocks arranged in the three main operating modes are shown in **fig 2.**

**Fig 2a** shows the 'record' mode used when the console's immediate function is principally recording lots of live sources simultaneously. The microphone/line input is accessible to all the response and dynamics modifying circuitry in the channel as well as being accessible externally through insert points. It passes through the main fader (or VCA if automated) and through to the multitrack routing matrix. The monitoring stereo busses are fed via their panpot from the secondary level control which is sourced from the multitrack input and return ('A' and 'B' switch) appropriate to that channel. All the usual monitoring functions (mute, solo) are available on this chain. PFL, though, is taken from the main signal path.

During 'mixdown', **(fig 2b)**, the machine return is applied to the main signal path and is mixed onto the stereo busses via the main (or VCA) fader, whilst multitrack routing is still accessible for versatility's sake, through the secondary level control. Very useful if you run out of effects sends! A function very closely related to mixdown is 'direct', shorthand for direct to stereo. The microphone preamp is recoupled to the main chain, everything else remaining the same as in 'mixdown', enabling direct live multisource mixes onto the stereo busses without having to access the multitrack routing.

'Overdub', **fig 2c**, is a half-way house between 'record' and 'mixdown'. This mode would be entered when most of the console is in 'mixdown' status, but individual tracks are still being laid or touched up. It is identical to 'record' with the exception that the main (VCA) fader is on the stereo busses feed in order to match functions with all the other channels which would be selected to mixdown (with all their main faders feeding the stereo busses). The ability to have all the main (VCA) faders on stereo mixdown whilst recording is still in progress means that the engineer can get a feel of the final mix and even start constructing sequences on the automation system during the 75th synthesiser overdub or vocal attempt. Two jobs for the price of one. If for any reason, valid or dumb, it is felt necessary to reverse the relative positions in the system of the main and secondary level controls, a button doing just that is available without otherwise upsetting the signal paths appropriate to the selected status. It is a 'local' reverse in that it still reverses the controls whether the channel status is defined by a console 'master' mode (in the case of the full electronic switching design) or by a channel command.

## PFL and solo

Immediately following the post-eq insert diff-amp is the take off for the 'pre-fade listen' feed which, upon activation, sends the signal onto auxiliary stereo monitoring busses whilst the 'PFL activate' buss is simultaneously pulled, causing those busses to override whatever else may be selected on the main monitoring module. This does not interrupt at all any signal paths, other than the monitor speaker (or headphone) feeds—hence it is described as 'non-destructive' channel monitoring.

The 'solo' facility though, is 'destructive'. It might seem a little in the sledgehammer-to-crack-a-walnut vein, but depressing a solo button mutes every other source feeding the main stereo mixing buss, leaving just that particular channel present at whatever level and panned stereo position it originally held in the mix. A refinement to this is the 'solo-unlock' button, which keeps any channel upon which it is depressed 'open' despite the presence of an active muting control voltage on the solo buss. This is especially useful for channels utilised as effect returns since it is then possible to monitor in 'solo' any channel with any effects in use with that channel, all at their relative levels and in-place stereo. Since effect sends are generally fed from post-fader feeds, all those feeds on channels other than the one in 'solo' will be obviously muted also, thus leaving all the effect sends free of extraneous clutter.

Note: the individual channel 'mute' function uses the same electronic switch as the 'solo', but when the channel 'mute' is depressed the channel stays muted regardless of a 'solo' or any other function. A control line is taken from the 'mute' as a facility to either instruct or receive commands from an automation system, if one is in use.

A 'solo' only interrupts feeds to the main stereo busses (which are the monitor busses in 'record' and 'overdub' modes and the main console stereo mixing busses during 'mixdown') leaving the feeds to the multitrack routing matrix intact.

An interesting subtlety in this particular area of the system is that the prefader auxiliary feeds (which are usually used for studio foldback monitoring feeds) are taken from before the solo/mute switching. This prevents musicians using these cue-feeds thinking they've gone deaf suddenly, just because somebody in the control room has hit a 'solo' for their own benefit. An incongruity avoided.

A deliberate choice was made not to transfer the channel in solo onto the auxiliary monitoring busses (such as is the PFL) although it would mean the 'solo' could then become similarly non-destructive. The design problems involved in then isolating all the other effect feeds in order that the returned effects into the monitoring were solely those appropriate to the solo channel would be very, very messy indeed. Even then, unless a completely duplicate set of effects were in use solely for this facility, it would mean 'robbing' the effects from the mix that you were attempting not to interfere with in the first place! Conclusion: why bother with a separate solo buss? Essential non-destructive channel monitoring is why PFL is provided.

It is not beyond the 'ken' of most people to realise that a destructive 'solo' is potentially lethal. Other than crashing a mix half-way through because an idiot producer decided he wanted to solo something, the most heart-strain inducing possibilities are in live PA. Now, a 'solo' function is a real boon during a PA soundcheck—but trying to work out why at the beginning of a show all you can hear is a MW kick-drum—no fun in the slightest! For these reasons, two approaches are taken here. Firstly, when a 'solo' is hit all the channels that are consequently muted lose their 'channel on' LED indicators. Secondly, a console master 'solo disable' is incorporated, which is essentially a master of the individual channels' 'solo unlock' function, for use during those delicate moments.

## Recorder control and routing

Other obscure switchery can be seen littering the periphery of the main system diagram. Adjacent to the machine 'A' and 'B' selector is a pair of switches—one ganged with that selector, and one ganged with the status switching, having continuity in 'overdub' mode. The object of this is to create a closing loop of contacts when the channel is in overdub and the monitor selector is switched to 'A' (as one has to when the actual recording part of an overdub commences). This closing loop may be interfaced with the multitrack recorder's record enable circuitry appropriate to that channel, hence allowing (provided the machine is primed) 'one-button' drop-in as soon as the 'A/B' selector is hit to 'A' and dropout (?) when it is returned to 'B'. An admitted disadvantage to this particular method is that it doesn't take into account the possible routing of the main channel path to a track other than the one to which its monitoring is related—that would require a subsidiary switching system to the routing matrix, which is another design exercise altogether. It is a relatively unimportant consideration since as a function it is most likely to be useful once the majority of the basic tracks are recorded, defined and being used with their appropriate monitoring chains.

Ganged to the mono/stereo switch in the multitrack matrix feeds are off/stereo/odds/evens switch control lines for steering the matrix logic—the 'off' functions being the mute facility for this feed. It is kept separate from the stereo busses' mute since it would be quite a shame to stop recording something just because you didn't want to hear it in the monitors for some reason.

A variety of formats for switching matrices will be described blow-by-blow at a later juncture, but for now there are principally two arrangements for the purposes of system description.

The stereo/mono channel outputs together with the logic control lines for 16 pairs of odds/evens or stereo feeds and not least the on/off commands are intended to feed a specific matrix card constructed around two 1 of 16 analogue multiplexer ICs. Whilst it may seem like a limitation only being able to access any two tracks at a given time, the author has yet to hear any heart-rendingly convincing arguments for doing otherwise. An alternative to this card, populated with individually controllable switching elements can be arranged to be free-access, but frankly some of the great advantages that these matrices convey—including simplified control, less panel space taken up with switches, and less money spent on same—are thrown away by doing so.

The multidevice card has an advantage that it may be populated for only as many groups as required, say 16, 20, 24 or 32. Both the multidevice and the large-multiplexer cards may be configured also into 1 into 32 switchers which, as will be discussed, is useful and advantageous in many system solutions, whilst the control functions and actual routing information on each card can be made to appear as a functional single 8-bit memory location for parallel data buss microprocessor control applications.

An overlap between differing system philosophies is apparent when the feeds to the multitrack matrix and the main stereo busses are scrutinised in **fig 1**. Hanging off the stereo busses' output is a pair of 4-way switches, intended as basic 8-track routing, which would not be needed if a stereo matrix routing system were in use. Similarly, the stereo matrix feeds would not be necessary in a system where the stereo 8-track routing were being employed. Taking this into account, the system PC card and metalwork are designed to house either, but not both (at least not without an uncomfortable squeeze). An interesting crossbreed between the two may be handy for those who need matrix routing facilities (for say 24 or 32 sources) but are unlikely to be using an automation system with its inherent subgrouping capabilities. By using a 1 into 32 (say) matrix card from the 'un-panned' matrix feed (permitting channel-to-any-track grouping) and having eight subgroups driven off the stereo busses' output, each with further 1 to 32 matrix cards fed from them, a useful alternative evolves. Reduction of any number of sources (either live or during mixdown) may be made across up to four pairs of stereo tracks on the multitrack by using the re-routing on eight sub-groups. Each of the subgroups of course may be panned as another source in its own right back into any of the other subgroups and thence again across any pair on the multitrack. This system of consolidating subgroups into a defined physical area of the console is, if anything, simpler to use and more identifiable operationally than a totally stereo freegrouping system and almost certainly better suited to the more panic-fraught operational circumstances of live recording and broadcasting.

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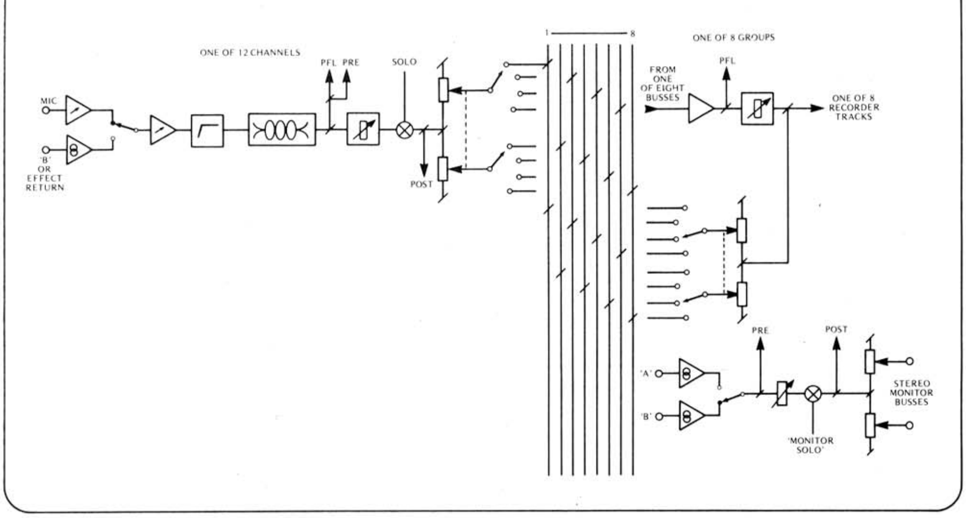
# Mixing console

FIG.3 SKETCH SCHEMATIC OF 8-TRACK FORMAT




## Auxiliary feeds

Auxiliary pre- and post-fader feeds again vary in configuration with opted design, the largest allowance being four single channel feeds (normally fed from post-fader for effects send use but selectable to pre- if required) and two separate stereo panable (another four feeds), normally fed from pre-fader for studio foldback or auxiliary stereo mixing but again also individually selectable to post-fade feed should need be. The pre-fader feed is derived (referring to **fig 1**) from directly after the post-eq break point return differential amplifier at the same point as the 'pre-fade listen' is taken. Immediately following the mute/solo electronic switch is the take-off for the post-fader feeds, so they also, in addition to the stereo busses' feed, become muted when any other channel is in 'solo' unless the 'solo-unlock' function is applied.

This total of 10 auxiliary busses (don't forget the PFL) are applied to mixing amplifiers in the console's auxiliary function (known in vernacular as the 'garbage') module.

## Monitoring selection

Fair warning must be given of a fairly radical approach toward master monitor selection. One of the main bugbears of complex console system design is the practical elimination of crosstalk, the worst variety of which is differing material crosstalk. The relatively dire crosstalk between adjacent tracks on a multitrack recorder is passable because usually the interfering signal is related musically and probably up in the mix in its own right.

This isn't true of many of the alternate sources that appear on master monitor switches—tape machine returns, effect returns, radio tuner, turntable, other studio outputs (maybe); any of which is dissimilar from and immediately noticeable at any level when present as crosstalk into the main recording signal paths, multitrack or stereo. Whilst the actual crosstalk figures in a well built mixer between these high level line sources and the few inevitably vulnerable spots in the main path might seem to measure quite well, time and time again its been seen that any crosstalk at the same or higher level than the residual noise of the path is subjectively objectionable. For console manufacturers this has been a slowly developing nightmare—as signal path noise levels subside with development more and more rubbish is left visible, or should we say audible.

It is not so much crosstalk within the monitoring itself that is the concern, it is crosstalk into the real, live, signal paths that is to be avoided. The self evident answer (once, as much has been done to clean up the mechanics of the crosstalk as the given design permits) is to remove all the dissimilar high level line sources from the console altogether, except when they are deliberately selected for monitoring purposes.

This is precisely the approach taken—the master monitor selector is in fact a remote rack mounted box containing a stereo input matrix, with the normal switching and priority functions controlled by a logic system user addressable on the console itself.

## An 8-track format

The requirements for a 12 into 8 small multitrack console imply a monitoring system other than that outlined earlier. Physical size is not as crucial, with the extra eight module widths for the group modules not representing any serious problems, given the facilities present on them. Multitrack monitoring is contained entirely on the groups. The questions of what needs to be where within this format should be largely answered in the rough block system drawing in **fig 3.**

Instantly vanished from the channel module are the secondary gain control and pan control appropriate to multitrack routing, but a mono channel output is maintained, derived from immediately pre-pan along with all the other post-fader auxiliary feeds. This is to facilitate external routing via the jackfield of channel outputs or possibly even via an electronic matrix should the basic 12-8-8 concept get 'stretched' to sufficient input channels to deem it worthwhile. No 'A'/'B' monitoring is allowed for, but the original multitrack return differential amplifier is (in the case of the channels designated for remix) paralleled in the frame with the appropriate group monitoring multitrack return input amplifier. On those channels not so designated, it is available as a line input for effect returns etc.

Eight-group routing is achieved in odd/even pairs via the main panpot, an extremely versatile method despite its basic simplicity, with any of the four pairs (pick a number, any number) being designated the main stereo mixing buss.

Sufficient isolation between these 'stereo pairs' to enable their use as eight single groups or subgroups is achieved largely as a result of the excellently low end-stop resistance values of the panpots specified in the design—previously with such a routing method it has been a case of crossing fingers and hoping for a good batch of pots. As the design stands, the crosstalk characteristics, reactive and resistive, are significantly better than those achievable on most multitrack recorders.

The groups are themselves reassignable and panable onto any of the other groups (being prevented from routing back into themselves for obvious reasons by the mainframe wiring) giving an altogether quite versatile subgrouping and routing arrangement. The groups also carry the full 'A'/'B' monitoring routine of the original in-line module complete with a full complement of pre- and post-monitor gain control auxiliary feeds.

Ordinarily, the main group fader is dedicated to the group gain whilst the monitor chain takes the secondary control, but the local fader reverse facility is retained enabling these functions to swap—again allowing the engineer to juggle the mix on the 'right sort' of faders whilst other recording is still occurring.

Each of the channels is capable of having its 8-track routing disabled in favour of direct routing to the monitor busses, this being intended specifically for those channels not dedicated to the multitrack recorder but carrying effects returns. As can be seen from **fig 3**, the post-fader auxiliary feeds in this mode are still contained in the channel path enabling tape or DDL repeat echoes or other re-entrant effects to be created in the monitoring independent of the main recording channels.

Should all the input channels be in use and effect returns still required (a likely course of events with the relatively small number of channels) two panable line inputs to the stereo monitoring busses are present on the monitor select module.

As regards subsidiary monitoring functions, there are two individual solo activation systems for the channels and groups, by virtue of the fact they are not being asked to perform the same function. The channel solo defeats (again with 'unlock' if needed) every other channel in order to ascertain how that particular channel's signal has fared through the console path. It would be quite daft to have a channel solo muting the groups through which it is being monitored, similarly so for a group monitor solo to mute the channels which it is intended to monitor! The solo function on the group modules is called 'monitor solo' to save confusion and, being solely in the monitoring chain, is non-destructive to the recording signal path. Pre-fade listen whether of channels or groups overrides any other monitor selection, but despite this monitoring priority is non-destructive to the main signal paths also.

## And so . . .

As can be noted from the progressively less description required for alternative formats, the restructurable basic module concept is obviously quite workable, also well beyond the examples outlined here. These described formats are by no means The Only Gospel Way of approaching the various applications even within the confines of the basic modules—other system arrangements will automatically suggest themselves as the elements are detailed in subsequent articles.

There is no pleasing everyone—this design approach will undoubtedly be accused of being simplistic by a few, 'too fussy' by others. Although the perfect 'systemless studio' is nowadays impracticable, this possibly offers the closest working approach. ■

# Designing a professional mixing console

**Steve Dove**

# Part Two ~ Broadcast Consoles

**Broadcast control consoles are a different breed to the standard recording console. In Part Two of this series Steve Dove examines the background to their design and configuration, and describes an on-air broadcast control console.**

WHILST the essence of broadcast mixers is simple enough—the combining of a few sources directly into a pair of outputs—it is all the rest of the system and monitoring necessary to make the mixer operational that is the complex part. As a rough guide, it's been said that recording consoles are 80% mixing, 20% system whilst broadcast is 20% mixing and 80% system.

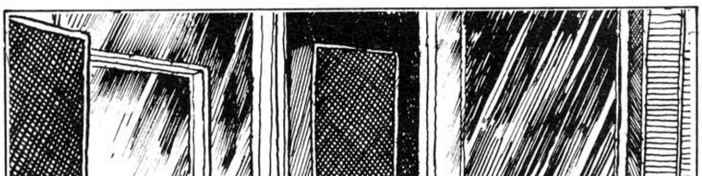
Generally the audio signal paths are (at least in relation to recording) laughably straightforward but some signal paths are totally alien to any other requirements. Despite the simplicity of the audio chain, the performance constraints and specifications expected by the broadcast authorities are exceedingly demanding—to the extent that the author is convinced there has never been a console manufactured by anyone in Britain that, when first switched on, has met totally either the IBA Code of Practice or an appropriate BBC spec. It isn't the purpose of this article to deliver a blow-by-blow account of how to make a mixer that passes the Code, those who know would realise the futility of the attempt and those who don't are best off out of it for their own sanity's sake.

In the early days of UK commercial radio there were virtually no manufacturers (other than those already supplying the BBC and export) capable of delivering what the new programme companies thought they needed or actually needed. Suppliers were difficult to come by for either systems (dual-purpose continuity/production control areas were unheard of until then) or even basic hardware (who made a stereo channel?). Also equipment budgets were awesomely small—station owners were typically small-town businessmen whose conception of a radio station was 'a chappy with a record player'—chief engineers of the time had a very steep uphill struggle.

The upshot of all this was that no-one, engineers, owners, presenters or manufacturers, really had a clue what was going to be an optimum working system.

Manufacturers, pleased to be free of the BBC's free-system 'all inputs and outputs accessible', found it possible to make the dedicated function small broadcast desks from modules already designed for excellently performing recording desks. Who else would measure recording desks when delivered before acceptance for noise, left/right crosstalk, If distortion, input and output transformer winding balance, etc?

*Bleary eyed engineers with crooked fags—cutters in one hand, soldering iron in the other—suffered late nights in many radio stations*



It's taken the last seven years of British commercial radio for the console manufacturers specialising in broadcast consoles to 'get it right'. Many grey hairs, late nights and early mornings have been suffered by engineers in new stations with cutters in one hand, soldering iron in the other, bleary-eyed with crooked fags hanging out of their mouths whilst the nice IBA chappies look on wistfully, drumming fingers.

A salutary tale to those who think it's an easy game.

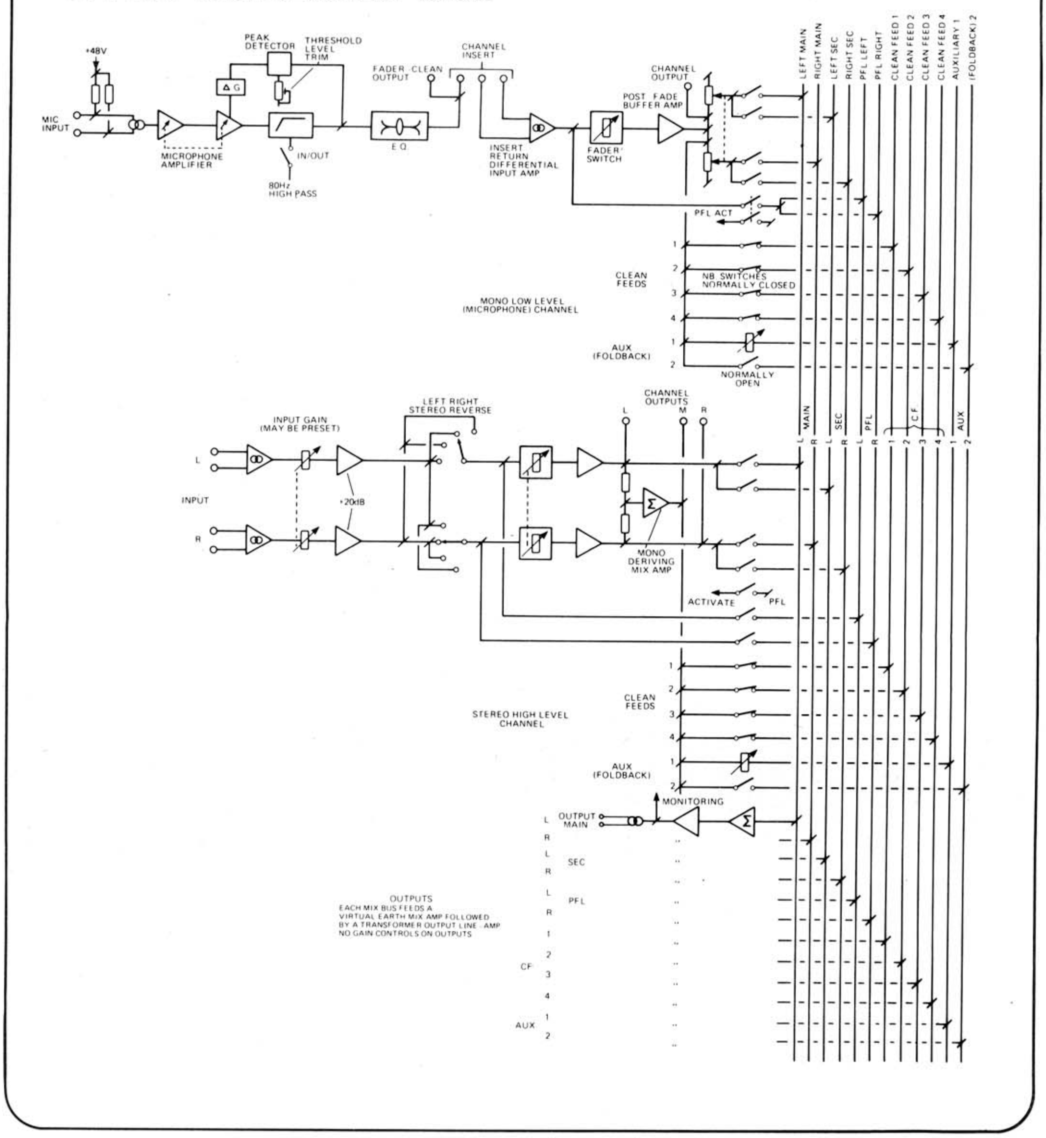
Although it is specifically only mixers intended for actual on-air broadcast that *have* to meet these specifications, they are meaningful determinations of path parameters before subjective apparence and as such realistically applicable to any audio-electronic signal path. It is illuminating, somewhat embarrassingly, to discover how few, especially recording systems, measure up.

Rarely is a broadcast mixer the only item of consideration in an on-air environment as it is always concerned with other control areas such as master control or even the central racks area. It is impossible to design a 'stand-alone' mixer—there are innumerable ties in both system and signal path to be considered.

## Remotes

Remote control of machinery such as tape and turntables is of far higher importance than in a normal recording environment—the mixer is the true centre of operation particularly in the case of a DJ-style self-operated

FIG. 4 BASIC SCHEMATIC BROADCAST CONSOLE




desk. Machine starts at least have to be ergonomically available on the mixer so that the operator does not have to change his position relative to a fixed point microphone. Extending this even further, a choice of start modes will probably also be required—fader, back-stop or press-button? In this realm of operator preference, cartridge machines pose an interesting question or three. Most cartridge machines are fully equipped for versatile remote operation with all the start/stop/fastforward, etc, function lines accessible along with tally light feeds for the various status—for a triple-stack that's quite a collection of pretty flashing lights to cram onto a compact channel module (nearly always the three triple-stack outputs are submixed into one stereo channel). Perfect for chief engineers with a *Star Trek* complex but not so for the cute, but dumb, chick doing the late show.

This consideration, along with a common DJ/operator preference for 'feeling' the cart-slot they are about to fire as a tangible reassurance that there is actually something *in* there, rather detracts from full remoting simply because it is possible. Similar thinking follows for 'proper' tape machines—perhaps the only function other than 'play' that needs to appear on a self-op console is a 'return to zero' indexing from an autolocate, in order to simplify the cueing up of a tape after prefade level checking. It is, however, again deeply suspected that given the time, the operator will physically check the machine and set up the cue by eye.

It must not be forgotten that self-op DJ-ing is possibly one of the most personal interractions between people and audio hardware and as fun a technological toy it may be, it also has to be straight ahead and logical. The consequences of failure or confusion in the studio can be several thousand listeners banging transistor radios on kitchen tables with puzzled looks.

## Production or On-air?

Two totally dissimilar approaches exist to broadcast control: (a) the console should have enough facilities to straighten out to broadcast standards any programme source that's thrown at it; (b) that all the programme material must be perfect before it hits the on-air desk regardless.

The first assumes that the technical competence of whoever is sitting behind the desk, combined with their conception of what sounds right is always adequate and reliable. The second, more realistically, assumes that this is not perhaps true, and benefits the DJ/operators by giving them less of a field of knobs to romp around in.

Naturally, provision has to be made somewhere on the station for rendering listenable less than adequate sources, but that control area doubles as a standby or alternate on-air control and cannot be greatly different in general layout and facilities to the original on-air desk. Fortunately, this seemingly unbridgable divergence in requirements is fairly simply resolved.

**Fig 4** is a block system schematic of a barebones broadcast console with minimal operator controls but adequate signal path flexibility and access to cater for most encountered broadcast situations. No on-desk equalisation is normally necessary, but facility for simple hf, lf and a single sweep frequency mid section is provided on the mono microphone channels together with a second order highpass 'rumble' filter. This equalisation is given only a limited range of $\pm 6$dB as if you have a microphone in a tightly controlled acoustic environment, such as a radio studio that needs tweaking more than that, it's time for a new one. It also prevents the DJ from doing horrible things to his own voice under the impression that it sounds wonderful, which it probably does to him with cans on and a bad cold.

A crash protection limiter is wrapped around the second section of the mic amp, catching an interviewee's lack of microphone technique, and usable (when the threshold is reduced into normal operating levels) as an effects limiter on the DJ's voice. He'll think that's wonderful too.

All in all, that should provide sufficient signal processing for any microphone source. Nevertheless a prefader breakpoint is provided, incidentally providing a fader-clean channel output feed, taken from the self-op microphone for talkback unless the channel is also used switched to an alternate source at any time.

The fader on the self-op mic position is seldom used as more than a soft switch and may be replaced by a conventional switch. This also solves the problem of what level to set the prefade monitoring feed from that channel. Ordinarily a mic channel fader is set to have calibrated gain at a certain amount of fader back-off, between 10 and 15dB, the PFL being set correspondingly. In the self-op position where the fader is wound right open, the PFL will be inaccurate in level by the amount of the back-off. A switch replacing the fader provides a defined amount of back-off, ie nil, to which the PFL may be set, thus avoiding the unfortunate situation of otherwise identical channels being non-interchangeable by virtue of differing internal level structures.

The amount of signal level headroom to be aimed for in a broadcast microphone channel provokes fairly interesting debate. With presenters well accustomed to reasonable microphone technique, virtually no headroom above the normal peak programme level is needed, but in an interview situation? The IBA code calls for headroom of at least 20dB above operating level with gain reduction to be made without incurring distortion by any level control accessible in the signal path. This, with supply rail voltages com-

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mon today, is reasonably simple to realise. They 'like to see' though, in the microphone channel, a headroom capability of 30dB above operating level (Daubney level). This implies running the mic channel at a level of about 10dB below standard line level, with the gain being made up either in the post fader buffer amp or actually directly on the mix buss. Naturally, if a desk output fader exists (and really there is no operational reason why one should), the desk will fail this stricter test since it would be impossible to eliminate clipping distortion in stages after the 10dB gain make-up has been introduced.

The stereo channel depicted on the main signal path schematic runs at unity gain throughout, headroom not being anything like the hassle it is on the microphone channel by virtue of the pre-processed nature of most line sources. A total of 20dB gain variation is made available at the front end to compensate for the widely varying levels from some prerecorded sources—for instance on disc between K-Tel compilation albums and mid-sixties Motown singles.

No insert point is provided since this can be simply overcome by making the two channel inputs jackfield accessible. From a practical maintenance viewpoint, this access is a necessity—there is nothing worse than grovelling about under and behind a desk in order to unplug a source to perform measurements on either the desk or the source, whilst finding the right test lead connectors to do so.

Mono derivation is achieved by a dedicated mix-amp for all the cleanfeed, auxiliary and channel outputs. Although it is entirely possible to derive most of the mono feeds via 'Y' resistor networks directly onto the mix busses, this has the drawback of worsening crosstalk, and in view of the number of feeds, makes an uncomfortably low impedance for the post-fader buffer amps to feed, unless they are beefed up to line-amps. The increased crosstalk is due partly to the greatly increased numbers of paths between the left and right channels, whilst to a large degree the increased ground currents resulting from the lower impedances will unearth (sorry) any further inadequacies in the grounding arrangements.

FIG.5 TYPICAL PANPOT ARRANGEMENTS

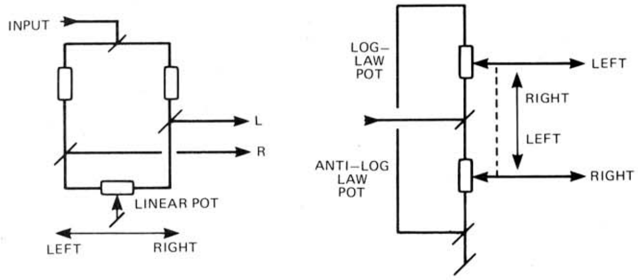
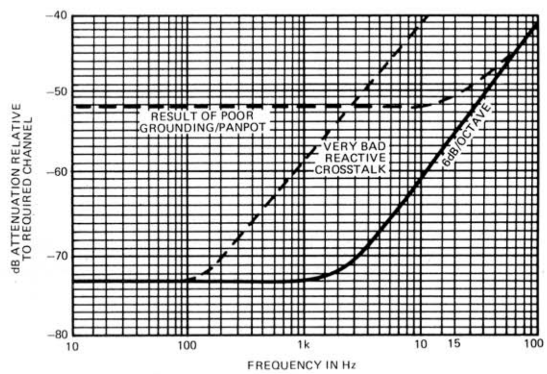



FIG. 5a FIG. 5b

FIG.6 TYPICAL MIXER LEFT/RIGHT CROSSTALK CHARACTERISTIC




## Crosstalk

Left to right crosstalk is possibly one of the more difficult specifications to better, particularly should conventional panpot arrangements be in use on the microphone channels. The two most common arrangements are shown in **fig 5**. The popular arrangement in **fig 5a** suffers from the fact that the wiper of the pot does not achieve total contact across the track, hence allowing a resistive path (ie a crosstalk path) to exist between the two ends of the pot and therefore the panned output. A less than perfect ground connection to the wiper will aggravate the situation. This is true also of the arrangement in **fig 5b**. Here, the effective ground resistance becomes added to the nearly inevitable end-stop resistance of the pot causing not only incomplete attenuation at full rotation, but again a resistive path between the two pan outputs. The latter is a more satisfactory panning system, but it does mean careful selection of the actual pots used in construction to avoid those with undue end-stop resistance. A 'bad news' pot can easily cause the channel to exceed the −50dB L to R, R to L crosstalk spec.

**Fig 6** shows a fairly typical crosstalk characteristic. The flat portion at the lf end is 'resistive' crosstalk, caused chiefly by panpot inadequacies and generally a less than perfect internal grounding path system. The rising bit toward hf is 'reactive' crosstalk, the result mainly of proximity and capacitance between the two paths through the channel console, being directly proportional to relative level and proximity, and inversely proportional to impedance (the lower the path impedances, the lower the crosstalk). It may be inferred that minimising reactive crosstalk is both an electronic and mechanical design headache. Relatively high impedances are unavoidable at the wipers of attenuators and a few other points of the path, whilst physical containment of the two separate paths within the same modules and printed circuit layouts limit how far you can physically remove them from each other.

Producing broadcast consoles that now do not need individual attention to better the IBA's tight −50dB at 15kHz spec has taken many years' experience, many mixers and much thought.

An old dodge that was originally mooted as a joke in the pub but which to everyone's amazement and hilarity actually worked, involved deliberately introducing a known amount of crosstalk in the channels, then 'tuning out' by inverse capacitative cross-coupling of the appropriate mix amps—the capacitances needed being so small that short pieces of twisted wire or screened cable sufficed. It was amazing how many people stared in disbelief then walked away shaking their heads, muttering, at the sight of a giggling loon gazing intently at a meter whilst purposefully snipping ⅛in at a time of a bit of wire just dangling in mid air apparently going nowhere.

Stereo source channels are occasionally more prone to crosstalk than mono channels, since rather than the stereo being derived at the tail end of the channel by a panpot, it is stereo throughout. Care in card layout, together with an awareness of even passive component idiosyncracies is the only method, short of using the rather wasteful one of using separate channels for the left and right. You would be surprised how much crosstalk can be introduced between two adjacent electrolytic capacitors, or even two mylar/polyester capacitors.

Despite how much care is taken over card design and buss systems, wiring to and from the card connectors to the external terminations is a major problem, especially if the wireman has an obsession about being neat and tidy, tying all the cabling into beautiful tightly bundled looms. Some you can, some you daren't.

Virtual-earth mixing busses are not entirely blameless for crosstalk, despite the fact that the impedance is, well, virtually earth. The signal is present as current, current generates a corresponding magnetic field where it is present (in this instance the buss) and the magnetic field induces a current into any adjacent, preferably parallel, bit of wire (in this instance the buss of the other stereo side); result—crosstalk. It's good practice for this reason to introduce a ground buss between each virtual earth buss, but this still doesn't help prevent an even sneakier path, present in mixers with ferrous chassis. A buss's magnetic field is capable, unlikely as it may seem, of introducing eddy currents into a steel chassis it may be in close proximity to, which can reintroduce currents into similarly located busses causing crosstalk and a designer to take up farming.

From the same root of inductive coupling comes the seemingly obvious advice not to mount output balancing transformers too closely to each other—this was the cause of many hours' fun trying to source a curiously vicious rise in lf crosstalk.

## Microphone channel

**Fig 7** shows a very useful channel subcircuit for use on microphone channels in on-air broadcast and broadcast production. It provides the usual channel facilities of PFL switching and channel mute via electronic switching, and also gives the person at the other end of the microphone some useful facilities.

All the activation lines are ground-

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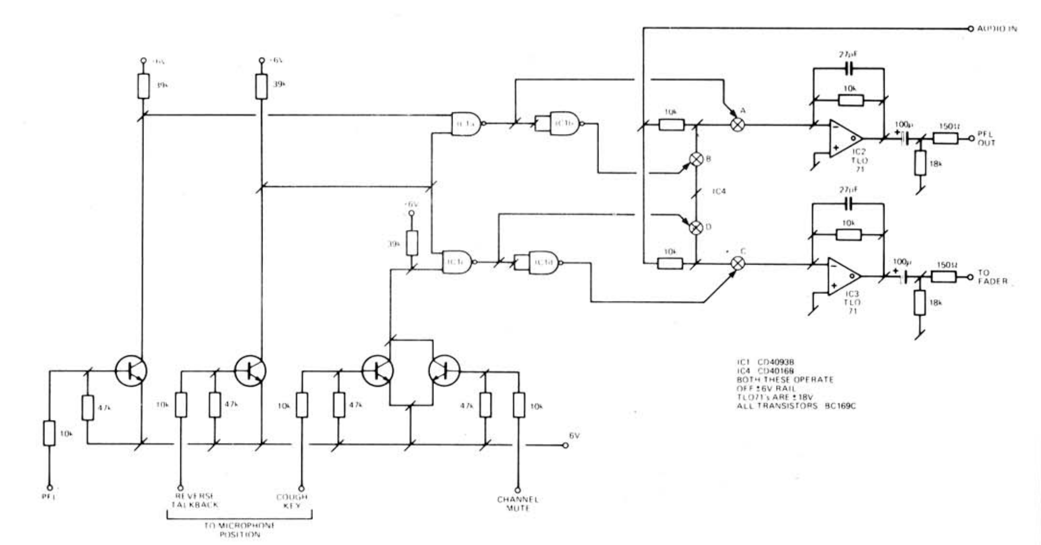
ing contacts, obviating the need to run switching pairs everywhere. Two control lines need to be run up to the microphone position, and switches provided for the reverse talkback and cough-keys. In operation, the reverse talkback button mutes the channel from air but applies it instead to the PFL system (which normally is a monitor override function) so that the presenter can give cues to the operator without having to attract his attention first, even if his mic is live. Activation control for the PFL system should be simple since there is a wide variety of control voltages and senses available in this circuit.

The cough-key is simply a parallel activating channel mute.

FIG.7 BROADCAST CHANNEL FUNCTION SUB CIRCUIT




## Why cleanfeeds?

Best described as 'everything but' feeds, the need for them can be illustrated by three fairly common operational functions; studio foldback, telephone balancing and network contribution.

Ordinarily studio foldback is by cans or a loudspeaker muted by any live microphones in the same area, but for certain types of shows it is desirable to leave the loudspeakers on continuously but with the live microphones removed from the feed. This is primarily for the benefit of a non-technical studio audience who may otherwise be put off and act less naturally as a result of things not sounding as they expect them to. Similarly in a 'guess that tune' type of format a desk output minus the microphone feed is required whilst the mic is live to pick up audience remarks. In a foldback arrangement, the 'cleanfeed send' is the programme minus mic loudspeaker feed, whilst the 'cleanfeed return' is the microphone itself, being remixed back into the programme.

Telephone talk-ins are perhaps the most common example of programme material needing cleanfeed manipulation. The telephone line termination is by a hybrid, whether it be a transformer arrangement or an electronic system exemplified by the Alice *TBU3* or Studer units. The object of the hybrid is to provide the subscriber's signal relatively free of colouration and interference from the signal that is being fed to him. It would be nonsense to try to send his own voice back down the line to him, so it must be removed from the studio feed to the TBU. The cleanfeed send is the studio output minus the subscriber return from the TBU, whilst the cleanfeed return is that subscriber return, being brought up into a channel which is routed so as not to appear in the cleanfeed send.

An approach taken originally to create a cleanfeed on ILR desks for either telephone balancing or network provision was what can best be described as 'sequential cleanfeeding'. This entailed grouping all the channels appropriate to the first cleanfeed into a pair of mix amps, from which the cleanfeed output was taken, whilst the cleanfeed return together with the original mix were remixed into a second, sequential, pair of mix amps from which a second cleanfeed could be taken and so forth. The two main drawbacks to this method are the number of mix stages the original sources are subject to (degrading noise and crosstalk) and the inability to cross-cleanfeed (ie have cleanfeed returns crossmixed into other cleanfeed sends) other than in the rigidly defined sequential order. The number of cleanfeeds was also inflexibly specified at the construction of the desk, being that once it was built, it would entail a major system upheaval to change.

Needless to say, an alternative method, parallel cleanfeeding, has far fewer problems and is of considerably greater flexibility, resembling faintly free-grouping multitrack recording routing. In this method, a cleanfeed buss consists of a mix of all the sources, with the exception of any channel (pick a channel, any channel) which can have its feed deselected from that buss. Obviously, the appropriate cleanfeed return may then be brought up to that channel, since it can no longer be routed out to itself, also it is present on all the other cleanfeed busses. There is no limit (other than the predetermined size of the system) to the number of channels that can be allotted to separate cleanfeeds and since all the cleanfeed busses carry all the sources (other than the ones specifically deselected from their respective busses) crossmixing between cleanfeeds is implicit and automatic.

Really, the parallel cleanfeed system can be treated in design as merely an extension of the auxiliary feed system, the only difference being that the sources have to be deselected from the buss rather than selected to it. Most broadcast auxiliary feed requirements are for either a source (or a small group of sources) in isolation, or for 'everything but' feeds. Rarely is a 'recording type' separately mixed feed needed so anything more than one control variable foldback feed is a wasted luxury.

## Broadcasting line amps

The thought of deliberately including high and low frequency roll-off filtering in any line output of a console must seem a total anathema to many studio engineers brought up in the 'broad is beautiful' era. In many circumstances, though, far from it. While not actually mattering whether filtering is present, the lack of it can occasionally have dire consequences —some occasionally potentially expensive. Immediately springing to mind are monitor loudspeaker and headphone feeds. Allowing possibly large amounts of energy through at frequencies the loudspeakers can't,

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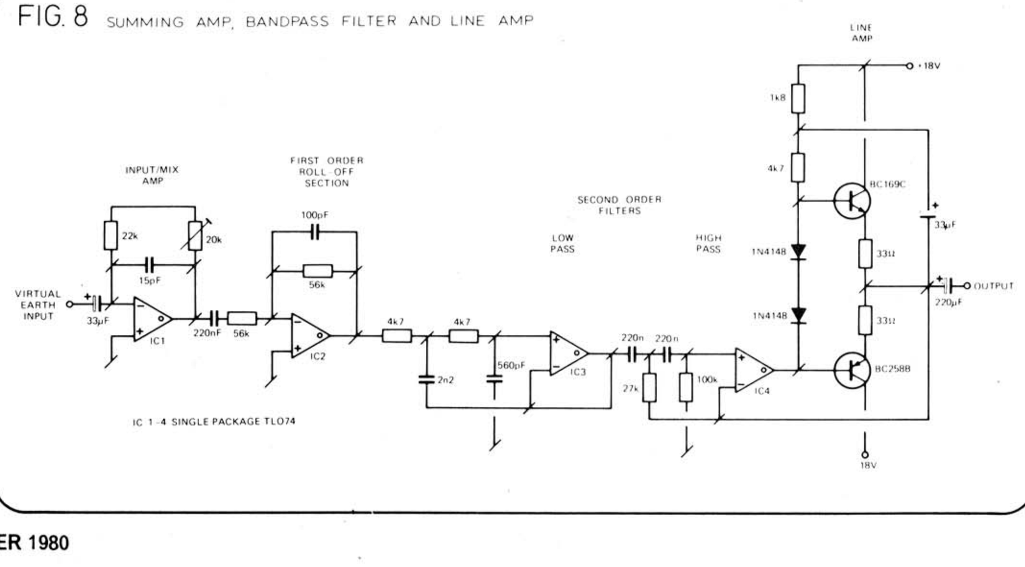
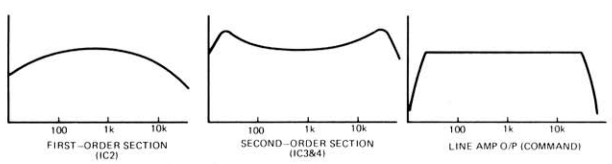
FIG. 8 SUMMING AMP, BANDPASS FILTER AND LINE AMP



# Mixing console

FIG 9 FILTER BANDPASS CHARACTERISTICS (FIG. 8)




and your ears won't, do anything about seems a singularly silly pursuit, however 'purist'. Such transducers are designed to operate within certain power and frequency handling parameters and making sure you don't ask them to exceed those is a prerequisite of their longevity!

The summing amp, line amp and filter in **fig 8** was originally designed as the main desk output feed for a series of BBC radio continuity consoles—the main design criteria for the filter being that the lf response was down 10dB at 10Hz. The somewhat embarrassed, mumbled reason for this was that the significant quantities of subsonic rumble, naturally present from mics and turntables, was sufficient to interfere with and create sporadic responses from the subsonic data signalling system used extensively on the network's PCM programme links. Disc jockey drops pickup on record, transmitter switches off in the Outer Hebrides, get the idea?

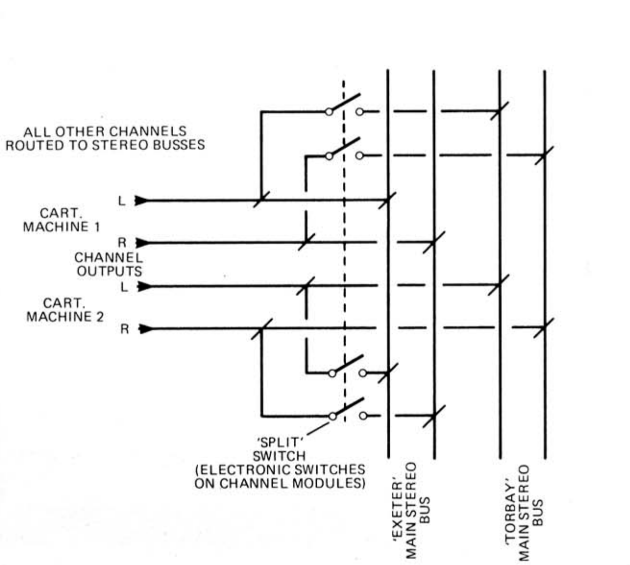
It is a fairly straightforward third order highpass and lowpass filter, with a gain adjustable mix amp at the front and a line output complementary transistor pair tacked onto the last filter amplifier. All the gain and frequency determining elements are kept isolated and non-interreactive for versatility, and with the values shown the ½dB down frequencies are close to 20Hz and 20kHz, with passband ripple of who cares. In order to achieve a rapid out-of-band fall-off slope characteristic, the lowpass and highpass sallen and key filters (IC 3 and 4 respectively) were calculated for quite a high 'Q' and hence 'peak' approximately 1dB just inside their passbands, **fig 9**. To prevent the headroom margin at those frequencies being eroded by those 1dB peaks, the single-order roll-off and 'peak-smoothing' sections were placed ahead, around IC2.

ICs 1 and 4 are in fact all in a single 14 pin *TLO 74* quad op amp package, making possible a fairly high card component density and hence a compact system element.

All in all, it's quite amusing to watch a sweep frequency through the filter plummet as it passes out of band! Multiple passes of the same signal through this, or any, filter will have of course, a compounding effect. Subsequent generations will take on more and more exaggerated characteristics of the filters, but if they are present only at main outputs 'grand mastering', or monitor feeds, this is quite unlikely to become operationally problematic.

This filter was optimised for maximally flat in-band frequency response, phase response secondarily. Naturally, any roll-off filter will create in-band phase anomalies in particular approaching its turnover frequency. The question of the audibility of phase shifting will rage unabated and inconclusively for many years yet but for the purposes of justifying these particular filters, it is worth remembering that most frequency selective attenuation occurring in nature is interrelated with phase shifts, often of greater severity than those introduced by this filter. Maybe the same can't as easily be said about the vicious 12th order (and upwards) anti-aliasing filters commonly used in digital signal processing, but that is a separate argument altogether. A common finding by many authoriatative studies (including those by Ma Bell) is that until group delay is extended until there is an apparent *timing* disassociation, relative phase shift isn't noticable. This corresponds to many, many *cycles* of shift, never mind degrees.

FIG. 10 SPLIT ADVERTISING SYSTEM




## Twin Stations

Coming into reality very shortly will be the new 'twin' stations in the British ILR network, the first of which being Devonair, operating in Exeter and Torbay. The system design for such a set-up is, shall we say, fraught, but an interesting problem was the requirement for split advertising between the two areas despite a fair proportion of the 'proper' programming being shared and operated from a single control area.

Each of the main on-air consoles was designed with two stereo mixing busses with separate groups designated for Exeter and Torbay respectively. Ordinarily with completely shared programming both stereo groups receive all the sources, but when a 'split' button is activated, the channel carrying a triple-stack cartridge machine is deselected from the Exeter buss, and another channel carrying yet another triple-stack deselected from Torbay. This enables the required function of separate advertising cartridge feeds to the two areas, **fig 10**.

In this somewhat extreme instance, the main desk output feeds are being treated as cleanfeeds, being to a specified degree clean of each other.

## Master control

In the large ILR stations with large engineering staffs, master control of station output (always desirable) becomes practical.

Traditionally, master controlling has meant engineer-driven programming with nominal, or basic control, being given to the presenter. Working on the basis that the presenter's desk for redundancy's sake, or operational need, is to be air-capable in its own right, this has meant that in most current arrangements the MCR desk merely has sub-mix corrective control over the presenter's desk, with provision for the addition of commercial and taped feeds in addition to the 'nasty business' of phone-ins and outside sources. This leaves the things that need detailed attention to the engineer and more time to the presenter to be creative.

In order not to arrive at a second desk situation, where all the controls are duplicated, sub-mixes of the various source characters (grams, carts, tapes, mics, etc) are derived from the presenter's desk channel outputs and taken to the MCR as stereo line sources. An exception to this may be the main and interviewee microphones which would be split at microphone or mic-amp level and applied to separate mic channels in both desks, providing both a 'safe' and engineer control over interview situations. A major advantage to this system is that the MCR desk can be a 'normal' desk or even identical to the presenter's desk, with the sub-mixes and microphones appearing merely as an additional set of sources. Should it be required to use the MCR desk on-air, presto, a few switches and it's ready.

A differing approach, borrowing from recording automation technology, is to take the presenter's desk main stereo output direct to air but to provide each channel on each of the two desks with a VCA fader system. The presenter's desk will behave exactly as normal unless the engineer takes control of the VCA from the appropriate fader on the MCR desk, either completely or in an 'update' mode which increments or decrements the channel level around that set on the presenter's desk. Subgrouping, if required, is extremely simple being attained through tying the appropriate dc control voltages together, allowing the engineer the ideal operational circumstance where his desk, which may be in all respects identical to the presenter's, to appear to be just like a normal studio desk, with the sources appearing in the same place on both desks, regardless of whether he has taken control of them or not.

Since the MCR control is done via dc and may be subgrouped, it would be entirely possible for the engineer to leave the necessary faders clear and use the audio signal paths of the desk for other purposes—editing, dubbing, even pre-recording an interview—whilst still keeping an eye and having control over the other desk. Exactly how popular this capability would be with the engineers is, one feels, an altogether separate issue.

Perhaps an ideal arrangement, even in smaller stations would be for identical 'twin' control areas each VCA equipped and capable of controlling the other as and when required, obviously with a security system to avoid the delightfully forseeable disasters such a system could promote.

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# Mixing console

**Due to editorial space constrictions we were unable to publish the PA console description in Part One of this series. Below we detail the PA console format which should be read in conjunction with last month's article.**

## A PA desk format

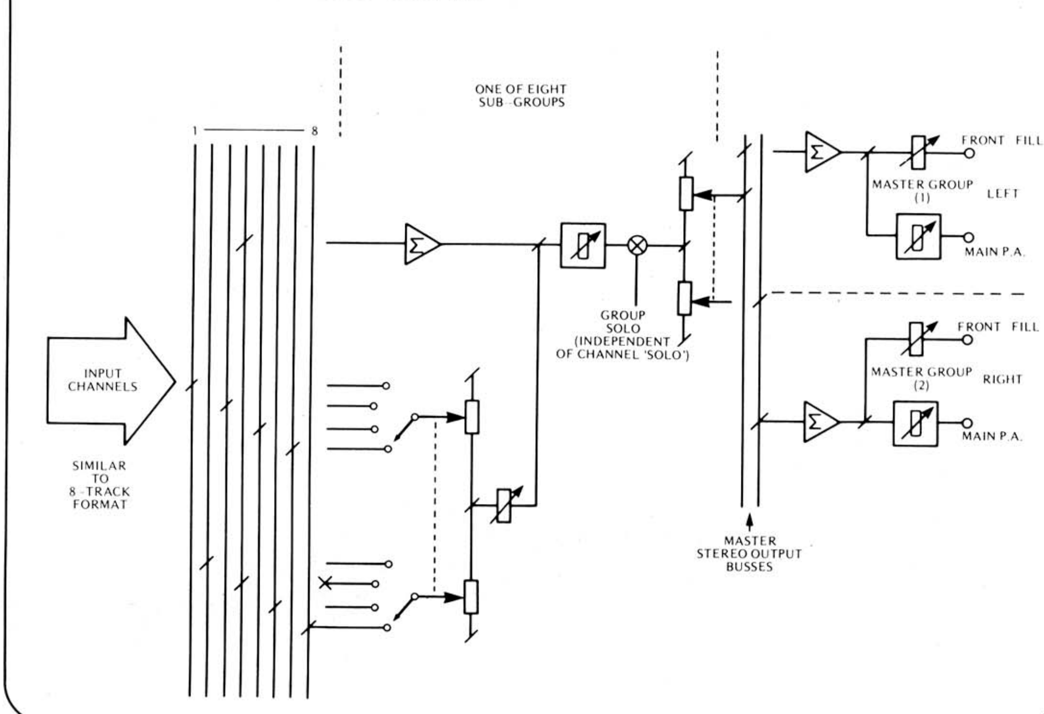
The PA desk format (outlined in **fig 11**) is fairly similar to the 12-8-8 format with the exception that no separate monitoring chain exists at all—the monitor loudspeaker system being very big, very loud and possibly 30ft above a stage 200ft away.

The channel is in fact identical to that discussed for the 12-8-8, including the separate solo system and 8-group routing. The groups, however, differ in that although there is still a group route-back facility, this now utilises the secondary gain control as well, whilst the main fader with its associated pan control feeds another pair of busses which constitute (together with another pair of group modules) the final desk output. The gain controls are again reversible at will.

No pan controls exist on the final group modules, but the two gain controls (non-reversible) remain in order to provide for control over two separate pairs of line-output feeds—nominally 'main PA' and 'front-fill'—often requiring individual setting.

Effect returns with this more than any other application tend to be problematic. All the channels inevitably end up used, so there is no possibility of bringing them back in that way and worse still, all the auxiliary feeds will probably be in use to effects, being that foldback is normally taken care of with a secondary stage console. For this reason, a separate module containing six 'bare-bones' effects returns, each with control of level and pan directly into the main desk output mixing busses is included as part of this mixer's 'backend' ■

FIG.11 SKETCH SCHEMATIC PA DESK BACK-END



# Designing a professional mixing console

Steve Dove

# Part Three~Op~amps, Friend or Foe?

FASHIONS change, the laws of physics don't. A simple and irrefutable statement, one would think. Unfortunately this industry, like most of the others which survive off the entertainment media, is populated with large numbers of persons who persistently refuse to believe it. Such are the individuals who are responsible for sweeping condemnations based on statements that tickle the sense of plausibility rather than sufficient breadth of comprehension and depth of knowledge to substantiate or explain them. So many of these proclamations are made for political and commercial reasons, totally unrelated to actual technological facts.

Such are the statements from which fashions are born—inertia sweeping them forward until the original criticisms have been well laid to rest but the engendered antipathy lingers on irrationally, supported dim-wittedly by those similarly incapable of substantiating their own opinions. Sadly, in an industry where abstract notions are a stock-in-trade and everyone has a pair of ears it is quite difficult to make clarifying statements based on facts—someone somewhere will always be at hand to propose yet another set of glazed-eyed contradictory waffle.

**"... people have got used to treating op-amp ICs as plug-in blocks of gain with little consideration for the fact that inside is a real, live collection of electronic bits which still have all the problems 'real' electronics always had ..."**

Consoles utilising integrated circuit op-amps have suffered from this exact syndrome, collecting a (sometimes deserved) dreadful reputation in the early days which has stuck.

This article is an attempt to explain the history and shortcomings of IC op-amps from conception to present day, to point out how some shortcomings are overcome and to provide reassurance that there is nothing really evil about those funny square black spiders after all. It is also an example to those prone to wistful opining that this, along with most other technology, is well understood and quantified, the concepts if not the details having been defined probably well before their birth.

## Devices

Many years ago, the author remembers deeply coveting then eventually giving in and forking out nearly five late 1960s pounds for a tiny transistor sized eight-legged queer-thing. At long last he actually held between quivering fingers a real, live Fairchild *UA709*!

This breakthrough opened up whole new avenues of creative ways to generate spurious oscillations. Many happy and otherwise hours were spent trying to get the wretched thing to do anything other than squegg. Never the most stable of creatures, the *709* once tamed provided a faltering education in the idiosyncracies of op-amp circuit design until expiring sadly and silently attempting to drive 15dBu into a screwdriver. Output stage protection was *not* one of its notable strongpoints.

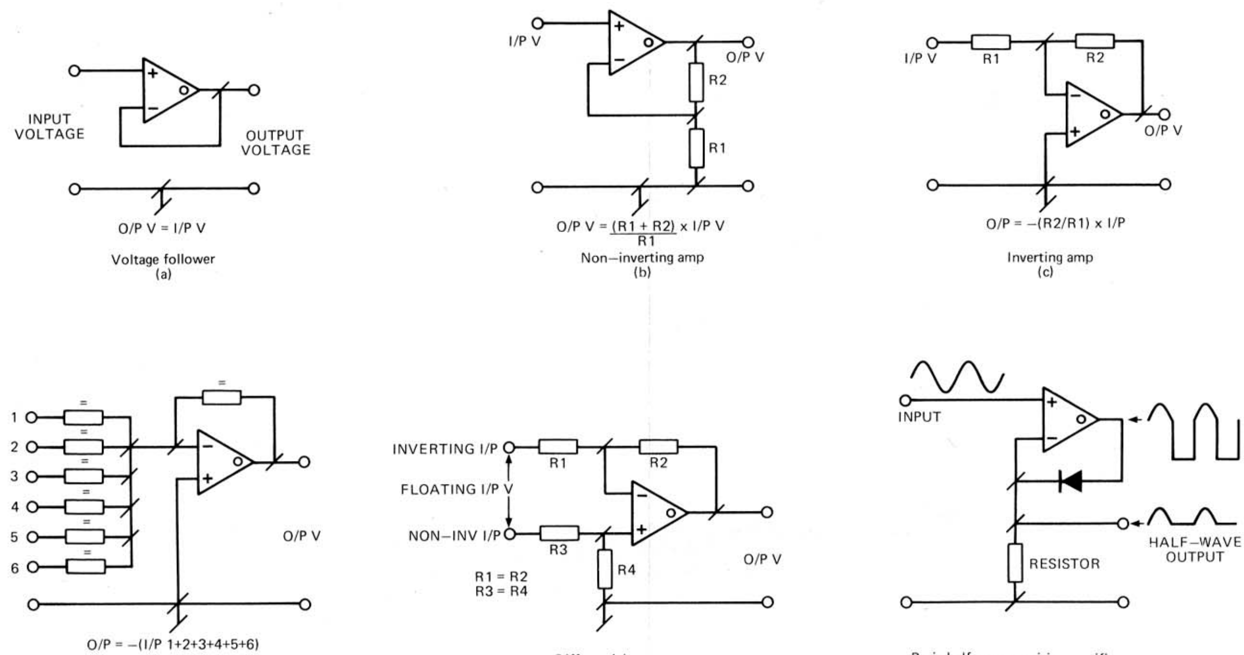
At this stage in the game, discrete transistor circuitry still ruled supreme in audio. The new fangled spidery things were eventually compensated sufficiently to remain operationally stable but little high-frequency loop gain remained to guarantee enough feedback for adequately low hf distortion. Also, crime of all crimes, they were so wretchedly noisy. Although their parameters could be set up to be acceptable for any set application and gain setting, the very nature of control in consoles is variable so the devices would almost inevitably end up operating away from their optimum.

Hot on the heels of the *709* came the now much loved and despised dirty old *741*. Best known in its plastic encapsulated 8-pin dual-in-line incarnation, it still took our industry many years to catch on to the fact that here existed a seemingly almost vice-free op-amp. Well, at least free of some of the *709*s vices, let's say—it was heavily internally compensated hence stable (unless you did something daft to it) the penalty for which was rapidly disappearing open loop gain with increasing frequency. There was just enough gain left in hand to get away with 20dB of broadband gain safely over a 20kHz bandwidth. Absolutely no mention will be made of the many (some well respected) mixer manufacturers who actually used them in mic-amps and mix-amps with anything up to 45-50dB gain around them . . .

Some IC manufacturers actually Came up with pleasant *741*s which were usably quiet and did not have output offset voltage problems on the scale of earlier devices. The *741* was also output protected to the extent of being short circuit proof. Sighs of relief all round.

Subsequent generations of op-amps to the *709* included the *748* (the

FIG.12a-f BASIC OP-AMP CIRCUIT CONFIGURATIONS




Voltage follower (a)

Non–inverting amp (b)

Inverting amp (c)

Virtual earth mix–amp (d)

Differential amp (e)

Basic half-wave precision rectifier (f)

uncompensated sister to the *741*) and the *301*—again some versions of which were excellent for the class of device. The *748* and *301*, being user-compensated did allow for more optimal parameter setting and in most circuits only required one capacitor to achieve this, as opposed to the *709's* necessary two resistor/capacitor networks.

This, although on the surface appearing to be of great convenience to the user, disguised the fact that far superior bandwidth and phase-margin performance could be obtained by carefully considering the nature of the compensation network. Rather than just a simple capacitor of sufficient value to hold the amplifier stable (which also turned the internal compensated transistor into a Miller integrator doing absolutely nothing for the device's speed) a more complex network such as a 2-pole C/R network **(fig 12)** improved matters greatly.

External feedforward whilst in use as an inverting or virtual-earth mixing stage also enabled a dramatic increase in bandwidth and hence speed over the more conventional compensation arrangements **(fig 13)**.

Full treatment of the compensation of the *301* family together with performance graphs are given in some manufacturers' data books, possibly the best being by Advanced Micro Devices a company who, oddly enough, don't seem to really specialise in op-amps at all.

All these early devices had one great failing though, one which has quite recently been leapt upon vigorously by the hi-fi fraternity and audio engineers alike in a frantic witch-burning ceremony for the like of which both categories are well noted, from time to time. Please stand up, the magic buzz-word (buzz-phrase?) *slew-rate*. Slew rate is the speed (measured usually in V/μs) that an amplifier output shifts at when a step source of extremely high speed is applied to the input. All the early generation op-amps had slew rates in the order of 0.5V/μs which by today's standards does not bear mentioning in polite company, but no-one really knew much better then.

The speed limitation was nearly always in the differential dc level-shifting stages of the devices, it being quite difficult to fabricate on the IC wafer ideal classes of transistors in configurations

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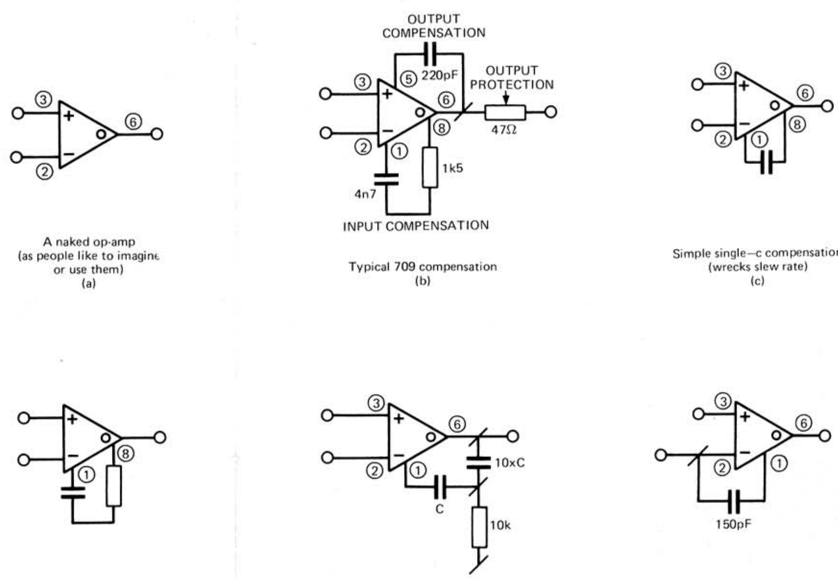
FIG.13a-f VARIOUS OP-AMP COMPENSATION TECHNIQUES




A naked op-amp (as people like to imagine or use them) (a)

Typical 709 compensation (b)

Simple single–c compensation (wrecks slew rate) (c)

Single–c with limiting resistor (extends bandwidth) (d)

Two-pole compensation (improves stability and bandwidth) (e)

Feedforward (considerably improves speed) (f)

(PIN NUMBERS FOR COMPENSATION BASED ON 301 TYPE)

# Mixing console

necessary to improve matters without compromising other device characteristics (such as input bias current which affects both input impedance and offset performance).

'Feedforward', in which a proportion of the unslewed input signal is fed around the relatively slow-responding lateral pnp stages, improving slew-rate and bandwidth appreciably, is used for example in the *LM318*. A device with still a notable number of devotees, with an achievable slew-rate by this technique of some 70V/μs.

It was in this area of slew-rate, combined with a significantly improved noise performance (again another parameter suffering from difficulty in fabricating appropriate devices in a relatively 'dirty' wafer) the next major breakthrough occurred in devices commonly used for audio applications; the Harris *911*. Although dramatically improved, the slew-rate was still not *that* fast and was also asymmetrical, being +5 and −2V/μs.

In recent years from the realms of the hitherto specialist domain of ultra-high input impedance instrumentation op-amps, has emerged a breed called 'Bi-FETs'. These have a closely matched and trimmed field-effect transistor input differential pair (hence the typically unimaginably high $10^{12}\Omega$ input impedance) and a very fast 13V/μs structure throughout. These wonderful creatures are typified by the Texas Instruments 'TLO' series and devices such as the National Semiconductors *LF356*. Selected versions can, when source impedance optimised, give noise figures bettering 4dB at audio—thoroughly remarkable for units costing very few pence more than a *741*.

The device speed has been achieved by the replacement of the conventional bipolar transistor differential input and level shifting circuit by the FET configurations. Incidentally, the intrinsic noise characteristic of these FET front ends is significantly different from that of bipolars and seems perceptually less objectionable. Needless to say, these are the devices around which most of the circuitry in this series has been designed, with minimal exceptions.

Talking of exceptions, there is one IC device that was designed specifically and optimised totally for inclusion in high quality audio equipment. With a quoted noise figure of better than 1dB, slew-rate again of 13V/μs and the ability to drive a 600Ω termination at up to +20dBm, the Signetics *NE5534* (TDA1034) is truly a chip amongst chips. It is also expensive.

This on its own is a perfectly valid reason for not using them everywhere, but more to the point, how many actual circuitry circumstances demand each and all of these characteristics? Not many and although a fairly detailed reasoning of design criteria is given in each of the circuit descriptions during the series, a brief explanation to put the minds of the 'purists' at rest who would otherwise demand using *5534*s thoughout, is in order here.

Noise in any competently designed and operated console can be attributed mostly to two sources, these being; (a) mixing ampliers with an appreciable number of sources and hence a lot of 'make-up' gain, but predominantly (b) the input stage, especially a microphone amplifier with a fair amount of gain in it. Once a background noise level is established from the front end stage (at a level obviously dependent on the amount of gain employed there) the difference in noise contribution between an amplifer with a typical unity gain noise of −120dBu and one of −115dBu is for the vast majority of considerations totally insignificant.

The output driving capability of the *5534* is not really worth putting to the test since conventional line-amp designs are still cheaper to construct than even the 1,000-off prices. The performance and ease of using the *5534* as a microphone amplifier far outweigh the hassle of a similarly performing discrete transistor design, which in this specific area is still its main close rival.

Unfortunately, the *5534*s are still the audio industry's favourite 'flavour of the week', and anything that isn't liberally peppered with them is regrettably considered déclassé.

In the realm of altogether more esoteric devices fall the purpose-designed encapsulated discrete amplifier modules such as the *JE990*, designed by Deane Jensen of Jensen Transformers. Many fascinating solutions to op-amp internal design problems, (some of which even IC designers evidently haven't realised existed) are implemented in this design whose features demand a total reappraisal of contemporary audio circuit design and philosophy. Optimum input source impedance (normally about 10kΩ with most IC and discrete amplifiers) is reduced to about 1kΩ by the use of an IC multi-parallel input transistor differential pair whilst small inductors in the emitter provide isolation from potential hf instability due to the gain/bandwidth characteristic of that first differential stage shifting with varying source impedances. Unity-gain noise is a staggeringly low −133.7dBu whilst the output is capable of delivering full voltage swing into a 75Ω load so permitting the use of exterior circuit elements of far lower impedance and hence reducing thermal noise generation due to them.

This elegant device inevitably carries an elegant price-tag. Its many attributes point the direction for design, it being the only direct improvement upon currently adopted techniques. It is a leap in advance of any devices available in IC form and also, to the author's knowledge, of any universal discrete circuitry elements used to date in console manufacture.

## Instability

An unexpected thrill facing designers as they upgraded to the newer, much faster devices was the tendency for all their previously designed circuits to erupt in masses of low-level instabilities like an attack of chicken-pox, even in what *had* been perfectly tame boards.

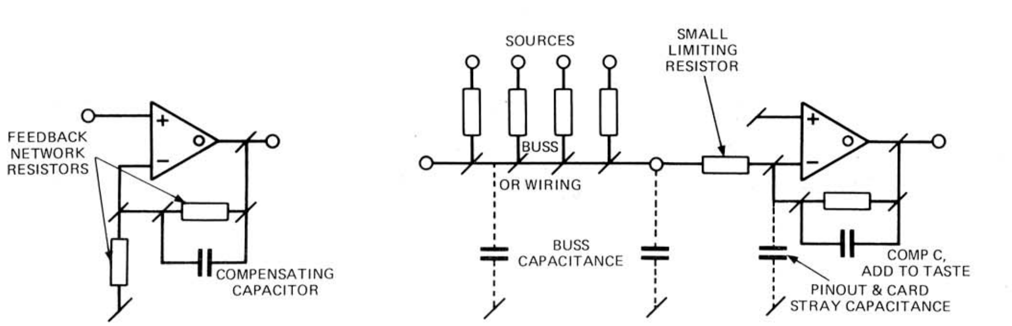
Layout anomalies, such as track proximity were a major contributor toward this so new layouts had to be generated with a whole new set of conditions added to the already hazardous game of card design. However, the real roots to this problem lay with the devices themselves and a lack of appreciation of the relationship between their internal configurations and the outside world. Everyone who had been brought up designing around *741*s and their ilk had got rather too used to treating them in a somewhat cavalier fashion and for good reason—there was precious little you couldn't do with them and without even showing a hint of oscillation. People got used to treating ICs as plug-in blocks of gain with little consideration for the fact that inside was a real, live collection of electronic bits which still had all the problems 'real' electronics always had. The reason the *741* was relatively impervious to user inflicted nasties is analogous to the fact that it's quite difficult to get anything that is bound, gagged and set in treacle to *not* behave itself.

Mistake number one with the new devices was believing that they were unity-gain stable because the data sheet said so. What that *really* means is 'does not burst into oscillation at unity gain' which is not the same thing at all.

It is important to maintain as large a margin as possible between the internally structured gain/bandwidth rolloff set for open loop and the rolloff around the external circuitry determining the closed-loop gain. This is in order to preserve sufficient phase margin at all frequencies the circuit has gain. Failure to do this can result in the feedback being shifted in phase sufficiently to become reverse-phase to that intended, positive feedback, and oscillation resulting. Even if the phase isn't shifted quite that far, the feedback is tending toward positive,

FIG 14a-b

FEEDBACK - PHASE - LEADING STABILITY COMPENSATION




Basic external feedback compensation (a)

Virtual-earth mixer with stray pinout & buss capacitances (b)

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and damped ringing when transients hit the circuit is a possibility. Also, these resonance effects are extremely high in frequency, typically many MHz, so any radio signal that gets as far as the circuitry will absolutely adore an amplifier that is critically resonant at its frequency! A reasonable phase margin to aim for at all gain frequencies is better than 45°, but in practice a compromise between desired circuit bandwidth traded-off against the need to tighten that bandwidth for phase-margins' sake can be fairly easily reached with the newer devices.

The normal and most flexible way to determine a circuit's closed-loop rolloff is by means of a feedback phase-leading capacitor across the main output-to-inverting-input feedback resistor, a typical arrangement is shown in **fig 14**.

A fairly common eroder of phase margin and progenitor of instability is stray capacitance from the amplifier's inverting input to ground. This capacitance, a combination of internal device, pinout and printed-circuit layout proximity capacitances, reacts against the feedback impedance to *increase* the closed loop gain at high frequencies. In normal circuits, even the typical 5 or so pF is enough to tilt up the closed-loop response well within the open-loop gain parameters, threatening stability. Far worse is the situation where the inverting input is extended quite some distance along wiring and a buss, as in a virtual-earth mixing amplifier—hundreds of pF may be present there. It can arise that despite a sizeable time-constant being present in the feedback leg, none of the expected hf rolloff occurs since it is merely compensating for the gain hike created by the buss capacitance. Ensuring required response and phase characteristics using any virtual-earth mixer can only be done properly with the finished system up and running *complete*, since any additional sources modify the impedance seen by the buss.

A small limiting resistor to define just how much this unwanted gain can rise may be added as close to the amplifier inverting input terminal as possible, but this is at the expense of the 'virtual-earth' point now having an impedance based on the value of that resistor. The resistor, incidentally, is also a measure of protection against rf on the buss being rectified by the input stage's junctions.

## Time domain

There is invariably a finite time taken for a signal presented at any amplifier's input to show an effect at the amplifier's output—the so called transit-time. This transit-time, as the frequency increases, becomes an appreciably greater proportion of the signal's wavelength and as such has to be taken into account due to its detraction from phase margin with increasing frequency.

Remember the great hoo-hah a few years ago about Transient Intermodulation Distortion? The effect that collected this name is due nearly totally to amplifier transit-times and not surprisingly, as is nearly always the case with 'fad' problems, has been known about and appreciated for as long as there have been negative feedback amplifier circuits—some 60 years. It is and always has been totally predictable.

TID is a direct result of the 'servo' nature of an amplifier with a large amount of negative feedback that is intended to provide a correction signal derived as a difference between the amplifier output and the applied input signal. Since there exists a time delay in the amplifier, the circuit has to 'wait' for that time before its correction signal arrives—the output during this time is uncontrolled and just flies off in the general direction the input tells it to. Once the correction arrives, the amplifier has to wait again to find out how accurate *that* correction was, see-sawing on until the amplifier output settles. Fortunately this all takes place rapidly (dependent on the amplifier external circuitry) but it still represents a discrepancy between input and output. It is an effect peculiar to amplifiers with large amounts of negative feedback (as is typical of most contemporary circuitry), it quite frequently displaying itself audibly especially in power amplifiers where the transit-time is quite long with the usual huge, slow output devices.

Amplifiers which rely on their own basic linearity, such as valve amplifiers, rather than on a servo non-linearity correction system are often held to be subjectively 'smoother', this certainly being a principle reason. Nowadays, though, with device speeds as they are, settling times are becoming insignificant in relation to the signal transients they are expected to cope with, so hopefully this nit has been well picked.

## Output impedance

Most newer devices, particularly the 'TLO' series of BI-FETs have a quite significant open-loop output impedance which although, by virtue of the enormous amount of feedback used, normally gets reduced to zero, is still present and included as part of the feedback path. Obviously, then, any reactive element at the output is going to materially affect the feedback phase and phase margin. And it does!

Any capacitance from the output to ground will form a feedback phase-lagging network, shifting the phase inexorably toward the point where the total amplifier and network phase shift reaches 180° at the inverting input (therefore a full 360° total) and the circuit oscillates. The frequency at which it oscillates is inversely relative to the capacitance value—it isn't unusual to find oscillations right at the edge of an oscilloscope's high frequency sensitivity with small values. Hanging a long bit of wire on the amplifier output (especially screened cable with its high screen to inner capacitance) is a surefire guarantee of instability for this very reason, with the added complication that there is a measure of inductance there, too. If you're really lucky, a long cable might start to look like a mismatched tuned stub at a frequency where the amplifier still has some gain. As a good stable rf signal generator it could probably win awards, but in a mixer . . . ?

Fortunately a simple cure for this is to buffer away the load from the output/feedback termination with a small resistor of typically 33 to 150Ω. This usually does it, but at the expense of headroom loss due to the attenuation from the buffer resistor against the load termination. Provided the load is greater than about 2kΩ, which it would really have to be in order to prevent getting close to current drive limiting in the IC output stage, this headroom loss should not exceed 0.6dB. An altogether more elegant way is to buffer off with a small inductance, giving increasing isolation with frequency and a phase shifting characteristic opposite to that of the (normally) capacitative load providing a total termination that is phase-constant at the higher frequencies. At the lower audio frequencies, of course, the inductive reactance is very low and the load sees the very low dynamic output impedance of the amplifier.

Both of these techniques also provide a measure of protection against the possibility of rf finding its way into the amplifier by means of rectification in the output stage or inverting input.

Some devices with a quite low output impedance before applied feedback (say those with unbuffered complementary emitter-follower output stages) are not likely to be phased as much by these problems (pun totally intentional) but it is just as well to habitually design in these considerations.

## Voltage followers

The above precautions, in addition to the feedback phase-leading capacitor, are now required circuit practice for using the newer fast devices in many op-amp configurations. It should be said here that because there is no facility for implementing phase-leading around the standard voltage-follower configuration and that this is the most criticial configuration for stability, it is not a preferred circuit element. The manufacturer will have designed the IC to be just stable enough at unity gain to be able to say so unblushingly. All hanging a compensation capacitor across the appropriate pins will do is slow up the slew-rate—better not to tempt fate.

Some good news. If it is the internal stage around which the external compensation capacitor is hung which is tending to instability, then the capacitor should cure it. Now the bad news. It rarely is that stage. If a previous stage, say the input differential amplifier is unstable, all the capacitor will do is slow up the amplifier and reduce the slew-rate to the extent that the oscillation is no longer *visible* at the output. It does not cure the instability. It's still there, hiding.

The use of a standard voltage-follower implies that in order to maintain the same system headroom in that stage, the input has to rise and fall to the same potentials as the output is expected to. It can't. In most op-amps, especially those with bipolar inputs, the differential input stages saturate or bottom significantly before the power supply rails are reached, which means that they not only cease to follow, but also will spend a considerable amount of time in unlatching. Once an amplifier internal stage has latched the feedback loop is broken and that stage has no assistance from the servo mechanism to unstick itself. Once the loop is re-established it has to settle again as if from a hefty transient before it can resume 'following'.

The IC manufacturers commonly specify the common-mode input voltage range, and it is precisely this limit that would be exceeded in use as a follower. For reference, it is ±13V for the *5534*, ±11.5V for an *LM318* and +15V to −12V for a typical BI-FET. All fall far short of the supply-rail maxima. Provided enough gain is put around the amplifier to prevent these common-mode limits being reached, there should be no latching hang-ups and the feedback network will also provide some 'meat' to hang closed-loop compensation around in addition to allowing the amplifier's full output voltage swing to be utilised.

Similar settling-time problems occur any time any stage is driven into clipping, but given the high supply-rail voltages, hence large headroom common today, clipping should be rare. Shouldn't it? ■

# Designing a professional mixing console

# Part Four

Steve Dove

## The Mixer Front-End

Like a sausage machine, a console is expected to accept any scraggy fodder in the way of input level and impedance whilst producing a nice uniformly consistent output capable of being deposited in the tightly-defined container that is a tape track.

Fortunately, industry standards provide at least some clues as to what mixers are likely to have stuffed up them. Nevertheless these standards can obviously do nothing to alter the physics of the operation of the assorted transducers and sources used and the disparity in the treatment required for say, a dynamic mic and a tape machine output totally precludes a 'universal' input stage.

Mixer front-end design tends to be a little like working on a grown-up jigsaw puzzle where all the important pieces perversely refuse to fit. It's really delightful to find some that fit beautifully — as in line-level input stages. This euphoria is ground away by the problems inherent in other areas — notably mic inputs.

Optimising noise performance in a dynamic microphone preamp is a performance, juggling a seemingly endless number of variables. A dynamic mic may be represented (a little simplistically) as a voltage source in series with a fairly lossy inductance representing an impedance midband typically of between 150 and 300Ω (**fig 15**). Being a transducer and, of necessity, mechanical in nature, many complex varying motional impedance effects contribute to the overall scene, but for most design purposes the specified electrical analogue suffices. The low impedance is primarily to mitigate high-frequency attenuation effects due to inevitable cable capacitance, which in practical circumstances mounts up to horrifying values of capacitance that the transducer must drive along with its load. Unfortunately the impedance is not low enough that it may be treated as a pure voltage source; there exists a tiny signal at a finite impedance that must be daintily ferreted out for optimum performance.

The jigsaw commences. Textbooks on electrical theory quite correctly state that to extract maximum *power* from a given source the optimum load is equal in value to the source impedance matched. This however, in the instance of a dynamic microphone, is of doubtful (if any) value — OK, we've squeezed all the energy possible from the generator but to what end? Given that electronic amplifiers of the type useful in low-noise applications are of relatively high input impedance (ie voltage amplifiers) and that the terminating resistance that largely defines the microphone's load is in fact dissipating most of our hard-won power. It is the source's output *voltage* capability that is of greatest value, not the power. So as can be seen in **fig 16**, 'matching' does a very effective job of sacrificing 6dB of signal level which naturally has to be made up in the succeeding amplifier. This does not imply that the noise performance is 6dB worse than possible since the source impedance as seen by the (assumedly perfect) amplifier is now a parallel of the mic and it's matching load, hence half the value of either. The thermal noise generation of this combined source is consequentially 3dB less, hence the noise performance is only degraded 3dB by such a termination. Still, who wants to throw away a good 3dB before we even start hassling with the amp?

Another good reason for not terminating with an equal or any fairly low resistance is the effect on microphone response and subjective quality. Having an inductive characteristic, the dynamic microphone capsule has an impedance that rises with frequency, predominantly at high audio frequencies where the inductive reactance of the source becomes large with respect to the coil winding resistance. When terminated with a relatively low resistance, the complex impedance of the capsule and the termination resistor form a single-order 6dB per octave lowpass filter, gracefully rolling off the top.

Combine that with a fairly hefty cable capacitance and you may delete the 'gracefully', since the complete network now looks like a rather lossy second-order filter. Still, regardless of termination method, we're stuck with cable capacitance — it's always a consideration.

All right. No termination resistor. The way to go is obviously as high a termination impedance as possible — but oh-oh, jigsaw time.

### Optimum Source Impedance

Amplifiers are not perfect. For noise criteria, the first device that the signal hits in the amp is the key one since the noise it generates usually masks, by a large margin, noise due to all succeeding stages.

All practical amplifying devices are subject to a variety of internal noise generating mechanisms including thermal noise generation, and this when measured gives rise to some important values, namely input noise voltage and input noise current.

For the most part, ordinary bipolar transistors are used as front-end devices both in discrete designs and op-amp IC packages so much of the following is specific to them.

These noise voltages and currents alter in both magnitude and ratio to each other with differing electrical parameters, especially collector current. Predictably, as this current decreases, so does the noise current (most of the noise is due to minor random discontinuities in device currents) and so the ratio between the noise voltage and current — or noise impedance — may be altered.

Thermal noise generation is common to all resistive elements, its amount being related to temperature and the bandwidth across which it is measured, an increase in either increasing proportionally the noise power generated. Under identical circumstances the noise *power* generated by any values of resistance is the same — differing resistor values merely serve to create differing ratios of noise voltage and noise current, the product of the two always equalling the same noise power. This particular noise phenomenon is totally unavoidable, it being the nature of atomic structure that when things get hot and bothered they grind and shuffle about randomly, creating electrical disturbances white in spectra (ie equal energy per bandwidth).

Even the real (resistive) part of the complex impedance of a dynamic mic generates thermal noise and this ensures that there is a rigidly defined noise value that cannot be bettered.

The difference between the noise floor defined by this thermal noise and the measured noise value of a practical system is known as the Noise Figure (NF) and is measured in dB. The noise output from a resistor is predictable, so a direct comparison of the noise voltage measured at an amplifier's output from a resistor applied to the amplifier input and the noise voltage expected of the resistor on its own is possible just by simply subtracting the measured gain of the amplifier. A measure of NF.

An interesting effect occurs when, with any given set of electrical parameters set up for the amplifier front-end device, the source resistance is steadily changed in value. A distinct dip in the NF occurs, see **fig 17**, and the value of resistor at which this dip occurs changes as the device parameters are changed (collector current primarily). For the usually predominant noise mechanism (thermal noise) a minimum NF occurs with a tiny amount of collector current (say 5 to 50μA) and a high source resistance (50kΩ up). Without diving into the mathematics, the nulling is balancing interaction between the external noise source and the internal voltage and current noise generators.

There is another major noise mechanism inherent to semiconductors, however, which is low frequency or lf noise — a burbly, bumping type noise caused by the semiconductor surface generating and recombining sporadic currents — most prevalent in 'dirty' devices but present to a degree in all. It is subjectively apparent and has to be considered. Measured alone, lf noise has its own set of collector current and source resistance nulls, usually far higher in current and lower in resistance than for thermal noise.

A compromise has to be struck. To make a generalisation, 100μA and 10kΩ for a typical low-noise PNP transistor seem about right. (PNP transistors are common in this area due to marginally better lf figures over NPN types.)

The source resistance value is that at which the device is optimally quiet for audio purposes and is known (surprise) as the Optimum Source Impedance. Incidentally, this impedance has absolutely nothing to do with the kind of circuit configuration the device may be in — whether it be in a common-base amplifier with

FIG. 15 SIMPLISTIC MODEL OF A DYNAMIC MICROPHONE.

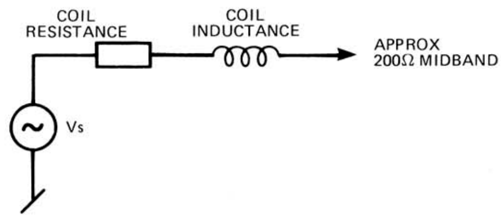



FIG. 16 'MATCHING' IMPEDANCES CAUSING LOSS.

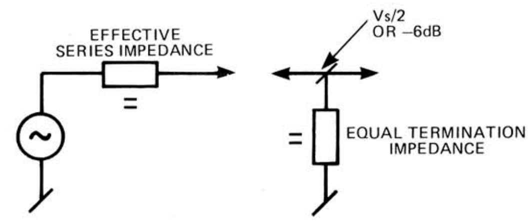



FIG. 17 NOISE FIGURE CURVES FOR A GOOD PNP FRONT-END TRANSISTOR FOR COLLECTOR CURRENT VS. SOURCE RESISTANCE.

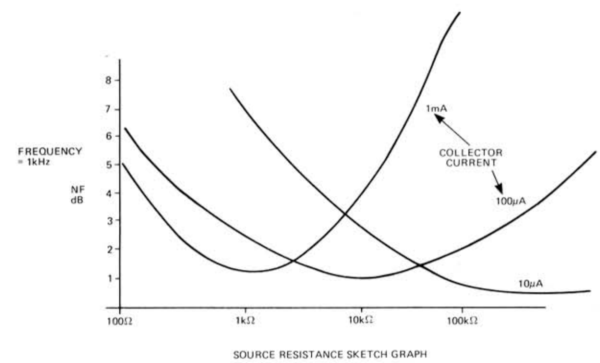
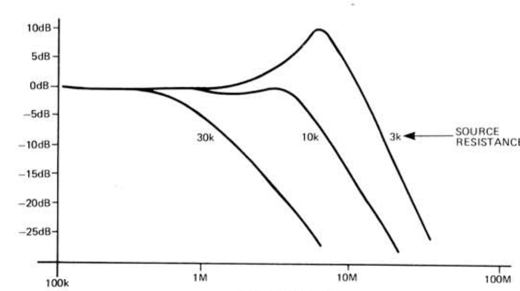



FIG. 18 GAIN VS. FREQUENCY FOR A TYPICAL FOLLOWER-CONNECTED OP-AMP HIGHLIGHTING EFFECTS ON RESPONSE OF SOURCE IMPEDANCE ON INPUT DEVICE.




an input impedance of 50Ω or in a totem-pole front-end with bootstrapping and a consequential input impedance of over 10MΩ — it doesn't matter. The source impedance for optimum noise performance stays at 10kΩ, or whatever, provided the collector current is the same in all cases. Optimum source impedance has nothing to do with input impedance.

This optimum impedance varies dependent on the type of input device used. For a field-effect transistor, the noise figure typically obtainable drops to an amazingly low value but unfortunately at an impedance of several dozen MΩ. Even supposing it was practicable to provide a source of that magnitude the whole arrangement would be so sensitive to any electromagnetic fields (such as RF) that even tiny amounts present would obliterate the noise advantage. The design and construction of capacitor microphones using FET front-ends highlights the hazards. The end results, more often than not, show such capacitor mics to be several dB noisier than a well-designed dynamic microphone/front-end combination.

Good bipolar transistors have OSIs in the region of 5 to 15kΩ, whether discrete or as part of an IC amplifier package. Ah! A piece of jigsaw that actually fits! By happy accident, these values closely coincide with the source resistance value that provides for optimum flatness of device transfer characteristics which helps a long way towards best frequency and phase linearity, hence stability in a typical high negative-feedback amp configuration.

**Fig 18** shows the effect of altering the source impedance into such an amplifier (using a conventional bipolar transistor input device) on output frequency response. The drop is due to the excessively high source impedance reacting against the device base-emitter and wiring capacitances to form a lowpass filter. The hf kink is a practical effect of the curious mechanism described last time, ie when a bipolar transistor is fed from an impedance approaching zero, its high frequency gain/bandwidth characteristic extends dramatically, radically altering the phase margin and consequentially the stability of an amp designed and compensated for more ordinary operating circumstances. The kink is a resonance within the amplifier loop caused by erosion of phase margin resulting from this mechanism, being only a very short step from oscillation.

As can be seen from the sketch graph, the response is maximally flat at a source resistance of around 10KΩ, about the same value as the OSI for noise performance.

A problem to reconcile. Our practical source impedance is nominally 200Ω for a dynamic microphone. The OSI for the best conventional input devices is around 10KΩ. How do we make the two fit?

## Microphone transformers

Please, don't go away. OK, you've heard some horrible stories about how nasty they are, but properly designed and used they do offer a good solution to the impedance matching and other problems.

Simplistically, a transformer is a magnetically soft core around which are two windings, the voltage ratio between the two being equal to the ratio of the number of turns on them. The impedance ratio is the turns ratio squared (eg a 10:1 turns ratio corresponds to a 100:1 impedance ratio) because power output cannot exceed power input and if the voltage is stepped up 10 times, the output current must be stepped down 10 times. Impedance, which is the ratio of voltage to current, is consequentially the square of the transformed voltage or current ratio.

Given this, it is a simple matter to calculate the ratio necessary to match the microphone impedance to the OSI that is actually being used and to a much lesser degree on the actual impedance of the microphone intended for use. Since few people are intense enough about the whole affair to bother measuring microphones, the assumption that 200Ω is a good mid-point serves well. The assumption that most bipolar input amplifiers have an OSI of between 5kΩ and 5Ω indicates that the transformer ratio should lie somewhere between 1:5 and 1:8.7.

Many consoles, notably some American ones, quite often use higher ratios (typically 1:10) probably in the naïve belief that the noise advantage of a step-up input transformer stems from the 'free gain' it affords, so the more the merrier. Although on a basic level it would seem to make sense that the less electronic gain you need to use the quieter the system must be, this fallacy completely belies the truth that the transformer merely allows you to *choose* and *alter* the impedance at which your amplifier is optimally quiet. Increasing the turns ratio makes the amplifier noisier.

Actually the 'free gain' can be more of a nuisance than a benefit. It is not unusual for mic inputs to receive transients exceeding +10dBu and mean levels of −10dBu especially in a nasty rock and roll environment. Even dynamic capsules can deliver frightening levels and this can pose headroom problems in the mixer front-end. A typical 1:5 transformer has a voltage gain of some 14dB (at 1:10 some 20dB) which would mean that even with no electronic gain after the transformer, normal mixer operating levels are being approached and exceeded. These circumstances make worrying about a dB or two of noise performance total nonsense and it just serves to point out that our poor mic front-end has to be capable, if not perfectly optimised for, elephant herds as well as butterflies.

Transformers have numerous limitations, inadequacies and problems resulting from their

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## Mixing console

physical construction that make their actual performance differ (in some respects radically) from that expected of the theoretical model.

### Core material

The heart of the transformer is the magnetically pliable material into and out of which the energy is induced. Virtually any material: nickel, steel, iron, ferrous derivatives and substitutes have the same basic limitations; saturation at a magnetic level beyond which they are incapable of supporting further excursion, and hysteresis — a crossover like non-linearity at low levels responsible for a significantly higher distortion than anything else likely to be found within a signal path.

These two effects at opposite ends of the dynamic spectrum mean that any transformer has a well-defined dynamic range within which it must be operated and which is quite less than the range of levels it would be expected to pass in use as a mic amp. This is especially true at low frequencies where the core is prone to saturation far earlier. Optimisation begins here. Is it to be designed for minimum hysteresis — hence low low-level distortion (butterflies) — or with lots of material tolerant of high signal levels (elephants?).

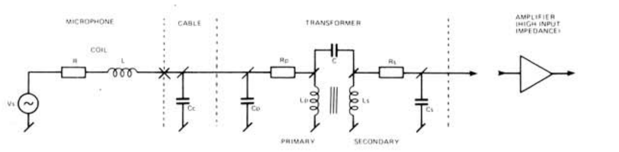
### Winding resistance

Windings are made of wire. Wire has resistance. Resistance means loss and lack of efficiency and noise performance. By the time there are enough turns on each of the windings to ensure the inductive reactances are high enough not to affect in-band use, the winding resistances can no longer be ignored.

### Winding capacitances

Capacitance exists between any things in close proximity, and that includes transformer windings — between each other, between adjacent turns and piles in the same winding and from the windings to ground. In this given instance it is nothing but bad news — capacitance

FIG. 19 TRANSFORMER COUPLING SHOWING MAJOR ELEMENTS




between windings means unwanted leakage and imperfect isolation, whilst winding self-capacitance reacts with the winding inductances to form resonances. Resonances, even if way out of the audio band, invite response trouble, also disturbing in-band phase linearity. Combinations of these capacitances greatly affect one of the most touted advantages of transformers, Common Mode Rejection (CMR).

### Common mode rejection

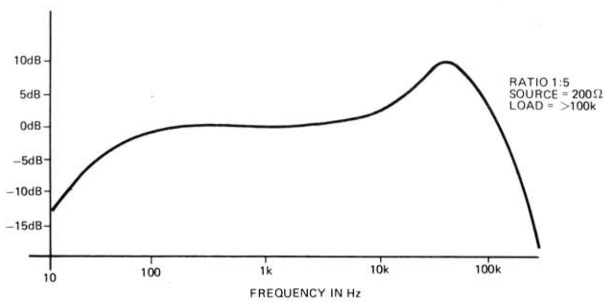
This is the ability of the transformer to ignore signals present in identical amplitude and phase on the two input legs, not transferring them across the secondary as differential information.

Principally, it is imbalanced distribution of capacitance along the length of the two windings, both with respect to each other and to ground that makes CMR less than perfect. Co-winding capacitance has the effect of directly coupling the two wiring masses together permitting common to differential signal passage worsening with increasing frequency at 6dB/octave. Electrostatic screening (a Faraday shield) between the windings helps alleviate, but certainly does not eradicate, co-winding capacitative coupling.

Further CMR worsening can be expected even if the two windings are perfectly balanced with respect to each other, if the primary winding is not end-to-end capacitatively matched with respect to ground. Any common mode signal from a finite impedance source (almost always the case) when confronted with such a capacitatively unbalanced winding sees it as being just that — unbalanced (becoming more so again with increasing frequency) hence again transferring input common mode signals across to become output differential information indistinguished from the wanted input differential source.

Broadcasters particularly are concerned with winding balance, not only on microphone transormers but on line-output transformers too, reasoning that common/differential transference is as likely to occur at a source as at an input.

FIG. 20 TYPICAL TRANSFORMER TRANSMISSION RESPONSE




### The real thing

**Fig 19** gives a better idea of what our poor little microphone's signal has to suffer. The winding capacitances (Cp and Cs) form lovely resonances with the inductances, whilst the transformed-up primary winding resistance (rp) added to the secondary's resistance (rs) merely serves to increase the microphone's effective source impedance, hence inefficiency.

The frequency response of the transformer fed from a 200Ω source and measured at high impedance across the secondary looks something like **fig 20**, the lf droop attributable to one or both of the winding inductive reactances becoming relevant to signal impedances, whilst the hf peak is the aforementioned secondary winding self-resonance. Usually the primary self-resonance is fairly well damped by the source impedance but occasionally added cable capacitance can play cruel tricks here, too.

### 'Zorbal' network

The mic amp itself, as discussed, has a pretty high input impedance (MΩ and up) whilst its optimum source impedance is quite defined at around 5 to 15kΩ (if your head still doesn't hurt too badly from accepting that something can simultaneously have two impedances).

It's good engineering practice to consider how the circuit behaves when the operating impedances are no longer defined by the mic ie when it is unplugged. Ordinarily, the sketch circuit of **fig 19**, with the mic disconnected, would probably scream merrily away in oscillation, as would any circuit with a high gain, high input impedance amp terminated only by the collection of vile resonances and phase shifting elements that are an open-circuit transformer. An open-circut impedance-defining resistor (Ro in **fig 21**)

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FIG. 22 FREQUENCY RESPONSE OF ZORBAL NETWORK

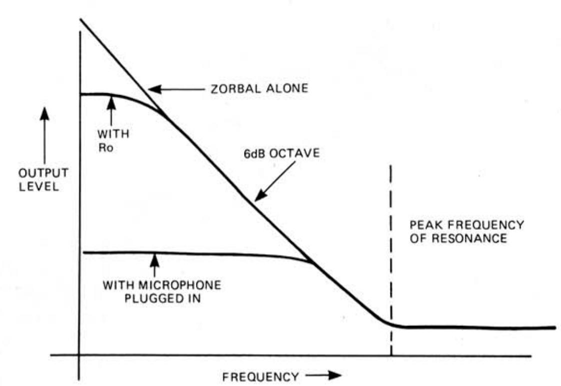
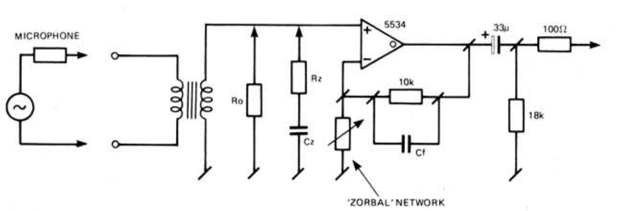



FIG. 21 BASIC MICROPHONE PREAMPLIFIER SHOWING COMPENSATION COMPONENTS



# Mixing console

of a value 10 or 20 times that of the amp OSL helps tame this, also marginally taming the secondary resonance.

There are a variety of techniques for dealing with this resonance, varying from pretending it doesn't exist through to actually *using* it as part of a front-end lowpass filter. The 'usual' way is to try to eliminate it as much as possible passively prior to the amp, the 'Zorbal' network in **fig 21** representing a typical approach. This is a series resistor/capacitor combination in conjunction with the open-circuit impedance defining resistor. They are calculated to produce a 'step' type response **(fig 22)** which when combined with the hump at the hf end of the transformer response produces a more acceptable roll off characteristic. Naturally, the inter-reaction between this network and the transformer's complex impedance is not quite that neat and tidy, the network capacitance reacting heavily with the transformer inductance, shifting the resonance frequency in the process. It is this which has led to the misconception that the capacitance somehow magically 'tunes out' the resonance.

Open circuit stability is dramatically improved since, as the sketch response drawing **fig 22**, here the network takes an even larger slice out of the overall hf response, keeping impedances at the nasty top end comfortably low.

From another tack, providing the compensating hf roll-off around a subsequent amp in the form of exaggerated feedback phase-leading, even around the mic amp itself (cf), has the advantage that the combination's noise performance at the higher frequencies remains unimpaired by an impedance mismatch resulting from a passive network.

Problems result in several areas. Compensation around the mic amp becomes limited when the electronic gain approaches unity whilst compensation around a late fixed-gain stage means that all stages prior to it, including the mic amp, have headroom stolen at the frequency of the resonance and to a degree of the magnitude of the resonance. This may or may not be a problem dependent on how far the lower side of the resonant curve invades the audio band.

The passive method reduces the magnitude of the resonance, the ultimate lowpass roll-off slope being that of the hf side of the resonance (which more accurately is a lightly damped LC lowpass filter anyway) which is some 12dB/octave. The active method uses an additional 6dB/octave curve in the compensation making a total 18dB/octave, but relies on the resonance being of a manageable quantity to begin with. Consequentially, a measure of both techniques is usually required, their balance and relationship being a long, long experimental process to optimise — for each different type of transformer at that, too.

This enforced filtering is of considerable advantage, helping to keep all sorts of unwanted ultrasonic garbage from finding its way into the mixer, and represents a major advantage of transformer inputs over solid-state varieties.

Further advantageous 'filtering' is the falling source impedance seen by the amplifier at extreme lf due to the winding inductive reactance reducing with frequency — this definitely helps combat the generation of excess lf noise.

There are regretttably two different amplitude response curves to be considered: one, the normal differential input, we have fairly thoroughly determined. The second, by virtue of its mechanism relying on imperfections within the main filter element itself — the transformer — rides completely roughshod over and oblivious to our carefully calculated filter responses. Common mode unrejected signals still appear at the amplifier input as if nothing had happened. Sick, isn't it?

FIG. 23 FRONT-END ATTENUATION-WORSENING NOISE FIGURE

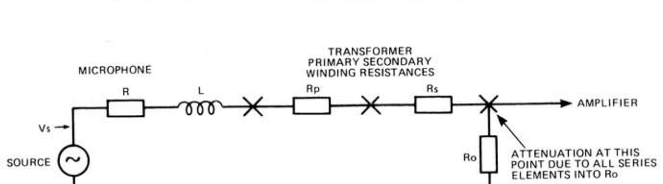
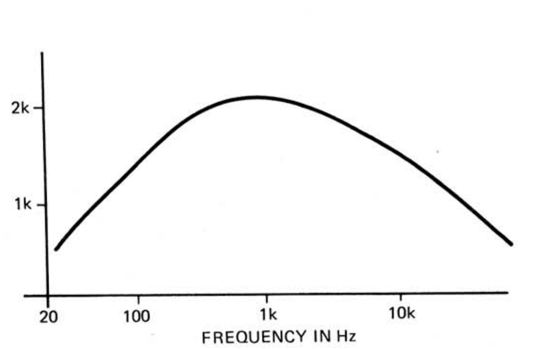



FIG. 24 TYPICAL INPUT IMPEDANCE CURVE




## Input impedance

As we determined earlier, we would end up with better noise performance and cleaner sounds if the microphone looked into a high, preferably infinite, impedance. Preferences apart, we have already had to define the reflected load (input) impedance by the resistor needed to keep the front-end stable under unplugged conditions (Ro) but at least it is an order of magnitude and above working impedances so its effect is small. It does, though, act as part of an attenuator of input signals along with the source impedance and winding losses **(fig 23)**. This is the major factor responsible for worsening front-end noise performance using transformers — any attenuation before the optimised amp directly degrades the noise figure, typically between 1dB and 6dB dependent on the transformer.

If the transformer was perfect, it could be assumed that the reflected impedance as seen by the microphone would be constant over the audio band — wrong! At the lf end **(fig 24)**, the diminishing inductive reactance (it tends to zero with frequency) becomes a term of greater importance, affecting parallel impedances, attenuation, and hence efficiency. Winding self-capacitances and the passive compensation networks are largely to blame for the hf droop although the list of contributing mechanisms is nearly endless.

A good rule of thumb is that the midband input impedance should exceed 10 times the source impedance, say 2KΩ. Any wild variation in this impedance is obviously going to result in frequency and phase response aberrations — this is probably the greatest single drawback to transformer front-ends.

Attenutator pads, regrettably necessary in many instances to preserve headroom and prevent core saturation with 'elephant' sources should maintain expected operating impedances when introduced. The transformer primary should still be terminated with a nominal 200Ω whilst the microphone should still look at 2KΩ or above. Departure from these will cause the microphone/ amplifier combination to sound quite different when the pad is thrown in and out, as would be expected from altering source and load impedances and complex filter characteristics.

## Transformerless front-ends

Bringing the amplifier optimum source impedance down to that of conventional dynamic mics is possible by means other than transformers. Reducing the ratio of amplifier-inherent voltage and current noises has this effect, being managed by somewhat of a fiddle — namely paralleling up lots of identical input devices — maintaining about the same noise voltage but proportionally increasing noise current therefore reducing the ratio between them (ie noise impedance).

The usual technique is to place two of these multi-device input front-end amps ahead of an electronic differential amplifier, as in **fig 25**. All the amplifier gain is made within the first pair of stages, differentially cross-coupled. This gain arrangement, rather than referring to ground, greatly assists the ignoring of common-mode signals. Differential input signals are amplified since the reference for each of the two amps is the other amplifier, tied to an

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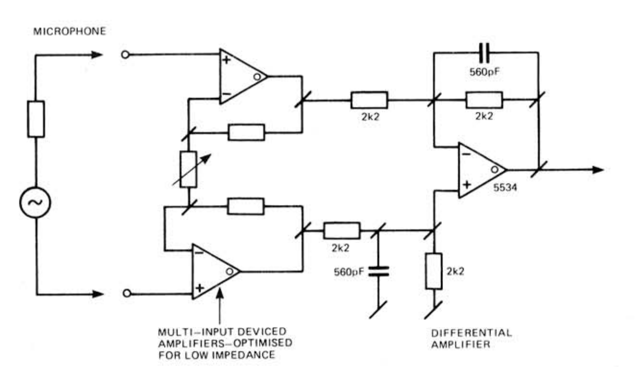
FIG. 25 BASIC TRANSFORMERLESS MIC-AMP ARRANGEMENT



# Mixing console

identical signal of opposite polarity.

If the input signals to the two amps are identical in phase and amplitude (common) the references for each of the amps are waving up and down identically to the signal — there is no voltage difference for the individual amps to amplify: presto, no gain. For ordinary differential input signals, the amplifiers operate conventionally, their 'ground' reference being a zero voltage point half way along the gain-determining variable resistor. This point is a cancellation 'null' between the opposite sense polarity swings of the two amps.

These amplifiers feed a conventional electronic diff amp running (usually) at unity gain, and in order to maintain stage noise as low as possible, the resistors are made as low in value as the devices can sensibly stand. Optimisation of impedances is not necessary since the outputs of the front-end pair can be assumed to be feedback zero impedance. This arrangement is unmistakably a bastardised 'instrumentation amplifier' which is a well-documented circuit configuration, the only thing of remark being the low impedance optimised front-end stages.

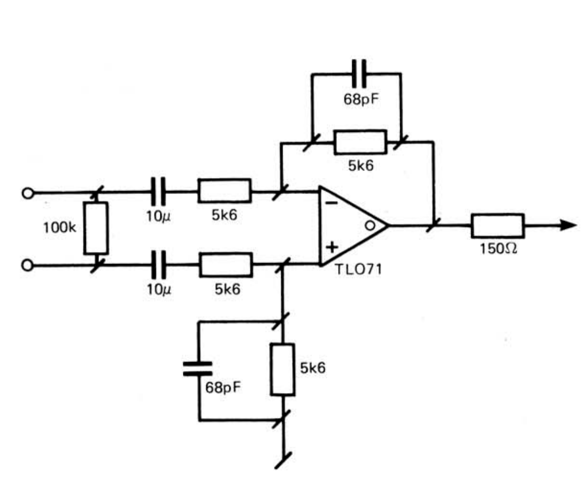
Although potentially offering far higher and flatter input impedances than transformer inputs there are, as always, hang-ups. Common-mode signals directly gobble up headroom in the first pair of stages even if those are operating as followers and they are subsequently cancelled in the diff amp. There is also the great danger that common mode signals (in addition to normal differential signals) can exceed the input swing capability of the input devices.

RF adores base-emitter junctions, this configuration giving it lots to play with. Filtering mic inputs sufficiently without sacrificing noise performance or input device hf gain (hence hf distortion, etc) is not a simple task.

Line inputs are commonly simple differential amplifiers, rather than unity-ratio transformers, similar to the one used in the transformerless mic amp, but with the resistor values elevated to bring the differential input impedance up to over the 10KΩ required of a bridging termination **(fig 26)**. Noise of these stages is directly attributable to these resistor values, so the lower the better. An instrumentation amp configuration would seem to offer possibly better performance for noise (the diff amp resistor values may be small) but entails the use of undesirable voltage followers (see Part 3, November issue) with potential stability problems, voltage swing limitations and unprotected (for RF) input stages. At least with a simple diff amp the impedances are comfortably low and the inputs buffered by resistors from the nasty outside world.

The dc blocking series capacitors have, unfortunately, to be large in value to maintain an even input impedance at the lowest used frequencies and, being necessarily unpolarised, physically large and expensive. A small price to pay, though, for such a delightfully simple but important circuit element. ■

FIG.26 AN ELECTRONIC DIFFERENTIAL INPUT AMPLIFIER




# More on Op Amps

Please note that **fig 12** was wrongly included in Part Three of this article (November issue) and is referred back to in Part Four. The reference to **fig 12** on page 63 should read **fig 13e**. We apologise for this error.

## Front-end instability

Altogether the most obscure potential instability causing effect relates directly to the behaviour of the input stage in bipolar front-end op-amps. The gain/bandwidth characteristic of the input differential stage is greatly dependent upon the impedance presented to the input, the gain/bandwidth increasing with reducing source impedance. There is the possibility that given an already critical circumstance, the erosion in phase margin due to this effect can cause instability. This can be mitigated by limiting the gain/bandwidth excursion by means of a resistor, typically 1kΩ, in series with the input. Ordinarily, this would have little effect on circuit performance but may, especially in say mic-amps, detract from noise performance which is largely dependent on the amplifier being fed from a specific source impedance, of which 1kΩ would be a sizeable proportion. However, it's usually fairly easy to arrange that the IC doesn't have a zero impedance at either of its inputs in the design stage.

Fortunately, this is a problem that FET-input op-amps do not have, owing to the far greater isolation between the FET gates and their channels.

A similar approach to that proposed for output isolation ie an inductor rather than a resistor, in series with the affected input is, on the surface, an equally good idea. The inductor's impedance would be low at audio (so not affecting noise criteria significantly) and high at rf where the low source impedance phenomenon does its dirty work. Unless the value is critically defined, an inductor of sufficient value to provide a usefully high reactance at rf is also likely to be self-resonant with circuit stray and its own winding capacitances at a frequency probably still within the gain/bandwidth capability of the amplifier. Attempting to solve one potential instability by introducing a resonant tuned circuit doesn't really win many points.

## Band limiting

One of the first great superficially appealing results of using the enormous feedback inherent from using op-amps at the relatively low gain requirements of the audio world was a close approach to 'dc-to-light' frequency response. The author well remembers hysterical peals of laughter as a new mixer's response was measured as still ±0 right to the end of the testing oscillator's ranges and the badly disguised puzzled looks and worried glances when we listened to it.

Those who have experienced design with discrete circuitry will not be surprised that this source impedance instability effect is also the reason emitter-followers were the most instability prone of the three basic transistor amplifier configurations, also that the cure was the same. Not only does the series resistor limit the source impedance before zero, it also acts together with any pinout and base-emitter capacitance as a lowpass filter helping to negate further external phase-shift that may detract from stability.

This base source-impedance instability is quite insidious in that it can either contribute to instability of the amplifier loop if it is already critical or be a totally independent instability local to the affected devices—nothing whatsoever to do with the characteristics of the external loop.

Most audio signals, especially live ones from microphones and tape machine returns with a high bias content, have present a fair amount of ultrasonic energy which would remain ultrasonic if it weren't for the progressively worsening linearity and propagation delay of the amplifier effecting cross-modulation of these out-of-band signals down into the audible range. The linearity worsens because the amplifier's open-loop gain is falling rapidly with frequency enabling less and less feedback to be used (the feedback being that which is keeping it linear anyway) whilst the finite transit-time of the amplifier becomes a significantly greater part of the signal's period as frequency increases, meaning that servo-loop inaccuracies in the circuit assume greater proportions in the output signal.

Slew-rate limiting occurs when the fastest signal rise-time the amplifier is expected to pass exceeds the speed of the fastest stage in the amplifier giving rise to intermodulation effects that are dependent upon both frequency and signal level. A common subjective result of this limiting is for the high-end of a drum-kit to change in character of sound with differing levels of lower frequency instruments on which it is 'riding'. Another favourite is the 'disappearing snare drum' in which again the sound radically alters with changing level.

So much for the expected result of 'improved transient response' through having a wide-open frequency response. As is now obvious with hindsight, deliberately limiting the mixer's input frequency response to a little more than the audio band results in an amazing clean up of the sound. By removing a lot of the inaudible signals that cross-modulate within themselves, and with in-band signals, you eliminate the cause of much of the lack of transparency and mush that had become the trademark of first-generation IC op-amp consoles.

Despite improved devices, this remains valid today. By band limiting the programme signal as

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early on in the chain as possible to reduce inaudible signals, there is far less chance of them generating unwanted audible products. A front-end single order lowpass filter, operating in conjunction with all the other lowpass effects of feedback compensation arrangements throughout the console should provide adequate minimisation of these products, given modern devices.

'Purist' arguments about the undesirability of any deliberate filtering seem rather futile in a world of real devices and final signal destinations such as tape (with its generally anything other than linear phase and frequency characteristics), disc (oh, disc), radio (rapid filtering above 15kHz) or digital processing/recording (very, very rapid lowpass filtering to avoid frequency 'folding' or aliasing).

## Where to?

Modern console design's proliferate use of amplifier elements has mushroomed in recent years with the availability of compact and extremely low cost IC op-amps. Increasingly complex functional blocks are becoming increasingly commonplace—if in order to improve their electrical and sonic characteristics it would mean an increase in size and cost of well over an order of magnitude would they still be quite as popular? In the 'good old days' of valves it was not through any lack of expertise that equalisers even of today's complexity did not exist, it was just the size and cost of the concept that would have made even reckless men shudder. Also, it is to be noted, they were not really thought necessary.

Could it be that the next level of enlightenment in consoles is going to demand more simple and concise systems traded for a far higher and thorough level of elemental electronic design?

## Op-amp theory

Whilst there is no intention of turning this article into a beginner's guide to electronics, a brief description of the four main amplifier configurations used in op-amp circuitry could be useful. Almost all the circuitry in this series is based around these four formats, with very few exceptions, so a grounding in their characteristics will greatly facilitate finding your way around and understanding later circuits.

The op-amp is a 3-terminal device (**fig 13a**) labelled '+' '−' and '0'. They are normally operated between a split or differential rail of some ±15 to 18V, the junction between the supplies being called 0V and tied to the ground reference. All voltage and currents are measured and quoted with respect to this ground. For simplicity's sake power supply connections aren't shown.

The '0' terminal is obviously the output and is normally quite a substantial configuration capable of delivering some 20mA of either polarity, or put another way is capable of delivering its full output voltage (a little less than the rail voltage) into a load resistance of 1kΩ. The '+' and '−' input terminals are known as a differential input. If the − input is held steady at any voltage, and the '+' input moved from that same voltage in a positive direction, then the output voltage will rise positively at a rate of the input voltage change *multiplied* by the op-amp's quoted open-loop gain (for a 'TLO' some 100,000). Hence for a 100μV rise in voltage on the '+' input, the output voltage will rise 10V. If the '+' input moves negative of the − input voltage then the output voltage will go negative with respect to ground, again by the input voltage change times the amp's open-loop gain.

It doesn't really matter which of the inputs is held stable while the other one moves. If the '+' input is tied and the '−' input moved, then the output will still move to an extent determined by the gain but in the inverse polarity to which the input is moving. The '−' input is called the 'inverting input' because the output voltage changes in the opposite direction to the input, whilst the '+' is called the 'non-inverting input' because it moves in the same sense.

The real trick to the op-amp is this—if the two inputs are tied together and moved up and down in voltage, with an idealised unit the output will remain stationary at about 0V. This is called the common mode rejection characteristic. It is only the difference in voltage between the two inputs that causes the output to move, to a resultant voltage of the differential input voltage multiplied by the open loop gain.

OK, so now we have an amplifier with about 100dB gain. Very useful (ahem!). How do we go about making it give more usable amounts of gain, and defining that gain?

**Fig 12a** shows the basic 'voltage follower'. As the '+' input voltage rises, the output voltage rises also, pulling up the '−' input with it. The output cannot rise too far, though, or its effect on the '−' input would be to drag itself down again. A state of equilibrium is reached where the '−' input voltage balances against the '+' input, therefore maintaining the output voltage very close to that of the '+' input. Since op-amp inputs are of relatively high impedance, this configuration is used extensively to buffer critical circuitry away from a heavy or widely varying load.

A variation on the follower is the non-inverting amplifier shown in **fig 12b**. Here the output is removed from the '−' input by an attenuating voltage divider R1 and R2. As the '+' input moves, the amplifier again sets up an equilibrium between the '+' and '−' outputs as in the follower, only now in order for the '−' input to move as far as the '+' input to achieve the balance, the output has to swing far enough to overcome the attenuator. Hence the output voltage is the input voltage multiplied by the ratio of the total feedback network (R1 + R2) to R1.

The inverting amplifier (**fig 12c**) is really the same configuration but upside down with the input signal injected into one of the gain determining resistors.

If the input voltage rises, then the '−' input will try to rise above the '+' input, tied to ground. The output will drop to pull the other end of the resistor chain down so that the '−' input is back in balance with the '+' input. This way the amplifier always maintains the '−' input at virtually ground potential. The output voltage is thus determined directly by the ratio of the two resistors (since the bigger a resistor, the greater the voltage needed across it to drive the same amount of current through it—Ohms law) with the output voltage inverted with respect to the input.

Any number of source resistors can be tied into the '−' input. The output will generate an equal and opposite (given, as in **fig 12d** that all the resistors arc equal) voltage to the sum of all the input voltages. The '−' input still remains very close in potential to ground, so any individual source resistor is looking at a virtual ground and its signal is therefore isolated from all the others. Tra-la! The virtual-earth mix-amp!

The last configuration (**fig 12e**) is really a combination of the inverting and the non-inverting amplifier configurations. This is the unity-gain differential amplifier. Here importantly, the output voltage is determined only by the voltage difference between the two inputs, irrespective of their relation to ground. If the non-inverting input is grounded, with the floating ground-free input source still connected between the inverted and non-inverted inputs the IC '+' input will be to all intents and purposes grounded, so the output voltage will just be an inverse of the input voltage. If the inverting input is grounded instead, the input voltage will be attenuated to half its value by R3 and R4. However, since the amp is now operating as a non-inverting follower with a gain of two (determined by R1 and R2) the output voltage is at unity gain and non-inverted with respect to the input signal. Ungrounding the inverting input and feeding the entire configuration with a floating source will give a unity output with respect to its differential input. One more neat feature. If the inv and non-inv inputs are tied together and the source voltage is applied to them with respect to ground, no output will result. Why? because the circuit is simultaneously behaving as a unity gain non-inv amp and as a unity gain inverting amp—result total cancellation. Well, nearly total. The degree of cancellation is a measure of the circuit's Common-Mode Rejection Ratio.

Nearly all the circuitry you are ever likely to come across in any op-amp orientated bit of gear is likely to be made of these or combinations of these circuitry elements. The main exception to this rule, at least in the circuits to come, are where the devices are being used as dc control signal conditioners such as voltage comparators in peak-overload indicators and limiter sidechains, and in the non-linear quasi-logarithmic curve generator for the PPM drive amp, together with the precision rectifiers.

As comparators, op-amps are really quite efficient. The purpose of a comparator is to wave a flag when voltage exceeds a specified level. Going back to the basic op-amp characteristics for a moment, we know that if our '−' input is tied to a reference voltage, when the '+' input is below that, the output will be negative and when above the output swings positive. The only ambiguity occurs when the two inputs are very close in voltage—in face within 300μV of each other, then it behaves as an amplifier with its enormous open-loop gain, the output dithering somewhere between the extremes of the output swing. Still, its unambiguous enough in the real world!

A minor snag to this is that some op-amp types (such as the *5534*) have diodes strapped internally between the '+' and '−' inputs, so trying to use them as comparators causes instant fry-up. Mind you, if you're lunatic enough to want to use a *5534* as a comparator, you deserve to blow it up!

Another basic circuit worth exploring here is a halfwave precision rectifier. The diode (**fig 12f**) is included in the loop from output to '−' input, the halfwave output being taken from the '−' input, not the op-amp's output. The output rises 0.7V to overcome the forward drop across the diode, whilst the input is positive, still enabling the '−' input to balance the '+' input, following for the positive half-cycle. In the negative half-cycle, the diode is reverse biased, the op-amp has no feedback and the output swings hard negative, but the '−' input stays tamely tied to ground by the resistor, having a value far less than the diodes' reverse leakage. ■

# Designing a professional mixing console

Steve Dove

# Part 5 ~ Signal Switching

SIGNAL routing within the channel and other areas of the system is a prickly problem that has always been an area of much discontent for designers, especially since the advent of in-line consoles and projected function programmability. There are the old standard belt and-and-braces answers—relay systems—but these have lost, justifiably, a lot of appeal in the light of current technological advances.

Unless they are of the dizzyingly expensive miniature IC package variety, relays tend to be big, heavy, eventually unreliable, mechanically noisy and a nuisance to implement electronically, demanding support circuitry such as back-emf protection diodes and drive transistors for a realistically operable system. The coils, being inductive in nature draw a surprisingly large instantaneous 'on' current and release an equally surprisingly large amount of back-emf energy upon deactivation. Both of these—through mutual-inductance coupling, dubious common ground paths (even as far back as the master ground termination in separated supply systems), twitchy power supplies and even (it has been known) mechanical microphonic effects—tend to impinge themselves on audio signal paths as clicks, splats and other assorted bumps in the night. Of course it's possible to have silent relay switching, but after designing in separate ground un-related power supplies of considerable 'heft', spatially separated the relays from the audio, preferably on another card, worked out the drive interfaces and liberally sprinkled the whole issue with diodes, resistors and capacitors to tame the transients, you'll wish you'd taken up making telephone exchanges instead.

Certain routing applications do implicitly require relays, and their lack of concern about the amount of dc and either common-mode or differential signals of absurd quantities that may accompany the audio in balanced networks. Such circumstances are to be found anywhere a telephone line is.

Primarily, then, this is almost specifically a broadcasters' problem, where many external high-quality sources appear down 'phone lines and are routed before hitting either the station's internal distribution amplifier system or even a desk line input directly. 'Outside Source Selection', as it's called, does not—fortunately—have the same splat-elimination constraints as intra-console switching, since the signal is nearly always of high level, balanced, is riding with at least a little dc which will unavoidably click upon switching, but most importantly the selector is very unlikely to be switched whilst actually on-air. The BBC argue otherwise, but even *they* admit to not having an answer to the noticeable resultant clicks.

So, what are the alternatives to relays?

## Electronic switching

The basic outline characteristics for an audio switch are simply that it has an infinite 'off' impedance and zero 'on' impedance, and that its control be isolated from the through path. In the real world, of course, some leeway has to be given, but fortunately not much in these basics, but more in subtleties.

Transistors are out, despite their high on-off impedance ratios, because they are essentially undirectional in current flow, and the control port (the base) is actually half of the signal path as well.

Field effect transistors have been and still are used extensively for switching. They again have a high on/off ratio, the control port (the gate) is of extremely high impedance and well isolated from the signal path, but the gate on/ off voltage levels are a bit awkward for interfacing with logic control signals. It is bidirectional, its channel path being essentially just a voltage-controlled resistor, but the 'on' resistance tends to vary a bit with the varying audio voltage across it (auto-modulation) hence distortion in the more basic FET switching configurations can be a problem.

Closely related to FETs are MOSFETs, a different chemical structure and physical construction, but of essentially similar characteristics with the pleasant exceptions that the gate is of even higher impedance and better isolated, the control voltage swing required also being easier to deal with. Complementary MOSFET elements (CMOS), connected back-to-back to form virtually ideal bidirectional analogue transmission gates are nowadays manufactured in all manner of variations and packages by IC manufacturers. Early versions of CMOS transmission gates had some rather untoward vices. They were 'raw' CMOS elements and one of their main attributes, the extremely high impedances in their 'off' states and of their control ports, made them liable to destruction by fairly everyday amounts of static electricity—cleaning the IC pins by rubbing them on a nylon pullover sleeve was just not on. Also, they tended to latch up easily should any of the MOS junctions inadvertently get reverse-biased into conduction. Most current devices are now gate-protected to prevent static-blatting, and the worst that happens with the audio signal exceeding the switch rail voltage by a small amount now is that the switch 'breaks over', not resulting in the terminal consequences it once did.

Perhaps the best-known and most-used switch of this kind is the *4016* (and its brother the *4066*, which is identical but for a lower 'on' resistance). It is a 14-pin dil package containing four independently-controlable CMOS transmission gates. Each gate can pass up to the IC's rail voltage (typically 18V) into a load impedance of down to 1kΩ at a rated distortion of about 0.4% in the most basic of switching formats. Obviously both the distortion figure and the headroom availability of 18dB above 0.775V (for 18V supply) are both woefully inadequate by today's expected console standards. Another less obvious pitfall is the decreasing switch isolation at high frequencies due to leakage capacitance.

## CMOS characteristics

**Fig 27a** gives a typical representation of the variation of a CMOS transmission gate's 'on' resistance with signal voltage applied to the gate. This variation in resistance is of course the source of the distortion. If we could restrict the signal voltage to within that nice (linearish) bit in the

FIG. 27 TYPICAL C-MOS TRANSMISSION GATE 'ON' RESISTANCE CHARACTERISTICS

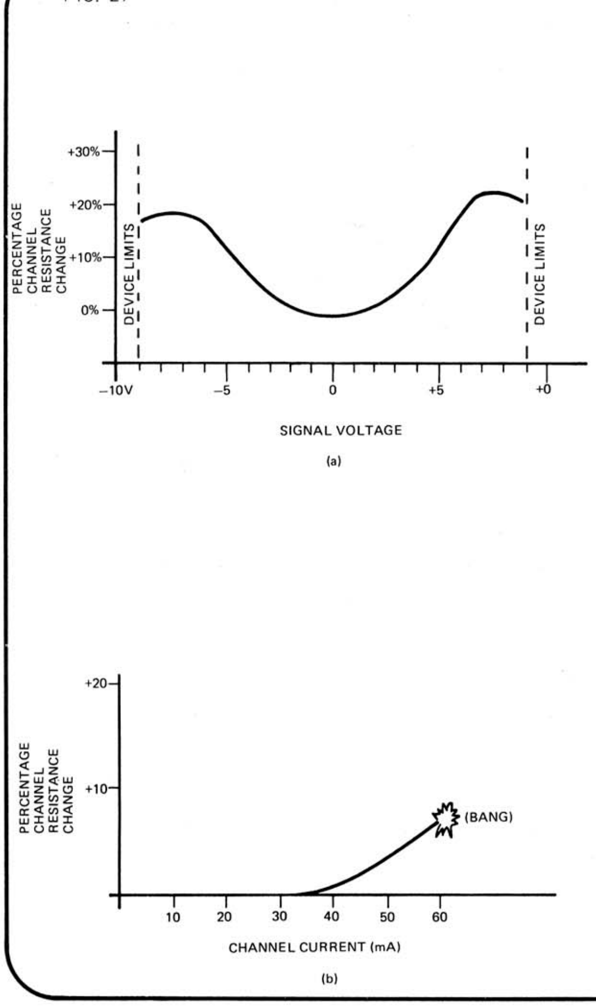
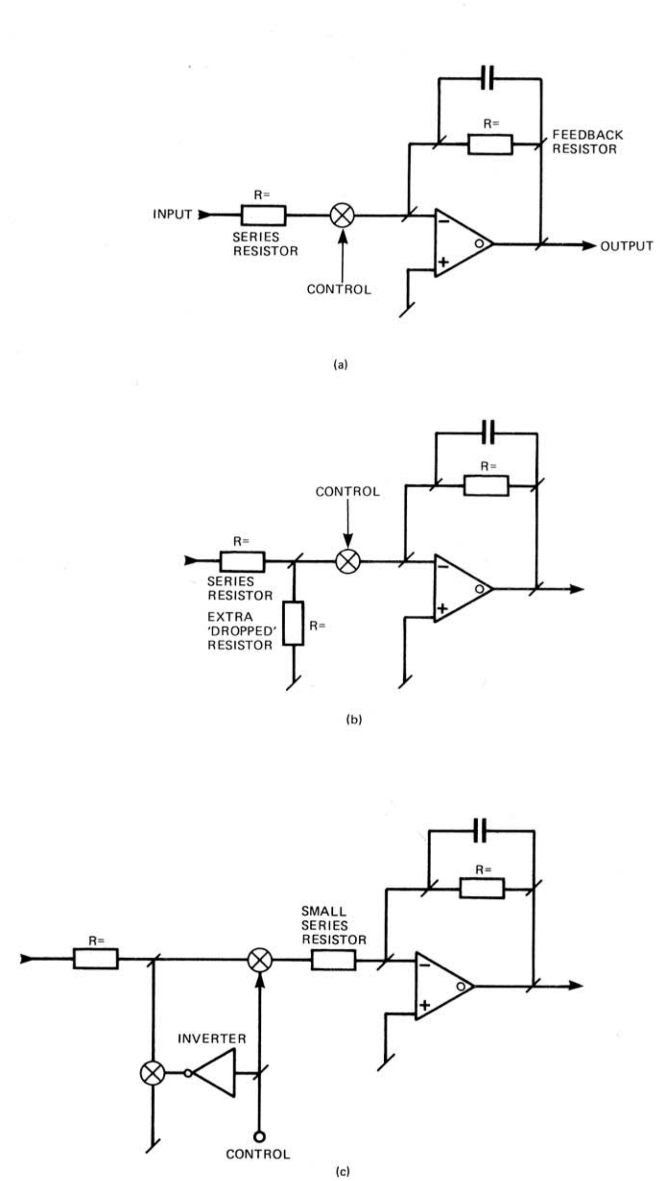



FIG. 28 SWITCHING ARRANGEMENTS USING C-MOS TRANSMISSION GATES




middle . . . or better still virtually eliminate the signal voltage altogether . . .

Placing the switching element right up against a virtual earth point as in **fig 28a** achieves this signal voltage elimination, the switch now behaving as a two state resistor almost perfectly. When closed, the 'on' resistance variation (which will be small anyway because of the very low voltage swing across it) will be effectively swamped by the (relatively) much larger series resistance. When open, the 'off' resistance extends the total series resistance to a value approaching infinity. In practice, the on/off ratio is not really adequate. Capacitance across printed circuit tracks and in the device encapsulation itself, combined with common-ground current and other essentially flat-response crosstalk mechanisms result in a cross-switch leakage characteristic ultimately rising 6dB/octave against frequency. Also, despite the fact that the distortion problem is now resolved, there still remains a headroom problem when the switch is open. If the source voltage presented to the series resistor exceeds that of the CMOS gates' supply rails, the gate will 'break over', turning on for that excessive portion of the input waveform.

Dropping a resistor equal in value to that of the series resistor from the junction of that resistor and the gate to ground **(fig 28b)** is a working approach. The maximum signal that can be present across the gate when 'off' is now half that previously, which is usually more than enough to prevent break-over. This 6dB loss is magically made up for in the 'on' mode because the signal's source resistance into the virtual earth amplifier is now halved (series resistance effectively in parallel with the dropped resistor). Incidentally, the crosstalk improves as a consequence by almost 6dB also, but—swings and roundabouts—the noise output from the amplifier is degraded by 6dB since we're asking it to provide that much gain. For many practical purposes, this switching configuration, with its performance as defined, is quite adequate. For instance, the noise and crosstalk characteristics are a good order of magnitude superior to any analogue multitrack recorder, so this element could be a good choice for a track assignment routing matrix.

A refinement of this element—in fact really an extension of the same principle—is shown in **fig 28c**. Here, a second analogue transmission gate replaces the dropped resistor, and is driven through an inverter from the control line for the original gate, arranging for it to be 'on' when the other is 'off' and vice-versa. When the original gate is 'on' there is very little potential across either of the gates, and similarly, when the second gate is 'on', since it is tying the series resistor to ground, crosstalk is dramatically improved, since when the element is 'off', any signal present at the series resistor faces the double attenuation of the series resistor into the 'on' second gate followed by the 'off' original gate into the virtual earth input of the amp. In the elements' 'on' mode, there is no input attenuation, hence no gain and no extra noise contribution from the amp. The only limitation now to this switching elements' cross-switch leakage characteristic is pc card layout and grounding arrangements. Given a good home this element is virtually unmeasurable. It does, however, have one naughty quirk that may preclude its use in some places. Unless a great deal of bother is gone to to arrange complementary on/off switching timing for the two gates, they are both momentarily partially 'on' together during a switching transition. This, for an instant, ties the virtual earth amp input to ground via the quite low 'half-on' impedances of the two 'seriesed' gates. This creates an instantaneous burst of extremely high gain in the

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amp which shows as a transient of noise, or worse still, as a 'splat' if any dc offset is present at the virtual earth point. It can be minimised, or at least the extent of the transient defined, by a small value resistor in series with the input (**fig 28c**). This will, of course, increase the signal voltage across the gates and hence increase the distortion, so a compromise has to be struck to suit the given application. However, excessive distortion should not be a problem.

In order to reduce the thermal noise contribution to the circuit noise performance, the resistances involved in switching should be as low as practically possible, consistent with device limitations and the ground current arrangements. The feedback resistor around the virtual-earth stage is limited by the op-amp's output drive capability, bearing in mind it has to drive its load, too! **Fig 27b** demonstrates a typical CMOS switching element's channel resistance variation with through current—it behaves quite nicely and linearly until about 40mA, which actually compares more than favourably with the output drive current capability of an op-amp. As a rule of thumb then, the resistors used around analogue gate switching circuits can be as low as 2.2kΩ.

## A practical matrix

The *4000* series of CMOS devices—which are very commonly used—have one important feature at odds with general mixer technology—their maximum supply voltages. The earlier *4000 'A'* series were limited to 15V total (as compared to the 36V total used in this particular console design) whilst the more recent buffered *'B'* series can stand 18V, with a bit more at a squeeze (Who said that!?). Whilst this is immaterial, given the virtual-earth switching technique, it is a pain having to provide for and derive a differential ±5V supply either centrally or on each card in addition to the main differential 18V rails. Many IC manufacturers however, notably Siliconix and Harris, produce analogue switching packages not only capable of running directly off the full mixer rail voltage, but also in switching configurations that can be directly and usefully applied to our purposes (despite the fact that they were designed for something else completely). **Fig 29** shows one mixer channel's worth of a digitally-assigned 32-track routing matrix, designed around a pair of Harris *HI506A* 16-way multiplexers.

The *506A* contains sixteen analogue transmission gates tied to one common 'output' (which we will cross out and pencil in 'input' instead). Each of the free ends of the gates do not pass go, and go directly to a mix buss each. They all share a common series source resistor via the input port. Since only one of these gates can be open at a time—the one corresponding to the binary 4-bit address code on the address inputs—there is no possibility of two or more busses being inadvertently shorted together. The device manufacturers proudly point out the 'break-before-make' delay in switching, meaning that a newly selected gate waits until the previous one has de-latched, so there isn't even a momentary switching short.

Crosstalk with this configuration (which you will notice is a variation between **figs 28a** and **c** is extremely good. Again there is the double attenuation of the series resistor into the 'on' gate (some 20dB to start with) followed by any 'off' gate into any of the other virtually zero impedance mixing busses. A slightly more critical crosstalk situation could exist when all the gates are turned 'off' (by tying the *506's* enable low (pin 18)) since the first set of attenuation no longer exists. This is why external switching elements (IC3a and IC3b) are arranged to tie

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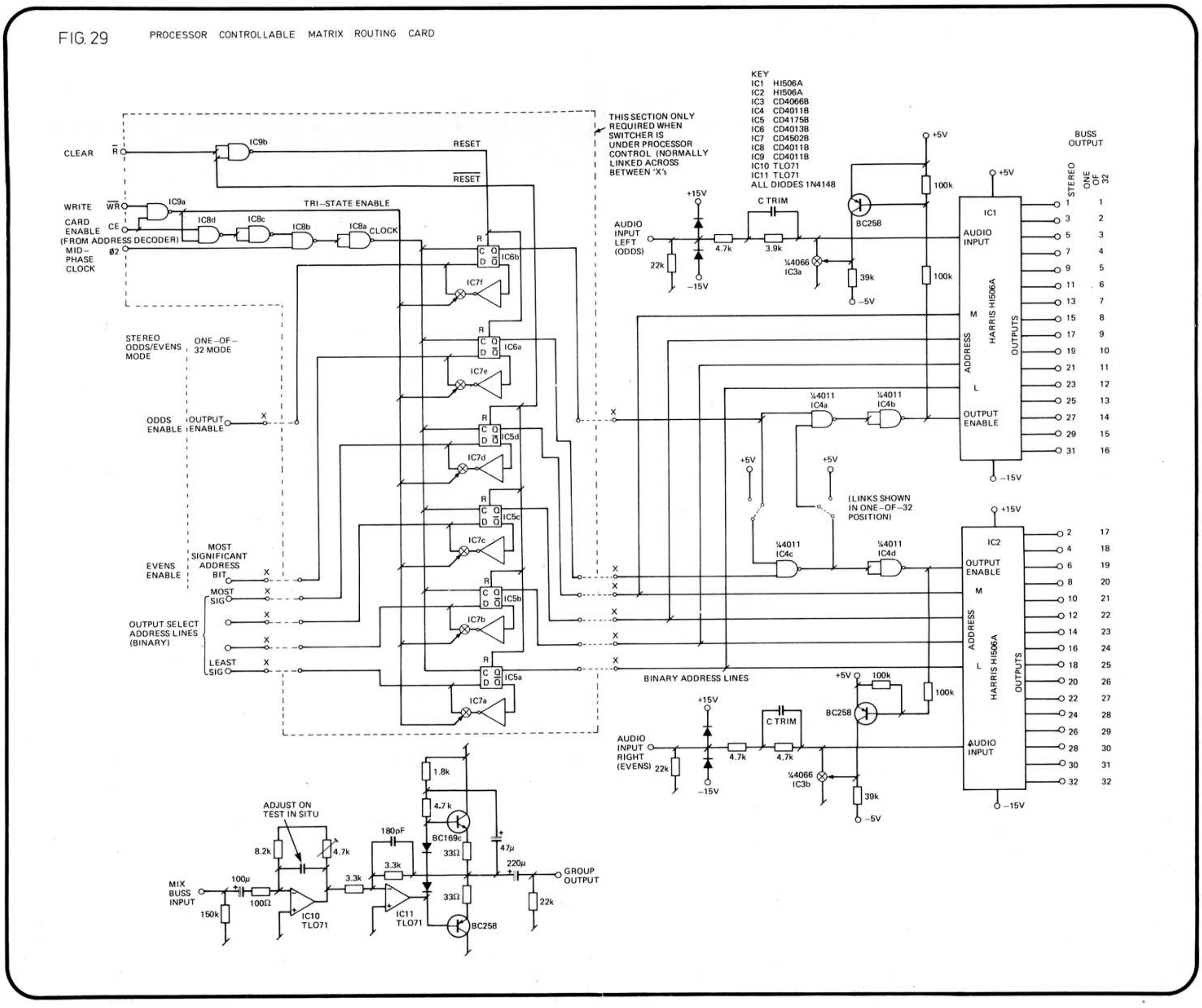


# Mixing console

**TABLE 1**

**PROCESSOR I/O LINES DEDICATION**

| | Line | Function | |
|---|---|---|---|
| CPU I/O | 1 | CARD ADDRESS | —least significant bit (LSB) |
| | 2 | CARD ADDRESS | |
| | 3 | CARD ADDRESS | (Seven binary bits allow up to 128 matrix routing cards to be separately addressed) |
| | 4 | CARD ADDRESS | |
| | 5 | CARD ADDRESS | |
| | 6 | CARD ADDRESS | |
| | 7 | CARD ADDRESS | —most significant bit (MSB) |
| | 8 | $\overline{R}$ $\overline{\text{RESET}}$ or $\overline{\text{CLEAR}}$ | (When this bit is *down*, all memory is cleared) |
| CPU I/O | 9 | $\overline{W}$ $\overline{\text{WRITE}}$ | (This bit down allows data on lines 11-16 to be stuffed into memory) |
| | 10 | $\phi 2$ MID-PHASE CLOCK | (Clock pulse delayed slightly to allow data and address to settle before enabling the memory) |
| | 11 | DATA | —least significant bit (LSB) |
| | 12 | DATA | Matrix card, output or selection |
| | 13 | DATA | Matrix card, output or selection |
| | 14 | DATA | —most significant bit (MSB) |
| | 15 | DATA | —evens enable (or MSB is one-of-32 mode) |
| | 16 | DATA | —odds enable (or ON/OFF) |

the end junction of the series resistor and the *506* inputs to ground whenever the enable lines are low.

Crosstalk is now completely down to the interconnections to this card, power supply decoupling, solid and correct ground paths but mostly to inductive and buss/earth/buss eddy-current coupling between the virtual earth busses themselves: yet another design area where performance is completely determined by mechanical considerations . . .

The same switching card may be configured—merely by changing two wire links—in two different routing formats. The first enables a stereo pair of signals (say the planned outputs of a channel) to be routed to adjacent pairs of outputs, ie 1 and 2, 7 and 8, 27 and 28, etc, where the odd numbers represent 'left' and the evens 'right'. Either odds or evens may be accessed singly by suitable feeds to the 'odds enable' and 'evens enable' control inputs. Quite obviously these also facilitate disabling (turning off completely) the routing.

A 4-bit binary control buss selects which pair of the possible 16 pairs may be accessed, so these six control lines are all that need to be extended to the channel module where simple switchery performs all routing requirements.

When the aforementioned wiring links are made in the fashion shown in **fig 29** the card becomes configured as a one-source-into-32 destination switcher, necessitating some control function changes. 'Evens enable' becomes the additional highest significant bit of the destination address code (five bits are needed for 32 combinations) whilst 'odds enable' turns into the switcher's enable/disable control. (The benefit, in both modes, of disabling the switcher when not actually in use is that it removes the feed totally from the destination busses, therefore not impairing their performance at all, whilst not disturbing a pre-selected routing set up on the address lines.)

With the same signal applied to both the audio inputs, it is now possible to access any one of the 32 busses.

## Processor control

The seemingly great mass of logic circuitry enclosed in the dotted lines allows the card to be controlled by a computer or central processing unit (CPU). All it really is, is six flip-flops acting as memory elements (so that the card can remember what the CPU has told it to do) and six tri-state buffers that, on request, tell the CPU what the card is actually doing. These little chunks of memory both save the CPU having to store the information somewhere else and also act as a very useful diagnostic aid to help find out what isn't doing what, where and why—something that anyone who has played with computers or large logic systems will immediately realise the value of!

For ordinary direct operation, this logic would be left off the card and linked across (between the 'x's on the diagram). The gaggle of NAND gates in the top left hand corner merely organises the CPU buss information to fire the appropriate clock, enables and resets to the memory elements.

CMOS *4000* series logic operating at 5V is not the fastest logic family in existence and would probably prove too slow for most recent microprocessor CPUS. This is not a problem in reality since the practical way of dealing with this would be to hang the entire switching matrix logic system off a bunch of the CPU input/output lines, masquerading as a local address/data/control buss system.

A nice convenient 16 I/O lines are required (two lots of eight—handy for micros) and these are formatted in **Table 1**. Being software controlled, the I/O lines may be timed a little more gently than the hardware-determined processor busses.

A separate address decoder card takes however many of the card-address bits are required (5 for 32, 6 for 64, 7 for 128) and generates the decoded feeds for the 'card enable' (CE) on each matrix card.

## Audio path

Enough of this digital junk. If you've survived without migraine, in the bottom left-hand corner of **fig 29**, hide a good old-fashioned analogue mix amp and line amp which are the group output stages for the channel to which the particular matrix card is relevant. Where else to put them except on the matrix card where they can't get any closer to the busses?

The mix buss input is tied on the back of the edge-connector to the buss it is responsible for—this ensures card replaceability and redundancy.

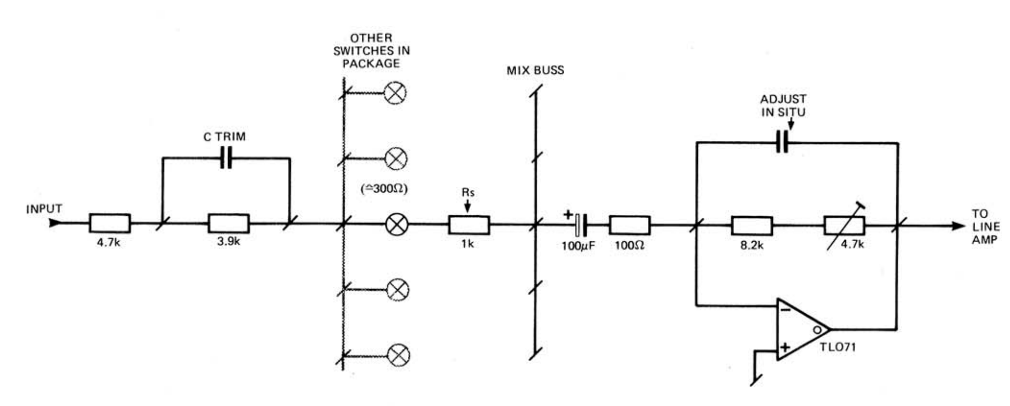
Note! No values are attributed to the feedback capacitor around the mix amp, since this not only has to compensate for the amp's own tendency to oscillation but for the added irritation to this of the buss impedance—an unknown until actual construction. Similarly, astute circuitophiles will note a 'bodgette' in the form of a capacitor across part of the switcher's input series resistance. This provides a variable hf 'kick' which can be of assistance in sorting out frequency and phase response quirks in particularly horrid buss systems—this is fortunately very rarely needed and is provided 'just in case'.

**Fig 30** shows the audio path through the switcher, devoid of frills. The mystery 1k resistor, Rs, which does not appear on **fig 29**, is internal to the *HI506A*, appearing on each of the switches 'inputs'. Although a minor nuisance in this application (it means the MOSFETs are not actually switching a 'zero' impedance) they are part of the device's internal protection against, principally, static electricity. A worthwhile sacrifice. Harris do make an unprotected version, the *HI506*, with all the switching elements exposed (which would mean that the device's 'on' impedance would be down to the 300-odd ohms of the element as opposed to the 1.3k-ish of the *HI506A*) but the use of an expensive unprotected IC on a plug-in card such as this has a very high cringe-factor.

The total source impedance before the buss is around 9.9k, which with the addition of the 100Ω buffering resistor becomes 10k before the virtual-earth input of the mix amp. A gain trim 4.7kΩ resistor in series with 8.2kΩ gives a gain determining feedback resistor swing of approx 8.2kΩ to 12.9kΩ, corresponding to a tweak swing of −1.7dB to 2.2dB.

The line amp is quite unremarkable, being a simple beefed-up inverting amplifier—necessary to maintain the absolute input/output phase relationship. ■

FIG. 30 SWITCHER AUDIO PATH



# Designing a professional mixing console

Steve Dove

# Part Six ~ When is a Ground not a Ground?

A HUMAN working visualisation of anything electronic soon becomes impossible without a mental image of the nice, solid, infinite, immovable, dependable god — Ground. Similar to the Messiah, this one has many names too: Earth, Ground, OV, Reference, Chassis, Frame, Deck, etc — each of differing interpretation but all ultimately, alluding to the great Omnipotent Nothingness.

Electrons couldn't care less about all this. They just go charging about as potentials dictate and any circuit will work perfectly well referred to nothing but itself (satellites, cars and flashlights work, don't they?). 'Ground' in this instance is but an intellectual convenience.

Interconnection of a number of circuit elements to form a system necessarily means a reference to be used between them. To a large degree it's possible to obviate a reference even then, by the use of differential or balanced interfacing, unless of course power supplies are shared.

So having proved that ground is seemingly only a mental crutch, why is it the most crucial aspect of system design and implementation?

## Wire?

**Fig 31a** shows a typical, ordinary long thin bit of metal known more commonly as wire and occasionally as printed-circuit track. However short it is, it will have resistance which, courtesy of Mr Ohm, means that a voltage will develop across it as soon as any current goes along it; similarly, Mr Maxwell says a magnetic field will develop around it — Bingo! Inductance. If it is in proximity to anything, it will have capacitance to that too.

So, **Fig 31a** actually looks more like **Fig 31b** with resistive and distributed reactive components. Admittedly these values are small and seem of little significance at audio frequencies but clues have already been laid (particularly in Part Three Op-Amps — Friend or Foe?) that believing the world ends at 20kHz is not so much myopic as stupid.

A radio engineer looking at **Fig 31b** would mumble things like "tuned line", "resonance" or "bandpass filter", maybe even (are you sitting comfortably) "antenna". Rf technology and thinking may seem abstruse and irrelevant to console design until it is considered that devices commonly used nowadays have bandwidths often many dozens, sometimes hundreds, of Megahertz wide. An even more frightening realisation is the enormous quantities of rf energy present as a consequence of our technological being.

A more obscure collection of equivalents is shown in **Fig 32**: **(a)** representing a wire into a bipolar transistor input; **(b)** a wire from a conventional complementary output stage; whilst for reference sake, **(c)** is a basic 'crystal-set' type radio receiver. Quaint, but for the presence now of considerably more V/m rf field energy compared to the heyday of 2LO. In all the three circumstances rf collected and/or delivered by the antenna/tuned line is rectified hence demodulated by a diode (being the base-emitter transistor junctions in **(a)** and **(b)**). As contrary as it may seem for demodulation to occur at an amplifier output, it is perhaps the most common detection mechanism with the demodulated product finding its way back to the amplifier input by means of the conveniently provided negative feedback leg.

Making our bit of wire fatter and thicker has the effect of lowering the resistance and inductance whilst increasing capacitance (greater surface area exposed to things nearby) so although the wire's resonant frequency stays about the same the dynamic impedance (hence 'Q') reduces. Whilst in general this is deemed to be a 'good thing', in some instances it can merely serve to improve the matching and coupling of the rf source to the resonance.

Carried to a not-quite-fatuous extreme, even the console frame constitutes a big fat resonant tank at a surprisingly low (mid-vhf) frequency—and frame resistance, however heavily constructed, cannot be disregarded and treated as a universal earth path. Some 'earth' eh?

For the purposes of practical design, these considerations perhaps become a little better defined. The reactive elements of capacitance, inductance with the attendant effects of resonance, and filtering are concerned with less obvious aspects such as electronic stability and proneness to radio demodulation, whilst resistance gives rise to most of the horrors usually lumped under the collective 'grounding problems'.

## A 'good earth'

The closest most of us get to earth is the big pin on a mains plug and fortunately for most purposes it is adequate provided just the one point is used as the reference — other points are likely to have slightly differing potentials due to dissimilar routing and resistances. Compared to a 'technical earth' (eg a water pipe (make sure the plumbing isn't plastic, *please*) alternatively a fortune in copper pipe hammered into the earth) conventional mains earth can have a surprisingly high potential — a volt or two even — considering it is principally a safety facility not ordinarily carrying current. Any potential implies resistance in the earth path which is bad news about something intended as a reference whilst also detracting from the safety aspect. Practically, though, it doesn't matter too much if everything is waving up and down a bit provided everything, including *even unrelated things in proximity* are waving up and down in the same manner. The potential is usually small thankfully, meaning that the 'earth' impedance is reasonably low to the extent it may be considered zero.

## Why earth anything?

With all our component system parts tied together by a reference 'ground' (the organisation of which is a bramble patch in itself, to be trod in later) and everything working as expected, the question arises of why it is necessary to refer our 'ground' to earth. If the internal grounding is completely kosher our system will operate perfectly, quietly and tamely regardless of what potential (with respect to earth) it is tied to, whilst if not tied, it will derive its own potential by virtue of resistive leakages, inductive coupling and capacitance to things in its environ. For an independently powered system (say batteries) these leakages and couplings will be of far higher impedance and hence easily swamped by human body impedance to earth. (We are, dependent on hand-clamminess and footwear reasonably coupled to terra-firma at between 5kΩ and a couple of MΩ at 50Hz).

If, as is most often the case, most of the system is powered off the ac mains this floating ground potential becomes of far lower impedance and consequentially much more capable of dragging current through the human load (it's the current that does you in, not the voltage). A tell-tale sign is a burring/tingling feeling as you drag a finger across exposed metalwork on something that is deriving its own ground potential.

The mechanism for this lower impedance is fairly straightforward. Mains transformers are wound with the optimum transfer of energy at 50 to 60Hz and very high flashover voltages, say 2 or 3kV, in mind — the finer points of transformers such as leakage inductance, interwinding and winding imbalance capacitance are all but disregarded, meaning they end up being horrific.

Being far greater in scale than ordinary ambient reactive couplings, they primarily dictate the floating ground potential at anything up to 240V ac or whatever the mains happens to be locally.

A strange practice by a few, predominantly American, manufacturers is to tie either or both the 'live' and 'neutral' mains ac lines to chassis via capacitors, typically of 1 to 100nF with the result that if the chassis is not directly earthed it rides at (in the case of both lines being tied) half the mains voltage. The capacitor values grossly swamp transformer leakages and give the chassis floating potential an uncomfortably (literally) low impedance. The chassis tingle changes from 'Mmm — interesting' to vile oaths with attendant flailing limbs.

A system composed of many separate mains-powered things will almost certainly hum, buzz and sound generally uneasy — seemingly in direct contradiction to the earlier statement that ". . . the system will operate perfectly . . . regardless of what potential it is tied to . . .". Being tied to lots of different potentials at lots of different points along a ground path is definitely not playing the game, sorry.

Each different mains transformer will have different amounts and permutations of leakage and hence propagate different potentials and degrees of mains-borne garbage into our otherwise perfect grounding path. Assorted ground potentials mean assorted ground currents meaning assorted noises.

Tying the entire grounding path to earth is the ultimate swamp-out of leakage impedances. A connection to a (nearly) zero impedance makes a nonsense of most other potential creating paths, most of which have reactances exceeding 1kΩ. Sledgehammer technique.

Ordinarily in such a multi-supply circumstance, regardless of earth termination, significant currents exist along the ground reference lines. The resultant inter-element noise and hum voltages (developed across the inevitable line resistances) quickly become intolerable in unbalanced systems — any wobbling of the ground reference becomes directly imposed upon the required signal.

Balanced, or pure differential transmission helps obviate these perturbances by rendering them common-mode in a system that is (theoretically) only sensitive to differential information. In reality, practical transformers can afford a good 70 to 80dB common-mode isolation at low audio frequencies but deteriorate in this respect at 6dB/octave with increasing frequency, to poor rejection (if any) around the winding resonance frequencies unless considerable effort is made to 'fudge' a more accurate balance externally. Although transformer balancing does effect a dramatic improvement

FIG. 31 WHAT IS A BIT OF WIRE?

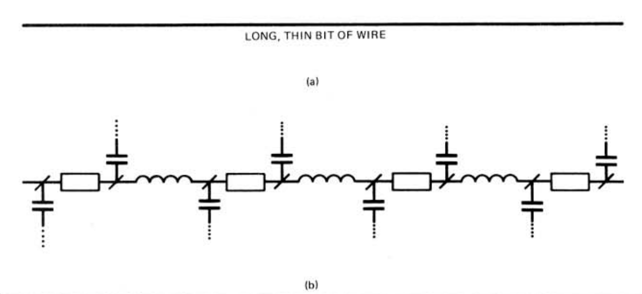
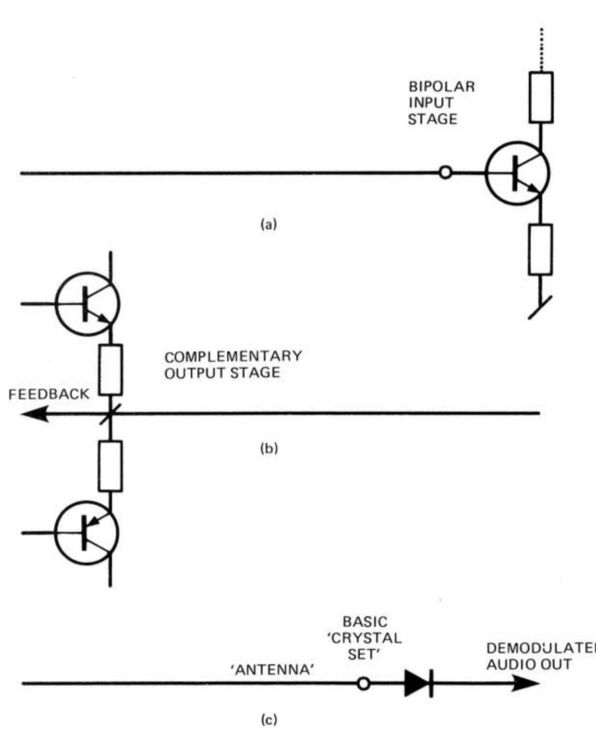



FIG. 32 COLLECTION OF EQUIVALENTS

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# Mixing console

in noise levels, it is far greater for fundamental hum (50 to 60Hz) than it is for other mains-borne noise. This explains why in 'dodgy' systems lighting dimmer buzz, motor spike noise or any source with a high hf energy or transient content is so persistent. The ubiquitous 'tizz'.

The golden rule is to treat any balanced system's grounding as if it was unbalanced — this minimises the inevitable reference ground currents whilst helping to unlearn that transformers are a panacea.

There is one good reason not yet mentioned for grounding to earth. The consequences of a piece of gear's chassis becoming inadvertently mains live potential, are obvious. Rather death to a fuse or breaker than one of us.

Let's assume (giggles) that the grounding for the studio control room is all sensible and that our console has a nice juicy solid earth termination. What about the intra-console grounding paths? This is perhaps the ultimate unbalanced signal path.

## Inside the console

Most conventional amplifier stages rely on a voltage difference between their 'input' and 'reference' in order to produce a corresponding output voltage (referred, naturally, to the input's reference). If the input is held steady, though, whilst the reference is wobbled, a corresponding (amplified) inverted wobble will appear at the output.

It is plain, then, that any signal the reference sees that is not also *common* to the input (eg ground garbage) will get amplified and summed into the output just as effectively as if it were applied to the 'proper' input. The obvious (and startlingly often overlooked) regimen to render extraneous garbage unimportant, is to ensure that the point at which an amplifier's source is referred is tied directly to the amp's reference, whilst the amp's output is only taken in conjunction with its reference. Successive stages daisy-chain similarly — source reference to amp reference, amp reference to destination reference, etc, etc.

This thinking is called. . .

## Ground follows signal

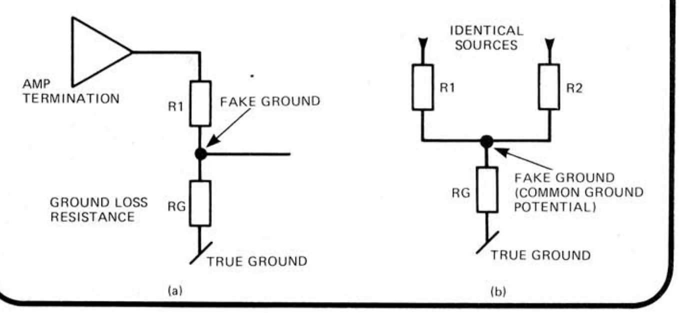
A classic maxim and one that has dictated the system design of nearly every console built. It was particularly true in the era of discrete semiconductor design, where 'ground' was not only audio ground but also the 0V power supply return. As an added complication the power supply positive rails, being heavily regulated and coupled to ground were an equal nightmare as they too became part of the grounding path. This could be fairly simply avoided though by spacing each circuit element away from the supply rail by an impedance considerably greater than that offered by the 'proper' ground path — achieved by either separately regulating or simply decoupling by a series resistor/parallel capacitor network.

Accelerating technology has for once, atypically, actually made life a bit simpler. Specifically, the trend toward IC op-amps with their required differential (+ve and −ve) supply rails. This, thankfully, removes electronic operating current from the audio system ground, whilst individual stage supply decoupling is rendered unnecessary (in most instances) by the excellent power supply noise rejection ratio of most popular op-amps. Nevertheless, correct grounding paths still apply, the removal of supply current just exposing and highlighting audio grounding subtleties.

Unfortunately, whilst op-amps have simplified matters in one respect, their ease of use and versatility have been largely responsible for the creation of enormous systems with so many stages, break points, mix busses and distribution networks that the simple daisy chaining of 'ground follows signal' becomes unwieldy if not unworkable. Alternate grounding schemes, such as 'star' grounding where every ground path and reference is taken to a central ground/earth tend to play an increasingly important role.

In practice, a necessary compromise between these two prime systems occurs in most console thinking. 'Daisy chain' applies mostly to 'on card' electronics (eg in the mic-amp/eq sections) whilst systems switching and routing rely on 'star'.

FIG. 33 AMPLIFIER OUTPUT LOAD




## Ground current summing

A principal grounding-related manifestation is crosstalk, or the appearance in a signal path of things that belong elsewhere. Other than 'air-borne' proximity related reactive crosstalk (Part Two) most unwanted visitations are by courtesy of the resistive ground path mechanism.

In **Fig 33a**, $R_1$ represents the load of an amplifier output (what it is in actuality, say the 10k$\Omega$ of a fader or a 600$\Omega$ line termination, is immaterial for the present). RG represents a small amount of ground path wiring, etc, loss resistance. It is quite apparent that the bottom end of the termination is spaced a little way from true ground by the wiring resistance — the combination forms a classic potentiometer network. The 'fake ground' has a signal voltage present of the amplifier output voltage attenuated by $R_1$ into RG.

In a practical circumstance with a 600$\Omega$ termination ($R_1$) and a ground loss (RG) of 0.6$\Omega$, the 'fake ground' will have present a signal voltage about 60dB down. Use of the 'fake ground' as a reference for any other circuitry is a sure-fire guarantee of injecting −60dB worth of crosstalk.

Two identical terminations sharing the same 'fake ground' (**Fig 33b**) happily inject a small proportion into each other by generating a common potential across the ground loss RG.

Should the second termination be far higher in impedance (say the 10k$\Omega$ of a fader) its contribution to the common 'fake ground' potential will be far less (−86dB) since the ground impedance is much smaller in relation to the source. Correspondingly, though, this higher impedance termination is more prone to be cross-talked into from the lower impedance contributors to the common ground.

## Typical problems

Let's take a fairly unusual (but definitely not unknown) grounding anomaly caused as a result of inattention to grounding paths. $A_2$ is a line amp feeding a termination of 600$\Omega$, into a lossy ground of 0.6$\Omega$ resulting in a 'fake ground' potential 60dB below the amp's output (**Fig 34**). An earlier stage in the chain ($A_1$), in this example a mic amp, with a considerable amount of gain has its feedback leg (amplifier reference) tied to the same 'fake ground'. Its

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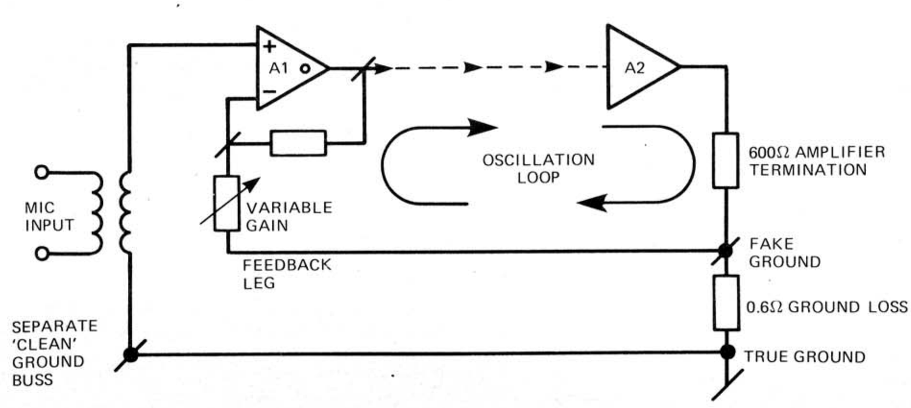
FIG. 34 POTENTIAL INSTABILITY MECHANISM AS A RESULT OF INCORRECT GROUNDING



input ground reference (here lies the cock-up) is taken from a separate buss supposedly to provide a nice 'clean' ground. This, of course, it does admirably, the buss being tied straight to 'true' ground and having no sources of great substance going to it.

Any signal present on the 'fake ground' is duly amplified by the mic amp (in its inverting mode), is attenuated at the line amp output back into the 'fake ground' and — you guessed it — as soon as the mic amp gain exceeds the output attenuation the entire chain bursts into glorious oscillation.

A very similar mechanism was responsible for an owner's criticism of his well-known type American console, that whenever he attempted to use the track routing on any channel module, the sound of that channel discernibly altered. It was found that ordinarily nothing in the channel drew much current, all ground impedance requirements being quite light. Until, that is, the track routing line amp was accessed with its load of routing resistors and terminated output transformer demanding a relatively large ground current. This output stage current shared the module's only ground access point (two parallelled connector pins) with all the rest of the module electronics — with the notable exception of the mic and line input transformer ground returns. The resultant feedback, although nowhere near enough to promote oscillation, did by virtue of the output transformer's phase shifting at both hf and lf frequencies result in distinct colouration.

A purist answer to these 'fake' and loop problems is to choose one grounding point for the entire console and to take every reference and ground return directly to it through separate ground wires (which must of course be coloured green in order to function correctly).

A few minor problems would ensue.

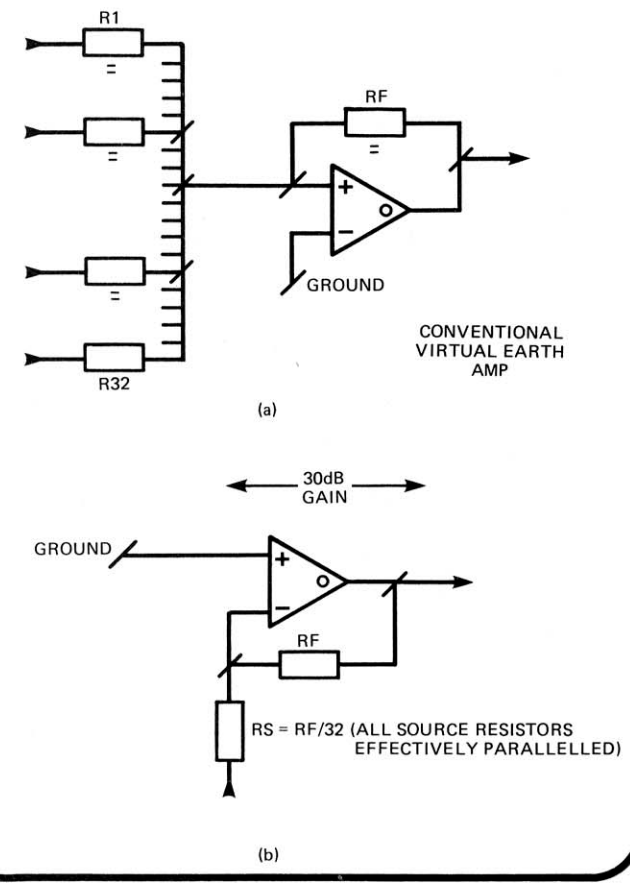
The enormous number of ground lines required would soon outstrip the capacity of the module connectors, the mass of wiring would cause apoplexy if not dark mutters of resignation from the wiremen whilst also severely aggravating the already desperate world shortage of copper. Fortunately, a working compromise suggests itself based upon separating the different classes of ground requirements by impedance.

It seems reasonably safe to tie fairly high impedance sources to a common ground point, buss or line, (since the ratio of their impedances is so great that resultant 'fake ground' potentials will be normally low enough to ignore). Anything that is likely to draw current (any kind of output or line amp stage) should go directly to ground, will not pass through any buss and not collect shared ground paths on the way.

Any ground buss must have a measure of resistance and must therefore be 'fake' to a certain degree, if not a truly festering pit of garbled nasties. If we do our sums right, ground buss signal levels can be kept acceptably low, say below −100dBu.

Smugly, we can expect to ignore figures like that — until we (almost inevitably) amplify them up. If you're wondering what crazy circuit arrangement unavoidably amplifies up garbage ground noise — it's called the . . .

FIG. 35 VIRTUAL EARTH MIX-AMP AS AMPLIFIER OF GROUND-BORNE GARBAGE




## Virtual earth mix amp

**Fig 35a** tells the story. As, say, a multitrack mix amp, it can typically have 32 sources applied to it — the through gain from any source being unity (assuming the source resistors equal the feedback resistor) but the real electronic gain of the circuit is 33 or about 30dB. Redrawing the circuit slightly, **Fig 35b** shows exactly what this 30dB is amplifying. Clue: that which is directly applied to the op-amp's non-inverting input. Yep, ground! True, it is also merrily amplifying the noise due to the resistors and the internal noise mechanisms of the device, but for our argument here it is amplifying *ground*.

In any reasonably sized console, providing no sources are grossly out of proportion to the majority, ground garbage is pretty random and 'noisy' in character — the result being that on being amplified up it serves to make the mix amp apparently noisier than would be expected from calculation. In suspect systems it has been found to be *the* predominant noise source. It is truly astonishing what loving care and attention to virtual earth mixer grounding can have on buss noise figures.

For mix amps, practical noise performance has little to do with the device employed and nearly everything to do with grounding.

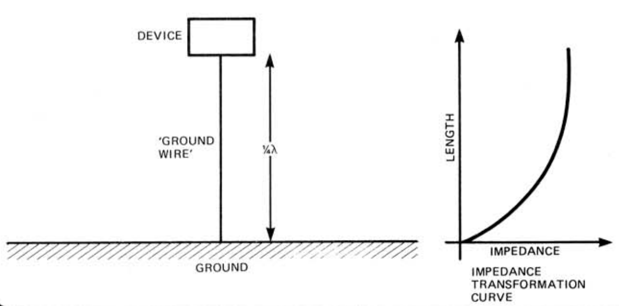
## . . . And higher up. . .

Noise generation due to grounds is not limited to the resistance predominant in the ground wiring at audio frequencies. At rf, well within the bandwidths of modern op-amps, even fairly short ground wires and busses can have very significant inductive reactances dramatically raising the effective ground impedance. This not so much reduces the isolation between the various stages as directly couples them together. All the inherent rf noise instabilities of the stages resulting become intermodulated (by the device's non-linearity at those frequencies) down to make their presence felt to audio earthlings as yet *more* audible and measurable noise.

A good 'shock-horror' example, which although described in simplistic theoretical terms, manifests itself sometimes dramatically in practice and can be called the 'Standing On One Leg Effect'.

The box in **Fig 36** represents a device that relies on the wire to be connected to the ground mass. Looks OK doesn't it? It is, apart from at the radio frequencies at which the wire is electrically ¼ wavelength or an odd multiple of ¼ wavelength. Our innocuous bit of wire turns into a tuned line transforming the 'zero' impedance of the ground to an 'infinite' impedance at the other end. Result — the device is totally decoupled from ground at those frequencies. Practical consequencies of this of course vary, from instability at very high frequencies on cards with long supply and ground leads to the author's most memorable encounter where an otherwise incurable case of TV signal demodulation in an electronic keyboard was fixed just by snipping a foot off the mains lead. ■

FIG. 36 THE 'STANDING ON ONE LEG' DISAPPEARING GROUND TRICK



# Designing a professional mixing console

Steve Dove

# Part Seven ~ Equalisers 1

THE TERM 'equalisation' is, strictly, a misnomer—it was originally utilised to describe flattening and generally putting to rights the response of systems in which by a matter of course or by design it had got a bit bent out of shape, eg telephone lines and tape machines. (In the latter case, the equalisation refers to the adjustment tweaks to the pre-emphasis and de-emphasis curves—not necessarily the curves themselves.)

In search of a name for the deliberate modification of amplitude and phase versus frequency response for 'the sound, man' and for the occasional genuine creative effect, the contraction 'eq' is well understood as both a noun and verb.

There is precious little in a modern studio that needs response modification to render it 'flat'—if there is, it needs mending or retiring, quick.

This sonic mutilation uses response curves, shapes and limits that have grown through an uneasy mixture of operator needs and technical expedience/feasibility—one of today's multi-parametric channel eqs would have needed nigh on a rack full of valves 20 years ago. Funny, they never seemed to need them then . . .

The delight (and maybe curse) of IC op-amp design is that active filter (hence eq) implementation and techniques have blown wide open, limited only by the largeness of the pcb and the smallness of the user's fingers.

Eq curves can be roughly lumped into three user categories: garbage disposal; trend and area. Highpass and lowpass filters that eliminate air-conditioning/mic stand rumble/breathing and excessive noise are obviously enough in the business of *garbage disposal.* Gentle 'hi-fi' type 'treble and bass' slopes and shelving establish response *trends,* whilst resonance-like 'bell' shaped lift and cut filters manipulate given *areas* of the overall response.

As the curves differ, so do the design techniques required.

**Purists call it Deliberate Frequency Selective Amplitude and Phase Distortion, studio operators call it Equalisation, textbooks and designers, a Royal Pain.**

## Single order networks

You can't build a house 'til you've got the bricks, so they say. **Fig 37** has the bricks, in the form of combinations of basic passive components with a rough guide to their input/output voltage transfer functions, with the assumption that Vi source impedance is zero and Vo termination infinite impedance.

Capacitative reactance decreases with increasing frequency, hence reacting against the resistance in a potentiometer-like fashion to increasingly 'short' the output to ground with increasing frequency in **Fig 37a** whilst steadily isolating the output from the input with reducing frequency (rising reactance) in **Fig 37b.**

Inductors have entirely the opposite reactive characteristics—inductive reactance is directionally proportional to frequency, so the curves in **Fig 37c** and **d** will be of no surprise at all.

"What about combinations of inductance and capacitance?" cry the anxious millions.

Shut up, sit down and wait a bit.

More useful curves are derived when the passive R, C and L elements are wrapped around an op-amp in the classic inverting and non-inverting amplifier modes—these are shown in **Fig 37e** to **l**. All the curves in **Fig 37** are normalised to unity gain and the same centre frequency, that being the frequency at which the curve departs significantly from flat. Standard arithmetic formulae normally consider or obtain a frequency at which the curve has departed 3dB from flat—the 3dB down point—and it is usually also where the phase has been shifted 45°.

To move the frequency at which the filter 'bites', any of the elements may be varied. Making them bigger makes the frequency lower, smaller-higher. An important point to remember is that whilst increasing inductance increases reactance at a given frequency, the inverse is true of capacitors. Bigger capacitor, smaller reactance.

There are an infinite number of combinations of element values to create the same curve at the same frequency. Say, in **Fig 37a** the value of the capacitor was reduced (increase in reactance) the filter curve would shift up in frequency. A corresponding increase in the series resistor value would result in the turnover frequency being restored to its original point. Identical filter, differing resistor/reactor combination. What *does* remain the same is the ratio or relationship between the two elements—it is only the filter *impedance* (the combination of resistance and reactance) that varies.

With the exception of a devious and evil few, any active filter's operation can eventually be sussed referring to these basic single order filter characteristics.

## Resonance

There is one particular combination of the two reactive elements (capacitance and inductance) that is of prime relevance to the construction of eqs. Shown in **Fig 38** it is a series connection of inductive and capacitative reactances.

In, for example, the context of a simple resistor/reactor filter **(Fig 37a)** the reactance not only causes an amplitude shift with frequency, but a related phase shift also. A fundamental difference between the two types of reactance is the direction of the output voltage (Vo) phase shift with respect to the source (Vi). More specifically, the capacitor in **Fig 37a** causes the output voltage phase to lag further behind the input as the rolloff progressively bites, to a limit of -90° at the dregs of the curve, whilst the inductor of **Fig 37** imposes an increasing voltage phase *lead* as the lf roll-off descends with a limit of +90° at maximum attenuation.

The two reactances therefore, in their pure unadulterated form effect phase shifts of +90° to -90°, in other words 180° apart, or in yet more words they are in *exact opposition* to each other. So?

So, referring again to **Fig 38** a slightly different light shines—the two reactances are working in direct opposition to each other, the inductive reactance is trying to cancel the capacitative reactance and vice versa. Arithmetically, it is (surprisingly enough) that simple — two reactance values may be directly subtracted from each other and the whole network treated as a single reactance of the same reactive character as the one predominant in the network.

As an example, if for a given frequency the inductive reactance is (+)1,200 (the + indicating the phase shift character of inductance) and the capacitative reactance is (-)1,500Ω, then the effective reactance of the entire network is that of a capacitor (-) of 300Ω reactance. With a change of frequency, the two reactances will shift, one up, one down giving another network reactance resultant. (As a by-the-way, because there are two reactive effects operating simultaneously in this network, it is said to have *second-order* characteristics.)

For any pair of inductor/capacitors at any frequency their two reactances will still be equal. Thinks! If you subtract two equal numbers the answer is nothing.

Eh?

At that frequency, the two reactances cancel completely resulting in a short circuit across the network terminals, no reactance, nothing (disallowing component losses).

A frequency-selective short circuit. Either side of that frequency of course one or other of the reactances become predominant again.

Like the single order networks, there's an infinite number of combinations of C and L that will have resonance (the two reactances will equal) at any given frequency and the relative values and rate of change in reactance either side of resonance hinges on the chosen combination. Say a given L/C ratio gives a reactance of 10kΩ detuned 10% from resonance. Changing the L and C (maintaining the same resonant frequency) to make their reactances

FIG. 37 SINGLE ORDER FILTERS

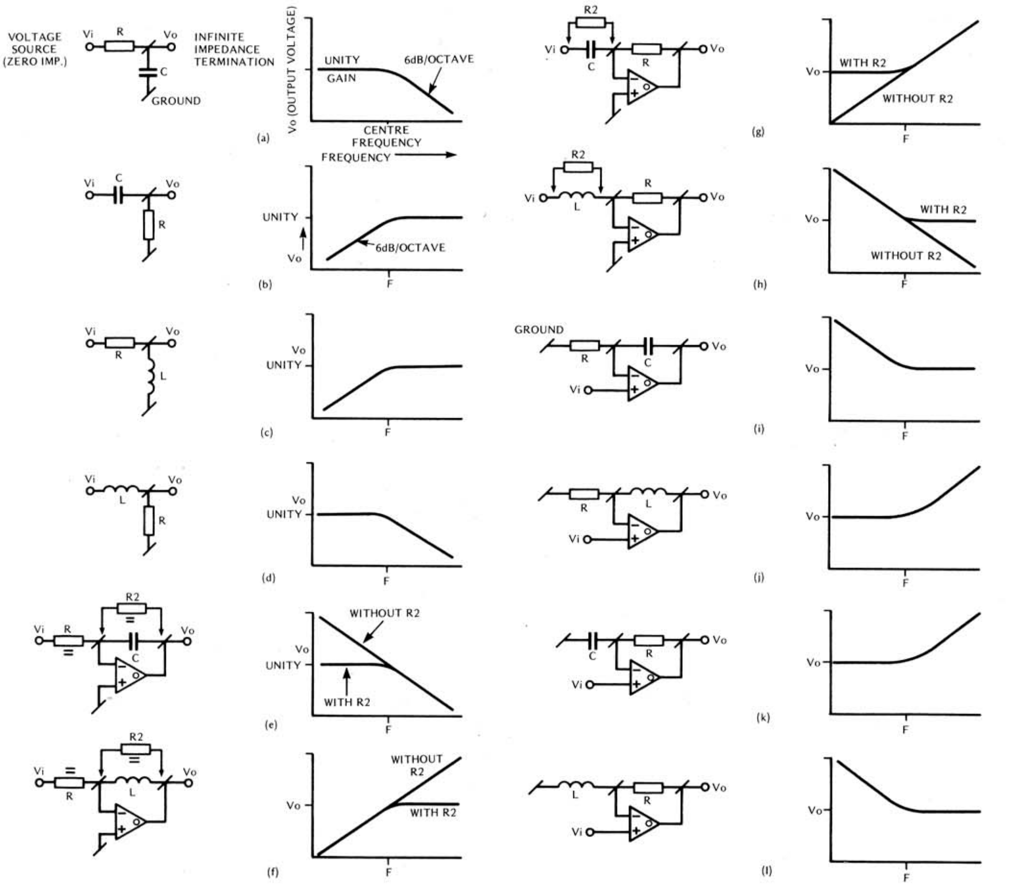
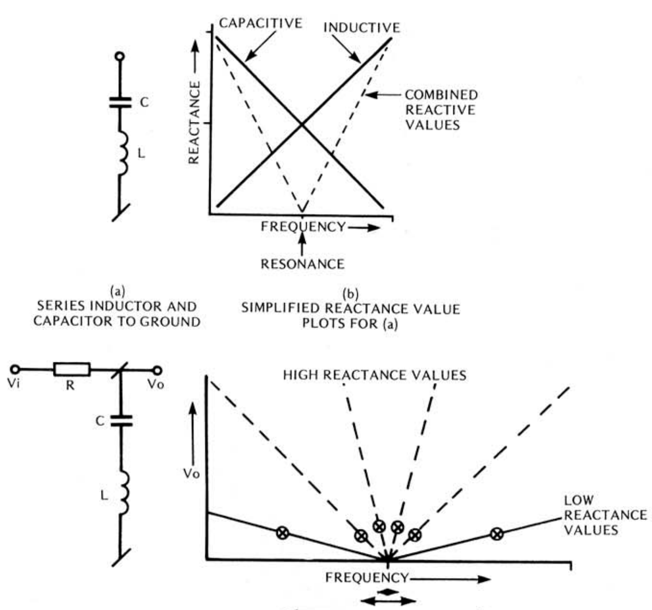



FIG. 38 RESONANT CIRCUITS




(a) SERIES INDUCTOR AND CAPACITOR TO GROUND

(b) SIMPLIFIED REACTANCE VALUE PLOTS FOR (a)

(c) NOTCH FILTERS

BANDWIDTH FOR VARIOUS REACTIVE VALUES (3dB DOWN POINT/45° PHASE SHIFT) (d)

a tenth of their previous value (making the L 10 times smaller and the C 10 times bigger) makes the 10% detune reactances only 1kΩ.

The relevance of this is rapidly seen looking at the notch filter circuit in **Fig 38c.** At any frequency, the series resistor forms a classic 'L' attenuator against whatever reactance the L/C network is 'pretending' to be, total attenuation occurring at the network's resonant frequency. The rate of reactance change upon detune is directly related to the reactances of the network's constituent L and C. A fixed value of series resistor looking into our two earlier mythical 10kΩ and 1kΩ reactance value networks will cause a detune slope 10 times more steep with the higher reactance network than with the 1kΩ one—in other words the higher reactance network has a sharper notch filter effect, less bandwidth and a higher Q than the 1kΩ network. By a factor of 10, surprise, surprise.

There exists a direct relationship between the network reactance, series resistance, the bandwidth and Q. Q is numerically equal to the ratio of elemental reactance to the resistance in a series tuned circuit, whilst the bandwidth (between the two '3dB down' points where the phase has been shifted ±45°) is the ratio of filter centre frequency to Q.

The greater the Q, the smaller the bandwidth.

Filter resonant frequency may be altered by changing either the L or C, whilst Q is subject to variation of the resistor or juggling the reactance of the L/C network.

## Creating inductance

It's most efficient (electrically and financially) in the majority of console-type circuitry for inductance to be simulated or generated artificially by circuits that are the practical implementation of a mathematical conjuring trick—known generically as 'gyrators'.

A true gyrator is a 4-terminal device that transmutes any reactance or impedance presented to one port into a mirror image form at the other port **(Fig 39).**

Hence a capacitor (with falling reactance vs frequency) is magically translated into a reactance of rising characteristic vs frequency at the output port, *voila!* inductance! The scale of inductive reactance generated may be easily and continuously varied by altering the internal gain-balance structure of the gyrator (in **Fig 39b** by changing the transconductance of the back-to-back amps).

A continuously variable inductor!

Real inductors—the things with miles of wire knotted around odd-shaped bits of ferrite or some such—have a justifiably bad name for audio design. They are big, heavy, they saturate easily, their core hysteresis causes distortion, they are subject to pick-up of nearby (and not so nearby) magnetic fields prin-

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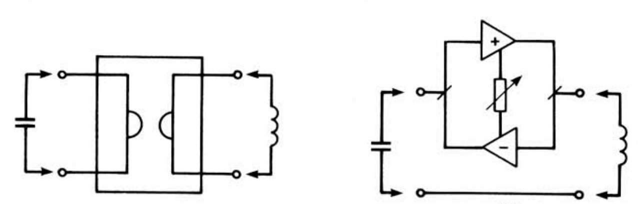
FIG. 39 GYRATORS



(a) 'BLACK BOX'

(b) USING OPPOSED TRANSCONDUCTANCE AMPLIFIERS cipally mains ac hum and rf unless well screened (which makes them even bigger and heavier) the windings are prone to break and they are e.x.p.e.n.s.i.v.e.

It is therefore quite easy to see why a means of avoiding them is popular! Naturally, the simulated inductive reactance is only as good (the inductor Q) as the quality of the capacitative reactance (the capacitor Q—determined by its leakage resistance) and the loading effect of the 'gyrator' circuit itself. Fortunately, for the purposes of normal equalisers very large Qs are not necessary so selecting capacitor types to this particular end is not really necessary.

An obvious extension of the continuously variable inductor is the continuously variable bandpass filter formed by adding a capacitor either in series or parallel with the gyrated inductor, forming series and parallel tuned circuits respectively making notch and peak filters. Whilst ideal for fixed frequency filters with the network's Q or sharpness defined by a resistor in series with the gyrator resonator, the idea falls down when the resonance frequency is moved.

If the frequency is moved *up,* the reactances of the elements at resonance become lower, consequently the ratio of the reactances to the fixed series resistor (this is the ratio that determines the Q) becomes smaller and the Q of the filter becomes broader in response relatively. In order to maintain the same Q over the projected frequency variation the series resistor has to be ganged with the frequency control—boring. Should it be necessary to make the Q a variable function also, as in a parametric-type eq section, it would mean devising a variable-variable resistor—brain-stranglingly boring. For this reason, parametric-type equaliser sections are ordinarily constructed around second-order active-filter networks, typically of the State-Variable variety.

Let's not write off gyration for function variable filters straight away —as we'll see they form in one way or another the second reactance in many active filters anyway.

True gyrators of the back-to-back transconductance amp type are, let it be said, an unmitigated drag to make, set up and use. Fortunately there are simpler ways of simulating variable reactances, if not pure reactance at least a predictable effect of a reactive/resistive network.

## The bootstrap

The simplest of the lot is shown in **Fig 40a,** with typical values shown for argument's sake. It relies on a wonderful trick called 'bootstrapping'. The principals behind this trick are shown in **Fig 41.** A 1kΩ resistor with a volt across it will pass 1mA, so says Mr Ohm in his well known law. Without changing the source potential of 1V, the bottom end of the resistor is tied to 0.8V. There is 0.2V across the resistor and so a current of 0.2mA flows through the resistor. Aha! The clever bit! The source (still at 1V) sees 0.2mA flowing away from it, the amount of current it would expect to see going to a resistor of 1V/0.2mA = 5kΩ. It 'thinks' it's looking at a 5kΩ resistor!

Continuing this, stuffing a potential of 1V (not the same source) at the bottom end of the resistor means there is no voltage across the resistor, therefore no current flow and our original source 'thinks' it's seeing an open circuit (infinite resistance) despite the fact that there is still a 1kΩ resistor hanging on it.

This phenomenon holds true with any source voltage, ac or dc, provided the instantaneous 'bootstrap' voltage is the same as the source. This implies in ac (eg audio) the bootstrapping is exactly in phase with the source—any phase difference creates an instantaneous potential difference across the resistor, current flows, etc, etc.

The 'fake inductor' works on frequency dependent bootstrapping, the terminal being almost totally bootstrapped to high impedance via the 150Ω resistor at high frequencies and the bootstrap voltage reducing (together with its phase being shifted) with falling frequency. At very low frequencies, no bootstrap exists, so the terminal is tied to ground via the 150Ω resistor and the effective zero output impedance of the voltage follower. The circuit emulates an inductor reasonably well—very low impedance value at low frequencies, increasing with frequency to quite a high, virtually open circuit, impedance.

Problem No 1 with this simple circuit is that at high frequencies a parallel impedance (consisting of the variable resistor/capacitor chain) hangs directly from the terminal to ground. Buffering the chain from the terminal by a follower eliminates that one **(Fig 40c).**

**Fig 40a** creates an analogue of an inductor with losses as shown in **Fig 40b.** The series resistor is the 150Ω bootstrap resistor—after all a 'proper' inductive reactance tends to zero at low frequencies, not 150Ω, therefore the resistor is effectively in series. The R/C network across the lot represents, again, the highpass filter impedance which upon the addition of the follower disappears to be replaced in **Fig 40d** by the follower's input impedance—a lot higher and enough to be ignored.

Losing the effective series impedance of the bootstrap resistor is hassle No 2. A fascinating circuit of wondrous properties but previously of little real worth smiles at us in **Fig 40e.** Bearing more than a little resemblance to a differential amp, this circuit can rotate the output phase through 180° with respect to the input around the frequency primarily determined by the highpass filter, $R_1$ $C_1$. Not only but also, the amplitude remains constant throughout.

How? This is dealt with in **Fig 40g** and **h** where the simplistic assumptions that a capacitor is open circuit at low frequencies and a short at

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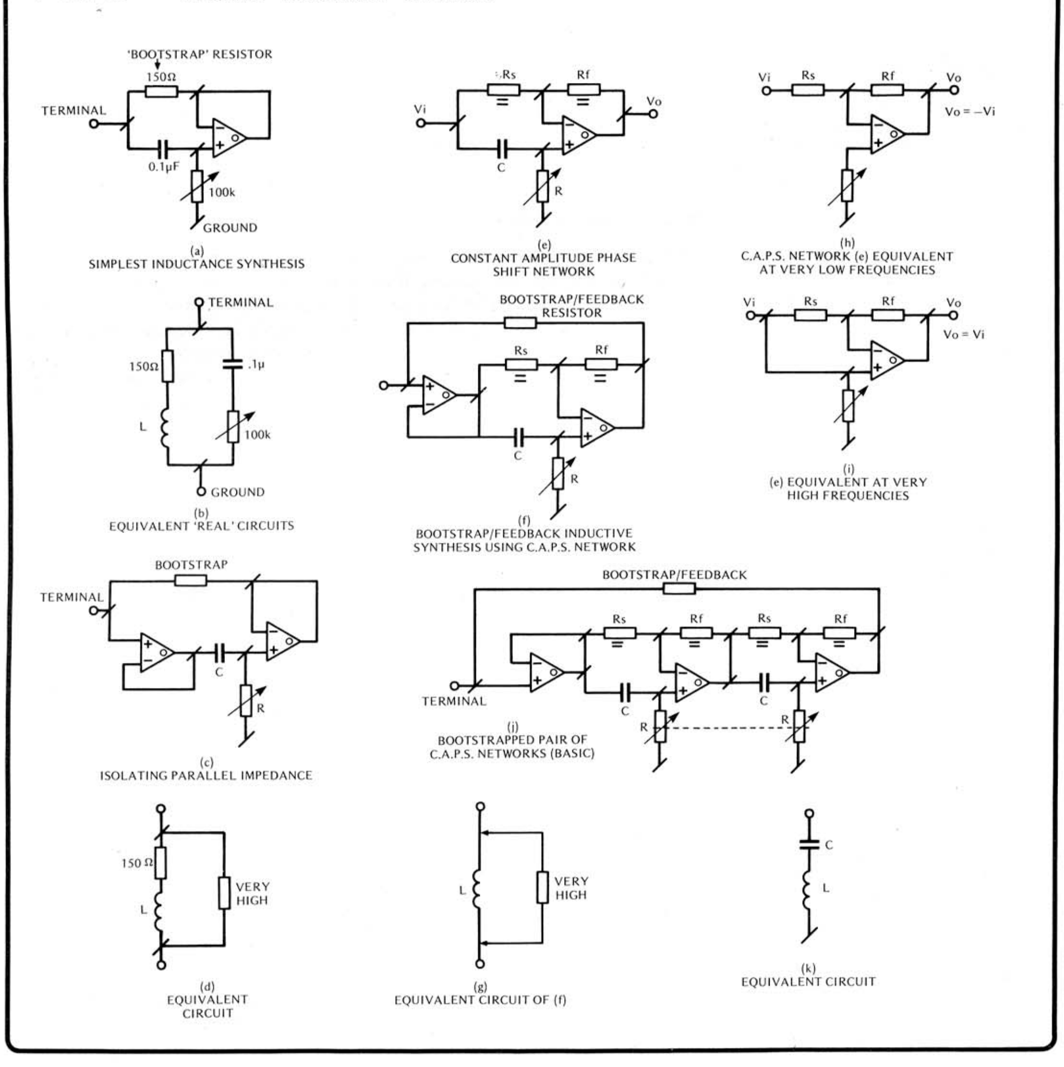
FIG. 40 INDUCTIVE REACTANCE SYNTHESIS



# Mixing console

high frequencies show that at lf the circuit operates as a straightforward unity-gain inverting amp (-180°) whilst at hf it operates as a gain-of-two non-inverting amp minus a gain of -1 due to the inverting amp chain Rf/Rs—in other words a unity gain non-inverting amp (0°).

At high frequencies bootstrapping back to the input is cool **(Fig 40f)** providing the expected lovely high impedance. The nice bit occurs at lf where the phase rotates around to -180°. The output amp generates an equal and opposite current along the 'bootstrap' to any which are supplied to the input terminal. Translated that means that the circuit has turned into an inverting amp and is treating the terminal as a virtual earth point via the 'bootstrap' (now 'feedback') resistor. Virtual earth means virtually zero impedance. Neat.

As a short footnote to this gyrator epic, consider what happens to either **Fig 40c** or **Fig 40f** if the highpass filter C/R is replaced by a lowpass filter by transposing $R_1$ with $C_1$ and vice versa. It may seem a bit dumb to use circuitry to imitate a capacitor—but a continuously variable capacitor . . .?

## Simulated resonance

We now possess all the variable-everythings we need to create single and second order filters. Tracking variable capacitors and inductors allow us to manufacture constant Q bandpass filters irrespective of frequency—this realisation itself brings a dawning of understanding in how the much-touted loop filters such as the state-variable actually operate. The clue lies with the 180° phase shift circuit **(Fig 40e)**. Connecting two such filters (with the variable resistor elements ganged) in series produces a remarkable circuit. At any frequency within the design swing it is possible for the circuit output to be 180° out of phase with the input—and only at that frequency. Combining the input voltage and the output voltage in a separate amp results in direct cancellation at that frequency and at no other—in short a notch filter with a nice resonant characteristic. Alternatively, bootstrapping the input from the output actually turns that input port into something that behaves exactly like a series tuned circuit to ground **(Fig 40j)**. Continuously variable in frequency with a constant Q to boot by virtue of the simultaneously tracking simulated inductor and capacitor maintaining exactly the same elemental reactances at whatever the resonant frequency is adjusted to. Same source resistance, same reactance, same Q.

Same Q definitely does not imply same *bandwidth*—as the resonant frequency changes, the bandwidth changes proportionally. Bandwidth is after all the ratio of frequency to Q.

Some active filters, such as the 'multi-feedback' variety (of which more anon, but not here) exhibit a constant bandwidth vs resonant frequency characteristic—meaning if it has say a 400Hz 3dB down point bandwidth at 5kHz, it will also show a 400Hz bandwidth when the resonant frequency is changed to 500Hz. A 10:1 variation of Q.

This, on the surface at least, appalling characteristic has been deliberately and usefully used in a mid-sweep eq providing the user with a broad low-end to 'fatten things up' automatically changing to a sharper filter higher up to pull out 'rings', sibilance and assorted screeches, which benefit from the high Q attenuation not molesting too much of the surrounding.

It has in fact received more praise than criticism, despite the obvious limitations.

## Mechanical filters

Achieving prominence in TV, radio and communication-type signal processing is the Surface Acoustic Wave type filter which has in the last few years almost totally displaced conventional multi-element inductor/capacitor resonant and bandpass arrays. Distantly related to the quartz crystal effect—resonance in a piece of solid material at a precisely consistent frequency, SAW filters establish precise acoustic transference and interference patterns across a piece of solid material. The filter is excited at one end and sensed at the other—a bit like a reverb plate, only the plate's characteristics would have to be precisely trimmed to transfer energy along only between very sharply defined frequency limits, say 1kHz to 1.2kHz only and to reject all others. An SAW block can be arranged to have an almost perfect rectangular response with a desperately plummeting fall-off either side of the bandpass.

Current technology places practical such filters at 10MHz and above, commonly being used for receiver intermediate-frequency band-shaping.

The extremely sharp fall-off shape still looks appealing when applied to audio thinking—90dB attenuation within 1kHz is typical for a narrow-band filter (25kHz) at 10.7MHz.

A practical technique for utilising these properties is sketched in **Fig 42a.** It is essentially a loop modulation/demodulation system, the double balanced modulators being constructed of transfer-matched hot-carrier diodes fed by identical amplitude and phase of the oscillator to minimise mod/demod discontinuity distortion.

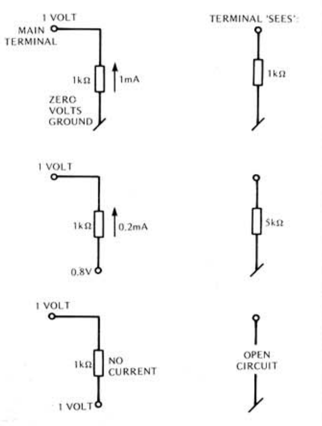
The oscillator is centred on 10.6875MHz, the lower frequency limit of the bandpass filter. This, in its 'resting' mode, disposes of the modulated signal's lower sideband over the filter's 'cliff edge'. Now consider what happens as the oscillator frequency moves up 50Hz—yes, everything below 50Hz is pushed over the precipice. Increasing the frequency to 10.6885MHz (1kHz up) acts as an astonishingly sharp 1kHz highpass filter. Softening the roll-off characteristic is achieved by a really cunning trick; the oscillator is *frequency modulated* ultrasonically (so as not to be audible) backwards and forwards over the part of the bandwidth that needs shallowing out—say from 1kHz down to 250Hz. **Fig 42c** shows the effect of an Equal Frequency Domain Duration (EFDD) ultrasonic waveform (such as a triangle wave) on the demodulated audio output.

A variation on the FDD technique is used to overcome a basic hang-up of the system—a fixed roll-off filter such as the SAW results in a wide range of effective roll-off rates when applied to the logarithmic-based audio band. The answer is to add a 'biasing' logarithmically asymmetric component to the ultrasonic FM. This ensures that at higher filter frequencies, the 'softening' FDD waveform is modulated wider so closely approximating at 1kHz roll-off the same number of dB/octave attenuation as down at 50Hz.

The fundamental problem with SAW filtering is in-band attenuation—the element itself has at best 18dB loss, there is 6dB loss due to one sideband being removed and each of the balanced modulators incurs –7dB conversion gain—a grand total of 38dB to re-establish in the make-up amp.

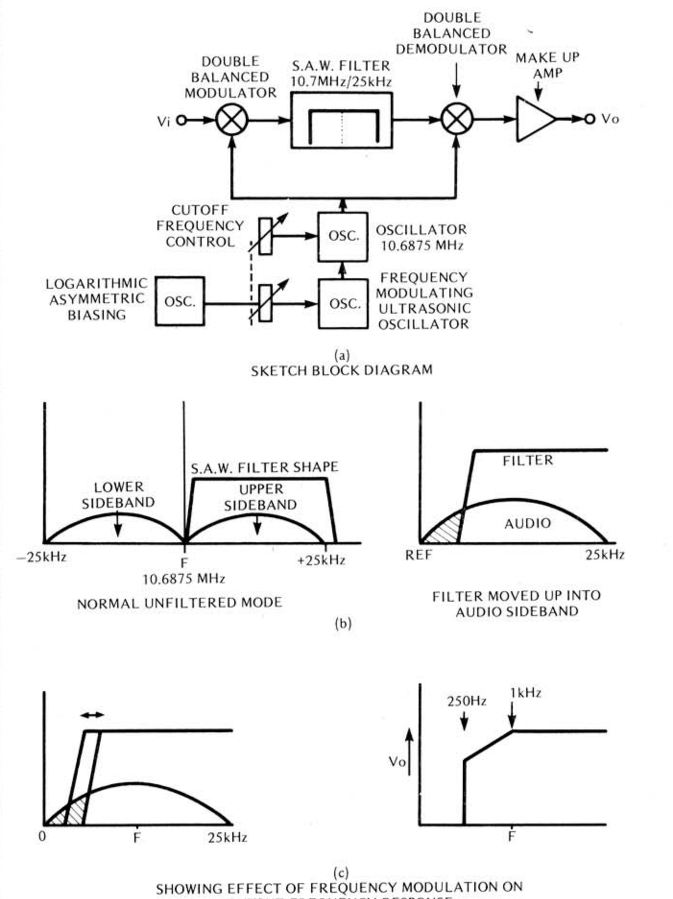
For the present, this technique is a bit noisy. ■

FIG. 41 BOOTSTRAPPING




NOTE: 1kΩ RESISTOR IS ATTACHED TO TERMINAL IN ALL CASES

FIG. 42



# Designing a professional mixing console

Steve Dove

# Part Eight ~ Equalisers 2

**We hope you enjoyed our April bonus at the end of Equalisers Part One "Mechanical Filters". Although plausible, as you will appreciate this system is an expensive and elaborate means of achieving a high-pass filter effect. Avid constructors should note that construction should have been completed by 12pm, April 1, and that the accompanying Fig 42 is erroneous. We promise to return to the straight and narrow henceforward, accordingly Fig 42 below and its accompanying text replaces the previous entry.**

METHODS of filtering come thick and fast once the basics are established. The development of a perfectly popular bandpass filter arrangement is shown in **Fig 42.** It starts out in life as two variable passive single-order filters of a common 'crossover' frequency point, ganged so that they track. Reconfigured slightly **(Fig 42b)** to minimise interaction, they are shown with their drive and sense amps. Wrapping the two networks around an inverting amp isolates them completely from each other, improving the filter shape. The bandpass Q is really rather low, well under one in fact, leaving it rather limited in scope for practical applications. A discretionary degree of positive feedback from the amp output back to the non-inverting input smartens up the Q, if a little unpredictably and more than critical of component tolerances.

Yes, viewers, it does look rather like a Wien Bridge oscillator, doesn't it? Attempting to get the Q too high proves the point unquestionably!

## Listening to Q

This raises the problems of excessive Qs. Fortunately, extremely high Qs (rather, any greater than 10), are unnecessary or unusable for eq purposes. The higher the Q becomes, the less actual spectral content of the signal it modifies so despite the fact that its peak gain or attenuation is the same as a lower Q filter, it seems to do subjectively *less.* Judicious care is required in setting up the filter to enhance or trim exactly what is required—accidental overkill is easy.

There comes a break-point with increasing Q where you are not so much listening to the filter's effect as to the filter itself. Resonant tuned circuits are to a degree electrical storage media, where energy inside the circuit shuffles backwards and forwards between the two reactive elements until the circuit losses waste it away. The greater the Q (and by definition the lower the included losses) the more pronounced the signal storage.

Whilst recently playing with silly-Q filters using synthesised L and C elements of virtually eliminated intrinsic losses, an 80Hz bandpass filter of some 0.3Hz bandwidth (Q about 250) continued to 'ring' for quite a few *seconds* after the input signal was removed—a beautiful sinewave slowly decaying away. Despite being good for a laugh, it's of no value at all in a practical eq. A transient hitting such a filter triggers a virtually identical series of decaying sinewaves at the frequency of the filter—who needs that?

Squarewaves sent through audio paths are good for kicking resonant ringing off at almost any frequency—it's a convenient means of unearthing inadvertent response bumps and lurking instabilities. The breakpoint—where you are starting to hear ringing as much as signal—is quite low, a Q of between 5 and 10.

## Squegg or slug?

It is not too difficult now to appreciate that resonant circuits and oscillators are very close cousins —often indistinguishable, except for maybe an off component value here and there. There are two fundamental approaches to achieving a resonant bandpass characteristic using active-filter techniques.

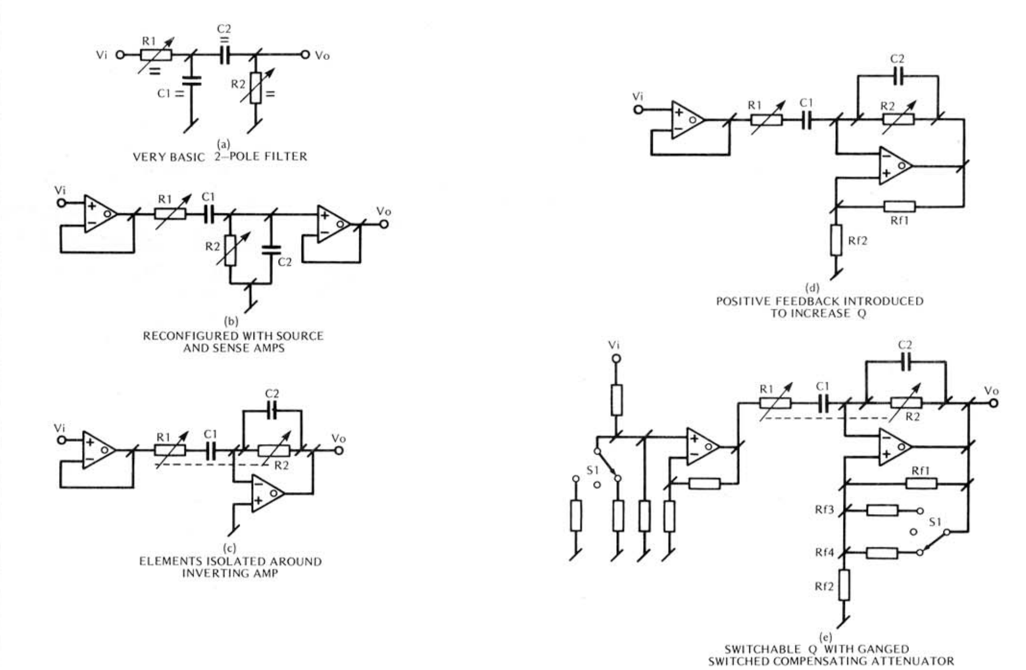
The first is to start off with a tame, poorly-performing passive network and then introduce positive feedback to make it predictably (you hope) unstable—the feedback exaggerating the filter character, increasing Q to the desired extent. A perfect example of this is the 'Wein Bridge' development of **Fig 42.** The major disadvantage of such methods is that the Q is disproportionately critical with respect to the feedback adjustments.

The second approach is to start off with an oscillator, then slug it until it's tame enough. This is the basis of the state-variable, the bi-quad and similar related loop-type active filters.

## The 2-integrator loop

Three inverting amps connected together in a loop, as in **Fig 43,** seems

FIG. 42 A BANDPASS FILTER DEVELOPMENT



a perfectly worthless circuit, as such it is. It's there to show (assuming perfect op-amps) that it is a perfectly stable arrangement—each stage inverts (180°), so the first amp section receives a perfectly out-of-phase (invert, revert, invert) feedback so cancelling any tendency within the loop to drift or wobble. Removing 180° of phase shift would result in perfect in-phase *positive* feedback; the result would require scraping off the ceiling.

Arranging for the 180° to be lost only at one specific frequency results in the circuit being totally unstable at just that one frequency—in other words it oscillates controllably. Creating the 180° phase loss is left to two of the inverting amps being made into integrators **(Fig 43b)**, so called because they are an electrical analogue of the mathematical function integration.

The integrator you may recognise from a single-order filter variation in **Fig 37.** It's not so much the amplitude response that's useful here as the phase response, which at a given frequency (dictated by the R and C values) reaches −90° with respect to the input. Two successive ganged-value integrators—presto, −180° shift.

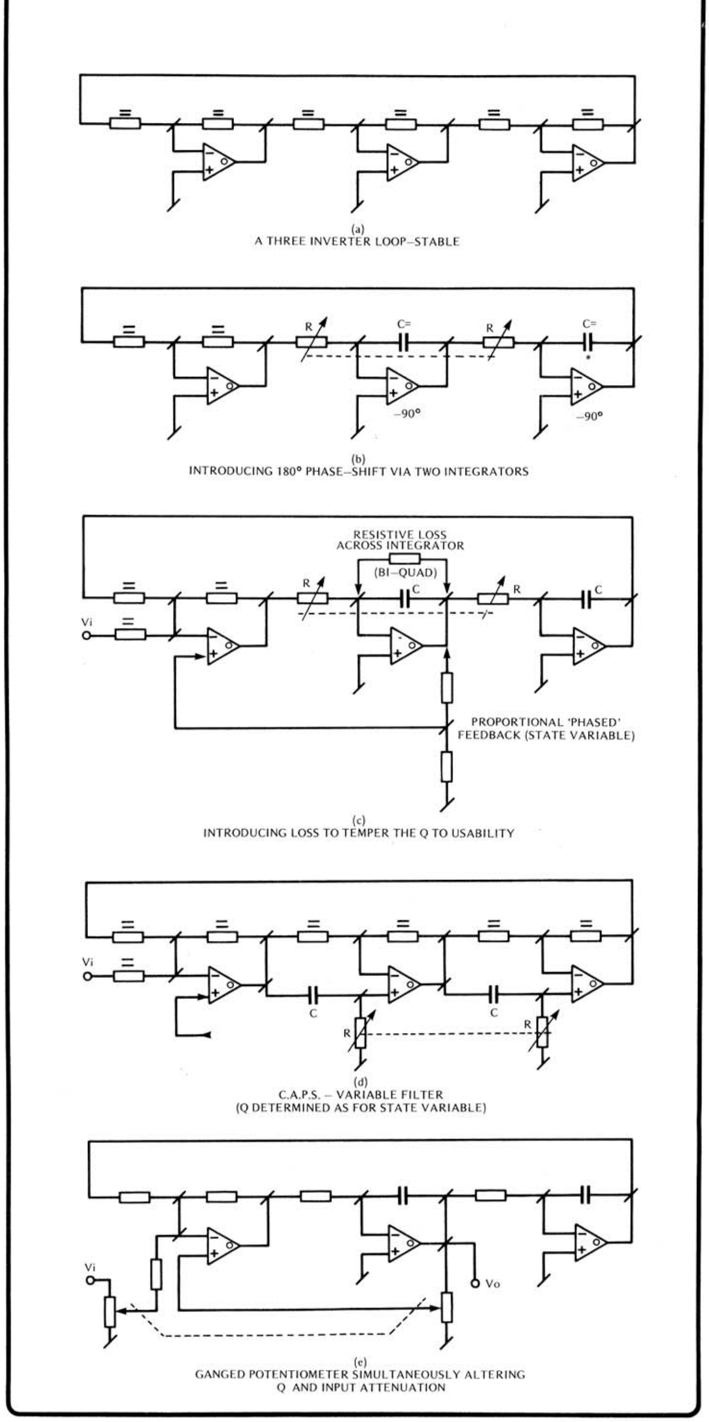
Slugging the loop to stop it oscillating can be achieved in a variety of ways:

(a) Trimming the gain of the remaining inverter—this is unduly critical like the 'Wein Bridge' for Q determination.

(b) Doping one of the integrator capacitors with a resistor **(Fig 43c).** This in essence is the 'bi-quad'. The Q is largely dependent on the ratio of the capacitative reactance to the parallel resistance, which consequentially varies proportionally with frequency. For fixed frequency applications the bi-quad is easy, docile and predictable.

(c) Phased 'negative' feedback. Not true negative feedback but taken from the output of the first integrator (90° shift). This provides an easily managed Q variation, constant and independent of filter frequency **(Fig 43c).** Forming the basis of the state-variable filter, this has turned out to be 'The Active Filter Most Likely To Succeed', if the majority of current commercial console designs are to be believed.

FIG. 43 2-INTEGRATOR LOOP FILTER




Panaceas, once you've kicked the stone over and grubbed around underneath it for a bit, are never quite as tasty afterwards.

Loop filters, such as described in **Fig 43c**, have a number of inherent bothers that are usually glossed over for the sake of the design's operational simplicity and elegance.

Each amp within the loop has a finite time delay, which together add up to significant phase shifts within the open-loop bandwidths of the amps—some get simply added to the delay imparted by the integrators but the total time discontinuity around the summing amp can promote instability in the multi-MHz region. Compensation for this around the summing amp can introduce further phase shifts upsetting the filter performance at high frequencies. Take your pick.

Two major problems are due to the nature of the integrator arrangement itself. They come to light at the extremes of the feedback capacitance's reactance, ie at very low and very high frequencies where respectively the reactances are virtually open-circuit and short-circuit.

Open-circuit at lf means the op-amp is 'infinitely' amplifying external resistor noise and internally generated thermal and (mostly) lf noise, plus any lf noise presented to the input along with the signal. In fact, and as far as the outside world is concerned, the entire loop feeds round and defines the 'gain', but each individual amp generates and amplifies *a lot* of lf noise.

At high frequencies, the reactance approaches short-circuit connecting the output back around to the inverting input. This arrangement, zero closed-loop gain, is about as critical in terms of device instability as you can get (see Part Three, Op-Amps—Friend or Foe? November 1980) since there is no possible way of further externally defining the closed loop characteristics beyond that of the integrating capacitor itself, which may or may not be adequate.

For audio frequencies, the integration capacitor value can be quite sizable—up to 1μF, say. If there isn't an immediate problem of the op-amp's current output capacity being incapable of charging such a capacitor instantaneously, there will almost certainly be a problem related to the device's open loop output impedance; this corresponds to a resistor in series with the device output, which obviously enough forms a time constant and a filter within the integrator capacitor. Time constant means *more* time delay in the loop, whilst the implicit lowpass filter around the immediate op-amp means a reciprocal rise in hf response, stealing from the op-amp's stability phase margin.

As tame as it may superficially seem, the state-variable is not an unconditionally or reliably stable arrangement, with out-of-band dynamic state nasties potentially degrading its sonic performance.

## Improving the loop

With the exception of inevitable loop effects, most of the undesirable things about the state-variable can be eliminated or mitigated by replacing the integrators by constant amplitude phase shift elements, **(Fig 43e)**, resulting in what could best be known as a CAPS-variable filter.

All the constituent elements being basically stable with provision for independent compensation—with no undefined gain for any of the spectrum—seems like a healthier format to start making filters around. Stabilisation is as for the state-variable.

There is another way of looking at the state-variable/CAPS-variable filter that will suddenly resolve the previous discussions on gyrators, L and C filters, series tuned circuits, etc, with the seemingly-at-odds approach of active filters.

Resonance depends upon the reaction of two reactances of opposite sense—180° apart in phase effect. Rather than achieve this in a differential manner, one element +90° with the other −90° at a given frequency, active filters achieve the total difference by summing same-sense phase differences, (−90°) + (−90°) ie, still 180°. Two reactive networks are still involved, hence it is still a second order effect. At the end of the day the principal difference is that such loop type active filters have their median 'resonance' phase displaced by 90° from their input as a result of both reactances 'going the same way', as opposed to the nil

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phase shift at resonance of a 'real' L/C network.

## Q and filter gain

Pretty much every resonant type active filter has the unfortunate characteristic of its gain at resonance being at least related and often directly proportional numerically to the Q of the filter. This means a filter with a Q of 10 usually has a voltage gain of 10 (or 20dB gain) at resonance.

Naturally, this is 'not on'. Even specifying a maximum Q of five only helps by 6dB of lift.

That represents a very sizable chunk of system headroom stolen at the filter frequency, also making the 'sum and difference' matrixing necessary to provide the usual 'lift and cut' facilities difficult to configure.

The obvious solution is to attenuate the signal going into the filter by the same amount as the gain (hence Q) expected of the filter. Arranging a continuously variable Q control that also attenuates the source appropriately is not a conspicuously simple task, at least with most filters. Perhaps the most straightforward example is shown in **Fig 43e**, a state-variable type filter with an attenuator in the 'slug-back' network altering the Q ganged with an attenuator ahead of the input/summing amp.

Within a couple of dB, this holds the resonant peak output constant over a considerable useful Q range.

Most other filters are not so obliging in terms of continuously variable-Q. Switching between a few values of Q whilst substituting appropriate input attenuation is quite often a practical and operationally acceptable solution, applicable to nearly any filtering approach. **Fig 42e** illustrates a further development of the Wein Bridge arrangement using this method to provide alternative Qs. The attenuator values are necessarily high in impedance to prevent excessive loading of the source, a factor which in some practical eq circumstances is important.

## Highpass filters

Two stone-age single order highpass filters are shown in **Fig 37** (Part Seven, Equalisers 1), the keys being the reducing inductive reactance to ground with reducing frequency in **Fig 37c** and the rising capacitative reactance against reducing frequency in **Fig 37b**.

How about combining the two, omitting the resistors as in **Fig 44a**? As expected, the two opposing reactances combining result in an ultimate roll-off twice as fast as one of the single orders but they have also resulted in a resonance peak at the point of equal reactance. Well, resonance Q is the ratio of elemental reactance to resistance, so deliberately introducing loss in the circuit in the form of a termination resistor tames the resonance to leave a nice, flat, in-band response **(Fig 44b).**

Substituting a basic 'gyrator' or simulated inductance for the 'real' one **(Fig 44c)** naturally works just as well, and even better than expected. The filter output can be taken straight from the 'gyrator' amp output to start off with, saving having to use another one as an output buffer.

Secondly, we can automatically introduce the required amount of loss into the inductor by increasing the value of the bootstrap resistor and get the resonance damping right. (Refer to discussion of 'gyrators', in Part Seven).

Thirdly, we can easily change the turnover frequency of the filter by varying what was the 'tuning' resistor. In doing this, of course, the elemental reactance to loss ratio will change causing the Q (hence damping factor) to change with it. No tears. The frequency change and required damping change are directly related and may be simultaneously altered with a ganged control—even if we do our sums right, with the two ganged tracks having the same value!

A slight redraw of **Fig 44c** gives **Fig 44d**—a more conventional portrayal of the classic Sallen and Key highpass filter arrangement. Well I never . . .!

As the Sallen and Key filter evolves, it is seen that an 'equal value' filter (where the two capacitors are equal, the two resistors are equal) results in a less than adequate response shape, with a fairly pronounced resonant peak of one or two dB—demanding further damping. An expedient method, working on the assumption that a few more resistors are cheaper than a special two-value ganged potentiometer, is to increase the damping by introducing gain into the 'gyrator' buffer amplifier (also providing a means of stability compensating it correctly)—see **Fig 44e.** A side effect of this technique of damping adjustment (which incidentally is independent of filter frequency) is that an input/output in-band gain is introduced. This may or may not be problematic. The 4dB-ish gain introduced necessary to render the filter frequency response maximally flat could be included in overall system gain, or alternatively a compensating attenuator could be instituted ahead of it. This could also be arranged to be a fixed frequency band-end single order highpass filter to accelerate the slope out-of-band.

## Second or third or more, order?

Because of their simplicity, it is tempting to go overboard on band-end filtering and it is mostly designers' faults because they rarely get a chance to listen to their results.

Without delving too deeply into psychoacoustics, the ear notices easily a third or more order filter being introduced for much the same reasons as a high-Q bandpass filter is obvious—severe modifications to the signal path's transient response and the introduction of 'ringing' type time-related components into the signal's spectrum.

An application where this effect is not overly objectionable is where the filters are defining bandwidth at audible limits (such as in the line amp/filter described in Part Two—Broadcast Consoles, October 1980). Within the audible band, though, the ear is quite merciless toward such noises.

The transient response modification is not the end of the story, it being that the drastic modification of the relationships between the fundamental frequency and the harmonics of instruments in the operating turnover area of the filter is likely to be interpreted as 'unnatural'.

Second order filters score in both respects—there is less transient response disturbance and less tonal characteristic modification—there are few who would dispute that they sound more natural and (ducking to avoid flying objects) 'musical'. A personal wrinkle, for which provision is made in the design, is to leave a small controlled amount of under-damped 'bump' in the filter frequency response. This has two consequences, one being slightly more rapid out-of-band roll-off but the other—a subjective effect—is that the extra programme energy introduced by the 'hump' serves to help offset the loss of energy in programme below the turnover frequency. The perceived effect upon introducing the filter is more of a 'change in sound' rather than a direct drop in lf response.

This raises an interesting possible line of debate: Should equipment in the recording chain (not just consoles) be designed and set up to be perfect according to conventional techniques and wisdom, or to sound 'right'? ■

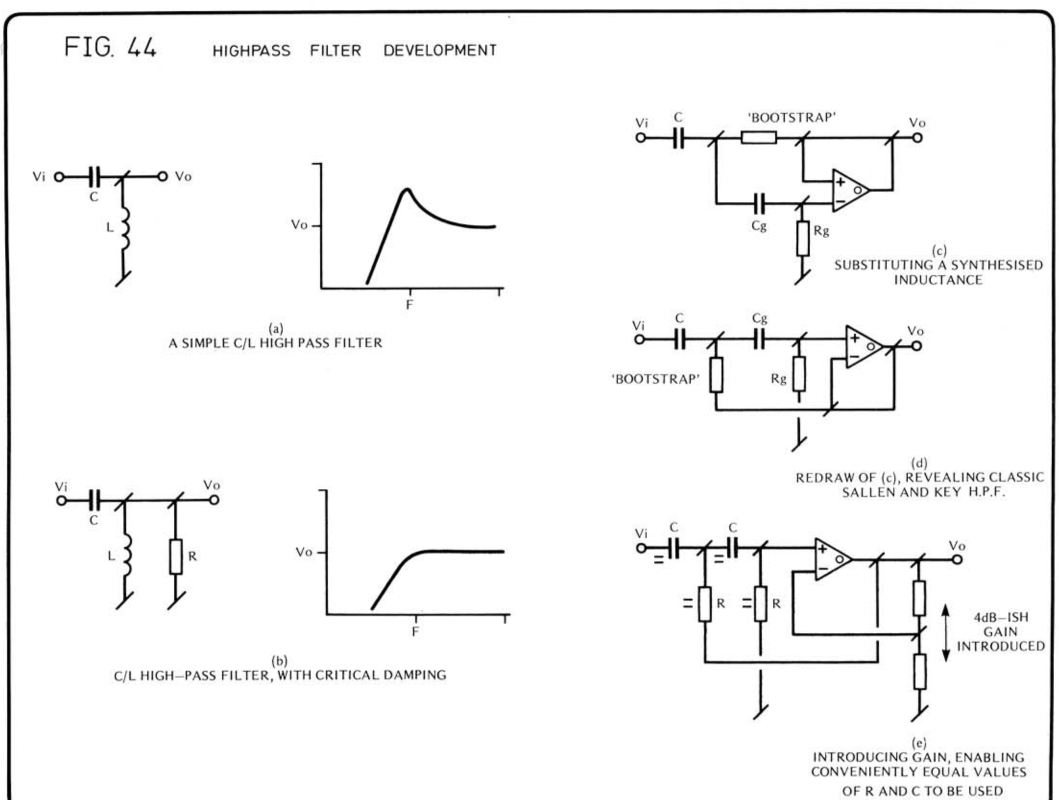



FIG. 44 HIGHPASS FILTER DEVELOPMENT

(a) A SIMPLE C/L HIGH PASS FILTER

(b) C/L HIGH–PASS FILTER, WITH CRITICAL DAMPING

(c) SUBSTITUTING A SYNTHESISED INDUCTANCE

(d) REDRAW OF (c), REVEALING CLASSIC SALLEN AND KEY H.P.F.

(e) INTRODUCING GAIN, ENABLING CONVENIENTLY EQUAL VALUES OF R AND C TO BE USED

# Designing a professional mixing console

Steve Dove

# Part Nine ~ Equalisers 3

**Parts seven and eight were primarily concerned with detailing active filter techniques useful as amplitude and phase response shaping elements. This part deals with methods of applying these elements to practical eq design.**

ACHIEVING bald response shapes of whatever nature—high pass, low pass or bell-shaped bandpass or notch—does not really constitute a usable eq system. The shape—even if variable in frequency and bandwidth—is either there or not, in or out, no subtleties or shades, and some means of achieving control over the strength of effect is vital. By far the most common (but certainly not the only) control requirement and one easily understood by operators is 'lift and cut', where the frequency areas relevant to the various filters are required to be boosted or attenuated by any variable amount within known limits. Determining these limits alone is good for an argument or two, dependant on such disparate considerations as system headroom, operator maturity (!) and obviously, application. An eq created specifically for horrific effects is not a subtle beast: 20dB of adjustment is not unknown (and not, unfortunately, unheard). 6dB, though, is often far more than enough, particularly in self-op on-air control suites. A general median accepted by most manufacturers is to provide between ±12 and ±15dB level adjustment on channel-type eq's.

## The Baxandall

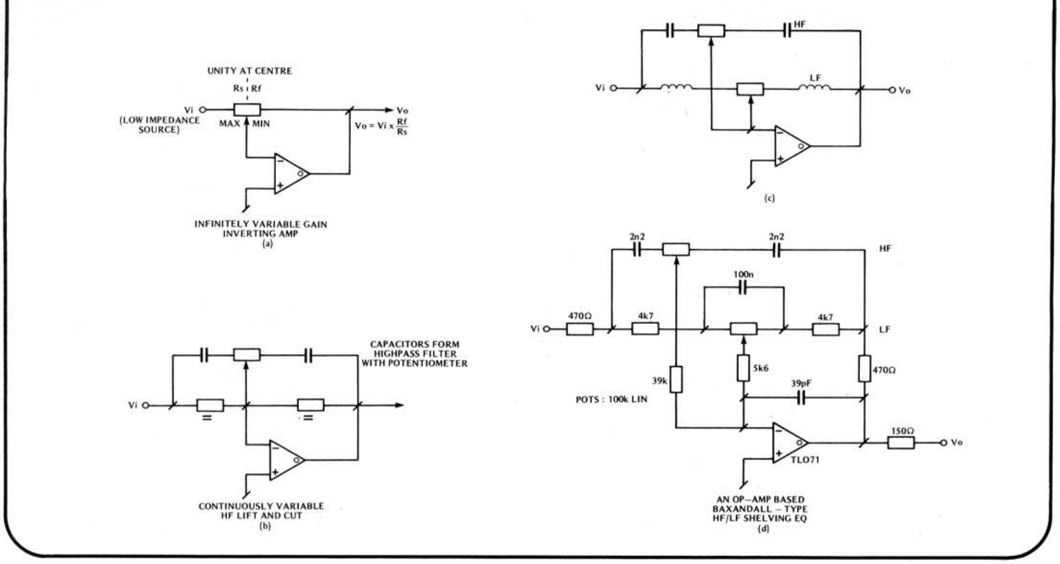
Hi-fi type tone controls needed similar basic operational hf/lf lift and cut facilities and a design for this dating from the 50's by Peter Baxandall has since been an industry standard in assorted and updated forms. A development of the Baxandall idea is represented in **Fig 45** based rather around today's more familiar op-amp technology rather than discrete transistors or valves. **Fig 45a** shows a 'virtual-earth' type inverting amplifier with the gain (being equal to the ratio of the feedback resistor Rf to the series resistor Rs) continuously variable from infinite loss (min) to infinite gain (max) with unity sat in the middle. If a fixed gain determining leg is introduced and the variable leg made frequency conscious **(Fig 45b)**, in this instance by crude single-order high pass filters (the series capacitors), the gain swing only occurs within the passband of those filters. The through gain for the rest of the spectrum is determined by the two fixed resistors—if this fixed chain is replaced by a second frequency-conscious network that does not significantly overlap the original one in bandwidth, the two chains independently modify their frequency areas **(Fig 45c)**. The fixed chain is only necessary where the gain is otherwise unpredictably defined by a frequency-conscious network.

The belt - and - braces low - pass arrangement of **Fig 45c** can be simplified into the more elegant **Fig 45d**, which more closely resembles the definitive Baxandall circuit. Rather than isolating the lf lift/cut chain with increasing inductive reactance, the control is buffered away with relatively small resistances and bypassed to high frequencies by capacitance. The control takes progressively greater effect at lower frequencies, as the rising capacitative reactance reduces the effective bypass. A further refinement is a pair of stopper resistors, small in value, that define the maximum lift and cut of the entire network.

Naturally, more complex eq can be configured around the same arrangement. A mid-frequency bell curve is easily introduced by any of the means in **Fig 46c**, giving a good clue how to avoid having to use a 'real' tuned circuit using dreadful inductors.

A variable signal either positive or

FIG 45 DEVELOPMENT OF BAXANDALL STYLE EQ



negative in phase sense to the source Vi can be picked off from a pot straight across the existing hf and lf chains, taken to an active-filter arrangement to derive the needed amplitude response shape, then returned into the loop at either the virtual-earth point (to which the hf and lf chains are tied) or to the non-inverting reference input **(Fig 46d)** dependent on whether the absolute phase of the filter is positive or negative respectively. Industry favourites seem to be this approach using either a Wein-Bridge bandpass **(Fig 42e** in part eight was evolved specifically to such an end) or a state variable type (for better or worse) as in **Fig 43e.**

Any number of such active chains may be introduced, provided two Great Hangups don't intrude excessively.

Hangup 1: Interaction between frequency groups. Hanging on two control chains that operate at the same frequency either adjustably or through overlap can at best be deceiving or at worst self-defeating. In the Baxandall (as with most other arrangements as we shall see) if maximum gain (say 15dB) is attained at a given frequency by one control, a second similarly tuned chain, cranked for maximum, will not give the expected additional 15dB gain—the overall loop is already operating close to the maximum gain defined by the stopper resistors. A notable measured result is for a sweep-mid bell curves' maximum lift and cut capability to be restricted at the extents of its range where it overlaps into the shelving hf/lf curves. It's maybe fortunate that providing the overlap isn't really silly; this interaction is not subjectively obvious.

A rough rule born from hard experience of squeezing most eq from least electronics is to not allow overlap incursion beyond the point where either curve has ± 6dB eq effect individually. Overlapping is best achieved from the comfort of another eq stage, although that too invokes other compromises.

Hangup 2: Noise. The basic Baxandall, using purely passive frequency-determining components, is a delightfully quiet arrangement. With controls at flat, it is theoretically only 6dB noisier than the unity gain noise of the amplifier —probably in the –100dBu region. Noise character varies with the controls much as one would expect of an amplifier the gain of which is directly manipulated at the frequencies in question—hf lift, more hf noise etc.

As soon as active filtering is involved, more noise is unavoidably introduced, often highly coloured and consequentially more noticeable. What is worse is that it's present all the time irrespective of control positions—even with its appropriate control at neutral centre, it is quite usual to hear a mid-sweep 'swoosh' in the noise changing with filter frequency. This is along with the strange spectral character of the noise emergent from some filters, notably the integrator-loop variety, resultant from unoptimised impedances and dubious stability almost inherent to the design.

## Alternatives

The source impedance versus feedback impedance ratiometric approach of the Baxandall is not the only way of achieving symmetrical lift and cut. A method enclosing the controls within the feedback legs of a non-inventing amplifier is developed in **Fig 47.** This has the advantage of leaving the op-amp's non-inverting input free, obviating the need for a preceding low-impedance source or buffer amplifier. Roundabout to

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FIG 46 METHODS OF INTRODUCING RESONANT FREQUENCY SELECTIVE ELEMENTS TO THE BAXANDALL

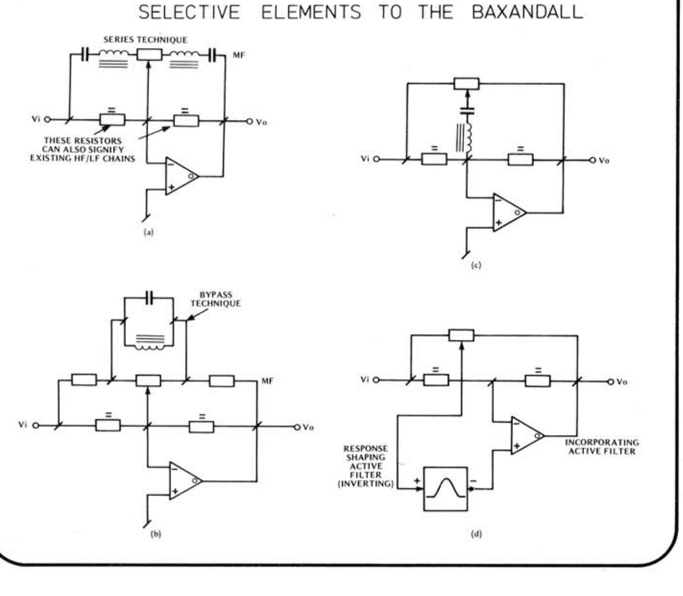
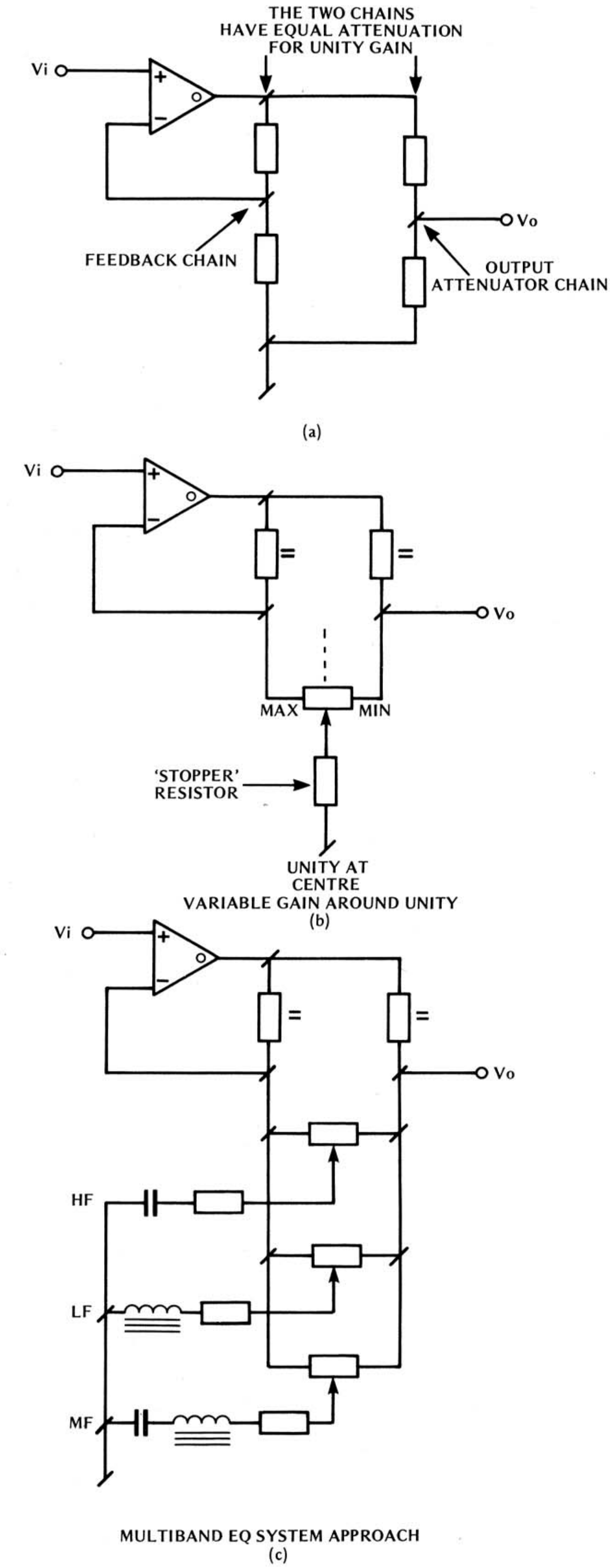



FIG 47 SWINGING OUTPUT EQ GAIN BLOCK



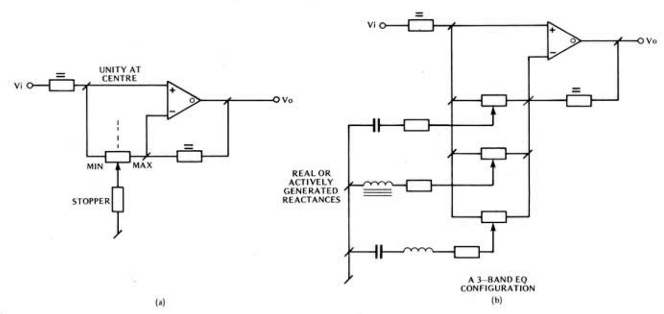
# Mixing console

this swing is the necessity of a buffer amp or quite high destination load impedance since the output is variable in impedance and included in the op-amp's feedback loop—heavy control modification, potential phase margin erosion with consequent instability, and certain headroom loss would be among the penalties.

Unity gain in **Fig 47a** is achieved when the attenuation in the feedback chain equals the output attenuation; the feedback attenuator causing the op-amp to have as much voltage gain as the output attenuator loses. Replacing the two bottom legs of the attenuators with a swinging potentiometer **(Fig 47b)** provides a lift/cut facility: when the pot is swung toward 'min', the feedback leg is effectively lengthened to ground and the amplifier gain consequentially reduced somewhat. Meanwhile the output attenuator is shortened considerably, reducing the output accordingly. At 'max', surprise surprise, the reverse occurs —the feedback leg is shortened, increasing the op-amp's loop gain whilst the output attenuator is lengthened, losing less of the available output. A small 'stopper' resistor defines the overall gain swing about unity which would otherwise range from zero to ear-plugs respectively.

Introducing reactances and/or complex impedances into the potentiometer ground leg (or legs as in **Fig 47c**) results again in lift/cut control over the frequency bands in which the reactances are lowest, ie hf for capacitors, lf for inductors (real or fake) etc. This arrangement—which has been spotted in a few odd places professionally, and in some Japanese hi-fi, has only one major drawback other than the previously mentioned output loading considerations. In order to achieve reasonable control dB-per-rotation linearity, the two attenuators (feedback and output) need to be of about 3dB loss each with the control at centre. This implies that the obtainable output voltage is 3dB below the output swing capability of the op-amp, meaning a headroom deficit of that amount in the equaliser stage, probably where it is most needed. Bad news!

FIG 48 'SWINGING INPUTS' GAIN BLOCK




Avoiding the headroom headache but utilising a rather similar technique, the 'swinging inputs' gain block of **Fig 48** is very promising. In fact, very forthcoming.

Here, the feedback attenuator remains unchanged but the output attenuator is shifted around to the op-amp's non-inverting *input*. At 'min', the input attenuation is quite vicious whilst the feedback leg is long, making the op-amp deliver only a small amount of gain. When the attenuation characteristics are reversed for 'max', the op-amp works at a high loop gain whilst the input is only slightly attenuated—unity is achieved at control centre where the input attenuation equals the amplifier's make-up gain.

There is a fascinating trade-off between noise mechanisms in this circuit arrangement. Assuming a maximum of three controls (for fairly standard hf, lf, and mid-sweep curves) before interaction becomes a major hassle, the amplifier can have

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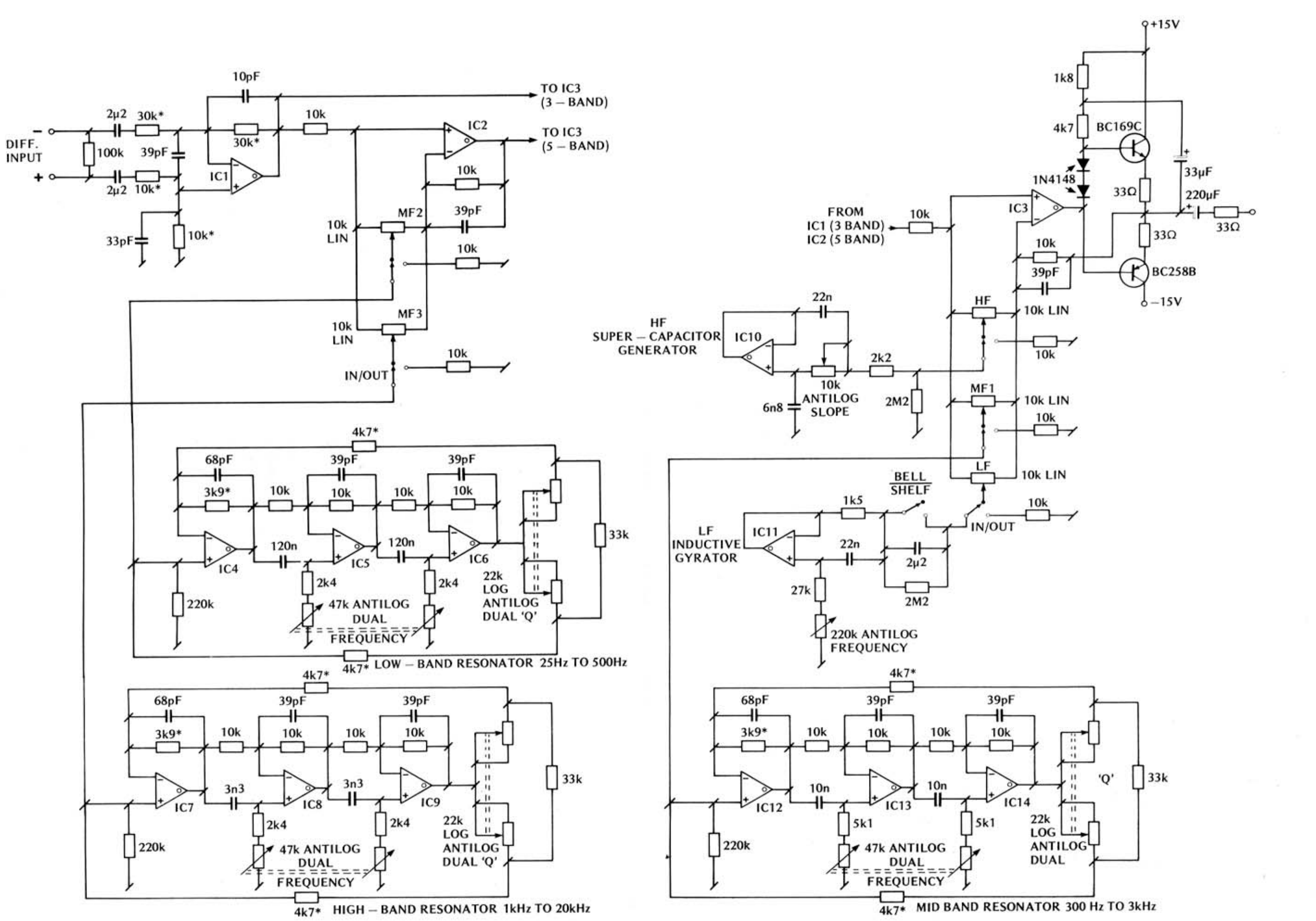
FIG 49 5-BAND EQ CIRCUIT DIAGRAM



between 10dB and 20dB of fairly frequency-conscious background gain (ie with all controls 'flat') rendering it, at first sight, significantly noisier than a Baxandall. Two big BUTS.

*But,* the impedances around the amplifier are considerably lower—in the order of a *decade* lower—reducing thermal noise generation due to resistive elements and op-amp internal mechanisms considerably.

*But,* noise generated by the active frequency determining filters is, with the controls neutral, injected equally into the inverting and non-inverting inputs of the op-amp. Differential amplifiers being what they are, common-mode signals (such as this equally injected filter noise) get cancelled out and do not appear at the output. Fun, this, isn't it?

Interaction can still intrude, care being required to prevent excessive frequency band overlap. Centre-tapped pots (the tap being grounded) eliminate many interactive effects but at the cost of increased invariable background gain (noise) and peculiar, almost intractable, lift/cut gain variation linearity versus control rotation.

## A practical eq

A three section parametric eq with additional versatile shelving-type hf and lf controls has its circuitry detailed in **Fig 49.** It is designed to be easily shortened to hf, lf plus a single mid-band parametric section for applications that don't demand the full—and by any standards fairly outrageous—complement of facilities. Each individual selection is switchable in or out to allow preset controls and simple in/out comparisons with tie-down resistors maintaining the unused filters' dc conditions to minimize switch-clicks. Even a brief gawp at the circuit reveals a major benefit—the signal path through the eq is via merely three op-amps, one (IC2) being an input differential amplifier and another (IC3) does duty as the output line-amp. In the shortened version this path is reduced to only two op-amps, IC1 and IC3 (which serves also as a swinging-input eq gain block). IC2 and its associated circuitry are unused. Most of the circuitry has been described elsewhere in this article or preceding ones so only fresh gruesome tit-bits will be explored now.

Perplexed by the strange-looking values around the differential input stage? Well, without lurching into tedious sums, those values provide unity differential in/unbalanced out levels whilst providing an identical impedance (with respect to ground) on each of the two input legs. Naturally, the more precise the component values, the better the common-mode rejection is likely to be.

## The first eq stage

IC2 is the first swinging-input stage which has two non-frequency-overlapping filters hanging off it, one section covering 25Hz to 500Hz: the other, 1kHz to 20kHz. Each filter network creates a complex impedance form against frequency that is a dead ringer (get it?) for a series L/C tuned circuit to ground. This fake tuned circuit (formed from two constant-amplitude phase-shift networks in a loop, named the CAPS-variable filter) reaches parameters proper filters cannot reach.

The centre frequency is smoothly variable, 'Q' remaining constant over the entire swing. The 'Q' itself is continuously variable between around 0.75 and 5 (very broad to pretty sharp, representing bandwidths of 1.5 to 0.2 octaves respectively). Positive feedback inside the loop (which defines the 'Q') is balanced against negative feedback (which controls minimum filter impedance, hence amplitude) interestingly enough relying on the input impedance of the 'swinging-input' stage as part of the negative feedback attenuator. Fortunately this is reasonably constant irrespective of lift/cut positioning.

In the absence of complementary square-law/reverse square-law dual gang potentiometers ideally required for the purpose, readily available log/antilog dual-gang pots slugged a bit to a reasonable approximation control the positive/negative feedback balance. As a result of this compromise, the crest amplitude due to the filter varies within ± 1dB as the 'Q' control is swept—however, in comparison to the dramatic sonic difference from such a 'Q' variation, this tends to insignificance.

The result of all this, at the output of IC2, is a pair of somewhat beauteous resonant-type curves of continuously variable place, height, depth and width.

## Second eq/line amp

A reasonably hefty pair of transistors are hung on the end of IC3 to provide a respectable line-drive capability, in addition to the amplifier's use as a 'swinging-input' eq section. There is by far and away enough open-loop gain in the op-amp/transistors combination (over a much greater bandwidth than mere audio) to cope with 15-odd dB of eq lift and output stage discontinuities.

Differing from the last eq stage, this one only has a single mid-frequency bell curve creator, operating over a range of 300Hz to 3kHz, together with deceptively simple-looking but fascinatingly behaving hf and lf impedance generators.

Gyrating inductance to create a conventional lf shelving response (variable in turnover frequency by a 220K antilog pot) is achieved around IC11. A fairly large (2.2$\mu$F) series capacitor forming a resonance is switchable in and out, the capacitor's value being carefully calculated to work with the circuit impedances to provide an extreme lf response that falls back to unity gain below the resultant resonant frequency. The matched capacitor value also ensures virtually unnoticeable disturbance to the curve above the resonant frequency—reliance is placed on the characteristics of such a capacitor/variable gyrated inductance network: the 'Q' reduces proportionally to increasing frequency. Typical resultant response curves (**Fig 50**) show just what all this means, demonstrating an extraordinarily useful bottom end control.

Unusual—that's one way to describe the hf impedance generator and its eq effect. It is essentially a 'super-capacitor', or 'capacitative capacitor', and if both of those are meaningless, it's a circuit that when in conjunction with a resistor causes a second-order response as would normally be expected of an inductor/capacitor combination—a slope of 12dB/octave as opposed to a single-order effect of 6dB/octave. It utilizes negative-impedance conversion—if this is sounding a bit weird and sci-fi, don't worry it works and **Fig 51** shows what it does at the eq output.

The response is 'hinged' about 1kHz, the control varying the frequency (between 5kHz and 20kHz) at which the gain reaches maximum (or minimum if the lift/cut control is 'cut'). The slope between 1kHz and the chosen maximum frequency is virtually a straight line representing a nearly constant dB/octave characteristic, with a nearly flat-top shelving characteristic.

In electronic terms, this is achieved by progressively disorganising the super capacitor until it gives in, eventually looking like a simple, single capacitor . . .

Unusual. ■

FIG 50 FREQUENCY RESPONSE OF LF SECTION (CONTROL AT MAX GAIN)

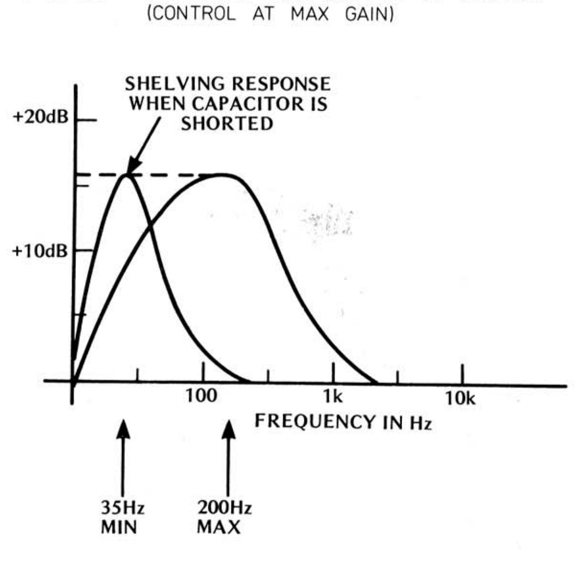
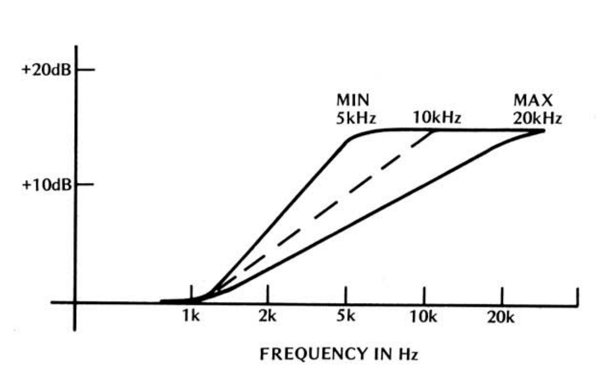



FIG 51 CHARACTERISTICS OF HF SECTION (LIFT CUT CONTROL AT MAX LIFT)



# Designing a professional mixing console

Steve Dove

# Part Ten~Monitoring

**Monitoring—the desk-related function, not the big nasty loud things nailed on the wall—is usually the last sub-system to be considered in console design. After all the programme paths have been established, zipping about everywhere and jumping through hoops, a suspicion dimly glimmers that it might possibly be useful to listen to them occasionally.**

IN THE context of recording systems this 'lapse' is definitely more common than broadcast where the monitor selection and routing often exceeds the programme path in complexity and functions. (If you get a chance, check out the monitor section of a BBC radio continuity desk and *think* about what is needed to achieve everything.)

Monitor system design, like much else, falls into two phases; deciding what it needs to do and the implementation.

At its simplest, monitoring consists of a power amp and speakers hung across the mixer's main output(s), auxiliary functions being either unused or preset—in PA work the PA actually *is* the monitoring, the only other function necessary being PFL and then only during 'panic mode'. Alternatively the monitoring demands for multitrack recording extend to effectively an entire secondary submixer replete with panning and pre/post foldback effect feeds, stand-alone soloing together with listen access to all desk send and return ports. The in-line console principle makes efficient use of electronics to combine often coincident signal path and monitoring requirements for the multitrack machinery, to the extent that it is operationally rare to listen to anything other than the stereo buss output. This serves as both the multitrack monitoring buss and the stereo mixdown buss.

Three distinct types of monitoring activity evolve in multitrack work:

(a) Mainline—the stereo buss, which encompasses the multitrack machine sources/returns and stereo mixdown.
(b) Transient—short-term check listening of individual channels for reassurance or adjustment, using prefade listen or solo functions.
(c) Auxiliary—access to the assorted foldback/effect feeds, effect returns, mastering machine and subsidiary 2-track and cassette machine returns.

From an operating point of view, that seems to be the division—from a technical stance it's a different matter entirely. The solo function is very closely related to the stereo buss—in fact using exactly the same single path throughout—and can be seen simply as a modification of it. PFL, though, despite a similar seeming operation (only prefade as opposed to postpan listening) actually requires an entirely separate buss and mixing system, its output switched to override the main path into the monitors. (It may seem a bit strange to go through all this palaver for a spot-check function that tells you less than the stereo in-place solo, until it's remembered that a solo disrupts the mix whilst a PFL is non-destructive.) System complexity disguised under operational transparency. Conversely, an operator usually has a psychological hook about the main stereo buss monitoring being the Great Holy Unblemished Gospel signal path, all the auxiliary functions being somehow tarnished and 'unclean'. Wrong.

In reality, the monitoring chain selects directly between all its sources, merely treating the stereo mix as one of the many.

An assumption is made that the 'solo' function is as outlined in Part One (September 1980) where, simply, if a console channel is 'soloed' all other sources contributing to the main stereo buss are muted, leaving the desired channel in isolation at its set level and panned position. An exception and extension to this is for other channels (principally those returning effects to which our soloed channel may be contributing) to remain unmuted in the stereo 'mix'.

The upshot of this is that 'solo' monitoring is inherent to the stereo mix path—if that path isn't selected to monitoring then neither is the solo. So although a solo overrides the main stereo mix (unless disabled altogether by the 'solo safe') it cannot override anything else, unlike the prefade listen.

PFL, although it could just be brought off as another monitored source, is made to simulate the 'solo' in single button touch operation, with the added advantageous capability of overriding everything.

This then gives us a logical priority

FIG. 52 BLOCK DIAGRAM MONITOR SELECT CONTROL LOGIC

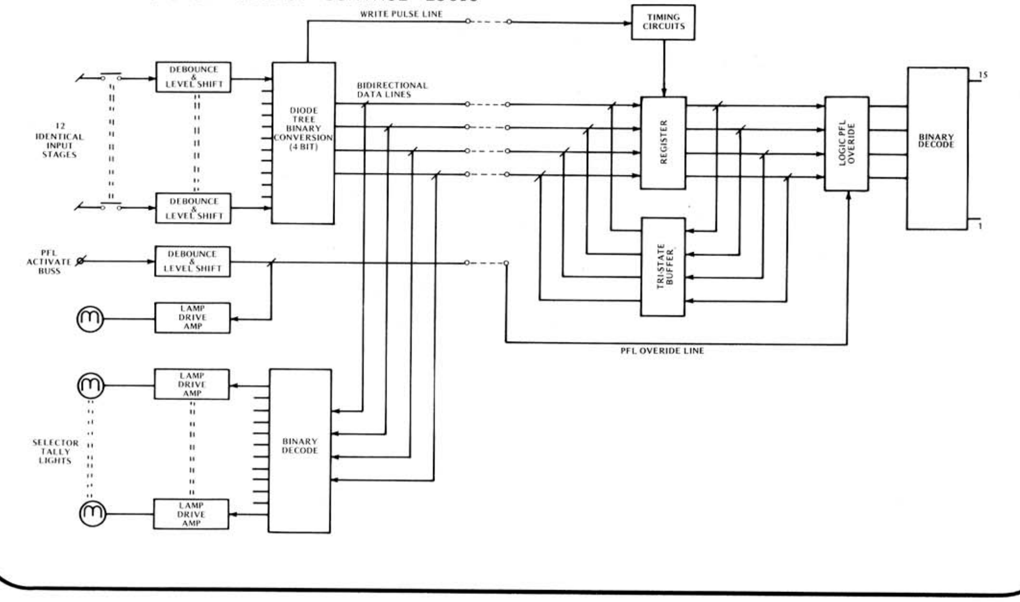
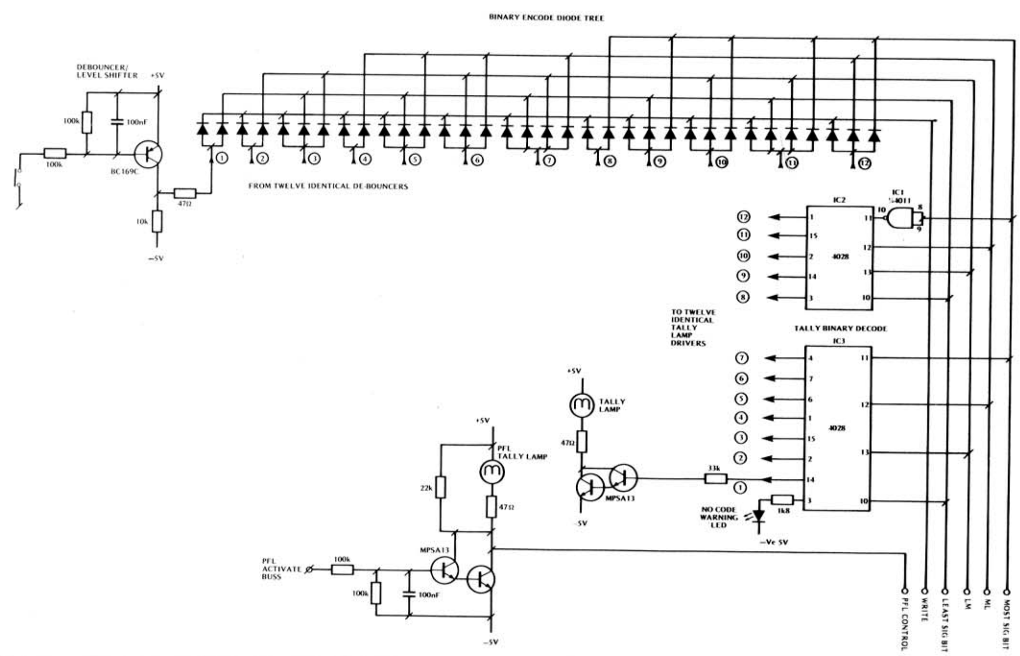



FIG. 53 MONITOR SELECT CONTROL LOGIC – DESK END




on which to base the monitoring routing system.

## Controls

OK. So now we've worked out how to get what signal and at what priority, into the monitoring chain—once it's there, what other torture do we put it through?

- Level control: volume to you, mate.
- Mute: so that you can turn the row off.
- Dim: so you can hear what people say.
- Mono: people still use it, you know.
- Phase Reverse: to make sure you haven't already done it inadvertently. (This function with the Mono makes for one of the quickest ways in history of lining up machine azimuth.)
- Split: (Huh?) a cunning frolic unashamedly pinched from broadcast monitoring technology. This routes a mono sum of the main stereo mix buss continually to the 'left' side of the monitor chain and a mono sum of whatever source is selected (including PFL override) to the 'right' side, providing simultaneous monitoring of two different sources—one of which would almost certainly be desk output anyway. Why?

Well, its origins lie in network radio, where an announcer on air has to talk up to a programme junction and smoothly hand over to another studio/network feed/ 'Independent Radio News'/whatever, at a cue. In order to do this, he has to be able to hear both himself and the network he is opting into to hear the lead-up and handover cue.

Other than its primary design use, the 'split' function is used considerably under other normal programming, affording random source monitoring without losing track of what the main desk output is doing. It's also used extensively in programme pre-recording and production enabling, with practise, real-time multisource edits without recourse to razor blades and that dreadful tape that curls up and sticks under your fingernails. A technique (who remembers this?) very reminiscent of jump-editing on discs.

This author is convinced that 'split' will find a niche in multitrack recording techniques. If nothing else, it will fulfil the requirement for single speaker mono monitoring, by simply selecting the 'right' side to a dead source.

- Idiot speakers: those nasty things gaffer-taped on the meter penthouse to do Dansette record player and Radio One impersonations, also affording a respite from the deafness-inducing 'normal' monitor speakers.

## Crosstalk

In a programme sense, two forms of crosstalk are relevant. The first, *related* crosstalk is a signal bleeding over into another signal path which is carrying a musically and temporally related signal, eg between the left and right of a stereo pair or between adjacent tracks of a multitrack. It happens and is fortunately not often subjectively obvious or embarrassing.

Crosstalk within multitrack recording systems is usually little short of horrifying. As a result of the large physical size of the console, ground paths are unavoidably long and ground currents generate (and cross-inject into other paths) crosstalk voltages across the resultant ground impedances. Capacitance between interconnecting cabling, looms, modules, busses, everything result in a reasonably dreadful overall crosstalk performance.

This is only mitigated by multitrack machine crosstalk between tracks—a safe order of magnitude worse than even a horrid desk ever could be. It is all tolerable and usable simply because all the crosstalk is related and blends in unnoticeably. The only area where this is not necessarily so is in monitoring, where a 'hostile' signal (say a delayed replay 'B' check of a master) can be screaming about in uncomfortable proximity to the main stereo mix paths.

Broadcasters face this problem all the time—all their sources are hostile (!) unless brought up on air.

This is *unrelated* crosstalk where the bleeding signal is totally dissimilar and irrelevant to the interfered signal. Basically, if any unrelated crosstalk is audible above system background noise, it will be noticed.

A fairly recent and insidious sort of unrelated crosstalk comes in the form of assorted chirps, buzzes and sizzles stemming from the relentless march of digits into mixer designs. SMPTE timecode and automation

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# Mixing console

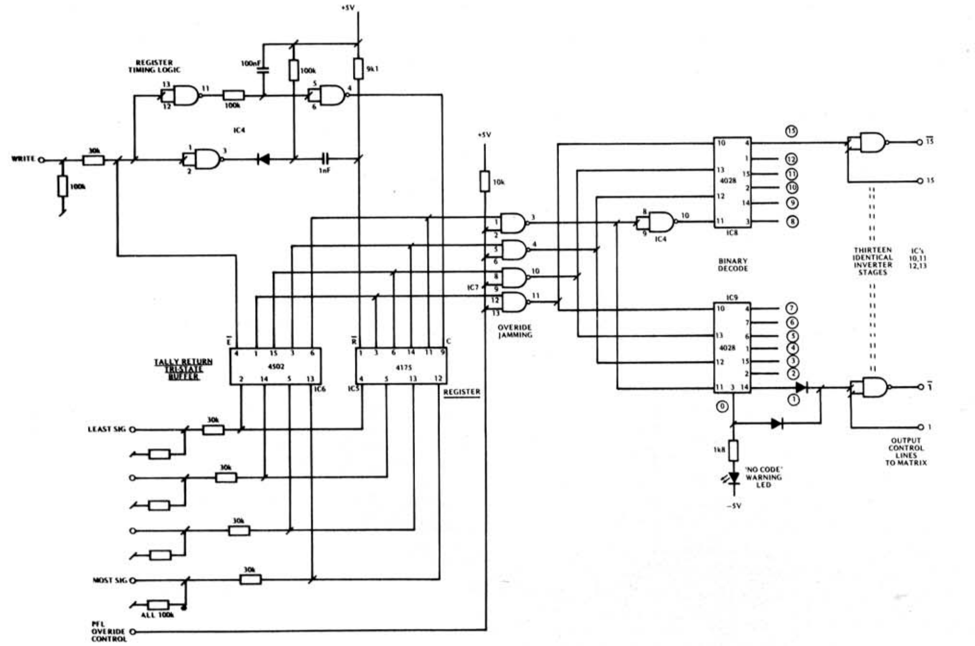
codes were bad enough, but trying to get computer clock droning out of mixing busses is not one of life's fun tasks.

A very reasonable quantitative measurement technique of all such effects is specified in the IBA local radio Code of Practice.

Originally the test for interchannel crosstalk (*ie* between any channels in a desk), it's also used for any dissimilar path crosstalk measurements. In short, it asks for better than 60dB of isolation at 6kHz between the paths, measured with a standard PPM with a CCIR 468 weighting filter in line. Since this CCIR curve has 12dB of gain at its crest (at 6kHz, surprise) the specification is actually calling for better than 72dB of isolation at 6kHz—not easy, and very realistic. Such a figure is not far above system noise floors, generally. Remember, it's a peak measurement.

This harping on about crosstalk is not without a point, as it is actually concerned with the physical construction of the monitoring switches.

FIG. 54 MONITOR SELECT CONTROL LOGIC— RACK END




## The switcher

A technique used to minimise crosstalk across broadcast monitoring and particularly outside source selection (whole hordes of hostile sources) is employed here.

The selecting switcher (based on electronics for outside source selection matrices) is contained in a rack mounting box near the jackfield, the routing controlled by digital logic level lines from the desk. In this way only a single pair of hostile signals is returned into the desk—far easier to engineer away from things that may be unduly influenced. All the required sources, including PFL, main stereo output, auxiliaries, 2-track machine returns, etc, are accessible on the field as a matter of course—short jumping links to the terminations on the switcher replace all the messy hassle of getting dozens of bits of signal via motherboard or hard wired connectors into the back of a conventional monitoring module. Ugh!

The monitor control electronics (mono, split, volume, etc) are conventionally located and constructed in the now mercifully un-hectic monitoring module.

## The data link

**(Figs 52, 53 and 54)**

Communication between the console controls and the rack is via a 6-line data buss—four lines of which form a one-in-16 binary code to select the lucky source, the other two being control bits. The write command line goes high immediately a different routing is selected, enabling the

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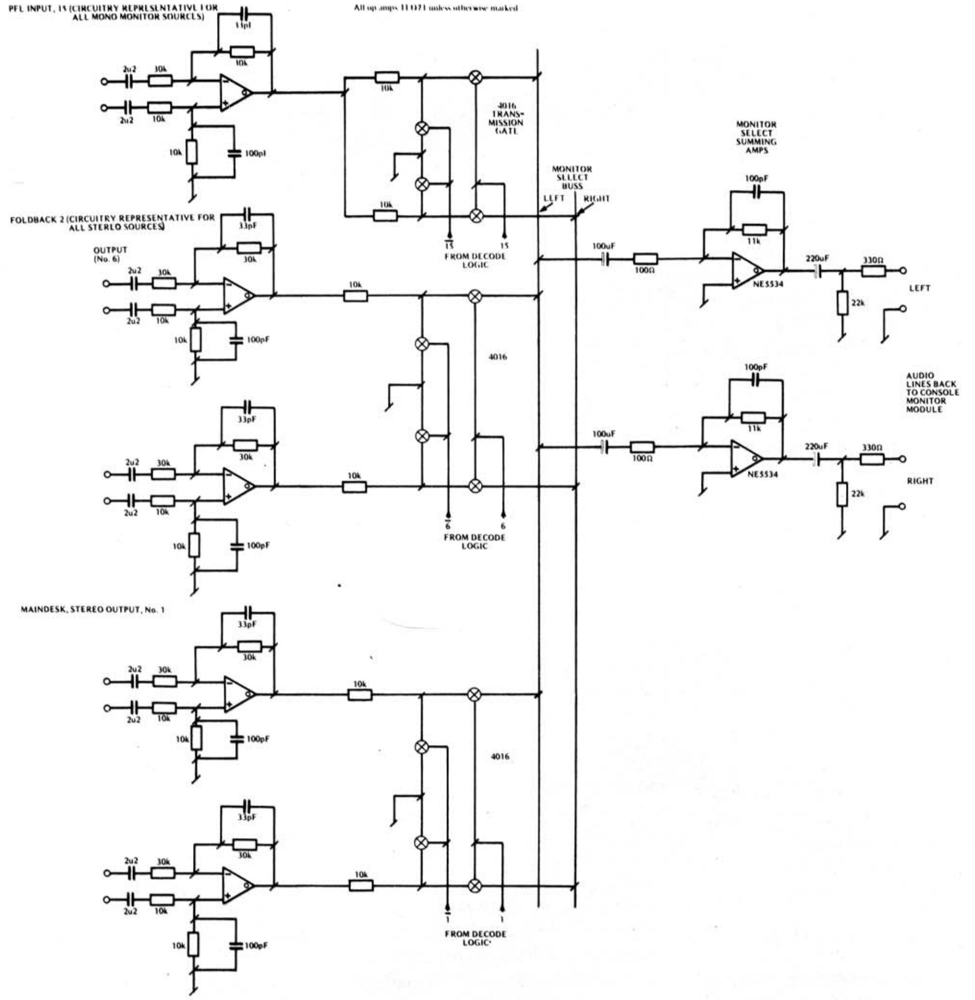
FIG. 55 MONITOR SELECT SWITCHER



# Mixing console

4175 register (IC 5) at the rack end to swallow whatever code is set up on the 4-bit code buss by the diode tree. Regardless of any other monitoring condition, if the PFL activate buss is grounded, the PFL override line drops low (it is ordinarily tied high) preventing the stored code in the register reaching ICs 8 and 9, the 4028 binary decoders. Instead, they see all data lines high and decode that as source 15—where the PFL (audio) is brought up on the matrix. This code 'jamming' is relieved when the PFL activate line is released, so the matrix reroutes back to the code stored in the register.

No apologies are offered for the somewhat 'agricultural' diode tree approach to binary encoding. It's simple, fairly inexpensive and bomb-proof. Similarly, why use a one-shot timing chip for the register timing logic when two resistors and a capacitor will do?

A nicety is that use of the code buss is made to send a tally code back from the rack end to the console end, at all times other than the instants when re-routing is occurring. This is achieved by the 4502 (IC ring. This is achieved by the 4502 (IC 6) tri-state inverter/buffer which stuffs the register output back up the lines to be decoded by a pair of 4028s (ICs 2 and 3) which then, via Darlington transistors, drive the lights. Thus, there is a readout of what the switcher is *actually* doing, not what it *ought* to be doing. The tri-state is disabled whenever the write line goes high, which of course signifies another bunch of digits steaming up the code buss the other way from the diode tree.

Sixteen selections are possible (0 to 15), but as shown below, only 14 are used. Code 15 is dedicated to PFL, Code 1 to main stereo mix. A default mode exists, whereby if the code buss logic takes a walk or becomes disconnected and no codes are being generated, the busses will almost certainly rest all low—code zero. Zero, when decoded by ICs 8 and 9, brings up warning LEDs on the monitor modules and rack front panels, whilst also pulling on source one (main stereo mix) through the switcher. At least it won't all go quiet on you.

## Audio path

The 12 normal routing selections are detailed in the following table. Some of these sources are stereo, some mono. **Fig 55** shows the only difference as being a mono source only needing one input amp, the output of which is split through the two analogue transmission gates into the stereo monitor select mix buss. Operation of this type of switching is detailed in Part Five (January 1981 issue).

**ROUTING SELECTION CODES**

| | |
|---|---|
| 0 | 'No Code' warnings, defaults to source 1 |
| 1 | Main desk stereo monitor/mix |
| 2 | Stereo mastering machine return |
| 3 | Two-track machine return |
| 4 | Stereo cassette machine return |
| 5 | Foldback 1 (stereo) |
| 6 | Foldback 2 (stereo) |
| 7 | Effect send 1 |
| 8 | Effect send 2 |
| 9 | Effect send 3 |
| 10 | Effect send 4 |
| 11 | Spare, access on field |
| 12 | Spare, access on field |
| 15 | Pre-fade listen desk output |

The monitor summing amp outputs are taken directly back from the rack back into the console monitoring module via a pair of ground-free differential input amplifiers and hence into the monitoring chain.

An interesting paradox—which fortunately is of no real effect in any of the intra-control area routing for which they are utilised in these designs—dealing with the input impedances of differential input amplifiers is worth mentioning here. With the simple one op-amp circuit (as at the front end of the matrix in **Fig 55**) it is possible to arrange the input impedance of the two legs to be equal for a differential (normal) signal *or* equal for common-mode (interfering) signals—but not both simultaneously. In other words, it doesn't work! In the former case, interfering signals are likely to be induced at dissimilar levels into the two legs causing a differential and hence transferred resultant of the common-mode signal. If, though, it's optimised for common-mode input impedance matching, the inescapable differential input impedance unbalance could cause a balanced *source* connected to it to have a worsened common-mode rejection ratio because of that intolerance—boring, eh!

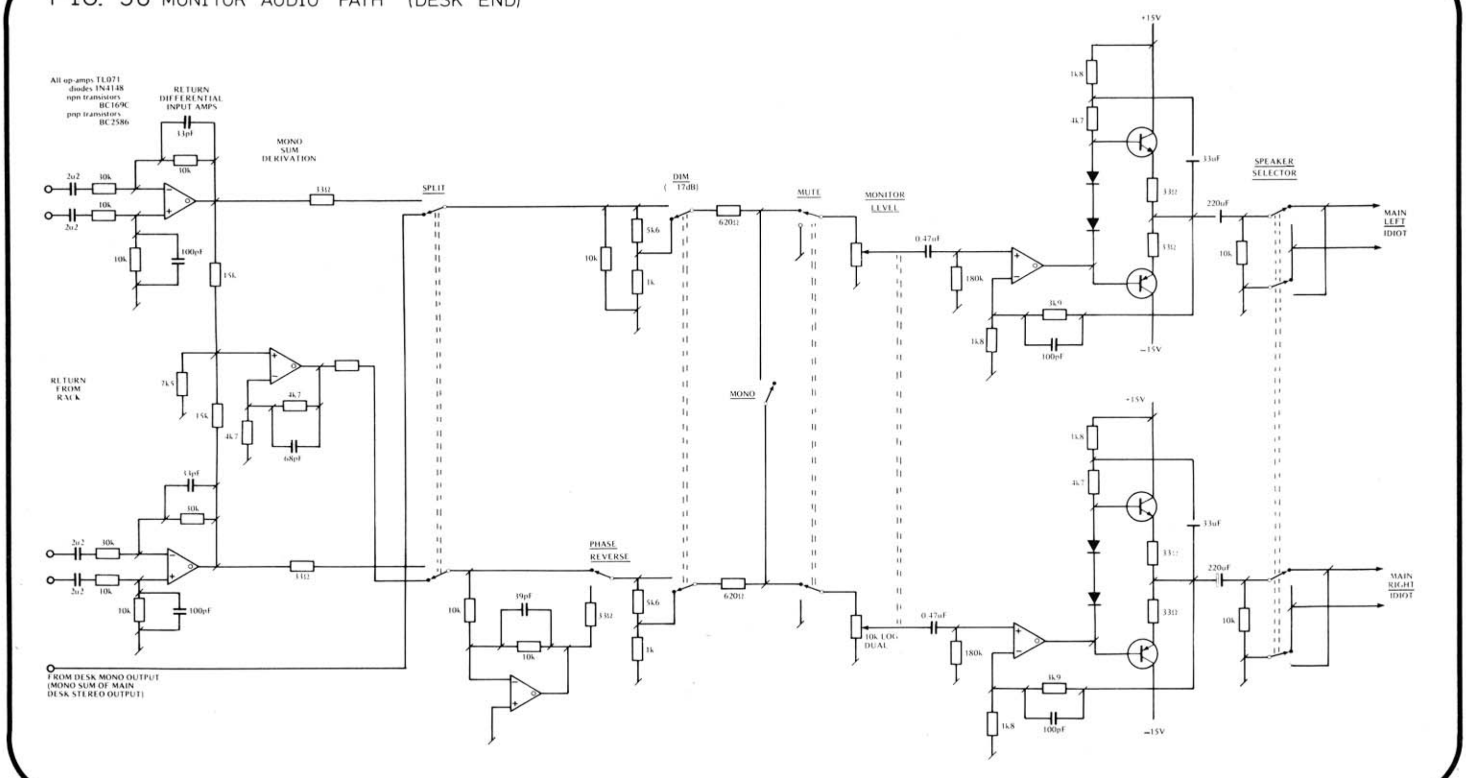
There are means, well documented elsewhere, by the use of active feedback paths or multiple op-amp 'instrumentation amp' type circuits, of avoiding these imbalance effects. Since this simple circuit is used solely to provide a measure of ground isolation from zero or very low impedance sources here, these effects are largely irrelevant to our purposes.

For the 'split' function, a mono sum is brought into the desk monitor chain of the main stereo mix. A mono sum is also derived of the switcher output immediately after the return differential input amps in the desk. Phase reverse is implemented by inserting a unity gain inverter into the 'right' path.

The rest of the monitor chain is self-explanatory **(Fig 56)**—dim attenuation may be varied simply by changing the value of the 1kΩ lower leg resistor.

Almost as a retaliation against the trend elsewhere in these designs to digital control, storage and remote capability, the monitor audio chain is unremarkably straightforward and conventional. Yes, it uses strange things called 'switches' as opposed to analogue transmission gates. This lapse, it is trusted, will be forgiven. ■

FIG. 56 MONITOR AUDIO PATH (DESK END)



# Designing a professional mixing console

Steve Dove

# Part Eleven ~ The Channel System

**Which would you rather have? An enormous jackfield and a BSc in knitting or a few cute little pushbuttons? This article describes the channel switching logic that reconfigures the signal paths for the various operating modes.**

A SYSTEM is a means of reducing the versatility of its component parts. Ideally, there should be no system but practicality dictates that there must be one. The thought is mortifying: hundreds of elements, the mic amps, diff input amps, line amps, equalisers, filters and routing matrices roaming loose and needing to be cobbled together for each individual operational requirement.

We need a saving grace and fortunately there is one. Engineering and balancing habits are pretty well entrenched giving rise to a few well defined, commonly used elemental combinations. Rationalising these combinations and arranging to be able to easily select them as necessary is a good compromise. We've not so much lost versatility as gained a family of operating modes.

The entire channel subsystem relies on the electronic switching elements used being entirely transparent—noiseless, distortionless, clickless and other impossibilities.

Noise due to the potentiometric CMOS switching employed here is very largely due to the individual summing amps, scaled by the gain asked of them. The impedances around these switches are low enough to fall somewhat below the optimum source impedance of the devices used. Noise resultant from them is defined to quite low (—100dBu or better) floor levels—fairly meaningless under the stampede of typical front-end or machine noise.

Distortion is primarily due to the CMOS transmission gates' automodulation, ie the path resistance varying with instantaneous signal voltage, but this at zero level is typically a nonsensical value. Both the harmonic and intermodulation products are almost unmeasurably low principally because of the near virtual ground operation of the active CMOS elements. No voltage swing, no automodulation.

The basic switching element, now, can be given the holy water treatment.

## Function modes

If possible, reference should be made to **Figs 1** and **2** from Part One of the series (September 1980) during this discussion of the channel system. These show the overall channel in block diagrammatic form and the various ways the circuit blocks are configured for the different functions expected of the channel in use. **Fig 1** has all the reconfiguration represented by diagrammatically accurate but forbiddingly incomprehensible mechanical switching. **Fig 57** replaces those in the main signal paths with

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FIG. 57(a) CHANNEL SYSTEM - SIGNAL PATHS

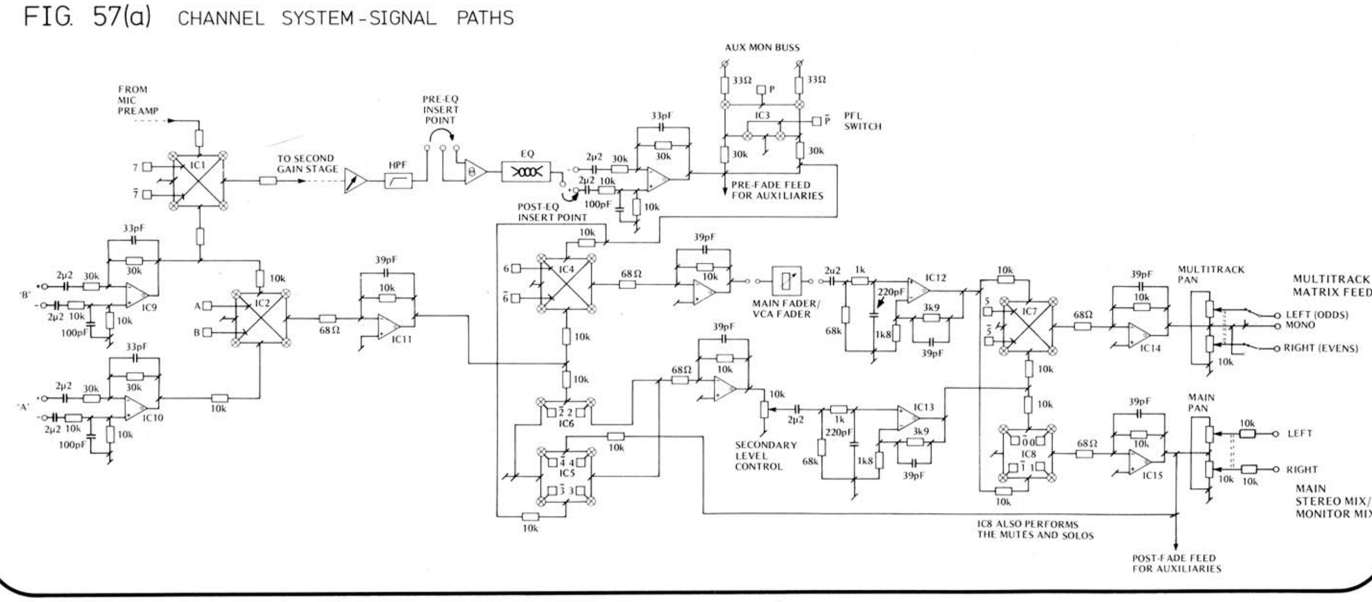
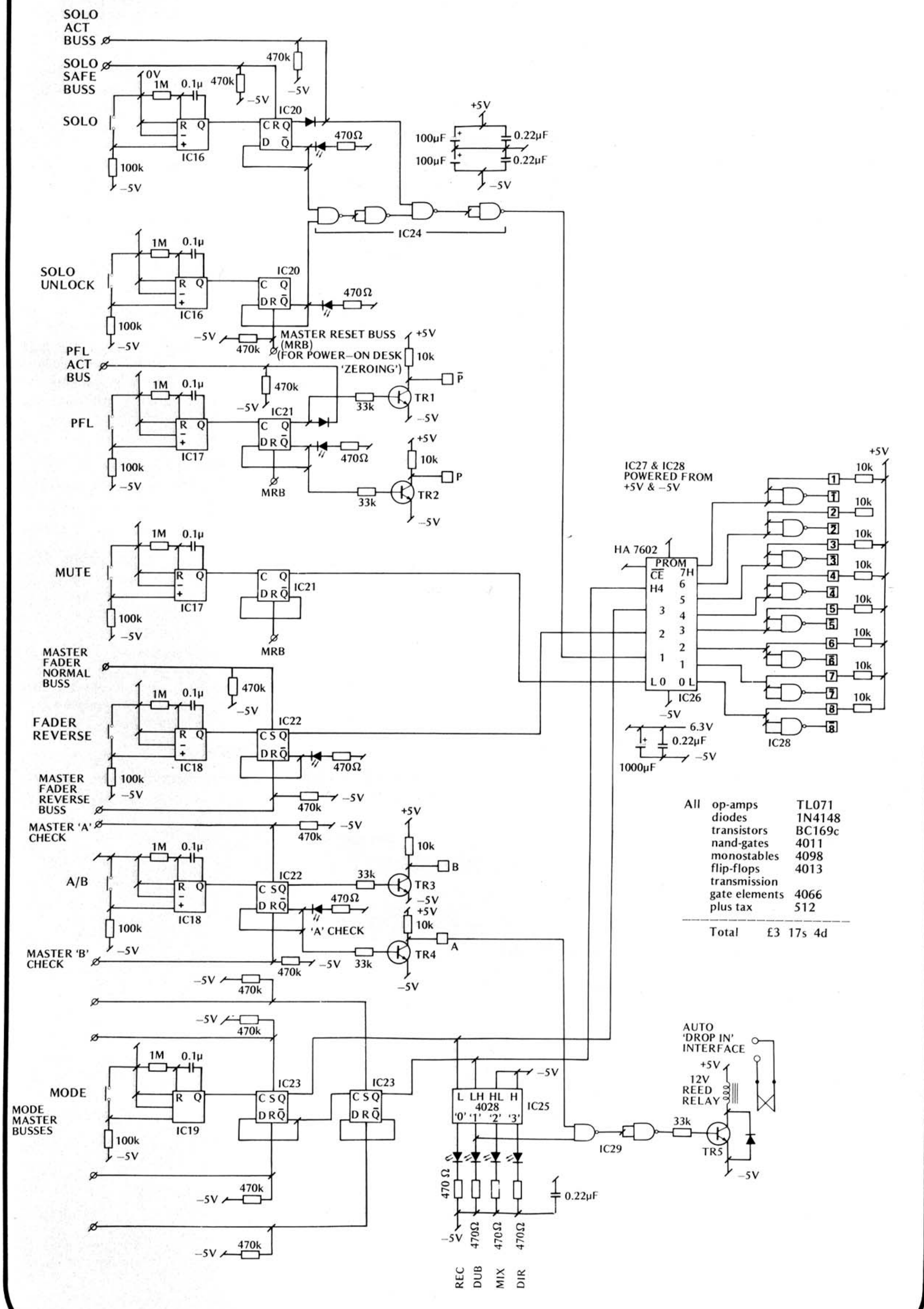


FIG. 57(b) CHANNEL SYSTEM – CONTROL LOGIC



electronic switching elements, which may seem more or less of a jungle dependent on whether you were brought up on hard-gold contacts or silicon.

Certainly there are fewer electronic switchpoints than there were mechanical. This rationalisation is primarily due to yet another incursion of esoteric (for audio) digital devices. It's osmotic—leave the wretched digital things lying around on the bench and they creep into circuits.

A simplified representation of the four basic channel operating modes is given in **Fig 58a** for recording, **Fig 58b** mixdown/direct to stereo, and **Fig 58c** overdubbing. The little 'x marks the spots' show the switching points.

As a brief resumé (Part One has the lowdown lower down) the main multitrack operating modes and their implementation in this system are outlined here.

## Recording (See Fig 58a)

Here the object is to get a 'live' source (eg mic) through the signal modification chain (ie limiting, equalisation) and on to a track or tracks of the multitrack machine. Level control on this path is by the main fader (or VCA fader if automation is applicable). Before and after monitoring of the tape track dedicated to the channel is routed onto the main stereo monitoring/mix buss via the secondary level control.

## Mixdown (See Fig 58b)

The machine return is brought through the modification chain and mixed onto the main stereo monitoring/mix buss via the main/VCA fader. The machine monitoring chain is disabled.

Those with sharp eyes or quick memories will notice a system design modification here, made as a result of user input since the original design's incarnation. **Fig 2b** (September 1980) shows the secondary level control feeding the multitrack routing independently of the main stereo mix via the main/VCA fader. Since a major justification for keeping the multitrack routing open during mixdown is to provide additional effects feeds, this would be far better served if the secondary level control is fed post main fader and post mute/solo switching. To enable this, a 'crossfeed' electronic routing is included **(fig 58b)**. However, independent control is restored when required if a 'fader reverse' is called.

Another mode, 'direct to stereo', is a derivative of 'Mixdown'. It enables live sources to be mixed straight on to the main buss obviating the need to use multitrack routing.

## Overdub (See Fig 58c)

A halfway house between 'record' and 'mixdown'. Intended for use when most of the desk is in mixdown but individual channels are being laid or touched up. Signal flow is as 'Record', only with the main/VCA and secondary level controls interchanged. The main/VCA fader in this mode therefore controls the monitor feed into the main stereo mix buss, which ties in with this fader's operation on all the other channels that are in 'Mixdown'.

A handy interlock exists in this mode to facilitate 'single button drop-in'. When the channel system function is selected to overdub and the monitoring path is set to 'A' check (machine input) a relay closing pair is made which may be plumbed into the machine's remote control access. Provided the track is 'armed' ready to record, hitting 'A' check automatically drops the machine in simultaneously. The increasingly prevalent use of machine synchronisers / timers / autolocators /coffee-grinders has dramatically eased multiple pass overdubs previously wearing on fingers and patience. To help it along a bit more, a control buss is run specifically to drop a channel in 'overdub' mode into 'A' check upon a given trigger from the aforementioned teasmade.

## Logic control

A separation is made in **Figs 57a** and **57b** between the analogue signal switches and their digital control electronics not purely because of the differing disciplines but for clarity's sake—lots of lines running all over the place.

Each top-panel switch is a momentary-action touch switch with an associated LED indicator (with the exception of the function mode switch—more later). The toggle push-on/push-off characteristic is provided by the basic debouncer/flip-flop circuit as in **Fig 59.** This action is not only fun and playworthy therefore fashionable, it scores in a couple of other important respects:

- cost, surprisingly. The combination of a small mechanically simple non-latching push-to-make switch and a fairly small amount of silicon bits has it nearly every time over latching pushbutton switches which are either downright klutzy, staggeringly expensive or slow-boat from Yokohama delivery; and
- versatility. Using electronic latching rather than mechanical catches makes remote/automatic function presetting and triggering a comparative doddle.

Debouncing is removing the ragged

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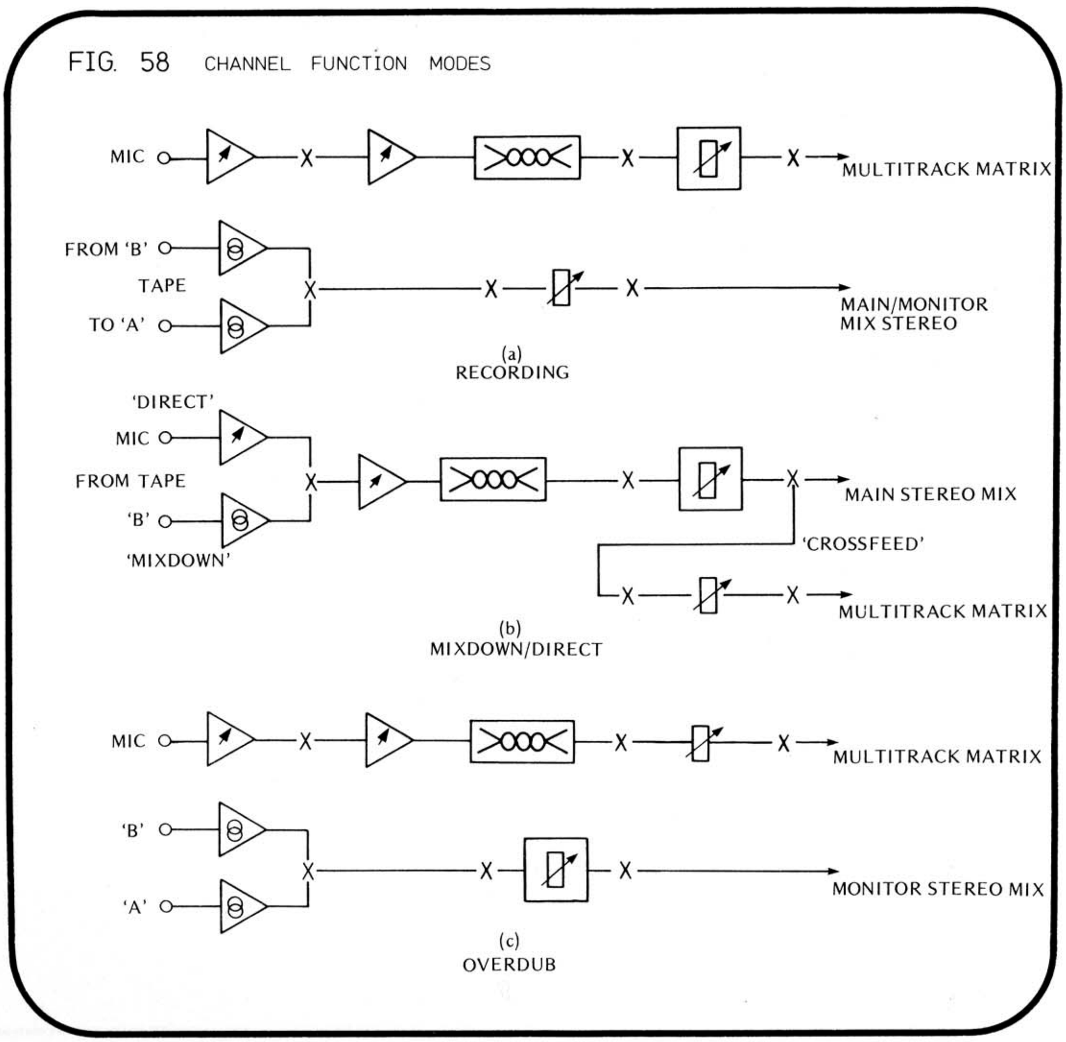
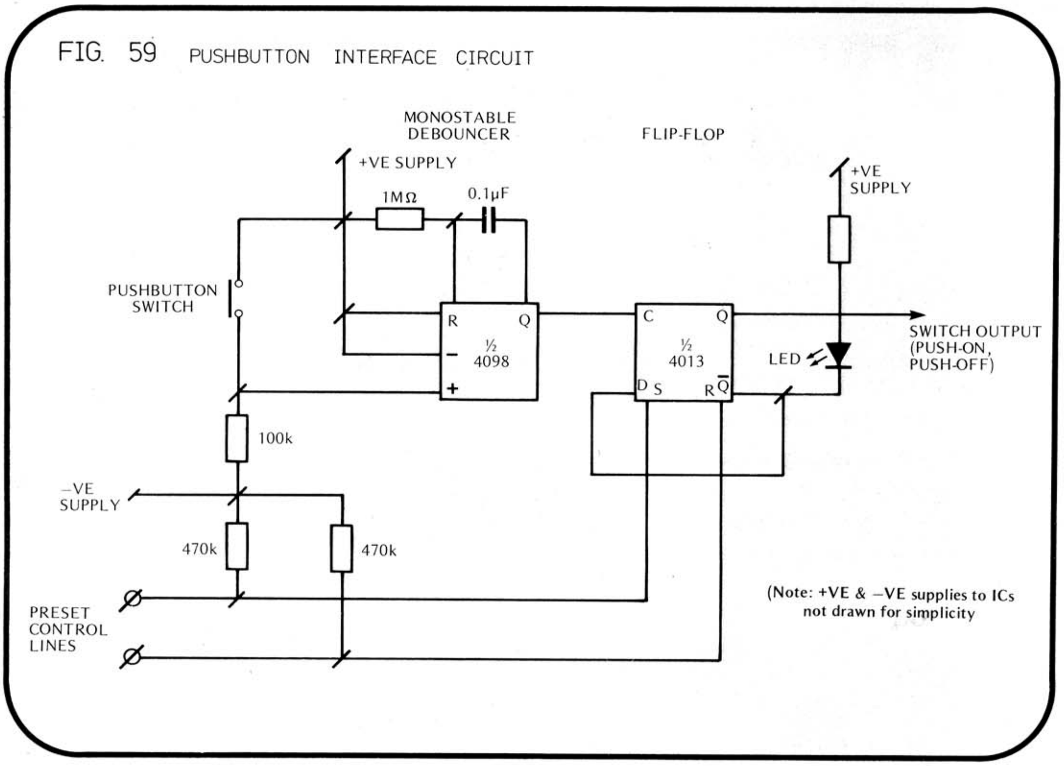





edges from a switching signal. Switch contacts do not, as one would expect and lustfully desire, simply make contact when pressed then break contact on release. The two bits of metal graunch against each other or bounce a few times whilst moving together or apart, resulting in a series of ragged spiky 'almost contacts' rather than simply touch or not touch.

Ordinarily, this doesn't matter too much, but if the switch is feeding a bistable flip-flop (as here) the fun begins. Flip-flops are usually 'edge-triggers', meaning a positive going state—another pulse flops it back and so on. A string of rapid untidy and unpredictable pulses, as kindly provided by nearly any mechanical switch, sends bistables bananas, with a vengeance.

Slugging the switch with time constants is nearly foolproof, but practically faultless is the arrangement in **Fig 59.**

The *4098* contains two monostables (which is handy since the *4013* contains two flip-flops). It can sense either positive or negative transitions, positive in this application, catch the very first input transition and stuff out a uniform, clean, predictable clock pulse for the flip-flop. Subsequent bounces and scrunches merely extend the output pulse slightly, but don't generate any spurious output transitions.

Flip-flops can have their outputs 'jammed' by stuffing the required state up 'Set' (making the Q output go positive) or 'Reset' (negative). Remote control on a plate.

There are a few unconventionalities in the logic design, all done in the name of reducing component count, largely obviating level-shifting transistors and other cheesiness, whilst maintaining the inviolable 'ground for active' law of control interfacing. (This is a common-sense rule that simply means that any accessible control line should just need to be taken to something reasonably groundish in order to activate whatever it's supposed to, not a specific voltage above or below ground. This helps avoid the "should this go to +5 or –24 volts —BANG! oops, sorry . . ." routine whilst greatly simplifying system design—grounds are omnipresent.)

The main reason for the unusual logic powering **(Fig 57a)** stems from the use of a bipolar PROM in the assignment logic. This needs a tightly controlled 5V supply unlike the CMOS ICs which will run off nearly anything with 'volts' written on.

## What's a PROM?

PROMS (or Programmable Read-Only Memories) are digital devices used extensively in computer technology for storing lots of individual items of information or sequences of information that are regularly referred to.

'Memory' is self explanatory.

'Read-Only' means that in normal operation it's only possible to retrieve the information that's stored, not to put in new information or modify contents.

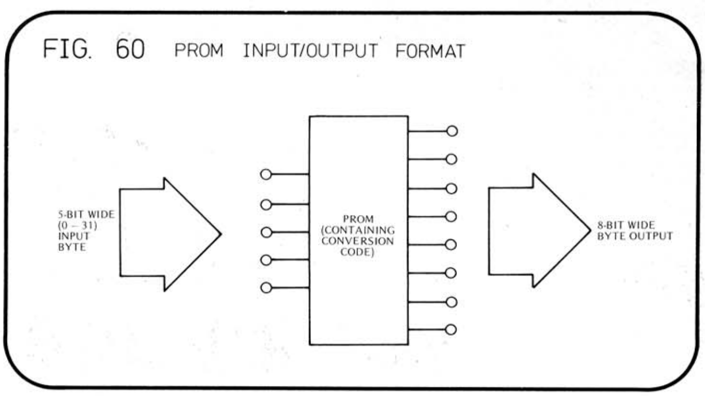
'Programmable' means that someone somewhere with the appropriate magic box can stuff already prepared information into the PROM for you. With some types of PROM he can even erase what's in there for you by giving it a quick holiday under a sun-tan lamp, then re-program from clean. The type we're using in this design can't be re-stuffed through, since the programming is achieved by literally blowing tiny internal fuses in the 'shape' of the data. This seeming inversatility is reasonable with such devices where the device cost is cheap compared with programming costs (human time).

The information stored is of course binary in nature—an '0' or a '1', up or down, there or not, etc, and the number of these binary bits contained in each PROM can be up to 65,000 odd, 8,192 and 16,384 being very common. For this channel system control, the PROM used stores a paltry (!) 256 bits which in fact is still a wee bit overkill, but they don't really come much smaller.

This 'baby-PROM', a Harris *7602,* is much like most adult PROMs in that the bits are organised internally in chunks eight wide as a digital word (byte). Eight happens to be the byte width of most popular microprocessors, that's why. In the baby-PROM there are 32 such bytes of stored data ($32 \times 8 = 256$) each being accessible with a specific 5-bit wide address code (given by the binary numbers from 0 to 31). This format is diagrammatically represented in **Fig 60.** For any of up to 32 'command' states, pre-programmed responses for the eight output lines are immediately accessible.

This particular type of baby PROM is usually used at the 'top-end' of microprocessor memory maps where a 'page' (256 bytes) is given over to the processor's function 'vectors', such as interrupts. As an example, if the processor receives a 'non-maskable interrupt' (NMI) it usually means "*Panic!*—the power is collapsing!" or similar. NMI makes the processor 'look' at a certain address in the baby-PROM's page, which tells it where to find in memory a program to 'save the environment', ie hide safely all the crucial operating data, quickly.

In the context of our channel system, the PROM outputs drive the analogue switches (organised per **Fig 57**) to route and control the channel and monitor signal paths through the system elements. This occurs in accordance with and under the command of the PROM address

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# Mixing console

inputs, which are indicators of selected channel function (Rec/Mix/Dub, etc), local or remote fader reverse commands and, importantly, mute and solo status.

Whatever's happening to *Studio Sound?* First a computer program, now (egad!) a truth table! Signs of the times, one supposes. Digits are here to stay. If you don't like such things, well it's downhill all the way from here.

This is the input/output truth table for the program burnt into our baby-PROM. The input and output binary lines are tagged as a guide to their function in the real world of the channel system. The 'O/P HEX' column is the numerical value in hexadecimal notation of each output word, necessary information for the nice little programming bod.

Insomniacs and other weirdos will be able to pass much time referring the truth table to **Figs 57a** and **57b** working out exactly what happens to the channel under all control conditions. Fear not, it all works.

That just about deals with all the software involved—relatively painless. Most of the control logic is still done in hardware, largely consisting of jammable debouncer/flip-flops. For the channel function control, a single pushbutton that steps through the four functions is realised by a simple 2-bit counter (IC 23 in **Fig 57b**). This generates a 2-bit code that feeds both the PROM control inputs and a *4028* binary to decimal decoder, IC 25, which drives the relative status indicating LEDs.

Solo, solo unlock and solo safe are dealt with in ICs 16, 20 and 24 but the relevant action on the analogue circuitry is still executed via the PROM. It can be deduced that the PROM's 'solo' command and mute do just the same thing—resulting in a fair number of duplicated and redundant program codes within the prom. At least this gives room for expansion or function modification if and when required, by simple card link changes and a differently programmed PROM.

PROM TRUTH-TABLE WITH BIT FUNCTIONS ANNOTATED

| 4 INPUT 1 | | | | | 7 6 OUTPUT 3 2 1 0 | | | | | | | | | |
|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|
| No. | H | BINARY | | L | HEX | H | BINARY | | | | | | L | |
| DECIMAL | CHANNEL FUNCTION MODE | | FADER REVERSE | SOLO | MUTE | HEXADECIMAL | MIC/LINE | MAIN FADER INPUT | MULTITRACK OUTPUT | CROSSFEED | SECONDARY FADER INPUT | | MAIN STEREO OUTPUT | | CHANNEL FUNCTION |
| ØØ | 0 | 0 | 0 | 0 | 0 | Ø5 | 0 | 0 | 0 | 0 | 0 | 1 | 0 | 1 | RECORD |
| Ø1 | 0 | 0 | 0 | 0 | 1 | Ø4 | 0 | 0 | 0 | 0 | 0 | 1 | 0 | 0 | |
| Ø2 | 0 | 0 | 0 | 1 | 0 | Ø4 | 0 | 0 | 0 | 0 | 0 | 1 | 0 | 0 | |
| Ø3 | 0 | 0 | 0 | 1 | 1 | Ø4 | 0 | 0 | 0 | 0 | 0 | 1 | 0 | 0 | |
| Ø4 | 0 | 0 | 1 | 0 | 0 | 6A | 0 | 1 | 1 | 0 | 1 | 0 | 1 | 0 | |
| Ø5 | 0 | 0 | 1 | 0 | 1 | 68 | 0 | 1 | 1 | 0 | 1 | 0 | 0 | 0 | |
| Ø6 | 0 | 0 | 1 | 1 | 0 | 68 | 0 | 1 | 1 | 0 | 1 | 0 | 0 | 0 | |
| Ø7 | 0 | 0 | 1 | 1 | 1 | 68 | 0 | 1 | 1 | 0 | 1 | 0 | 0 | 0 | |
| Ø8 | 0 | 1 | 0 | 0 | 0 | 6A | 0 | 1 | 1 | 0 | 1 | 0 | 1 | 0 | OVERDUB |
| Ø9 | 0 | 1 | 0 | 0 | 1 | 68 | 0 | 1 | 1 | 0 | 1 | 0 | 0 | 0 | |
| 10 | 0 | 1 | 0 | 1 | 0 | 68 | 0 | 1 | 1 | 0 | 1 | 0 | 0 | 0 | |
| 11 | 0 | 1 | 0 | 1 | 1 | 68 | 0 | 1 | 1 | 0 | 1 | 0 | 0 | 0 | |
| 12 | 0 | 1 | 1 | 0 | 0 | Ø5 | 0 | 0 | 0 | 0 | 0 | 1 | 0 | 1 | |
| 13 | 0 | 1 | 1 | 0 | 1 | Ø4 | 0 | 0 | 0 | 0 | 0 | 1 | 0 | 0 | |
| 14 | 0 | 1 | 1 | 1 | 0 | Ø4 | 0 | 0 | 0 | 0 | 0 | 1 | 0 | 0 | |
| 15 | 0 | 1 | 1 | 1 | 1 | Ø4 | 0 | 0 | 0 | 0 | 0 | 1 | 0 | 0 | |
| 16 | 1 | 0 | 0 | 0 | 0 | B2 | 1 | 0 | 1 | 1 | 0 | 0 | 1 | 0 | MIXDOWN |
| 17 | 1 | 0 | 0 | 0 | 1 | BØ | 1 | 0 | 1 | 1 | 0 | 0 | 0 | 0 | |
| 18 | 1 | 0 | 0 | 1 | 0 | BØ | 1 | 0 | 1 | 1 | 0 | 0 | 0 | 0 | |
| 19 | 1 | 0 | 0 | 1 | 1 | BØ | 1 | 0 | 1 | 1 | 0 | 0 | 0 | 0 | |
| 20 | 1 | 0 | 1 | 0 | 0 | 89 | 1 | 0 | 0 | 0 | 1 | 0 | 0 | 1 | |
| 21 | 1 | 0 | 1 | 0 | 1 | 88 | 1 | 0 | 0 | 0 | 1 | 0 | 0 | 0 | |
| 22 | 1 | 0 | 1 | 1 | 0 | 88 | 1 | 0 | 0 | 0 | 1 | 0 | 0 | 0 | |
| 23 | 1 | 0 | 1 | 1 | 1 | 88 | 1 | 0 | 0 | 0 | 1 | 0 | 0 | 0 | |
| 24 | 1 | 1 | 0 | 0 | 0 | 32 | 0 | 0 | 1 | 1 | 0 | 0 | 1 | 0 | DIRECT |
| 25 | 1 | 1 | 0 | 0 | 1 | 3Ø | 0 | 0 | 1 | 1 | 0 | 0 | 0 | 0 | |
| 26 | 1 | 1 | 0 | 1 | 0 | 3Ø | 0 | 0 | 1 | 1 | 0 | 0 | 0 | 0 | |
| 27 | 1 | 1 | 0 | 1 | 1 | 3Ø | 0 | 0 | 1 | 1 | 0 | 0 | 0 | 0 | |
| 28 | 1 | 1 | 1 | 0 | 0 | Ø9 | 0 | 0 | 0 | 0 | 1 | 0 | 0 | 1 | |
| 29 | 1 | 1 | 1 | 0 | 1 | Ø8 | 0 | 0 | 0 | 0 | 1 | 0 | 0 | 0 | |
| 30 | 1 | 1 | 1 | 1 | 0 | Ø8 | 0 | 0 | 0 | 0 | 1 | 0 | 0 | 0 | |
| 31 | 1 | 1 | 1 | 1 | 1 | Ø8 | 0 | 0 | 0 | 0 | 1 | 0 | 0 | 0 | |

## Logic meets analogue

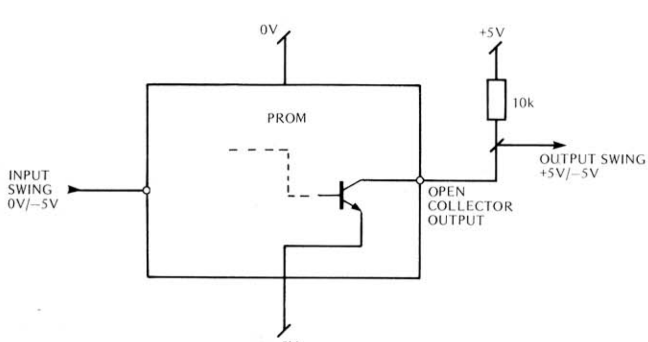
The *7602* PROM hangs between logic ground and –5V (of the split ± 5V logic supply) thus necessitating all its input feeds to be similar in swing — 0 to –5V. All the drive logic—flip-flops, debouncers and master buss logic is similarly powered.

Why?

Analogue transmission gates such as the design of **Fig 57** are required to pass (and stop) analogue signals referred to ground and therefore of both polarities, so the gates have to be fed from a split rail (in this instance the ± 5V logic supply).

Converting between the 0/–5V logic and the ± 5V control voltage swing needed by the gates is done fairly cheekily by using the open-collector output drives of the PROM (**Fig 61**). Open-collector is exactly that—there is no positive output pull-up internal to this PROM, the idea being that it may be paralleled with other open-collector devices in a 'wired-OR' buss configuration. When the output transistor is turned off, the collector is at high impedance whilst, when on, it forms a very low resistance path to the negative rail. Advantage is taken here of the high-impedance state to 'pull' the collector up an extra 5V above the PROM's internal supply —up to the +5V rail in fact. When the transistor turns on the collector dutifully zaps down to the –5V rail. It doesn't care what's at the other end of the load 'pulling' resistor provided it isn't of excessive potential (20-25V as a guess).

FIG. 61 UNIPOLAR TO BIPOLAR CONTROL SWING CONVERSION UTILISING PROM OPEN-COLLECTOR OUTPUT




Some of the analogue switches are driven directly off the PROM outputs, whilst others have the necessary inverse-switching feed provided by a conventional inverter.

As a note to the unwary, bipolar memories such as the *7602* drink a lot of juice and splash around large amounts of this current when being switched. This explains the large amount of decoupling festooned around it and the logic supply rails generally. Needless to say, the analogue transmission gates are referred to *audio* ground, not the click-infested logic ground, despite the fact that they are powered off the logic supply rails.

## To Schmitt or not to Schmitt

Throughout the entire console design, a large number of *4011* quad dual-input NAND gates are used, even to the exclusion of other device types such as inverters, where the two NAND inputs are strapped creating an inverting buffer. This is largely for convenience and to minimise inventory types, incidentally resulting in a cost advantage (as a result of the greater number bought) over acquiring lots of small quantities of differing IC types.

*4093*s are a plug-in replacement for the *4011*, with the difference of their Schmitt-trigger action—handy in cleaning up dirty bits and useful sometimes for switch de-bouncing. As a general rule, a *4093* may be used anywhere a *4011* is. A proving exception to this is the caution which must be observed when plugging them into positions where they are used as inverters for analogue transmission gates. The hysteresis (about 2V) intrinsic to the Schmitts can result often in the two potentiometric switching elements both being momentarily in similar states during switching both 'off' or both 'on'—until the Schmitt threshold is reached by the rising/falling control voltage. At this point the gates flip rapidly into their correctly opposing states.

The simultaneous states manifest themselves as switching clicks and splats; both 'off' leaves the series gate vulnerable to signal breakover and potential death from high source programme levels. Both 'on' ties the virtual-earth following amp input via a low impedance to ground, causing the amp to have an abrupt and enormous burst of high gain.

Superior devices, in odd circumstances such as this, do not necessarily mean better performance! ■

# Designing a professional mixing console

Steve Dove

# Part Twelve ~ The Channel Front~end

**Console input stages largely determine the measurable performance standard for the entire system — the quoted noise, distortion, bandwidth and phase characteristics. How these relate to how the console *sounds* is an altogether separate affair — optimisation is not the same as maximising or minimising parameters.**

OPTIMISING the front-end sound is down to shrewd judgement in balancing the nearly endless electronic operating conditions such that adequate performance obtains over the wide range of input signals expected and common. Any wrinkles such as there may be, are arranged to exert influence only possibly under quite extraordinary operational conditions.

This front-end design ties in with two previous articles in this series: Part Four (December 1980) where it fulfils a logical extension and practical conclusion to the design philosophies expounded there, and Part Eleven (September 1981) where it completes the entire channel system from mic in to mix out. It would seem a bit lonely without the support and context lent by the two other articles.

## Origins

The mic-amp is a somewhat developed version of a basic front-end design (**Fig 62**) which is in grave danger of becoming an industry standard. The precise origins of this rather clever two op-amp arrangement are obscure, but it's been around a few years.

Initially most striking is the manner in which a single-track potentiometer is used to simultaneously vary the gain of two amplifying elements — the front-end, non-inverting stage and the succeeding inverting amplifier. Since the first stage is (as far as its inputs are concerned) a conventional non-inverting amplifier, transformer input coupling is not any more problematic than with simpler mic-amps (eg **Fig 21**, December 1980).

With maximum gain distributed between two stages, large gain is possible without any danger of running out of adequate steam at high frequencies for feedback purposes. This incidentally also makes for reasonably simple stabilisation of the amps.

Other than the obvious neatness of one-pot gain control, two nice features are inherent in the design which are delightful from the points of view of system level architecture and of operation respectively.

## Level architecture

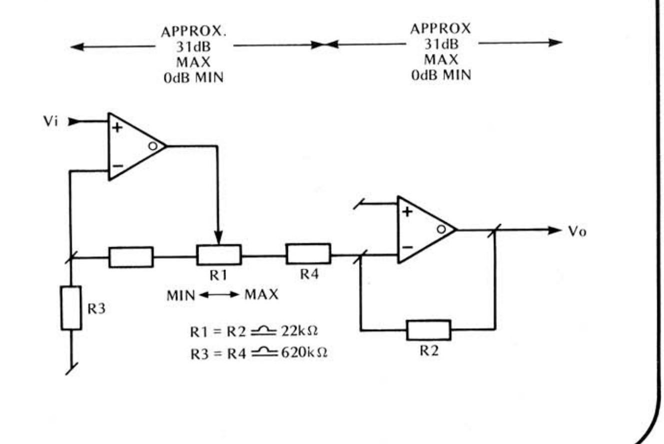
System level architecture is largely concerned with operating all the elements of a system at the optimum levels and/or gain for noise and headroom, ie at a comfortable place somewhere between the floor and the ceiling. Where gain is involved, it's important that the resultant noise be due primarily to the gain stage that has been optimised for noise (or rather lack of it) such that it can then mask all the other hopefully minor contributions. At no point in the gain swing — particularly at minimum gain — should it be necessary to attenuate away unwanted residual gain. The amount of attenuation gets directly subtracted from overall system headroom — what good is 24dB of headroom everywhere else, if you've only got 16dB in the front-end?

In this respect circuits similar to **Fig 62** score well. The graphs of **Fig 63** show why. **Fig 63a** represents the gain in dB of a simple non-inverting amp varying with the percentage rotation of an appropriately valued linear pot in its feedback leg. This is like the gain/rotation characteristic of the first amp of **Fig 62.** Similarly, **Fig 63b** is the gain/rotation plot for a linear pot as the series element in an inverting amp, such as the second gain stage of **Fig 62.**

For the first half of the rotation, the first stage provides all the gain swing and most of the gain, only about 6dB being attributable to the inverting stage at mid-point. Toward the end of the rotation, the position reverses with the front-end remaining comparatively static in gain, the extra swing and gain coming from the inverting stage.

Noise criteria are met, since the first (optimised) stage always has more than enough gain to allow its noise to swamp the second stage with the exception of at minimum gain. There it hardly matters anyway because the front-end noise contribution is going to be at a similar level to the overall system noise-floor ie *really* quiet! The impedances around the second stage largely determine that amp's noise performance and this is such that it need not be considered in relation to input noise at any sensible gain setting.

FIG. 62 BASIC MIC-AMP DESIGN




Headroom is satisfactory as no attenuation after the first gain-stage is needed for any gain setting.

An operation goody can be gleaned from **Fig 63c**. This is the combined gain/rotation curve for the total two op-amp circuit. Note that for a very large percentage of rotation around the middle (where it's most often used) the dB gain change per rotation is as good as linear. It just gets a bit cramped at the top and bottom but you can't win 'em all.

For reference a little later on, it may be noted that there are two available resistors $R_2$ and $R_3$ which may be used to modify the gain structure independently of the potentiometer.

As a microphone amplifier, the fairly high optimum source impedance of the op-amp used (a Signetics *NE5534* in this instance) has to be matched to the likely real source impedance of some 150 to 200Ω. No apologies are offered for the use of transformer input coupling (Part Four, December 1980).

A Sowter type *3195* (1:7 ratio) is used here. Many circuit values (marked with an asterisk in **Fig 64**) — some in some quite unexpected places — are dependent on the specific transformer type in use. Other excellent transformers, notably the Jensen *JE-115-K*, can be very successfully used provided the differing ratios are taken into account in level calculations. Phase and response trimming values will vary significantly — with Deane Jensen's *JE-115-K* it is in fact simpler than with Dr Sowter's.

Despite the circuit's apparent simplicity, a lot of effort has gone into defining the front-end bandwidth and straightening out the phase response at audible extremities. Taming the hf resonance is quite tiresome.

On the front of the transformer hangs the usual stuff to make the mic-amp useful in this world of capacitor microphones: a 20dB input attenuator and phantom power —

48V via 6.8k per leg carried common-mode along the mic line. This should see to all-comers.

A line-in option is brought in via the transformer also, featuring far stiffer input attenuation (about 36dB) whilst simultaneously disabling much of the gain swing of the first amp. The resultant gain swing of 35dB (between −25dBu and 10dBu input level) with a bridging-type input impedance of some 13kΩ should accommodate most things that the mic input or machine-return input differential amp can't or won't. A small equalisation network is used in the attenuator to bolster up the extreme lf phase response.

Common-mode rejection in the transformer is dependent mostly on the physical construction of its windings. The Sowter, in common with most other transformers, may be in need of compensation by deliberately reactively unbalancing the primary winding. Jensen transformers are uncannily good in this respect — no tweaks usually being necessary.

Less than perfect CMR shouldn't cause any ill manifestations in a typical recording environment with fairly short input leads. A high rf field of any sort, or an application with very long leads or worse yet, a multicore, is far more likely to create problems with untrimmed inputs than with those properly balanced. Vulnerability is greatly increased to all types of common-mode nasties including noise on the phantom power supply feed. Indeed this is a common compounding of faults on a desk that exhibits consistently noisy inputs.

A minor compromise is necessary in the first stage to prevent it gasping with exhaustion on extremely high input levels. Ideally, the op-amp's output has to look into an impedance of 600Ω or greater (this being the lowest impedance it can drive full output voltage swing into). Maximum gain state isn't really a problem — if the first stage was gasping out of puff into the second stage input stopper resistor, the filter output would be some 30dB into clipping and someone might notice.

No, the dodgy bit would be at minimum gain where the first stage is operating almost as a follower, its output load being some 770Ω of the remaining feedback path to ground. That's safe. Unfortunately, it would be nicer if that small resistance were a lot smaller since it is contributing unwanted thermal noise to the otherwise beautifully optimised front-end. Before you rush for the smelling salts, the degradation in calculation is only minor points of a decibel and in practicality is easily lost in the grey mist that always surrounds the marriage of calculation with practical noise measurement.

The idea of using a front-end stage that turned into a follower under operating conditions did cause trepidation at first but it has proved stable without any obvious trace of ringing within its bandwidth. This is probably because it is only being asked to look into safe cosy unreactive loads. The most horrid and evil things that will make any incipiently unstable circuit squeal in horror have left it quite cold — amongst the instruments of torture have been pulse generators/storage scope and rf sweep generator/spectrum analyser. The 22pF compensation capacitor is more an act of conscience than a practical necessity.

Down from the nether world of Megahertz, the mic-amp is quite stable at audio even with the mic unplugged and input unterminated; the input 'zorbal' is designed to work in conjunction with the fairly low input impedance of the *5534* (150kΩ the book says).

## The limiter

Elaboration on the simple two op-amp mic-amp element consists of arranging an automatic gain control element in the feedback loop of the second amplifier and following that with a variable turnover frequency highpass filter.

A photo-resistor device has its resistive end strapped across the normal gain determining feedback resistor. Its resistance drops in value from very high (MΩ) inverse relation to the photo-diode current to a limit of around 300Ω at about 20mA diode current. This resistance swing provides plenty of gain swing in the second amp for use in a peak limiter arrangement.

Selectable to be able to pick off from either the highpass filter output (as an input limiter) or from after the post-eq breakpoint downstream (as a channel limiter) the limiter side-chain is true symmetrical peak-detecting. A positive-going and a negative-going level detecting comparator are switchable between 'clip' detection (2 to 3dB before system headroom) or 'programme' level (nominally +8dBu but actually internally tweakable up and down.

A bi-colour LED blinks red to indicate limiting in action and when the limiter is disabled, it blinks green to signify that the selected level (clip or programme) is being reached or exceeded. In this 'indicate' mode, the limiter integration time-constant is deliberately shortened to make the green flashing similar in character to the red flashing in 'limit'. (The difference is due to the nature of servo-loops, of which a feedback limiter such as this is an example. In limit, the loop is self-regulating, the gain-control element holding back the audio level so that it's just tickling and 'topping up' the side-chain. In indicate, the loop is broken and there is no such regulation. The green light stays on whenever the threshold is exceeded and tends to hang on for a bit whilst the time-constant capacitor

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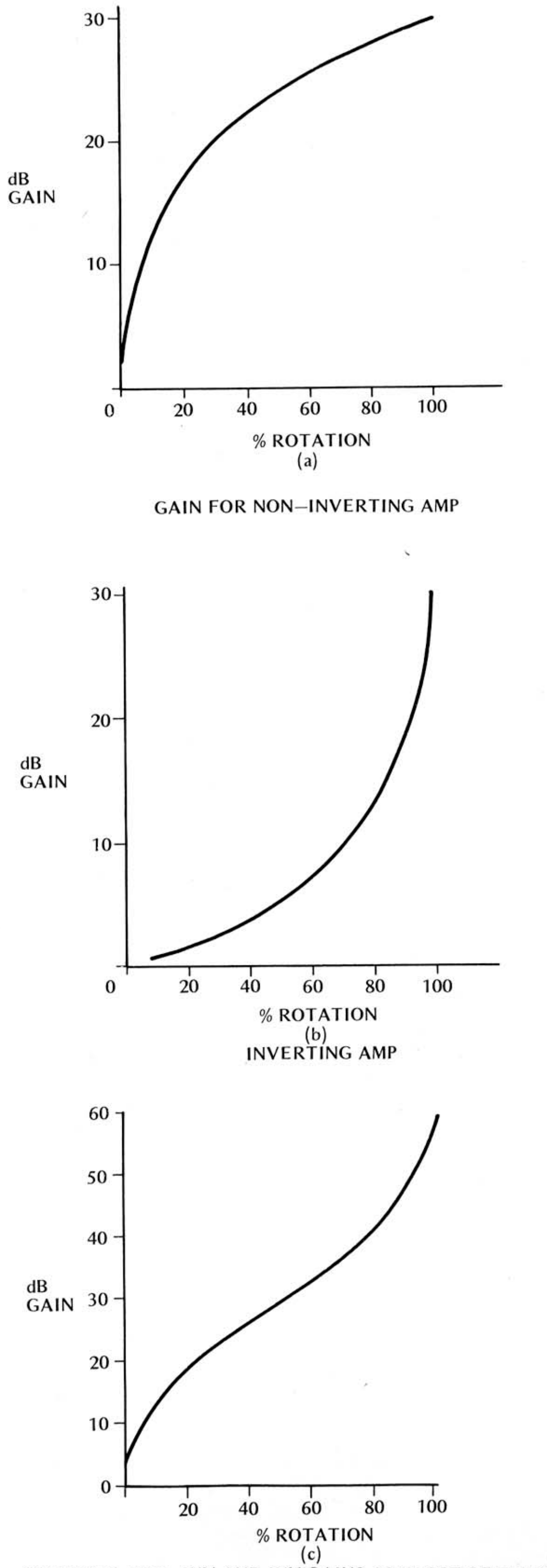
FIG. 63 GAIN vs ROTATION FOR LINEAR POT




COMBINED NON–INV AND INV GAINS AS IN CCT OF FIG 62

# Mixing console

discharges. With even a minor overload, this hangover can extend for quite a few finger-drumming seconds — hence the shortened time-constant.)

As an experimental aid both the attack and release time constants for the side-chain are on presets although it's suggested that once comfy settings are found, fixed value resistors of the measured value be substituted for them. Two reasons: good presets are unbelievably pricey these days; trying to get a number of channels exactly the same by ear is a mission for fools.

One thing this limiter isn't, is subtle, being designed primarily as a protection limiter. The comparators deliver a rail-sized wallop to the integrator upon threshold, softened a bit by the attack preset in conjunction with the comparators' output impedance. This rather unusual approach is to help 'wake up' the photo-resistor which has a relatively leisurely response time — the combination can be adjusted to slow enough that it doesn't *clip* whilst fast enough to prevent an audible *snatch*. Overshoot is generally well within 1dB on normal programme, given a release time long enough to prevent ridiculous pumping.

As a rough guide, if it's intended to use the limiter for sporadic transient protection it's best to aim for short attack and release times, bearing in mind that such settings will behave more as a clipper to the lower frequencies. For continual effect use, longer time constants will be less gritting and more buoyant. This side-chain arrangement certainly behaves differently to more conventional FET or VCA linear proportional systems and needs a slightly different approach in setting up.

From a design viewpoint, there is an awful lot of spikey current hammering about — into the integrator and through the LEDs of the indicator and photo resistor. This current is kept well away from ground, where flowing through the low level mic-amp ground path it sounds reminiscent of only slightly louder than a machine gun. Best to keep it all in the supply rails, where it belongs

## Highpass filter

Constructed around the front-end's line output amplifier is a second-order highpass filter. It is a completely ordinary Sallen & Key type filter, arranged to use a dual-gang equal value potentiometer to sweep the 3dB-down turnover frequency from between 20 and 250Hz. A click-stop switch at the lf end (anti-clockwise) negates the filter, replacing it with a very large time constant, single-order dc decoupler. The filter and the decoupler are both tied to reference in order to minimise clicks — fortunately the *TL071* used in the filter uses barely any input bias current, so there is little developed offset voltage from that source to worry about.

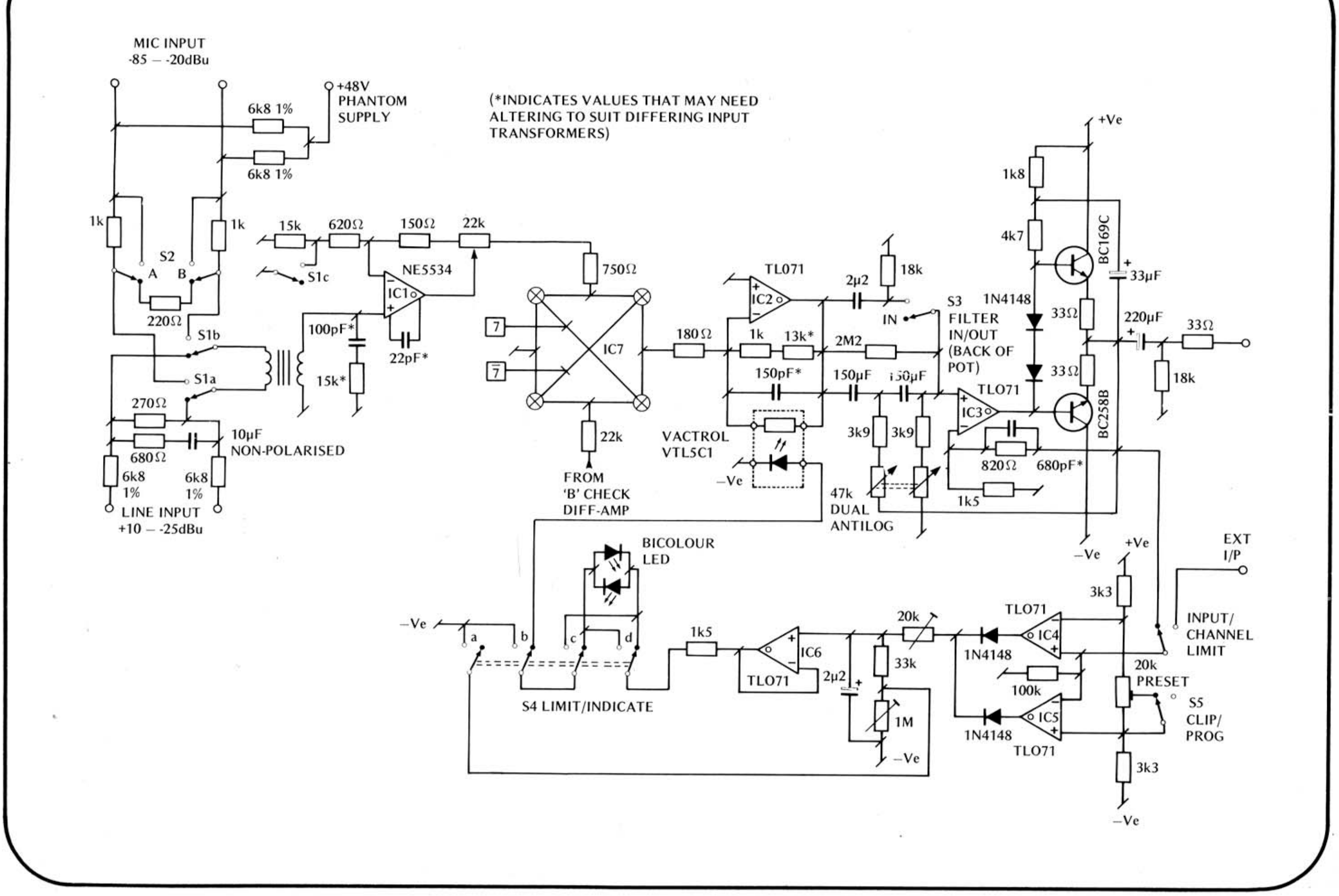
Being an equal-value filter, the Q or turnover would be very lazy indeed if the feedback were not elevated in level to compensate for the upset resistor ratio. Here a compromise is struck. A low Q gives a very gentle roll-off and not very rapid turnover, but causes least phase disturbance. A high Q gives fastest roll-off at the expense of frequency response — pronounced bumps — and frantic phase response exhibited as ringing and smeared transients. Just like your monitors. Oops, sorry.

An uneasy medium lies where the in-band frequency response is maximally flat and for this a feedback hoist of around 4dB is needed. This gain is taken across the filter as a whole, with the second stage of the mic-amp arranged to sustain a 4dB loss to compensate — it all works out in the end, with no compromise of headroom. With minimum gain set, there is still about unity electronic gain front to back. An added convenience of gain is that it provides a better chance of shoring up feedback phase margin — quite important in a line-amp that may have to drive a lot of heavily capacitative cable. Also, it provides yet another single-order lowpass pole to help iron out the mic transformer's hf resonance!

## Where does it go?

Somehow or other, this fine front-end has to get glued into the channel system. That is what the analogue CMOS switcher is doing in the middle of it. It is the self-same switch that is looking lonely in the top left-hand corner of **Fig 57a** (September 1981). The switch's purpose is to route either the mic-amp or the machine return ('B'-check) diff-amp output into the main signal path under command of the channel-mode switchery via the PROM interpreter. Unity gain from diff-amp output to filter/line-amp output is achieved by the appropriate switch-source resistor. ■

FIG. 64 CHANNEL INPUT AMPLIFIER, HIGHPASS FILTER AND LIMITER



# Designing a professional mixing console

Steve Dove

# Part Thirteen~The Back~end 1

**A little of something for everybody—this part concludes the channel module description with details of the auxiliary feeds, the audio path summing and output stages—whilst not for the weak of heart there is a treatment of how to make virtual-earth mixing actually work.**

Much of what follows refers back to Parts Three (November, 1980) and Six (March, 1981) of this epic series. So for inveterate *Studio Sound* hoarders I suggest you dig the copies out before we continue.

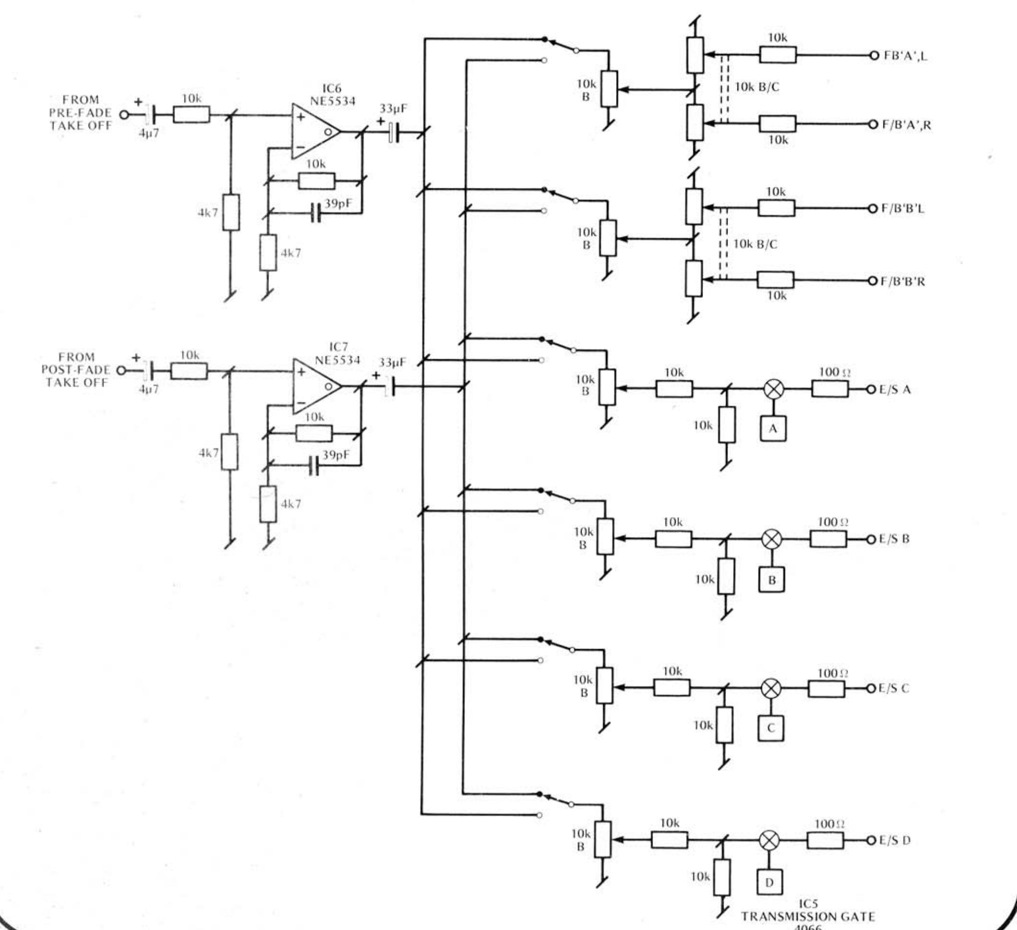
## Auxiliary channel feeds

Two prefade (pre-mute) feeds are provided on each channel, each with level control and pannable across a stereo pair of mix busses. This provides a versatile facility enabling separate stereo submixing, two rather flash stereo foldbacks or four separate feeds. Each of the pairs is selectable to post fade should extra effect feeds be needed during a heavy mixdown, whereupon they will be subject to channel mutes also.

Four individual post-fade effect feeds are individually mutable, locally or remote, individually level controlled and selectable to pre-fade (should foldback requirements get a bit silly).

Pre/Post switching is done via 'real' switches—in fact push-pull switches operating concentrically within the level control potentiometers, saving hassle and panel space (in normal, out reverse). There is no conceivable time the function needs to be remoted so there is no drawback there to mechanical switching. Effects feeds, though, are quite often twitched during mixes; consequentially, remotes are brought out to a rear multiway socket to facilitate linking to automation should the need be. Local activation is achieved through the debounce/latch arrangement used extensively in the channel mode switching **(Figs 65a and 65b),** the latch output driving a very simple single element transmission gate per feed to buss—isolation, crosstalk and noise criteria are not particularly critical on these feeds, but still come out quite creditably. The console switch-on reset master reset buss (MRB) cancels all these feeds leaving a clean slate rather than the alternative unpredictable hordes of 'ons' and 'offs' in the event of a power interruption or control 'zeroing'.

FIG. 65a AUXILIARY SEND CHANNEL AUDIO PATH




## Summing modules

Much of the actual mixing within the system so far described is self-contained—multitrack routing, when achieved via a matrix, allows multiple sourcing to any chosen 'group' or machine track. A stereo mixdown of all the channels is possible by this method by selecting them to an arbitrary pair of tracks

FIG. 65b EFFECT SEND LOGIC CONTROL

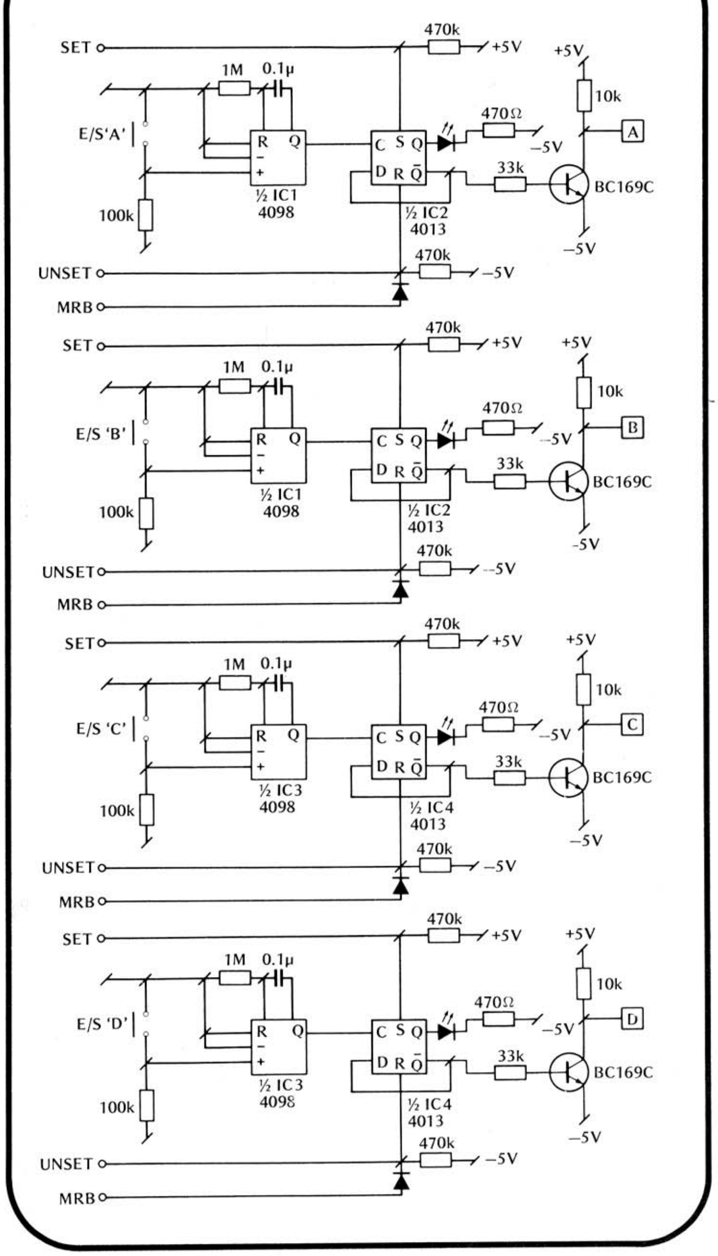
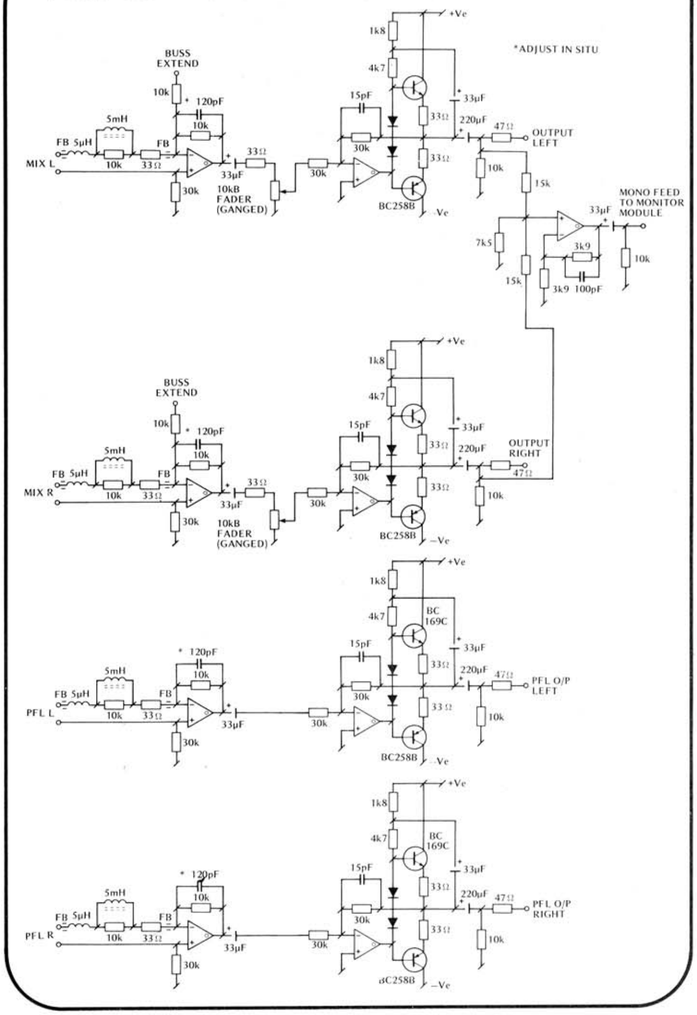



FIG. 66 MASTER GROUP MODULE




across which the mastering machine is hung. This is in fact the mixdown technique used in many console systems whether in-line or discrete monitoring; although entirely feasible it is not the manner in which this particular system is intended to be used, unless your really out to make life hard for yourself.

Stereo mixdown is achieved in the same busses as the multitrack monitor mix, the 'solo' monitor function making its happy home here too.

A master group module contains the mix-amps, fader and line-amps pertaining to the stereo buss together with sundry other related things, like mono summing (Required for a monitor feed) and clean auxiliary buss access for extending the monitor mix (for effect returns or temporary extra channels).

The circuit diagram of **Fig 66** in its simplicity belies the hidden design which is in the relationship of the circuitry to its mechanical and electrical environment.

This is where the care and feeding of op-amps (Part Three, November 1980) and grounding paths (Part Six, March 1981) really pay dividends —or not if you aren't careful. Mix-amp stages—with large numbers of permanently assigned sources such as in the main mix busses —are as crucial to the overall well-being of a console as any front-end stage. In a typical situation, as a unity gain virtual earth mixing stage with 33 sources (channels plus access) the amplifier is being asked for 30dB-ish broad band gain—as much as any other stage in the chain including both the mic pre-amp or secondary input stage.

That this mix-amp gain is sometimes referred to as 'noise gain' is not accidental. Unless care is taken to balance fader-back channel noise contributions against this self-generated mix-amp noise, the latter could well predominate and arbitrarily determine the noise-floor for the entire console. Similarly, channel noise contribution due to gainy buffer-amps should equal or outstrip mix-amp noise. Other compromises start waving flags to delight and amuse. Self-noise generation in the mix-amp is predominantly the amplified thermal noise of the source and feedback resistances, device input current noise and surface generation/recombination noise. The last two can be minimised by device choice—minor quirks simply solved. Thermal noise is physics and here to stay until the universe cools off a bit (who said entropy?). Common sense on first glance says make the mix resistors as low in value as possible, but too low a value would cause quite large signal—hence ground—currents to be thundering about, and on a less technical level necessitate yet another tier of buffer amplifiers to feed the busses after the pan controls. The buss feed resistors are also deliberately used to modify the law of the log/reverse-log pots used for the pan. Whilst not materially affecting the centre-pan attenuation, this trick can help the subjective linearity of an image sweep across stereo versus control rotation, which can otherwise be a little too concentrated at the ends of the control. A law unto itself.

Ordinarily though, the mix resistors are of such a value that in the context of a complete mixer the combined effectively paralleled resistance is well below the optimum source impedance of the mix-amp device used, so the primary noise modes are those previously mentioned. This isn't too difficult with FET front end devices such as

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the *TLO71* with their ludicrously high OSI. These devices have a couple of other major benefits in this application by virtue of their FET inputs. Input current, hence input current noise, is extremely low and being FETs they don't have many low-frequency junction and surface noises inherent to bipolar devices. It seems a paradoxic absurdity to use an ultra-high input impedance device for 'zero' impedance mixing.

Things can get a bit startling if the resistance/OSI relationship is awry. Above the OSI, device input noise voltage becomes an increasingly important noise contribution. Many years ago in a mixer design with bipolar device mix amps and quite high mix resistors, the measured buss noise was actually *lower* on a 20-channel version than on the 10-channel original. It wasn't until many visions had passed of Nobel Prizes and Rolls-Royces that we sussed what was happening. Increasing the number of source resistors reduced the buss impedance above the amplifier's OSI, through it and eventually below it where input noise voltage was no longer contributing.

Theoretical source impedance and device contributions tell less than half the story in a practical design, they may be quantifiable in the isolation of a test bench but thrown into a system they can all seem a bit meaningless. Part Six (March 1981) gave an insight—it's largely down to grounding and out-of-band considerations.

## Curly things

The funny curly things between the buss and the amplifier input in **Figs 66 and 67** are inductors—remember them? These are only small ones though; don't panic. A simplistic view is that they are there to stop any rf on the mix buss finding its way into the electronics, but this is only part of their purpose. The ferrite beads and small chokes (about 5$\mu$H) are there to increase the input impedance and hopefully help decouple the buss from the amp at vhf and mid-vhf respectively, whilst the larger inductance creates a rising reactance in one phase sense to mitigate the falling reactance of the buss capacitance. If left unchecked this capacitance would cause the mix-amp extreme hf loop gain to scream off into the ionosphere turning it in to a lovely rf oscillator on the way. Feedback phase-leading around the amp stops the gain *rising* but if it weren't for some series loss, accidental or deliberate, in the input leg it would be insufficient to hold the amps' phase margin within the limits of stability at the bandwidth extreme where device propagation delay becomes significant in the loop. A small series resistance can provide this loss whilst also defining the maximum gain to which the circuit can rise, whilst the parallel inductor/resistor combination improves on this in a few important respects.

The inductor is calculated to present low in-band (<20kHZ) reactance, allowing the mix-amp to operate on the buss in a virtual earth (zero impedance) configuration. The reactance rises gently at the audio hf end, imparting little frequency response anomaly but a definitely beneficial partial phase-straightening against the inevitable effect of heavy feedback phase-lead compensation.

At even higher frequencies, the inductive reactance continues to rise until the combined network impedance is limited by the resistor, itself a high enough value to afford a sizeable choke to buss nasties and to define amplifier out-of-band gain to a reasonably low value. It is low enough, however, to stop the inevitable inductor/buss capacitance resonance getting completely out of hand. Making an L/C oscillator is one way of preventing spurious instability, one supposes.

Whilst FET inputs are far less prone than are bipolar inputs to the intermodulation and direct demodulation effects that cause rf interference to appear out of nowhere, this fairly healthy brace of filtering may be helpful to those unfortunates living within spat-upon distance of Crystal Palace/Empire State Building/Mount Wilson or

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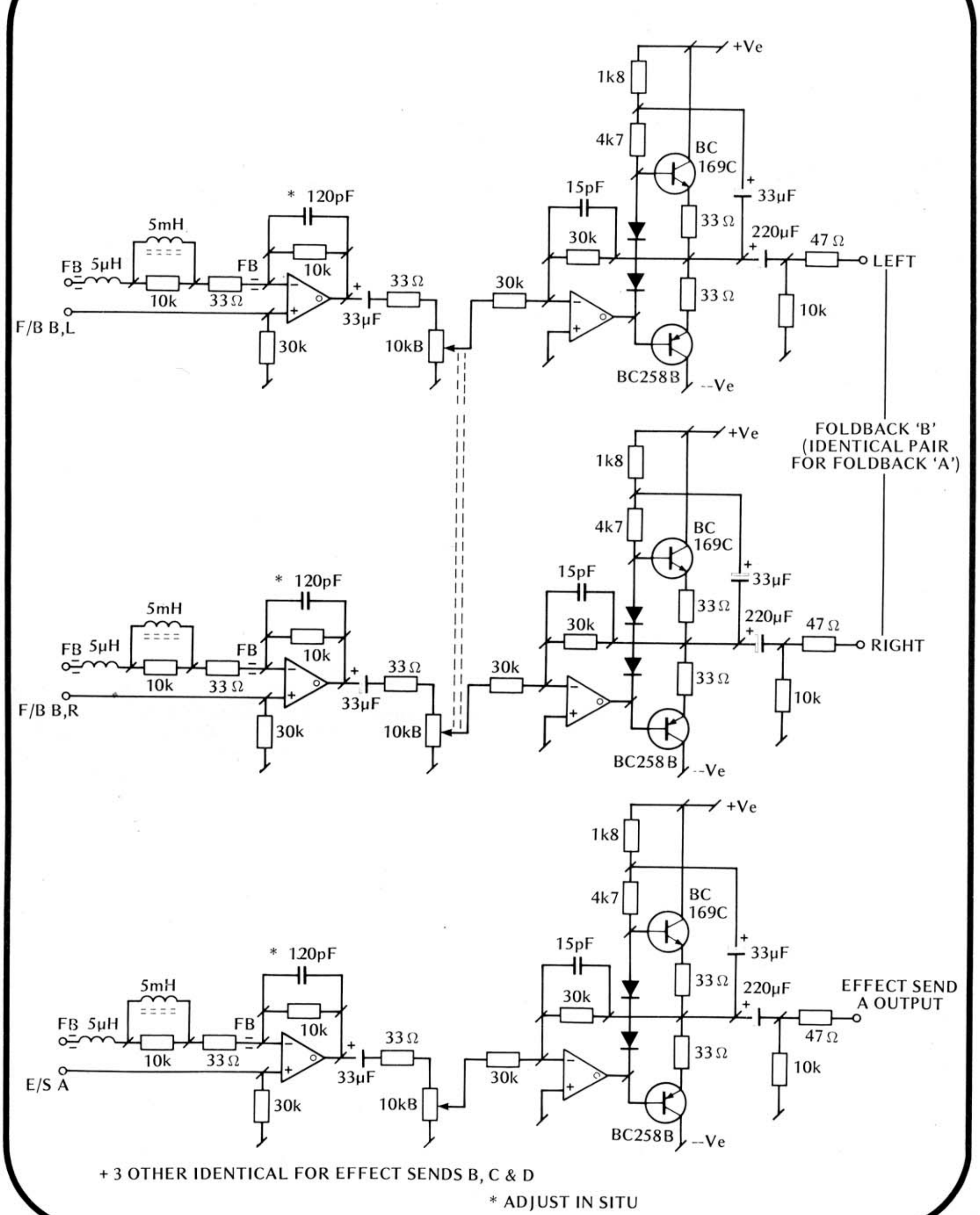
FIG. 67 AUXILIARY GROUP (GARBAGE) MODULE



some other unsociable source of vhf, megawatts.

## Alternative mixing

There are of course alternatives to single-buss virtual-earth mixing. Passive resistor mixing **(Fig 68)** is quite viable for fixed-assignation systems that aren't going to be chopped, changed or switched in and out of. A major advantage is that buss capacitance is merely something to be taken into account of in terms of response and phase, rather than directly imperiling the stability of the mix-amp. For passive mixing, the mix-amp is just a buffer amp to make up the loss in the resistor tree and rf filtering becomes simple with known filter source and load impedances together with the ability to refer against ground. Primary hang-up is that the buss is unbalanced, has an impedance at audio (albeit fairly low), and hence lays itself wide open to induced garbage and capacitatively coupled crosstalk. Despite this, it is a method used with considerable success for many years in quite a few production mixers.

Distributed or devolved mixing **(Fig 69)** uses local mix-amps to sum blocks of, say, eight channels, the outputs of these local amps then being taken to a common summing point. This quite neatly obviates having to deal with a long buss but does create a practical problem of locating the distributed summers, preferably where it doesn't mean having to dismantle the mixer to get at them!

Both passive and devolved systems have the advantage that large amounts of the 'buss' can be run in screened single cable, the extra capacitance not having the awful consequences it would with virtual earth.

FIG. 68 PASSIVE MIX AMP ARRANGEMENT

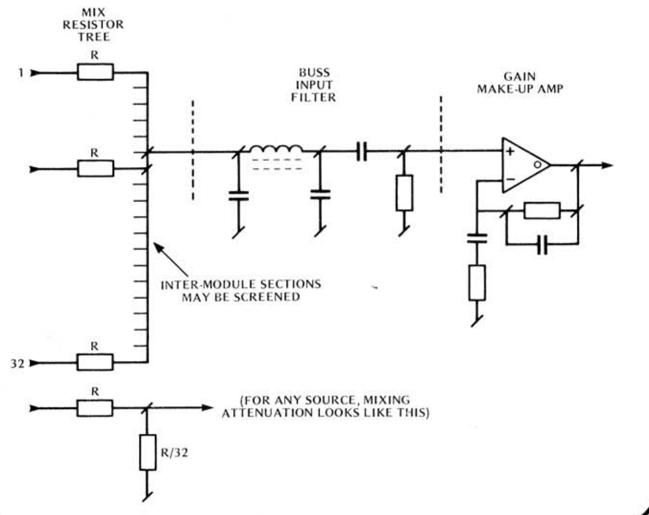
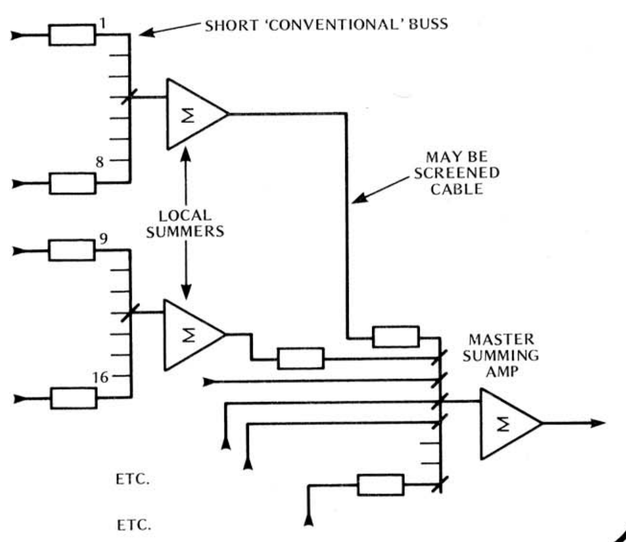



FIG. 69 DISTRIBUTED OR DEVOLVED MIXING




For consistency all busses would be run devolved meaning sub-mix facilities for the PFL busses, four effect sends, four foldbacks, the main stereo/monitor mix and the eight subgroups (if used) together with having to arrange the master mix for each of those at the grouping end. Aarrghh! life's too short, really!

## Virtues of earthing

The console's virtual-earth mixing busses all end up in identical mix amp/attenuator/line amp configurations. Exceptions are the mono sources (effect sends) which have individual master level controls rather than ganged stereo attenuators and the PFL (which does not need a level control anyway, being a purely monitoring function). These back-end stages are homed in two of the very few one-off modules in the design: stereo monitor/mix (with the master fader) and the PFL summing occupy the Master Module whilst the remaining auxiliary functions are summed in the Auxiliary Master (or Garbage) Module. **(Figs 66 and 67).**

The outputs, each low impedance unbalanced, are taken to the jackfield, where they are normalled to their appropriate destinations and directly bridged by the differential inputs of the monitor selector switching matrix (adjacent to the field). It is assumed that the studio system will operate on the unbalanced out/differential or balanced input principle—output transformers need to be added if not.

Grounding paths for virtual-earth mixing—especially in long mixers—are always the final arbiter on how far down the system noise floor will go and how susceptible the mix stage is to extraneous fields and earth currents. In this age of digits ground paths are especially crucial. Remember (Part Six) how the 'ground' on the non-inverting input of an op-amp mix-stage gets amplified up by the 'noise gain' of the stage? This implies that a ground noise of –100dBu will end up at about –70dBu for a 32 source mixer—barely adequate. A simple but so, often ignored rule with virtual earth stages is to make sure that the ground reference has got the same 'dirt' on it as the signal and vice versa—Yes Ground Follows Signal. If both ground and signal have the same garbage in the same phase, there's a fighting chance that it'll get ignored as common-mode and not amplified in the mix-amp. Thus for each mix-buss, there is a parallel ground buss being fed by the last relevant ground reference from each channel. Avoiding a major buss length ground loop (otherwise known as a single-turn transformer), this means that all the heavyweight signal current in the fader/mute/mode switchery has a direct wire to central ground whilst the mix-amp has a respectable output referenced ground to work against, clean of channel signal currents but representative of the say buffer-amps' reference in the case of stereo mix. The mix amp does *not* take a direct system central ground of it's own.

As a quick aside, signal path grounding in the channel is greatly simplified by the 'ground decoupling' afforded by the differential return amplifiers at the eq input and post-eq break-point return. With such a complex system it would be rather trying without them.

## Automation asides

Thankfully automation system fader modules now have separate ground terminations for at least audio, control voltage manipulation and logic. This is a welcome change from some earlier generation systems where the voltage control line shared the audio ground as a reference, occasionally with some less than healthy results. The favourite must be ground currents from heavily modulated channels—eg kick drum—twitching the ground potential against which the VCA's were referenced hence cross modulating into all the other channels. That one's fun. So is trying to lose logic chatter and wheezes—a reverse effect of the same mechanism.

Nowadays, installing or building-in a fader-and-mute automation system to a console design such as this is little more complex than putting in standard Penny & Giles faders, reading the handbook and making sure you plug into the big box the right way up. As for audio interfacing, it 'looks' just like a conventional fader—a top, a bottom (ground) and a wiper. In fact with the Melkuist in bypass that is *exactly* what you've got—a standard log fader.

Valley People's *Fadex* is a little different in that the control element is always a voltage controlled *amplifier* that can be arranged to give gain obviating the need for a post fader buffer amp to allow for fader back-off. In practice though, and especially in this design, that amplifier performs other functions also such as output drive and (specifically here) bandpass filtering. It's worthwhile considering however, if it is to be designed in from the outset. ■

# Designing a professional mixing console

Steve Dove

# Part Fourteen ~ The Back ~ end 2

**This more-or-less final part of the series (some notes and addenda will follow in the next issue) discusses alterations to the circuitry for non-standard operating levels, suitable ICs for use in the design, what will happen when components fail, and where console design is likely to go from here. Computer design techniques are also discussed.**

HAVING completed the major aspects of this design study, it really only remains to consider a few areas which might otherwise have got lost in a confusing welter of circuitry. Nothing particularly complicated, but a few important items like what happens when components decide to give up the ghost (and what it will sound like), metering and questions of headroom, component selection and so on. But first we should look at something rather basic: namely, what happens when you switch the beast on in the morning, and how the master function circuitry operates.

## Function masters

**Fig 70** shows the devastatingly simple console master function circuitry. All the clever stuff is done in the channels, allowing this bit to be little more than switch contacts. No debouncing is necessary since the master busses directly actuate the set and reset 'latch' functions of the channel function registers.

Lockouts are arranged on the fader main/reverse selection and master monitor 'A'/'B' switching to prevent both the relevant control busses being heaved at the same time: this could otherwise lead to some very odd things happening inside the channel signal routing. Similarly a ground follow-through lockout arrangement is used on the master function mode selection—otherwise the consequences of more than one button being pushed simultaneously would be only to select a virtually random mode, rather than get the channel electronics really upset.

Note that all the switching is to ground from the logic −5V supply rail. This interfaces with the majority of the channel logic as described in **Fig 57b** (September 1981). An important feature is the Master Reset Buss and its control. Ordinarily, a heap of random logic circuitry dependent on flip-flops and latches (of which this console is an ace example) would, on power-up, tend to settle into whatever state these registers 'felt like' at the time. This all depends on device symmetry, temperature, humidity, phase of the moon and colour of hessian you've got on the walls, and worse still is not usually repeatable. An intriguing exception to this is the knack of CMOS flip-flops to come back up in their previous state after a short power disablement—probably a function of small charge storage—but it wouldn't be wise to rest the defence of the nation on it.

Wisdom, common sense and sore ears dictate that on power-up the desk should come on neutral, all channels muted and with monitoring functions such as PFL and solo disabled. As well as providing a frame of reference from which to start re-using the console, it saves all the aggro of finding the one stupid function that's killing the monitoring. Console mode and basic monitoring conditions can be set up just by hitting the relevant master controls.

TR1 of **Fig 70** grounds the Master Reset Buss for as long as the 22$\mu$F capacitor takes to charge up—around a quarter of a second. This charging takes place when the −5V logic rail appears. Should the supply collapse, the capacitor is rapidly discharged via D1 ready to re-initialise the MRB signal as soon as power reappears.

Although it would be extremely simple to do, no top-panel master reset control is made available. Why? Because sooner or later someone's going to hit that button at exactly the wrong moment, that's why.

## Device idiosyncracies

ICs come and ICs go, and some just lie around on the floor waiting to be trodden on by bare feet. Not a long time ago, the Signetics *5534* was a rare and expensive beastie that took board meetings and ransoms to possess. During the course of this series the production-quantity cost has plummeted by the good offices of basic economics to pocket-change value—today, it's rare to find a piece of gear in our industry that's not bristling with them.

Despite this, there is no real justification for using these devices anywhere in this design other than where they have already been specified (principally the mic preamp). Their main advantage, low input noise, is of little if any value elsewhere, the noise floor being dependent rather on design criteria, the system mechanical construction and grounding than the choice of amplifiers.

The *5534* is also now multiple-sourced from different manufacturers and, it now being proved that a sizable market exists for such devices, a rash of competitive and upgraded op-amps has been provoked. Most promising, to the extent that it is worthwhile using them in these circuits instead of the specified *5534*s, are recent devices from Analog Systems (distributed by Pascall Electronics Ltd in the UK). Apart from a *5534* look-alike, called the *MA 342*, their *MA 322* measures quieter for thermal ('white') noise in the mic amp circuit (Part 12, October 1981) and has a quite significantly lower turnover frequency for its lf noise. Being relatively new, it's pricey, but then so was the *5534* . . .

On the level of more journeyman type op-amps, most major manufacturers are doing Bi-FET devices similar to the Texas *TLO* series.

Many of the circuits described rely a little on the extremely high input impedances of the Bi-FET devices and hence the very low bias currents required. Gaily stuffing in bipolars may result in generated output offset voltages which could manifest themselves in extreme instances as switch clunks and 'scrapy' pots. Also the feedback phase-leading compensation may or may not be adequate for devices other than the Bi-FETs, especially some bipolars with less than tasty internal poles. Should you be tempted to use more conventional bipolar devices, particularly in quad packages, it is also worthwhile examining their characteristics when inputs or outputs are taken above or below the supply rail potentials. If the device structure under such circumstances is unprotected and turns into a silicon-controlled-rectifier which deftly shorts the supply rails together—as a certain well-known make or two tend to—you are better off without them. Unless of course you like short, sharp bangs and bits of flying molten plastic.

A similar SCR failure mode exists within the CMOS logic family potentially exhibiting itself in transmission gates. Logic itself is ordinarily working within defined supply rails but with transmission gates such as the *4016*, *4066*, etc, there is the possibility that the audio they are switching can exceed the rails. The thing which saves them in the virtual-earth/potentiometric switching system employed here is that they are (a) fed from a reasonably high source impedance so that not much current can flow and (b) the audio level architecture is such that excessive 'breakover' levels are unlikely if not impossible.

A slightly more obscure potential 'failure' mechanism exists with the PROM used in the channel routing logic (**Fig 57b**). 'Failure' is in quotes because it would really only do what it's supposed to. The programming path for the PROM is via the output ports where once an address is selected, the required data is blasted in in short spikes which weaken or blow the approprate bits of the internal diode matrix. Any potential at the open-collector outputs in excess of about 12V opens this programming path and even short accidental over-voltages can 'soften' or even wreck the stored pattern. With only ±5V for the logic rails in **Fig 57b** the PROM is quite safe given good regulated supplies.

It would be quite unfortunate if a little amusing for the desk to reprogram itself should you be tempted to be adventurous or less than thorough with the logic power supplies.

## Meters and headroom

Metering is a subject that has been well avoided. There are plenty of proprietary meters of the popular standards and types, plus quite a few strange ones too. It's all down to personal preference and the information one hopes to glean from the assorted needles, twinkly lights and cathode rays dancing before your eyes.

Without jumping into the snake pit argument of average versus peak-reading instruments it is relevant to state that the choice will directly affect the operational levels, the level architecture, machine line-ups and various tweaks, notably the input stage limiter threshold. Out of habit, this console was designed with standard PPMs in mind, where the peak operational level throughout the system is expected to be PPM 6, or +8dBu. Line-up level, ie the system and output level for which the front-end gain stage is calibrated, is 0dBu, PPM 4. This will suit any current or expected PPMs whether to *BS 4297: 1968* specification or the new-fangled bureaucratically-doctored EBU spec.

'Proper' American broadcasters have taken quite a fancy to a mutant PPM which is similar in dynamic characteristics to *BS 4297* only with the level for the various marks elevated by 8dB, the marks being given actual level values (up to a 'max' of +16dB whereupon it's painted red) instead of the familiar 1-7. This is, it is given to be believed, so that the signal levels generated from control areas using these meters are similar to those from older areas using (curiously non-standard) +8dBm-referred VU meters. Such are the levels they are used to sending down inter-studio and telephone lines. Buzby would clutch his little feathered heart and fall off his pole, claws smouldering.

The elevated-level PPM is an idea with some merit where most material dealt with is pre-recorded and fairly predictable in level, thus not requiring an awful lot of headroom.

Users of PPMs and VUs tend to fall into the respective category types of "We'll only peak up to the 3% tape distortion point," and "Let's wind it up 'til just before it comes back sounding bent." VUs are very good for giving an idea of subjective loudness and not worrying you about transients which can often be anything up to 20dB above the indicated value.

Given standard +4dBm referred VU meters that means that under normal operational circumstances, headroom in any console is perilously skinny. Various ways of dealing with potentially inadequate headroom are in use (see **Fig 71**). A favourite is to run the entire console system at a depressed level, usually −4dB, the necessary 4dB make-up at the end being done passively by an output transformer ratio step-up. It's a bit cheesy for a couple of reasons—there is a transformer there that otherwise needn't be and as with any transformer step-up arrangement it is overly critical to termination impedance. The frequency response could suffer awfully with a heavily reactive load such as a long line.

Headroom is mostly a problem in input channels, before the channel

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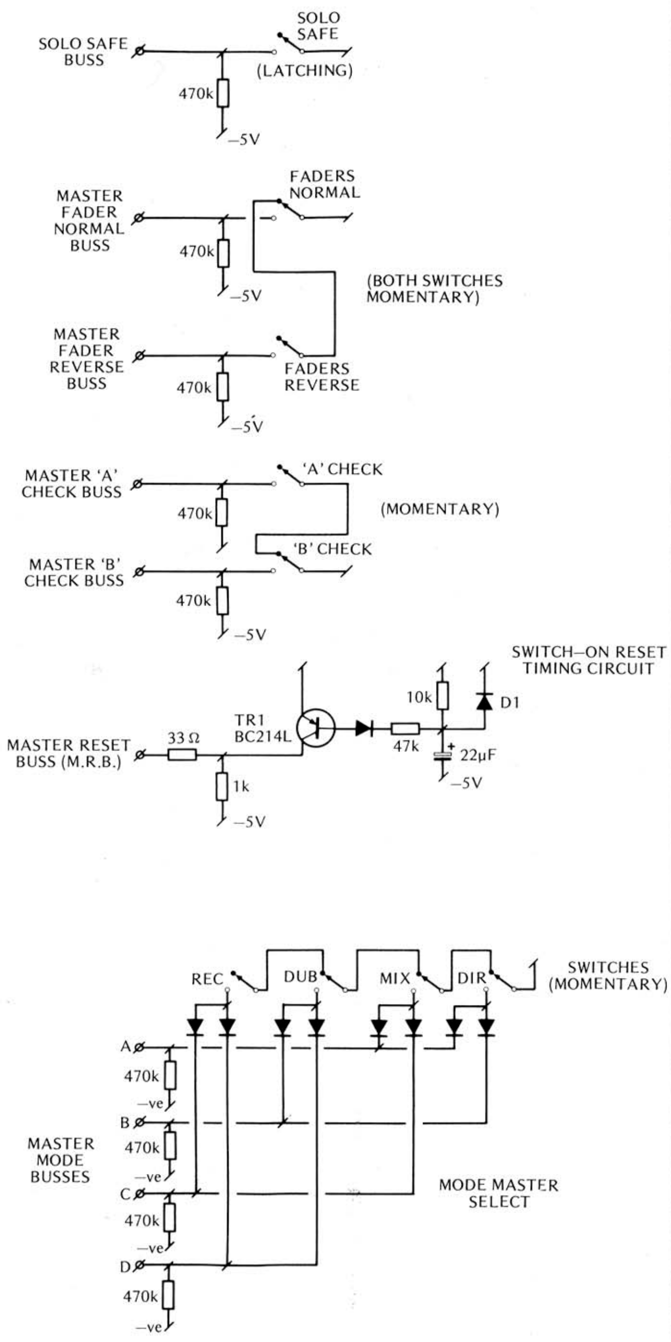
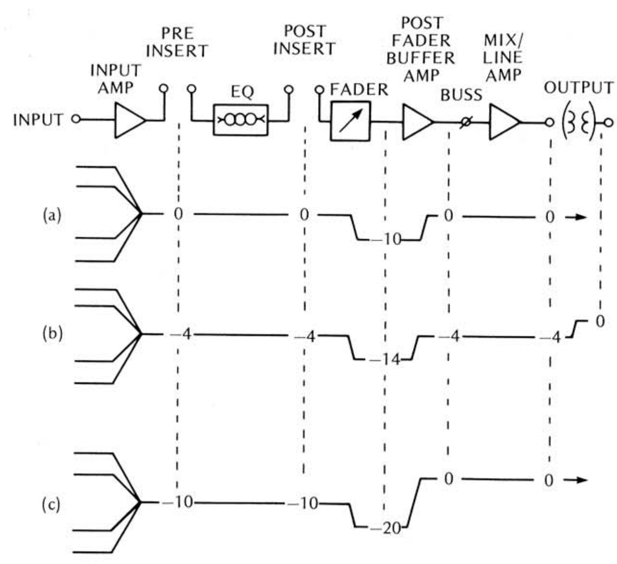
FIG. 70 DESK MASTER FUNCTION CONTROL CIRCUITRY




ALL DIODES 1N4148

FIG. 71 SYSTEM LEVEL ARCHITECTURES




(a) FLAT – 0dBu referred throughout
(b) DEPRESSED – – 4dBu throughout, level made up passively by output transformer
(c) CHANNEL DEPRESSED −10dBu in channel, 0dBu output

gain controlling element—the fader. Both ragged unpredictable input sources and equaliser gain gobble up the non-margin—hopefully beyond that point the levels and hence the mix are easily and well regulated by the faders. Dropping channel operating level by 6dB or 10dB helps matters tremendously, the gain make-up being either in the mix-amps or the post-fader buffer amps, the latter being normal. This does compromise buss-noise (quiescent desk output noise) but seeing that the main justification for doing it is the high level of signals thundering around, it's roundabout and swings time. This depressed channel system is worthwhile in any circumstance, regardless of metering type, where there is likely to be a Great Unknown lurking on the end of an input line.

Cons to the pros are that all the channel insert points operate at the depressed (say −10dBu) level which may or may not give pain with some less than versatile outboard toys, but more immediately of concern is that other internal channel circuits will need adjusting.

Machine line-in feeds from the 'A' and 'B' input differential amps will need to be dropped by 10dB—this is easily accomplished by altering the values of the resistors around electronic switches to scale down a factor of 3.16 (10dB's worth) **(Fig 72a)**. The PFL buss mix-amp gains will need to go up 10dB (the extra buss noise here is no great crime) and an extra 10dB of gain put into the pre-fader auxiliary feed buffer amps. Re-establishing main path gain to unity is simply achieved by upping the gain of the post-fader buffer amps **(Fig 72b)** by changing the feedback bottom leg resistors from 1k8 to 430Ω. This provides for 10dB of fader back-off and the necessary 10dB reinstatement.

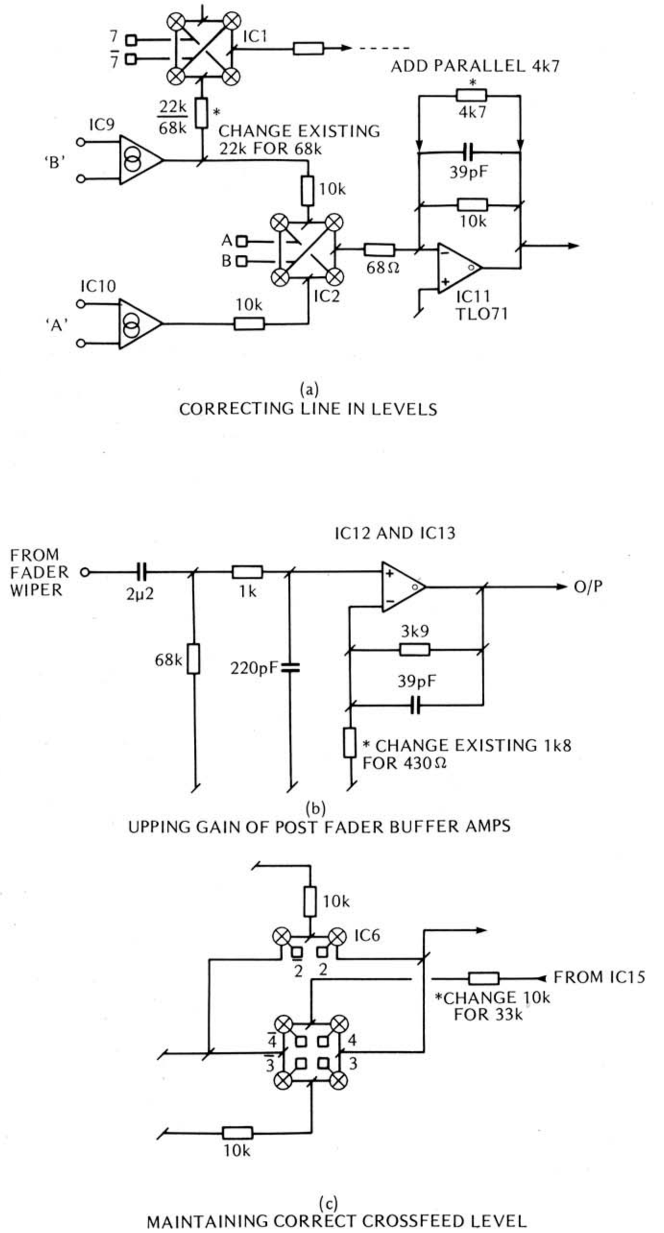
If all that sounds complicated, just bear in mind that it's all achieved with resistor changes—no surgery.

No, it doesn't matter that the machine monitor differential input amps are still operating at normal undepressed level. The 'A' check is directly monitoring a desk output which is at normal level anyway, so no headroom problem. As for the 'B' check: if you've got more level coming back from the machine than you're putting in ('A' check) then it's time for realignment and a rap on the back of the knuckles.

It is entirely possible to recalculate the values around the diff amps to drop 10dB and still maintain input balance but that would greatly increase the number of component changes necessary to alter channel system level. Which is no mean consideration should you choose to do so on a desk full of 32 channels.

FIG. 72 COMPONENT CHANGES TO OPERATE CHANNEL AT A DEPRESSED (-10dBu) LEVEL

(Refer to Fig 57(a) September 1981)




## Design by computer?

There is a whole breed of design engineers, who were nurtured and blossomed before the great pocket calculator revolution, for whom active-filter and equaliser design became an intuitive art, not totally unrelated to the master gardener's 'green thumb'. The innovations of calculators and subsequently micro-based computers somehow took the fun out of it all, to be replaced only by the zealous determination that your filter is point nothing, nothing, nought Hz inaccurate, gleaned from a neatly tabulated column of figures or a graph on a computer VDU. That is, until you try to actually *make* it with standard value 5% resistors and 10% capacitors . . . you end up reasonably close anyway, but no closer than a few years ago when a cautious squint and head-scratch were the customary design aids.

Sure, the mathematics of the networks were as well known then as they are now—they were just as unwieldy too.

Pencil and papering the 3dB-down frequency of a multiorder (the most basic formula) is thoroughly tedious arithmetic and doesn't *really* tell you what you want to know anyway, which is when the response departs from flat (say ±½dB). Far simpler to throw it together and juggle bits 'til it works as hoped. Euphemistically described as 'empirically determining values' . . .

Computers have turned the heavyweight and ponderous sums of filter design, previously at all pains avoided, into an intellectual game. Once the filter maths has been written in a digitally digestible form, the machine's great strength—iterative calculation—takes over, either plotting on a screen graph or tabulating a given filter's input/output amplitude transfer characteristic, output phase and input admittance for each of the standard ISO ⅓-octave frequencies in the audio band extrapolated also from 10Hz to 100kHz. The programs written cover variously in one form or another, single and double order filters, gyrators and loop filters.

Originally intended as an exercise in small-computer-programming, the ragged trail of sleepless, tireless nights, cups of coffee and exasperated domestic companions suffered for these programs was all worth it for a design tool that has since proved useful to the extreme of indispensability.

Again, as an indication of the weight of number-trundling involved in a second order filter calculation, the humble micro takes about half-a-second (running in BASIC) to do a single frequency plot. (A simple machine language addition by comparison takes about 2$\mu$s.) All the Sallen & Key compound filters (eg front-end, highpass filter as discussed in Part 12, and the blanket bandpass line amp in Part 2) had their values determined using these programs as did much of the eq circuitry.

Understanding the impracticality of finding a 34.162kΩ resistor, all the values are rounded out to their nearest standard value (*E24 Series* in the case of resistors). A basic premise is that the capacitors used in the filters are correct at their standard values and that resistors are changed to suit the filter shape. Weird resistors are far easier to find than weird capacitor values.

## Raw components

In case the idea of rounding off values to the nearest standard value sticks in your throat somewhat, a quick rationalisation is in order. In hard, solid practical terms borne out over years of measurement, component tolerances of 5% on carbon film resistors and 10% on *good quality* polyester and mylar capacitors are usually reliable and quite pessimistic. A good rule of thumb is that networks created from these can usually be relied upon to be ±0.2dB (for resistors in a purely gain-determining context) and ±5% or within 0.1 octave in frequency. Since most of the frequency determination circuitry is continuously variable anyway these tolerances end up being drowned in

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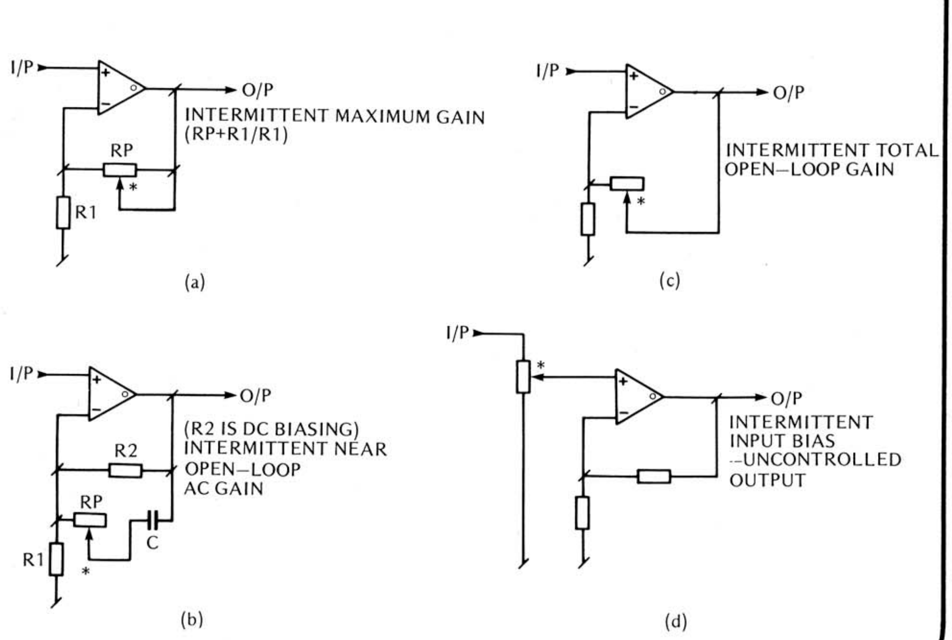
potentiometer, knob and graticule line-up inadequacies.

For any given manufacturer's batch of components, particularly capacitors, the values all tend to 'lean' one way within the tolerance, hence making consistency between modules quite reasonable.

Those who want to 'know' everything is consistent beyond the technically and operationally very acceptable gain and frequency tolerances quoted above can run to the cost of 1% resistors and capacitors or maybe even tighter . . . in the case of fixed-step switched networks whose primary virtue is consistency and repeatability close tolerances would be necessary, since they would then be forming part of not so much an equaliser but more a calibrated frequency/amplitude selective amplifier. A distinction.

Even so, long 'daisy-chain' resistor networks used as switched elements in such animals can quite happily be constructed from our old friend 5% resistors with astonishingly accurate intervals leaving only the capacitors to be highly spec-ed or selected.

FIG. 73 SKELETON CIRCUITS SHOWING EFFECTS OF POTENTIOMETER WIPER FAILURES




## Dodgy pots

Potentiometers (pun totally intended) tend to be laws unto themselves with several vices, the principal nasties including:

- wildly varying total track resistance;
- law inconsistency between supposedly identical pots;
- non-zero end stop resistance—disastrous in a panpot particularly;
- non-monotonicity—meaning the resistance versus rotation does not change smoothly up or down, but lurches in rapid up and down steps whilst in transition either way. Caused by a rough carbon surface (which gets worse with age), this one wrecks any notions of resetability!
- intermittent wiper lifting off track.

Any one of these, of course, makes a complete nonsense of careful, if not neurotically precise, fixed component value calculations for gain-sets and equalisation—even few of the best potentiometers affordable for console use really warrant surrounding with better than 5% resistors and 10% capacitors.

Generalisations are odious and some of the newer series of conductive plastic pots as exemplified by the Bourns *80* and *90 Series* are a delight to design with and use, displaying none of the above vices to an extent that could be irritating.

A particularly horrid wrinkle with many cheap and not-so-cheap pots is a predisposition with age for the wiper to break free of the track, over a chunk of debris. Dependent on the circuit context, a variety of loud effects can result. In any arrangement where the pot is included in a feedback loop, or worse still is also part of the dc biasing loop of the amplifier it is controlling, a wiper break will cause:

- the amplifier to operate at a much greater gain (defined by track resistance in conjunction with R1 in **Fig 73a**);
- go open loop at audio frequencies (nearly infinite gain)—**Fig 73b;**
- grandaddy of the lot, go open loop at dc as well, causing the amplifier output to fly to one or other of the supply rails—**Fig 73c.**

None of these are particularly quiet!

In the context of many complex active filters, a lifting wiper can not only create any of the above effects, but also the reactances forming the filter turn miraculously into timing elements creating one hefty great relaxation oscillator, sending plus elephant dB of something usually nastily ultrasonic screaming through the desk, steaming power amplifiers, gently smouldering monitor speakers and your ears. For a short while.

It suddenly becomes a very expensive pot.

## Where to from here?

Life is a continual story of change, he waxed lyrical. Despite a faltering last few years, recession and excruciating-to-the-point-of-admirable mismanagement of the media that like it or not support us, our industry is still alive and changing. Bearing in mind that this mixer and series were mooted and first development undertaken two years ago, it is gratifying to see that it is all still current and relevant.

The writing, or rather the digits, are on the wall and it is quite certain that signal processing technology will end up inextricably intertwined with high-speed mainframe computer systems, via an intervening period, downing now, of digital control of analogue electronics. The emphasis is rightly shifting to the control console being *just that* rather than a box full of electronics—which will find its more natural home in a rack elsewhere.

What you see on the market around us now, this mixer design being part of that family, are the *Last of the Great Analogue Mixers* which will be as fondly remembered in years to come as *AXBT* microphones, valves, Michael Miles and round fader knobs.

A tear will drop from your eye as you take your grandkids to the Science Museum.

It's a little worrying to think that those responsible for mixer design in the mid-future will not be electronics engineers with a pair of ears and a knack of translating the languages of sound and electrons, but will be those mathematicians and programmers sparkling enough to translate frequency domain and temporal characteristics into seemingly alien and brain-freezing digital algorithms. OK, maybe it isn't worrying—except for electronics engineers—it is merely exchanging one set of technologists for another. A less obvious danger is that no longer will a console signal path be intuitively simple to grasp and even follow in tangible bits of circuit—digital signal processing is and will remain arcane and very much a 'black box' activity.

Assuming the music industry as we know it still exists, engineer and producer mentality will have to become more 'lady-driverish' and totally user polarised, those who will be most successful being those who possess or can afford the best or most elegant or best sounding mainframe operating system software. Hardware—the equipment—will become almost incidental. Chastening! ■

**Subject to demand, it is our intention to produce a booklet containing reprints of all the parts of Steve Dove's series. We expect such a booklet to cost around £1.50 to £2.00, depending on the number of copies we need to produce. This booklet will be offered prior to publication by means of a form in the magazine, but we need to know in advance how many copies will be required, so that it can be produced in the most economical way. Would readers who are interested therefore let us know *in writing* as soon as possible if they will require copies. *Do not send any money at this time.* The booklet will be produced if and when there is sufficient interest to make publication of a reprint substantially less expensive than a bunch of photocopies.**

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