PERFORMANCE LIMITS IN CONTEMPORARY CONSOLE DESIGN

by John Roberts

If we take a look at how consoles have evolved over the years, we will observe two trends, not only has the console grown in input/output capacity in response to popular tape machines, but control flexibility has increased even faster.

A single input strip today has more electronic circuitry than an entire console of 15 years ago and the trend is not slowing. Microprocessors have started popping up in consoles, combined with automation to give console memory and simple computational powers. I will not attempt to predict where all this will lead, but if a console ever acquires taste we're all in trouble!

This unchecked demand for more capacity and complexity, has forced the circuit designer to become more reliant on miniature integrated circuits to meet reasonable package size and power requirements. To better understand the realizable signal quality capability of a modern console we must look at the basic IC op-amp.

Operational Amplifiers

In the last ten years the operational amplifier has progressed from a laboratory curiosity to a low cost building block, capable of outstanding performance and incredible circuit densities. Generous application of negative feedback effects, ruler flat frequency response, and barely measurable distortions.

How They Work

To better understand integrated circuit op-amp performance in different configurations, let's first look at their open-loop gain characteristics. Think of the op-amp as a black box with a floating differential input and single-ended output.

Since a typical high performance IC op-amp can have open-loop gains exceeding 100 dB (100.000:1), the output will swing rail-to-rail for differential inputs as small as .0003 volts peak-to-peak. It's immediately obvious that they were not designed to be used this way as the same op-amps have typical input offset voltages ten times that!

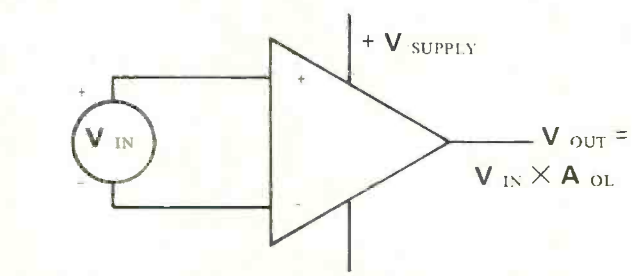

Figure 1: Floating Differential Input/Single Ended Output

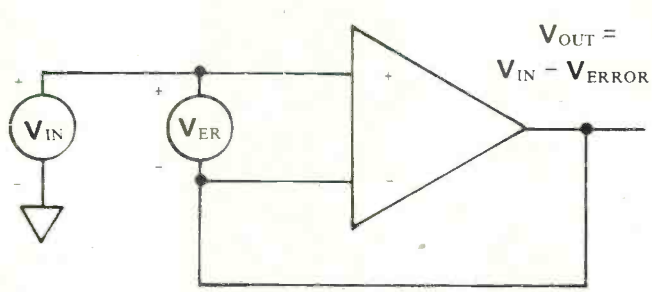

Figure 2: Op-Amp Using Negative Feedback

— the author —

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Enter Negative Feedback

Connecting the op-amps output directly back to its minus input dramatically changes this circuit's character. (You may recognize the standard unity gain follower.) One-hundred-per-cent negative feedback causes the output to very closely follow the input. If we now drive the output rail-to-rail we notice something rather curious. The same .0003 volts peak-to-peak shows up at the input as a differential error voltage. In this unity gain mode the error voltage subtracts directly from the input voltage causing a very small gain error (.001%). While it may seem unreasonable to worry about by an error voltage 100 dB below the output, let's see what happens in other gain configurations.

Figure 3 shows an op-amp configured as a non-inverting gain stage. As before, we can assume the negative feedback will force the minus input to follow the plus input. Knowing that these inputs offer high impedances to the feedback network, we can use Ohms Law to calculate the current flowing through R_G:

$$(V_{IN} - V_{ER})/R_G = I_{RG}$$

Since this current must be supplied by the output through R_F we can again use Ohms Law to calculate the voltage drop across R_F and thus the output voltage:

Since

$$I_{RG} = I_{RF}$$

$$V_{RF} = R_F\,(I_{RG})$$

$$V_O = V_{IN} - V_{ER} + V_{RF}$$

$$V_O = (V_I - V_{ER})\,(1 + R_F/R_G) \qquad \text{[EQ.1]}$$

From inspection of this equation we can make two useful observations; 1) the gain of a non-inverting amp can be predicted by: $G = 1 + (R_F/R_G)$ [EQ.1A], 2) the error voltage is referenced to the input and amplified along with the signal.

Now let's look at another popular configuration (inverting amp/summer).

In this case the input signal is connected to the minus input through R_G. Once again the negative feedback will force the minus input to follow the plus input which is connected to ground. The minus input becomes a virtual ground, and we can again use Ohms Law to predict the output voltage. Since the

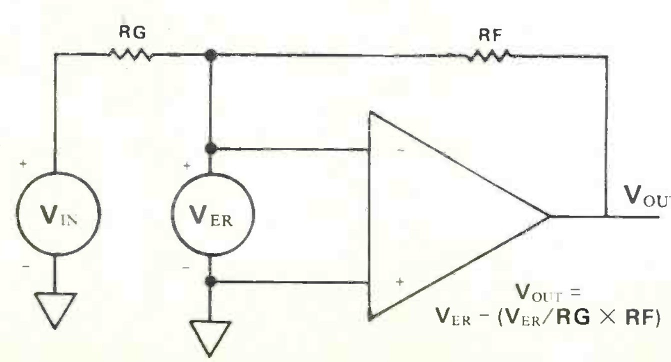
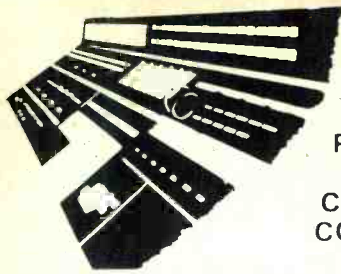


Figure 3: Non-Inverting Gain Stage



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minus input will be held to within the error voltage of ground we can define the current in R_G as:

$$I_{RG} = (V_{IN} - V_{ER})/R_G$$

and again:

$$I_{RG} = I_{RF}$$

therefore:

$$V_O = V_{ER} - (V_{IN} - V_{ER})\,(R_F/R_G) \qquad \text{[EQ.2]}$$

From this we can derive the general inverting gain equation:

$$V_O = -V_{IN}\,(R_F/R_G) \qquad \text{[EQ.2A]}$$

However, the error voltage in this case is amplified by 1 + (R_F/R_G). You will recognize this as the gain equation for the non-inverting configuration [EQ.1A]. This observation allows us to simplify analysis of input error effects by always considering them in series with the plus input. A very useful characteristic of the virtual ground is it's ability to sum together multiple inputs at different gains with no interaction.

V out now becomes:

$$V_O = [V_{IN1}/_{RG1} + V_{IN2}/_{RG2} + V_{IN3}/_{RG3}]\,[-R_F] \qquad \text{[EQ.3]}$$

To properly analyze the error voltage contribution we assume resistors R_{G1} through R_{G3} are connected to ground. The gain seen by the error voltage is:

$$G_{EQ} = V_{ER}\,(1 + R_F)/(1/R_{G1} + 1/R_{G2} + 1/R_{G3}) \qquad \text{[EQ.4]}$$

The disturbing thing to notice here is that the error voltage is amplified by one plus the sum of all the gains. This error gain can rapidly become significant in the case of a recording console where 40 to 50 signals can be summed together at one time.

The Real World

Until now we have been considering the performance of ideal op-amps. Let's take a look at a real integrated circuit op-amp. At very high frequencies (1 - 10 MHz) inter-stage delays and phase shift add to where the output is lagging the input by 180°. Normally one would not take phase response up with the radio frequencies as being a very significant performance factor, but look at what happens when that phase-shifted output is applied to the minus input. That wonderful negative feedback suddenly becomes positive feedback, exciting tweeter smoking oscillations. To insure stability, the product of open loop gain and feedback factor must be kept less than unity for frequencies where the time phase shift approach 180°. Contrary to what you might guess the worst case for stability is the unity gain follower (Figure 2). In all the other gain configurations the feedback is attenuated by the ratio $R_G/(R_F + R_G)$ [EQ.5]. This feedback factor improves stability, since we do not often encounter feedback factors of −100 dB (a closed loop gain of 100.000:1) we must roll-off the op-amps open loop gain. The most commonly used technique is to integrate a capacitor right onto the chip across an intermediate gain stage.

Since this compensation cap reduces the slew rate as well as the usable open loop gain, some amplifiers do not compensate for unity gain stability allowing the user to optimize the compensation for a given feedback factor and desired margin of stability. A TL074 is unity gain stable while still providing 13 μv/sec. A NE5534 is only stable at gains above 10 dB, adding the 22PF external compensation capacitor required for unity gain stability drops the slew rate drops from 12 v/μsec. to 7 v/μsec.

More Error Voltage

So far we have only looked at the error voltage in terms of open loop gain. To properly assess its significance we must look more closely. The error voltage is a catch-all term composed of all the ways op-amps vary from the ideal. This error voltage contains a noise term, a distortion term, a DC offset term, and a loop-gain error term.

Input Noise

Since the op-amps input noise term interacts with the source and feedback network impedances, it is usually specifed as a voltage term and a current term. The total input noise being the sum of the noise voltage plus the noise current times the

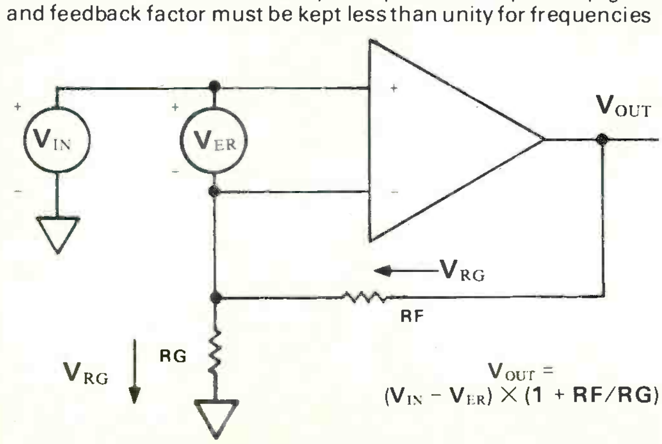


Figure 4: **Inverting Gain Stage**

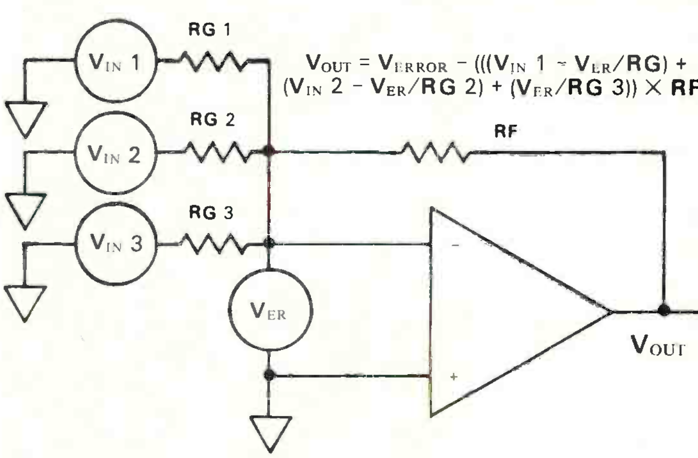


Figure 5: **Inverting Amp Summer**

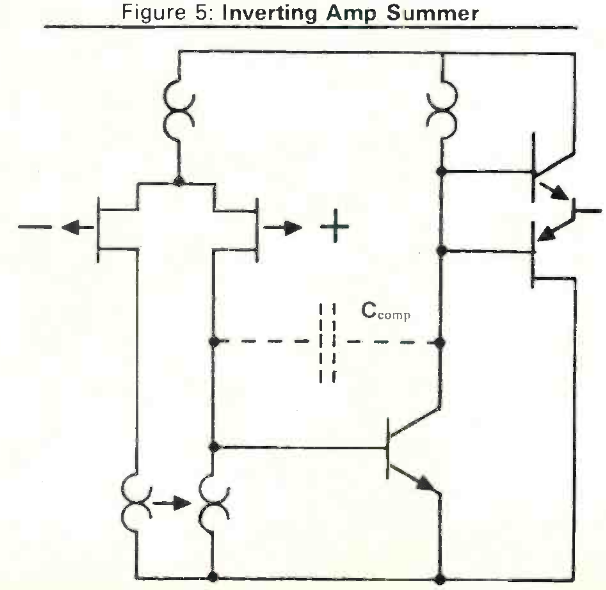
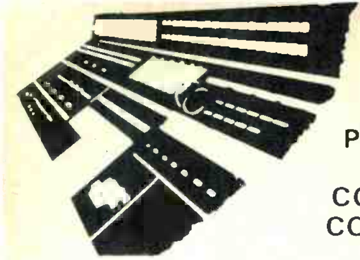


Figure 6: **Typical Op-Amp Internal Schematic**



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source/feedback network impedance, plus the thermal noise (Johnson) of the source/feedback network. This total input noise voltage is referred to the input and amplified along with the signal. As shown by the earlier example (Figure 5), the noise is amplified by the gain seen by the plus input. This gain is often referred to as the noise gain [EQ.1A.4].

DC Offset

Like the noise term, DC offset also interacts with the source/feedback network impedances and is specified in terms of a voltage term and a current term. The offset also contains a bias current term which must be applied to the difference between the impedances present at the plus and minus inputs. This term is referenced to the input and amplified by the noise gain.

AC Error Voltage

The amplitude of the AC error voltage can be calculated directly from the open loop gain plot (Figure 7). Above the compensation pole, the AC error voltage leads the output by 90° and rises 6 dB per octave. Like the other error terms this is also referenced to the input and amplified by the noise gain.

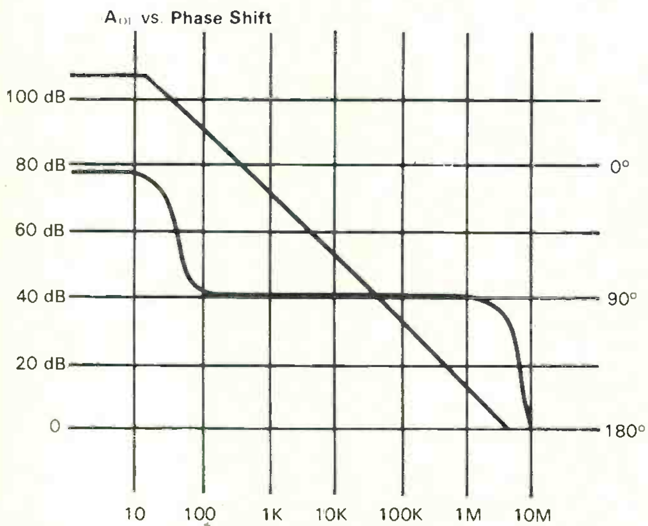


Figure 7: A_{OL} **vs. Phase Shift**

Distortion

The dominant source of distortion in op-amps is the increasing non-linearity of the input differential stages transconductance with amplitude. Since the amplitude seen by this input stage is the inverse of the open loop gain there is a significantly rising distortion versus frequency term. Again, this error is referenced to the input and amplified by the noise gain.

Typical Values

To get a better feeling for the significance of these errors, let's look at some typical values. These numbers are generalized; the exact values will depend on source/feedback network impedances and device selection.

Bi-FET (J-FET Input TL074)	
Offset Voltage	3 mV
Offset Current	5 ρA
Bias Current	30 ρA
Noise Voltage	18 ηV/$\sqrt{Hz}$
Noise Current	0.01 ρA/$\sqrt{Hz}$
Typical Noise	2.5-3 μV (-110 dBv)
A_{OL} @ 100 Hz	105 dB
A_{OL} @ 10 kHz	55 dB

High Performance Bi-Polar (NE5534)	
Offset Voltage	0.5 mV
Offset Current	20 ηA
Bias Current	500 ηA
Noise Voltage	4.5 ηV/$\sqrt{Hz}$
Noise Current	0.5 ρA/$\sqrt{Hz}$
Typical Noise	0.5-1 μV (-120 dBv)
A_{OL} @ 100 Hz	100 dB
A_{OL} @ 10 kHz	70 dB

Distortion

It is very difficult, if not impossible, to come up with some typical specs for distortion. It is further confused by typical measurement techniques. As you will recall from our discussion of the gain error term, it is the output voltage times the inverse of the open loop gain. Since the compensation cap forms an integrator above the pole frequency the error voltage is the inverse, or time derivative, of the output divided by the input stage transconductance. (For those of you whose calculus is as rusty as mine, the time derivative is simply the slope or dV/dt of a waveform. We can assume the input stage to be linear for small signals (frequency << slew limiting).

The problem occurs when we attempt to measure distortion with a sine wave. The slope or derivative of a sine wave is a cosine wave. Since a cosine wave also happens to be a sine wave phase shifted 90°, these two sine waves combine to form a new sine wave slightly reduced in amplitude and phase shifted from the original. Your megadollar distortion analyzer doesn't even see that error voltage. Right now some of you are saying, "If it cancels out, so what!" Well, let's take another example: this time instead of a slippery old sine wave let's look at a triangle wave. This time the slope error voltage is a square wave. Now there's no way you're going to add a square wave to a triangle wave and get anything but a distorted triangle wave! I am encouraged by the amount of interest in other than single tone sinewave testing (sine + sine and sine + square), as demonstration of a dissatisfaction with the status quo. The major difference in the new tests is the presence of more high frequency content forcing the input stage into its non-linear regions.

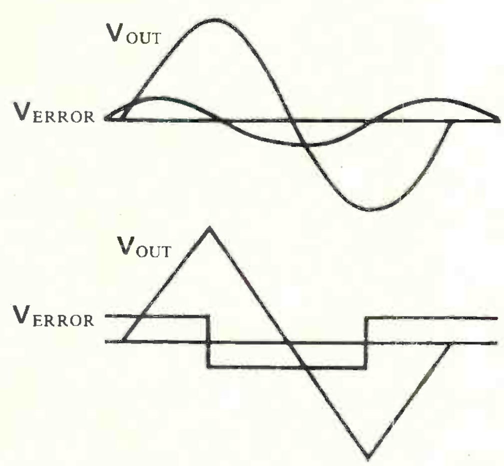


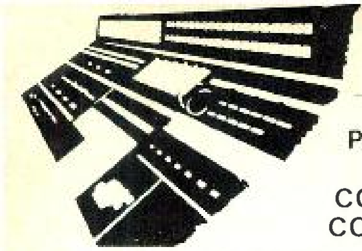
Figure 8: V_{OUT} **as it relates to** V_{ERROR}

The distortion that does turn up in the traditional measurement is the distortion on the error voltage reduced by its relative amplitude. A popular equation to describe this distortion is:

$$D_{CL} = D_{OL}\,(A_{CL}/A_{OL}) \qquad \text{[EQ.6]}$$

To demonstrate the ability of negative feedback to reduce distortion, imagine either one of the two example op-amps operating at a closed loop gain of 20 dB. Even 100% open loop distortion would not approach 1% at the output until above 20 kHz. Since typical open loop harmonic distortion is clearly orders of magnitude lower, this measurement usually returns statistically small numbers below 20 kHz.

However, all of these errors are referenced to the input and will be increased by the noise gain. Amplify any of these terms



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by the 60 dB typical of a mike pre-amp and they will get significant. A large console's summing amp with 50 inputs is operating at a noise gain of 34 dB. In conclusion these op-amps are very good but not perfect. When used at low noise gains (20 dB) they should provide excellent performance.

Slew Rate

You cannot easily predict amplifier slew rate requirements from manufacturers spec sheets and the accepted sine wave frequency response of a human ear. Both op-amps chosen for this discussion are capable of slew rates on the order of 12 - 13 V/μsec., or almost ten times the slew rate so predicted. This slew rate is not as excessive as first appears. Manufacturers often spec maximum slew rates with the input stage saturated or over-driven. Above input saturation the output is no longer able to follow the input and the negative feedback is not working its magic. Even more importantly, the input stage has a rapidly increasing non-linearity as you approach this saturation point. For reasonable control of distortion it is desirable to afford as much margin as is practical. Eating away at this margin from the other direction is the fact that in live music situations you will encounter slew rates corresponding to sine wave frequencies well in excess of the 17 - 20 kHz considered audible. Without debating the audibility of these higher frequencies they are capable of overloading slower input stages and causing intermodulations. When in doubt choose the higher slew rate amplifier as it usually predcits a larger region of linear operation.

Now To The Console

These advances in technology have for the most part made console design a little easier. The op-amps' near ideal performance in low gain applications, simplifies most of the routine processing and routing. Needless to say, the rest of the industry has not been standing idle. Advances in media (metal tape) and recording technology (analog — Stephens/digital — 3M, Sony/noise reduction — Dolby, dbx, etc.) promise dynamic ranges on the order of 90 dB. For the console to remain transparent to such program we must better that 90 dB by some margin. Further incentive to maintain these dynamics is provided by recent developments in consumer playback equipment. While the question of how much dynamic range is usable or even desirable deserves consideration, it is better decided by the producer on an artistic basis whenever possible.

As you will recall from our earlier dicussion of feedback networks, op-amps' noise and distortion can be referenced to their inputs for analysis. The high signal gains required by microphone pre-amps rule out single op-amp approaches. The other problem area in console design is the master summing busses. Although each signal may be mixed in with a gain of unity, the input referred noise and distortion adds to each one being effectively multiplied by the number of inputs. Since it is not uncommon for a console to mix as many as 50 inputs, the single op-amp approach is again inadequate.

Microphone Preamplifiers

Until recently the state-of-the-art in microphone pre-amps consisted of a transformer gain stage followed by a traditional op-amp or discrete gain stage. The transformers almost noiseless gain and impedance matching ability was able to scale the microphone's 100 - 200 ohm source impedance and -60 dBv signal up to levels that could be handled by subsequent stages. Properly executed a transformer pre-amp design is capable of a 3 dB noise figure.

Being within 3 dB of a 200 ohm resistors' Johnson (thermal) noise is respectable but the Trans-Amp™ improves upon even that. The Trans-Amp combines a discrete transistor gain stage coupled to a unity gain stable op-amp. Negative feedback is brought from the output of the op-amp to the emitter of the transistor gain stage.

The closed loop gain of the circuit in Figure 10 is $1 + (R_F/R_G)$. The open loop gain becomes:

$$G_{OL} = A_{OL} + R_{OL}\,(1/R_G + 1/R_F)$$

If we vary R_G to change our closed loop gain, the open loop gain changes also since R_G is common to both equations. This topology is capable of very large $\pm$160 dB open loop gains since the feedback factor equal R_{OL}/R_F the amplifier is stable at all gain settings. The input noise is dominated by the transistor stage which is optimized for the low source impedance. The Trans-Amp will deliver distortion performance approaching the unity gain configured op-amp and noise performance within 1 dB of the source impedance.

The Summing Amp

It is not as easy to define what was the state-of-the-art in summing amp design. So as not to offend anybody I'll describe several popular approaches.

The first approach involves designing a dedicated op-amp. A discrete op-amp can be designed with lower input noise and better high frequency phase characteristics than an integrated circuit. Optimizing the compensation capacitor for the improved phase margin and known feedback factor can result in substantial increases in usable open loop gain. Therefore, lower error voltage and distortion.

To simplify analysis of the noise and distortion performance of a summing op-amp let's consider all the input resistors terminated to ground and the noise/distortion in series with the plus input.

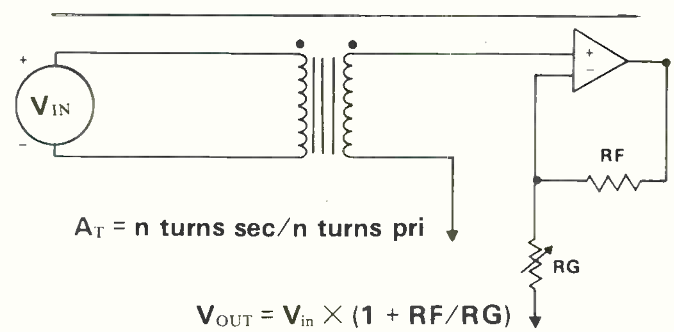

Figure 9: **Op-Amp With Tranformer Input**

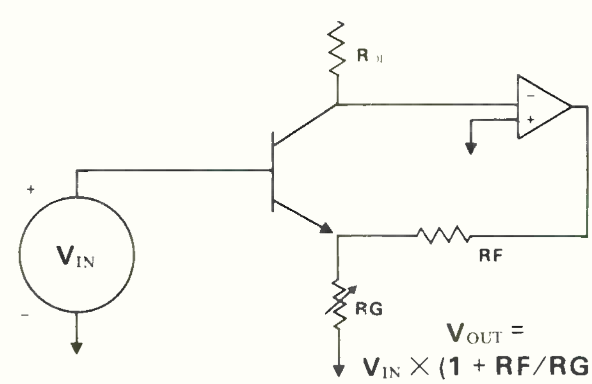

Figure 10: **Trans-Amp Example**

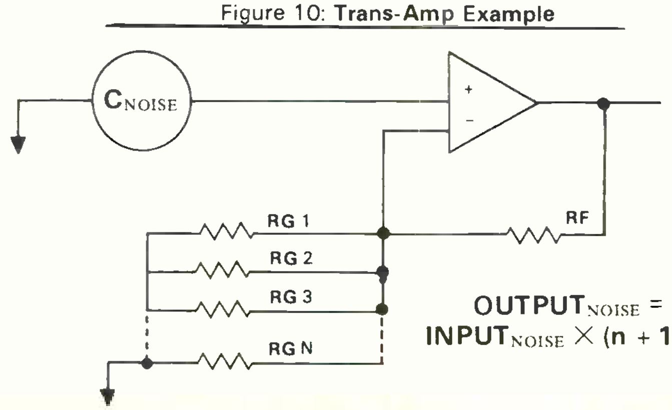

Figure 11: **Dedicated Op-Amp Summing Design**

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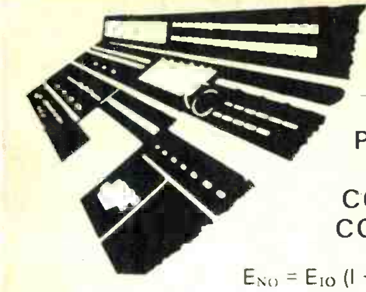

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$$E_{NO} = E_{IO}\,(I + R_F/(1/R1 + 1/R2 + 1/RN))$$

$$E_{NO} = E_{IO}\,(N + 1)$$

$$D_{CL} = D_{OL}\,(A_{CL}/A_{OL})$$

If we take the example of a 48 input mix buss, the noise and distortion is being amplified by 49 times (34 dB).

A second popular approach is to break up the buss into 2 or 4 sub-busses, and then re-combining those outputs. This approach takes advantage of the non-coherent addition of each sub-busses' noise and distortion.

Output noise =

$$E_{on} = \sqrt{[4\,(13en)^2 + 5\,(en)^2]}$$

$$E_{on} = \sqrt{[676(en)^2 + 25(en)^2]}$$

$$E_{on} = \sqrt{701(en)^2}$$

$$E_{on} = 26.5(en)$$

The law of diminishing returns takes over rapidly as the four sub-buss approach only buys you a (5.3 dB) improvement over the single op-amp approach. Approaches one and two can be combined for an added noise improvement. The distortion improvement would probably wash since you have to increase the compensation caps to make up for the lower sub-buss feedback factors.

Once again the Trans-Amp topology has merit. Grounding the input and connecting the buss directly to the emitter of the input

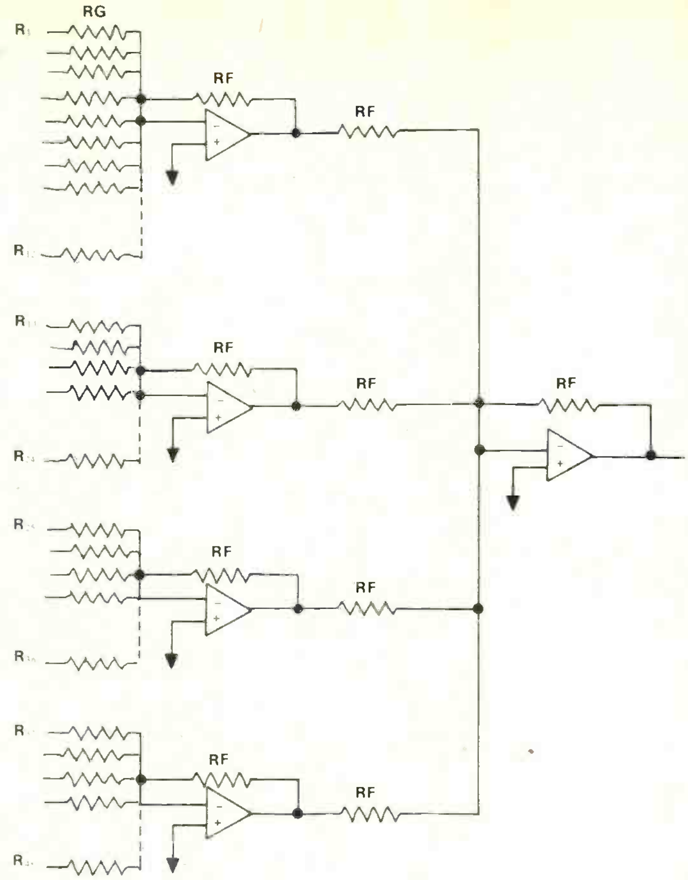


Figure 12: **Multiple Buss Summing Design**

gain stage reaps the same open loop to closed loop gain tracking with the possible qualification that the Trans-Amp will not excell when a small number of inputs are being summed. The Trans-Amp will outperform the previous examples for large buss structures.

The best solution to the summing buss problem sounds almost too simple to be true.

If we replace the summing resistors with current sources, the N term completely drops out of the noise/distortion equations. (Assuming the practical current sources have relatively high output impedances.)

Output noise =

$$E_{ON}\,(1 + RF/\infty) = (1 + 0)$$

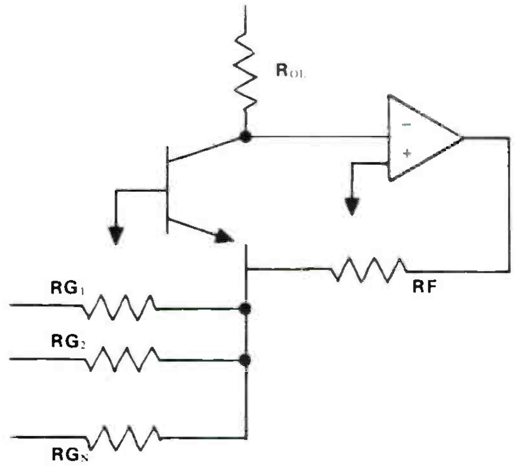
$$E_{ON} = E_{IN}\,(1)$$

The complete absence of an N term in the noise gain equation does feed the imagination. Hundreds of signals can be summed together with no measurable degradation in quality.

As if the advantages of a noiseless summing amp aren't enough, there are other practical benefits. In the traditional resistive summing amp, minute differences in ground potential (like hum and buzz) are amplified by the full noise gain of the summing amp. You quickly find out how clean your signal ground is when you throw 34 dB of gain on it! Another benefit occurs when the current sources are switched electronically. Non-linearitys in device on-resistance have no effect on the signal current being passed.

Since we are already dealing with currents the buss can be connected directly to the input of a current ratioing VCA (such as an Allison EGC-101) for improved buss headroom and control flexibility.

Not unlike archeology it is interesting to examine what lies below this freshly removed layer of noise/distortion. The next level or noise floor merits discussion as it describes the limit of what a console can deliver. Let's consider the summing amp to be noiseless. Suppose we choose to sum together 48 channels of program. We can assume their noise to be incoherent and add as the sum of the squares. Therefore:



Figuer 13: **Trans-Amp Summing Design**

Output noise =

$$E_{ON} = \sqrt{48(en)^2}$$

$$E_{ON} = 6.9(en)$$

$$E_{ON}\,(dB) = 20\log_{10}/E_{ON}/+ 16.8$$

To see what that means in actual noise terms, suppose those 48 channels were all dbx or digital at fader positions such that the channel noise in each channel is −90 dBv. Adding 48 of these together in a perfect summing amp gives us an output noise floor of −73 dBv. We are still getting better than 90 dB dynamic range, but not by a whole lot.

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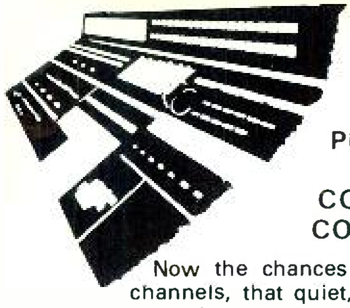

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Now the chances of your ever encountering that many channels, that quiet, are not very likely. But let's carry this example one step further. Suppose all 48 channels are muted or noise-gated off. (A possible situation with today's automated consoles.) The buss output noise is now dominated by the residual noise of the current sources. Since the practical implimentation of a usable voltage-to-current convertor (current source) contributes noise on the order of −105 dBv. Forty-eight of these add to become −88 dBv at the buss output. The VCAs (Allison EGC101) unity gain noise contribution of −88 dBv further degrades the output noise another 3 dB to −85 dBv. If we then make the buss output driver capable of a +27 dBv output (Note: This requires discrete components as most op-amps will not drive 600 ohms to 50 volts peak-to-peak.) the buss output noise floor is +27 dBv − (−85 dB) or −112 dB below clipping! Because of this ideal summation, every time you double the number of inputs the output noise only rises 3 dB. One-hundred channels could be summed together with a dynamic range of 109 dB. Two-hundred inputs with 106 dB, etc.! I feel comfortable calling this buss structure noiseless, or at least transparent to 90 dB dynamic range programs.

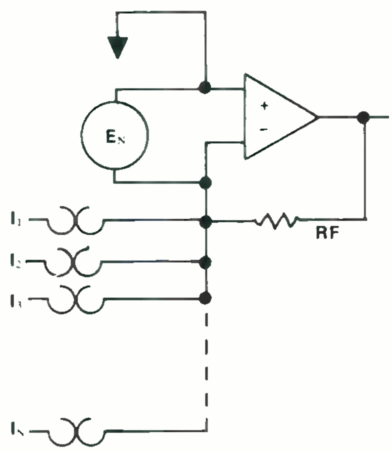

Figure 14: **Current Source Summing Design**

If we define the best dynamic range any one channel of program can deliver as: Analog with metal tape (Stephens), 80 - 90 dB. Digital (Sony), 90 - 93 dB.

Any one channel mixed in at unity gain will dominate, and this doesn't even consider the fact that you first have to get 80 - 90 dB out of your room. Easier said than done.

As good as this new hardware is, all it takes is one sloppy track to sabotage this effort. The engineer/producer will ultimately determine the dynamic range of the finished product.

Without getting too far away from this discussion of hardware, I would like to offer a few suggestions:

- 1. Use only as many tracks as you need. Just because you're paying for all 45 don't feel compelled to use them, or if you must, you can leave them turned off during mixdown.
- 2. If you can afford the extra tracks, double up your lead vocals or other noise critical tracks. (Putting the same signal on 2 tracks improves signal-to-noise 3 dB. On 4 tracks, 6 dB.)
- 3. If you are confident that you won't want to change the equalization later, apply all mid-range boost when laying down the track. Cut during mixdown. (Note: When in doubt don't do it. Even the best symmetrical equalizer can not always be corrected.)
- 4. If you have automation, noise gates, or even lots of hands, keep all secondary tracks down when not contributing to the mix.
- 5. Use compression sparingly and only on the track(s) that need it.
- 6. Effects mix levels should be set at the return and not the send. This way the effect is always operating at its best signal-to-noise ratio.
- 7. If you're making a disco version of the "1812 Overture," disregard 1 through 6.

While these suggestions will help you maintain dynamic range, they are not inviolate and should only be followed when they don't interfere with artistic considerations.

Meters

The general availability of high resolution bar-graph displays have improved the readability of the typical console's meters. It's a lot easier to scan across a line than focus on 32 jumping needles. Well, that's obvious. However, there still seems to be a certain amount of schizophrenia in the industry between PPM (peak program) and VU (average responding). In brilliant attempts to please most of the people most of the time, console meter packages vary from switched, to side-by-side, to switched side-by-side. Well, guys, there is a better way. Drawing upon the definition of what peak and VU actually represent you can make some very useful assumptions. Not only will peak rarely fall below VU, but you can rely on it to provide a useable margin (3 dB in a sine wave). Displaying the VU as a connected bar and the

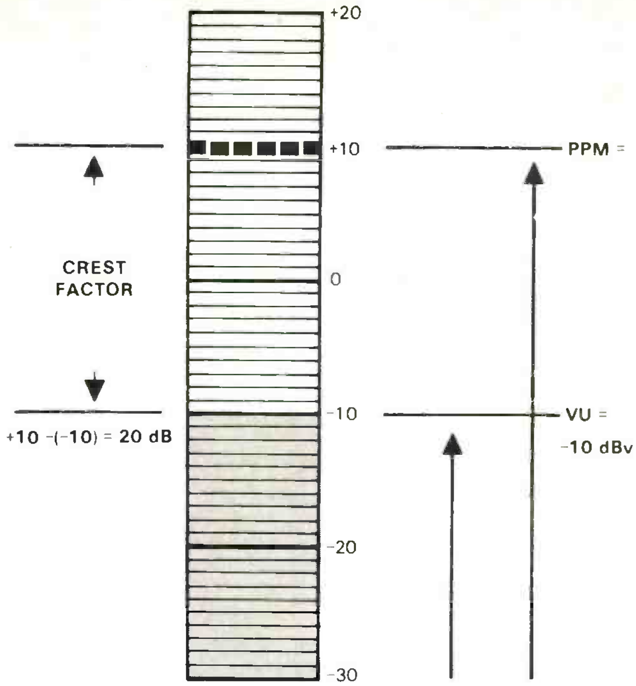

Figure 15: **Illustration of a meter which simulaneously reads peak and average.**

peak value as a dot floating above the bar, you can successfully display both characteristics on one meter.

The simultaneous display of peak and VU against a dB scale allows you a direct readout of crest factor (ratio of peak to VU), useful when compressing a track or even when deciding what tracks could use compression.

Caveat

By the way, almost everything discussed in this article that is of any value is either patented, patent applied for, or proprietary. Loft Modular Devices (Roberts), U.S. Patent No. 4,166,245. Allison Research, Inc. (Buff), U.S. Patents Nos. 3,237,028, 3,293,450, and 3,714,462. Unauthorized use prohibited by law. □□□

Glossary

Ohms Law: Ohms Law defines the relationship between voltage, current, and resistance. If you know any two you can calculate the third.

I = Current in amperes
V = Voltage in volts
R = Resistance in ohms
I = E/R, or E = R/I, or R = E/I

A OL = Open Loop Gain. How much the output changes for given input change.

A CL = Closed Loop Gain. Voltage that would appear at ouput as predicted by feedback if inserted directly at plus input.

"Johnson" Noise Voltage: Noise caused by thermal agitation of electrons in resistors. The amount of noise voltage is directly related to the resistance. Therefore, a 100 K resistor has much more noise than a 100 ohm resistor. The significance of this voltage depends on the circuit. A 100 ohm resistor can be noisy when at the input of a mike pre-amp, while a 100 K resistor may be insigificant in a unity gain op-amp.

Transconductance: Is the relationship of an output current to an input voltage. In the case of an op-amp input stage it refers to the current delivered to the next stage for a given differential input voltage.

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