A low-noise preamplifier with variable-frequency tone controls

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Preamble

In 1996 I published a design for what is commonly known as the Precision Preamplifier in Electronics World; one of its main features was a treble/bass tone control with the frequencies variable over a 10:1 range. When I mentioned on the DIYaudio board in May 2012 [1] that I had developed an improved version of this tone-control, I was surprised at the enthusiasm for its immediate publication.

I could of course have just written up the tone-control, but it occurred to me that since it inherently phase-inverts it needs another inversion before or after it to preserve absolute phase. A stage that merely inverts to correct phase has never appeared to me an efficient use of components, so I thought that it would be better to make use of it as a Baxandall active volume-control; I have therefore designed one intermediate in complexity and performance between the original 1996 one-path volume control and the 4-path volume control used in my recent Elektor preamp.[2] This 2-path volume control is placed after the tone-control because in normal use it is likely to be set for a gain less than unity and so noise from the earlier stage will be attenuated.

You can see where this is heading- with a comprehensive tone-control and an active volume-control we almost have a line-only preamplifier. We just need to add an input select switch. Actually things are slightly more complicated than that (surprise) as the tone-control can have quite a low input impedance and some sort of input buffer is needed. It could be a simple voltage-follower, but a balanced input amplifier will cost very little more and give the great advantage of rejecting noise from ground currents. Job done?

Well, maybe, but a conventional unity-gain balanced input amplifier made with four 10 kΩ resistors and a 5532/2 is a relatively noisy thing, giving some -104 dBu noise out. This would make it the noisiest block in the signal path by quite some margin, which is not helpful for a design intended to showcase a low-noise tone-control. The noise can be reduced significantly by adding a little more complexity in the form of two unity-gain buffers to drive a balanced input amplifier built from much lower value resistors. With 820 Ω resistors and using LM4562 sections in all three positions we get -113.9 dBu; almost a 10 dB improvement and now significantly quieter than the tone-control.

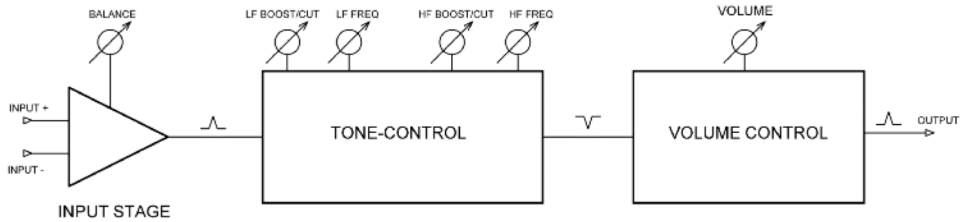

Figure 1Preamplifier block diagram. Spikes show the phase of the signal.

However, our work here is not yet done. I think it is difficult to claim that a preamplifier is usable without a balance control; we don't all have symmetrical listening spaces. On the other hand, it seems inelegant to add an extra stage just for the simple balance function, so there is a strong incentive to try to build it into one of the three existing stages. The tone-control is relatively complex and I don't relish the task of trying to add variable gain (non-interactive with all the tone functions, of course) to it. Trying to bodge it onto the active gain control without losing its supreme property- that volume depends on angular control setting and nothing else does not appear promising. That leaves us with the balanced input stage. If this is configured so that it can either attenuate or add gain, it makes a very effective vernier balance control; 'vernier' in that it has more than enough range to move the stereo scene fully from right to left, but is not capable of fading one channel to zero. That is quite unnecessary. I have used such vernier balance stages many times, but I think I have moved the technology on another step, as you will see below.

We wind up in **Figure 1** with something much simpler than the Elektor preamp; not with the same remarkable noise performance, but still better than almost anything else around. And it has the unique feature of the variable-frequency tone-control. A 4-pole input select switch is placed before the input amplifier.

Until recently my design policy has been to use a careful mix of cheap 5532s and expensive LM4562s, the latter only being used when they made a significant improvement to the performance. The price of LM4562s has continued to slowly decrease, so for this design I decided to use them throughout.

Balanced input amplifier

The variable-gain balanced input stage for the Elektor preamp was the state of the art- or at any rate the state of my art- when it was designed in 2011. Having thought more on the subject since, I think it can be improved.

As is now well-known, the great merit of the Baxandall active volume-control is that its control law depends only on the setting of a linear pot and not on the value of the pot track or the ratio of that track to fixed resistors. Since pot tracks are usually specified at ±20 %, dependence on their value leads to significant errors. With the Baxandall control excellent inter-channel balance is obtained with inexpensive pots. You can take it from me that customers really do complain if there is image shift when they adjust the volume control.

But what about the other preamplifier controls? The tone-control uses the same principle up to a point; the boost/cut controls are linear pots and their effects is independent of their track resistance, but the track values of the HF and LF frequency controls do have some effect on the frequency. It is, principle at least, possible to design out that track-resistance-dependence, but to do it without compromising noise and distortion might be quite a challenge. However, we can eliminate track-resistance-dependence in the balance control...

Figure 2a is the standard balanced input. **Figure 2b** is the low-noise variant, in which the addition of the two input buffers allows much lower resistances to be used around the differential amplifier, greatly reducing the effects of Johnson noise and opamp current noise. With the values shown, and using all LM4562s the improvement is 9.1 dB over a conventional 10 kΩ & 5532 balanced input as in Figure 2a. Making the gain of a balanced input stage variable is not just a matter of making two of the four resistors variable; the four-gang pot is never a welcome component, and in this application it would need to have quite impossible mechanical and electrical accuracy to give a good Common-Mode-Rejection-Ratio. (CMRR)

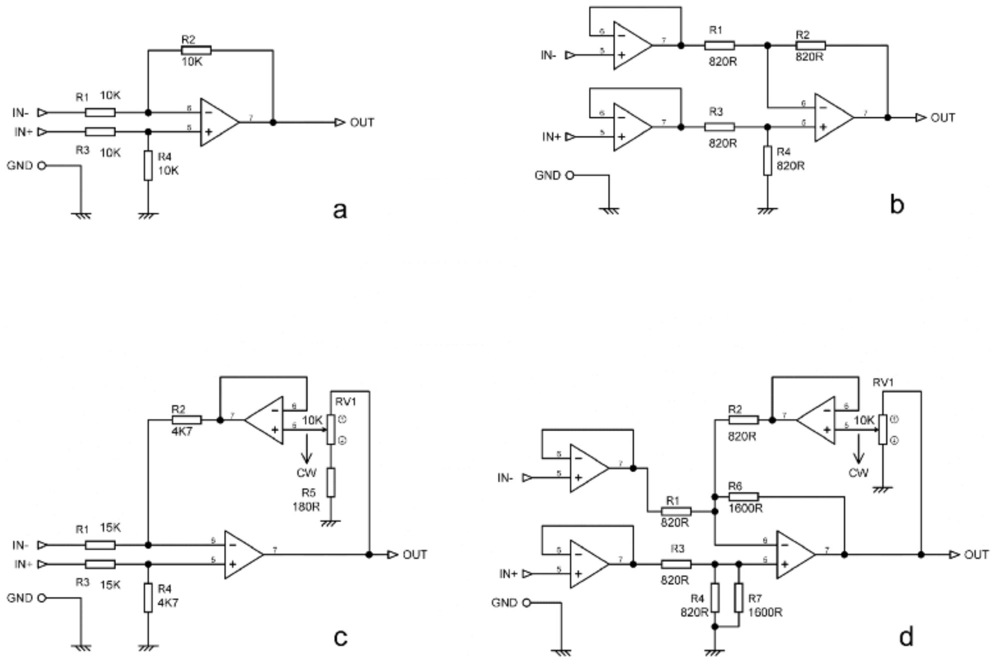

Figure 2 Balanced input stages: a) Conventional; b) with buffers for low-noise; c) Variable gain with CMRR preserved d) Low-noise, variable gain, and no pot-dependence. Gain increases as control is moved clockwise.

Figure 2c shows the way to do it. Since the feedback arm is driven from a constant and very low impedance, the CMRR does not vary at all when the gain control is altered. This is in almost every way a very satisfactory configuration and I have used it times without number in mixing consoles. If resistors R2 and R4 are made less than R1 and R3 then the stage can attenuate as well as amplify, which is ideally what we want for a balance control. The values in Figure 2c are optimised for a general purpose line input gain control rather than a balance control, the maximum gain being +25 dB and minimum gain -10 dB. This means a reverse log pot is required to get middle gain at middle control setting. For the more restricted gain range of a vernier balance control a linear pot is satisfactory so the uncertainties of dual slope controls are avoided. The resistors have relatively high values as input buffers are not used. However, input buffers can be added just as in Figure 2b, allowing noise reduction.

A snag remains. You will note that the maximum gain of Figure 2c is set by the end-stop resistor R5 at the bottom of the pot. The pot track resistance will probably be specified at ± 20 %, while the resistor will be a much more precise 1%. The variation of pot track resistance can therefore cause quite significant differences between the two channels. While this is less important than the volume control matching, in that it only causes a fixed balance error rather than varying image-shift, it would be nice to eliminate it. If we could lose the end-stop resistor then the pot would be working as a pure potentiometer with its division ratio controlled by the angular position of the wiper alone.

Obviously if we simply do away with the end-stop resistor the gain is going to be uncontrollably large with the pot wiper at the bottom its track. We need a way to limit the maximum gain that keeps other resistors away from the balance pot.

I was just going to bed at dawn the other day (the exigencies of the audio service, ma'am) when it occurred to me that the answer is to add a resistor that gives a separate feedback path around the differential opamp only. R6 in **Figure 2d** provides negative feedback independently of the gain control and so limits the maximum gain. It is not intuitively obvious (to me, at any rate) but the CMRR is still preserved when the gain is altered, just as for Figure 2c. You will note in Figure 2d that two resistors R4, R7 are paralleled to get a value exactly equal to R2 in parallel with R6. Figure 2d shows resistor values that give a minimum gain of -3.6 dB and a maximum gain of +5.8 dB, and low value resistors are used with input buffers once more to reduce noise. The gain with the pot wiper central is -0.1 dB which I would suggest is close enough to unity for anyone. This concept eliminates the effect of pot track resistance on gain, and I propose to call it the Self Input. Alright?

In Figure 2d two resistors R4, R7 are paralleled to equal R2 in parallel with R6. The nominal values must be exactly equal for the best CMRR; in practice we are at the mercy of resistor tolerances, but we should at least start out with an exact nominal value. The first prototype gave a CMRR of -56.5 dB, (flat 20 -20 kHz) which is rather better than you would expect from 1% resistors; they are commonly more accurate than their official tolerance. If it came out as, say, -30 dB you could be sure that something was wrong with either the design or construction.

Figure 3 shows the final schematic for the balanced input amplifier, and incorporates not only the extra components required for practical use, such as EMC filters, but also some extra minor improvements. In Figure 3, making R7 equal to R8, R10 equal to R11, and R9 equal to R12 is the simplest way to get the nominal resistance values equal for good CMRR. However, so long as the ratio between the feedback and non-inverting arms is correct the resistors need not have the same value. If we scale down all the resistors in the non-inverting arm by the same ratio we reduce the effective resistance at the non-inverting input of A2b by that factor and so reduce both Johnson noise and the effect of the current noise of A2b at its non-inverting input. Dividing them all by ten would be simple and effective as regards noise but places far too much loading on the buffer A1b, so we have to use a smaller factor. We have to divide the resistors by this smaller factor, and here some minor troubles begin. Assuming we stick with E24 resistor values, we are not going to find a combination that gives the exact ratio we want, even though there are two in parallel which gives us some extra freedom.

We will here divide the resistor values by roughly a factor of two, so that R is changed from 820 Ω to 470 Ω. This is a conservative approach to avoid too heavy a loading on A1b. Applying the ratio 470/820 to the combination of R11 and R12 in parallel we get a target value for their combined resistance of 310.7436 Ω; you will see the need for such precision in a moment. The closest approach we can make with two E24 values is R = 620 Ω and R = 620 Ω, giving a nominal combined value of 310 Ω. The measured CMRR was now only -43.6 dB. Since different resistor specimens were used this doesn't actually prove anything, but it is rather worrying. 310 Ω is only 0.239% low compared with 310.7436 Ω, but that represents a quarter of the tolerance of a 1% resistor.

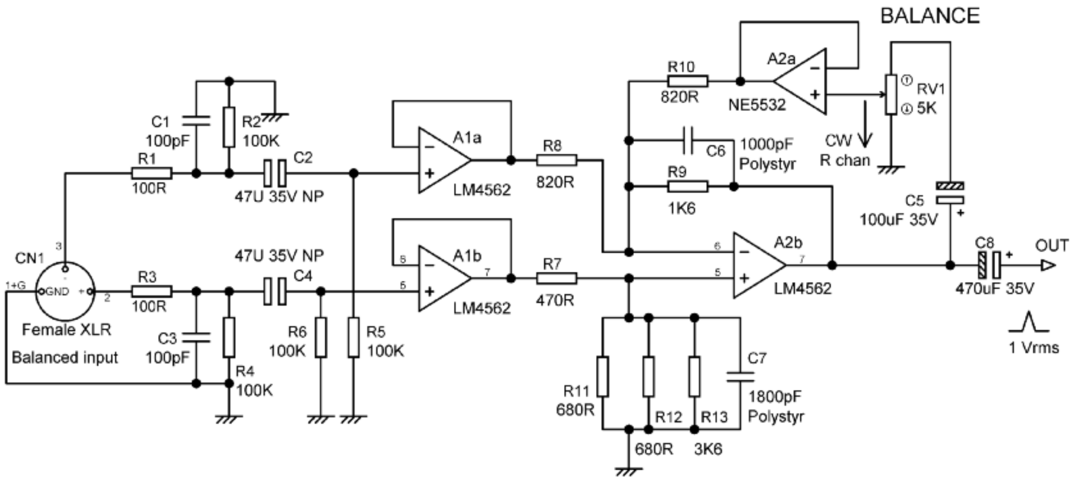

Figure 3 Final arrangement of the variable-gain balanced input amplifier. The balance pot is a linear type.

We can get much closer to the desired value if we permit ourselves three parallel resistors. R11 = 680 Ω, R12 = 680 Ω, and R13 = 3k6 gives us a nominal combined value of 310.6598 Ω, which is only 0.0270% low, much less than the resistor tolerance. Using this network gave a measured CMRR of -49.3 db, which is rather more satisfactory. The nominal value is now so close to that required that the individual tolerances of the resistors will completely determine the CMRR behaviour.

As I have written elsewhere, [3] a resistance made up of two equal resistors, either in series or parallel, will on average be more accurate by a factor of √2 than the resistor tolerance because random errors in the values tend to partially cancel. Three equal resistors are more accurate by √3, four equal resistors by a factor of √4 =2, and so on. Because 3k6 is much larger than the two 680 Ω resistors it has little effect on the final value so the accuracy improvement here is only better than √2 by a tiny amount. Still, it all helps.

So, after a somewhat effortful reduction of resistance, how much lower is the noise? Frankly, the results are not stunning. Noise is only reduced by 0.5 dB at minimum gain and 0.3 dB at higher gains. There is some scope for scaling down R7 etc further without overloading A1b, but very little is likely to be gained. Nonetheless the improvement, although small, should be wholly robust and costs only one extra resistor. A bit of work now and slightly lower noise forever.

Table 1 gives the noise performance after scaling down the resistor values. I think you will agree it is fairly quiet.

	Min gain	Mid gain	Max gain
Stage gain dB	-3.6	-0.1	+5.8
Noise out dBu	-116.6	-112.4	-107.3
EIN dBu	-113	-112.3	-113.1

Table 1: Balanced input stage output noise at 3 gain settings. R7 = 470 Ω. Corrected for AP noise at -119.2 dBu; measurement bandwidth 22 Hz – 22 kHz, rms sensing, unweighted.

It is however slightly regrettable that the EIN is about 0.7 dB worse at the central balance setting, where it is most likely to be used. This is because the pot wiper sees the maximum impedance when in the middle of the track, so this is the worst case both for Johnson noise from the pot, and the effect of A2a current noise flowing in that impedance. Their calculated contributions are of the same order. There doesn't seem to be much that can be done about this except to use a 1 kΩ pot as in the Elektor preamp, with a suitable increase in the value of C5. I prefer to keep all the pots the same value.

The final step is to add the capacitors C6, C7 to define the HF bandwidth at the first practical opportunity. The values shown (which are larger than usual) give -1.0 dB at 100 kHz. The value of C7 has to be scaled up in the same ratio as the resistors R7 etc were scaled down; fortuitously, 1800 pF is close enough to the exact value to maintain good CMRR up to 20 kHz and beyond. Polystyrene capacitors

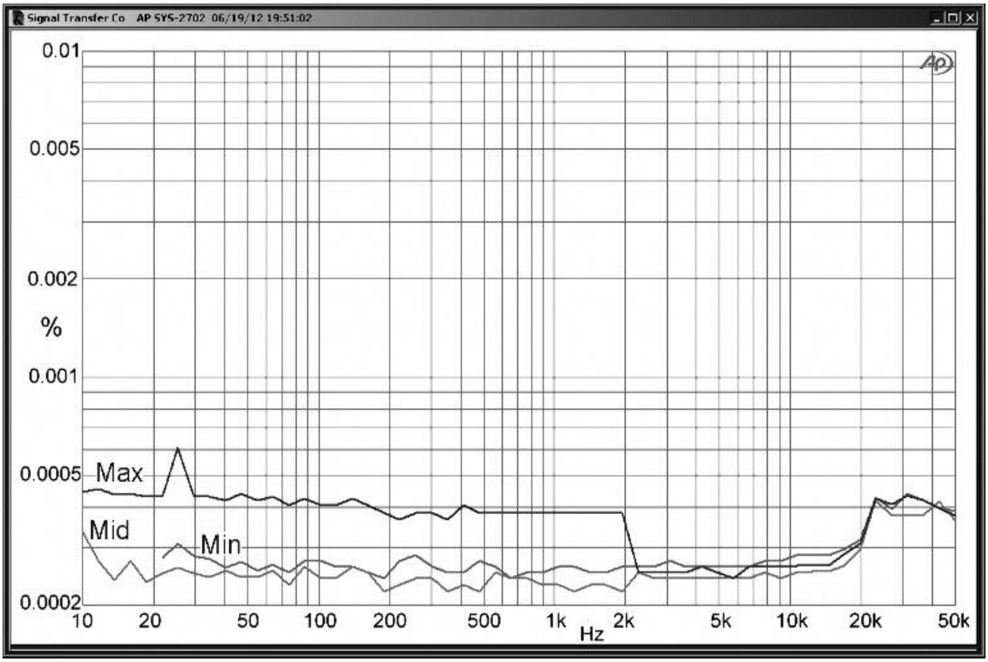

Figure 4 Balanced input amplifier distortion at minimum gain, mid gain, and maximum gain. 9 Vrms out for each case. The step in the upper trace is solely due to internal gain-switching in the testgear.

should be used, preferably with a 1% tolerance. Be aware that the circuit may show signs of parasitic oscillation if these capacitors are not installed. It will be stable without the capacitors if A2 is a 5532, but the noise performance is then 3 dB worse.

The distortion of the balanced input amplifier (with balanced input signal) is shown in **Figure 4**. The output level of 9 Vrms is at least 6 dB above the highest level the stage is likely to handle.

The variable-frequency tone control

The tone-control stage has two bands- bass (LF) and treble, (HF) and in that respect resembles a conventional Baxandall tone control. However it is a great deal more flexible; in each band the frequency at which control begins is variable over a ten to one range, making it much more useful for correcting speaker deficiencies etc. Very few manufacturers have offered this facility. The only one that comes immediately to mind is the Yamaha C6 preamp (1980-81) which had LF and HF frequency variable over wide ranges (they could overlap in the middle) and Q controls for each band as well.
Some improvements have been made to the tone-control compared with the 1996 design; they are:

- All opamps changed from 5532 to LM4562 to reduce noise and distortion
- Circuit redesigned to use 5k linear pots throughout instead of 10k linear, again to reduce noise
- Frequency control laws improved so the middle of the frequency range corresponds with the

middle position of the control (centre-detent).

- Number of expensive 220 nF polypropylene capacitors reduced
- DC blocking added to LF frequency control to stop DC flowing through it
- Noise in tone-cancel mode reduced

The boost/cut range is +/-10 dB, the LF frequency range is 100 Hz - 1 kHz, and the HF frequency range is 1 kHz - 10 kHz. The response curves do not level out at their boosted or cut level, but smoothly return to unity gain outside the audio band. This is sometimes called an RTF (Return-To-Flat) characteristic; in the mixing console world it would be known as a 'peaking' EQ. It is a valuable feature because boosting 10 kHz is one thing, but boosting 200 kHz quite another, and can lead to stability or EMC problems. Likewise a touch of boost at 40 Hz may improve a loudspeaker response, but boosting 5 Hz by any amount is likely to be a bad idea. The RTF time-constants are fixed so the boost/cut ranges are necessarily less than +/-10 dB toward the frequency extremes, where the RTF effect starts to overlap the variable boost/cut frequencies. The measured responses for maximum boost and maximum cut at minimum, middle, and maximum frequency settings are shown in **Figures 5 and 6**
The schematic of the tone-control is shown in **Figure 7**.

The tone-control stage consists of an inverting amplifier A3a, and two separate frequency-selective paths that provide LF and HF control. The LF path is essentially a variable-frequency first-order low-

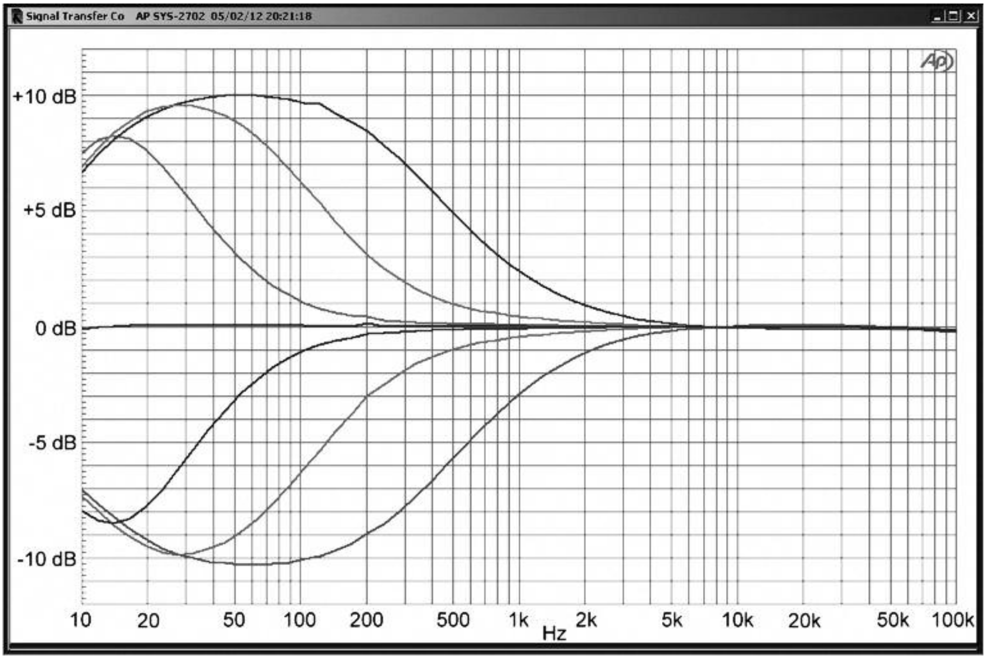

Figure 5 LF tone-control frequency response, max cut/boost, at minimum, middle, and maximum frequencies.

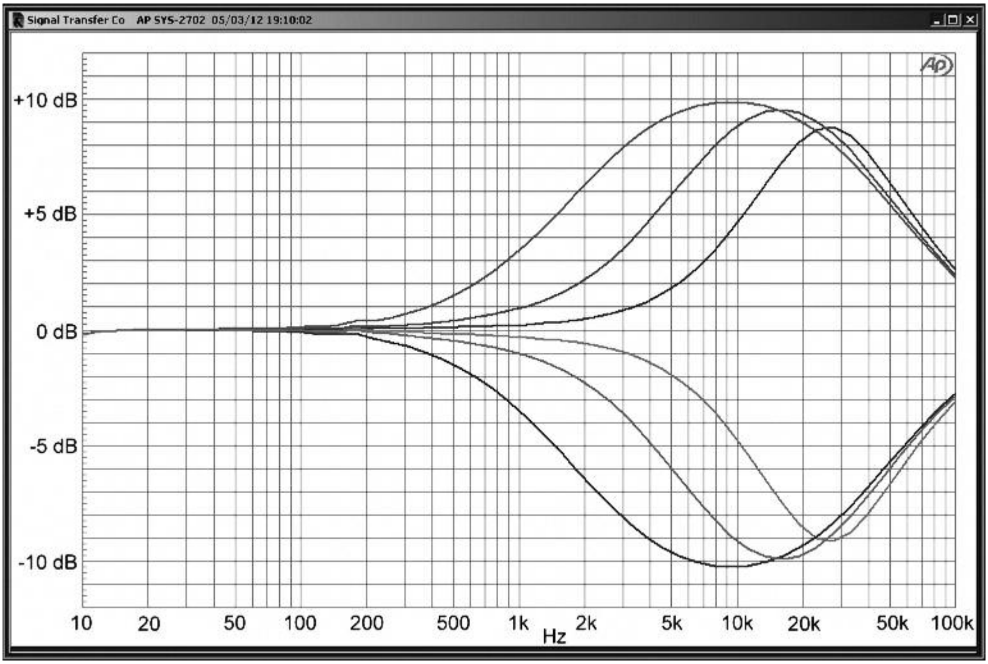

Figure 6 HF tone-control frequency response, max cut/boost, at minimum, middle, and maximum frequencies.

pass filter, and the associated LF cut/boost control can only act on the frequencies this lets through. Similarly the HF path is essentially a variable-frequency high-pass filter. The filtered signals from the LF and HF paths are summed and returned to amplifier A3a via its non-inverting input. Returning these signals at unity gain would give excessive levels of maximum cut and boost, so an attenuation of 9 dB is introduced into the return path. This limits the maximum cut and boost to +/- 10 dB. This attenuation is introduced by the loading of R22 on the combination of R27 and R33. The gain of in the two paths is unity before this attenuation is applied, and so there are no problems with them clipping before the output of A3a. It is therefore possible to put the 9 dB loss after the two paths rather than before, so it very handily attenuates the noise from them. The loss attenuator is made up of the lowest value resistors that can be driven without distortion, to minimise both the Johnson noise thereof and the effect of the current noise of the non-inverting input of op-amp A3a.

The Tone Cancel switch SW1 disconnects the return signal from any contribution to the main path, preventing 5 out of 6 opamps from contributing noise. R22 is also shorted to ground, removing its Johnson noise and the effect of the current noise from A3a. Tone-cancel usefully reduces the stage output noise by about 4 dB, depending on the HF freq setting. It leaves only A3a in circuit; if this was switched out its phase-inversion would be lost and (assuming absolute phase in the whole signal chain is preserved with the tone-control active) a phase error introduced. This tone-cancel configuration also has the advantage that the signal does not briefly disappear as the cancel switch moves

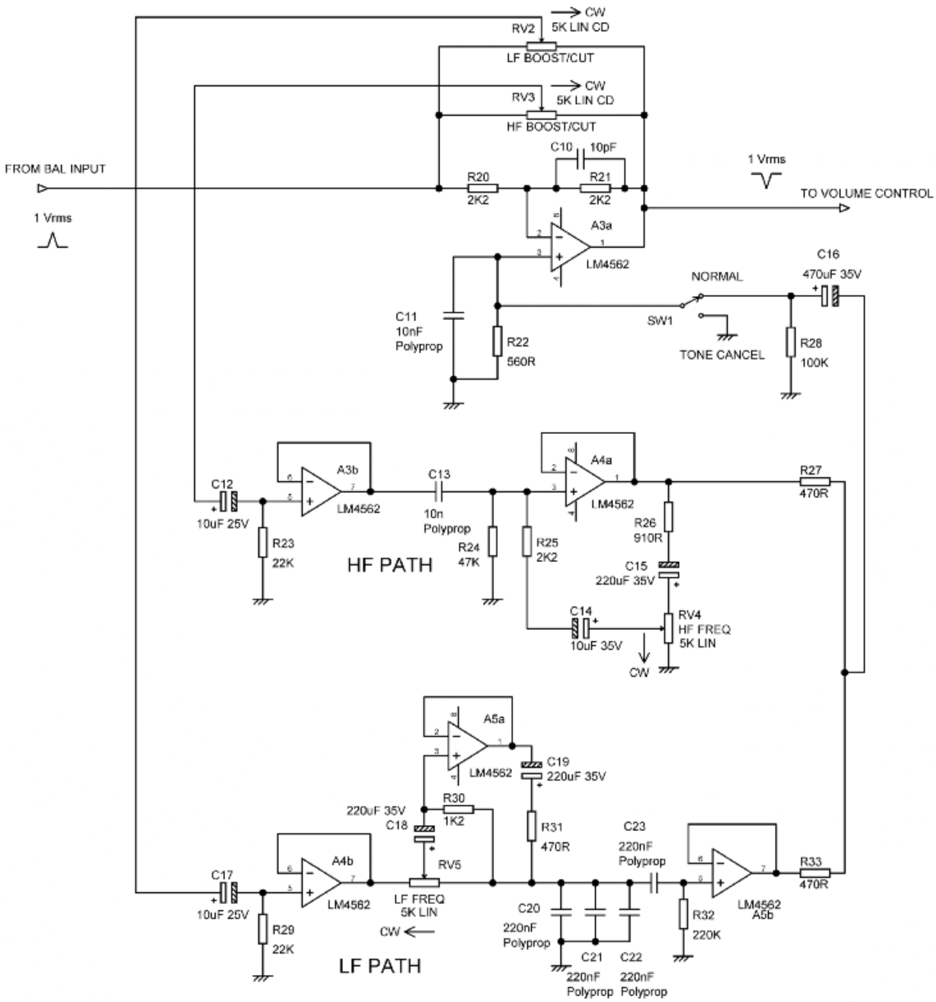

Figure 7 The tone-control schematic, showing the two separate paths for HF and LF control. All pots are 5 kΩ linear.

between its two contacts. This minimises transients due to suddenly chopping the waveform and makes valid tone in/out comparisons much easier. Such interruptions are bound to occur when an entire stage is by-passed.

The LF path begins with buffer A4b which prevents loading on the cut/boost control RV2 and so avoids boost/frequency interaction. The low-pass RC time-constant capacitance is the parallel combination of C20, C21 and C22; the use of three capacitors increases the accuracy of the total value by √3. The associated time-constant resistance is a combination of RV5, R30, and R31. The LF frequency law is made approximately logarithmic by A5a; at minimum frequency RV5 is fully anti-clockwise, so the input of buffer A5a is the same as at the C20 end of R31, which is thus bootstrapped and has no effect. When RV5 is fully clockwise R30 and R31 are effectively in parallel with RV5 and the turnover frequency is at a maximum. The presence of R30 gives a roughly logarithmic law; its value is carefully chosen so that the centre of the frequency range is at the centre of the control travel. Sadly, there is some pot-dependence here.

The original design had DC excluded from all the pots except the LF frequency control, which had about 2 mV across its track due to the offset voltage of A5a; the control noise from this was audible at very high gain and so it had to be dealt with. Putting a blocking capacitor after A4b does not work as this leaves A5a without any DC reference; adding a bias resistor to ground immediately after the blocking cap also did not work too well as a tendency for the DC conditions to wander around a bit remained. Putting C19 in series with the output of A5a instead worked nicely. There is absolutely no effect on the LF response.

The HF path begins with buffer A3b to prevent loading on cut/boost control RV3. C13 with R24 and R25 make up the basic high-pass time-constant. The effective value of the time-constant resistance can be altered over a 10:1 range by varying the amount of drive that R25 receives from A4a, the potential divider effect and the rise in source resistance of RV4 in the centre combining to give a good approximation to a logarithmic frequency/rotation law R26 is the frequency endstop resistor that limits the maximum effective value of R24; its value is carefully chosen so that the centre of the frequency range is at the centre of the control travel. C11 is the HF RTF capacitor- at frequencies above the audio band it shunts the sidechain signal to ground, so the HF cut/boost control no longer has any further effect. The HF frequency setting has a strong effect on the noise performance; this is true even when the HF boost /cut control is set to flat, as the HF path is still connected to A3a.

Compared with the HF path, the LF path contributes very little extra noise to the tone stage because most of its circuitry is before the low-pass filter, which almost eliminates its noise contribution; only the noise from buffer A5b returns unfiltered. This can be seen in Table 2. The RTF time-constant for the LF path is set by C23 and R32, which block very low frequencies and so limit the lower extent of LF control action. C23 was originally 1 uF, which would be very expensive in a no-distortion polypropylene type, and so it was reduced to 220 nF and resistor R32 increased accordingly to maintain the same roll-off frequency. The impedance at this point is therefore quite high at low frequencies and care should be taken that C23 does not pick up electrostatic hum.

A complete evaluation of the performance of the tone-control is a lengthy business because of the large number of permutations of the controls. If we just look at extreme and middle positions for each control, we have maximum boost, flat, and maximum cut for both HF and LF, and maximum, middle and minimum for the two frequency controls, yielding 3 x 2 x 3 x 2 = 36 permutations. It is a pain to measure every possibility so the measurements here are restricted to those that put the greatest demands on the circuitry. (eg HF freq at minimum) Table 2 shows the noise output of the tone control at these settings:

HF level	HF freq	LF level	LF freq	Noise out dBu
Flat	Min	Flat	Max	-105,1
Flat	Mid	Flat	Max	-106,2
Flat	Max	Flat	Max	-107,2
Flat	Max	Flat	Mid	-106,8
Flat	Max	Flat	Min	-107,1
Flat	Min	Flat	Min	-105,6
Max boost	Min	Flat	Min	-100,5
Max cut	Min	Flat	Min	-107,4
Max cut	Min	Flat	Max	-107,6
Flat	Min	Max boost	Max	-103,8
Flat	Min	Max cut	Max	-105,9
Flat	Min	Max cut	Min	-105,6
Flat	Min	Max boost	Min	-105,0
Tone-cancel				-110,2

Table 2: The noise output of the tone control at various settings. Not corrected for AP noise at -119.2 dBu. (Difference negligible). Measurement bandwidth 22 Hz – 22 kHz, rms sensing, unweighted.

The final improvement planned for the tone-control was to reduce the noise in the tone-cancel position by reducing the resistors R20 and R21 from 4k7 to 2k2. The output noise with tone-cancel engaged is -110.2 dBu with R20, R21 = 4k7, and with R20, R21 = 2k2 it drops handily to -112.6 dBu. Since purists will no doubt do most of their listening with tone-cancel engaged this seems a useful modification. But...

It is essential to bear in mind that circuits like this tone-control can show unexpected input impedance variations. A standard Baxandall tone-control made with 10 kΩ pots can have an input impedance that falls to 1 kΩ or less at high frequencies where the capacitors have a low impedance. It is not obvious but the alternative tone-control configuration used here also has serious input impedance variations.

Looking at the circuit with the original 4k7 resistors, it would be easy to assume that because the input terminal connects only to a 4k7 resistor and two 5 kΩ pots the input impedance cannot fall below their parallel combination ie 4k7 || 5k || 5k = 1.63 kΩ. It would also be quite wrong, because while the other end of the 4k7 resistor is connected to virtual ground, the two 5 kΩ pots are connected to the stage output. When the controls are set flat, or tone-cancel is engaged, this carries an inverted version of the input signal. The effective value of the pots is therefore halved, with zero voltage occurring halfway along the pot tracks. The true input impedance when flat is therefore 4k7 || 2.5k || 2.5k = 987 Ω, which is confirmed by simulation.

When the tone-control is not set flat, but to boost, then at those frequencies the inverted signal at the output is larger than the input. This makes the input impedance lower than for the flat case; when the circuit is simulated it can be seen that the input impedance varies with frequency inversely to the output amplitude. Conversely, when the tone-control is set to cut, the inverted signal at the output is reduced and the input impedance is higher than in the flat case. This is summarised in Table 3, with R20, R21 = 4k7.

HF level	HF freq	LF level	LF freq	Input impedance Ω
Flat	Min	Flat	Mid	987
Flat	Mid	Max boost	Mid	481
Flat	Mid	Max cut	Mid	1390
Max boost	Mid	Flat	Mid	480
Max cut	Mid	Flat	Mid	1389

Table 3: Input impedance of the tone control at various settings.

As you can see from the table, the input impedance falls to the worryingly low figure of 480 Ω at maximum HF and LF boost. It is true that in this case the gain is +10 dB, and so the input voltage into this impedance cannot exceed 3 Vrms without the output clipping, limiting the current required of the preceding stage, but it is not clear that driving a low voltage (3 Vrms) into an impedance that would cause distortion when driven with a high voltage (9 Vrms) gives suitably low distortion. An investigation of this appears to be work that needs doing. The planned reduction of R20, R21 from 4k7 to 2k2 to reduce noise with tone-cancel engaged brings the minimum input impedance down to 388 Ω, and this now looks like rather less of a good idea unless you are planning to use the tone-control after a stage with plenty of drive capability.

The electrolytic capacitors in the tone-control are used solely for DC blocking, and play no part at all in determining the frequency response. This would be highly undesirable, both because of the wide tolerance on value, and the distortion generated by electrolytics when they have significant signal voltage across them. The only design criterion for these capacitors was that they should be large enough to introduce no distortion at 10 Hz and the original values were chosen using the precision algorithm "looks about right" though I hasten to add this was followed up by THD measurements to confirm all was well (See also 'Electrolytic capacitor distortion' at the end of this article).

The distortion of the tone-control was measured for various other permutations of maximum boost/cut and maximum/minimum frequency. There were no surprises there so I will not use up valuable space displaying the results; if there is interest I will put them on my website at douglas-self.com. Suffice it to say that THD at 9 Vrms in/out never gets above 0.001% except now and then it just barely exceeds it at 20 kHz. General THD levels are in the range 0.0003 – 0.0007 %.

Figure 8 shows the distortion performance when the controls are set flat, for active and with tone-cancel. The slightly higher level with tone active is due solely to extra noise from the LF and HF paths.

I have spent some time on the design issues of the tone-control because it is the original raison d'e-tre of this article. There are many things to consider, and no doubt the design could be further improved with some more work. In terms of mixing console design this is a simple EQ circuit; a big console would have four separate control bands instead of two, and each band would have variable Q as well as frequency. The top and bottom bands would also have a peak/shelving switch allowing defeat of the RTF feature. The complexity is considerable.

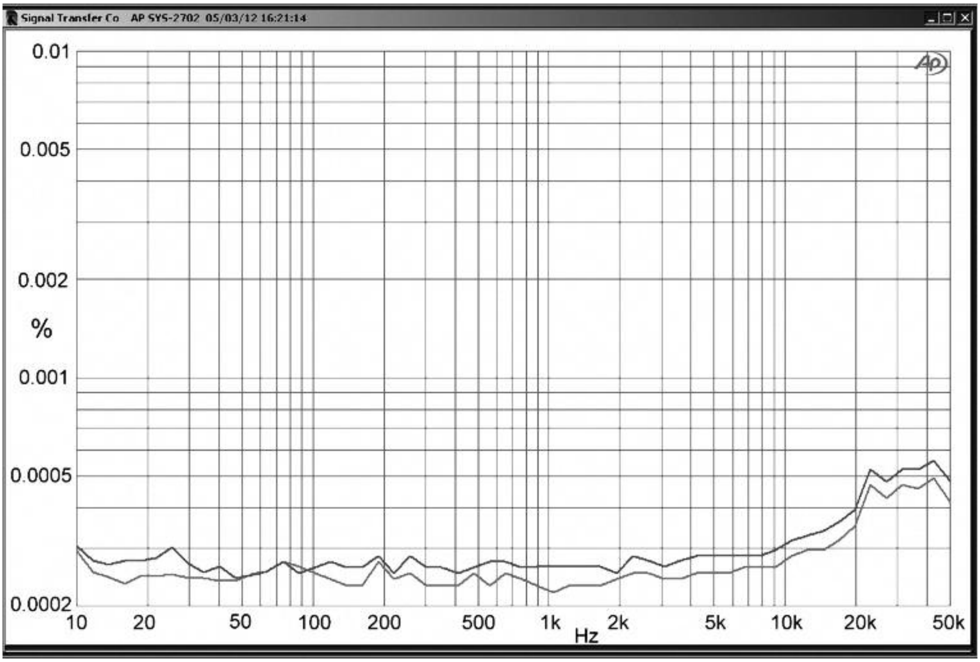

Figure 8 Tone-control distortion at 9 Vrms output. Lower trace is for tone-cancel mode, upper trace is for tone-control active but set flat.

The active volume control

As previously mentioned, the great value of the Baxandall active volume-control is that its gain depends only on the angular setting of the control, and not the ratio of the track resistance (with its ±20% tolerance) to other resistors, or the dubieties of two-slope "log" pots. This is because the maximum gain is set not by end-stop resistors but by an amplifier stage with its gain set by two fixed resistors. As a result even quite ordinary dual linear pots can give very good channel matching; the one problem that the Baxandall configuration cannot solve is channel imbalance due to mechanical deviation between the wiper positions. A disadvantage is that the gain/rotation law is determined solely by the maximum gain, and it is not possible to bend it about by adding resistors without losing the freedom from pot-value-dependence.

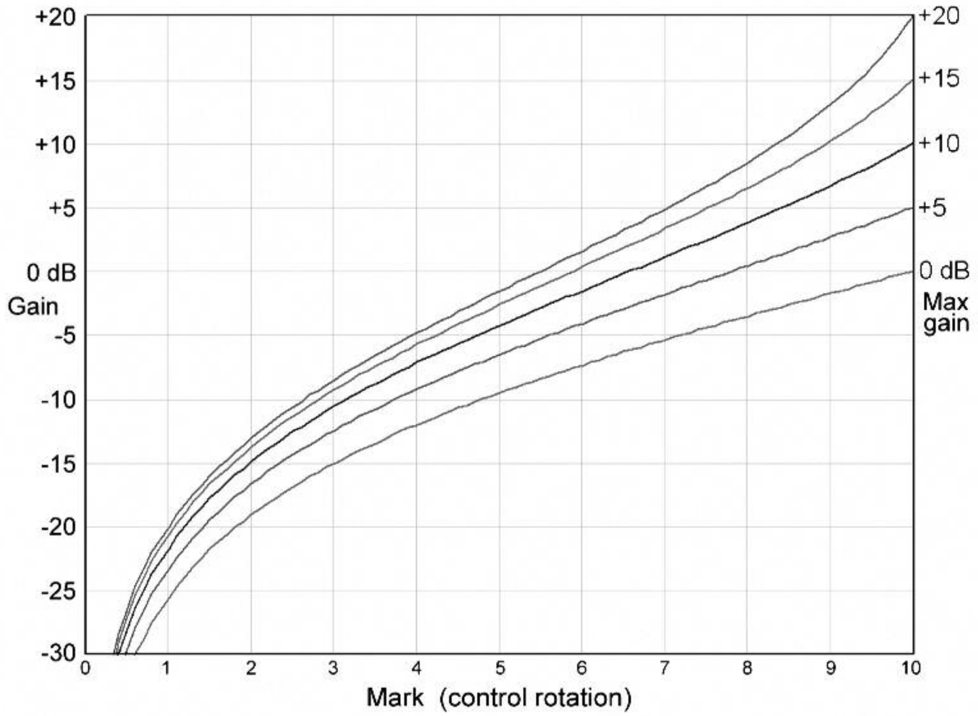

Figure 9 The gain law of a Baxandall volume control stage depends only on the maximum gain; here 0, +5, +10, +15 and +20 dB.

Figure 9 shows how this works. Pot rotation is described here as Marks from Mk 0 to Mk 10 for full rotation. (No provision is made for rock bands seeking controls going up to Mk 11 [5]). Changing the maximum gain has a much smaller effect on the gain at the middle setting (Mk 5). In this preamplifier design a maximum gain of +10 dB is used, giving -4 dB with the volume control central.

Figure 10 shows the volume stage. Noise is reduced compared with the 1996 design by the use of a 5 kΩ rather than a 10 kΩ pot, and by doubling the shunt-feedback gain stages so their noise partially cancels and gives a 3 dB advantage. A6a is a unity-gain buffer that prevents the gain stage loading the pot, and A6b, A7a are the shunt-feedback stages whose gain sets the maximum gain of the complete control. Their outputs are averaged by the 10 Ω resistors R43, R46. At a first glance the Baxandall volume configuration looks like a conventional shunt feedback gain control, but the vital difference is that the maximum gain is set by R41, R42 and R44, R45. The stage inherently gives a phase inversion which here neatly cancels that of the inverting tone-control stage, so preserving absolute phase. It requires a low-impedance drive such as an opamp output if it is to give the designed gain range and freedom from law-dependence on the pot value.

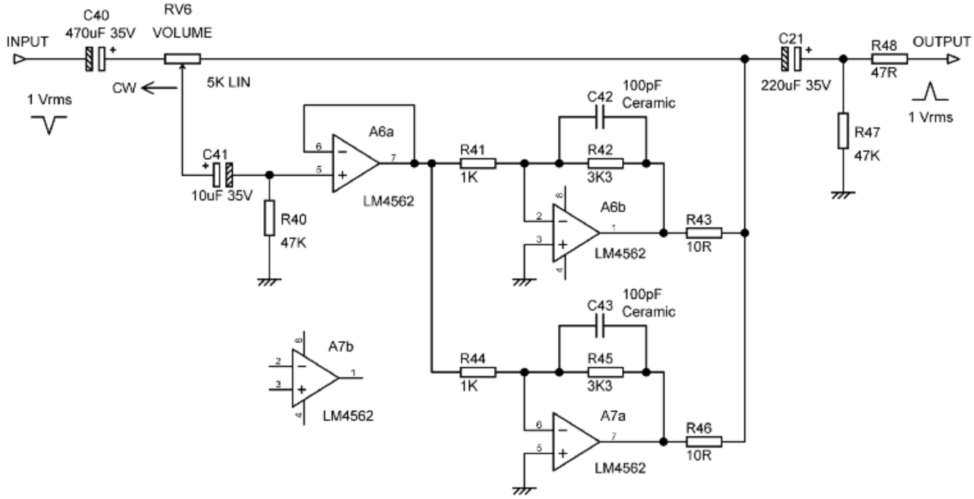

Figure 10 Baxandall volume-control with single buffer, dual gain stages. Maximum gain is +10 dB.

With circuits like this that are not wholly obvious in their operation, it pays to keep a wary eye on all the loading conditions. There are three loading conditions to consider:

Firstly, the input impedance of the stage. This varies from the pot track resistance at Mk 0, to a fraction of this set by the maximum gain of the stage. It falls proportionally with control rotation as the volume setting is increased. **Figure 11** shows how the minimum input impedance becomes a smaller proportion of the track resistance as the maximum gain increases. With a maximum gain of +10 dB the minimum input impedance is 0.23 times the track resistance, which for a 5 kΩ pot gives 1.162 kΩ. Since the preceding stage is based on an LM4562 it will have no trouble at all in driving this.

Secondly, the loading on the buffer stage A6a. A consequence of the gain of the 2-path A6b, A7a stage is that the signals handled by the unity-gain buffer A6a are never very large; less than 3 Vrms if output clipping is avoided. This means that R41, R42 and R44, R45 can all be kept low in value to reduce noise without placing an excessive load on unity-gain buffer A6a.

Thirdly, the loading on the gain stages A6b, A7a. At Mk 10 the loading is a substantial fraction of the value of the pot, but it gets heavier as volume is reduced, as demonstrated in the rightmost column of Table 4. We note thankfully that the loading stays at a reasonable level over the mid volume settings. Only when we get down to a setting of Mk 1 does the load get down to a slightly worrying 383 Ω; however at this setting the attenuation is -21.6 dB, so even a maximum input of 10 Vrms would only give an output of 830 mV. We also have two opamps in parallel to drive the load, so the opamp output currents are actually quite small. Note that the loading considered here is only that of the pot on the gain stages. The gain stage opamps also have to drive their own feedback resistors R42 and R45,

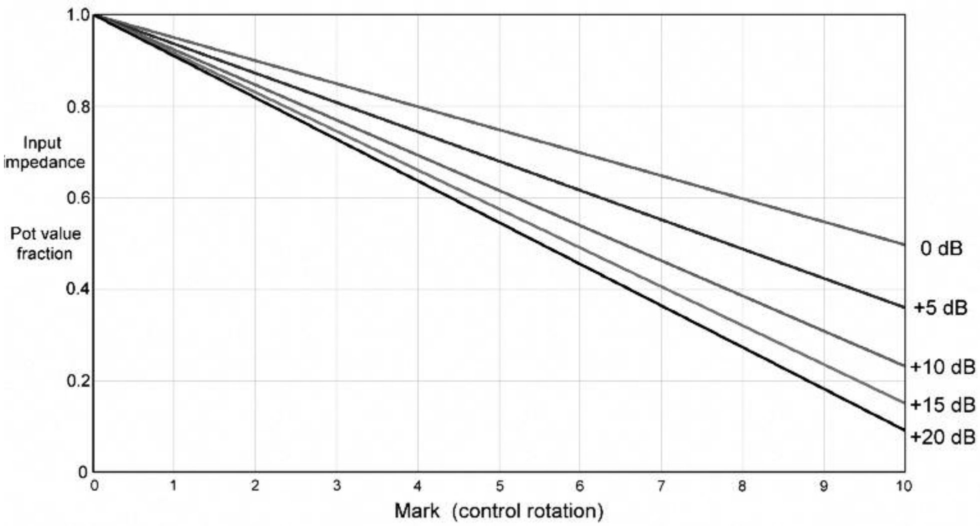

Figure 11 The input impedance of the volume control stage as a proportion of the pot track resistance falls more rapidly when the stage is configured for higher maximum gains.

and whatever load is connected to the preamp output sockets, so this should be taken into account. I have heard doubts expressed about a possible rise in distortion at low volume settings, because the gain stages A6b and A7a see a very low impedance. This may be so, but the current to be absorbed by the gain stages is very limited because almost the whole of the pot track is in series with the input at low settings. To prove there is not a problem here, I set the volume to Mk 1 and pumped 20 Vrms in, getting 1.6 Vrms out. The THD residual was indistinguishable from the GenMon output of the AP SYS-2702. In use the input cannot exceed 10 Vrms as it comes from an opamp.
To push things further, I set the volume to Mk 0.2, (ie only 2% off the endstop) and shoved 20 Vrms in to get only 300 mVrms out. The THD+N residual was 0.0007%, composed entirely of noise with no trace of distortion. I then replaced the 4562s in the A6b, A7a positions with Texas 5532s (often considered the worst sort for distortion) and the results were just the same except the noise level was higher giving a THD+N of 0.0008%. As noted, table 4 shows the gain and the noise output at the various control settings.

I haven't bothered to fill in all the entries for input impedance, as it simply changes proportionally with control setting as seen in Figure 12, ranging from the minimum of 1162 Ω to 5 kΩ, the resistance of the pot track.

Control position (Mk)	Gain dB	Noise output dBu	Input impedance Ω	Opamp load Ω
10	10,37	-109,0	1162	3900
9	6,98	-110,3	1547	3456
8	4,03	-111,4	1929	3069
7	1,29	-112,7		2687
6	-1,38	-113,7		2305
5	-4,11	-114,9	3083	1918
4	-7,07	-115,7		1534
3	-10,48	-117,1		1151
2	-14,83	-118,4		767
1	-21,61	-119,8		383
0	-infinity	-121,3	5000	

Table 4 Volume stage gain, noise, input impedance, and gain stage loading versus control setting. Corrected for AP noise at -119.2 dBu. Measurement bandwidth 22 Hz – 22 kHz, rms sensing, unweighted.

Figure 12 shows the distortion performance at maximum gain (Mk 10) and 9 Vrms out. A small amount of third-harmonic is discernable in the noise of the THD residual. Results at Mk 9 and below are indistinguishable from the testgear output apart from a small amount of added noise.

Overall performance

Now to look at the noise and distortion performance when the input/balance, tone control, and volume control blocks are plugged together. The noise will accumulate down the signal path in an orderly rms-sum fashion, but distortion products may reinforce or cancel making the outcome unpredictable.

The noise from the balanced input amplifier with the balance control central is -112.4 dBu. The noise from the tone-control depends on its settings, but if set for a flat response is -107.2 dBu with the HF frequency control at maximum and -105.6 dBu with the HF frequency control at minimum. If we add the balanced input noise to those figures we get -106.0 dBu (HF freq max) and -104.8 dBu (HF freq min). The difference is now only 1.2 dB; we will take the worst case of -104.8 dBu and see what happens when we send that to the volume control.

Figure 13 shows how the noise from the volume-control stage increases with gain. When added to the noise coming from upstream scaled by the stage gain, we get the "total noise" line. The noise output with the control central (it will probably spend most of its time around there) is a satisfyingly low -108 dBu. The dotted line shows the upstream noise scaled by the volume gain but with the volume stage noise not added; in other words the result with an ideal noiseless volume control.

The point of this sort of noise analysis is to find out if any stage is dominating the noise situation, and if so, is it worthwhile spending time and money making it quieter? The volume stage noise contri-

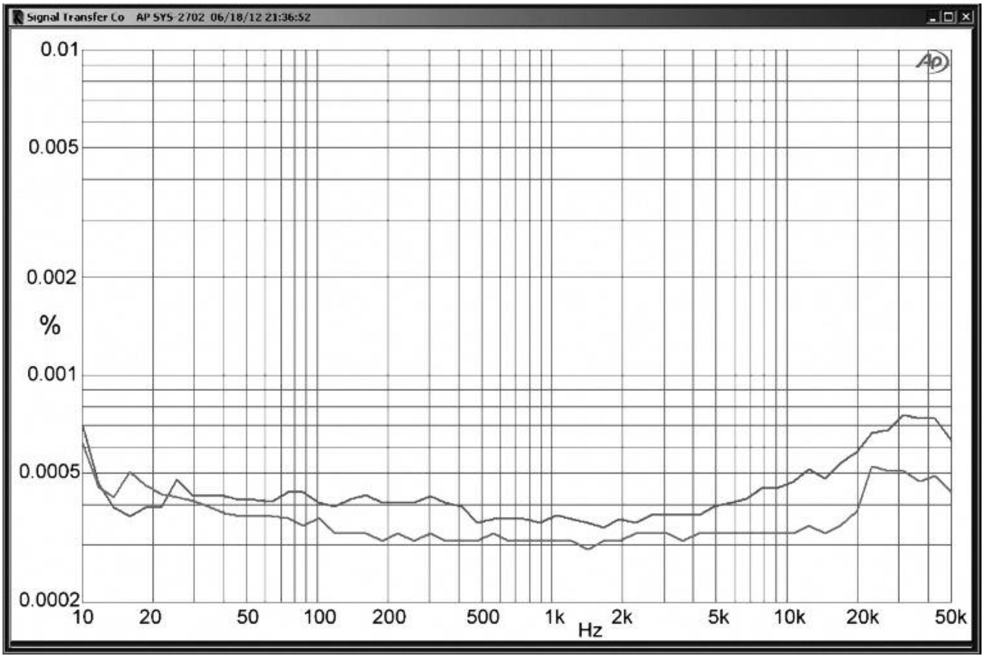

Figure 12 Volume-control distortion at max gain. Upper trace is volume-control output; lower trace is testgear reference output. (GenMon) 9 Vrms output.

bution is negligible at higher settings, but becomes more significant as the volume is turned down, until at a setting around Mk 3, the contributions are equal. The gain here is 25 dB below the +10 dB maximum, and the total noise out only -116 dBu. There does not seem to be a pressing need to make the volume stage quieter, which is a pity as it would be relatively straightforward. If we want to reduce noise then the tone-control stage is the obvious candidate for development.
It would certainly be possible to do this. For lower noise in Tone-Cancel mode, the 5 kΩ boost/cut pots could be scaled down to 1 kΩ, as in the Elektor preamplifier, and the extra opamps required to drive the resulting low impedances would give partial cancellation of their voltage noise. For lower noise with the tone-control active, which is more useful for most of us, the impedances in the HF and LF paths could also be scaled down by 5:1; a bigger ratio is difficult due to the problems of sourcing dual-gang pots with a value of less than 1 kΩ. The HF path is the priority as it generates much more noise than the LF path. Such a modified tone-control would however be a formidable piece of electronics, with some dauntingly large polypropylene capacitors, and it would probably be hard to reduce the noise by more than 6 dB before the number of opamps required got out of hand.

A little-known disadvantage of the Baxandall configuration, and indeed of any active-gain control, is that the output noise does not go to zero when the volume control is set to Mk 0. It is very low here, at -121.3 dBu, and much quieter than a fixed-gain amplifier following a passive volume control, but it is not zero. The shunt amplifiers in the active-gain stage work at a noise gain of one when volume is set to Mk 0, so the voltage noise of the opamp(s) will always appear at the output, even though the signal gain is zero. The only obvious way to solve this is to combine an active-gain stage with a final passive attenuator, to form a distributed volume control system. With the passive attenuator at zero there is a short-circuit across the output and no noise, unless you want to get pedantic about the Johnson noise of the internal wiring (of the order of -155 dBu, I would estimate). This does of course require 4-gang pots for stereo. Distributed volume techniques using two passive controls with a fixed gain stage in between have been used in high-end Japanese preamplifiers in the past.

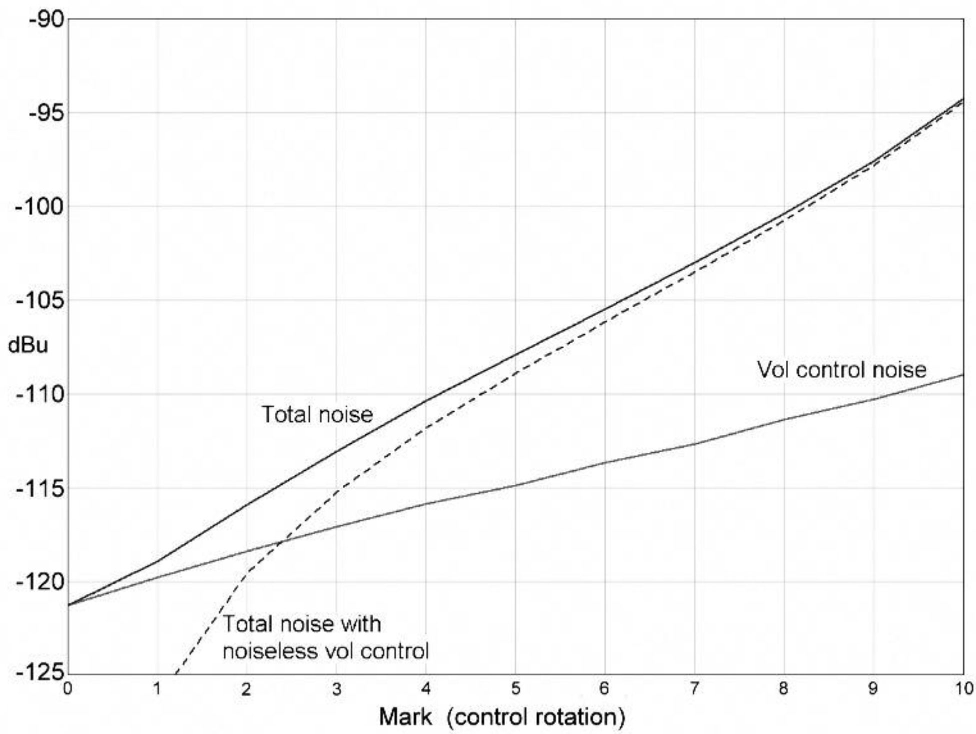

Figure 13 Total noise output and volume control stage noise versus volume control setting. The dotted line shows the noise output with a noiseless volume control.

The overall distortion performance of the complete preamplifier is summarised in **Figure 14**. It keeps below 0.001 % from 10 - 20 kHz at 9 Vrms, which is well above normal operating levels. The volume was set to Mk 6 (gain -2 dB) to keep a high level throughout the signal chain.

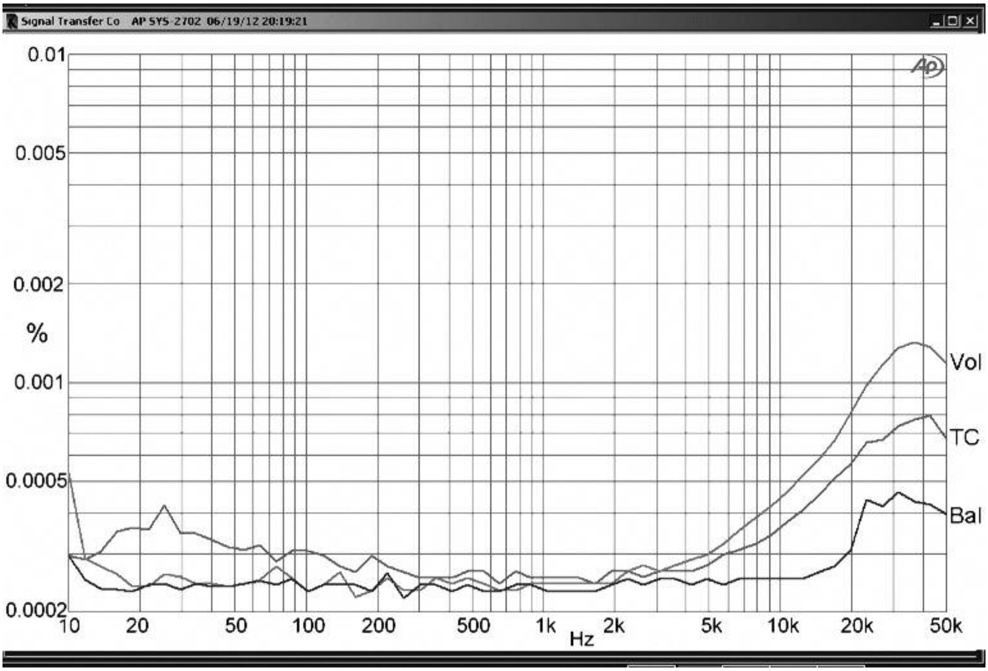

Figure 14 Distortion of complete preamplifier at 9 Vrms, measured at the outputs of the balanced input, (Bal) the tone-control, (TC) and the volume-control. (Vol) Balance mid, tone-control in but flat, volume = Mk 6.

Variations

You will note that in Figure 11 there is a spare LM4562 section left floating helplessly in space. This could be used as an inverter to provide a fully-balanced output, which raises the signal level over the interconnect by 6 dB and so gives more immunity from ground noise when used with a suitable input [6]. Alternatively, the section could be added to the volume control in parallel with the existing shunt-feedback gain stages to further reduce noise.

A CMRR trim preset could be added to the input/balance stage, with one of R11, R12 or R13 in Figure 3 replaced by a preset in series with a fixed resistor. This allows the CMRR to be trimmed to better than -80 dB up to at 500 Hz, and 70 dB at 2 kHz; it falls off at higher frequencies with opamp open-loop gain. Sophisticated testgear is not required as you are simply tuning for minimal signal; any sort of oscilloscope should be adequate.

It would be possible to build the preamplifier using 5532s rather than LM4562s, without changing any other components. The noise will be greater, and distortion will inevitably be somewhat higher, but I think that the loading conditions are light enough to allow this to work reasonably well.

In conclusion

What started as a simple showcase for the improved variable-frequency tone–control has evolved into a complete preamplifier design, with new material on both balanced line inputs and active volume-controls. Hope you like it.

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Electrolytic capacitor distortion

While writing this article, I decided to take a more penetrating look at those electrolytic capacitor values. It might be that some of them were grossly over-sized and could be smaller and cheaper types. Electrolytic capacitor distortion is examined in Small Signal Audio Design, [4] where I reached the tentative conclusion that a criterion of less than 80 mV rms signal voltage of across the capacitor would make the THD unmeasurably low. The circuit of figure 7 was simulated with the four combinations of max/min LF frequency, and maximum cut/boost, and a 1 Vrms input. The voltage across each of the capacitors was then evaluated.

As expected, C14 and C15 in the HF path never see more than a few hundred microvolts across them, and it appears that C15 in particular could be substantially reduced in value. In the end I decided not to spend the time working out how much as the cost saving is trivial, and it helps with purchasing to keep as many components as possible the same value.

In the LF path, the worst case signal voltage across C8 is 2.6 mV at 55 Hz (max LF freq, max boost) and the worst case for C11 is 6.7 mV at 57 Hz. (max LF freq, max boost) If we scale these up for a maximal 9 Vrms output at 50 Hz, (which means a 3 Vrms input) the worst case overall is 20 mV across C11, which is comfortably less than our 80 mV criterion.

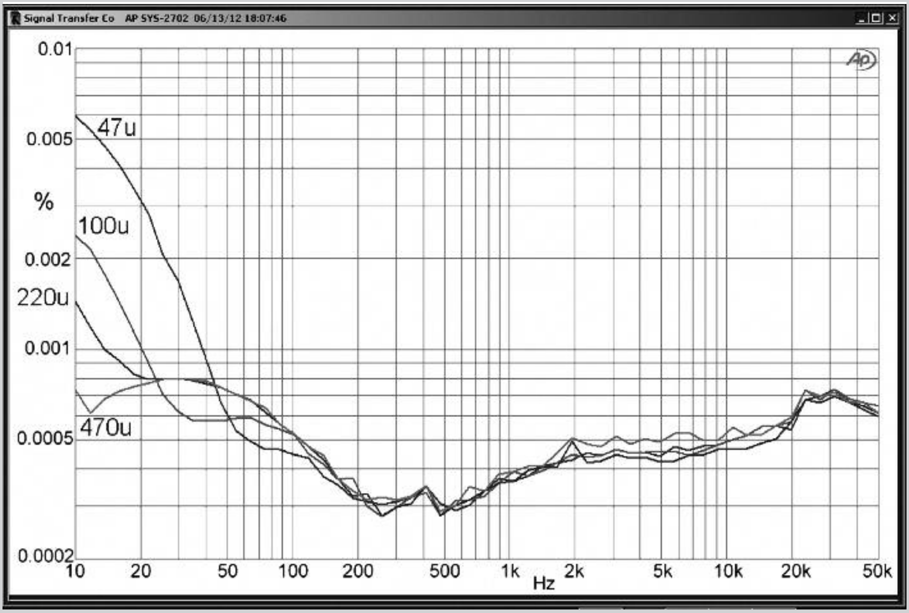

Figure A Tone-control distortion with varying values of C16. 9 Vrms output; max LF boost, max LF freq.

In the original 1996 design, the worst offender was in neither the HF or LF path; it was the 220 uF return capacitor C16, which had 75 mV across it at 10 Hz (max LF freq, max boost). This is a result of the relatively high turnover frequency of R22 and C16 which is 1.29 Hz. With a maximal 3 Vrms input the voltage across C16 scales up to 222 mV at 10 Hz, (it falls from that value as frequency increases) and we can expect added distortion at low frequencies as shown in **Figure A**. When we do the actual measurements we find that the distortion increase is very modest with a 220 uF capacitor, but the situation is definitely improved by increasing it to 470 uF. The results for 47 uF and 100 uF are also given as a Dreadful Warning against skimping on coupling capacitor size.