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DC Servos and Digitally-Controlled Microphone Preamplifiers

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ABSTRACT

Microphone preamplifiers for professional audio applications require a very wide range of gain and low noise in order to provide a high-quality interface with the vast number of available microphones. In many modern systems the preamplifier gain is controlled indirectly via a digital interface in discrete steps. Often dc servo amplifiers are employed as a means of keeping the dc gain fixed to avoid large changes in output offset voltage while the audio band gain is varied. The resulting highpass filter response varies substantially as a function of the preamplifier gain. We investigate the frequency and time-domain effects of this. We also investigate several approaches to minimize these effects.

1. INTRODUCTION

Microphone preamplifiers used in professional audio applications are expected to provide appropriate signal conditioning for a very wide variety of microphones. While “professional” microphones generally share the characteristics of having balanced outputs and source impedances substantially less than 1 kohm, the sensitivity of a sample of 43 available models was found to vary over a 37 dB range [1]. This, along with the wide range of applications, from a microphone placed just inside the resonant head of a rock drummer’s bass drum to one placed ten feet above a solo violinist, results in a very wide range of signal levels that must be amplified with minimal added noise or distortion.

In modern digital audio systems, the analog microphone preamplifier serves as the primary interface between the microphone and an analog-to-digital converter (ADC), with all further signal processing performed in the digital domain. Since user control over the digital signal processing (DSP) is necessarily digital, adding digital control over the microphone preamplifier itself yields a uniform interface for the end user, as well as potential additional features such as storage of previous settings and remote location of the actual analog preamp.

Most examples of digitally-controlled microphone preamplifiers allow gain adjustment in discrete steps rather than over a continuous range since such an approach is much more amenable to maintaining low input-referred noise. (The reasons for this are beyond the scope of this paper.) The currently available integrated circuit solutions [2], [3], [4], [5] all include an on-board DC servo amplifier to maintain the offset gain (defined as the gain between the input offset voltage of an amplifier to the amplifier output, also often referred to as the noise gain) of the preamplifier at a constant value regardless of the ac gain. This approach is an alternative to simply ac-coupling the amplifier's feedback network. In this paper we will explore some of the tradeoffs involved with the servo approach.

2. MAINTAINING CONSTANT OFFSET GAIN

2.1. AC-Coupling the Feedback Loop

One of the most common topologies for professional microphone preamplifiers is the three-opamp instrumentation amplifier. The "front end" of this circuit, which typically provides most, if not all, of the signal gain, is shown in Figure 1. Note that while there are certainly other topologies used, the following analysis generally applies to them as well. Further, the "opamps" A_1 and A_2 may be of many types, including current- or voltage- feedback amplifiers in discrete or integrated form. Since the circuit shown in Figure 1 has unity common-mode gain, it is typically followed by a stage that will reject common-mode signals (the third opamp) but this is not germane to the discussion at hand.

If the open loop gain of amplifiers A_1 and A_2 is large enough, for frequencies well above the highpass filter formed by the C_C/R_B networks, the circuit has a closed-loop differential signal gain A_d equal to:

$$A_D = \frac{v_{out}}{v_{in}} = 1 + (2R_F / R_G)\,. \qquad (1)$$

While dc input voltages (including phantom power voltages) are blocked by the C_C capacitors, $V_{OS,}$ the input offset voltage (the difference between the individual input offset voltages of amplifiers A_1 and A_2), is amplified by the differential gain. A 1 mV input offset voltage will result in a 1 V output offset when the amplifier is set for 60 dB of gain. The output offset, of course, varies directly with gain. Moreover, when the gain is varied the change in offset can become audible. This can take the form of a "woof" in the case of a continuous change in gain (such as when resistor R_G is a variable resistance) or ""clicks" and "thumps' when the gain is varied in discrete steps.

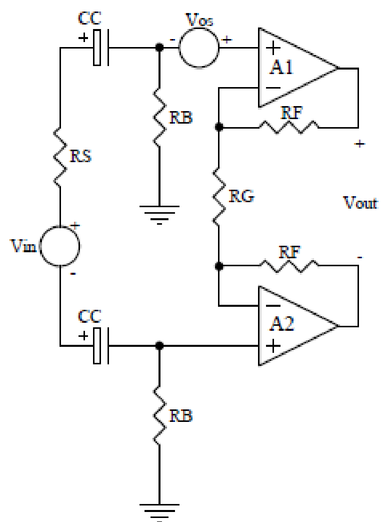

Figure 1 - Preamplifier front end

The simplest approach to preventing audible artifacts with gain changes is to ac couple the feedback loop by adding a capacitor C_G in series with R_G, as shown in Figure 2.

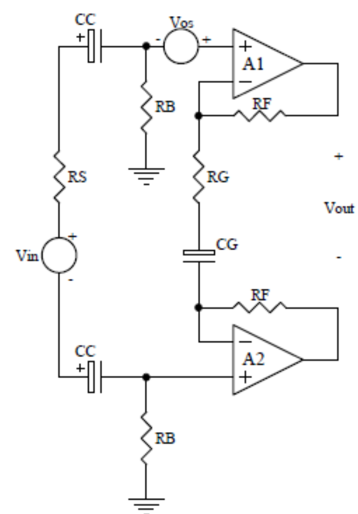

Figure 2 - Front end with ac-coupled feedback

The differential offset gain, $V_{OUT}/V_{OS,}$ is now unity (at DC) for any values of the resistors in the feedback network. The overall frequency response for both the differential offset gain and the differential signal gain is a first-order shelving highpass with a pole at:

$$f_p = \frac{1}{2\pi R_G C_G}, \tag{2}$$

and a zero at:

$$f_z = \frac{1}{2\pi (R_G + 2R_F) C_G}. \tag{3}$$

The differential gain is unity at low frequencies and rises at 20 dB per decade between f_z and f_p to the value in equation (1). (Note that we have ignored the effects of the input coupling capacitors C_C for the moment. Since they are isolated from the feedback network, their effects on the response may be analyzed separately.) This approach is very effective at eliminating audible artifacts during gain changes. Its drawbacks are mainly in the physical implementation. In order to minimize the thermal noise contribution of the feedback network, the resistances must be kept relatively low. At high gains the value of R_G will often be on the order of 10 ohms. In order to maintain response at the low end of the audio band C_G must be quite large – often a few thousand microfarads. While this capacitor need not have a high voltage rating since little voltage is ever developed across it, it can still be one of the largest components in the preamplifier circuit. This, along with its electrical location in the circuit (essentially right at the preamplifier inputs) makes it vulnerable to picking up electro-magnetic interference from sources such as nearby digital circuitry or switch-mode power supplies.

2.2. DC Servo

In order to overcome these shortcomings, many designers choose to employ a dc servo amplifier to maintain constant dc offset gain. This usually takes the form of an integrator circuit with its input monitoring the preamplifier output, and its output driving the preamplifier input as shown in Figure 3. The fully differential integrator A_3 in Figure 3 is one of several possible implementations. By inspection, assuming high enough open loop gain, we see that A_3 will force the differential output voltage V_{OUT} to equal its own input offset voltage V_{OSB} at dc regardless of the preamp gain.

This approach has the advantage of allowing the use of smaller capacitors if the impedances around the servo amplifier are kept high. The output of the servo amplifier is typically fed to the preamplifier input via a substantial voltage divider (the R_D/R_B networks in Figure 3) in order to minimize the noise contribution of the servo amplifier, so any interference picked up by the servo amplifier is attenuated as well.

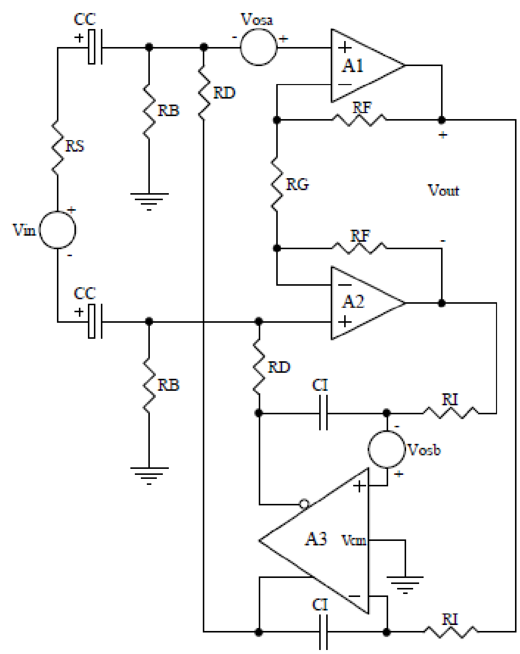

Figure 3 - Front end with dc servo

In order to analyze the effects of the addition of the servo on the differential gain of the circuit, it is useful to examine its “half-circuit” as shown in Figure 4. This simplified circuit will give identical results to the circuit in Figure 3 for all differential behavior, which is what we are interested in.

3. TRADEOFFS IN THE SERVO APPROACH

3.1. Frequency-Domain Effects

Since the servo circuit of Figure 4 couples directly to the preamplifier input network including R_B, C_C, and R_S, we cannot treat their frequency response effects separately. The overall frequency response from $V_{IN}/2$

to $V_{OUT}/2$ is that of a second order highpass filter, the characteristics of which vary greatly with the passband signal gain set by R_F and R_G. The natural frequency f_0 and Q factor of the highpass are:

$$f_0 = \frac{1}{2\pi}\sqrt{\frac{A_D}{(R_D + \frac{R_S}{K})R_I C_C C_I}} \tag{4}$$

and

$$Q = \frac{1}{2\pi f_0 (\frac{R_I C_I}{K A_D} + R_S C_C)}, \tag{5}$$

where $A_D = 1 + (2R_F/R_G)$, the passband gain, and $K = R_B/(R_B+R_D)$. While these expressions are not greatly insightful, they do at least indicate that the natural frequency of the highpass filter is proportional to the square root of the passband gain A_D, and that the Q is also proportional to the square root of A_D as long as $R_1C_I/KA_D >> R_SC_C$, which is usually the case for typical designs.

An example and frequency response plot gives a more intuitive view of what happens as the preamplifier gain is changed. Figure 5 shows the simulated frequency response of the circuit in Figure 4 with the following component values: Rs = 0, R_B = 1.2 kΩ, C_C = 47 uF, R_D = 470 kΩ, R_I = 100 kΩ, C_I = 470 nF, and R_F = 5 kΩ. Resistor R_G is swept to vary the passband gain A_D in 10 dB increments from 0 dB to 70 dB.

At gains below about 40 dB it is apparent that the poles contributed by the servo circuit and the input ac-coupling network (C_C and R_B in parallel with R_D) are separated on the real axis in the S plane. The initial 20 dB/decade rolloff with a -3 dB point of 2.8 Hz is due to the input ac-coupling network. The filter response transitions to a steeper 40 dB/decade rolloff at lower frequencies, but the point at which this occurs varies widely in frequency as the passband gain is varied. It is worth noting that, at 0 dB gain, the servo circuit highpass pole is at very low frequency, around .01 Hz. This has a significant impact on the filter's step response, which we explore in the next section.

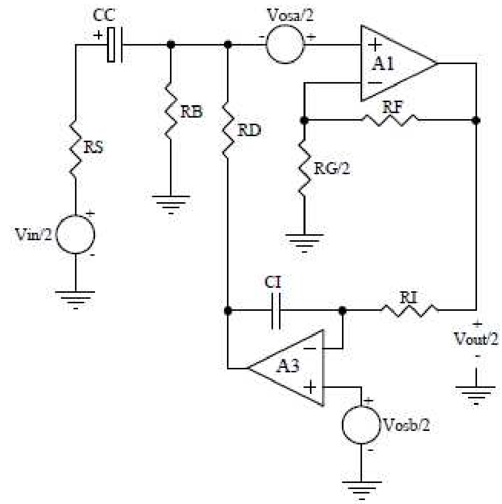

Figure 4 - Front end with dc servo "half circuit"

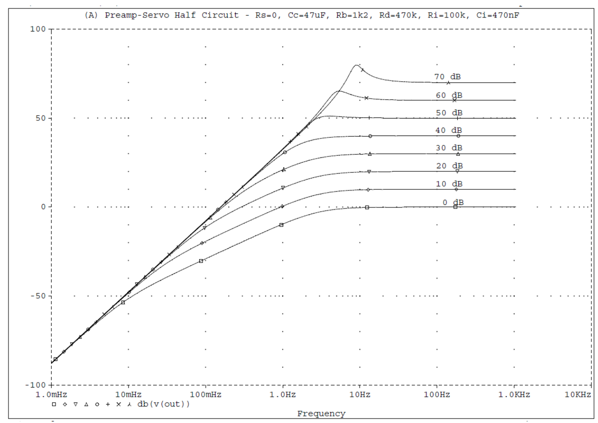

Figure 5 - Frequency response vs. gain of figure 4 circuit – (See text for component values)

At gains above 40 dB the pole-pair becomes complex and we see the effects of higher Q as gain increases above 50 dB. The 60 dB curve exhibits a peak of 5.2 dB at 5.4 Hz (Q = 1.75 from equation (5)), while the 70 dB curve has a 9.8 dB peak at 8.8 Hz (Q = 3.1).

Such peaking at infrasonic frequencies, particularly at very high gain settings, can make the preamplifier susceptible to overload from sources such as nearby traffic or HVAC systems, so this behavior should be avoided. Unfortunately, it is easily missed in most audio measurements that start at 20 Hz and go up in frequency from there.

From equation (5) we can glean that increasing either R_I or C_I, or decreasing K will decrease the Q factor at any given passband gain with roughly equal effectiveness. Figure 6 shows the resulting response when C_I is changed from 470 nF to 4.7 uF and R_D is changed from 470 kΩ to 240 kΩ. At 70 dB of gain the peaking has been reduced to 3.3 dB at 4.4 Hz (Q = 1.37), and the response at 60 dB gain is nearly Butterworth (Q =0 .77). Unfortunately this improvement in peaking has the effect of slowing the servo system. Comparing the response curves for gains below 40 dB in Figure 6 to those in Figure 5, we can see that the initial rolloff due to the input ac-coupling network has remained fixed, while the transition to the 40 dB/decade slope has moved further down in frequency.

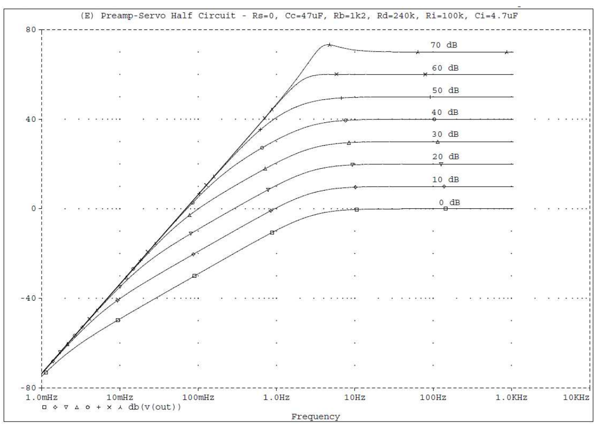

Figure 6 - Frequency Response vs. Gain of Figure 4 Circuit – C_I = 4.7 µF, R_D = 240 kΩ (See text for other component values)

3.2. Time-Domain Effects

Figure 7 shows the "half-circuit" from Figure 4 with the component values used to produce Figure 6. Switch S_1 has been added to vary the passband gain from one gain setting to a higher one by connecting R_{G2} in parallel with R_{G1}, as might be done in a digitally-controlled gain scheme.

As stated earlier, assuming ideal characteristics for amplifiers A_1 and A_3, in the steady state servo amp A_3 will force the output of A_1 to equal its input offset voltage $V_{OSB}/2$. If S_1 is initially open and R_{G1} is an open circuit, A_1 is at unity gain. A_3 must force the junction of R_B and R_D to a voltage equal to $V_{OSB}/2$ – $V_{OSA}/2$. If resistor R_{G2} has a value of 5 Ω, when switch S_1 is closed, the passband gain goes from 1 (0 dB) to 1000 (60 dB). At this instant the voltage at the output of amplifier A_1 will jump from $V_{OSB}/2$ to $1000*V_{OSB}/2$. It will decay back to $V_{OSB}/2$ with a time constant commensurate with the 60 dB highpass response shown in Figure 6. This time response for the circuit in Figure 7 is shown in Figure 8 where S_1 closes at t = 10 msec, and $V_{OSB}/2$ is 1 mV. The dc level does not settle for more than 200msec.

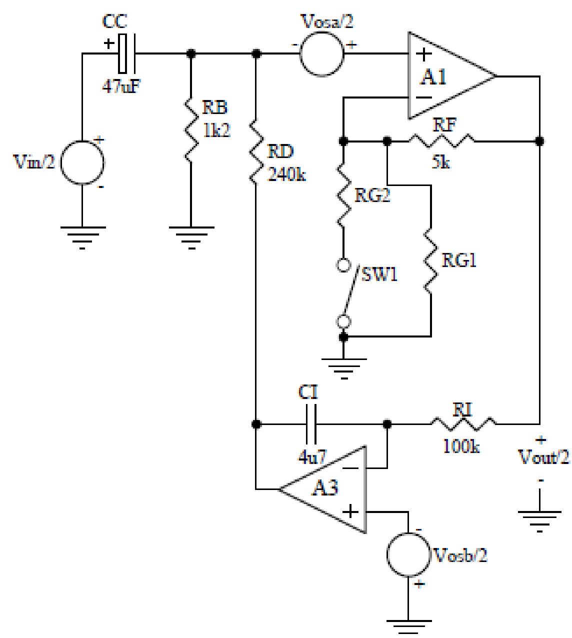

Figure 7 - "Half circuit" with switched gain

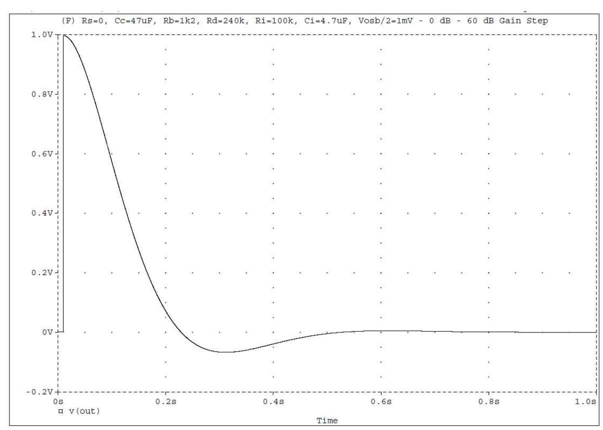

Figure 8 - 60 dB gain-step response, $V_{OSB}/2$ = 1 mV

Clearly a 1 V spike at the preamplifier output is unacceptable unless the system output is muted at some point further along. For this reason controller software for digitally-controlled microphone preamplifiers is often written to prevent such large single-step gains changes. Instead, the gain is ramped in smaller steps in a closer approximation of an all-analog implementation using a potentiometer for gain control. However, the time-domain behavior of the servo circuit must still be taken into account when implementing such a scheme.

Figure 9 shows the time response of Figure 7 with R_{G1} open circuited, this time with R_{G2} = 12.1 k Ω and $V_{OSB}/2$ = 1 mV. Here the gain switches from 1 (0 dB) to 1.413 (3 dB). The voltage at the output of amplifier A1 starts at 1 mV, as before, and when switch S_1 closes at t = 10 msec, the voltage jumps to $1.413 * V_{OSB}/2$ = 1.413 mV. While the instantaneous dc voltage change is much smaller with this smaller gain step, the settling time is much larger, because it is now governed by the much slower dynamics of the low-gain frequency-response curves seen in Figure 6. The output voltage does not return to near its final destination for over *4 minutes*.

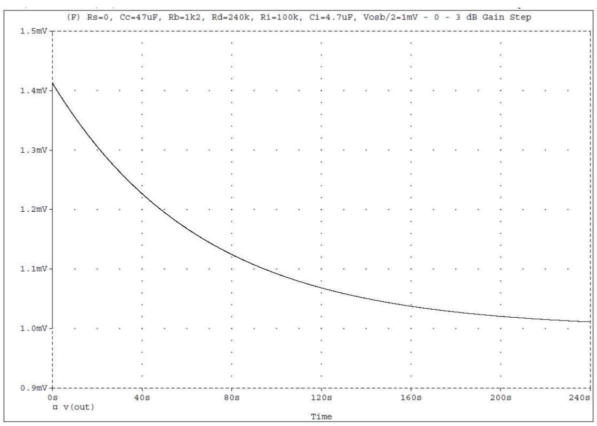

Figure 9 - 3-dB gain-step response, $V_{OSB}/2$ = 1 mV

While a 400 µV spike at the preamplifier output may not be much of a disturbance, if the gain is ramped further without waiting for settling between each step, the net output offset voltage change builds. In the case of successive 3-dB gain steps, the output voltage will change by 1.414 times the output voltage at the instant each gain change is made. This is illustrated in Figure 10. Here additional switches have been added to the Figure 7 circuit to create 20 3-dB gain steps from 0 dB to 60 dB. The first step from 0 dB gain to 3 dB gain occurs at t = 25 msec, and successive 3-dB steps occur at equal 25 msec intervals. $V_{OSB}/2$ is again 1 mV. This is not a lot better than the single step, with a maximum change in offset of 680 mV.

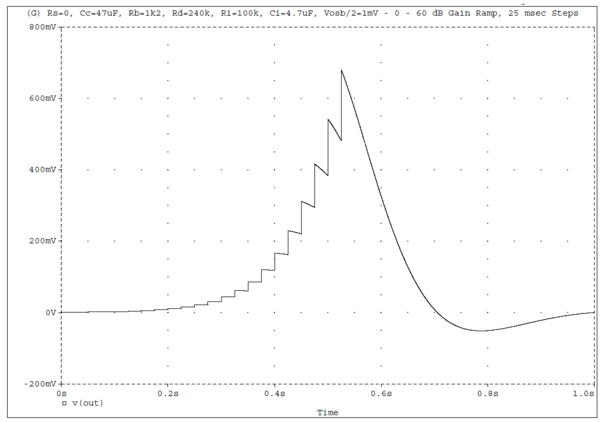

Figure 10 – 60 dB ramp in 3-dB steps 25 msec apart

Figure 11 shows how slowing the ramp down by using 250 msec between steps reduces the maximum output offset voltage change to 62 mV but the change takes 5 seconds to complete.

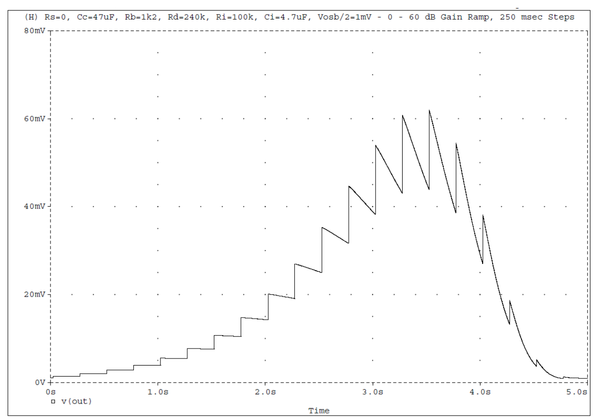

Figure 11 - 60 dB ramp in 3-dB steps 250 msec apart

3.3. Mitigation Approaches

3.3.1. Variable Step Spacing in Time

Note that after the step at 3.525 seconds in Figure 11 (the 45-dB step) the output voltage begins decreasing. The servo time constant at these higher gains is small enough that the voltage decays by more than a factor of 1.413 between transitions. This suggests that we might be able to optimize the ramp by tailoring the time between transitions according to the gain setting.

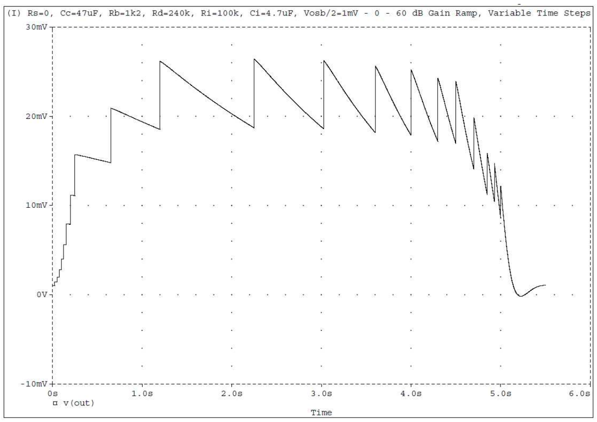

Figure 12 – 60 dB ramp in 3-dB steps with variable timing

Figure 12 shows the results of an empirical attempt at this approach fitted within a 5 second total time to allow comparison with the constant-interval ramp of Figure 11. The first steps at the lowest gains occur at short intervals of 25 msec, since the very long servo time constant at these gains means that there is little benefit to waiting longer. At intermediate gains as the servo time constant gets smaller, the intervals between steps are increased. After about 33 dB we take advantage of the faster time constant to keep the step changes in output relatively constant while decreasing the time between steps. Compared to the response in Figure 11, the maximum output offset shift is reduced from 62 mV to 26.4 mV, and the maximum step change is reduced from 18.1 mV to 12.9 mV. It should be noted that these output voltage changes will all be proportional to the servo amplifier's input offset voltage. The 1 mV value for V_{OSB2} used here as an example corresponds to a 2 mV input offset voltage for the full differential circuit, which is somewhat greater than the maximum specification for the integrated circuit implementations cited in references [2], [3], [4], and [5], but certainly not uncommon for many of the opamps typically used for discrete implementations.

3.3.2. Two-Speed Servo

Since the servo time constant is responsible for the slow settling after each discrete gain change, we investigated making the servo time constant smaller just for the duration of the gain-change ramp. Figure 13 shows a possible implementation for the "half circuit" introduced in Figure 7. Switch S_2 and resistor R_{I2} are added to speed up the servo by approximately a factor of 20. It should be noted that at 60 dB gain, with S_2 in its speed-up position (closed), the Q of the highpass filter formed by the servo circuit and input coupling network is 3.45. This would result in a substantially peaked response with attendant ringing in the time-domain if the circuit were left in this state. For the following example S_1 is turned on coincidentally with the first gain step, and is turned off coincidentally with the final gain step in the ramp. The intervals between gain changes are staggered, starting at 250 ms and gradually decreasing to 50 msec over the course of the 3-second long ramp.

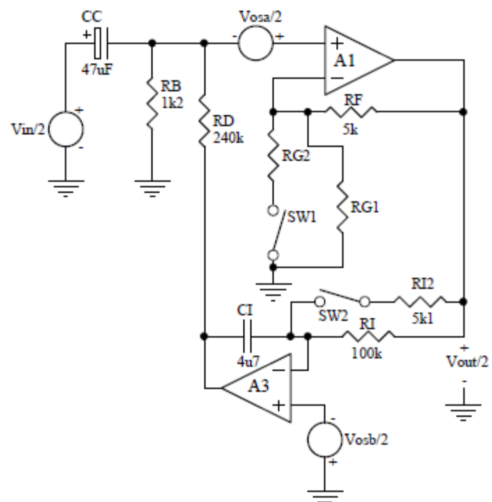

Figure 13 – "Half-circuit" with 2-speed servo

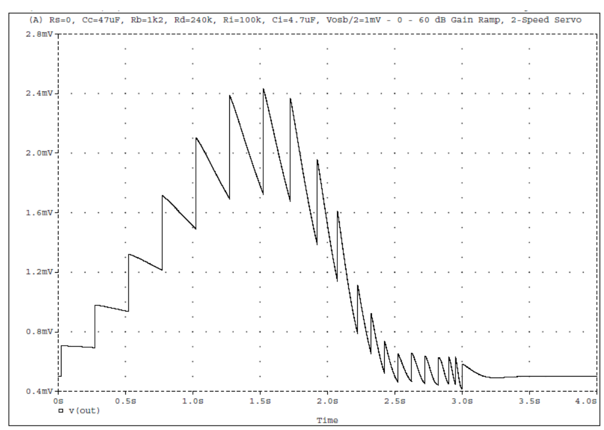

Figure 14 – 60 dB ramp with 2-speed servo

Figure 14 shows the time response at the output of amplifier A1. The maximum overall output offset shift is 1.9 mV, and the maximum step change is 700 µV. This approach certainly seems to bear further investigation.

3.3.3. Back to AC-Coupling

As we stated in section 2.1, ac-coupling the feedback loop is very effective at suppressing any audible artifacts associated with gain changes. For applications that demand the fastest gain changes with minimal artifacts, this may be the preferable approach. It's worth noting that the two digital gain-control devices cited in references [2] and [3] can be used without their internal servo, and ac-coupling of the feedback loop can be arranged. Consult the manufacturer for application details.

4. CONCLUSIONS

The use of dc servos to minimize changes in output offset during changes in gain of digitally-controlled microphone preamplifiers has some desirable advantages, but also some drawbacks when compared with the venerable approach of using capacitors to ac-couple the amplifier's feedback loop.

The combination of the servo amplifier and the input ac-coupling network typically used to block phantom-power voltages creates a second-order highpass filter that can result in frequency response peaking below the audio band. Component values should be chosen for the desired range of gains to ensure reasonable response at the maximum gain setting, keeping in mind that too much damping at high gains will lead to very slow response at low gains.

The servo amplifier maintains a constant preamplifier output offset voltage only in the steady state. It produces a transient change in offset for each gain change equal to the instantaneous output voltage multiplied by the gain change. The time required to settle back to the steady-state value is a function of the gain, and can be very slow at low gains. This can lead to the requirement for slow ramp times for large gain changes.

We have shown two approaches to mitigate these time-domain effects. The first is to adjust the time interval between gain steps appropriately for the time constant at the current gain. This approach shows modest improvement over a constant time interval for all gain steps. The second approach is to implement a two-speed servo that is sped up during gain changes. This approach shows a more dramatic improvement and is deserving of further investigation.

Finally, the most demanding applications may be best served by ac-coupling the preamplifier's feedback loop.

5. ACKNOWLEDGEMENTS

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6. REFERENCES

[1] Gary Hebert, Joe Lemanski, Interfacing Digitally-Controlled Microphone Preamplifiers to A/D Converters, presented at the 133rd AES Convention, October 2012

[2] THAT Corporation, THAT 5171 High-Performance Digital Preamplifier Controller IC Datasheet, Document 600133 Rev. 07

[3] THAT Corporation, THAT 5173 Audio Preamplifier Digital Controller IC Datasheet, Document 600155, Rev. 02

[4] Texas Instruments, PGA2500 Digitally Controlled Microphone Preamplifier Datasheet, Rev. A, SBOS289A − November 2003− Revised December 2003

[5] Texas Instruments, PGA2505 Digitally Controlled Microphone Preamplifier Datasheet, Rev. B, SBOS396B–March 2009–Revised June 2009