HIGH PERFORMANCE CURRENT REGULATORS REVISITED

A Special Note: This test data is still very preliminary, and subject to changes. It is being shared for purposes of commentary and critique.

Several readers wrote in with comments and questions related to the use of the LM317 and MOSFET current sources within parts one and two of “Sources 101”[1, 2]. These experiences (not all of which were good) led to the following discussion of two “all-weather” current sources/sinks. These current sources are suitable for virtually any type of audio application—that is, either power related or within the signal path, at currents from just a few to 100mA.

TWO “ALL-WEATHER” CURRENT SOURCES/SINKS USING MOSFET CASCODES

The key development that came about deals with the improved MOSFET cascode current sources of **Fig. 1**, both using the DN2540 depletion mode MOSFETs. There are really a couple of parts to this story-within-a-story. One is that it was in part stimulated by some e-mail exchanges, the first of which used a MOSFET current source in a power amp. More importantly however, some later e-mail discussions revealed disappointing performance with the original *Fig. 13C*[2], *when this circuit was used as an active load for a triode preamp stage*. This was traced to an excessive white noise output, which in turn was due to the use of the LM317 within Fig. 13C. In this circuit the LM317 establishes the basic current operation of the circuit, insofar as noise properties.

Although my own original use of the Fig. 13C circuit was within a power supply regulator, I made a definite mistake in not establishing a caveat on noise vis-à-vis the LM317, *if used within signal path circuits*. Its noise level is such that while it might be OK for use as a push-pull power stage cathode bias (i.e., high level), it definitely isn’t suited for lower level stages. In fact, the feedback I got was this: The Fig. 13C circuit was indeed OK in the high level tube applications, but not in preamps—*due to a high residual hiss level*. I apologize for not anticipating this type of use, and regret any inconvenience caused.

Fortunately, there is a simple and very attractive alternate solution—just use one of the two following advanced DN2540 cascode circuits. And, the second part of the story comes from the stellar performance achieved with these cascodes. It took some special means to see just how good they really are!

The schematic of the new circuits is shown in **Fig. 1**, which consists of a conventional cascode (left), plus what I think is a very useful wrinkle, a bias-multiplying cascode (right). The left circuit isn’t really new, because it is a relatively straightforward application of FET device cascoding, and this setup is similar to what my correspondents used. As discussed[1], it is usually desirable in this type of circuit to have the Vgs (off) of the upper device (M2) be specified as substantially higher than the lower part (M1). But, if two identical parts are used just as shown, the D-S voltage seen by M1 becomes the Vgs of M2—not always best for highest rejection. Nevertheless, with a simple circuit such as this, and no other high Vgs (off) family parts to use, you use what’s available, the same part for M1 and M2. And, it still works, quite well, in fact.

The slightly more complex circuit to the right adds a second pair of resistors, which provide a DC gain for the gate bias of M2, as:

$V_{DS}(M1) = V_{GS}(M2) * (1 + R3/R4)$

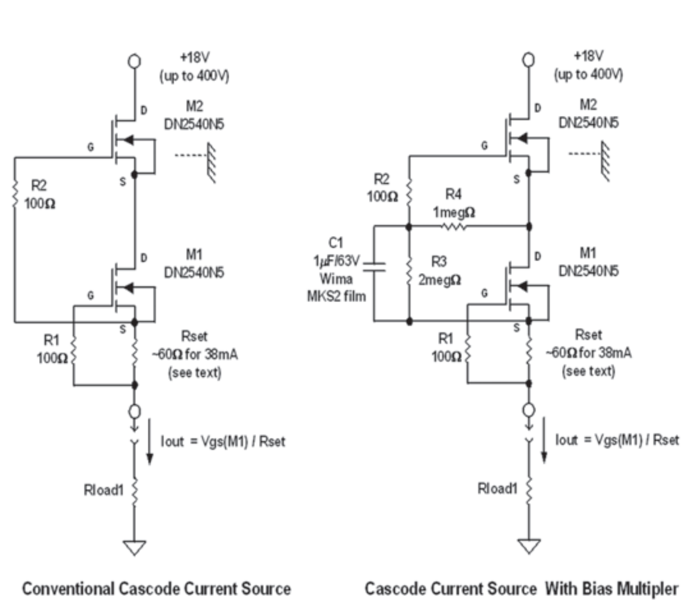

FIGURE 1: DN2540 Cascode MOSFET Current Sources.

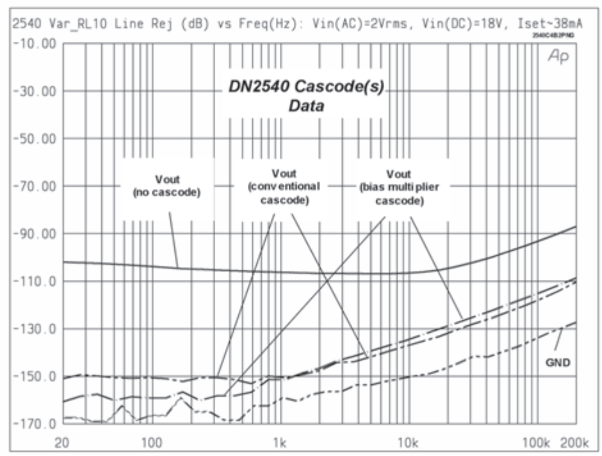

FIGURE 2: The performance of *Fig. 1* DN2540 MOSFET current sources for both conventional and bias-multiplied modes is excellent. With a new higher resolution setup, rejection of the conventional cascode is ≅ 150dB, while that of the bias-multiplier cascode is ≅ 160dB at low frequencies. For reference, a re-tested plot of no-cascode DN2540 operation shows performance comparable to original data (Fig. 12B from Sources 101 Pt. 2).

So, if the DC bias of M2 is, for example, 2V, the resistor ratio shown increases it to 4× this, or 8V. This multiplied bias then becomes the V_{DS} of M1. This higher bias should enhance the M1 rejection characteristics for input changes, as it moves M1 V_{DS} further into its pentode region. The capacitor C1 bypasses R3, for lowest overall noise.

This approach allows noiseless biasing of this cascode, i.e., selective biasing without the use of Vs dividers, or batteries. When I first tested the bias-multiplying cascode, I couldn't see any improvement vis-à-vis the conventional type (left). Both circuit forms showed performance virtually indistinguishable from the setup noise floor.

But, I had also made some further setup improvements, which have a direct bearing on measurable results for these circuits. Here's what I did to change the test setup from that previously described[1].

1. Raised the test circuit drive to the DUT to Vin(AC) 2V RMS, which adds 6dB of dynamic range vis-à-vis the original 1V RMS setup.
2. Changed test load resistor Rload1 to 10Ω, a step which adds 20dB more sensitivity to the Vout error voltage. To be valid, this step assumes that all Zs being measured are >> 10Ω, which they are in all cases thus far (in this DN2540 case, it turns out that the measured Z is on the order of 100MΩ or more).
3. Created new Audio Precision software to do a rescaling of the output data files, so that all measured data points in a given sweep are reduced by 20dB (the ratio of the Rload1 change, or 10:1). This effectively brings the setup calibration back to a 1Ω load reference. It also offsets the displayed dynamic range 20dB, downward. The total *display* dynamic range shown is still 160dB. It effectively adds 20dB more to the lower end of the measuring dynamic range, as the noise is effectively reduced relative to the Vout signal. In these plots, the 0dB reference trace isn't shown, but it is/was calibrated prior to a test run. The working test range was also calibration-checked at low levels, with a 10MΩ resistor, resulting in a nice flat line at –139.9dB (just about as it should be, with a measured -140dB being ideal).

The resulting test plots of **Fig. 2** are for the Supertex DN2540N5 TO-220 part(s) operated at 38mA, a current level that corresponds roughly to similar data[2]. It is easy to see the additional dynamic range here, which extends to –170dB. The plot which is repeated for the non-cascode 2540 operation is almost identical to the original 2540 plot at a 30mA level (Fig. 12B[2]).

A focus of this particular experiment was to discern whether the bias-multiplying cascode version (**Fig. 1**, right) really did make a difference in terms of rejection, vis-à-vis a more conventional version (**Fig. 1**, left), where the upper FET's gate is tied directly to the lower FET's source. Lo and behold from **Fig. 2**, yes, it does seem to make a difference! The rejection is better with the extra DC bias, just as theory says it should. The amount is about 10dB, not an overwhelming amount, but also clearly an improvement. This difference might change to even more between the two operative modes, were M1/M2 to see additional drain voltage. Here each sees at the most about 9V, which isn't very high for a 400V part. Interestingly, I couldn't see this differentiation at all, with the older setup and Rload1 = 1Ω. With the load of 10Ω, it is easy to see.

SELF-GENERATED NOISE

All of the aforementioned tests are relevant to synchronous noise of a current source circuit, which is noise that is related to AC components of the unregulated power source. But, another important noise type is *self-generated noise,* which is the form generated by a device (or circuit) with a steady DC bias. It is this type of noise that led to the relatively poor performance of the LM317 cascode circuit in the preamp stage. Generally this type of noise—white noise—is relatively constant over a wide range of frequencies, and typically manifests itself (audibly) as hiss.

In a test series planned for a future *audioXpress* article, I tested several different current source circuits for self-generated noise, including the LM317 and DN2540, with a test DC operating current of 10mA. I measured the noise components of the output current with a 100Ω resistor, an appropriate scaling amplifier, a 10kHz low-pass filter, and an RMS responding voltmeter.

To summarize in a few words, the DN2540 at 10mA came in more than 50 times lower in noise vis-à-vis the LM317, for similar conditions. Expressed in terms of noise density in pA/√Hz, the LM317 was over 1000 pA/√Hz, with the DN2540 less than 14pA/√Hz. Note: *These figures couldn't be called exact, in particular for the DN2540, where the noise is on a par with the input amplifier of the test setup. However, the main point here is simply that the DN2540 is many times lower in noise than the LM317.*

CONCLUSION

So, these two MOSFET cascodes can be termed "all-weather" current sources, because they can operate as either current sources or current sinks. They are also all-weather from the standpoint of a tremendous current range capability, from just a few mA up to 100mA or more. Plus, of course, there is the wide voltage range, up to 400V. Suitable for a wide range of solid-state or tube uses, in power or signal circuits, the **Fig. 1** configurations are the most versatile of any discussed. I recommend them without any major reservations. Although the TO-220 package of the DN2540N5 could be a downside for low power uses, there is a handy alternate, the TO-92 DN2540N3 (same internal chip).

ACKNOWLEDGMENTS

Henk Huitema, John Tracy and Ben Veltmaat provided useful feedback on their experiences using these current sources, within original correspondences taking place in Sept 2007. My thanks to all of them. In their final setup, they all reported successful use of the **Fig. 1** (left) cascode. This cascode type has previously been discussed[5].

REFERENCES

1. Walt Jung, "Sources 101: Audio Current Regulator Tests for High Performance, Part 1," *audioXpress*, April 2007, pp. 10-23.

2. Walt Jung, "Sources 101: Audio Current Regulator Tests for High Performance, Part 2," *audioXpress*, May 2007, pp. 8-17.

3. Walt Jung, Response to correspondence from Thomas Bohley, Chris Paul, and John Popelish, Xpress Mail *audioXpress*, September 2007, pp. 56–58.

4. Walt Jung, Response to correspondence from Marc Whitney, Xpress Mail *audioXpress*, September 2007, pp. 58–60.

5. Ixys IXCP10M45s forum thread at: www.diyaudio.com/forums/showthread.php?s=9c690653b3b7ba542b30545a1556ab9c&threadid=102335&perpage=25&highlight=&pagenumber=4.

THE IMPASSE PREAMP

I really enjoyed reading "The ImPasse Preamplifier" (*aX* 2/09) by Stuart Yaniger. I particularly got a kick out of the section headings alluding to lyrics from one of my favorite rock songs, and to presidential campaign slogans as well. But I do have a few questions and comments.

The article lauds the low input capacitance that the medium mu 6SN7 affords a gain of 20 circuit. It cites 80pF with a 6SN7, and a whopping 300pF from the high mu D3a. That may be so, but I'd like to point out that the high mu 12AX7, with a 1.7pF grid-plate capacitance, would do better with this than either of the other triodes in a gain of 20 circuit. Its input capacitance would be only 34pF. If you don't like the 12AX7, there are many other high mu tubes with low grid-plate capacitance—the 6SL7 and 12AT7 come to mind.

The article also points out quite correctly that the CCS in the plate circuit helps to produce a gain stage with a terrific PSRR. The problem is that the phase inverter output stage has a lousy plate PSRR; it passes on just about the full noise of the supply. And this noise is strongly attenuated at the cathode, so there is no common mode cancellation by the differential F4 power amps. With the given inverter output stage, the best PSRR of the overall circuit would be obtained if the gain stage cut the power supply noise about in half.

Given a 350V DC supply, a gain stage of 20 with a plate voltage of 120V DC and a PSRR of 0.5 can be implemented simply with a 12AX7. Use a cathode resistor of 5k11, a plate resistor of 237K, and add a 392K resistor between the plate and ground. Tap off a grid bias about 1k5 from the cathode stage's grid. Advantages of this approach are lower input capacitance, a simpler circuit, and that only 0.22W are dissipated in the top plate resistor as opposed to 1.5W with the 6SN7.

While it is true that the AC voltages at these points would be identical, the impedances would certainly not be. The triode's cathode impedance is about 1/gm, while its plate impedance is much, much higher. (You might partially compensate for this by adding resistance in series with the article's C4, the cathode coupling capacitor.) But if you're going to match the inverter stage's cathode and plate resistors to 0.1%, you'll want to do the same with the input resistances of the F4s you're driving. Alternatively, you could compensate for your F4s mismatch in the inverter stage's resistors.

I have always thought that this tube should be considered as a possible audio standard. Its transconductance, mu and maximum cathode-heater voltage are higher than those of the 6DJ8, and it doesn't have the 6DJ8's unusual construction that is sometimes blamed for a strident sound. Is the 6KN8 still commercially available?

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Stuart Yaniger responds:

I thank Mr. Paul for his interesting letter and the very cordial and positive correspondence that it triggered.

His comments regarding system power supply rejection are very well taken and I agree with them. As a practical matter, with the regulator shown, system signal-to-noise is better than -100dB with respect to 20V, so the point that I made in the article and his clarification are more academic questions. But indeed, it does show the need for a reasonably quiet supply or some means of canceling supply noise if this is not the case.

Indeed, it is possible to reduce the gain, and hence the Miller effect, of a high mu triode by loading down its plate sufficiently and adding some degeneration. Unfortunately, the side effect is decreased linearity—a triode is most linear when the load line is horizontal (constant-current), and decreasing the effective AC plate load swings the load line more vertical. In the case of the 12AX7, because the plate resistance of the tube is so high, the situation is exacerbated—with a CCS plate load, the tube is superbly linear. With a load resistor only two or three times the plate resistance, the modulation of plate resistance with current swing is rather severe. In a quick test of Mr. Paul's proposed circuit, linearity was degraded by an order of magnitude. I should note that the CCS shown in the article only uses about $4 worth of parts.

The question of source impedances of a split load is a vexing one which generates much heat, and it is important to note my stated restriction that plate and cathode be loaded equally. Much of the issue depends on what you use as the definition of "source impedance." If it is defined as the equivalent Thevenin impedance derived from voltage change when a test load is applied to the output, indeed both the plate and the cathode show the same source impedance as long as the restriction is honored. Identical test loads must be applied to plate and cathode. If this is the case, it may easily be experimentally verified that both sources are identical and are approximately equal to the reciprocal of the tube's transconductance.

Adding a series resistor to the cathode leg will ensure that, with equal test loads, the source impedances will be unbalanced and it will break the symmetry—the frequency response with identical reactive loads (e.g., cable capacitances) will now *not* be equal on both legs. A good mathematical derivation of this slightly surprising result may be found in Morgan Jones's *Building Valve Amplifiers,* 3rd edition, on pages 405-406 (available from Old Colony Sound Lab, www.audioXpress.com). However, Mr. Paul has given me the opportunity to issue a warning that I neglected to make in the article—the cabling from each phase to the power amplifier must be the same! If the cable carrying one phase is longer than the other, the reactances will be different and the frequency response and balance will suffer disproportionately.

On the matter of resistor matching, as long as the input resistors of the F4 are more than a decade larger than the plate and cathode resistors of the output phase splitter of the Impasse, their tolerances are a decade less critical. Though it certainly won't hurt to match them more tightly, standard 1% values will work fine.

The 6KN8 is a very interesting 6DJ8 variant. Its gain is higher, but its linearity is somewhat lower when used as a voltage amplifier. This, of course, balances out when used with 100% degeneration, as is the case here.

Again, I thank Mr. Paul for the most interesting discussion and for the excellent and interesting contributions that he has made over the years. I was delighted and honored that he would read my article and take the time to analyze and comment on it. *aX*