In this, the second part of four articles, Kendall Castor-Perry introduces the driver circuit for what he calls the "Class i" output stage. He shows through detailed analysis that it can deliver perfectly resistive output impedance, through complete cancellation of transistor non-linearity.

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Introducing the Class i driver
We saw in part 1 that using a long-tailed pair form of power stage driver could give us a nice way of creating the small bias voltages needed to set the quiescent current. We also saw, though, that the linearity performance of such a stage is exquisitely dependent on the value of this bias voltage, and is never actually perfect. In this part, we'll look at a method for fixing this robustly.

If we don't want the current in the inactive half of the output stage to fall to zero, we need that half to transform topologically into a constant current source/sink with a predictable value. And the offset-equipped differential input amplifier discussed in part 1 immediately suggests a route to doing this. Consider the circuit of figure 4 compared to part 1’s figure 1. The base of the input transistor has been connected instead to the output terminal, driving a current out to a load.
Figure 4: UGB driver reconnected as a constant current source

The feedback transistor runs at \( m \) times the current in the input transistor. Because of the resulting built-in offset voltage of the imbalanced long-tail pair, the effect is to produce a circuit that is unwilling to deliver lower than a certain value of current, set by this offset voltage across \( R_e \). And that sounds like what we want out of half of our output stage; lots of current when we want current, and non-zero current (soaked up by the other half, of course) when we don't. There's just one problem, though - we lost our input terminal, in order to make this happen.

So, here's the clever part: just put another transistor back in, to act as an input path. We get the circuit shown in figure 5. The resultant input transistor 'doublet' is the characteristic signifier of the Class i driver. Depending on the relationship between the voltages at the two bases, the current-splitting in the long-tailed pair is controlled either by the input signal or by the connection back to the output node.
We'll presently see that the resulting circuit neatly hands off control from one half to the other as the load current direction changes. But what makes it special is that it actually does work exactly, and can deliver solid design equations that can be used for quantitative work. It is possible to dimension a circuit that not only essentially eliminates both the switching and transconductance modulation components of crossover distortion, but also works out of the box in production, over temperature, with absolutely no need for any trimming components or complicated control loops.

The output impedance is indeed purely resistive; under these admittedly rather ideal conditions, the output stage cannot introduce any current-dependent distortion. Only two parameters need to be chosen; the emitter resistor and a parameter $K$ (equal to the ratio of the mirrors in the collector feeds). In a practical discrete circuit, this ratio is set by using unequal degeneration resistors in the collector mirrors. Early effect in the mirror transistor can be neglected as long as several hundred mV is dropped across those resistors.

Patent searches in the late 1990s indicated that of the many investigators of what amounts to an ‘analogue OR gate’ operation required for this kind of current control [11], Nakayama [12] got closest to the Class i configuration. However, his patent (filed 1983, granted 1985) shows elaborate circuits of unnecessary complexity, and I feel that he missed the atomic elegance and ideal operation of the Class i doublet.

**Formal analysis of the Class i stage**

The circuit for analysis is based in figure 5a. To simplify the analysis for this article, we'll assume that the transistors in the long-tailed pair all have identical saturation currents. We'll assume no degeneration (the effect can be added later) and we'll also ignore any base current flowing in the attached power devices. Voltage $V_o$ is the drop across the stage when current is taken out of it, $V_{out}-V_{in}$.
We first write down expressions for $I_{c1}$ (flowing in the Class i input doublet) and $I_{c2}$ (flowing in the feedback transistor), which sum together to give the tail current $I_t$ (flowing through source $I_1$ in **figure 5a**):

\[
I_{c2} = I_s \exp \left( \frac{V_o + I_t R_E - V_E}{V_t} \right) \tag{1}
\]

and

\[
I_{c1} = I_s \exp \left( \frac{V_o}{V_t} \right) + I_s \exp \left( \frac{-V_E}{V_t} \right) \tag{2}
\]

with

\[
I_t = I_{c1} + I_{c2} \tag{3}
\]

$V_t$ is the thermal voltage, $kT/q$ for an ideal transistor with unit emission coefficient. Now from equation 2 we have

\[
I_{c1} = I_s \exp \left( \frac{V_o}{V_t} \right) + I_s \exp \left( \frac{-V_E}{V_t} \right) \]

\[
\therefore I_{c1} = I_s \exp \left( \frac{-V_E}{V_t} \right) \cdot \left( 1 + \exp \left( \frac{V_o}{V_th} \right) \right) \]

\[
\therefore I_s \exp \left( \frac{-V_E}{V_t} \right) = \frac{I_{c1}}{1 + \exp \left( \frac{V_o}{V_th} \right)} \tag{4}
\]

and we can substitute this back into equation 1:

\[
I_{c2} = I_s \exp \left( \frac{V_o + I_t R_E - V_E}{V_t} \right) \]

\[
\therefore I_{c2} = I_s \exp \left( \frac{V_o}{V_t} \right) \cdot \exp \left( \frac{I_t R_E}{V_t} \right) \cdot \exp \left( \frac{-V_E}{V_t} \right) \]

\[
\therefore \frac{I_{c2}}{I_{c1}} = \exp \left( \frac{I_t R_E}{V_t} \right) \cdot \frac{1}{1 + \exp \left( \frac{V_o}{V_th} \right)} \tag{5}
\]

The discussion on the UGB driver indicated that we could bias the collectors with either a current source or a mirror. Let’s assume the general case where we use both, i.e., the value of the current in the working collector (feeding the output device) is the sum of the output of a current mirror driven by the other collector (as shown in **figure 5a**) and a constant current $I_0$ (not shown). We can solve this equation using equation 3 to get expressions for the device currents involving only the constant current source values:
So plugging equation 7's expressions into equation 5 and rearranging:

\[
\frac{mI_T + I_0}{I_T - I_0} \left(1 + \exp\left(-\frac{V_o}{V_{th}}\right)\right) = \exp\left(\frac{I_T R_G}{V_{th}}\right)
\]

and defining

\[K = \frac{mI_T + I_0}{I_T - I_0}\]

we get

\[
I_L = \frac{V_{th}}{R_G} \ln K + \frac{V_{th}}{R_G} \ln \left(1 + \exp\left(-\frac{V_o}{V_{th}}\right)\right)
\]

which is our precious relationship between the current this half-stage is delivering, and the voltage that appears at the output. We can see immediately that the quiescent current when the output voltage drop is zero is

\[
I_2 = \frac{V_{th}}{R_G} \ln K + \frac{V_{th}}{R_G} \ln (1 + 1) = \frac{V_{th}}{R_G} \ln (2K)
\]

and the limiting current when the output voltage is many times the thermal voltage is

\[
I_{\text{lim}} = \frac{V_{th}}{R_G} \ln K + \frac{V_{th}}{R_G} \ln (1 + 0) = \frac{V_{th}}{R_G} \ln K
\]

This shows that solely by choosing the currents in the Class i driver's input pair we can set both the quiescent and the minimum current levels (they differ by a constant that's only dependent on the emitter resistor value). Note that the current as defined here is always positive for all values of output drop.

The minimum current can be chosen to be quite low even with low value emitter resistors. If all current sources in the amplifier have the same tempco, \(K\) is independent of temperature. Current mirror tolerance can be made pretty independent of transistor parameters. So the minimum and quiescent current settings are really quite stable over large production volumes.

The thermal voltage is proportional to absolute temperature, and therefore so are the quiescent and limiting currents in the output devices. Don’t be tempted to make the emitter resistors out of normal metal wire with PTAT (proportional to absolute temperature) resistivity to compensate. It will lead to signal-dependent distortion, because the resistors will vary with load current.
Formal analysis of the Class i stage (cont.)

Going back to equation 8 and developing the matching expression for the other half of the stage, we can write, with + and – suffixes for sourcing and sinking halves:

\[
I_{L+} = \frac{V_{cc}}{R_E} \ln K_e + \frac{V_{cc}}{R_E} \ln \left(1 + \exp \left(-\frac{V_O}{V_{bh}}\right)\right)
\]

and

\[
I_{L-} = \frac{V_{cc}}{R_E} \ln K_e + \frac{V_{cc}}{R_E} \ln \left(1 + \exp \left(+\frac{V_O}{V_{bh}}\right)\right)
\]

Now the actual current into the load is just the difference between these two. Assuming that the K and emitter resistor values are the same for the two half-stages, we have

\[
I_L = \frac{V_{cc}}{R_E} \ln \left(1 + \exp \left(-\frac{V_O}{V_{bh}}\right)\right) - \frac{V_{cc}}{R_E} \ln \left(1 + \exp \left(+\frac{V_O}{V_{bh}}\right)\right)
\]

\[
\therefore I_L = \frac{V_{cc}}{R_E} \ln \left(1 + \exp \left(+\frac{V_O}{V_{bh}}\right)\right)
\]

\[
\therefore I_L = \frac{V_{cc}}{R_E} \ln \left(1 + \exp \left(-\frac{V_O}{V_{bh}}\right)\right)
\]

\[
\therefore I_L = \frac{V_{cc}}{R_E} \ln \left(1 + \exp \left(+\frac{V_O}{V_{bh}}\right)\right)
\]

\[
\therefore I_L = \frac{V_{cc}}{R_E} \ln \left(1 + \exp \left(-\frac{V_O}{V_{bh}}\right)\right)
\]

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\therefore I_L = \frac{V_{cc}}{R_E} \ln \left(1 + \exp \left(+\frac{V_O}{V_{bh}}\right)\right)
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\[
\therefore I_L = \frac{V_{cc}}{R_E} \ln \left(1 + \exp \left(-\frac{V_O}{V_{bh}}\right)\right)
\]

\[
\therefore I_L = \frac{V_{cc}}{R_E} \ln \left(1 + \exp \left(+\frac{V_O}{V_{bh}}\right)\right)
\]

\[
\therefore I_L = \frac{V_{cc}}{R_E} \ln \left(1 + \exp \left(-\frac{V_O}{V_{bh}}\right)\right)
\]

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\therefore I_L = \frac{V_{cc}}{R_E} \ln \left(1 + \exp \left(+\frac{V_O}{V_{bh}}\right)\right)
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\therefore I_L = \frac{V_{cc}}{R_E} \ln \left(1 + \exp \left(+\frac{V_O}{V_{bh}}\right)\right)
\]

\[
\therefore I_L = \frac{V_{cc}}{R_E} \ln \left(1 + \exp \left(-\frac{V_O}{V_{bh}}\right)\right)
\]

This rather simple result delivers the Class i punch: the output impedance of the full stage, assuming matching between the devices, has a theoretically perfect linear output impedance equal to the value of one emitter resistor. And if this is the case, the stage cannot cause any non-linearity, whatever the load current. Note, by the way, that the sign in equation 12 results from our choice of variables – it is a positive output resistance!

The alternate configuration using a complementary power device driven from the collectors of the input transistors analyzes out to exactly the same result; the characteristics of the output devices don’t enter into these expressions at all, and create only second-order effects. And interestingly, when the effect of device base current is taken into account, we see that both configurations also behave the same way.

The emitter follower-based circuit steals current from the feedback transistor, whereas the complementary form injects extra current into the input transistors. In either case, base current is only an issue when one half of the circuit is strongly conducting, and it causes a small additional
curvature to the output as the offset of the driver stage changes (the effective value of $K$ changes a little at high output currents). It has little effect in the region where both halves are conducting and doesn’t disrupt the good linearity in the crossover region.

In part 3, we'll look at a complete example amplifier in simulation, and explore the benefits that the class i driver can confer. We'll also look at some of the take-care points that you'll need to attend to in a practical embodiment that needs to deliver the excellent potential linearity performance.

References
[10] Owen Jones, private communication

About the author
As well as being The Filter Wizard, Kendall’s a skilled circuit designer, product engineer and all-round analog expert. He's a widely published communicator of theory and method in many electronics disciplines. His fascination with electronics and audio dates back to boyhood. He spent 21 years designing filters, precision instrumentation, signal processing equipment and music systems. Then 11 years selling and designing in analog semiconductors. He's created advanced products to chase signals across many domains, extract the information from them and do something useful with it.

These days, Kendall is with Cypress, doing system architecture, product definition and strategic application analysis for their precision analog and mixed-signal devices. And also, of course, supporting customers with all sorts of analog and digital filter designs. He's also educating and mentoring a new generation in the ways of analog and systems thinking. Striving to improve the world one dB at a time!