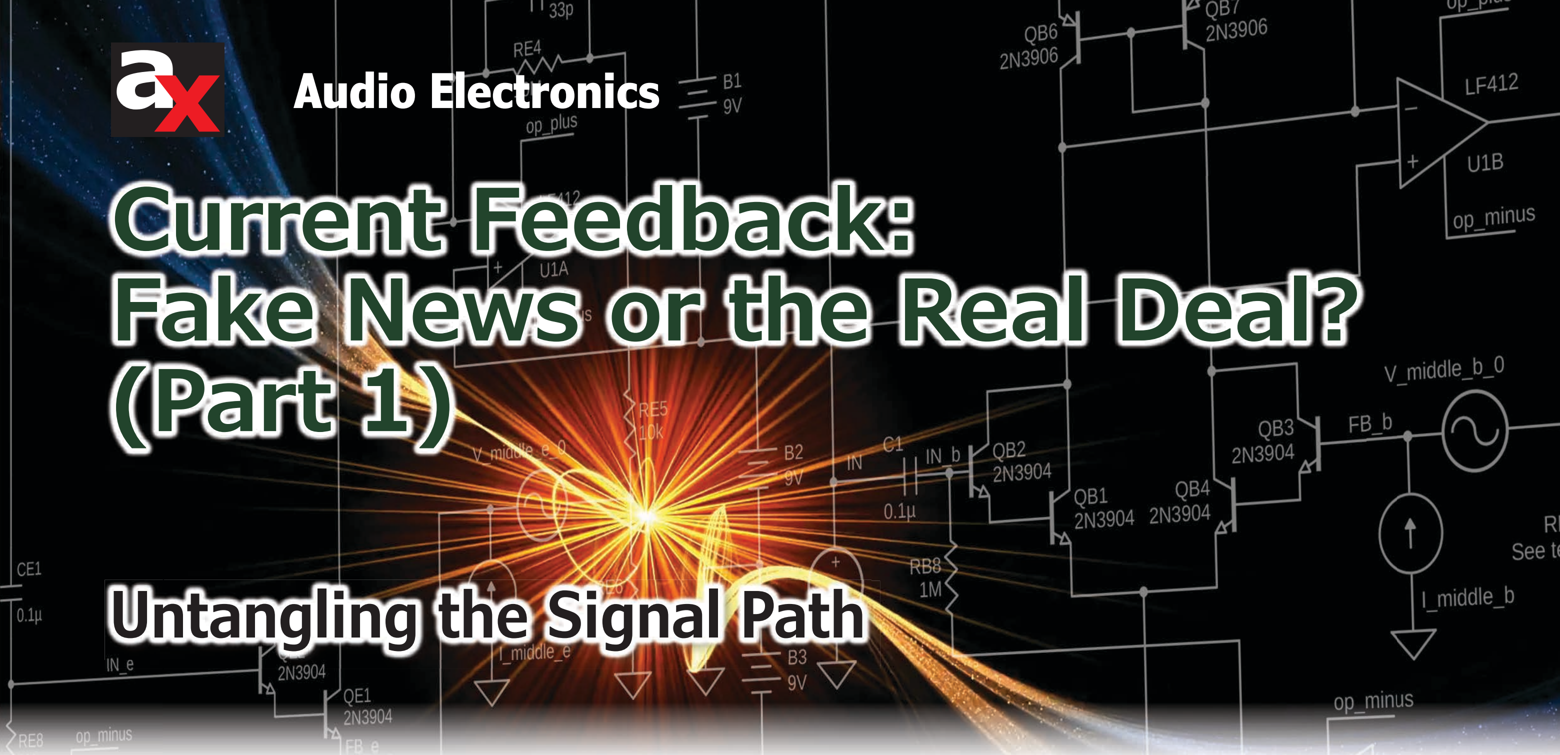

Current Feedback: Fake News or the Real Deal? (Part 1)

Untangling the Signal Path

Recently, the very existence of current feedback as opposed to voltage feedback has been called into question. This article employs bench tests, simulations, and equations derived from various circuits under investigation to address this controversy.

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Audio electronics is no stranger to controversies. Several examples come to mind.

First, it's generally agreed that if a circuit's output has a voltage, the impedance of that output can be determined. First, the voltage is measured. Then the output is connected to a load of known value and the voltage is measured again. A simple calculation reveals the impedance. But we can also set the output's signal to zero and drive that output with a current. The ratio of the resultant voltage to that current gives another impedance. Must these "output" and "input" impedances be identical? Thevenin's theorem provides the answer.

Second, the ground-referenced impedance at a Cathodyne phase splitter's cathode is widely accepted to be somewhere near $1/g_m$, the reciprocal of the triode transconductance. But is the impedance at the anode the same? B+ power supply noise rejection tests at that electrode answer the question.

Third, recently, the existence of current (as opposed to voltage) feedback has been challenged, specifically in ICs that the semiconductor industry has come to call "Current Feedback Amplifiers." In this article, I hope to shed some light on this wrangle.

A Diversionary Thought Experiment

Before we consider feedback, let's engage with something simpler. There is a functional equivalency between certain signal sources. A voltage source V in series with impedance Z is indistinguishable from a current source I in parallel with Z—that is, if we exclude from Z = V / I values of zero and infinity. Suppose that one of these sources drives a load L. Does the load see more of one source type than the other? Well, if L is less than Z, it's a better approximation to say that a current I flows through it than that a voltage V appears across it. The reverse is true if L is greater than Z. And if Z equals L, one approximation is as good (or as bad) as the other. We'll see that this relationship of source to load impedances and that of amplifier feedback networks to feedback inputs are similar.

Some Amplifier Input Structures

Figure 1 shows two different Darlington pair input stages. Let's refer to them by the junction to which feedback is applied: e-fed (emitter-fed) and b-fed (base-fed). The transconductance term g is approximated as shown in Figure 1. Perhaps surprisingly, each stage has the same value of g when all Darlingtons are identically biased. The e-fed Darlington sees the full differential voltage applied between FB and IN. Although each b-fed Darlington sees half of that, the PNP current mirror adds these b-fed half currents. And so each stage offers the same signal output current. Let's follow these currents to their sources. In the b-fed case, they come from "steering" the I_{dc} bias current. Negligible currents arrive from the FB (feedback) base—hence, voltage (no current) feedback. But in the e-fed case, no steering of I_{dc} is possible—the feedback portion of the signal current is sourced exclusively from the amplifier's output through the feedback network. This is one possible definition of current feedback.

Consider the impedances of the FB inputs. The e-fed circuit's is approximately 1/g Ω (see the Figure 1.) The b-fed's impedance is larger by a factor of β^2, where β (often greater than 100) is the ratio of a transistor's collector to base current. With all Darlingtons biased at 1 mA, $Z_{FB\_e} \approx 52\ \Omega$, while Z_{FB_b} is much larger. Such different input stages can lead to amplifiers with some noticeably different characteristics.

Amplifiers with E-fed and B-fed Input Stages

In Figure 2, the e-fed amplifier is on the left and the b-fed one is on the right. For functionally corresponding components, e-fed and b-fed reference designators differ only in whether they contain an "e" or a "b". Henceforth, when possible, we will generalize by suppressing those letters. Current sources are implemented with R_1 and Q_5. R_8 through R_{11}, C_2 and U_2 form servos that keep Out near 0 VDC. U_1 provides gain and low impedance outputs. +9 VDC at the U_1 (+) inputs establishes +9 V at the Q_1 and Q_2 collectors. Rb_7 works with Rb_4 to help shift Out close to 0 VDC. We shall see later that R_4 and C_3 have been selected to provide matching loop gains for the two amplifiers at 100% feedback (infinite R_6.) For the moment, we'll set C_3 and all sources with "middle" in their names to zero. This makes the voltage sources short circuits and the current sources open ones.

And now for something completely different (with apologies to Monty Python.) In analyzing these designs, I believe it would be wise to take guidance from an episode of *The Adventures of Young Indiana Jones*. In the 1920s, Indiana Jones (Indy) was quite taken with Jazz. He had an opportunity to play

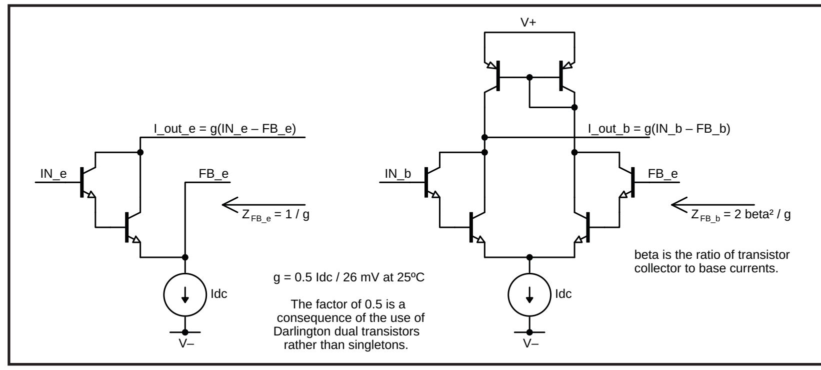

Figure 1: Input stages to which feedback is applied to an emitter (left, e-fed) and a base (right, b-fed).

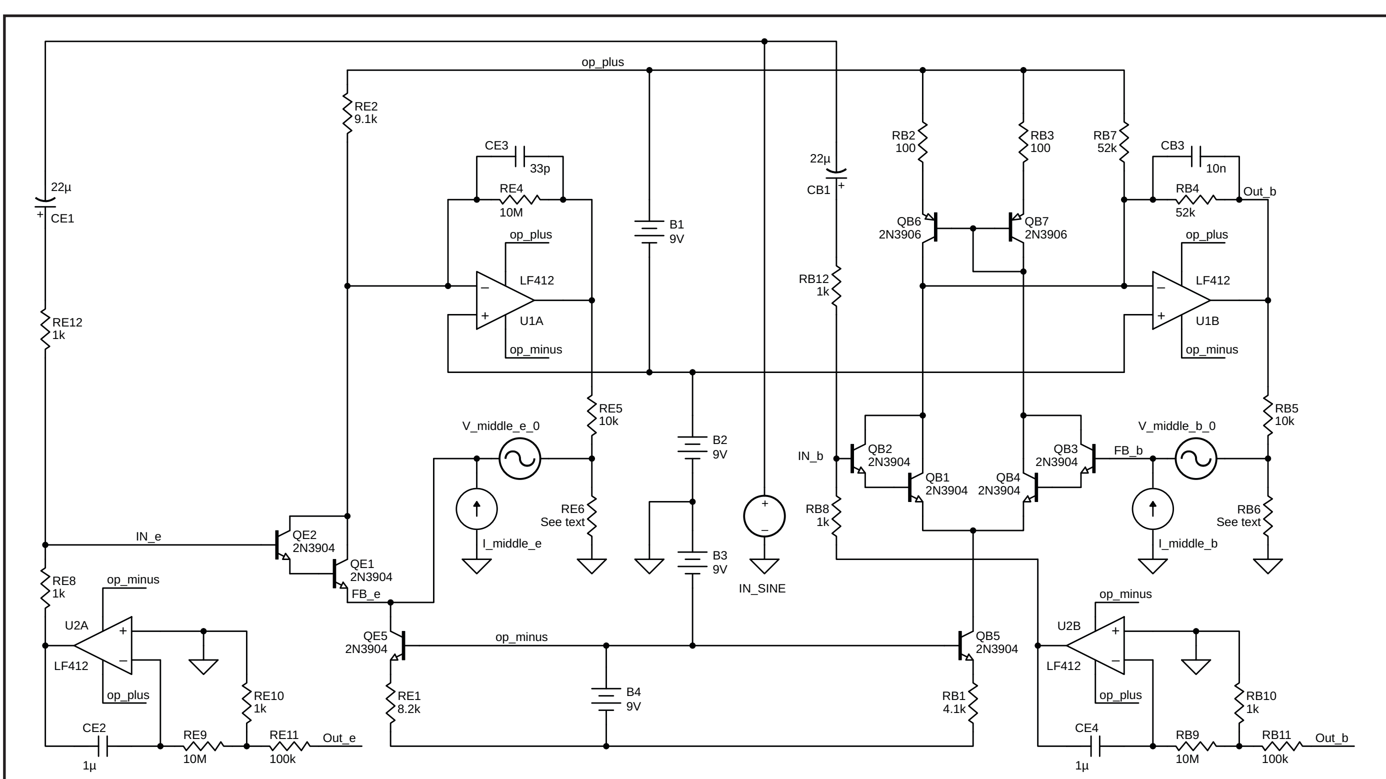

Figure 2: Amplifiers composed of e-fed (left) and b-fed (right) input stages

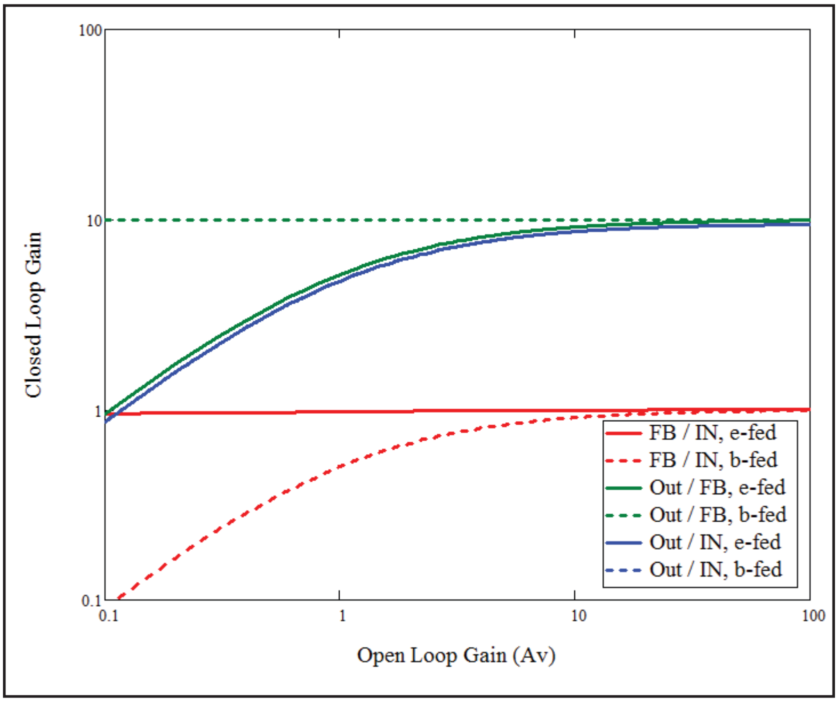

Figure 3: Graphs of equations 5 through 7 as a function of A_v for R_4 as shown in Figure 2 and R_6 = 1100 Ω.

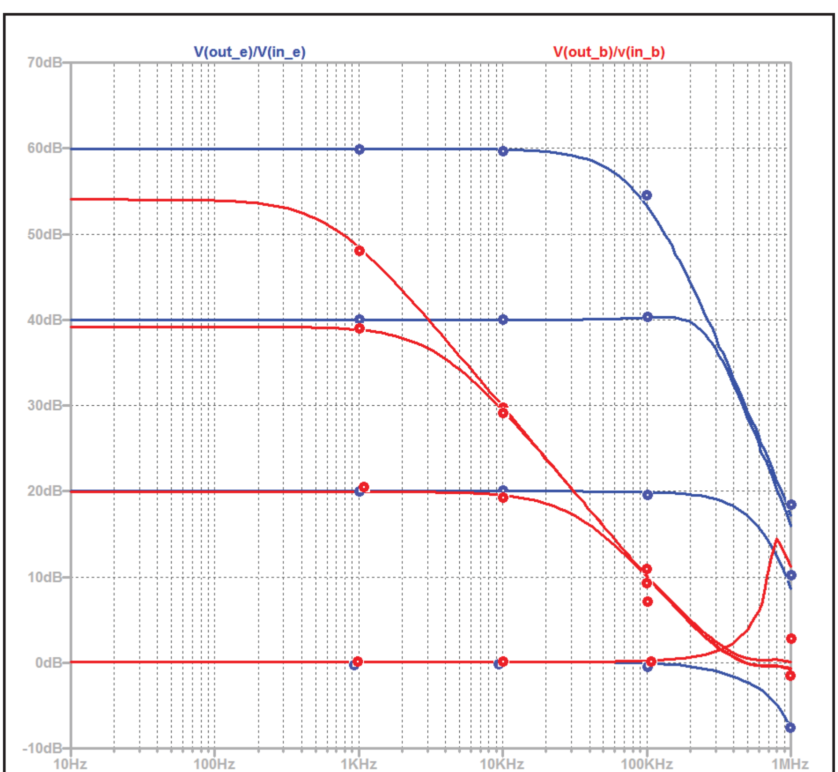

Figure 4: Measured and calculated closed loop gains of the b-fed and e-fed amplifiers shown in Figure 2 for G = 1 + R_5/R_6 = 0, 20, 40, and 60 dB.

saxophone with some of the greats of that era. But they remonstrated him for not knowing a song well enough to improvise on it. He was told, "You've got to know it forward and backward, inside out and upside down before you can understand it well enough to play it," or something to that effect. Let's follow that advice.

We can determine the open loop gains A_v of the Figure 2 circuits by inspection by setting the source IN_SINE to zero, disconnecting R_5 from Out, and driving R_5 from a voltage source V_test (not shown.) We obtain:

$$A_v = \frac{Out}{V\_test} = g \times R_4 \times \frac{R_b}{R_b + R_5}$$

where $R_b = R_6 \parallel Z_{FB}$ [1]

Solving this equation for R_4 while setting R_6 to infinity and A_v to 1,000 allows us to calculate the values for the R4s that you see in the schematic:

$$R_4 = \frac{A_v \times (Z_{FB} + R_5)}{g \times Z_{FB}} \qquad [2]$$

For the following equations, the first is obvious and the second comes from conservation of current at the FB node:

$$Out = g \times R_4 \times (IN - FB) \qquad [3]$$

$$\frac{FB}{R_6} + \frac{(FB - Out)}{R_5} + \frac{(FB - IN)}{Z_{FB}} = 0 \qquad [4]$$

From the above, we can derive three equations of interest:

$$\frac{FB}{IN} = \frac{1}{1 + \dfrac{1}{A_v + \dfrac{(A_v + 1)}{Z_{FB} \times \left(\dfrac{1}{R_5} + \dfrac{1}{R_6}\right)}}} \qquad [5]$$

$$\frac{Out}{FB} = \frac{1 + \dfrac{R_5}{R_6}}{1 + \dfrac{1}{A_v \times \left[1 + \dfrac{Z_{FB}}{R_5} + \dfrac{Z_{FB}}{R_6}\right]}} \qquad [6]$$

$$\frac{Out}{IN} = \frac{\left(\dfrac{R_5}{R_6} + 1\right)}{\left(1 + \dfrac{1}{A_v}\right)} \qquad [7]$$

Figure 3 shows these as dashed (b-fed) and solid (e-fed) lines as A_v varies, with R_6 arbitrarily set to 1100 Ω to make G = 1 + R_5 / R_6 ≈ 10. In these circuits, Z_{FB_e} measures 52 Ω and Z_{FB_b} >> R_6.

At high values of A_v, results are what we expect: FB is approximately equal to IN, and Out is G ≈ 10 times both. But look at what happens at low gains. The blue Out/ V_in traces are indistinguishable because A_v varies in lockstep for these two circuits. They also coincide with the solid green OUT/FB e-fed curve (I've purposely offset the blues slightly in the graph so that both they and the solid green can be seen.) The red dashed b-fed FB/IN is very much less than unity, though it remains approximately 1 throughout in the solid red e-fed case. The Out/FB dashed green b-fed curve holds steady at 10, but the solid green e-fed one falls to 1. The differences between these circuits, while obvious at low open loop gains, are simply masked at the high A_v to G ratios, which we are well advised to employ.

Combining equations [3] and [4] yields:

$$\frac{Out}{IN} = \frac{\left(\frac{R_5}{R_6}+1\right)}{\left[1+\frac{\frac{R_5}{Z_{FB}}+\frac{R_5}{R_6}+1}{g \times R_4}\right]} \quad [8]$$

The K = g × R_4 term is the voltage gain element in our design. Since we desire Out/IN to equal G, good practice requires K to be large enough that the term in the square brackets is very close to unity. However, when we account for the effects of C_3, (see Figure 3; C_3 is in parallel with R_4 and therefore affects K), we see that K falls with frequency. Varying G by varying R_5 will affect the coefficient of the 1/K term. This will change the frequency at which our "good practice" falters. However, the small value of Z_{FB} in the e-fed design affords us an opportunity. We simply hold R_5 constant and vary R_6. While R_6 remains greater than Z_{FB}, there is little variation in the coefficient of 1/K and therefore, little variation in bandwidth. Additionally, equation [1] shows us that this keeps the loop gain relatively constant. This strategy is unfortunately unavailable in the b-fed case. Here, Z_{FB} is typically quite large, R_5/Z_{FB} is a lot less than 1, and the 1/K coefficient varies with G. That explains the constant gain-bandwidth product typical of b-fed input stage designs.

The simulation in **Figure 4** of the Figure 2 circuits accounts for the effects of C_3 (we hold $R_4 \times C_3$ equal for both amplifiers) and demonstrates the results as we vary R_6 so that G = 0, 20, 40, and 60 dB. The small circles on the graph represent values measured on the test bench, where the Q1-Q2 and Q3-Q4 combinations are replaced with single package Darlingtons, BC517s. The peak of the b-fed curve in Figure 4 around 1 MHz is due to inadequate compensation. Overall, the curves are a bit difficult to believe. Did we really design for the same A_v in both amplifiers? We can check this by zeroing IN_SINE, disconnecting the R_5s from the Outs and driving the R_5s from a separate signal source V_test (again, not shown). The simulation and bench test results appear in **Figure 5**.

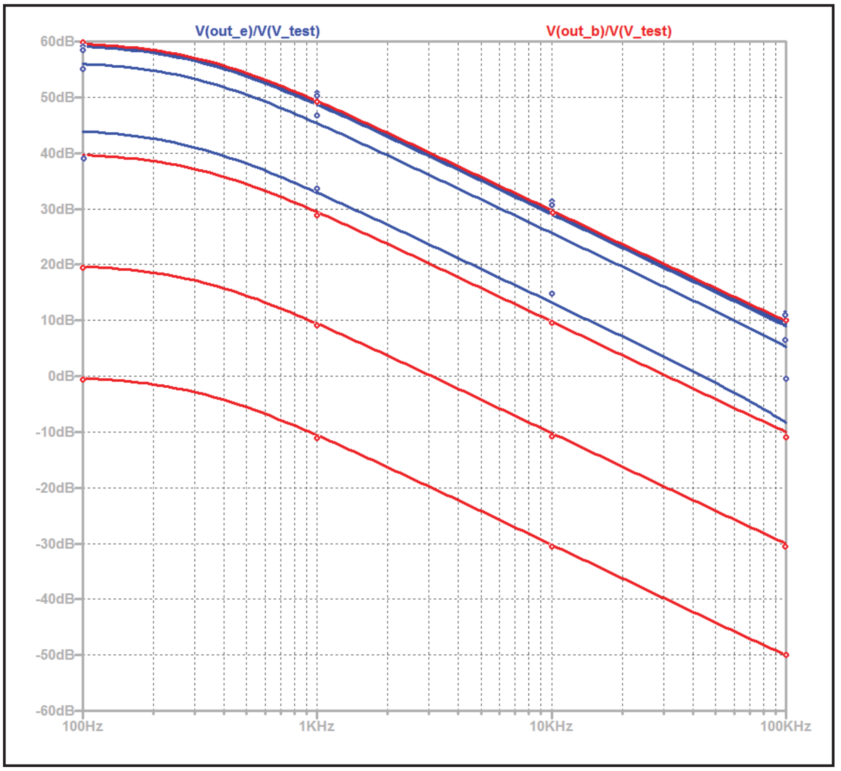

Figure 5: Measured and calculated open loop gains of the b-fed and e-fed amplifiers of Figure 2. IN_SINE is set to zero and R_5 is disconnected from Out and driven by a separate signal source, V_test (not shown).

The top blue (e-fed) and red (b-fed) curves in Figure 5 (which correspond to R_6 = infinity and G = 0 dB) confirm that the two amplifier open loop gains are practically identical under these circumstances, and are 1000 = 60 dB at low frequencies. The b-fed amplifier behavior is unremarkable. Its response is inversely proportional to G = R_5/R_6 + 1, as expected. But look at the e-fed performance! A_v does not fall as 1/G as G increases as we would expect, but rather much more slowly. And in Figure 4, it reaches a closed loop gain of exactly 60 dB, even though A_v itself is only 60 dB (when R_6 = infinity.) What is going on here?

The e-fed performance can be explained by the feedback network looking like a degenerative resistor to Q_1. Unlike in the b-fed case, it prevents the full voltage IN – Out/G from appearing across Q_1 and Q_2. But as we reduce G by reducing R_6, we also reduce the value of the degenerative resistance. More of that voltage appears across Q_1 and Q_2. This acts to increase the gain generated by this Darlington, partially counteracting the fall in loop gain. It is

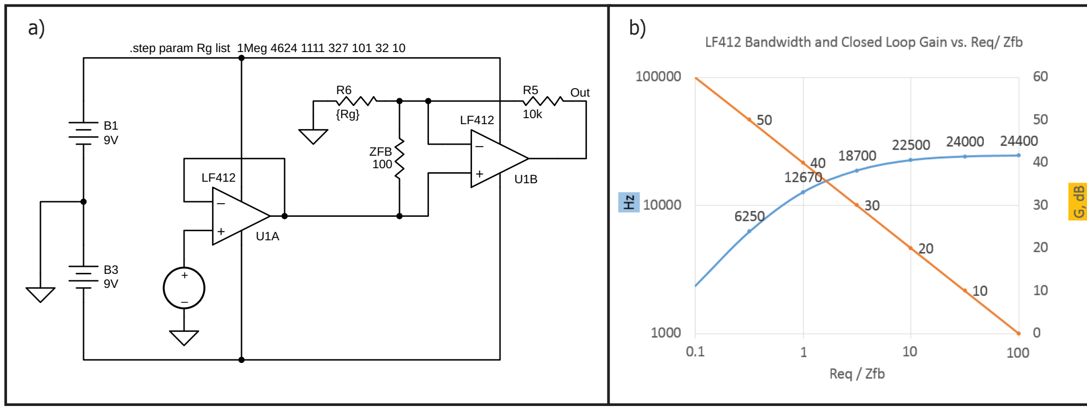

Figure 6: (a) Closed loop gains of an LF412 op-amp for $G = 1 + R_5/R_6 = 0$, 10, 20, 30, 40, 50, and 60 dB. (b) Z_{FB} drastically lowers the impedance at the inverting input of this b-fed-like op-amp.

almost anticlimactic to note that Figure 4 confirms that the e-fed design has a nearly constant bandwidth for G = 0 db (R_6 || R_5 = infinity || 10 kΩ) and 20 dB (R_6 || R_5 = 1100 || 10 kΩ.) Both resistances are greater than 1/g = 52 Ω. But bandwidth begins to fall when R_6 || R_5 is 100 Ω || 10 kΩ and more dramatically when 10 Ω || 10 kΩ, which are comparable to or less than 1/g. This evokes our "diversionary thought experiment" at the start of the article and brings us to...

Indy's Final Gambit

There's at least one more way to look at these designs. In 1975, R.A. Middlebrook described methods to precisely determine circuit loop gains. He established the following relationship between the total loop gain T of a given circuit, and its purely voltage T_V and current T_i loop gains:

$$(1+T)^{-1}=(1+T_i)^{-1}+(1+Tv)^{-1} \qquad [9]$$

There are a number of things that surprise about this relationship. First, although the value of T is unchanged by the chosen point of analysis within a given loop, this selection will affect T_v and T_i (although [9] will still hold.) For this reason, we apply our analysis at FB where we find the question of feedback type to be relevant. Second, because the applicable terms are in the denominators, the smaller of the voltage and current loop gains will have the stronger influence on the total loop gain T. Referring again to Figure 2, Middlebrook requires us to zero IN_SINE and to alternately activate V_middle and zero I_middle, and then to activate I_middle and zero V_middle. When V_middle is active, T_v is the negative of the ratio of the voltage at the junction of R_5 and R_6 (FB – V_middle) to that at the emitter of Q_1 (FB). To determine T_v, we first write the following current conservation equation at the FB node:

$$\frac{FB-0}{Z_{FB}}+\frac{FB-V\_middle}{R_6}+\frac{FB-V\_middle-Out}{R_5}=0 \qquad [10]$$

Combining equation [3] (with IN_SINE = 0) and [10], we obtain:

$$T_v=\left(g\times R_4+\frac{R_5}{Z_{FB}}\right)\times\left(\frac{R_6}{R_6+R_5}\right) \qquad [11]$$

With only I_middle active, T_i is the negative of the ratio of the current flowing from the junction of R_5 and R_6 (FB / Z_{FB} - I_middle) to that flowing into the emitter of Q_1 (FB / Z_{FB}). To determine T_i, we first write the current conservation equation at the FB node:

$$\frac{FB-0}{Z_{FB}}+\frac{FB}{R_6}+\frac{FB-Out}{R_5}-I\_middle=0 \qquad [12]$$

Combining equation [3] (again with IN_SINE = 0) with [12], we arrive at:

$$T_i=\left(1+\frac{R_5}{R_6}+g\times R_4\right)\times\frac{Z_{FB}}{R_5} \qquad [13]$$

The term responsible for amplifier gain is $g \times R_4$ = K. When equations [11] and [13] describe properly designed circuits, at low and moderate frequencies K dwarves any term with which it is summed. Under these conditions, we see that:

$$\frac{T_v}{T_i}\approx\frac{Req}{Z_{FB}},\ \text{where } R_{eq}=R_5\,||\,R_6 \qquad [14]$$

About the Author

Christopher Paul was born in the US the year before the Nobel prize was awarded for the invention of the transistor, the development of which he claims no credit for. His love of music from an early age led to some puzzling experiments with musical instrument amplifiers, and from there to an interest in electronics. Chris finds it interesting to delve into relatively simple circuits and to investigate and write about aspects of their performance that are little known or appreciated. He enjoys deriving equations that can be used to evaluate and design them, although he admits to being in therapy for this. Chris has worked for companies whose products are in the fields of communications, electronic article surveillance, and enterprise hand-held computers. He holds two M.Sc. degrees in electrical engineering from Brooklyn Polytechnic (now part of New York) University.

This tells us that when $R_{eq} > Z_{FB}$, the voltage loop gain exceeds the current loop gain. And when that inequality is reversed, the current loop gain is larger. However, the larger the loop gain, the less the feedback. And so, $R_{eq} > Z_{FB}$ implies that current feedback predominates at this point in the circuit and is the major influence on total loop gain. And, $R_{eq} < Z_{FB}$ implies that voltage feedback predominates at this point.

These results lead to an interesting question: Are the types of feedback and associated phenomena dependent on the kind of input stage, or only on the above inequality? Easy enough to find out...

Hold On to Your Hats

Figure 6 shows the results of a simple bench test. Start with an amplifier with a b-fed-like input—the LF412 (actually, its JFET inputs draw even less current than a b-fed.) Can we coax it to behave as if it had an e-fed input? The graph answers, "yes." Bandwidth is practically constant while $R_{eq} = R_5 || R_6 > Z_{FB}$, but falls with R_{eq} when that inequality fails.

Apparently, circuits have distinctly different characteristics depending on this inequality, regardless of input stage type.

Current Feedback Shakes Another Tree

Many of us are familiar with the four types of signal routings shown in **Figure 7**. The source of the feedback is the output, either the voltage across the load (shunt-derived) or the current through it (series-derived). But what is the feedback destination? It must be the op-amp feedback (inverting) input, right? There is no other place to "feed back" a signal to that will influence circuit operation! Typically, we are told that series applied is voltage feedback. So voltage is applied, but no current flows into the FB input, right? Wrong, if this is an e-fed device. The signal current flowing through such a transistor must come through the feedback network only. So how can this be exclusively voltage feedback? We are also told that shunt applied is current feedback. So this current flows into the op amp FB input, right? Not if we are working with a b-fed or FET input device (which accepts negligible current) it isn't. So how is this current feedback?

My takeaway is this: Describing signal routing as series/shunt/applied/derived is appropriate and a sufficient means of distinction. Insisting that each type is always associated with only current or only voltage feedback belies the differing natures of op-amps. It's bad enough that "current feedback" could be defined alternatively by input stage type or by Middlebrookian inequalities—consider that to some, it has also meant load current (series derived) feedback!

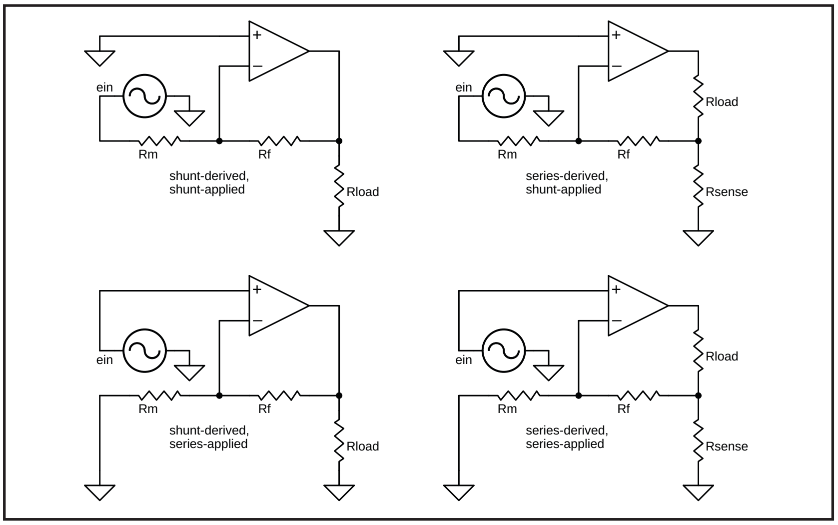

Figure 7: The four arrangements of signal routings

So Where Does This Leave Us?

Regardless of input stage type, while $R_{eq} > Z_{FB}$, we can maintain a constant bandwidth while varying the circuit's closed loop gain by adjusting only R_6. This is easier if the amplifier has an e-fed rather than a b-fed input stage, but it is not strictly necessary. With e-fed stages, as G is increased by reducing R6, loop gain falls more slowly than the 1/G of the b-fed case. This greater than expected loop gain likely has beneficial effects when it comes to limiting the increase in distortion at high closed loop gains, at least from the output stage.

The matching of A_v at G = 0 dB for the two input stage type amplifiers was done for purposes of comparison. Nothing presented here is meant to suggest that one type of input stage inherently leads to wider bandwidth or better performance in an amplifier.

In the second part of this article, we'll look at extensions of these input stage types as employed in modern op-amps and see how they lead to even further distinctions in performance. Op-amps built around some of these are what the industry has taken to calling Current and Voltage Feedback Amplifiers. We'll also delve further into feedback types.