Complete cancellation of nonlinearity can in theory be accomplished by pre-distortion [10] or feed-forward [11] schemes. In a conventional feedback system a complete cancellation is not possible, due to finite loop gain. A feedback topology that uses infinite loop gain to completely cancel nonlinearity is described here.

Like the other schemes, it is dependent on matching properties. Another drawback is that the infinite loop gain can exist at just one frequency. At other frequencies the loop gain is finite but boosted, making most implementations conditionally stable. The scheme is not as complicated as predistortion, and an advantage over the feed-forward approach is that the problematic summing at the output is avoided.

4.4.1 The Topology

The basic topology of an inverting voltage amplifier with feedback boosting is shown in figure 4.5.

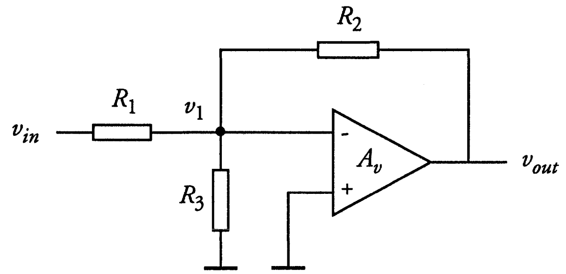

Figure 4.5: The basic inverting topology

Nodal analysis gives:

$$\frac{v_{in}-v_1}{R_1}+\frac{v_{out}-v_1}{R_2}-\frac{v_1}{R_3}=0 \Rightarrow \frac{v_{out}}{R_2} = -\frac{v_{in}}{R_1}+v_1\left(\frac{1}{R_1}+\frac{1}{R_2}+\frac{1}{R_3}\right)$$
$$= -\frac{v_{in}}{R_1}+\frac{v_{out}}{A_v}\left(\frac{1}{R_1}+\frac{1}{R_2}+\frac{1}{R_3}\right) \tag{4.6}$$

If we select R_3 such that the last term in (4.6) cancels, then the transfer from v_{in} to v_{out} becomes linear and independent of the gain A_v. The gain of the system is then completely determined by the resistors R_1 and R_2. The gain is identical to that with an ideal amplifier having infinite gain.

$$R_3 = -(R_1 \| R_2) \tag{4.7}$$

The last term in (4.6) cancels if R_3 is selected according to (4.7). The negative value does not necessarily result in instability, since the Miller effect of R_2 and the amplifier makes the resistance of the node v_1 positive.

The negative resistance can be implemented as shown in figure 4.6, where it consists of the lower amplifier and R_3, R_4 and R_5. Capacitor C_c is a phase-compensation capacitor that might be necessary depending on the bandwidths of the amplifiers.

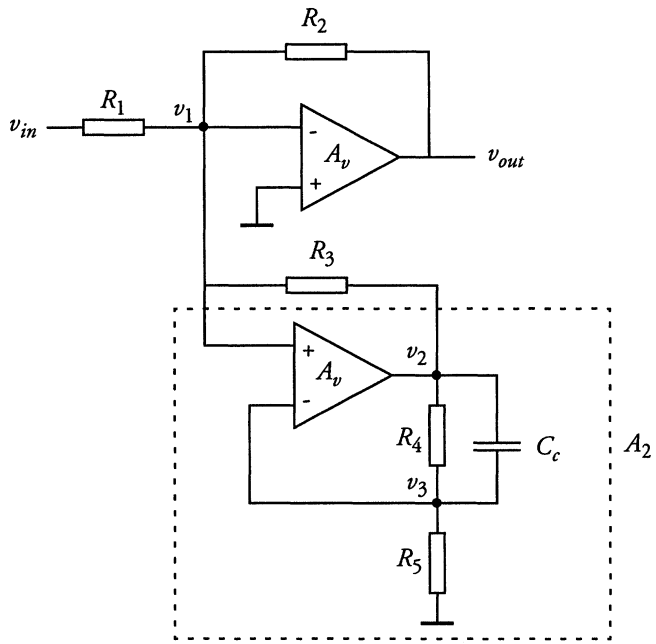

Figure 4.6: The complete inverting topology

4.4.2 Analysis

Assume the lower amplifier with feedback resistors R_4 and R_5 in figure 4.6 to have a frequency response with a single dominant pole at p_1 (4.8).

$$A_2 = \frac{A_{dc2}}{1 + s/p_1} \tag{4.8}$$

The negative resistance, R, is then

$$\begin{aligned} R &= -\frac{R_3}{A_2 - 1} = -\frac{R_3}{\dfrac{A_{dc2}}{1 + s/p_1} - 1} = \frac{-R_3(1 + s/p_1)}{A_{dc2} - 1 - s/p_1} \\ &= R_{dc}\frac{1 + s/p_1}{1 - s/(p_1(A_{dc2} - 1))} \quad , \qquad R_{dc} = -\frac{R_3}{A_{dc} - 1} \end{aligned} \tag{4.9}$$

At high frequencies the resistance becomes positive and equal to R_3. This phase-shift is accomplished by the left halfplane zero and the right halfplane pole. The right halfplane pole and negative resistance do not necessarily cause instability, as can be seen by calculating the feedback β of the first amplifier. If R_{dc} is selected according to (4.7), β becomes

$$\beta = \frac{R_1 \parallel R}{R_2 + R_1 \parallel R} = \dots = \frac{R_1}{R_1 + R_2} \cdot \frac{1 + s/p_1}{s/p_1} \cdot \frac{A_{dc2} - 1}{A_{dc2}} \tag{4.10}$$

The first factor in (4.10) represents the normal feedback, without negative resistance. The rest represents the boost in loop gain. At DC the single s-term in the denominator makes the loop gain infinite. Above p_1 the loop gain is almost unaffected. This pole has to be located well below the frequency where the loop gain reaches one, in order to preserve the phase margin. The whole arrangement results in a conditionally stable amplifier, if the loop gain already rolled off with at least single pole rate in parts of the boosted frequency range.

If there is a matching error, this will show up as a constant term next to the s in the denominator, reducing the loop gain at low frequencies, see figure 4.7.

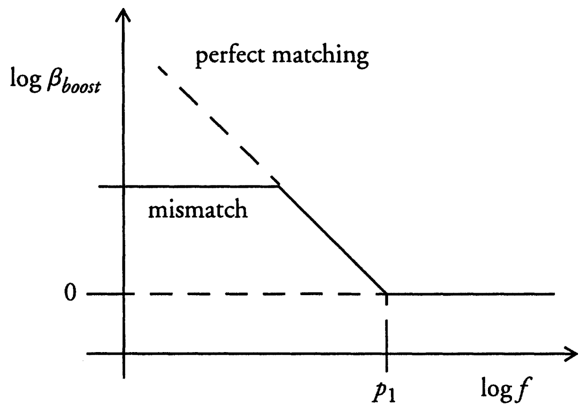

Figure 4.7: The boost in the feedback (extra loop gain)

Even in the case of perfect matching, there will not be zero nonlinearity. The nonlinearity of the second amplifier will cause some errors, but since this amplifier does not have to drive a large load, and since it operates on very small signals (the error signals of the primary amplifier), the nonlinearity will be extremely small.

If we apply some figures we can see how small the signal level at the second amplifier is. Assume the DC-gain to be as low as 1000, and the second amplifier to have the closed-loop gain A_{dc2} equal to 10. The signal at the output of the second amplifier is then 100 times smaller than at the output of the first amplifier. The errors due to mismatch are therefore likely to dominate.

4.4.3 Detailed Stability Analysis

To show the stability of the configuration the characteristic polynomial with single pole amplifiers can be examined.

$$\begin{cases} A_{v1} = \dfrac{A_{o1}}{1 + s/p_1} \\ A_{v2} = \dfrac{A_{o2}}{1 + s/p_2} \end{cases} \tag{4.11}$$

Nodal analysis gives the coefficients of the characteristic polynomial:

$$\begin{array}{ll} \text{constant term:} & (1 + \beta_2 A_{o2})\left(\dfrac{1}{R_2} - \dfrac{1}{A_{o1}}\left(\dfrac{1}{R_1} + \dfrac{1}{R_2}\right)\right) + \dfrac{A_{o2} - 1 - \beta_2 A_{o2}}{R_3 A_{o1}} \\ s\text{-term:} & \dfrac{1}{p_2}\left(-\dfrac{1}{R_2} - \dfrac{1}{A_{o1}}\left(\dfrac{1}{R_1} + \dfrac{1}{R_2} + \dfrac{1}{R_3}\right)\right) + \\ & \dfrac{1}{p_1} \cdot \dfrac{A_{o1} - 1 - \beta_2 A_{o2}}{A_{o1} R_3} - \dfrac{1}{p_1} \cdot \dfrac{1}{A_{o1}}\left(\dfrac{1}{R_1} + \dfrac{1}{R_2}\right) \\ s^2\text{-term:} & \text{negative} \end{array} \tag{4.12}$$

β_2 is the feedback of the second amplifier. The configuration is stable if all the coefficients have the same sign. Assuming large A_{o1} and A_{o2}, the stability conditions (negative coefficients) can be simplified:

$$\begin{array}{ll} \beta_2 R_3 > \dfrac{R_2}{A_{o1}} & \text{(DC Miller condition)} \\ R_3 > R_2 \dfrac{GB_2}{GB_1} & \end{array} \tag{4.13}$$

The first condition is easy to fulfil. The Gain-Bandwidth product of the second amplifier has to be chosen so that the second condition is fulfilled with some margin.

4.4.4 Experiment and Simulations

An experiment using two LF351 operational amplifiers was performed. They were used since they have a JFET-input stage with a very high input resistance. One of the amplifiers was heavy loaded and the other was used as a negative resistance.

A distortion meter was used to measure the nonlinearity. The amplifier was phase-compensated not to overshoot on a square wave. The result of the measurement can be seen in figure 4.8. As expected, the feedback-boosting enhances the linearity at low frequencies.

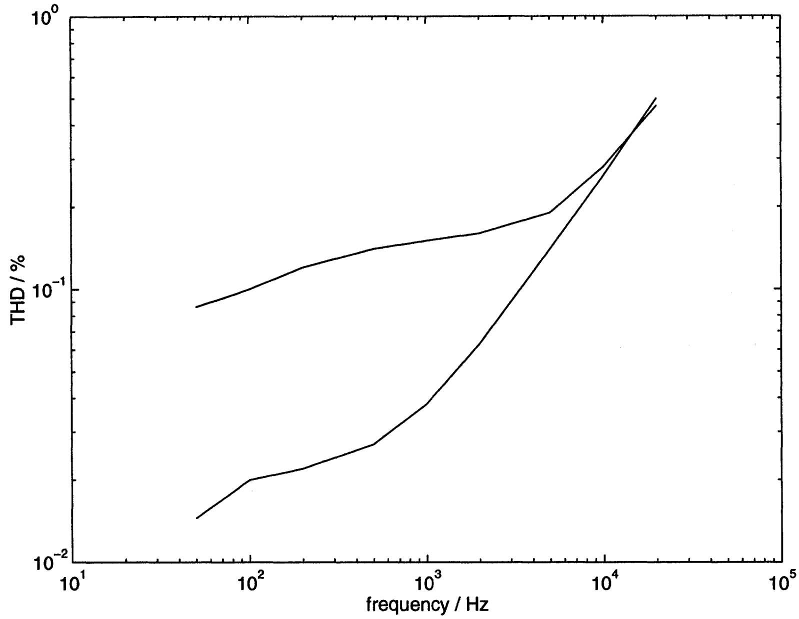

Figure 4.8: Measured THD with and without feedback boosting

The distortion and noise of the instrument made it hard to measure the low distortion at low frequencies. Some simulations were therefore made to verify the low-frequency linearity.

Two integrated two-stage amplifiers in 0.8μm CMOS were used in the simulations. The negative resistance amplifier was made a bit smaller than the other one. The results of the simulations can be seen in figure 4.9. At low frequencies the THD performance was improved several thousand times. Note the correspondence between the distortion reduction in this figure and the feedback boost in figure 4.7 with perfect matching.

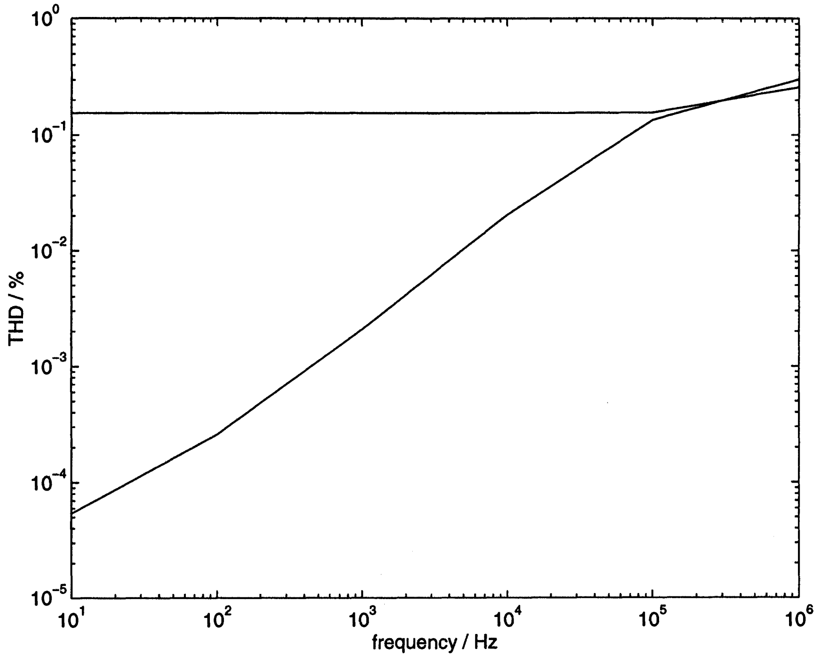

Figure 4.9: Simulated THD with and without feedback boosting (0.8μm CMOS two-stage amplifiers)

4.4.5 Other Topologies

The feedback boosting technique is not limited to inverting voltage amplifiers. A transresistance amplifier can be realized by omitting the resistor R_1 in figure 4.5. Formula (4.7) is then replaced by

$$R_3 = -R_2 \tag{4.14}$$

It is also possible to realize a non-inverting amplifier with feedback boosting, figure 4.10a. The value of the negative resistance should be selected according to (4.7). The negative resistance has to be floating, which can be realized as in figure 4.10b. The value of the negative resistance is

$$R = -R_4\frac{R_5}{R_6} \tag{4.15}$$

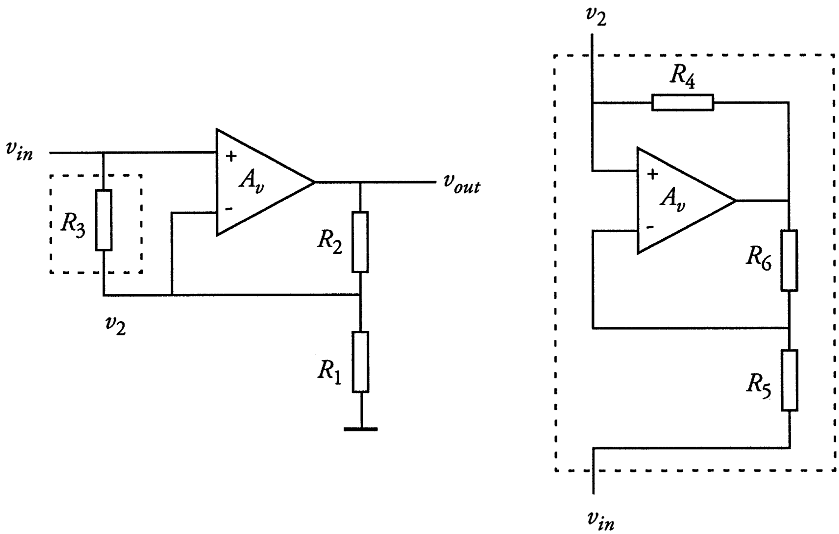

Figure 4.10: (a) A non-inverting topology (b) A floating negative resistance

4.4.6 Distortion Cancellation at Signal Frequencies

The compensation amplifier must have zero phase shift at the frequency where the loop gain is to be infinite. The compensation current can then compensate completely for the error, as the currents are in phase (anti-phase). A compensation amplifier with a low-frequency zero was investigated. The phase-shift was then zero at a non-zero frequency.

Some simulations were made using the same two-stage amplifiers as before. The second amplifier was tuned to have zero phase-shift at 10kHz, using a capacitor in its the feedback network. When sending a 5kHz tone through the circuit, the resulting 10kHz (second order) distortion was very small, which shows that the circuit works as intended.