Wideband fully differential amplifier noise improved using active match

Michael Steffes - June 09, 2013

Since its first appearance in 1999, the single to differential application of wideband fully differential amplifiers (FDAs) have used a resistor to ground as part of the input match at the cost of higher input referred noise voltage. If that resistor could be removed, with an input impedance match set solely by the path into the summing junction, a much lower input referred noise should be possible. This would be a viable option when the input match can be maintained to very high frequencies via a >1GHz common mode loop bandwidth. The design equations for both approaches will be shown here with a comparison of input referred noise parametric on target gain developed.

Single to differential conversions using a fully differential amplifier

One of the more useful functions supported by the growing range of FDA amplifiers is to convert from a single ended source to the differential output required by all modern ADC inputs. These can be either DC or AC coupled designs. When DC coupled, a careful attention to input common mode range is required and bipolar supplies can be useful in that case for many FDAs. For higher speed requirements, single supply is more common and an input match to some source impedance is often desired to limit reflections and/or SFDR degradation. While single supply FDAs can provide a DC coupled path, an AC coupled approach will be shown here to remove input common mode range considerations from this development. These same results will apply to a DC coupled design as long as the inputs stay in range. A typical AC coupled implementation of a doubly terminated 50Ω input design is shown in figure 1. This is targeting an example design set to a gain of 5V/V starting with a 499Ω feedback element and using a free Spice simulator to generate the schematic (reference 1).
Figure 1: Gain of 5V/V (14dB), 50Ω input impedance, AC coupled, Single to Differential Design

Several considerations are common to this type of circuit -

1. The feedback resistors are equal
2. The input impedance is set as a combination of Rᵣ and the impedance looking towards R₉₁.
3. The impedance looking towards R₉₁ will be increased over just the R₉₁ value by the action of the common mode loop within the FDA (reference 2). That loop acts to hold the output common mode voltage fixed which will then cause the input common mode voltages to move with the input signal increasing the apparent input impedance looking into R₉₁.
4. The R₉₁ resistor is set to get differential balance as R₉₁ + Rₛ||R₉₂.
5. With R₉₁ set, the noise gain (NG) for this circuit is 1+Rₛ/R₉₂.
6. With the AC coupling on the input paths, the DC I/O operating voltages default to the internally developed Vᵦᵦ reference (1.2V for this 3.3V single supply device). That Vᵦᵦ controls the output common mode voltage but, since there is no DC current path back to the input, it will also set the DC input common mode operating voltage.

The particular example above is using a very low noise, 4GHz gain bandwidth, FDA – the ISL55210 where, in this case, the design started by selecting an Rᵣ value then proceeded to solve for the Rₛ and R₉₁ elements. There has been very little vendor guidance in splitting that input match contribution between the Rₛ and R₉₁ elements. The tradeoff is that driving the R₉₁ element down (Rᵣ up) will reduce the input noise and extend the bandwidth (for voltage feedback based FDAs). Going in that direction then depends more on the common mode loop bandwidth to set the input match into the R₉₁ path (reference 2). While most reported approaches to deriving the resistor values for the circuit of figure 1 have either been iterative or approximate, picking an Rᵣ for a target gain (Aᵥ) and input impedance (Rₛ) can be manipulated into a quadratic solution for Rᵣ (reference 3).

\[
\frac{R₁^2 - R₁}{2Rᵣ(2 + Aᵥ) - RᵣAᵥ(4 + Aᵥ)} - \frac{2RᵣR₂^2Aᵥ}{2Rᵣ(2 + Aᵥ) - RᵣAᵥ(4 + Aᵥ)} = 0 \quad \text{Equation 1}
\]

Solving the coefficient denominator for zero will give a minimum Rᵣ to send Rᵣ to infinity depending then only on the R₉₁ input path for the match. For the example here, this solves to 160.71Ω.

\[
Rᵣ_{\text{min}} = \frac{RₛAᵥ(4 + Aᵥ)}{2(2 + Aᵥ)} \quad \text{Equation 2}
\]

As the Rᵣ is reduced towards this Rᵣ_{min}, the Rₛ elements will decrease while Rᵣ will →∞. Using equation 1 to get Rᵣ as a decreasing Rᵣ is selected, then the 2 other resistors are set by these expressions -
Noise analysis for the single to differential FDA

Once a set of resistor values are determined using these design equations, those can be placed into a noise analysis circuit to arrive at a total output differential spot noise. All of the elements contribute to the noise as shown in the circuit of figure 2 where the noise terms are shown as spot noise voltages and currents.

\[
R_{g1} = \frac{R_f - R_z}{2 \frac{A_v}{1 + \frac{R_z}{R_f}}}
\]
\[\text{Equation 3}\]

\[
R_{g2} = \frac{R_f}{2 \frac{A_v}{1 + \frac{R_z}{R_f}}}
\]
\[\text{Equation 4}\]

Noise analysis for the single to differential FDA

For the situation here where the \(R_f\) and \(R_g\) elements are equal, and the current noise terms are equal, the total output noise expression is very simple as shown in equation 5 where \(NG\) is the Noise
Gain equal to $1 + R_f/R_g$ (page 14 of the ISL55210 data sheet)

$$E_0 = \sqrt{(e_{ri} NG)^2 + 2(i_{ri} R_f)^2 + 2(4kTR_f NG)}$$  

Equation 5

Any wideband, Voltage Feedback based FDA can use this design flow to move the implementation resistor values down to the minimum set allowed by equation 2. Table 1 shows some of the lowest noise wideband Gain Bandwidth Product (GBP) FDA’s to apply to this analysis.

<table>
<thead>
<tr>
<th>FDA parts</th>
<th>$E_{ni}$ (nV/√Hz)</th>
<th>$I_{in}$ (pA/√Hz)</th>
<th>GBP (MHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>ISL55210</td>
<td>0.85</td>
<td>5.00</td>
<td>4000</td>
</tr>
<tr>
<td>LTC6406</td>
<td>1.60</td>
<td>2.50</td>
<td>3000</td>
</tr>
<tr>
<td>ADA4930</td>
<td>1.20</td>
<td>3.00</td>
<td>2800</td>
</tr>
<tr>
<td>LTC6409</td>
<td>1.10</td>
<td>8.80</td>
<td>10000</td>
</tr>
</tbody>
</table>

Table 1: Some modern FDAs and key parameters

Stepping the $R_f$ down for the example design of figure 1, re-computing the other resistor values, and then the input referred noise gives the results of table 2. The resistor values (exact here) would be the same for any of these 4 example devices to give a gain of 5V/V from the input of $R_t$ with a 50Ω input match (reference 4). Input referring the output noise from equation 5 by a gain of 5 gives the estimated input spot noise for each device in Table 2 (where these are still including the assumed 50Ω source noise collapsed into the $R_g$ elements of figure 2).

<table>
<thead>
<tr>
<th>$R_f$</th>
<th>$R_t$</th>
<th>$R_{g1}$</th>
<th>$R_{g2}$</th>
<th>NG</th>
<th>ISL55210</th>
<th>LTC6406</th>
<th>ADA4930</th>
<th>LTC6409</th>
</tr>
</thead>
<tbody>
<tr>
<td>500.00</td>
<td>71.18</td>
<td>88.11</td>
<td>117.48</td>
<td>5.26</td>
<td>2.16E-09</td>
<td>2.51E-09</td>
<td>2.27E-09</td>
<td>2.50E-09</td>
</tr>
<tr>
<td>462.31</td>
<td>73.79</td>
<td>80.43</td>
<td>110.23</td>
<td>5.19</td>
<td>2.07E-09</td>
<td>2.44E-09</td>
<td>2.19E-09</td>
<td>2.39E-09</td>
</tr>
<tr>
<td>424.63</td>
<td>77.14</td>
<td>72.72</td>
<td>103.06</td>
<td>5.12</td>
<td>1.98E-09</td>
<td>2.36E-09</td>
<td>2.10E-09</td>
<td>2.27E-09</td>
</tr>
<tr>
<td>386.94</td>
<td>81.60</td>
<td>64.97</td>
<td>95.97</td>
<td>5.03</td>
<td>1.88E-09</td>
<td>2.27E-09</td>
<td>2.01E-09</td>
<td>2.16E-09</td>
</tr>
<tr>
<td>349.25</td>
<td>87.81</td>
<td>57.16</td>
<td>89.02</td>
<td>4.92</td>
<td>1.77E-09</td>
<td>2.18E-09</td>
<td>1.92E-09</td>
<td>2.03E-09</td>
</tr>
<tr>
<td>311.55</td>
<td>97.09</td>
<td>49.26</td>
<td>82.26</td>
<td>4.79</td>
<td>1.66E-09</td>
<td>2.07E-09</td>
<td>1.82E-09</td>
<td>1.90E-09</td>
</tr>
<tr>
<td>273.83</td>
<td>112.46</td>
<td>41.22</td>
<td>75.03</td>
<td>4.61</td>
<td>1.54E-09</td>
<td>1.96E-09</td>
<td>1.70E-09</td>
<td>1.76E-09</td>
</tr>
<tr>
<td>236.19</td>
<td>142.95</td>
<td>32.95</td>
<td>68.99</td>
<td>4.37</td>
<td>1.41E-09</td>
<td>1.82E-09</td>
<td>1.57E-09</td>
<td>1.61E-09</td>
</tr>
<tr>
<td>198.50</td>
<td>233.38</td>
<td>24.21</td>
<td>55.39</td>
<td>4.04</td>
<td>1.25E-09</td>
<td>1.65E-09</td>
<td>1.41E-09</td>
<td>1.43E-09</td>
</tr>
<tr>
<td>160.81</td>
<td>67658.77</td>
<td>14.32</td>
<td>54.28</td>
<td>3.50</td>
<td>1.06E-09</td>
<td>1.41E-09</td>
<td>1.20E-09</td>
<td>1.21E-09</td>
</tr>
</tbody>
</table>

Table 2: Swept table of resistor values and resulting noise.

Decreasing the $R_f$ design value will be monotonically reducing the noise due to both decreasing resistor noise contributions and decreasing NG. The minimum value of 160.71Ω moves $R_f$ to infinity giving the lowest possible input noise and noise gain. The decreasing NG (=1+A_v/2 when $R_t$ is open) will also be extending the bandwidth for these voltage feedback devices. One tradeoff to moving these resistors down would be the ability of the common mode control loop bandwidth to hold the active input match over frequency looking into $R_{g1}$ acceptably close to $R_s$. In the limit of $R_t \to \infty$, that 14.3Ω $R_{g1}$ in the last row of table 2 will be transformed into a 50Ω active input impedance by the
common mode loop. Another concern might be the increased output stage loading due to lower $R_f$ values that will added to the actual differential load to possibly impair the harmonic distortion performance.

Plotting the input referred noise vs $R_f$ from table 2 gives figure 3. There is obviously a huge benefit towards reducing the noise by reducing the $R_f$ as far as consistent with desired input match frequency span and loading considerations. Starting from just picking $R_f = 500\Omega$ and proceeding down to the minimum 161Ω value for these design targets drops the total input spot noise from about 2.15nV/√Hz to 1.06nV/√Hz using the lowest noise ISL55210. Backing out the noise voltage delivered by the 50Ω source impedance to the matched input (that is still included in this 1.06nV/√Hz minimum) gives an input referred noise for just the amplifier stage of 0.96nV/√Hz.

![Input En vs Rf values](image)

**Figure 3: Input referred noise comparisons vs. target $R_f$ value**

**Eliminating $R_t$ and using only an active match design**

**Eliminating $R_s$ and using only an active match design**

Taking this analysis to its limit, eliminating the $R_s$ element to ground, uniquely solves for a set of required resistor values. Solving for the required $R_f$, given a target input impedance matched to $R_s$ and a gain from $R_{g1}$ to the differential output, gives the simplified design equations 6 to 8 where equation 6 is just the $R_{f\min}$ expression of equation 2 repeated.

$$R_f = \frac{A_v(A_v + 4)R_s}{2(A_v + 2)}$$

**Equation 6**
And then, \( R_{g2} = R_s + R_{g1} \) \hspace{1cm} \text{Equation 8}

Placing these expressions into the output noise calculation of Equation 5 using \( NG=1+A/2 \) in this simplified design, and working into a Noise Figure (NF) expression, gives equation 9 (reference 5).

\[
NF = 10 \log \left( 2 + \frac{4}{A_t} + \frac{\left( e_{si} \left( \frac{1}{2} + \frac{1}{A_t} \right) \right)^2}{kT R_s} + \frac{1}{2} \frac{i_n R_s A_t + 4}{A_t + 2} \right)
\]  

\hspace{1cm} \text{Equation 9}

Starting from the gain of 14dB (5V/V used earlier) and stepping the gain up in 2dB steps for a fixed input impedance of 50Ω, and using the 0.85nV/√Hz with 5pA/√Hz current noise for the ISL55210 from table 1, gives the required resistor values and resulting noise of Table 3.

<table>
<thead>
<tr>
<th>Gain (dB)</th>
<th>Gain (Av)</th>
<th>Rf</th>
<th>Rg1</th>
<th>Rg2</th>
<th>Noise gain</th>
<th>Eo</th>
<th>Noise Figure</th>
</tr>
</thead>
<tbody>
<tr>
<td>14</td>
<td>5.01</td>
<td>161.04</td>
<td>14.26</td>
<td>64.26</td>
<td>3.51</td>
<td>5.32E-09</td>
<td>7.51</td>
</tr>
<tr>
<td>15</td>
<td>6.31</td>
<td>195.71</td>
<td>12.03</td>
<td>62.03</td>
<td>4.15</td>
<td>6.36E-09</td>
<td>7.06</td>
</tr>
<tr>
<td>16</td>
<td>7.94</td>
<td>238.53</td>
<td>10.06</td>
<td>60.06</td>
<td>4.97</td>
<td>7.67E-09</td>
<td>6.88</td>
</tr>
<tr>
<td>18</td>
<td>10.00</td>
<td>291.67</td>
<td>8.33</td>
<td>58.33</td>
<td>6.00</td>
<td>5.30E-09</td>
<td>6.36</td>
</tr>
<tr>
<td>20</td>
<td>12.59</td>
<td>357.83</td>
<td>6.85</td>
<td>56.85</td>
<td>7.29</td>
<td>1.13E-08</td>
<td>6.08</td>
</tr>
<tr>
<td>22</td>
<td>15.85</td>
<td>440.62</td>
<td>5.60</td>
<td>55.60</td>
<td>8.92</td>
<td>1.39E-08</td>
<td>5.86</td>
</tr>
<tr>
<td>24</td>
<td>19.55</td>
<td>544.25</td>
<td>4.56</td>
<td>54.56</td>
<td>10.98</td>
<td>1.71E-08</td>
<td>5.67</td>
</tr>
<tr>
<td>25</td>
<td>25.12</td>
<td>674.23</td>
<td>3.69</td>
<td>53.69</td>
<td>13.56</td>
<td>2.12E-08</td>
<td>5.51</td>
</tr>
<tr>
<td>30</td>
<td>31.62</td>
<td>837.60</td>
<td>2.97</td>
<td>52.97</td>
<td>16.81</td>
<td>2.63E-08</td>
<td>5.39</td>
</tr>
<tr>
<td>32</td>
<td>39.81</td>
<td>1042.88</td>
<td>2.39</td>
<td>52.39</td>
<td>20.51</td>
<td>3.27E-08</td>
<td>5.28</td>
</tr>
<tr>
<td>34</td>
<td>50.12</td>
<td>1301.05</td>
<td>1.92</td>
<td>51.92</td>
<td>26.06</td>
<td>4.08E-08</td>
<td>5.20</td>
</tr>
</tbody>
</table>

Table 3: Swept Gain 50Ω active match element values and ISL55210 noise analysis.

This first row of values closely matches the earlier results in the last row of table 2. These resistor values would be correct for any voltage feedback FDA while the output noise and Noise Figure are predicted using the ISL55210 input spot noise numbers. As is normally the case, increasing the gain will be reducing the input referred noise at the cost of reduced bandwidth as shown by the increasing noise gain (V/V). Continuing the gain of 5V/V design from figure 1, but eliminating the \( R_t \) element and using the values in the first row of Table 3, gives the simulation circuit of figure 4.
Figure 4: Gain of 5V/V, 50Ω input, active match with wideband FDA.

With a noise gain =3.5V/V in this circuit, this should give >1GHz bandwidth for this 4GHz gain bandwidth device. While the simulations here are pretty accurate, this circuit, over a wide range of gains and input impedances, can also be easily tested using the ISL55210-ABEV1Z Active Balun Evaluation Board.

Figure 5: Frequency response shape for Vout/Vin from the simulation circuit of figure 4

Note the extremely fine scale on this simulation, showing <0.3dB rolloff from 1MHz to 1GHz where the low frequency rolloff is being set by the blocking capacitors. One final check is to look at the input impedance to see if in fact the common mode feedback loop is transforming that 14.3Ω $R_{g1}$ into something close to 50Ω. Changing the simulation circuit of figure 4 to a current source input with a shunt 50Ω and probing the input voltage on an AC simulation should give close to 25Ω if the circuit is working correctly. Manipulating that data into just the impedance looking into $R_{g1}$ gives figure 6. The simulated response closely matches the expected 50Ω with an increasing impedance at higher frequencies as the common mode loop bandwidth rolls off. The match here exceeds 34dB return loss.
through 1GHz - far higher in frequency than previously available FDAs. This closely matches the measured input impedance for this circuit (reference 6)

![Simulated Input Impedance at Rg1](image)

**Figure 6: Input impedance for current source input to figure 4**

**Summary and conclusions**

Wideband FDAs offer a useful circuit block for converting single to differential in high dynamic range signal processing designs. A closed form solution for the termination element to ground offers an easy means to assess tradeoffs in the split between that element and the series resistor into the summing junction. Increasing that $R_t$ element decreases the other resistor values (for a fixed target input match and gain) which will then extend the bandwidth and reduce the noise. In the limit, removing $R_t$ and depending only on the $R_{g1}$ element and the common mode loop to set the input impedance will give the lowest noise and widest bandwidth response for any voltage feedback FDA. This application works best using FDAs with very high bandwidth common mode loops. This approach can possibly be used to replace an RF amplifiers’ single ended I/O plus balun solution with just this active balun configuration of the ISL55210. It has the added benefit over balun designs of isolating the load from the source impedance. The simple design equations shown here give considerable design flexibility in the input impedance and gain by changing just 4 resistors values.

**About the author**
Michael Steffes With 27 years involvement in high speed amplifier design, applications, and marketing, Michael Steffes has introduced over 80 products spanning 5 companies while publishing >40 contributed articles. Current focus is on high efficiency high speed ADC interfaces, DSL/PLC line interface solutions, and online design tool development.

References:

1. Intersil’s free Spice and power simulator, iSim PE (registration required)
2. “Get a wideband matched input impedance with ultra-low noise using the active match capability of a new type of amplifier”, Michael Steffes
4. These resistor values match those delivered by the ADI diff amp calculator – recognize the gain there is from the source and includes the divide by 2 to the input of Rt.
5. Contact the author for the step by step derivation of this noise figure expression
6. “Designer's Guide to the ISL55210-ABEVAL1Z Active Balun Evaluation Board”, Michael Steffes, Intersil application note AN1831 (Click on Documents tab and look under Evaluation Boards)