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On the application of frequency selective common mode feedback for multifrequency EIT

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Abstract

Common mode voltages are frequently a problem in electrical impedance tomography (EIT) and other bioimpedance applications. To reduce their amplitude common mode feedback is employed. Formalised analyses of both current and voltage feedback is presented in this paper for current drives. Common mode effects due to imbalances caused by the current drives, the electrode connections to the body load and the introduction of the body impedance to ground are considered. Frequency selective narrowband common mode feedback previously proposed to provide feedback stability is examined. As a step towards multifrequency applications the use of narrowband feedback is experimentally demonstrated for two simultaneous current drives. Measured results using standard available components show a reduction of 62 dB for current feedback and 31 dB for voltage feedback. Frequencies ranged from 50kHz to 1 MHz.

Keywords: bioimpedance, common mode error, common mode feedback, current drive, imbalance, multifrequency

(Some figures may appear in colour only in the online journal)
1. Introduction

The established method of electrical impedance tomography (EIT) (Boone and Holder 1996, Bayford 2006) uses an array of sensing electrodes as shown in figure 1(a). The body is a non-uniform conducting medium. Current is injected into two electrodes and the voltage across pairs of electrodes is detected and amplified. Algorithms using the resulting voltage measurements deduce the disposition of different components in the body (Bayford 2006, Adler et al 2009). Measurements from a current drive with simultaneous multiple frequencies are also employed (McEwan et al 2007). Increasing the number of electrodes and employing drives with more than one sinewave frequency enable better definition of the body components. When there are a large number of electrodes the voltages across the distant sense electrodes due to the current drive can be very small. The amplitude of the differential voltages depends on the application, but voltages down to about 0.05 mV can be expected.

Current drives are generally floating, i.e. have high impedance to ground, to provide an optimal regime to enable sense amplifiers to detect very small differential signals. In figure 1(a) to aid analysis the distributed nature of the body has been simplified to an equivalent circuit containing four impedances (Rosell and Rui 1992). These represent the attenuation of the differential signals between the drive electrodes and the sense electrodes.

Any mismatch or imbalance particularly in the driving circuit or contact electrodes generates a common mode signal at the same frequency as the differential signal which has to be accommodated by the sense amplifier. Due to the high output impedance of the current drive (Bertermes-Filho et al 2000, Ross et al 2003, Hong et al 2009), common mode signals of around 0.5 V are possible and would require a common mode rejection ratio (CMRR) of between 140 dB and 107 dB for 0.1% accuracy of the measurement (Rosell and Rui 1992, Boone and Holder 1996). The high CMRR required of the sense amplifier can be reduced by applying common mode feedback before the sense amplifier input. Mismatches in the sense amplifier will also cause unwanted common mode signals but are not considered here. The common mode voltage experienced at the electrodes is assumed to be the same on all electrodes which for common mode signals allows the circuit to be further simplified to figure 1(b). In this figure the effect of the body impedance to ground, $Z_b$, has been added as a centre tap which assumes that $Z_b$ does not cause mismatch. There is always a $Z_b$ component either through body capacitance to earth and/or a reference electrode but provision of a reference electrode in an EIT system can distort the field distribution being monitored and may be undesirable. The common mode voltage $V_{cm}$ is derived by adding the voltages on two sense electrodes ($=2V_{cm}$).

This paper formalises the analysis, which previously has not been explicitly addressed, of common mode effects due to imbalances caused by current drives, mismatch of electrode connections and the effect of the body impedance to ground. It also examines the introduction of narrowband common mode feedback to address problems of stability encountered in conventional wideband EIT feedback systems. In section 2 the analysis and comparison of the application of common mode current or voltage feedback is discussed. Section 3 introduces narrowband common mode feedback and its transient response. Section 4 presents measured results of dual frequency current and voltage feedback. Section 5 discusses the results and conclusions are in section 6.

2. Common mode feedback

2.1. Feedback options

To reduce any large common mode voltages common mode feedback has been proposed by Rosell and Rui (1992). Both voltage and current feedback are possible. Figure 2 shows...
feedback options. Components $Z_e$ have been added to represent the electrode contact impedances and the ideal floating current drive in figure 1 has been replaced by the practical realization of two balanced current drives (figures 2(a) and (c)).

Common mode feedback can use either voltage feedback or current feedback (Hwang 2005, Hwang and Park 2005). Referring to figure 2, the options shown are:

(a) ‘Balanced’ current drive using current feedback to the electrode $Z_e$. In practice there is always a small mismatch $\Delta I$ of the two current drives.
(b) Single current drive using current feedback. The balancing current is supplied by the feedback transconductor $G$. This is a simpler version of option (a) and the feedback loop gain required is much higher.
(c) Centre tapped feedback (Casas et al 1996, Anton and Riu 2010). Voltage feedback $A$ is connected to the centre tap of the ‘balanced’ current drive.
(d) Single current drive using voltage feedback. The balance current is supplied by $A$. The feedback loop gain must be much higher compared with (c).

The filter defines the frequency characteristics of the feedback loop.

2.2. Feedback options

Using the example of figure 2(a) for balanced drive when there is no feedback and assuming $Z_o >> Z_e$ the common mode voltage $V_{cm}$ is

$$V_{cm} = \Delta I Z_o \frac{Z_L/2 + 2 Z_e \| Z_b}{Z_o + Z_e \| Z_b}$$

where $Z_o$ is the output impedance of each of the ‘balanced’ current drives and $\Delta I$ $I$ is their mismatch error. The effect of the value of $Z_o$ on the common mode voltage is illustrated by considering the extreme cases:

For $Z_b = \infty$ then $V_{cm} = \Delta I Z_o$; for $Z_b = 0$ then $V_{cm} = \Delta I Z_L/2$.

Since $Z_o >> Z_L$ when there is no feedback the common mode voltage can be significantly reduced by a low value of $Z_b$.

The common mode voltage can be reduced by applying current feedback as in figure 2(a).
For extreme values of \( Z_b \), equation (3) can be simplified to:

\[
V_{cm} = \Delta I \frac{Z_o}{1 + GZ_o}.
\]

For \( Z_e = 0 \), the equation simplifies to:

\[
V_{cm} = \Delta I.
\]

A similar analysis of the voltage feedback option in figure 2(d) (assuming \( Z_c = 0 \)) yields:

\[
V_{cm} = \frac{I}{2} \frac{(Z_l/2) + (Z_l/2)|Z_b|}{1 + A(1 + \frac{GZ_l/2}{Z_c})}.
\]

For extreme values of \( Z_b \), equation (6) can be simplified to:

\[
V_{cm} = \frac{IZ_l}{2} \frac{1}{1 + 2A}; \quad \text{for } Z_b = 0 \quad V_{cm} = \frac{IZ_l}{4} \frac{1}{1 + A}.
\]

If \( A \gg 1 \) both equations in (7) simplify to:

\[
V_{cm} = \frac{IZ_l}{4A}.
\]
In (4) and (7) when the loop gain is sufficiently large the common mode voltage has been reduced to a value independent of $Z_b$. Similar results occur in the other two options in figure 2.

2.3. Stability

The stability of the feedback is identified by the phase margin at the frequency when the feedback loop gain is unity. From (4) the current feedback loop gains $A_L$ are:

$$A_L = \frac{G_Z}{1 + \frac{\Delta Z L}{Z_o}}$$

For $Z_b = \infty$ $A_{Loo} = G_Z o$; for $Z_b = 0$ $A_{L0} = G_Z l$.

Since $G_Z o$ or $G_Z l$ will have more than one pole and $Z_o >> Z_L$ the low frequency gain of $A_{Loo}$ is high and there will be more instability compared with $A_{L0}$. The loop gain variation in (7) is far less than (4). If there are wide variations in $Z_L$ and $Z_o$, introduction of a dominant pole to ensure stability must be based on the worst case values expected. The dominant pole affects the settling time.

2.4. Comparison of feedback options

Table 1 summarizes the steady state formulae for the common mode voltage $V_{cm}$ using the different feedback options assuming $Z_b = \infty$, $Z_o >> Z_c$, $Z_l$, $1/G$, and where $Z_{c1} = Z_c(1 + \Delta Z_c)$ and $Z_{c2} = Z_c(1 - \Delta Z_c)$.

<table>
<thead>
<tr>
<th>Assumption</th>
<th>Current feedback balanced drive (figure 2(a))</th>
<th>Centre tapped voltage feedback balanced drive (figure 2(c))</th>
</tr>
</thead>
<tbody>
<tr>
<td>Without $Z_o$, $I(1 \pm \Delta I)$</td>
<td>$GZ_o &gt;&gt; 1$ $A &gt;&gt; 1$</td>
<td>$\Delta I Z_o$</td>
</tr>
<tr>
<td>Matched $I$, $Z_o(1 \pm \Delta Z_c)$</td>
<td>$\Delta Z Z_o$</td>
<td>$\Delta Z Z_o$</td>
</tr>
<tr>
<td>No feedback, $I(1 \pm \Delta I)$</td>
<td>$\Delta I Z_o$</td>
<td>$\Delta I Z_o$ Independent of $Z_L$</td>
</tr>
<tr>
<td>Independent of $Z_L$</td>
<td></td>
<td>Independent of $Z_L$</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Assumption</th>
<th>Current feedback single drive (figure 2(b))</th>
<th>Voltage feedback single drive (figure 2(d))</th>
</tr>
</thead>
<tbody>
<tr>
<td>$Z_o(1 \pm \Delta Z_c)$</td>
<td>$GZ_o &gt;&gt; 1$ $A &gt;&gt; 1$</td>
<td>$I(Z_o + 2Z_{c2})$</td>
</tr>
<tr>
<td>No feedback, $I(1 \pm \Delta I)$</td>
<td>$\frac{I Z_o}{2}$</td>
<td>$\frac{I(Z_o + 2Z_{c2})}{2}$</td>
</tr>
<tr>
<td>Independent of $Z_L$</td>
<td></td>
<td>Function of $Z_L$ and $Z_{c2}$</td>
</tr>
</tbody>
</table>

* In voltage feedback $Z_{c2}$ cannot be ignored.

In (4) and (7) when the loop gain is sufficiently large the common mode voltage has been reduced to a value independent of $Z_b$. Similar results occur in the other two options in figure 2.

In voltage feedback $Z_{c2}$ cannot be ignored.
3. Common mode feedback

3.1. Narrowband multiplier filter for selective frequency feedback

Potential instability can be resolved by using a narrowband multiplier filter as in figure 3 (Hwang 2005, Rahal et al 2009). Figure 4 shows simulations of the frequency and transient response characteristics of the circuit in figure 3 using ideal components from the ABM library in OrCAD®. The single pole response to a step is most clearly seen at $V_{c1}$ shown in figure 4(b). The locking frequency is $f_c$ and the low pass filters (RC = $T_o$) have poles $f_o = 1/T_o$. Amplifiers $A_{dc}$ can provide voltage gain at low frequencies. When $A_{dc} = 1\,\text{V V}^{-1}$ at the locking frequency the filter gain factor is 0.5. Outside the locking frequency the filter gain is attenuated. The low pass filters in figure 3 have three functions: (i) they define the narrowband bandwidth, (ii) they suppress the $2f_c$ component generated at the outputs of the first multipliers, and (iii) they provide a dominant pole to allow stability in feedback. In the steady state since one of the inputs of the second stage multipliers is $A_{dc}$, the outputs are independent of $f_o$. The $2f_c$ ripple at the output of the first multiplier is suppressed at $V_{c1}$ by a factor $2f_c/f_o$. In the example of figure 4, $f_c = 100\,\text{kHz}$ and $f_o = 1.6\,\text{kHz}$. The bandwidth in figure 4(a) is $2f_o = 3.2\,\text{kHz}$. The resulting ripple is 7 mV p–p. Under steady state conditions at the locking frequency $f_c$ the filter transfer function has a gain of $0.5A_{dc}$ and the common mode feedback analyses in section 2.2 apply. The step response is dominated by the poles of the low pass filters whose time constants are $T_o = 1/f_o$, defined by a tradeoff between bandwidth chosen, accuracy and feedback stability.

3.2. Step response with common mode feedback

In figure 2 by adding the multiplier filter into the feedback circuit (‘Filter’ in figure 2) a stabilising dominant pole can be introduced at the locking frequency $f_c$ with very little phase shift. When in a feedback circuit, the step response (at the locking frequency) operates in a similar fashion to the step response of a standard operational amplifier with a dominant pole in its feedback path (Rahal et al 2009). For (4) and (7) the Laplace form is

$$V_{cm}(s) = M \frac{1}{s} \frac{1}{1 + \frac{A_L}{1 + sT_o}}$$

where $M$ is a value depending on the feedback circuit, and $A_L$ is the loop gain of the feedback at the locking frequency.

The step response is
In (5) and (8) it has been shown that the final steady state value of $V_{cm}$ is hardly altered over a wide range of $Z_b$. The transient term in (11) can therefore be used to determine the time taken to reach to within a certain range of the final value.

Figure 5 shows the time taken to reach within 1% of the final value as a function of the loop gain $A_L$. For fast step response it is an advantage to have high feedback gain even though this results in the high initial values of $V_{cm}$ as shown in (4) and to a lesser extent (8). From (4) using common mode current feedback the step response is very susceptible to $Z_b$ variation compared with common mode voltage feedback (7) as shown in simulation in figure 6. Note that the linear analysis shown here breaks down when considering times below the period of the operating frequency of the narrowband filter. Feedback can also break down if there is insufficient attenuation in the low pass filter when very high loop gains are used.

4. Experimental results using two simultaneous frequencies

Only one frequency can be processed using a multiplier filter feedback. For two or more simultaneous frequencies each frequency must have its individual multiplier filter and feedback circuit. To demonstrate the practicality of the principle a single current drive with two simultaneous frequencies using either current (figure 2(b)) or voltage (figure 2(d)) feedback were demonstrated. Figure 7 shows the implementation of the basic common mode feedback circuit. It has two narrowband multiplier filters as shown in figure 3 (Hwang 2005). The circuit constructed uses discrete component devices. $Z_L = 1 \text{k} \Omega$ and $Z_e = 500 \Omega$ were used. The choice of resistive load components was to ensure any frequency effects observed were due to the common mode structure. There was no $Z_b$ (potentially the least stable condition).

Components from Analog Devices (AD8066, AD8130, AD835) and Texas Instruments (OPA861) were installed following advice supplied in their application notes. Frequencies ranged from 50kHz to 1MHz. Where possible devices were ac decoupled. The dc offsets in
the circuits surrounding the multipliers and low pass filter (integrator) were trimmed to within about 2 mV. This provided a suitably low baseline in spectra measurements (figures 8 and 9). Test equipment used two four-channel signal generators (TGA12104, Thurlby Thandar) and a spectrum analyser (E4411A, Agilent Technologies). The amplifier $A_v$ provided a gain of 40 V V$^{-1}$ before the multiplier filter (which had an attenuation of 0.5) to ensure that errors in the multipliers due to offsets and mismatch were minimised. The multiplier filters were adapted from those in Langlois et al 2014. The input drive sinewave input was restricted so that the response could be observed over a reasonable frequency range before reaching the input limit of the multipliers. For current feedback the combination of $A_v$ before the filter and the output transconductor $G_2$ provided a transconductance $G$ (figure 2) of 0.022 A V$^{-1}$. The transconductors $G_1$ and $G_2$ had output resistances of about 50 kΩ. For voltage feedback the transconductor $G_2$ was replaced by an AD8066 unity gain voltage buffer. Common mode measurements attenuated by approximately 0.2 in the connection to the spectrum analyser were monitored at $V_s$ in figure 2 (where $V_s = 2V_{cm}$).

The performance when applying two sinewave signals simultaneously is shown in figures 8 and 9 for current and voltage feedback, respectively. Figures 8(a) and 9(a) show the common
mode signals without feedback. For the current feedback the feedback transconductor input was set to 0 V. For the voltage feedback transconductor \( G_2 \) was replaced by an AD8066 configured as a unity gain voltage buffer.

Figures 8(b) and 9(b) show the common mode signals with feedback. Common mode reduction was about 62 dB for current feedback and 31 dB for voltage feedback compared with 67 dB and 32 dB from table 1. The amplitude of the input ac currents differed for current and voltage feedback measurements. The input currents, of the order of 20 \( \mu \)A, were adjusted to provide very approximate similar feedback common mode amplitudes.

Current mode feedback operated at lower frequencies than voltage feedback. A 3 dB drop in common mode attenuation occurred at 300 kHz. In current feedback from table 1(b) the common mode attenuation is \( G_{Z_{cm}} \). Capacitive strays, \( C_s \) in figure 7, at higher frequencies
will significantly reduce \( Z_o (= C_s || R_o) \) and therefore the common mode attenuation. Since \( R_o = 50 \, \text{k}\Omega \) the 300kHz pole indicates \( C_s \) is about 10pF. For voltage feedback from table 1(b) the common mode attenuation is \( 1/2A \), and is (to a first order) independent of both \( Z_o \) and \( Z_L \) and therefore can operate at frequencies mainly limited by the high frequency limit of the voltage feedback \( A \) (figure 2).

An example of the spectrum for dual frequency feedback is shown in figure 10 for voltage feedback. To observe the spectrum of the common mode feedback a single sinewave was swept over the frequency range. The minimum frequency difference between a chosen frequency \( f_c \) and its adjacent frequency \( f_n \) causes a component \( (f_c - f_n) \) at the output of the first multiplier (figure 3) which must be attenuated by the integrator RC in the same manner as the \( 2f_c \) frequency component. The integrator RC (primarily chosen to sufficiently attenuate the lowest expected \( 2f_c \) component to suppress distortion (Langlois et al 2014)) will sufficiently attenuate the \( (f_c - f_n) \) component if \( (f_c - f_n) > 2f_c \). A designed lowest operation frequency of 10kHz limits the minimum frequency difference to 20kHz.

5. Discussion

Without feedback the application of a connection \( Z_o \) from the body to ground can substantially reduce the common mode voltages. However on applying feedback, analysis and simulation show that the resulting steady state common mode error voltage changes very little with \( Z_o \). It is the step response which can be significantly affected.

The measured results show that a useful common mode reduction can be expected with dual frequency input signals. With single pole filters in the multipliers (figure 3) there is some harmonic breakthrough due to the remaining ripple on \( V_{c1} \) illustrated in figure 4(b). The addition of extra frequencies each with its appropriate narrowband frequency common mode feedback will add further harmonic components as a result of the frequency difference between two adjacent applied frequencies. This limits how close adjacent frequencies can be allowed for adequate suppression by the low pass filter RC in figure 3. The sum of input currents from multiple frequency signals cannot exceed an upper safe limit, so each signal must be reduced proportionately. The addition of unwanted signals and the reduction of drive current (and therefore differential output signals) will limit the number of frequencies possible.
Outside the locking frequencies there is little feedback. In the case of current feedback the load then remains floating, which allows capacitive coupling from external devices. For voltage feedback outside the chosen frequencies one side of the load is held (via an electrode) at earth. In practice impedances, notably of the electrodes and the load, and strays will modify the results.

Measurements have been performed with standard discrete components. The loop gain for voltage feedback was relatively low compared with that of current feedback. There were limitations, in particular in the multipliers, figure 3, with their built in dc offsets affecting the low pass filter accuracy. The loop gain can be increased by increasing the gain of $A_{dc}$ in the multiplier filters in figure 3. An integrated circuit design using a switching multiplier (Godoy and Dawson 2008) as the first stage multiplier and dc amplifiers with dc offset compensation (Enz and Temes 1996) can be used to boost the loop gain.

6. Conclusion

The problem of large common mode voltages at higher frequencies in bioimpedance applications can be suppressed by adding narrowband common mode feedback. It has been successfully employed using two drives with different frequencies simultaneously. Formalised analysis of both current and voltage narrowband feedback has been presented for current input drives. The limit of the number of additional current drives will be dictated by the accumulating background noise generated by the imperfections in the narrowband feedback components. Experimental results for two simultaneous frequencies using standard available components show a reduction of 62 dB for current feedback and 31 dB for voltage feedback.

Acknowledgments

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Appendix A. Analysis of the balanced current feedback (figure 2(a))

The balanced current feedback in figure 2(a) has been selected as an example to demonstrate the process of analysis. The other feedback analyses follow a similar procedure. Figure 11 shows the balanced current feedback of figure 2(a). For simplicity the output impedance \( Z_{oG} \) of the transconductor \( G \) is assumed to be infinite and the gain of the filter to be unity. The loop gain of the feedback is then \( Z_{o}G \). It is also assumed that \( Z_{o} \gg Z_{e}, Z_{L}/2 \).

The current drives with their output impedances are replaced by their dual voltage drives. For analysis, the contributions of \( I_{+} \) and \( I_{-} \) to \( V_{+} (=2V_{cm}) \) can be split and, using the principle of superposition, then combined. Figure 12 shows the two circuits.

Referring to figure 12(a), since \( Z_{o} \gg Z_{e} \) the voltage \( V_{x+} \) is

\[
V_{x+} = I_{e}Z_{o} \left[ \frac{Z_{e}/2 + (Z_{o} + Z_{L}/2)}{Z_{o} + Z_{L}/2 + (Z_{o} + Z_{L}/2) \| Z_{b}} \right] \approx I_{e}Z_{o} \left( \frac{Z_{e}/2 + Z_{o} \| Z_{b}}{\sum Z} \right)
\]

(A.1)

where \( \sum Z \approx Z_{o} \| Z_{b} \). Similarly the voltage \( V_{x-} \) is

\[
V_{x-} \approx I_{e}Z_{o} \left( \frac{Z_{o} \| Z_{b}}{\sum Z} \right)
\]

(A.2)

Similarly in figure 12(b)

\[
V_{x-} \approx -(I_{e}Z_{o} + Z_{o}G_{V_{x}}) \left( \frac{Z_{e}/2 + Z_{o} \| Z_{b}}{\sum Z} \right)
\]

(A.3)

and

\[
V_{x+} \approx -(I_{e}Z_{o} + Z_{o}G_{V_{x}}) \left( \frac{(Z_{L}/2) + Z_{o} \| Z_{b}}{\sum Z} \right)
\]

(A.4)

By the principle of superposition

\[
V_{x} = V_{x+} + V_{x+} + V_{x-} + V_{x-}
\]

(A.5)
and using (A.2)–(A.5) yields

\[ V_I = \frac{2\Delta I Z_0}{Z_0 + Z_o \| Z_b} \left( \frac{Z_L/2}{Z_b} + 2Z_o \| Z_b \right) - \frac{(GVZ_0)(Z_L/2) + 2Z_o \| Z_b}{Z_o + Z_o \| Z_b} \]

(A.6)

where \( 2\Delta I = I_+ - I_- \).

Rearranging (A.6) gives

\[ V_{cm} = \frac{V_I}{2} = \Delta I Z_0 \left( \frac{Z_L/2}{Z_b} + 2Z_o \| Z_b \right) \frac{1 + GZ_0(Z_L/2) + 2Z_o \| Z_b}{Z_o + Z_o \| Z_b} \]

(A.7)

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