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A High-Voltage Bipolar Transconductance Amplifier for Electrotactile Stimulation

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Abstract

This article describes a high-performance transconductance amplifier specifically designed for electrotactile (electrocutaneous) stimulation. It enables voltages up to ± 600 V to be produced at the output which will allow the psychophysiological performance associated with stimulation of the fingertip using various stimulation waveforms to be studied more thoroughly. The design has a transconductance of up to 20 mA/V, an 8.8-M Ω output resistance, and can provide output currents up to ± 20 mA. A complete schematic diagram is presented along with a discussion of theory of operation and safety issues as well as performance and derating plots from the implemented design.

Keywords

Current control; electrocutaneous; electrotactile; transconductance amplifier

I. Introduction

ELECTROTACTILE stimulation passes a localized electric current through the skin in order to stimulate underlying touch nerves and thus evoke tactile sensations. This process can be used in sensory substitution—receiving via one sense information that is normally received by another sense. In the case of electrotactile stimulation, it is often vision or hearing information that is received by touch. The principles and applications of electrotactile stimulation and sensory substitution have been summarized in several review articles [1-4,5 (Chapter 7), 6,7 (pp. 12:42–48), 8,9].

The impedance of the skin can be quite variable due to environmental conditions, sweat gland activity, skin type, etc., and also decreases significantly with stimulation current [10-16]. Therefore, the stimulation current, rather than the voltage, is usually controlled. The most challenging skin location for electrotactile stimulation is the fingertip, for which the resistive component of the impedance (which also has a capacitive component) has been reported as high as 320 k Ω [17]; we have observed even higher values in our lab. It is therefore important that the proposed circuit have a very high output impedance (ideally infinity) so that the delivered current is insensitive to variations in load impedance. High voltage capability (compliance, up to 700 V, more typically 150–300 V) is also required, as is a slow rate consistent with the particular application or experiment; we set 50 V/ μ s as a design goal. These

requirements (explored in more detail in [17]) are difficult to meet using extant commercial instrumentation, published circuit designs and semiconductor devices.

Our high-voltage bipolar transconductance amplifier (HVB-TA) is based on an earlier design we have used in our lab for 15 years [18], which had a maximal voltage capability of ± 120 V, or with a modified asymmetric power supply, up to 240 V positive or negative pulses, but not both. It had an output resistance of 2 M Ω , adequate for electrotactile stimulation on hairy skin but not for glabrous fingertip skin, which is much thicker and has a much higher impedance. Modifications to this circuit include use of stacked darlington transistor pairs to achieve higher voltage capability using low-power semiconductors¹ and the addition of an additional feedback loop to achieve a higher output resistance.

The only other transconductance amplifier design we are aware of that approaches these requirements was described by Poletto [17]. That design used high-voltage operational amplifiers arranged in a Howland current pump configuration to achieve voltage-to-current conversion. The chosen semiconductor devices (Apex Microtechnology PA85A) have a maximal rail-rail power supply voltage of 450 V and output swing of approximately 400 V peak-peak, or ± 200 V for bipolar pulses. Achieving higher potentials (± 400 V bipolar, reconfigurable for higher-potential asymmetric pulses) requires a bridge configuration, necessitating either floating the load or the entire amplifier circuit. The former option is not desirable for safety and performance reasons. If coaxial stimulation electrodes are used, the typically large return electrode, tied to one side of the bridge, will cause the average body potential of a human subject to achieve one-half of the total stimulation potential, causing a hazard should the subject touch an external conductor and causing a significant displacement current to flow otherwise. Conversely, floating the amplifier necessitates a complex isolation scheme for both the power supplies and the input circuitry. The latter was achieved using a combination of analog and digital optical isolators to preserve both pulse level and sharp edge transitions. This scheme, however, limits the use of arbitrary wave shapes and necessitates separate analog and digital inputs.

Poletto's design has excellent feedback-based current control, fast rise times (< 1 μ s), and is capable of delivering indefinitely long current pulses undistorted. However, several disadvantages prompted us to develop an alternative circuit: (1) The output potential swing is too low for full-range (± 700 V) biphasic pulses on the fingertip, (2) The input waveform is limited and requires separate analog and digital inputs, (3) The operational amplifiers are expensive ($> \text{US\$}200$) and power-hungry (> 20 mA quiescent current, yielding 9 W power dissipation with ± 225 V rails), and (4) the circuit topology is very sensitive to component selection and physical layout to prevent instability and failure of the operational amplifiers. Electrotactile stimulator designs by other investigators [e.g. 19,20-22], being designed for specific applications, do not have specifications adequate for our research.

The design presented here is capable of producing $+600/-800$ V, ± 20 mA, bipolar current-controlled waveforms given the appropriate input voltage waveform. With ± 800 V rails, the quiescent current (power) of each amplifier channel is approximately 1.0 mA (1.6 W). Each amplifier channel incorporates quite a few discrete components (28 transistors, 4 analog integrated circuits, and several dozen passive components) but even with predominantly through-hole mounting the printed circuit board size is only 10.7×14.3 cm. Up to 16 amplifier channels can fit in a standard 6 in (15.2 cm) high, 19 in (48.3 cm) wide equipment rack. Each amplifier incorporates provisions for adjusting gain, output resistance, and compensation for

¹We chose low-power semiconductors because they have better amplification characteristics than high-power devices in the current region of interest, 0.5–20 mA. Low-power devices are available with breakdown potential of only 400 V, necessitating the voltage division scheme.

stray capacitance in the electrode wiring. We include here a complete schematic diagram, theory of operation, and detailed performance specifications measured for various resistive and cutaneous electrode loads. Table 1 shows a list of key specifications (see performance section for further details).

II. THEORY OF OPERATION

By definition, a transconductance amplifier takes a voltage as an input and produces a directly proportional current at the output. An ideal transconductance amplifier has an infinite output resistance so that the output current is independent of load impedance, and the output current waveform would have an identical shape as the input voltage waveform. A real TA has a finite output resistance, reducing the accuracy of current control, and there is furthermore noticeable rounding of the current waveform due to the effect of stray capacitance. To compensate for these imperfections, feedback control loops have been implemented and will be discussed below, along with the general operation of the TA itself. Note that to simplify circuit analysis among the transistors, the operation of the positive half of the circuit will be described and it can be assumed that the negative half will behave similarly.

Four figures provide both a complete schematic drawing as well as illustrate the key feedback loops in the circuit. Figure 1 shows the essential TA circuit, and is called the Base Amplifier. Figure 2 shows the Complete Amplifier, including the additional feedback loops compensating for stray capacitance and output conductance. Figure 3 conceptually shows the four feedback loops. Figure 4 shows the power supply.

Current Control Loop

In the Base Amplifier (Figure 1), the input voltage v_i is converted into a current by the loop formed by U_{sumB} , Q_8 , R_9 and C_{if} (bounded by dashed line). Q_7 is diode-wired and thermally-bonded to Q_8 to minimize thermal drift in the quiescent operating current, which is adjustable by varying R_1 . The loop is also represented in Figure 3 by the block containing C_{if} . This current feedback loop causes the pulse voltage at v_{test} to be nearly equal to the input voltage, v_i . It can be assumed that either the positive or negative half of the circuit will be turned on at an instance in time, but not both. Therefore, the current through R_9 will either take the path through R_7 or R_8 , but not both². Using

$$i_{E8} = \frac{v_{test}}{R_9} \quad (1)$$

and

$$i_{C8} = \frac{\beta}{\beta+1} i_{E8} \quad (2)$$

this input voltage is converted into a current. ($\beta > 90$ for the current range of interest, 1–20 mA). This current conversion and the current mirror to be described next operate similarly to our earlier design [18].

Current Mirror

The Wilson current mirror formed by Q_{31} , Q_{32} and Q_{17} reflects Q_{31} 's collector current, i_{C31} , to Q_{17} 's collector current, i_{C17} . Note that i_{C31} is a function of i_{C8} , and the output current, i_{out} , is a function of i_{C17} . This is important because it allows the load to be grounded and therefore

²Actually, each circuit half (e.g. R_7 or R_8) has a dc quiescent current of 0.3–0.5 mA to prevent crossover distortion. Currents in this analysis represent deviations from this quiescent operating point. R_7 and R_8 serve to stabilize the dc operating point but do not contribute to voltage-to-current conversion.

multiple electrodes can share a common return path. The Wilson current mirror has three characteristics that suit it well for this application. First, it has better matching of input and output current than other methods. Second, it is less dependent on transistor β variations. Third, it does not require the β of Q_{17} to be matched to that of Q_{31} and Q_{32} and Q_{17} can therefore be implemented by a high-voltage transistor [18]. Assuming all the transistors in the circuit have a similar current gain, and Q_{31} and Q_{32} are a matched pair, Q_{17} 's collector current, i_{C17} , can be represented as a function of i_{C31} by:

$$i_{C17} = \frac{1}{1 + \frac{2}{\beta^2 + \beta}} i_{C31} \quad (3)$$

Also, if Equation (2) is applied to transistors Q_2 , Q_4 , Q_6 , Q_{18} , Q_{20} and Q_{22} , then

$$i_{C31} = \left(\frac{\beta}{\beta + 1} \right)^3 i_{C8} \quad (4)$$

and

$$i_{C17} = \left(\frac{\beta + 1}{\beta} \right)^3 i_{out}. \quad (5)$$

Static Transfer Function

The static transfer function of this part of the circuit can be formed by combining (1)–(5):

$$g_m = \frac{i_{out}}{v_i} = \frac{\beta^8}{(\beta + 1)^6 (\beta^2 + \beta + 2)} \left(\frac{1}{R_9} \right). \quad (6)$$

Since $R_9 = 100 \Omega$ and $\beta \approx 100$, this is about 9.33 mA/V. When measured, g_m was approximately 9.6 mA/V. Notice that in (6), g_m is relatively independent of load resistance and power supply voltage. Also, β may vary with temperature, but a 10% change in β will result in only a 0.6% change in g_m .

DC stabilization loop

The current feedback loop is capacitively-coupled by C_{if} and hence will not transfer any dc voltage. However, R_{dc} (i.e. $R_{dca} + R_{dcb} + R_{dcc}$) provides negative-feedback dc from the output, as shown in Figure 3. The circuit's large voltage gain forces the dc component of out_{dc} to track the small dc component seen at v_i , for any conceivable electrotactile stimulation waveform. The resulting net dc load current is therefore minimized.

Output ac Coupling

If one of the output transistors were to fail, a dangerous dc current could flow from the high-voltage supply to the subject. To prevent this, capacitors C_1 and C_2 couple the electrode to output transistors Q_{22} and Q_{25} . Any residual dc drop that may develop across these capacitors is equally distributed by R_{24-25} .

Bias Point Stabilization

The quiescent current through Q_8 and Q_{10} is maintained at about 0.4 mA by R_1 and R_7 . It can be adjusted through R_1 , which is implemented as a variable resistor. Q_7 is diode-wired and thermally linked to Q_8 to reduce bias point variations due to temperature changes. A similar quiescent current can be assumed to travel down the output stack of transistors (Q_{17} – Q_{23}). Also, a small current will go through R_3 – R_6 and R_{15} – R_{18} . The quiescent current delivered to the “positive” input stack of transistors (Q_1 – Q_6) and resistors (R_3 – R_6) is therefore about

$$i_{bias} = \frac{15 - V_{7BE}}{R_1} + \frac{V_3}{8 \times 10^6} \quad (7)$$

or approximately 0.4 + 0.1 mA for $V_3 = 800$ V, which means that the total quiescent current measured through R_{14} is double that, or about 1.0 mA.

As mentioned earlier, the Base Amplifier has a nonzero output admittance that reduces the accuracy of current control and distorts the current waveform. The Complete Amplifier in Figure 2 shows the addition of two feedback loops that compensate for these effects, as well as a differential input configuration for noise reduction.

Output Stray Capacitance Compensation Loop

A stray capacitance, C_{st} , ranging from 40 to 200 pF, may develop at different points in the circuit, output cable or electrode array. This essentially creates a low-pass filter at the output and distorts the output current waveform at the electrode load, Z_L . To compensate for this, $C_f = C_{f1}C_{f2}/(C_{f1} + C_{f2})$ provides a differentiated positive-feedback path from the TA output to its input. In the equation

$$I_L(s) = I_o(s) - sC_{st}V_o(s) - g_{st}V_o(s) \quad (8)$$

$$= g_m V_i(s) - sC_{st}V_o(s) - g_{st}V_o(s), \quad (9)$$

where the $g_{st}V_o(s)$ term is a function of the conductance due to the circuit having a non-infinite output resistance and the $sC_{st}V_o(s)$ term is a result of the stray capacitance, the stray capacitance compensation signal component is averaged into $V_i(s)$. This component is labeled as v_3 in the circuit schematic and it is a function of $V_o(s)$ by

$$\begin{aligned} V_3(s) &= \left(1 + \frac{R_{f4}}{R_{f3}}\right) \left(\frac{sC_f R_{f5}}{1 + sC_f R_{f5}}\right) \{R_{p3}\} V_o(s) \\ &= \left(\frac{122.95s}{(1.67)(10^9) + s}\right) \{R_{p3}\} V_o(s). \end{aligned} \quad (10)$$

Note that $\{R_{p3}\}$ refers to the potentiometer's wiper's fractional position from ground (i.e. 0–1). Equation (9) then becomes:

$$I_L(s) = \frac{g_m}{3} (V_1(s) + V_2(s) + V_3(s)) - sC_{st}V_o(s) - g_{st}V_o(s) \quad (11)$$

$$= \frac{g_m}{3} \left[\begin{aligned} &V_1(s) + V_2(s) + \\ &\left(1 + \frac{R_{f4}}{R_{f3}}\right) \left(\frac{sC_f R_{f5}}{1 + sC_f R_{f5}}\right) \{R_{p3}\} V_o(s) \\ &- sC_{st}V_o(s) - g_{st}V_o(s) \end{aligned} \right] \quad (12)$$

$$= \frac{g_m}{3} \left[\begin{aligned} &V_1(s) + V_2(s) + \\ &\left(1 + \frac{R_{f4}}{R_{f3}}\right) \left(\frac{sC_f R_{f5}}{1 + sC_f R_{f5}}\right) \{R_{p3}\} \\ &- \frac{3sC_{st}}{g_m} \end{aligned} \right] V_o(s) - g_{st}V_o(s). \quad (13)$$

Since sC_f is very small relative to R_{f5} , $\left(\frac{sC_f R_{f5}}{1 + sC_f R_{f5}}\right)$ can be simplified to approximately $sC_f R_{f5}$, and so

$$I_L(s) = \frac{g_m}{3} \left[\left[\frac{V_1(s) + V_2(s) + \left(\left(1 + \frac{R_{f4}}{R_{f3}} \right) (C_f R_{f5}) \{R_{p3}\} - \frac{3C_{st}}{g_m} \right) V_o(s)}{s} \right] - g_{st} V_o(s) \right] \quad (14)$$

Therefore if $\{R_{p3}\} \approx \frac{3C_{st}}{g_m C_f R_{f5} \left(1 + \frac{R_{f4}}{R_{f3}} \right)}$, the effect of stray capacitance, C_{st} , on the output current waveform will nearly disappear, independent of the load impedance, Z_L . Note that with $\{R_{p3}\} = 1$, up to 229 pF of stray capacitance may be compensated for³.

Output Resistance Compensation Loop

The ideal current source will have an infinite output resistance, R_{out} , so that the same current will be always be delivered regardless of the load. The actual R_{out} of the circuit without compensation is about 717 kΩ, which is too low⁴. As was mentioned previously, the variable that represents this non-ideality is g_{st} , which can be seen in Eq. (14). In order to compensate for this, a small positive-feedback voltage is brought back to the input at v_2 from the output by

$$\begin{aligned} v_2 &= \left(1 + \frac{R_{f2}}{R_{f1}} \right) \left(\frac{R_{f6}}{R_o + R_{f6}} \right) \{R_{p2}\} v_o \\ &= (7.74) (10^{-4}) \{R_{p2}\} v_o \end{aligned} \quad (15)$$

where $R_o = R_{oa} + R_{ob} + R_{oc}$. R_{p2} is adjusted to attenuate this feedback voltage and acquire the highest R_{out} possible without causing the circuit to become unstable for an open-circuit load. The measured R_{out} is then about 8.8 MΩ, which is acceptable. Combining (14) and (15), the mathematics behind this compensation loop is clarified in the load current equation by:

$$I_L(s) = \frac{g_m}{3} \left[\left[\frac{V_1(s) + \left(\left(1 + \frac{R_{f2}}{R_{f1}} \right) \left(\frac{R_{f6}}{R_o + R_{f6}} \right) \{R_{p2}\} + \left(\left(1 + \frac{R_{f4}}{R_{f3}} \right) (C_f R_{f5}) \{R_{p3}\} - \frac{3C_{st}}{g_m} \right) V_o(s)}{s} \right] - g_{st} V_o(s) \right] \right] \quad (16)$$

$$= \frac{g_m}{3} \left[\left[\frac{V_1(s) + \left(\left(1 + \frac{R_{f2}}{R_{f1}} \right) \left(\frac{R_{f6}}{R_o + R_{f6}} \right) \{R_{p2}\} - \frac{3g_{st}}{g_m} \right) V_o(s)}{s} + \left(\left(1 + \frac{R_{f4}}{R_{f3}} \right) (C_f R_{f5}) \{R_{p3}\} - \frac{3C_{st}}{g_m} \right) V_o(s) \right] \right] \quad (17)$$

Therefore, if $\{R_{p2}\} \approx \frac{3g_{st}(R_o + R_{f6})R_{f1}}{g_m R_{f6}(R_{f1} + R_{f2})}$, the term containing g_{st} is eliminated.

³Under-compensation, i.e. $\{R_{p3}\}$ too small, results in slow dynamic response; overcompensation, i.e. $\{R_{p3}\}$ too large, results in ringing or oscillation. The maximal stray capacitance compensation was tested by setting $\{R_{p3}\} = 1$ and adding capacitance across a 500-kΩ load until ringing disappeared, approx. 600 pF. This exceeds the theoretical value of 229 pF, probably due to stray circuit board capacitance in parallel with the tiny feedback loop capacitor C_f (6 pF).

⁴ R_{out} may be calculated by first measuring the output current amplitude for a certain input voltage pulse with a purely resistive load, and the output current amplitude for the same input voltage using a different resistive load. Using current division equations,

$$R_{out} = \frac{I_{o1}R_{L1} - I_{o2}R_{L2}}{I_{o2} - I_{o1}}.$$

Gain Adjustment

Potentiometer R_{p1} is implemented as a front panel knob which allows for adjustment of the gain. Jumpers W_4 - W_7 have been included to easily adjust the gain of the instrumentation amplifier to 1, 2, 4 or 8 V/V. Also, $R_{gaintrim}$ was included in this design so that the gains of each channel can be calibrated to be equal for a certain gain knob setting.

Differential Input

Finally, the instrumentation amplifier, U_{inst} , provides a differential input and thus most common-mode noise in the input cabling is rejected. Therefore, the input component that is summed to form v_i is a function of $(v_{in+} - v_{in-})$. Note that due to the very high input impedance of U_{inst} the inputs will pick up a lot of noise if left floating. Hence R_{pin} and R_{nin} ensure that the inputs will not float.

Overall Transfer Function

$V_I(s)$ in (17) is equal to $(V_{in+}(s) - V_{in-}(s))\{R_{p1}\}$, and noting that $V_o(s) = I_L(s)Z_L(s)$, the circuit's entire static transfer function can be written as:

$$\frac{I_L(s)}{V_{in+}(s) - V_{in-}(s)} = \left\{ R_{p1} \right\} g_m / \left[3 - g_m Z_L(s) \left[\left(1 + \frac{R_{f2}}{R_{f1}} \right) \left(\frac{R_{f6}}{R_o + R_{f6}} \right) \{ R_{p2} \} - \frac{3g_{st}}{g_m} + s \left(\left(1 + \frac{R_{f4}}{R_{f3}} \right) (C_f R_{f5}) \{ R_{p3} \} - \frac{3C_{st}}{g_m} \right) \right] \right] \quad (18)$$

Note that the denominator of this equation is of the form:

$$3 - g_m Z_L(s) \left(K_1 \{ R_{p2} \} - K_g g_{st} + K_2 \{ R_{p3} \} - K_c C_{st} \right) \quad (19)$$

where $K_g g_{st}$ is the term due to the finite output resistance, K_1/R_{p2} is the adjustable term to compensate for that, $K_c C_{st}$ is the stray capacitance resulting term and K_2/R_{p3} is the corresponding adjustable compensation term. If the compensation terms are adjusted correctly, it is evident that the majority of this denominator will cancel out and the equation simplifies to the form of an ideal TA, with terms dependant on the load impedance eliminated:

$$\frac{I_L(s)}{V_{in+}(s) - V_{in-}(s)} = \frac{\{ R_{p1} \} g_m}{3} \quad (20)$$

Feedback Paths

Figure 3 summarizes the feedback loops of the circuit in a more conceptual manner. The current-control loop is formed by the C_{if} block. Likewise, the dc-stabilization loop is formed by the R_{dc} block. These two negative feedback paths implement voltage-to-current conversion and maintain dc stability. Two positive feedback paths compensate for stray capacitance (C_f and K_{Cst} blocks) and output circuit conductance (R_o and K_{Ro} blocks).

Semiconductor Selection

With maximal ± 800 V power supply voltages, Q_2 , Q_4 , Q_6 , Q_8 , Q_{10} , Q_{11} , Q_{13} , Q_{15} , Q_{17} , Q_{18} , Q_{20} , Q_{22} , Q_{25} , Q_{27} , Q_{29} and Q_{30} need to withstand 400 V, yet they must operate at low currents of 0.1 to 50 mA. Although the specified V_{CE} breakdown voltage of 400 V is adequate, the ones used were individually tested to withstand at least 500 V to reduce the possibility of failure if the maximal rail voltages are used. The Darlington pair configuration is used with the majority of these for its high gain characteristic. Matched pair transistors Q_{31-32} , and Q_{33-34} , were selected for their collector current rating as well as flat β and matched characteristics from 0.1 to 50 mA. The only ones found to meet our needs are only available in a surface mount package.

Lastly, as previously mentioned, Q_7 and Q_9 are glued to Q_8 and Q_{10} , respectively, in order to provide a thermal link.

Power Supplies

The circuit uses both low-voltage (± 15 V) and high-voltage (± 800 V) for operation as shown in Figure 4. In order to minimize output transients, proper power sequencing must be observed. To implement this, a three-position power switch (off, low-voltage, high-voltage) is used. When switched from “off” to “low-voltage”, only the ± 15 V supply is turned on. This activates the feedback paths and amplifier, thus giving the circuit time to stabilize at its dc operating point. When switched to “high-voltage”, the ± 800 V supply is applied to the now stable circuit. If the power supplies were sequenced in the opposite order or even at the same time, the high voltage would be applied to an essentially unstable circuit, and the output is momentarily unpredictable.

To increase usability, two high-voltage power supply schemes are available. One, implemented by miniature DC-DC converters, is typically intended to be used with up to three channels due to current limitations.⁵ The second high-voltage supply scheme uses external high-voltage power supplies.

III. SAFETY FEATURES

There are a number of safety features incorporated into the design to prevent the subject from electrical injury, which are explained below:

Redundancy

The TA circuit schematic shows that three faults (e.g. shorted C_1 , shorted C_2 and a shorted output transistor, assuming other transistors would then fail in a chain reaction) are required to connect the subject to the high voltage and pass the full power supply current to them. If using the internal power supplies, the maximum output current from each is 5 mA. External power supplies may also have the ability to limit their output current. Note that these three failures are a worst-case scenario and extremely unlikely. For less than three failures, less than 0.1 mA DC will be applied to the output.

Charge-Limiting Output Coupling

The output coupling capacitors are of low enough value to ensure that the charge flowing into any single electrode under any condition will not exceed 160 μC . This is less than 290 μC , the *Underwriters Laboratories* limit for capacitive discharges for a 0.1 μF capacitance [5].⁶

Power Supply

A 120 VAC-12 VDC converter which brings the power to the circuit is a medical-grade separate external unit. Therefore, dangerous AC voltages are kept apart from the actual instrument.

⁵The specified HV converters deliver 5 mA, while each amplifier channel consumes 1.7 mA from the HV rails for a 20-mA, bipolar, 1% duty cycle output. This is a high current and duty cycle for electrotactile stimulation. The amplifier channels may operate at even higher duty cycles (see Table 1), but this requires more power and therefore the three-channel limit is reduced.

⁶Our design is not UL tested or rated and is intended to be used by knowledgeable personnel for experimental purposes. The 160- μC fault rating assumes that one output capacitor is shorted, and that the other capacitor is pre-charged to 800 V by maximal conduction in one output polarity transistor stack, and that the opposite polarity stack subsequently shorts (an extremely unlikely scenario). The 0.1- μF capacitor then reverses polarity, requiring 160 μC for the resulting 1600-V voltage swing.

IV. PERFORMANCE

We built and tested the HVB-TA design, which has been in use in our lab for over one year for research completed and in progress [23,24,16]. Fig. 5 shows a photograph of one complete amplifier channel (including the Base Amplifier in Fig. 1 and the feedback loops and input circuitry in Fig. 2). The power supply is on a separate module, not shown. The circuit was implemented using mostly through-hole components on a 2-layer printed circuit board. Surface-mount techniques and a more efficient layout could be used to achieve a more compact design.

Figures 6 and 7 show the actual current waveform delivered to various resistive and fingertip electrode loads in response to a square input voltage (v_i) pulse with a risetime of approx. 2 μ s. As expected, higher-valued resistive loads showed larger current risetimes, consistent with greater RC time constants, C representing primarily the uncompensated stray capacitance in the output circuit and cabling. Electrode loads, which have significant capacitive components [16], showed faster rise times than purely resistive loads of the same resistance value. Figure 8 shows the electrode voltage for the same stimulation conditions used for Fig. 7. The faster rise time at higher currents is attributable to a reduction of both resistive and capacitive components of the electrode-skin load at higher currents [25, 16].

Figure 9 shows the measured maximal voltage and current capability of the HVB-TA for various resistive loads similar to typical fingertip resistances. (Significant portions of these curves would result in painful stimulation if real electrodes were used.) Although these derating curves reflect primarily the fundamental voltage and current limits of the circuit, certain asymmetries not observed in simulation indicate the possibility of improvement by better physical layout for reduction of stray capacitance. See legends for Figs. 6-9, for further details on the performance of the HVB-TA as actually built.

V. CONCLUSIONS

The design presented above is a stable and reliable transconductance amplifier to be used in an electrotactile stimulation system for the fingertip. It is an improvement to past designs because of its stability and ability to reliably produce bipolar current pulses up to ± 600 V. Feedback loops have been incorporated to optimize the output for typical stimulation waveforms. These features will allow electrotactile stimulation of the fingertip to be studied more thoroughly in future experiments.

Biography

Matthew A. Schaning received the B.S. degree in biomedical engineering from Marquette University, Milwaukee, WI in 2004, and the M.S. degree in biomedical engineering from the University of Wisconsin, Madison, in 2005. He studied biomedical instrumentation and sensory substitution. Since 2005, he has been an Engineer in the Research and Development Department, Energy-Based Devices, Covidien, Boulder, CO. His current research interests include tumor ablation and instrumentation.

Kurt A. Kaczmarek (M'86–SM'05) received the B.S. degree in electrical engineering from the University of Illinois, Urbana, in 1982 and the M.S. and Ph.D. degrees, both in electrical engineering, from the University of Wisconsin- Madison, in 1984 and 1991, respectively. From 1984 to 1986, he was a Senior Engineer with Baxter International, Deerfield, IL. He is currently a Senior Scientist at the University of Wisconsin- Madison, where, since 1992, he has studied the mechanisms and perception of electrical stimulation of touch. His interests include tactile displays, sensory rehabilitation and augmentation, teleoperation, and neurorehabilitation.

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Fig. 1. Base Amplifier. This is the main voltage-to-current converter circuit. A positive input voltage v_i is converted into a current signal in Q_8 by the components encircled by dashed lines (negative inputs are handled by Q_{10} ; see main text). The matched transistors Q_{31} and Q_{32} reflect the current in the transistor stack above Q_8 back down the transistor stack below Q_{32} . Therefore, the output current i_L is close to v_i/R_9 . The darlington-configured transistor stacks divide voltage evenly to achieve high voltage amplifier capability with readily-available semiconductor devices.

Fig. 2. Complete Amplifier. This circuit shows how the Base Amplifier is supplemented by feedback loops that compensate for stray capacitance and conductance in the Base Amplifier and output cabling. The instrumentation amplifier PGA206P is by Texas Instruments.

Fig. 3. Feedback Loop Summary. The overall HVB-TA has four feedback loops. The current control loop and dc stabilization loop are integral parts of the Base Amplifier (Fig. 1) and essential to its operation. Two additional loops external to the Base Amplifier compensate for stray capacitance and conductance in the output circuit and related cabling.

Fig. 4. Power Supply. Power supply modules HVP1P and HVP1N are by Pico Electronics (Pelham, New York), power supply module BST-15/100-D12 is by Datel/Murata (Mansfield, Massachusetts), and digital displays SP400 are by Lascar Electronics (Erie, Pennsylvania).

Fig. 5. Photograph of one complete amplifier channel. The printed circuit board has dimensions of 10.7×14.3 cm and is mounted on an equipment rack module ($13.3 \times 16.9 \times 2.5$ cm). At left are the front panel gain knob and BNC input and output connectors (not included in module dimensions). The trim potentiometers on the top and bottom left edge of the board are the gain trim, bias current, and resistance and capacitance compensation adjustments. The connector on the right side of the board receives power from a separate power supply module ($13.3 \times 16.9 \times 5.0$ cm) mounted in the same rack.

Fig. 6.

Dynamic response for a nominal 0.5-mA, 250-μs current pulse delivered to resistive (1–1000 kΩ) and typical fingertip electrode (dotted line) loads. This low (and subthreshold) current enables time response to be compared across a wide range of load resistances without exceeding the HVB-TA output voltage limit. The stainless steel fingertip electrode had a coaxial structure with a 0.9-mm-diameter center electrode and an annular surround with inner and outer diameters of 4.8 and 6.4 mm, respectively. The Fing-dry data were recorded within one minute after the electrode was placed onto a dry patch of fingertip skin. Fing-wet (hydrated) conditions were recorded 5 min after an electrode was placed on fingertip skin, allowing sweat to build up under the electrode and reduce skin resistance. Current was measured using a 300-MHz digital oscilloscope and a 100-Ω current-sampling resistor in the return current path; for the electrode, this was the surround. The HVB-TA R_0 compensation loop was adjusted to minimize the difference in asymptotic maximal current between 1-kΩ and 500-kΩ resistive loads, i.e., to minimize the effect of load resistance. This is equivalent to maximizing the output resistance of the circuit. The C_0 compensation loop was adjusted for best dynamic response (i.e., fastest rise time without overshoot) for a 500-kΩ resistive load. C_0 compensation is slightly dependent on load resistance, as evidenced by the slight overshoot with 1 and 10 kΩ resistive loads.

Fig. 7.

Dynamic response for 50- μ s current pulses delivered to a well-hydrated fingertip electrode (see Fig. 6 legend) with nominal currents ranging 1–20 mA (sub-threshold to painful). The top trace shows the response for a 1-k Ω resistive load (showing only the rising phase for clarity; the falling phase is similar).

Fig. 8. Electrode voltage resulting from the 50- μ s, 1–20 mA current pulses shown in Fig. 7 (well-hydrated fingertip electrode). A X100 high voltage probe was used with the oscilloscope to record these data.

Fig. 9.

Maximal undistorted output current and voltage measured for resistive loads of 1, 10, 20, 30, 50, 70, 100, and 200 $k\Omega$. These values varied only minimally over pulse widths 20–250 μs and pulse rates 10–100 Hz. The output was primarily current-limited for low load resistance and voltage-limited for high load resistance. The legend shows the symmetrical $\pm HV$ rail voltages for the output circuit. The somewhat lower output limits for negative pulses and high rail voltages was not observed in simulation. We assume that this effect is due to stray capacitance in the assembled circuit.

Table 1

Specifications. Maximum ratings are based on component manufacturer's published ratings. Performance specifications are measured; see details in main text and Figures 6-9.

	Specification	Notes	Value	Units
INPUT	Transconductance	Knob and jumper adjustable	0–20	mA/V
	Minimum Input Voltage	To produce up to ± 20 mA at output	± 1	V
	Maximum Input Voltage	To prevent damage	± 40	V
	Input Resistance		10	k Ω
OUTPUT	Output charge limit	For user safety	160	μ C
	Maximum Current	Nominal; see derating curves	± 20	mA
	Output Resistance	Without R_o compensation feedback	717	k Ω
		With R_o compensation feedback	8.8	M Ω
	Rise Time	For 100 k Ω load	< 10	μ s
	Slew rate	For 100 k Ω load	45	V/ μ s
	Maximum Voltage	Asymmetric and depends on load and rail voltages;	+800,	V
		see derating curves	–600	V
	Overshoot	Step transition into 100 k Ω load	<5	%
	Settling Time	Step into 100 k Ω load, settle to 2% final value	22	μ s
	Load grounding	One side must be grounded	---	---
	Linearity	Percent deviation from delivered current	≤ 8	%
		Percent deviation from full scale output	≤ 1	%
	Pulse width	Maximum	250	μ s
	Duty cycle	Pulse width * rate	10	%
Power supply current (No input signal)		LV	25	mA
		HV	0.4	mA
Power supply current (100Hz, 100 μ s, 20mA bipolar signal)		LV	30	mA
		HV	1.7	mA